**Jacob Makhinson,** *N6NWP*
1100 N. Sunset Canyon Drive
Burbank, California 91504

# "DEMPHANO"

## *A device for measuring phase noise*

Phase noise, and its effects on receiver performance, has received a great deal of attention in professional and amateur radio literature in the past few years. *The ARRL Handbook* provides a fine treatise on this subject.[1]

While frequency synthesizers are becoming common in all commercial amateur radio transceivers and some home-built equipment, for some stringent high-performance applications requiring high dynamic range, a free-running LC VFO may still be the device of choice.

Defining the phase noise requirements of the VFO is an essential task in the initial receiver design stage. An oscillator with a high level of close-in phase noise may reduce the receiver's ability to separate closely spaced signals. In addition, due to the so-called reciprocal mixing,[1] the noise sidebands of the oscillator can mix with strong off-channel signals causing in-band interference. This interference may overwhelm the weak signal, essentially causing desensitization of the receiver. On the other hand, a high level of far-out phase noise may raise the receiver's noise floor at the receiving frequency causing degradation of the receiver's dynamic range.

The purpose of this article is to provide the designer with a practical method of accurate phase noise measurement. An example illustrates the method for specifying the phase noise requirements for a VFO. A free-running LC VFO has been designed as a part of a portable 20-meter band transceiver. A 12-pole crystal filter described in **Reference 2** provides adjacent-signal rejection of 100 dB at 2.2-kHz offset. From **References 3** and **4,** the maximum acceptable close-in phase noise level at 2.2 kHz offset are:

$$L_c = P - 10\log(BW_{-100}) = -100 - 10\log(4400) = -136 \text{ dBc/Hz} \quad (1)$$

where:

$L_c$ = VFO close-in phase noise spectral noise density in dBc/Hz at offset $BW_{-100}/2$ Hz

$P$ = crystal filter rejection in dB at offset $BW_{-100}/2$ Hz

$BW_{-100}$ = crystal filter bandwidth in Hz at the specified rejection level

At larger frequency offsets, the phase noise level should be such that the phase-noise-governed dynamic range (PNDR) is equal or better to the spurious-free dynamic range (SFDR).

Assuming the PNDR equals the SFDR at 110 dB in a 2.5-kHz IF noise bandwidth, from **Reference 3** the maximum acceptable far-out phase noise level is:

$$L_f = -SFDR - 10\log(BW) = -110 - 34 = -144 \text{ dBc/Hz} \quad (2)$$

where:

$L_f$ = VFO far-out phase noise spectral density in dBc/Hz at large frequency offsets (100 kHz)

$BW$ = estimated IF noise bandwidth in Hz

## Block diagram

Phase noise measurement techniques are described in detail in the literature.[5,6,7,8] The scope of this article is limited to the method involving two frequency sources maintaining phase quadrature. As a general phase noise measurement tool it offers the best versatility and measurement accuracy. The block diagram of the measurement system is given in **Figure 1**.

The device under test (DUT) is a free-running LC VCO tuned to 6.144 MHz to match the frequency of the reference oscillator. The reference oscillator is built around a standard 6.144-MHz microprocessor crystal. It is assumed that the phase noise contribution of the reference source is insignificant compared to that of the VFO. One of the two oscillators must have provisions for electronic tuning to accomplish quadrature phase locking. In this example, the DUT is equipped with tuning for convenience; a tunable VXO as a reference

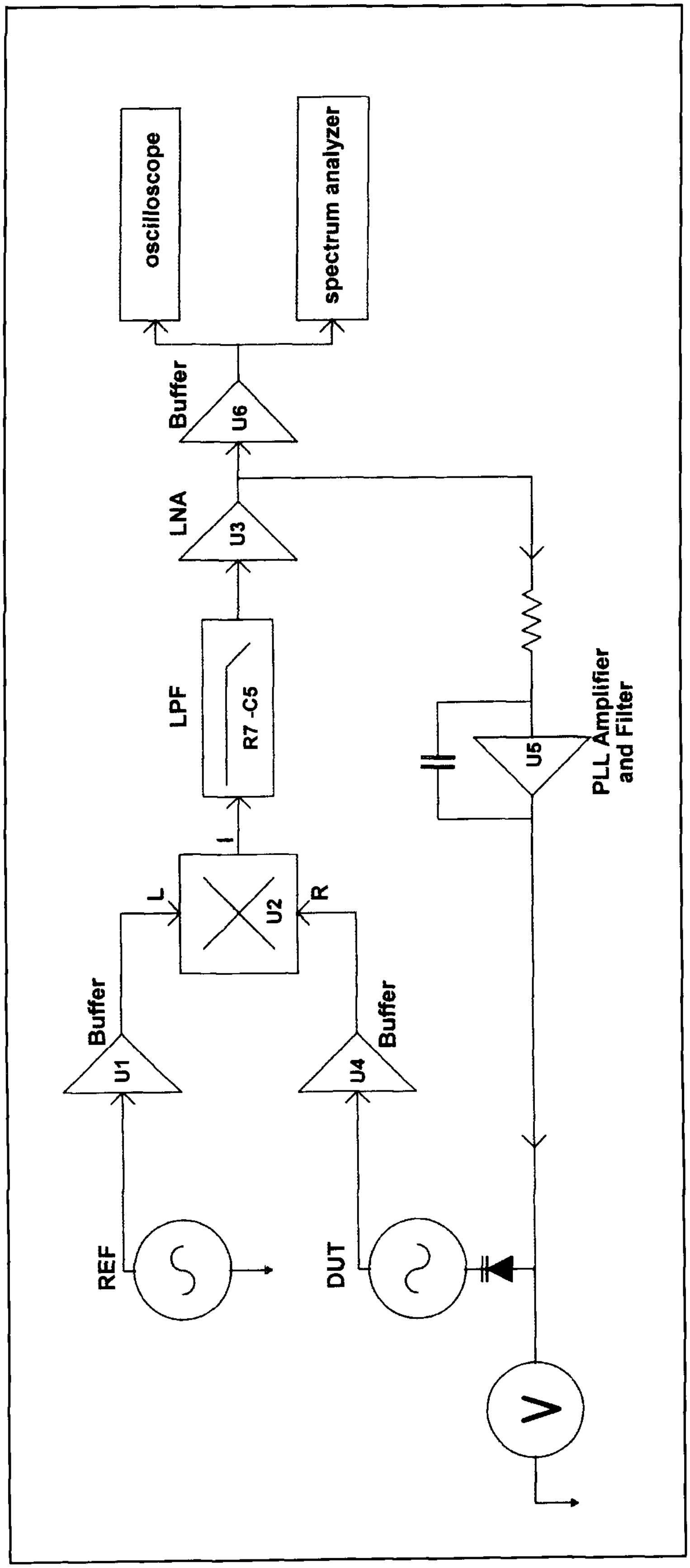


**Figure 1. Block diagram.**

source may be more appropriate if the DUT is a frequency synthesizer. Buffers U1 and U4 prevent injection locking of the two frequency sources. Mixer U2 serves as a phase detector. To properly use a double-balanced mixer as a phase detector, it is important that a quadrature is established and the mixer is used in its linear range.[8] Low-pass filter R7-C5 removes the conversion products that lie outside of the frequency range of interest: 0 to 100 kHz. The LNA built around U2 ensures that the full dynamic range of the spectrum analyzer is used, and the noise floor of the measurement system is not limited by the noise floor of the spectrum analyzer.

Buffer U6 is required to drive the 50-ohm input of the spectrum analyzer. The oscilloscope serves as a convenient way to monitor the progress of the PLL toward a quadrature lock. U5 is the loop amplifier—it produces the control voltage that causes the VCO to tune until it is at the same frequency as the reference source; the error voltage at the phase detector output returns to the nominal value of zero volts.

## Schematic diagram

The schematic diagram of the measurement system appears in **Figure 2**.

Buffers U1 and U4 have sufficient bandwidth to cover the HF band; measurements over the VHF band are possible if the buffers are replaced with devices with wider bandwidth. Resistors R1 and R12 discourage parasitic oscillations. Resistor R5 and the output impedance of U1 present a 50-ohm termination to the RF port of mixer U2. Resistor R16 and the output impedance of U4 present a 50-ohm termination to the LO port of the mixer.

U2 is a low-distortion high-LO level mixer. It ensures linear operation with signals up to 4 volts p-p at the J1 connector. The crystal oscillator should be able to provide a 8 to 10 volts p-p voltage swing at J2 to ensure that the required +17-dBm drive is applied to the LO port of the mixer. Resistor R6 presents a 50-ohm termination to the IF port of the mixer.

Networks R7-C5 and R23-C19 form a 2-pole low-pass filter to remove undesired frequency conversion products. A simple RC filter was chosen over an LC filter to ensure a flat noise gain response over the frequency band of interest. The -1 dB point of the filter is set to 150 kHz; it limits the phase noise measurement frequency range to 100 kHz.

U3 is a low-noise amplifier. Its noise contribution to the overall system noise can be neglected. The non-inverting configuration was chosen over the inverting configuration to preserve the flatness of the noise gain response. Resistors R10 and R11 set the gain of the amplifier; R11 is the gain control used during the calibration process.

R17, R20, C18, and U5 comprise the so-called Type 2 second-order loop. Being the most popular loop type, it offers almost infinite DC gain and independent selection of natural frequency of the loop and its damping factor. The proper selection of loop components is important and deserves a few comments. From the PLL theory from **References 5** and **9**:

$$\omega_n = (K_v * K_d/\tau_1)^{1/2} \ [1] \quad \zeta = (\tau_2 * \omega_n)/2 \ [2] \qquad \textbf{(3)}$$

where:

$\tau_1$ = R17*C18 and $\tau_2$ = R20*C18
$\omega_n$ is the natural frequency of the loop
$\zeta$ is the damping factor
$K_v$ is the VCO Gain factor (tuning sensitivity) in radians per second per volt
$K_d$ is the phase detector Gain factor in volts per radian

Because the PLL suppresses noise within its loop bandwidth, the -3 dB-point of the closed loop transfer function $f_{-3\,dB}$ should be chosen below the lowest offset frequency to be analyzed.

To accommodate the offset frequency range from 10 Hz to 100 kHz, the $f_{-3\,dB}$ is set to 3.5 Hz. The value of the damping factor is not critical for this application and is set to 0.7. The -3 dB-bandwidth of a Type 2 second-order loop is a function both $\omega_n$ and $\zeta$:[5,9]

$$\omega_{-3\,dB} = \omega_n * \{2\zeta^2 + 1 + [(2\zeta^2 + 1)^2 + 1]^{1/2}\}^{1/2}$$

$$\text{for } \zeta = 0.7 \quad \omega_{-3\,dB} = 2 * \omega_n \qquad \textbf{(4)}$$

Resistors R17 and R20 can be found from **References 1** and **2:**

$$R17 = (K_v * K_d)/(\omega^2_n * C18);$$
$$R20 = 2\zeta/(\omega_n * C18) \qquad \textbf{(5)}$$

where:

C18 is set to 100 µF
$\omega_n = (2\pi/2) * f_{-3\,dB} = \pi * 3.5 = 11$ rad/sec
$\zeta = 0.7$
$K_d$ = 0.38 V/rad - is obtained from Mini-Circuits data sheets
$K_v = 12 * 10^3$ rad/sec/V

The VCO Gain factor $K_v$ is obtained by applying a DC voltage to the VCO frequency control terminal and measuring the frequency deviation $\Delta f$ in Hz per 1 volt of control voltage change $\Delta V$.

$$\text{Gain factor } K_v = (2\pi\Delta f)/\Delta V \text{ rad/sec/V} \qquad \textbf{(6)}$$

The VCO gain factor is a non-linear function of the control voltage and must be measured around the nominal value (around 1 volt DC; the Q of the varactor diodes is too low below 0.7 volts, and the tuning sensitivity is too low above 2 volts DC).

The calculated resistor values are R17 = 390 k and R20 = 1.2 k.

With these loop component values, the acquisition time is very long. The speed of the acquisition can be improved by widening the loop bandwidth. This is a form of so-called aided acquisition[9] and is accomplished by placing a much smaller resistor in parallel with R17.

SW1, a 3-position toggle switch, places R22 in parallel with R17 to achieve a fast phase lock ("LOCK" position). SW1 should be in the middle position during the measurement period ("NORMAL" position). In the third position ("CALIBRATION"), a short is placed across the feedback components R20 and C18. In this mode, the phase lock is broken and a beat signal is produced at J3.

Finally, U6 serves as a buffer. Resistor R27 and the output impedance of U6 form a 50-ohm termination required at the input of the spectrum analyzer.

## Table 1. Phase noise versus frequency.

| Offset from carrier (Hz) | Measured phase noise (dBc/Hz) | SSB phase noise (dBc/Hz) |
|---|---|---|
| 10 | -75* | -81 |
| 100 | -99 | -105 |
| 1k | -128 | -134 |
| 2.2k | -138 | -144 |
| 10k | -155 | -161 |
| 100k | -159 | -165 |

*The noise measurement at 10 Hz is affected by the proximity of the -3-dB point of the loop bandwidth. Because the loop has a suppressing effect on the noise within the loop bandwidth, the actual noise level at 10 Hz is slightly higher.

## Calibration

The system calibration should be performed before every noise measurement to ensure measurement accuracy. The HP3585A spectrum analyzer is used for the phase noise measurement. In its "Noise Level" mode, it measures the rms noise spectral density in the frequency range between 0 and 40 MHz. The noise-level reading indicates the noise spectral density at the marker, normalized to a 1-Hz noise power bandwidth. The spectral noise density at the instrument's noise floor is -147

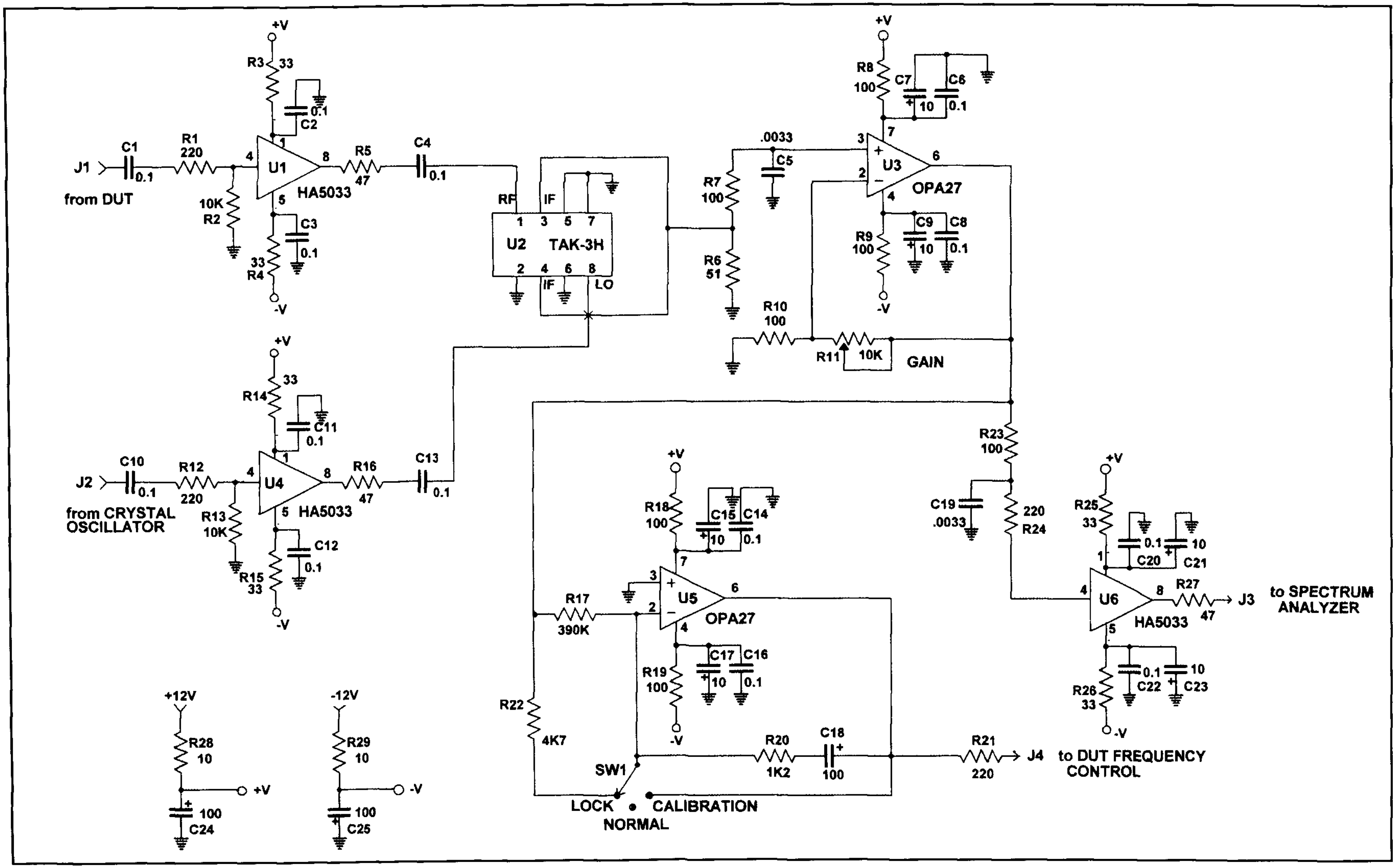


Figure 2. Schematic diagram.

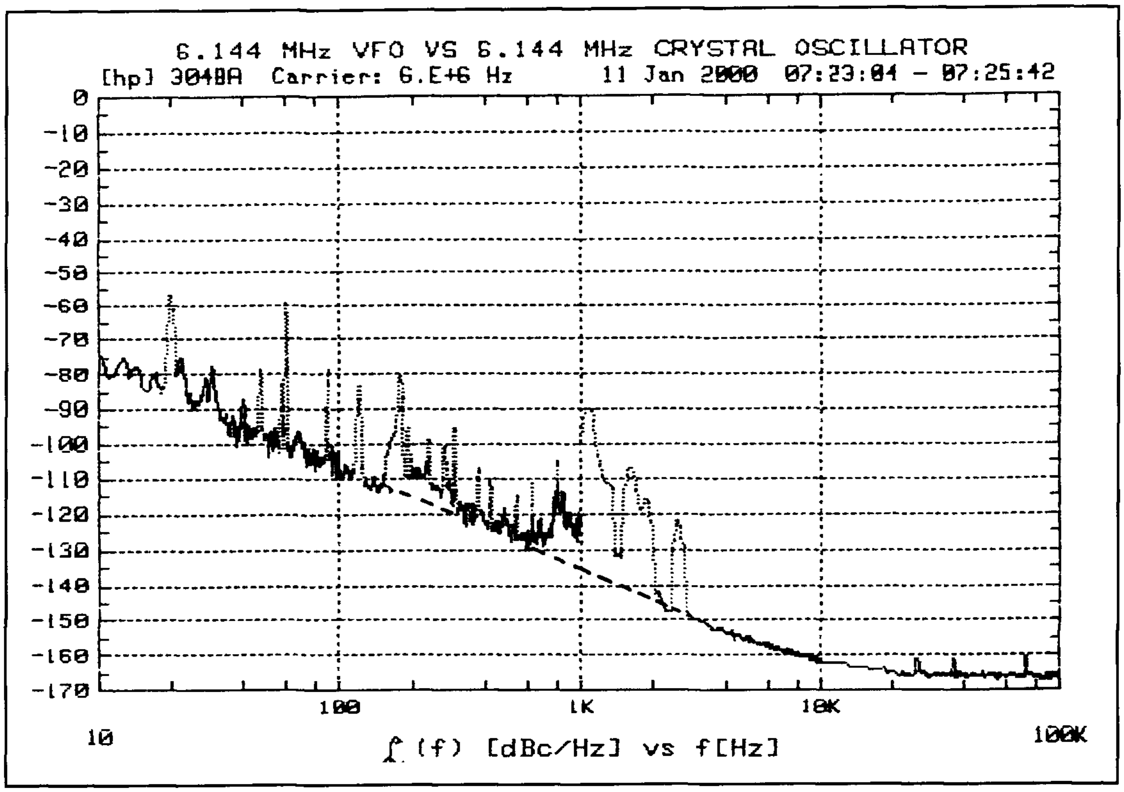


**Figure 3. HP3048A phase noise plot.**

dBm/1 Hz. Because the phase noise is measured relative to the carrier (dBc/1 Hz), the measurement range can be widened by raising the carrier level. If the carrier level (or the beat signal in our case) is raised to +20 dBm, the measurement range is widened to 167 dB (-167 dBc/1 Hz), which should be sufficient for most practical applications.

The calibration procedure is given below:

1. Apply power to the system and allow sufficient time for the frequency of the VFO to stabilize.

2. Toggle the switch momentarily to the "CALIBRATION" position; verify the presence of a beat frequency at J3 using the oscilloscope.

3. Disconnect J3 from the oscilloscope and connect it to the 50-ohm input of the spectrum analyzer.

4. Vary the VCO frequency to set the beat frequency anywhere between 5 and 50 kHz.

5. Set the analyzer marker to the peak of the signal; set the beat signal level to +20 dBm using the "GAIN" control R11.

6. Activate the "OFFSET" function and press the "ENTER OFFSET" button. The noise measurement floor has been moved down this way from -147 dBc/1 Hz to -167 dBc/1 Hz.

7. Vary the VCO frequency to set the beat frequency to 12 kHz; $f_{BEAT} = f_{REF} - f_{VFO}$.

8. Toggle the switch momentarily to the "LOCK" position; the beat signal disappears if lock is established.

9. Monitor the value of the control signal at J4 with a voltmeter. The voltmeter reading should be between 1.0 and 2.0 volts DC, otherwise adjust the VCO frequency slowly to bring the value of the control signal into the desired range. If the lock is broken in the process, re-establish it by toggling the switch again. If the voltmeter is AC powered, disconnect it prior to the measurement.

10. Activate the "NOISE LEVEL" function and the "dB (1 Hz)" reading should appear; it is equivalent to "dBc (1 Hz)."

11. Activate the "RANGE" function and set the value to -25 dBm. The instrument noise floor (after the "OFFSET" procedure) can be verified by disconnecting the signal from the input connector. The system is ready for a phase noise measurement.

## Measurement

The "NOISE LEVEL" function built into the HP3585A spectrum analyzer greatly simplifies the phase noise measurement process. The correction factors associated with equivalent noise bandwidth and detector type are automatically included in the readout.[11] The noise value is measured 100 times and averaged to reduce the

## Table 2. The components of the phase noise measurement fixture.

| Component | Part Number | Note Number |
|---|---|---|
| board | Vector 8007; Digi-Key #V1049-ND | 1 |
| connectors J1 - J4 | BNC, Receptacle Vertical PC, Amphenol, ARF1066-ND | 2 |
| buffers U1,4,6 | Harris, HA3-5033-5, DIP | 3 |
| amplifiers U3,5 | Burr-Brown, OPA27GP-ND, DIP | 2 |
| mixer U2 | Mini-Circuits, TAK-3H | 4 |
| potentiometer R11 | Philips, 10k, Multiturn, CT9W103-ND | 2 |
| resistors | 0.25 watt, 5% | surplus |
| capacitors, ceramic | Panasonic, type X7R, 5 or 10 % | 2 |
| capacitors, electrolytic | AVX, TAP series, tantalum, 16 volts DC 10 % tolerance | 3 |

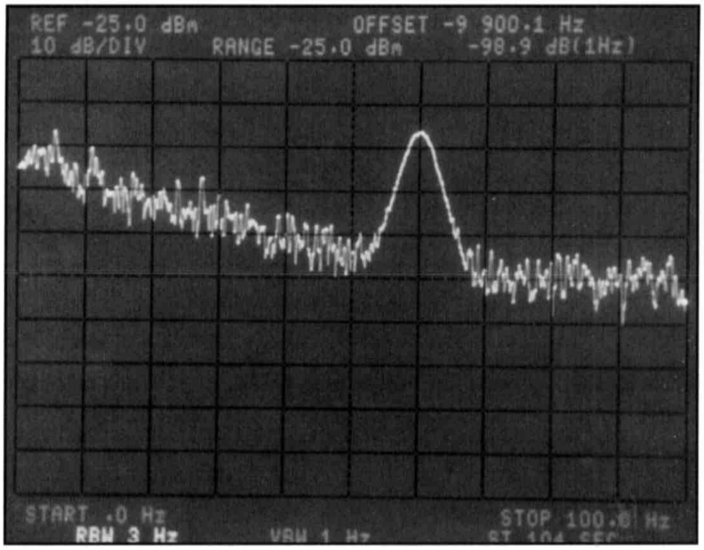


**Photo A. Phase noise measurement, 100-Hz sweep. Display marker is at 100 Hz.**

variance of the reading. The HP3585A can perform linear sweep only; it does not provide sufficient frequency resolution in a single sweep over 4 decades of desired frequency coverage (from 10 Hz to 100 kHz). Therefore, the desired frequency range is covered in three separate sweeps: 0 to 100 Hz, 0 to 10 kHz, and 0 to 100 kHz.

The measurement procedure is given below:

1. Set the "STOP FREQUENCY" to 100 kHz, "RESOLUTION BW" to 100 Hz, and "VIDEO BW" to 30 Hz. Place the marker at 100 kHz and record the readout in "dB/1 Hz".

2. Set the "STOP FREQUENCY" to 10 kHz, "RESOLUTION BW" to 30 Hz, and "VIDEO BW" to 10 Hz. Record the readout at 1 kHz, 2.2 kHz, and 10 kHz. The noise measurement at 2.2 kHz is required in order to verify whether the initial close-in phase noise level requirement has been met.

3. Set the "STOP FREQUENCY" to 100 Hz, "RESOLUTION BW" to 3 Hz, and "VIDEO BW" to 1 Hz. Record the readout at 10 Hz and 100 Hz. The 60-Hz leakage should have no effect on the measurement, but if it is excessive, a battery-operated power supply should be considered.

**Table 1** shows the results of the phase noise measurement. The spectrum analyzer measures noise from both noise sidebands, which add linearly at the mixer output. The SSB noise of the original RF signal is obtained by simply subtracting 6 dB from the measured results.[6,7,8,10]

**Photos A**, **B**, and, **C** show the HP3585A measurements results for three different frequency sweeps.

A phase noise test using the HP3048A Phase Noise Test System was performed to verify the accuracy of the measurement results in **Table 1**. The HP3048 phase noise plot is provided in **Figure 3**. The phase detector method has been used where the 6.144-MHz oscillator served as the reference oscillator. The few breaks in the phase noise curve are due to injection locking. Examination of the HP3048 phase noise plot shows excellent agreement with the measurement results in **Table 1**.

## VFO

The schematic diagram of the VFO is shown in **Figure 4**. Because the VFO was designed for portable operation, the design criteria were simplicity, a minimum number of parts, and low-power consumption.

L1, the tapped tank coil, is wound around an iron-powder toroidal core; the unloaded Q of the coil exceeds 300. C4, C5, C7, and C8 are NPO ceramic capacitors. C6 has a temperature-compensating characteristic to compensate for the positive temperature coefficient of the toroidal core. C2, an air-variable capacitor, tunes the frequency from 6.0 to 6.35 MHz to accommodate a 20-meter transceiver with an 8.0-MHz IF. The loaded Q of the resonator is kept high by using a tapped coil and loose coupling to the gate of Q1 through C8. J310 was selected as the active component because of its very low noise level in the HF frequency range. Resistor R5 is adjusted to set the Q1 drain current equal to 4 mA.

Q2-Q3 is a push-pull buffer. It has excellent linearity and low output impedance. Resistor R6 discourages parasitic oscillations, and resistor R12 sets the output impedance of the buffer equal to 50 ohms.

Two back-to-back varicaps, D1 and D2, and associated circuitry serve as a means of adjusting the VFO frequency to match the frequency of the reference oscillator during the phase noise measurement. Switch SW1 disconnects the adjustment circuit after the measurement is completed.

U1 and associated components provide the VFO with a regulated 9-volt supply. Bias current setting resistor R13 sets the regulator standing current equal to 2 mA. The overall VFO current drain is under 9 mA.

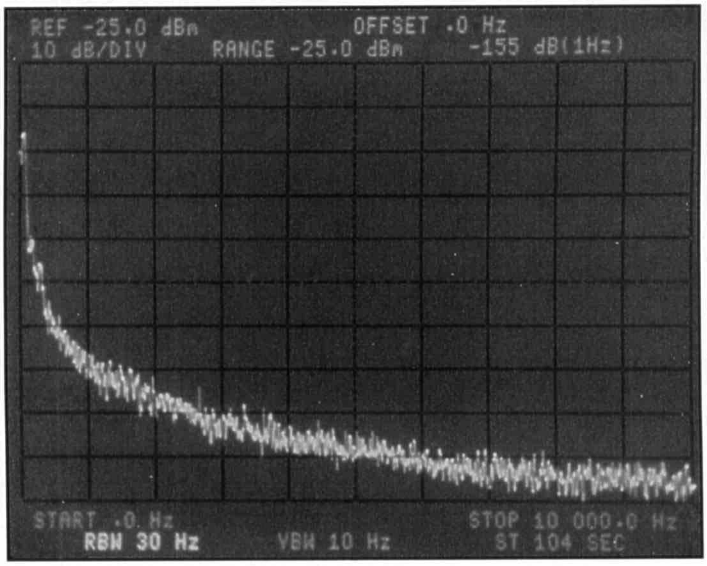


**Photo B. Phase noise measurement, 10-kHz sweep. Display is at 10 kHz.**

Despite the simplicity, the phase noise of the VFO is quite impressive. The design goals were met with a healthy margin. Several guidelines intended to minimize the phase noise level were implemented in this design:

1. High unloaded Q of the resonator,
2. Light coupling between resonator and FET to avoid saturation and maintain high loaded Q,
3. Low L/C ratio,

## Table 3. Details of the VFO components.

| Component | Part Number | Note Number |
|---|---|---|
| aluminum box | Rolec, 806-0007, 3.1 x 4.7 x 2.4 inches | 3 |
| board | Vector, 8007; Digi-Key # V1049-ND | 1 |
| BNC connector | BNC, Receptacle Vertical PC, Amphenol, ARF1066-ND | 2 |
| toroidal core, L1 | T-94-6 | 5 |
| FET, Q1 | J310 | 3 |
| diodes, D3, D4 | 1N4148 | 3 |
| transistor, Q2 | 2N4401 | 3 |
| transistor, Q3 | 2N4403 | 3 |
| varicaps, D1, D2 | NTE612, ECG612 | 3 |
| voltage regulator, U1 | LM317MP | 3 |
| RF chokes, L2, L3 | Delevan, DN2568-ND, 100 uH | 2 |
| resistors | 0.25 watt, 5% | surplus |
| capacitors, ceramic | Panasonic, type X7R, 5 or 10 % | 2 |
| tank capacitors | Philips, COG 100 volts, 13xxPH-ND | 2 |
| t° comp. capacitor, C6 | Philips, N750 100 volts, 1309PH-ND | 2 |
| capacitor air-variable, C2 | Oren Elliott, N-50, 15 blades-5160 | 6 |
| trimmer capacitor, C3 | Xicon, 2.7-10pF, 242-2710, NPO | 7 |
| feed-thru capacitors, C15,C16 | 1000 pF | surplus |
| capacitors, electrolytic | AVX, TAP series, tantalum, 16 volts DC, 10% tolerance | 3 |

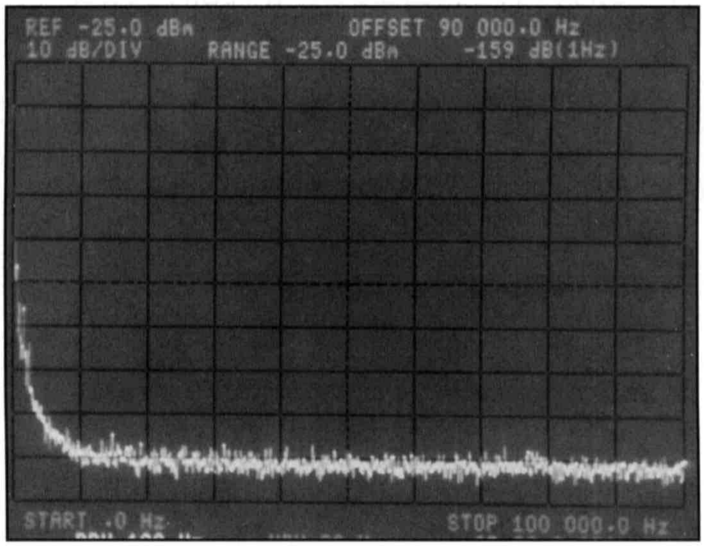


**Photo C. Phase noise measurement, 100-kHz sweep. Display marker is at 100 kHz.**

4. Active device with a low noise figure and low phase perturbation (J310),

5. Unbypassed source resistor to reduce flicker noise,

6. No gate clamping diode,

7. Energy coupling directly from the resonator (via C9).

More information on minimizing the phase noise can be found in **Reference 5**.

The implementation of the 6.144-MHz crystal oscillator is not critical. Practical examples of crystal oscillators are provided in **References 1** and **12**.

## Construction

### A. Phase noise measurement fixture

The components of the measurement fixture are mounted on Vector 8007 board (perforated, with a solid copper plane on one side). The components are mounted on the copper side, which serves as a ground plane. The ground side of bypassing components is soldered directly to the ground plane.

Input (J1-2) and output (J3-4) terminations are implemented via BNC connectors soldered directly to the board. At this frequency range, there was no need to enclose the board in a metal box, but it should be considered at higher frequency ranges.

**Table 2** serves as a guide for procuring the necessary components.

### B. VFO

The VFO is enclosed in a diecast aluminum box to ensure mechanical rigidity and provide RF-tight construction. The DC power and the control voltage are delivered via feed-thru capacitors to avoid RF leakage. The air-variable capacitor is attached to one of the side walls of the box with three screws. The VFO output is delivered via a BNC connector attached to another side wall.

The VFO components are mounted on a piece of a Vector board. Layout isn't critical, but the component leads are kept short. The tank coil L1 is built by tightly winding 22 turns of #16 AWG enameled copper wire on a T-94-6 iron-powder toroidal core tapped at three turns and 14 turns from the cold end. The coil is covered with low-loss polystyrene Q dope for enhanced stability. The inductance of the coil is 3.26 μH.

SW1 is a wire jumper that is removed after completion of the testing procedure.

The VFO component details are provided in **Table 3.** Dealer information is provided at the end of the article.

## Summary

Commercial equipment that is capable of measuring low levels of phase noise (like the HP3047 or HP3048 Phase Noise Test System) is quite expensive and far beyond the reach of most amateurs.

The measurement method described above offers a much more economical alternative. It presumes access to the HP3585A spectrum analyzer. Although discontinued years ago, this versatile instrument is still quite common and can be found in most well-equipped labs. It can also be found on the surplus equipment market at a price under $8,000.

In the absence of the HP3585A, any low-frequency spectrum analyzer can be used; however, a different calibration procedure would have to be developed using the rationale described previously. In addition, several correction factors associated with analog spectrum analyzers[6,7,8,10,11] would have to be applied to the readout to ensure measurement accuracy. ■

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