Receiver Phase Noise Measurement

Phase Noise at Low Frequencies

No mixer has perfect port-to-port isolation, and some of a receiver's local-oscillator signal leaks through into the IF. If we tune a general-coverage receiver, with its antenna disconnected, to exactly 0 Hz, the local oscillator is exactly at the IF center frequency, and the receiver acts as if it is tuned to a very strong unmodulated carrier. A typical mixer might give only 40 dB of LO isolation and have an LO drive power of at least 10 mW. If we tune away from 0 Hz, the LO carrier tunes away from the IF center and out of the passband. The apparent signal level falls. Although this moves the LO carrier out of the IF passband, some of its noise sidebands will not be, and the receiver will respond to this energy as an incoming noise signal. To the receiver operator, this sounds like a rising noise floor as the receiver is tuned toward 0 Hz.

To get good noise floor at very low frequencies, some professional/military receivers, like the Racal RA1772, use very carefully balanced mixers to get as much port-to-port isolation as possible, and they also may switch a crystal notch filter into the first mixer's LO feed. Most general-coverage radios inhibit tuning in the LF or VLF region. It could be suggested by a cynic that how low manufacturers allow you to tune is an indication of how far they think their phase-noise sidebands could extend!

Phase Noise Measurements

There are several different ways of measuring phase noise, offering different tradeoffs between convenience, cost and effort. Some methods suit oscillators in isolation, others suit them in-situ (in their radios).

If you're unfamiliar with noise measurements, the units involved may seem strange. One reason for this is that a noise signal's power is spread over a frequency range, like an infinite number of infinitesimal sinusoidal components. This can be thought of as similar to painting a house. The area that a gallon of paint can cover depends on how thinly it's spread. If someone asks how much paint was used on some part of a wall, the answer would have to be in terms of paint volume per square foot. The wall can be considered to be an infinite number of points, each with an infinitesimal amount of paint applied to it. The question of what volume of paint has been applied at some specific point is unanswerable. With noise, we must work in terms of *power density*, of watts per hertz. We therefore express phase-noise level as a ratio of the carrier power to the noise's power density. Because of the large ratios involved, expression in decibels is convenient. It has been a convention to use *dBc* to mean "decibels with respect to the carrier."

For phase noise, we need to work in terms of a standard bandwidth, and 1 Hz is the obvious candidate. Even if the noise is measured in a different bandwidth, its equivalent power in 1 Hz can be easily calculated. A phase-noise level of –120 dBc in a 1-Hz bandwidth (often written as

"–120 dBc/Hz") translates into each hertz of the noise having a power of 10^{-12} of the carrier power. In a bandwidth of 3 kHz, this would be 3000 times larger.

The most convenient way to measure phase noise is to buy and use a commercial phase noise test system. Such a system usually contains a state-of-the-art, low-noise frequency synthesizer and a low-frequency spectrum analyzer, as well as some special hardware. Often, a second, DSP-based spectrum analyzer is included to speed up and extend measurements very close to the carrier by using the Fast Fourier Transform (FFT). The whole system is then controlled by a computer with proprietary software. With a good system like this costing about $100,000, this is not a practical method for amateurs, although a few fortunate individuals have access to them at work. These systems are also overkill for our needs, because we are not particularly interested in determining phase- noise levels very close to and very far from the carrier.

It's possible to make respectable receiver-oscillator phase-noise measurements with less than $100 of parts and a multimeter. Although it's time-consuming, the technique is much more in keeping with the amateur spirit than using a $100k system! An ordinary multimeter will produce acceptable results; a meter capable of indicating "true RMS" ac voltages is preferable because it can give correct readings on sine waves *and* noise. **Figure 1** shows the setup. Measurements can only be made around the frequency of the crystal oscillator, so if more than one band is to be tested, crystals must be changed, or else a set of appropriate oscillators is needed. The oscillator should produce about +10 dBm (10 mW) and be *very* well shielded. (To this end, it's advisable to build the oscillator into a die-cast box and power it from internal batteries. A noticeable shielding improvement results even from avoiding the use of an external power switch; a reed-relay element inside the box can be positioned to connect the battery when a small permanent magnet is placed against a marked place outside the box.)

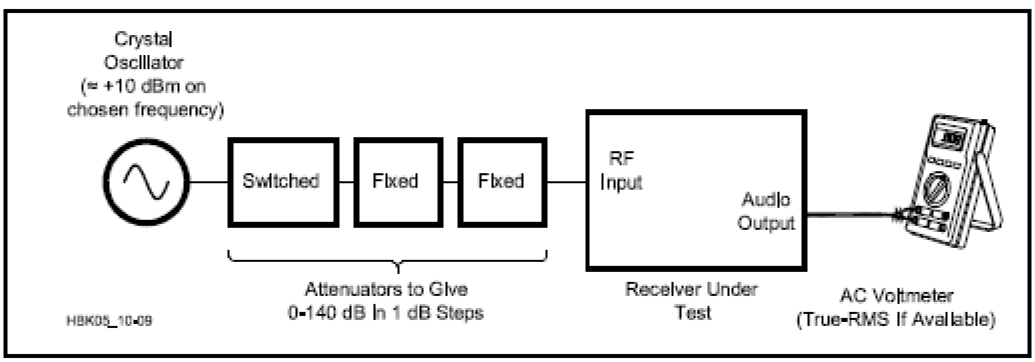

Figure 1 — Setup for measuring receiver-oscillator phase noise.

Likewise, great care must be taken with attenuator shielding. A total attenuation of around 140 dB is needed, and with so much attenuation in line, signal leakage can easily exceed the test signal that reaches the receiver. It's not necessary to be able to switch-select all 140 dB of the attenuation, nor is this desirable, as switches can leak. All of the attenuators' enclosure seams must be soldered. A pair of boxes with 30 dB of fixed attenuation each is needed to complete the set. With 140 dB of attenuation, coax cable leakage is also a problem. The only countermeasure against this is to minimize all cable lengths and to interconnect test-system modules with BNC plug-to-plug adapters (UG-491A) where possible.

Ideally, the receiver could simply be tuned across the signal from the oscillator and the response measured using its signal-strength (S) meter. Unfortunately, receiver S meters are notoriously imprecise, so an equivalent method is needed that does not rely on the receiver's AGC system.

The trick is not to measure the response to a fixed level signal, but to measure the changes in applied signal power needed to give a fixed response. Here is a step-by-step procedure based on that described by John Grebenkemper, KI6WX, in March and April 1988 *QST*:

1. Connect the equipment as shown in Figure 1, but with the crystal oscillator off. Set the step attenuator to maximum attenuation. Set the receiver for SSB or CW reception with its narrowest available IF filter selected. Switch out any internal preamplifiers or RF attenuators. Select AGC off, maximum AF and RF gain. It may be necessary to reduce the AF gain to ensure the audio amplifier is at least 10 dB below its clipping point. The ac voltmeter or an oscilloscope on the AF output can be used to monitor this.

2. To measure noise, it is important to know the bandwidth being measured. A true-RMS ac voltmeter measures the power in the noise reaching it. To calculate the noise density, we need to divide by the receiver's *noise bandwidth*. The receiver's –6-dB IF bandwidth can be used as an approximation, but purists will want to plot the top 20 dB of the receiver's bandwidth on linear scales and integrate the area under it to find the width of a rectangle of equal area and equal height. This accounts properly for the noise in the skirt regions of the overall selectivity. (The very rectangular shape of common receiver filters tends to minimize the error of just taking the approximation.)

Switch on the test oscillator and set the attenuators to give an AF output above the noise floor and below the clipping level with the receiver peaked on the signal. Tune the receiver off to each side to find the frequencies at which the AF voltage is half that at the peak. The difference between these is the receiver's –6-dB bandwidth. High accuracy is not needed: 25% error in the receiver bandwidth will only cause a 1-dB error in the final result. The receiver's published selectivity specifications will be close enough. The benefit of integration is greater if the receiver has a very rounded, low-ringing or low-order filter.

3. Retune the receiver to the peak. Switch the oscillator off and note the noise-floor voltage. Turn the oscillator back on and adjust the attenuator to give an AF output voltage 1.41 times (3 dB) larger than the noise floor voltage. This means that the noise power and the test signal power at the AF output are equal — a value that's often called the *MDS* (*minimum discernible signal*) of a receiver. Choosing a test-oscillator level at which to do this test involves compromise. Higher levels give more accurate results where the phase noise is high, but limit the lowest level of phase noise that can be measured because better receiver oscillators require a greater input signal to produce enough noise to get the chosen AF-output level. At some point, either we've

taken all the attenuation out and our measurement range is limited by the test oscillator's available power, or we overload the receiver's front end, spoiling the results.

Record the receiver frequency at the peak, (f_0), the attenuator setting (A_0) and the audio output voltage (V_0). These are the carrier measurements against which all the noise measurements will be compared.

4. Now you must choose the offset frequencies — the separations from the carrier — at which you wish to make measurements. The receiver's skirt selectivity will limit how close to the carrier noise measurements can be made. (Any measurements made too close in are valid measurements of the receiver selectivity, but because the signal measured under these conditions is sinusoidal and not noise like, the corrections for noise density and noise bandwidth are not appropriate.) It is difficult to decide where the filter skirt ends and the noise begins, and what corrections to apply in the region of doubt and uncertainty. A good practical approach is to listen to the audio and tune away from the carrier until you can't distinguish a tone in the noise. The ear is superb at spotting sine tones buried in noise, so this criterion, although subjective, errs on the conservative side.

Tune the receiver to a frequency offset from f_0 by your first chosen offset and adjust the attenuators to get an audio output voltage as close as possible to V_0. Record the total attenuation, A_1 and the audio output voltage, V_1. The SSB phase noise (qualified as SSB because we're measuring the phase noise on only one side of the carrier, whereas some other methods cannot segregate between upper and lower noise sidebands and measure their sum, giving DSB phase noise) is now easy to calculate:

$$L(f) = A_1 - A_0 - 10_{\log} (BW_{noise}) \qquad (1)$$

where

$L(f)$ = SSB phase noise in dBc/Hz
BW_{noise} = receiver noise bandwidth, Hz
A_0 = Attenuator setting in step three
A_1 = Attenuator setting in step four

This equation begins with the difference between the attenuation necessary to reduce the peak carrier signal to the MDS level (A_0 in step three) and the attenuation necessary to reduce the phase noise to the MDS level (A_1 in step four). Subtracting the bandwidth correction term results in the noise power per Hz of bandwidth with respect to the peak carrier signal. Note that this equation does not depend on the absolute power level of the carrier signal as long as it remains constant during the test.

5. It's important to check for overload. Decrease the attenuation by 3 dB, and record the new audio output voltage, V_2. If all is well, the output voltage should increase by 22% (1.8 dB); if the receiver is operating nonlinearly, the increase will be less. (An 18% increase is still acceptable for the overall accuracy we want.) Record V_2/V_1 as a check: a ratio of 1.22:1 is ideal, and anything less than 1.18:1 indicates a bad measurement.

If too many measurements are bad, you may be overdriving the receiver's AF amplifier, so try reducing the AF gain and starting again back at Step 3. If this doesn't help, reducing the RF gain and starting again at Step 3 should help if the compression is occurring late in the IF stages.

6. Repeat Steps 4 and 5 at all the other offsets you wish to measure. If measurements are made at increments of about half the receiver's bandwidth, any discrete (non-noise) spurs will be found. A noticeable tone in the audio can indicate the presence of one of these. If it is well clear of the noise, the measurement is valid, but the noise bandwidth correction should be ignored, giving a result in dBc.

Table 1 shows the results for an ICOM IC-745 as measured by KI6WX, and **Figure 2** shows this data in graphic form. His oscillator power was only –3 dBm, which limited measurements to offsets less than 200 kHz. More power might have allowed noise measurements to lower levels, although receiver overload places a limit on this. This is not important, because the real area of interest has been thoroughly covered. When attempting phase-noise measurements at large offsets, remember that any front-end selectivity, before the first mixer, will limit the maximum offset at which LO phase-noise measurement is possible.

Table 1
SSB Phase Noise of ICOM IC-745 Receiver Section
Oscillator output power = –3 dBm (0.5 mW)
Receiver bandwidth (Δf) = 1.8 kHz
Audio noise voltage = –0.070 V
Audio reference voltage (V_0) = 0.105 V
Reference attenuation (A_0) = 121 dB

Offset Frequency	*Attenuation (A_1) (dB)*	*Audio V_1 (volts)*	*Audio V_2 (volts)*	*Ratio V_2/V_1*	*SSB Phase Noise (kHz) (dBc/Hz)*
4	35	0.102	0.122	1.20	–119
5	32	0.104	0.120	1.15	–122*
6	30	0.104	0.118	1.13	–124*
8	27	0.100	0.116	1.16	–127*
10	25	0.106	0.122	1.15	–129*
15	21	0.100	0.116	1.16	–133*
20	17	0.102	0.120	1.18	–137
25	14	0.102	0.122	1.20	–140
30	13	0.102	0.122	1.20	–141
40	10	0.104	0.124	1.19	–144
50	8	0.102	0.122	1.20	–146
60	6	0.104	0.124	1.19	–148
80	4	0.102	0.126	1.24	–150
100	3	0.102	0.126	1.24	–151
150	3	0.102	0.124	1.22	–151
200	0	0.104			–154
250	0	0.100			–154
300	0	0.98			–154
400	0	0.96			–154
500	0	0.96			–154
600	0	0.97			–154
800	0	0.96			–154
1000	0	0.96			–154

*Asterisks indicate measurements possibly affected by receiver overload (see text).

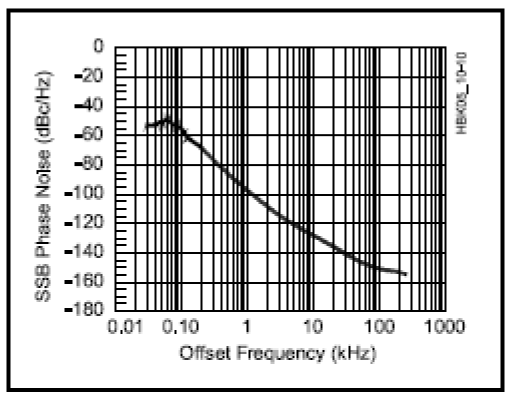

Figure 2 — The SSB phase noise of an ICOM IC-745 transceiver (serial number 01528) as measured by KI6WX.

Measuring Oscillator and Transmitter Phase Noise

Measuring the composite phase noise of a receiver's LO requires a clean test oscillator. Measuring the phase noise of an incoming signal, whether from a single oscillator or an entire transmitter, requires the use of a clean receiver, with lower phase noise than the source under test. The sidebar in the *Handbook's* **Oscillators and Synthesizers** chapter, "Transmitter Phase-Noise Measurement in the ARRL Lab," details the method used to measure composite noise (phase noise and amplitude noise, the practical effects of which are indistinguishable on the air) for *QST* Product Reviews. Although targeted at measuring high power signals from entire transmitters, this approach can be used to measure lower-level signals simply by changing the amount of input attenuation used.

At first, this method — using a low-frequency spectrum analyzer and a low-phase-noise signal source — looks unnecessarily elaborate. A growing number of radio amateurs have acquired good-quality spectrum analyzers for their shacks since older model Tektronix and Hewlett-Packard (now Agilent) instruments have started to appear on the surplus market at affordable prices. The obvious question is, "Why not just use one of these to view the signal and read phase-noise levels directly off the screen?" Reciprocal mixing is the problem.

Very few spectrum analyzers have clean enough local oscillators not to completely swamp the noise being measured. Phase-noise measurements involve the measurement of low-level components very close to a large carrier, and that carrier will mix the noise sidebands of the *analyzer's* LO into its IF. Some way of notching out the carrier is needed, so that the analyzer need only handle the noise sidebands. A crystal filter could be designed to do the job, but this would be expensive, and one would be needed for every different oscillator frequency to be tested. The alternative is to build a direct-conversion receiver using a clean LO like the Hewlett-Packard HP8640B signal generator and spectrum-analyze its "audio" output with an audio analyzer. This scheme mixes the carrier to dc; the LF analyzer is then ac-coupled, and this removes the carrier. The analyzer can be made very sensitive without overload or reciprocal mixing being a problem.

The remaining problem is then keeping the LO — the HP8640B in this example — at exactly the carrier frequency. 8640s are based on a shortened-UHF-cavity oscillator and can drift a little. The oscillator under test will also drift. The task is therefore to make the 8640B track the oscillator under test. For once we get something for free: The HP8640B's FM input is dc coupled, and we can use this as an electronic fine-tuning input. As a further bonus, the 8640B's FM deviation control acts as a sensitivity control for this input. We also get a phase detector for free, as the mixer output's dc component depends on the phase relationship between the 8640B and the signal under test (remember to use the dc coupled port of a diode ring mixer as the output). Taken together, the system includes everything needed to create a crude phase-locked loop that will automatically track the input signal over a small frequency range. **Figure 3** shows the arrangement.

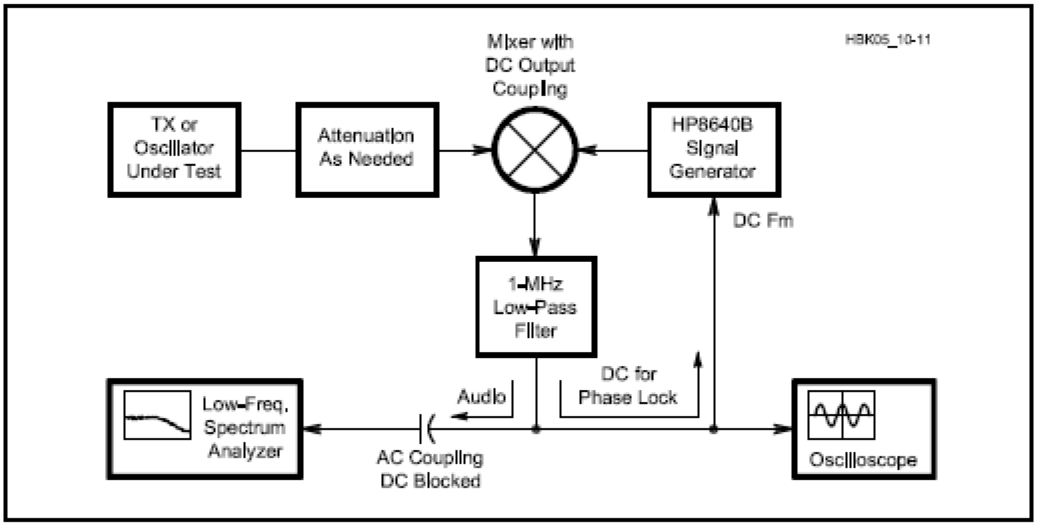

Figure 3 — Arrangement for measuring phase noise by directly converting the signal under test to audio. The spectrum analyzer views the signal's noise sidebands as audio; the signal's carrier, converted to dc, provides a feedback signal to phase-lock the Hewlett-Packard HP8640B signal generator to the signal under test.

The oscilloscope is not essential for operation, but it is needed to adjust the system. With the loop unlocked (8640B FM input disconnected), tune the 8640 off the signal frequency to give a beat at the mixer output. Adjust the mixer drive levels to get an undistorted sine wave on the scope. This ensures that the mixer is not being overdriven. While the loop is off-tuned, adjust the beat to a frequency within the range of the LF spectrum analyzer and use it to measure its level, "A_c" in dBm. This represents the carrier level and is used as the reference for the noise measurements. Connect the FM input of the signal generator, and switch on the generator's dc FM facility. Try a deviation range of 10 kHz to start with. When you tune the signal generator toward the input frequency, the scope will show the falling beat frequency until the loop jumps into lock. Then it will display a noisy dc level. Fine tune to get a mean level of 0 V. (This is a very-low-bandwidth, very-low-gain loop. Stability is not a problem; careful loop design is not needed. We actually want as slow a loop as possible; otherwise, the loop would track and cancel the slow components of the incoming signal's phase noise, preventing their measurement.)

When you first take phase-noise plots, it's a good idea to duplicate them at the generator's next lower FM-deviation range and check for any differences in the noise level in the areas of interest. Reduce the FM deviation range until you find no further improvement. Insufficient FM deviation range makes the loop's lock range narrow, reducing the amount of drift it can compensate. (It's sometimes necessary to keep gently trimming the generator's fine tune control.)

Set up the LF analyzer to show the noise. A sensitive range and 100-Hz resolution bandwidth are appropriate. Measure the noise level, "A_n" in dBm. We must now calculate the noise density that this represents. Spectrum-analyzer filters are normally *Gaussian*-shaped and bandwidth-specified at their –3-dB points. To avoid using integration to find their true-noise power bandwidth, we can reasonably assume a value of 1.2 × BW. A spectrum analyzer logarithmically compresses its IF signal ahead of the detectors and averaging filter. This affects the statistical distribution of noise voltage and causes the analyzer to read low by 2.5 dB. To produce the same scale as the ARRL Lab photographs prior to May 2006 *QST*, the analyzer reference level must be set to –60 dBc/Hz, which can be calculated as:

$$A_{ref} = A_c - 10 \log (1.2 \times BW) + 62.5 \text{ dBm} \qquad (2)$$

where

A_{ref} = analyzer reference level, dBm
A_c = carrier amplitude, dBm

This produces a scale of –60 dBc/Hz at the top of the screen, falling to –140 dBc/Hz at the bottom. The frequency scale is 0 to 20 kHz with a resolution bandwidth (BW in the above equation) of 100 Hz. This method combines the power of *both* sidebands and so measures DSB phase noise. To calculate the equivalent SSB phase noise, subtract 3 dB for non-coherent noise (the general "hash" of phase noise) and subtract 6 dB for coherent, discrete components (that is, single-frequency spurs). This can be done by setting the reference level 3 to 6 dB higher.

Low-Cost Phase Noise Testing

All that expensive equipment may seem far beyond the means of the average Amateur Radio experimenter. With careful shopping and a little more effort, alternative equipment can be put together for pocket money. (All of the things needed — parts for a VXO, a surplus spectrum analyzer and so on — have been seen on sale cheap enough to total less than $100.) The HP8640B is good and versatile, but for use at one oscillator frequency, you can build a VXO for a few dollars. It will only cover one oscillator frequency, but a VXO can provide even better phase-noise performance than the 8640B. There is free software available so that you can use your PC soundcard as an LF spectrum analyzer, though you may want to add a simple preamp and some switched attenuators.

Pontius has also demonstrated that signal-source phase-noise measurements can be accurately obtained without the aid of expensive equipment. (B. E. Pontius, "Measurement of Signal Source Phase Noise with Low-Cost Equipment," *QEX*, May-Jun 1998, pp 38-49.)