

# Application and measurement of ultra-low-noise oscillators

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## Introduction

Low phase noise signal sources are an essential component of many high-performance electronic systems, and range from single-frequency generators to multi-channel synthesisers. Common to almost all of these is that the output frequencies are derived from one or more crystal oscillators, by means including frequency multiplication and division, phase-locked loop (PLL), direct digital synthesis (DDS) etc.

An oven-controlled crystal oscillator (OCXO) is typically used as the 'master oscillator' from which all output frequencies are derived. It follows that the phase noise performance of the signal source is ultimately limited by that of the master oscillator.

The expression governing a synthesiser's phase noise is:

$$L(f_{\text{out}}) = L(f_{\text{ref}}) + 20\log(f_{\text{out}}/f_{\text{ref}}) \text{ dBc/Hz}$$

where  $f_{\text{out}}$  is the synthesiser's output frequency and  $f_{\text{ref}}$  is the master oscillator frequency.

This represents the theoretical ideal. In practice, the other components in the synthesiser will add noise, with the exception of higher-frequency oscillators locked to the master oscillator, which may have a lower far-out noise floor than the multiplied-up master oscillator.

The lowest-noise OCXOs available have phase noise floors of -185dBc/Hz or better. If the master oscillator frequency is in the region of 100MHz, then the optimum loop bandwidth for locking VCOs is likely to be >1MHz. The OCXO's performance is therefore critical to the synthesiser's phase noise from very low offsets out to 1MHz and beyond.

SC-cut crystals are generally favoured for high-performance OCXOs. Compared with AT-cut crystals they can be driven at higher power, offer lower close-in phase noise, age more slowly and have lower  $df/dT$  close to the turnover temperature.

As a general rule, the bigger the crystal, the lower its phase noise and aging rate. For this reason, overtone crystals are typically preferred in high-performance oscillators. In the 80-150MHz range, 5th-overtone crystals are commonly used.

The best crystal oscillators are capable of very low phase noise, but a number of pitfalls await the unwary system designer, any of which can seriously degrade the oscillator's performance. Some of the more common ones are discussed below.

## Tune voltage noise

With 100MHz 5th-overtone SC-cut crystals, tuning range of a few parts per million can readily be achieved without compromising the phase noise. The oscillators in Pascall's OCXO, OCXOF and XMN series have a guaranteed minimum range of  $\pm 6$  ppm with respect to the specified frequency. With long-term aging typically in the region of  $10^{-7}$  / year, this provides for many decades of use.

In ultra-low-noise oscillator designs, the close-to-carrier phase noise is generally dominated by the crystal's noise. Crystal selection is essential for low phase noise, as there can be more than 20dB variation even within a single batch of crystals. Although the guaranteed performance of the highest specification 100MHz Pascall oscillators is -137dBc/Hz at 100Hz offset they typically achieve better than -140dBc/Hz, with the best crystals closer to -150dBc/Hz, as illustrated in Figure 1.

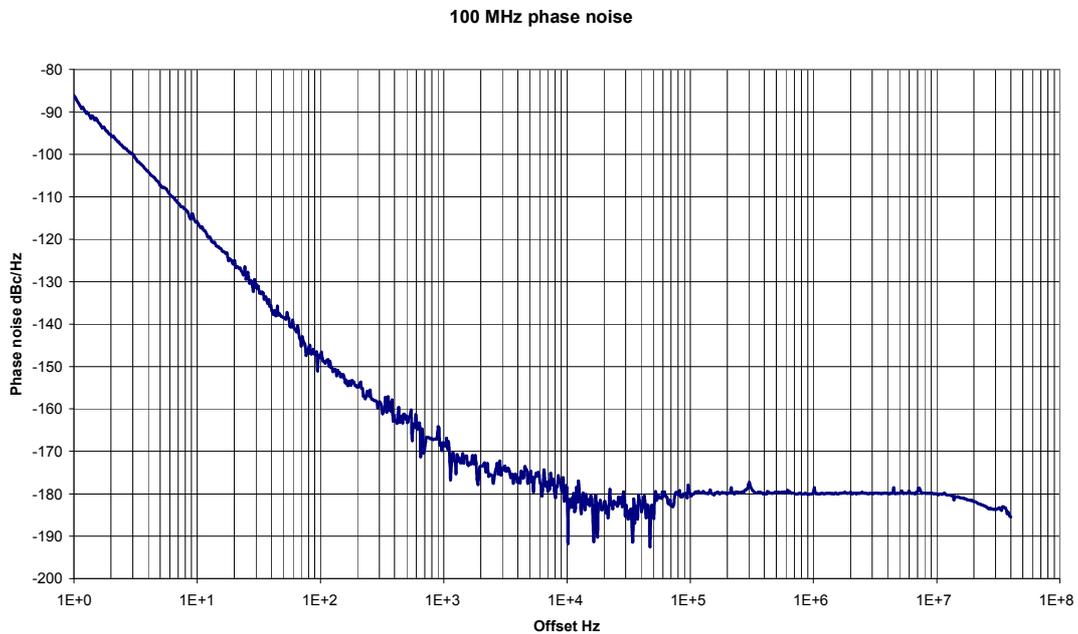


Figure 1

100MHz phase noise

This phase noise is achieved together with more than 20ppm total tuning range, with corresponding tune sensitivity typically in the range 2 to 2.5 ppm/V ( $\equiv 200$  to 250 Hz/V). With this combination of extremely low phase noise and moderately high tune sensitivity, care is needed to ensure that the oscillator's performance isn't degraded by noise on the tune input.

For a voltage-tuned oscillator, single sideband (SSB) phase noise due to noise at the tune input

$$= 10\log(V_n^2 K_o^2 / 2f_m^2) \text{ dBc/Hz} \quad (1)$$

where  $V_n$  is the RMS noise voltage per  $\sqrt{\text{Hz}}$ ,  $K_o$  is the tuning sensitivity in Hz/V and  $f_m$  is the offset frequency in Hz.

Depending on how the oscillator is tuned, common sources of noise may typically include voltage references, regulators, op amps and resistors.

If the tune voltage comes from a voltage reference or LDO regulator for example, the noise voltage density at 1Hz might typically be of the order of  $1\mu\text{V}/\sqrt{\text{Hz}}$ . If this is applied to an oscillator with  $200\text{Hz}/\text{V}$  tuning sensitivity, the resultant phase noise at 1Hz offset will be:

$$10\log((1\text{E-}6)^2 \times 200^2 / 2) = -77\text{dBc}/\text{Hz}$$

In the case of the oscillator shown in Figure 1, this would increase the phase noise by  $\sim 10\text{dB}$  at 1Hz offset.

It is therefore important to consider the impact of voltage sources at the design stage. If a potentiometer is used to tune the oscillator, the wiper - track contact may be an additional source of noise. If fast tuning isn't a requirement, the simplest solution is often to use an RC filter to reduce the added phase noise to an insignificant level over the offset range of interest.

The master oscillator in a synthesiser may be locked to an external reference. In this case, the effect of the PLL components must be taken into account.

If a PLL IC is used, its phase noise contribution can be estimated based on its specified normalised phase noise floor and  $1/f$  noise. The semiconductor manufacturers generally provide software to model PLL phase noise. However, with currently available ICs, setting the loop bandwidth much above 1Hz will degrade the phase noise of the best 100MHz OCXOs.

Improved performance may be obtained with a PLL built using high-speed CMOS, or for lowest noise an analogue loop. Care must be taken with the loop filter design to ensure that its noise doesn't limit the performance. The optimum loop bandwidth is unlikely to be more than a few 10s of Hz, so when choosing an op amp, its  $1/f$  noise will often be more relevant than its noise floor.

Tuning the oscillator from too high a source resistance will also increase its phase noise. The noise voltage density of a resistor is  $\sqrt{(4kTR)} \text{ V}/\sqrt{\text{Hz}}$ , where  $k$  is Boltzmann's constant and  $T$  is the temperature in Kelvin.

Applying this to equation (1), the added phase noise due to the resistor is:

$$10\log((4kTR) \times K_o^2 / 2f_m^2) \text{ dBc}/\text{Hz}$$

Figure 2 shows the calculated effect of various values of source resistance on an oscillator with the phase noise of Figure 1 and  $200\text{Hz}/\text{V}$  tune sensitivity.

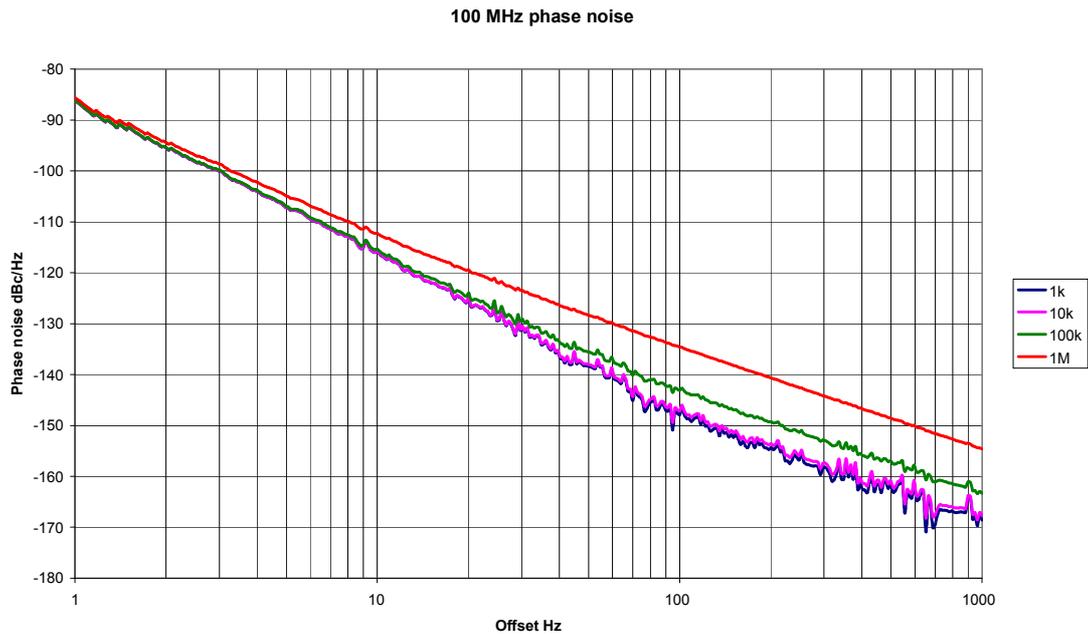


Figure 2 Effect of source resistance on phase noise

White noise at the tune input produces a  $-20\text{dB} / \text{decade}$  phase noise slope. The inherent slope of the OCXO below 100Hz offset is  $-30\text{dB} / \text{decade}$ , so source resistance is less significant at lower offsets. If high source resistance is unavoidable, fitting a capacitor at the tune input to attenuate noise above a few 10s of Hz will normally solve the problem. In a PLL it can be made the last element of the loop filter.

When designing the loop filter and oscillator tuning circuitry, don't forget resistor  $1/f$  noise, which is generated whenever there is DC current in a resistor. Increasing the resistance and / or increasing the voltage across a resistor will increase the  $1/f$  noise voltage.

Resistor types vary widely in their levels of flicker noise [1]. To illustrate this, the circuit shown in Figure 3 was used to tune a low-noise 100MHz oscillator. The close-in phase noise was measured with 2 different types of 0603-size resistors used for R1 and R2; thick film and metal film.

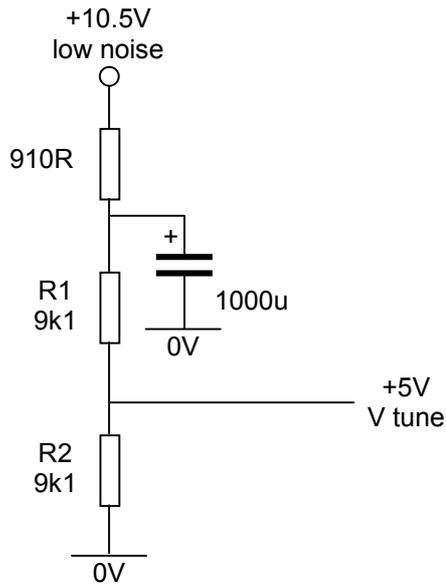


Figure 3 Resistor flicker noise demonstration circuit

The plots in Figure 4, measured using the circuit of Figure 3, give an indication of the typical performance degradation that can be caused by resistor flicker noise. Phase noise was measured with the resistors at +25°C and +85°C; the oscillator was at room temperature for all the measurements.

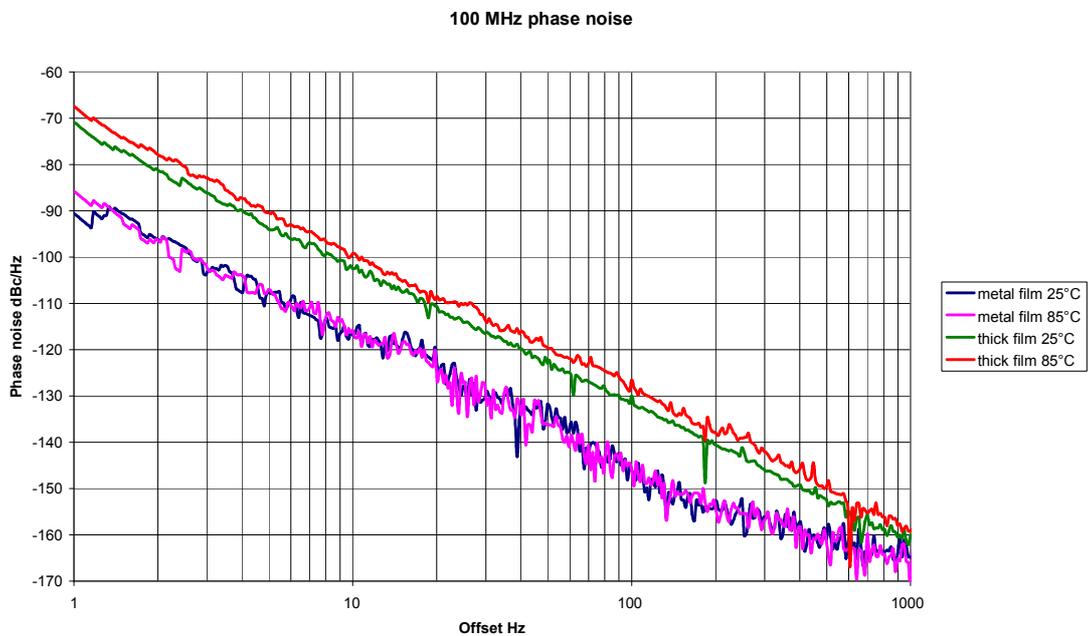


Figure 4 Effect of resistor flicker noise on phase noise

Beyond ~100Hz offset the metal film measurements are limited somewhat by the test set (Agilent E5052B with 100 correlations), but still clearly show the effect of the thick film resistors.

The oscillator's tune sensitivity was ~215 Hz/V at 5V. Using equation (1):

$$L(f_m) = 10\log(V_n^2 K_o^2 / 2f_m^2) \text{ dBc/Hz},$$

the phase noise measurements indicate resistor noise voltage density (at 1Hz) in the region of  $2\mu\text{V}/\sqrt{\text{Hz}}$  at 25°C and  $3\mu\text{V}/\sqrt{\text{Hz}}$  at 85°C for the thick film resistors. By contrast, the expression  $\sqrt{(4kTR)}$  predicts only  $\sim 9\text{nV}/\sqrt{\text{Hz}}$  white noise for 4.55k ohm. This demonstrates that care should be taken with the choice of DC-carrying resistors when their low-frequency noise can significantly affect the oscillator's phase noise. It has been observed that higher power (i.e. physically larger) resistors of a given type and value exhibit lower levels of flicker noise [1].

### **Power supply and ground**

Low-noise OCXOs generally include internal regulators which provide enough low-frequency ripple rejection to reduce input supply noise to an insignificant level. However, they are less effective at rejecting signals in the >100 kHz range typical of modern switchmode power supplies. It is therefore advisable to add external LC filtering when powering an OCXO from a switching supply.

It is preferable to have a series choke as the first element, in order to limit ground ripple current. If a shunt capacitor is fitted first there will be large ground currents, which may cause significant ground voltages within the oscillator, resulting in modulation of its output signal. (Oscillators will typically have  $\geq 1\mu\text{F}$  directly across the supply input as part of the regulator circuit.)

OCXOs typically use either thermistors or a low-power analogue temperature sensor in a feedback loop, in order to control the heater current and hence maintain constant oven temperature. The temperature sensing devices can be quite noisy, giving rise to significant noise current drawn from the supply.

The OCXO's internal layout will generally be designed to minimise ground noise current within the oscillator. However, it is important that any ground impedance common to the supply and tune inputs be kept low, to prevent this noise being superimposed on the tune voltage. As an extra precaution, it is advisable to place any low-pass filtering of the tune voltage close to the tune input.

### **Output load**

Ultra-low-noise OCXOs are generally designed to drive into 50 ohms. The phase noise performance may be degraded by grossly incorrect load impedance; high impedance in particular may cause the output stage to saturate, leading to increased noise floor and greater susceptibility to external interference, such as supply ripple.

## Frequency multiplication

In many applications, an OCXO is multiplied up to a higher frequency. In frequency synthesisers this is then typically mixed with lower frequencies, to give the required output frequency range and step size. With this architecture, the synthesiser's phase noise is primarily dependent on the master oscillator and multiplier.

In principle, the phase noise of a multiplied signal

$$= L(f_{\text{ref}}) + 20\log(N) \text{ dBc/Hz}$$

where  $L(f_{\text{ref}})$  is the phase noise of the input signal and  $N$  is the multiplication ratio.

In practice, achieving this relationship is far from trivial, particularly when the input signal has very low noise. There is an unavoidable conversion loss associated with passive multiplication, followed by further loss in the bandpass filter needed before the multiplied frequency can be amplified to the desired level. If the signal level from the filter is too low, the noise floor will then be dominated by the following amplifier.

This problem can be alleviated by driving the multiplier with higher power. However, even this adds noise. Consider an oscillator with  $-185\text{dBc/Hz}$  phase noise floor. If the multiplier drive amplifier's phase noise floor is also  $-185\text{dBc/Hz}$ , the signal will already be degraded by 3dB before it goes into the multiplier.

Choice of amplifiers is important. GaAs devices for example, can have very low noise figures, but tend to exhibit high levels of flicker-of-phase noise. This can increase the phase noise floor out to several MHz offset. Increasing the input power will not reduce this flicker noise.

Silicon or InGaP bipolar MMICs typically have lower flicker noise. However, driving them in or close to compression will increase their noise floor; driving them into saturation has been observed to cause a severe increase in flicker noise. MMICs with internal active bias circuitry appear to be somewhat worse in this respect than traditional resistively biased devices.

Active multipliers typically have lower conversion loss, or may even have conversion gain. However, they are less suitable for very low noise applications, as the non-linearity which the active device must display in order to generate harmonics, inevitably increases its noise.

Pascall's XMN / XMNP signal sources provide a solution by integrating an ultra-low-noise OCXO with Schottky diode based multipliers and filters, offering output frequencies up to around 12GHz. Sub-harmonic products are very low, and an output is provided at the oscillator's fundamental frequency in addition to the multiplied output.

The phase noise floor at the multiplied output is typically equivalent to  $-183\text{dBc/Hz}$  when referred to the oscillator's frequency. As an example, Figure 5 shows the phase noise of a 120MHz + 2.4GHz Pascall signal source.

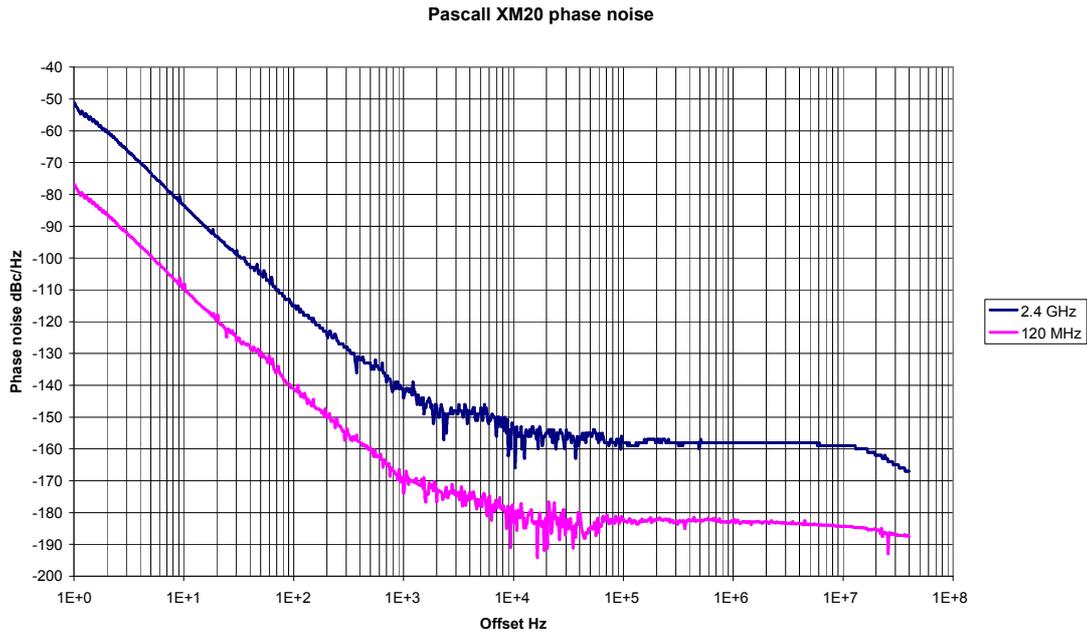


Figure 5 2400MHz and 120MHz XM20 phase noise

### Phase noise measurement

Measuring the performance of ultra-low-noise OCXOs can present problems. Most phase noise test systems use a PLL method as shown in Figure 6, in which the device under test (DUT) and a reference source at the same frequency are locked together. A double-balanced mixer is used as the phase comparator, and the two sources are locked in quadrature. In this condition, the mixer has maximum phase discrimination and minimum amplitude discrimination, and hence relatively little sensitivity to AM noise.

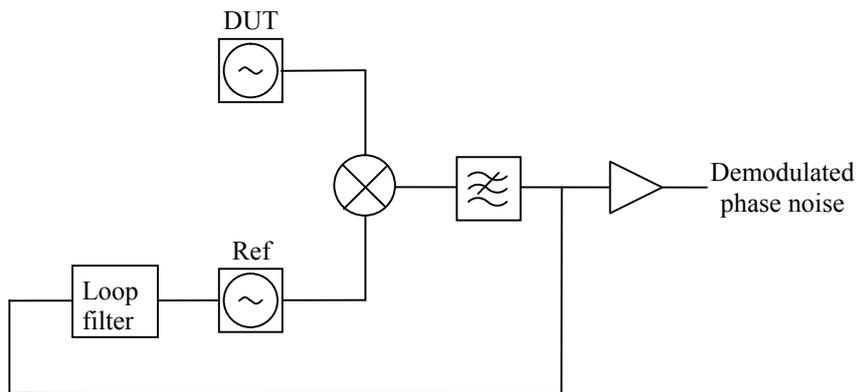


Figure 6 Phase noise measurement using PLL

The output of the mixer represents the phase difference between the DUT and the reference source. Inside the loop bandwidth the phase difference between the two

sources is reduced by the loop gain. Calibrating the phase detector constant and the tuning sensitivity enables the loop parameters to be calculated, and hence correction for loop gain can be applied.

Ideally the reference phase noise should be much lower than that of the DUT, so the system effectively measures only the phase noise of the DUT. However, in an integrated phase noise test set the reference source is generally a synthesiser, which will typically be much noisier than a low-noise OCXO. The traditional solution is to use a second oscillator, of the same type as the DUT, as the reference source.

If we assume that the outputs of the sources are uncorrelated, then the magnitude of the RMS phase difference between them is the same as the sum, and the measurement represents the combined phase noise of the two oscillators.

This leaves uncertainty, in that it isn't possible to identify the individual contribution of each oscillator. Adding a third oscillator and making 3 measurements (A+B, A+C, B+C) generates 3 simultaneous equations, which can be solved to give the phase noise of each oscillator. However, there is still a problem, in that the test system's own noise floor is most likely to be higher than that of a low-noise crystal oscillator. As it appears in each measurement, it isn't possible to obtain a measurement of the oscillator alone.

Over the past few years a number of dual-channel phase noise test sets have become available, in which the DUT signal is split and measured separately against two uncorrelated reference sources. Cross correlation is used to attenuate the reference and test system noise. Figure 7 illustrates the basic configuration.

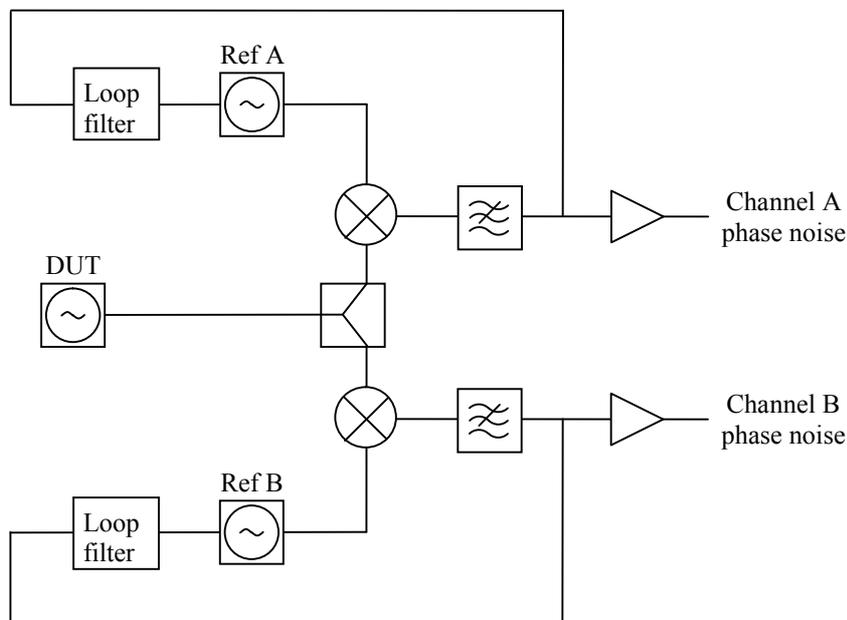


Figure 7 Phase noise measurement with cross correlation

This can be regarded as two independent test systems, each of which measures the combination of the DUT, a reference source and its own residual phase noise. As the two measurements take place concurrently, the DUT's phase noise is common to both channels. However, the reference and system noise is uncorrelated between channels. Cross correlation and averaging preserves the DUT noise while attenuating the reference and system noise [2].

Noise attenuation =  $5 \log(N)$  dB, where N is the number of correlations.

For example, 10 correlations gives 5dB noise reduction, 100 correlations gives 10dB, etc. Thus in principle, any reference or system noise can be reduced to insignificance, given enough correlations. The disadvantage is that every 5dB improvement increases the measurement time by a factor of 10; the effect of this is illustrated later.

To minimise the number of correlations required for a low-noise measurement, the reference sources should ideally have at least similar performance to the DUT. They also need to be voltage tuneable, to enable them to be phase locked.

Some phase noise test sets include synthesisers which can be used as the reference sources, removing the need for the user to provide two voltage-tuned oscillators. This simplifies the procedure, at the expense of increased measurement time. The Agilent E5052B for example, is a versatile signal source analyser with phase noise measurement as its main function.

In the E5052B and various other self-contained phase noise test sets, the relatively high far-from-carrier noise floor of the synthesisers is ameliorated by the ability to perform many correlations in a short time. At very low offsets, the synthesisers follow the profile of their low-noise reference oscillators. However, in the intermediate region, very long measurement times are needed to reduce the phase noise below that of a low-noise DUT. Even 10,000 correlations may not be enough.

Figure 8 illustrates this, using a 640MHz Pascall XM5P signal source, which combines a phase-locked 128MHz OCXO with a x5 multiplier and bandpass filters. The 'direct' plot was measured in the normal way on an E5052B, with 4,000 correlations, requiring 11 hours' measurement time. Even with so many correlations the plot shows anomalies at offsets between 3kHz and 150kHz.

The 'mixed' plot also used the E5052B, but with the 640MHz signal split 2 ways, each of which was mixed with a separate low-noise 600MHz source to give 40MHz IF. The 40MHz signals were then connected to the 2 channels of the E5052B. All 3 sources were locked to a single 10MHz reference with <1Hz loop bandwidth, so this did not affect the measured phase noise.

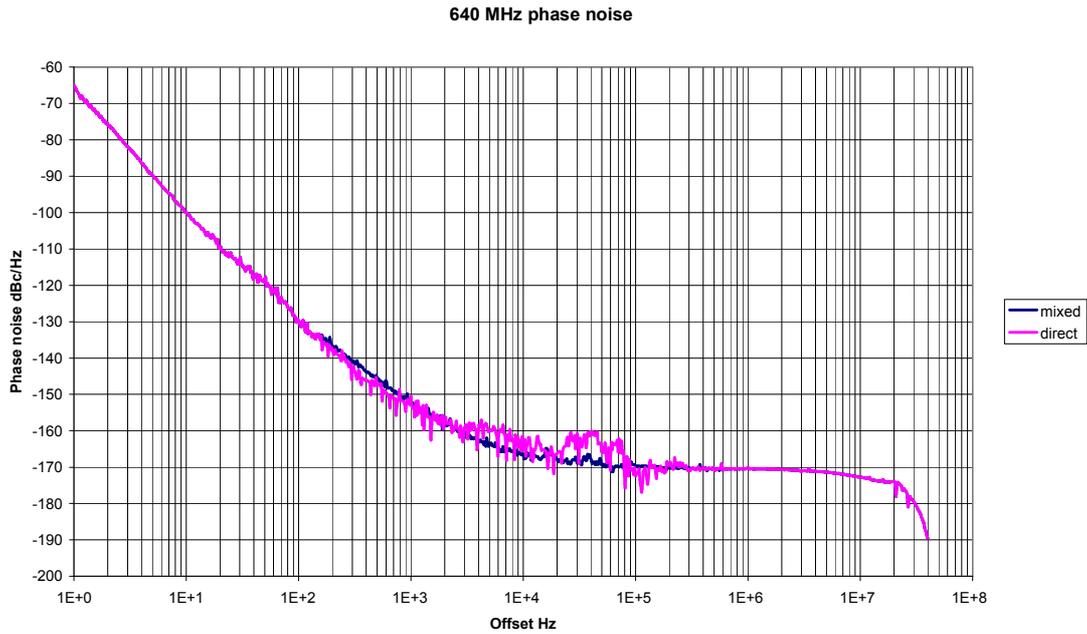


Figure 8 Pascall XM5P 640MHz phase noise measurements

With only 40MHz input frequency the E5052B's phase noise is much lower, and it was possible to measure the DUT accurately in less than 10 minutes. This illustrates the difficulty of measuring very low phase noise when relying on a test set's internal synthesisers.

Another potential source of error is noise on the tune voltage. Phase noise test sets often provide a programmable tune voltage source for the DUT. This will typically have a low noise floor, but may become much noisier at low frequencies. On the E5052B for example, the control voltage noise is  $1\text{nV}/\sqrt{\text{Hz}}$  at 10kHz. However, if it is used to tune an ultra-low-noise crystal oscillator, it can seriously degrade the performance at low offsets. Figure 9 shows its effect on a low-noise 100MHz oscillator.

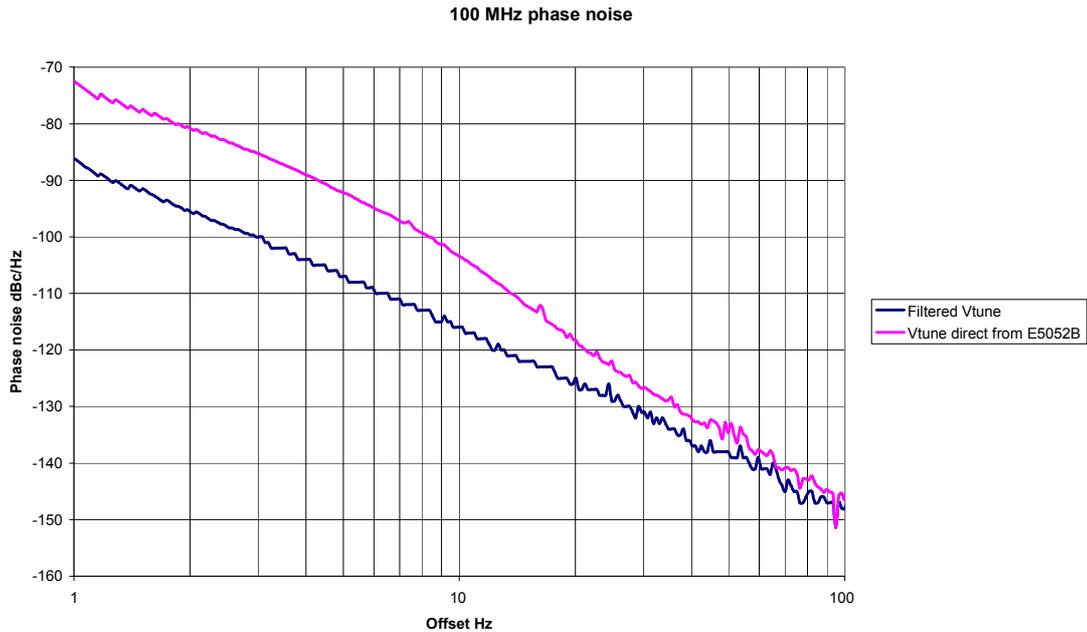


Figure 9 100MHz Pascall oscillator tuned by E5052B

Figure 9 shows a 100MHz Pascall oscillator with 5.25V tune voltage, at which its tune sensitivity was  $\sim 215\text{Hz/V}$ . The tune voltage was supplied by the DC CONTROL output of an E5052B. On one trace the E5052B output was connected directly to the oscillator's tune input; on the other it was first passed through an RC low-pass filter. Up to 15dB increase in phase noise can be seen when  $V_{\text{tune}}$  was supplied directly by the E5052B, compared with a low-noise filtered voltage.

For the DUT, this is easily fixed by RC filtering at the tune input. However, in the case of a 3-oscillator measurement, the same approach can't be used for the reference oscillators as it will change the PLL parameters, leading to gross inaccuracy and probably loop instability. Cross correlation will reduce the reference noise but might lead to unacceptably long measurement time.

One approach that has been used successfully for reference tuning, is to make a low-noise variable DC source and add it to a 'potted down' tune voltage output from the test set. The DC source is used to tune the oscillator on to frequency, ready for the test system to do the required calibration and close the loop. A 10:1 pot down for example, will give 20dB reduction in the close-in noise added by the test set.

At its simplest, this may be made from a bench power supply, 3 resistors and a capacitor, as in Figure 10.

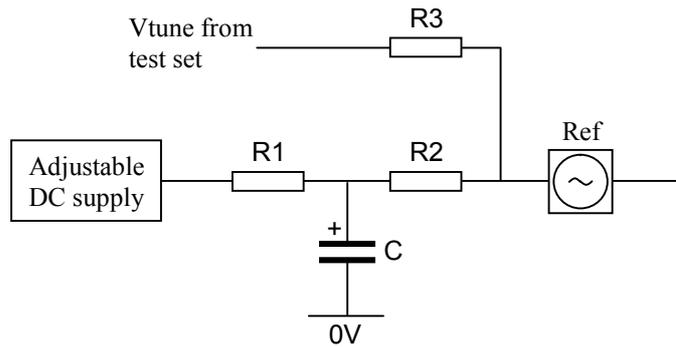


Figure 10 Tune voltage pot down

In the circuit of Figure 10, R1 and C filter the tune supply noise. If C has low reactance at the lowest frequency of interest, Vtune is attenuated by the potential divider formed by R2 and R3. To minimise calibration errors, R1 should be much less than R2. As an example, if R1 = 220 ohm, R2 = 2k2, R3 = 20k and C = 47000 $\mu$ F, noise on the DC supply will be attenuated by 36dB at 1Hz and test set Vtune noise attenuation will be 20dB. Resistors should be metal film types to minimise flicker noise.

For this circuit to give accurate results,  $1/(2\pi CR1)$  must be much less than the loop bandwidth and the lowest measurement frequency of interest. There is also a calibration error within the loop bandwidth, due to the finite ratio of R2:R1. With the values above, this could be up to 0.74dB. These potential problems may be eliminated by inserting a low-noise unity-gain buffer between R1/ C and R2.

## Conclusions

Designers of ultra-low-noise oscillators take great care to minimise the main causes of phase noise. However, it isn't possible to make an oscillator entirely immune to outside influences, so the system designer must pay attention to factors such as tune voltage noise, supply ripple and grounding if its full potential is to be realised. In addition, measuring very low phase noise requires care if the true performance of an oscillator is to be determined.

## References

- [1] F. Seifert: Resistor Current Noise Measurements, April, 2009
- [2] J. Breitbarth: Cross Correlation in Phase Noise Analysis, Microwave Journal, February 2011