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Some Thoughts on Designing Very High Performance VHF Oscillators

Building a very high performance oscillator requires some careful engineering design work.

A *QEX* article by Colin Horrabin about part of the HF7070 receiver retriggered my interest in VHF oscillators / VCOs. (The Development of the Low Phase Noise Double Tank Oscillator, Colin Horrabin, G3SBI, *QEX* Nov/Dec 2014.)[1] He claimed that a type of push-pull oscillator would improve the phase noise roll-off from 20 dB/dec to 40 dB/dec, and he also referred to some receiver measurements made by Rob Sherwood. The data points I reviewed do not support this theory, and the reciprocal mixing tests are not conclusive, because two signal generators were used. The correct comment is that the type 2, high-order phase locked loop inherently has a 40 dB/dec roll off, not the oscillator.

The single resonator oscillator using lumped elements by itself is a good solution. The slope of the radiation resistance of a quarter wave resonator does not change if a half wave resonator will be chosen, so a push-pull oscillator is not better.

The symmetrical oscillator proposed by Horrabin just uses twice the inductance, and the two capacitors, now in series, have half their individual value. In simple terms, Horrabin changed the LC ratio, which cannot have any influence on the phase noise nor the slope. The loading from the transistor may now be different.

The best way to get the phase noise evaluation right is to use a dedicated phase noise system like the Rohde & Schwarz FSUP 26 phase noise tester, spectrum and signal analyzer that the ARRL Lab has to make their measurements. At the same time, it is useful to calculate the best possible phase noise based on physics and using a low flicker noise FET. FETs in oscillators are limited to about 500 MHz because of their cut-off frequency. For higher frequencies SiGe HBT (heterojunction bipolar transistors) are superior, and because modern communications equipment uses PLL systems with sufficiently wide bandwidth, the flicker corner frequency inside the loop bandwidth is of less concern. Outside the loop bandwidth the loaded *Q* of the resonator determines the phase noise. If Colin Horrabin's paper is correct, the roll off has to be 20 dB/decade or 40 dB/decade but not 30 dB/decade, which would be due to flicker noise. A VCO with 1 kHz loop bandwidth was quoted. I will comment on this later.

Only for oscillators using the evanescent mode and distributed elements, like (multiple) coupled lines, the configuration results in an increased operating *Q*, which for lumped circuit components is not possible. Using coupled transmission line structures (distributed components) is a better choice. At VHF, this is prohibitive because of size.

An evanescent wave is a near-field wave with an intensity that exhibits exponential decay without absorption as a function of the distance from the boundary at which the wave was formed. Evanescent waves are solutions of wave equations, and can in principle occur in any context to which a wave equation applies. They are formed at the boundary between two media with different wave motion properties, and are most intense within one third of a wavelength from the surface of formation.

As evidence of how moving from lumped to distributed techniques can improve oscillator performance at frequencies where *LC* tank circuits become problematic, Figure 1 compares the difference in phase-noise performance obtainable using a resonator consisting of an ideal 2 nH inductor and a ¼ λ transmission line (11 Ω, 90° long at 2.6 GHz, attenuation 0.1 dB/meter) with the transistor biased by constant-current and constant-voltage sources for a simulated BJT Colpitts oscillator operating at 2.3 GHz. This is a result of the magnetic coupling, which does not exist for lumped (discrete) inductors.[2] The articles described in Notes 3, 4, 5, and 6 address this topic in practical applications.[3, 4, 5, 6]

The 1 kHz loop bandwidth would be dangerous because mechanically introduced microphonics would then not be suppressed. A 10 kHz loop bandwidth is much more opportune. Better synthesized local oscillators (LOs) use multiple loops and direct digital synthesis (DDS) systems, which allow such wide loop bandwidth. Many modern receivers and transceivers apply this technique.[7, 8] Even better today, software defined radios (SDR) can have excellent phase noise performance. (See the R&S EB-500 9 kHz to 6 GHz receiver: **http://n1ul.com/eb500.htm**.)

[1]Notes appear on page 40

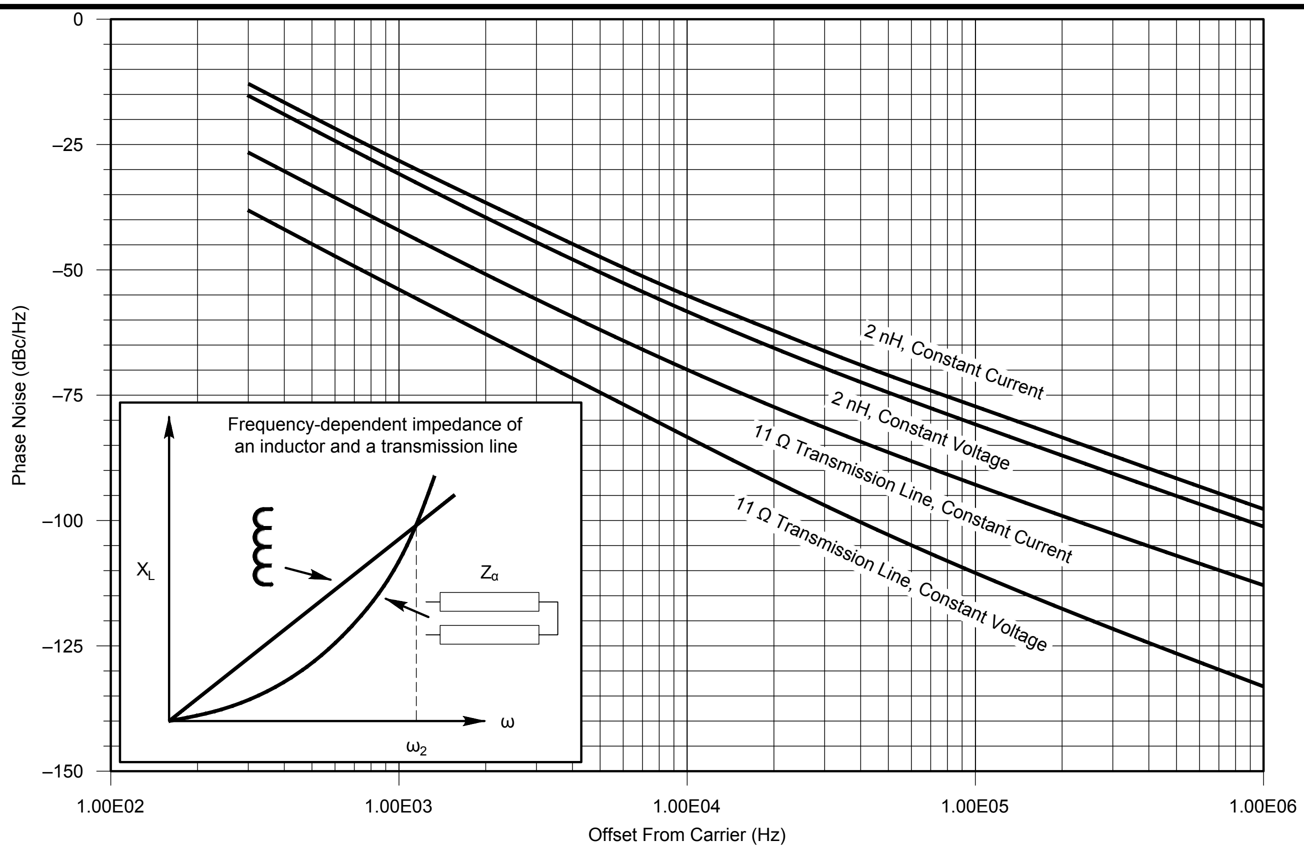

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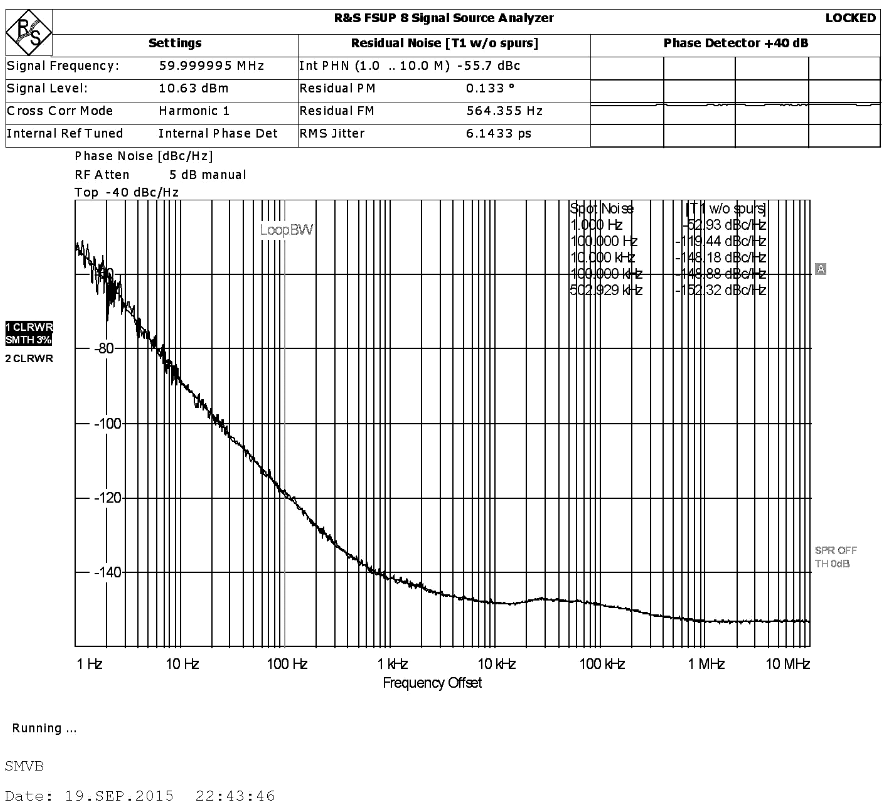
Figure 1 — This graph shows the phase-noise performance of a 2.3 GHz BJT oscillator with a resonator consisting of an inductor (2 nH) and a ¼ λ transmission line (11 Ω, approximating the behavior of a dielectric resonator) with bias from a constant-current source and a low-impedance, resistive constant-voltage source.

Figure 2 — This screen shot from the Rohde & Schwarz FSUP signal source analyzer shows the measured phase noise of a synthesized 60 MHz LO. The –150 dBc/Hz limit is a result of the buffer amplifier. It requires a special design to obtain better values. Even at +10 dBm output there **is a practical limit of about –175 dBc/Hz for a VCO and buffer. Some additional useful information can be found in the articles of Notes 9 and 10.**



Transceivers with a first IF between 45 to75 MHz, require such VHF oscillators. This paper will try to demystify this topic and will show the correct mathematics, proven schematics and measured data. It is partly based on *RF/Microwave Circuit Design for Wireless Applications* (see Note 2). Figure 2 shows the measured phase noise performance of a modern receiver that uses a 60 MHz LO. This measurement was made using the aforementioned R&S FSUP signal analyzer.

Some Equations

David Leeson was the first to help us understand the mechanics of phase noise, based on a low pass filter approach in 1966.[11] Dieter Scherer and others improved the model further.[9, 10, 12, 13, 14]

Phase noise is defined in terms of the noise spectral density, in units of decibels below the carrier per hertz, and is based on Equation 1 by Leeson, Scherer and Rohde.

$$£\left(f_m\right)=10\log\left[\frac{P_{\text{sideband}}\left(f_0+f_m,\ 1\ \text{Hz}\right)}{P_{\text{carrier}}}\right]=10\log\left[S_\varphi\left(f\right)\right]$$

[Eq 1]

$$£\left(f_m\right)=10\log\left\{\left[\left[1+\frac{f_0^2}{\left(2f_mQ_L\right)^2\left(1-\frac{Q_L}{Q_0}\right)^2}\right]\left(1+\frac{f_c}{f_m}\right)\frac{FkT}{2P_0}+\frac{2kTRK_0^2}{f_m^2}\right]\right\}$$

[Eq 1A]

where:
$£(f_m)$ is the ratio of the sideband power in a 1Hz bandwidth at f_m to total power in dB
f_m is the offset frequency from the carrier
f_0 is the carrier frequency
f_c is the flicker corner frequency
Q_L is the loaded Q of the tuned circuit
Q_0 is the unloaded Q of the tuned circuit
F is the noise factor
k is Boltzmann's constant
T is the temperature in Kelvins
P_0 is the average power at oscillator output
R is the equivalent noise resistance of the tuning diode
K_0 is the oscillator voltage gain.

When adding an isolating amplifier, the noise of an LC oscillator is determined by Equation 2.

$$£\left(f_m\right)=0.5\times10\log\left[\left(S_\varphi\left(f_m\right)\right)\right]$$

$$£\left(f_m\right)=0.5\times10\log\left\{\frac{\left[a_RF_0^4+a_E\left(\frac{F_0}{2Q_L}\right)^2\right]}{f_m^3}+\frac{\left[\left(\frac{2GFkT}{P_0}\right)\left(\frac{F_0}{2Q_L}\right)^2\right]}{f_m^2}+\left(\frac{2a_RQ_LF_0^3}{f_m^2}\right)+\frac{a_E}{f_m}+\frac{2GFkT}{P_0}\right\}$$

[Eq 2]

where,
G = compressed power gain of the loop amplifier
F = noise factor of the loop amplifier
k = Boltzmann's constant
T = temperature in kelvins
P_0 = carrier power level (in watts) at the output of the loop amplifier
F_0 = carrier frequency in Hz

f_m = carrier offset frequency in Hz

$Q_L (= \pi F_0 \tau_g)$ = loaded Q of the resonator in the feedback loop

a_R *and* a_E = flicker noise constants for the resonator and loop amplifier, respectively.

The problem with this design equation, which everyone likes to quote, is that it works after the fact. That means the designer does not know the output power, the flicker corner frequency, and the large signal noise figure, and finally, because the right part of the equation is the noise from the tuning diode, the value of the equivalent noise resistor, *R*!

Influence of the Tuning Diode

It is possible to define an equivalent noise resistor, R_{aeq}, which when inserted into Nyquist's equation, determines an open-circuit noise voltage across the tuning diode.

$$V_n=\sqrt{4kT_0R\Delta f}$$

[Eq 3]

where:
$kT_o = 4.2 \times 10^{-21}$ at about 300 K
R is the equivalent noise resistor
Δf is the bandwidth.

Practical values of R_{aeq} for carefully selected tuning diodes are in the vicinity of 200 Ω to 50 kΩ. We can now determine the noise voltage, V_n.

$$V_n=\sqrt{4\times4.2\times10^{-21}\times10{,}000}=1.296\times10^{-8}\,V\sqrt{Hz}$$

This noise voltage generated from the tuning diode is now multiplied with the VCO gain, K_o, resulting in the RMS frequency deviation.

$$\left(\Delta f_{rms}\right)=K_0\times\left(1.296\times10^{-8}\,V\right)\text{in 1 Hz bandwidth}$$

[Eq 4]

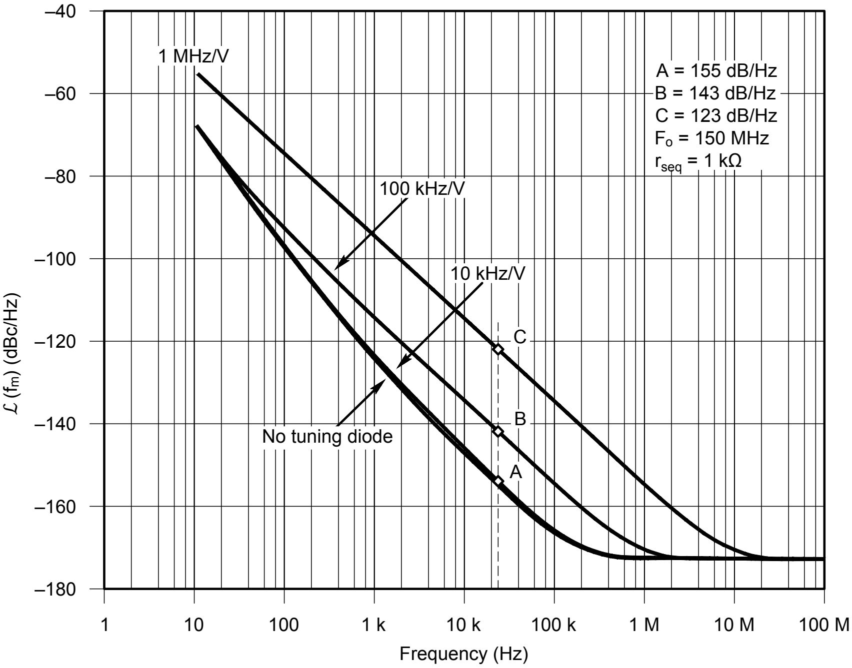


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Figure 3 — This graph shows the influence of the diode noise of a VCO at 150 MHz.

To translate this into an equivalent peak phase deviation, we will use Equation 5.

$$\theta_d = \frac{K_0\sqrt{2}}{f_m}\times\left(1.296\times 10^{-8}\right)\text{rad in 1 Hz bandwidth}$$

[Eq 5]

Or, for a typical oscillator gain of 100 kHz / V:

$$\theta_d = \frac{0.00183}{f_m}\text{rad in 1 Hz bandwidth}$$ [Eq 6]

For f_m = 2.4 kHz (typical spacing for adjacent-channel measurements for good SSB RF radios), then $\theta_c = 732 \times 10^{-9}$. This can be converted now into the SSB signal-to-noise ratio:

$$£\left(f_m\right) = 20\log_{10}\frac{\theta_c}{2} = -128\ \text{dBc / Hz}$$ [Eq 7]

The tuning diode adds significant noise, so if the above mentioned 1 kHz bandwidth for the PLL is used, at 2.4 kHz, the oscillator dominates.

Figure 3 shows the influence of the diode noise of a VCO at 150 MHz. In the case of lines B and C on the graph, you can see that the tuning diode greatly ruins the overall phase noise regardless of a high loaded *Q*!

The flicker frequency component also has a huge influence on the phase noise. Figure 4 shows the noise contribution of the flicker noise in a circuit with fixed *Q*. At 1 kHz offset, the phase noise deteriorates by 10 dB.

We can calculate the phase noise from circuit parameters, and using large singal parameters, or deriving these with the help from Bessel funtions, we specifically obtain Y21 for a large signal.

The total effect of all the four noise sources can be expressed as Equation 8.

$$£\left(\omega\right) = 10\log\frac{4KT}{\omega L\times Q}\left\{\frac{1}{2}\left[\frac{1}{2\omega_0 C_{eff}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2_{\text{Resonator}}$$

$$+4KTr_b\left\{\frac{1}{2}\left[\frac{C_1+C_2}{C_2}\right]\left[\frac{1}{2Q}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2_{\text{Base Resistance}}$$

$$+\left[2qI_b+\frac{2\pi K_f I_b^{AF}}{\omega}\right]$$

$$\left\{\frac{1}{2}\left[\frac{C_2}{C_1+C_2}\right]\left[\frac{1}{2Q\omega_0 C_{eff}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2_{\text{Flicker Base Current}}$$

$$+2qI_c\left\{\frac{1}{2}\left[\frac{C_1}{C_1+C_2}\right]\left[\frac{1}{2\omega_0 Q C_{eff}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2_{\text{Collector Current}}$$

[Eq 8]

We will use the example from the 2 Part *Microwave & RF* article, "Large-Signal Approach Yields Low-Noise VHF/UHF Oscillators."[15]

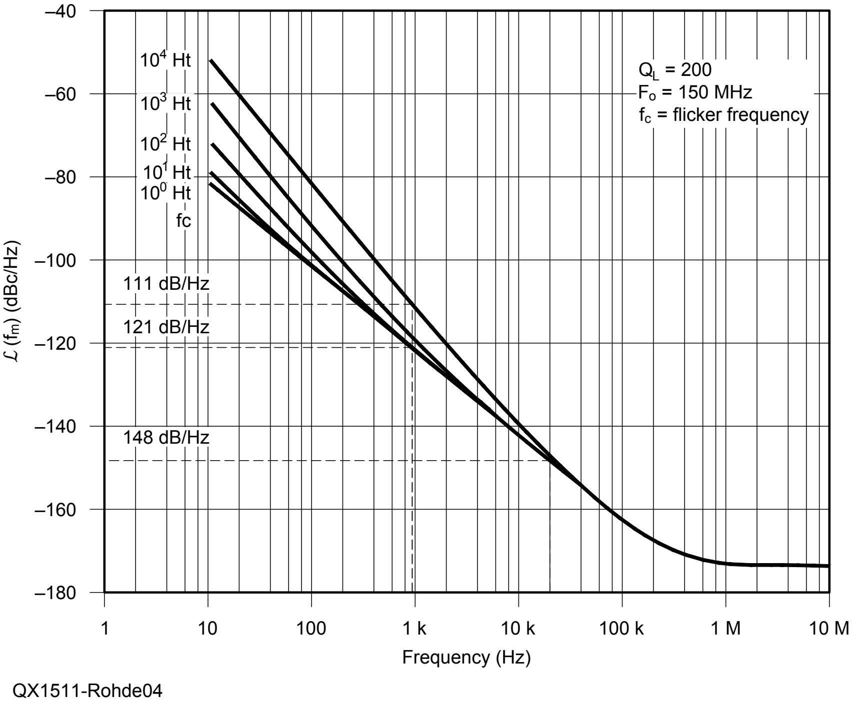

Figure 4 — Here is the phase noise contribution of the flicker noise to the oscillator noise.

[16] The schematic for this circuit is shown at Figure 5, and the measured phase noise of this 144 MHz oscillator is shown in Figure 6.

From the resonator, $R_P = 7056\ \Omega$ ($\omega L \times Q$)
Q of the resonator = 200 (Q of the inductor at 144 MHz)
Resonator inductance = 39 nH
Resonator capacitance = 22 pF
Collector current of the transistor, I_c = 10 mA
Base current of the transistor, I_b = 85 µA
Flicker noise exponent, AF = 2
Flicker noise constant, $K_f = 1 \times 10^{-12}$
Feedback factor, n = 5
Phase noise at 10 kHz:

$$PN_{(ibn+ifn)_i}\left(\omega_0 + 10\text{ kHz}\right) \approx -134.2\text{ dBc / Hz}$$

$$PN_{Vbn}\left(\omega_0 + 10\text{ kHz}\right) \approx -151\text{ dBc / Hz}$$

$$PN_{nr}\left(\omega_0 + 10\text{ kHz}\right) \approx -169.6\text{ dBc / Hz}$$

$$PN_{icn}\left(\omega_0 + 10\text{ kHz}\right) \approx -150.6\text{ dBc / Hz}$$

$$P_{out} = 5\text{ dBm}$$

The value for $K_f = 1 \times 10^{-12}$ is vaild for small currents, and in Equation 8 the main phase noise (measured) contribution is the resonator loss. For higher frequencies and higher output power (higher DC current, the flicker and DC current contribuion to the flicker noise will dominate. At 30 mA and higher, a typcal K_f factor of 1×10^{-7} is common.

Going back to the large signal phase noise analysis , the Equation 9 is really the most modern result.

$$\pounds(\omega) = 10\log\left\{\left[k_o + \left(\frac{k^3 k_1\left[\frac{Y_{21}^+}{Y_{11}^+}\right]^2 [y]^{2p}}{\left[Y_{21}^+\right]^3 [y]^{3q}}\right)\left(\frac{1}{\left(y^2+k\right)}\right)\right]\left[\frac{(1+y)^2}{y^2}\right]\right\}$$

[Eq 9]

where:

$$k_0 = \frac{kTR}{\omega^2\omega_0^2 L^2 C_2^2 V_{cc}^2}$$

$$k_1 = \frac{qI_c g_m^2 + \frac{K_f I_b^{AF}}{4\omega} g_m^2}{\omega^2\omega_0^4 L^2 V_{cc}^2}$$

$$k_2 = \omega_0^4\left(\beta^+\right)^2$$

$$k_3 = \omega_0^2 g_m^2$$

$$k = \frac{k_3}{k_2^2 C_2^2}$$

where k_1, k_2, and k_3, are constant only for a particular drive level, with $y = C_1 / C_2$, making k_2 and k_3 also dependent on y, as the drive level changes.

This Equation is derived in *Communications Receivers* (see Note 8).

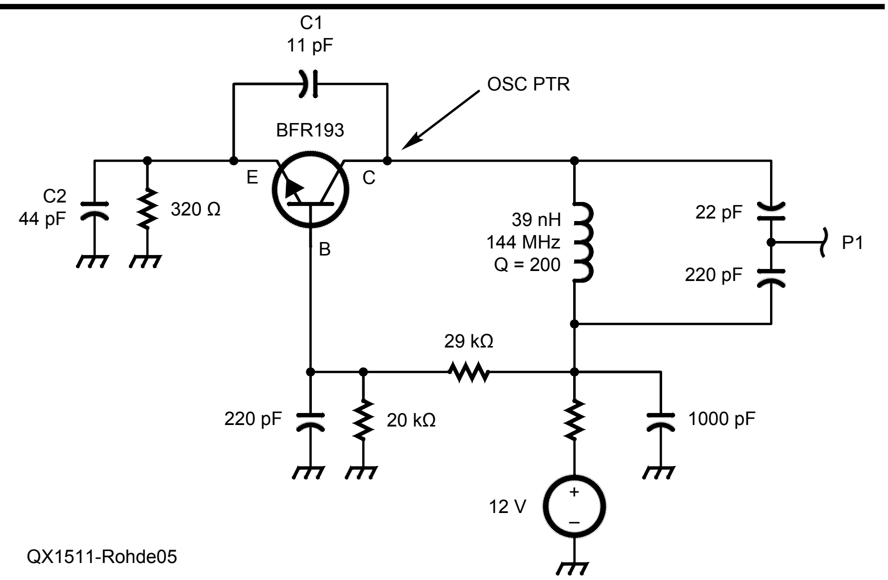

Figure 5 — This schematic shows a 144 MHz oscillator design at 60 MHz. This design is from "Large Signal Approach Yields Low-Noise VHF/UHF Oscillators," published in *Microwaves & RF*. See Notes 15 and 16.

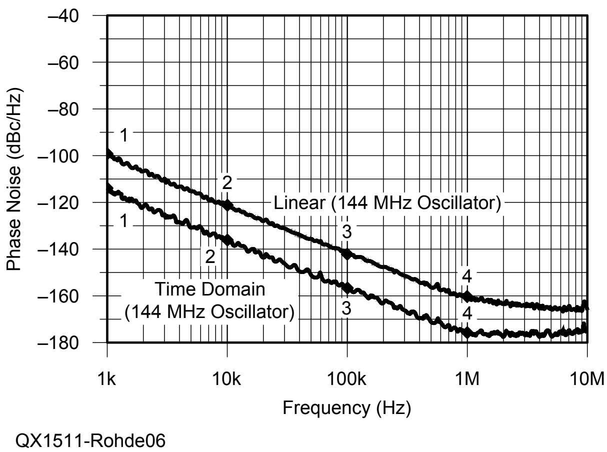

Figure 6 — This is the meassured phase noise of the 144 MHz Oscillator design of Figure 5, based on state of the art linear design, and based on an optimized design using large signal parameters. See Notes 15 and 16.

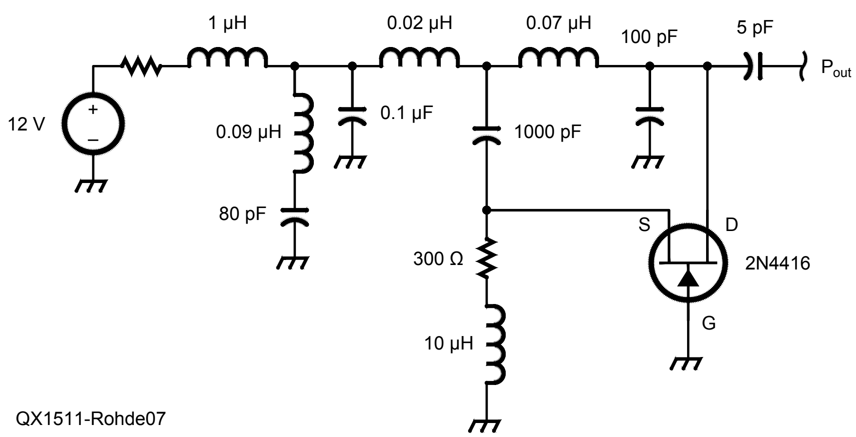

Figure 7 — This circuit is a possible simulation of the 60 MHz oscillator described by Colin Horrabin.

Another phase noise calculation approach is noted by Hajimiri in "A General Theory of Phase Noise in Electrical Oscillators."[17] It is quoted by academicians frequently because it is an elegant way, but for actual design activities it is useless. It is mentioned here for completeness. Also see *The Design of Modern Microwave Oscillators for Wireless Applications: Theory and Optimization*.[18]

The Circuits

In order to verify the noise quoted by Colin Horrabin, an FET circuit with 2 tuned LC circuits was prepared for simulation using the familiar 2N4416 JFET. Its data was obtained from the non-linear data provided by Philips for CAD applications, such as *SPICE* or *Harmonic Balance* based simulators.[19]

The power supply voltage is applied via a 1 µH RF choke, and in order to validate the claim, the 0.1 µF capacitor in the analysis could be toggled between this value and 0.1 fF = 0.1×10^{-15} F, in practice a value of zero. The result showed no difference in phase noise. There was a discussion about why the simulator did not agree with the expectations, but the phase noise values published by Colin Horrabin did not support the claim either. This topic was addressed in the beginning of this paper. Interestingly enough, if the circuit is made asymmetrical (see the 80 pF and 100 pF capacitors in Figure 7), and the tap is not grounded, a better phase noise results.

The simulation data agree fairly well with the published data, and no correction for the noise of the tuning diode was made. Figure 8 shows the predicted phase noise of the Figure 7 oscillator.

It is now of interest to design a better VCO. This has been achieved with the design shown in Figure 9. The noise improvement comes from the constant current source (5.6 kΩ) in the source; the higher voltage drop is compensated by the positive voltage at the transistor gate.

Figure 10 shows a circuit diagram of an ultra low noise 60 MHz FET oscillator design that uses a 2N4416 FET. The circuit uses a helical resonator, as shown in Figure 11. The original cicuit was modified and is using six aditional diodes for a wider tuning range, and the parallel combination of the diodes, because of no noise correlation, results overall in a lower noise contribution. Figure 12 shows the phase noise simulation for this oscillator circuit, and Figure 13 shows the measured result from the actual circuit.

The diodes make the VCO noisier below 100 khz, but because the loop bandwidth typically is wider, this compensates the noise. If we look at Equation 8, we will find that the major noise contribution is the loaded *Q* of the resonator. If by some magic the loading of the transistor drain impedance could be reduced, the noise would be less. Here, flicker noise is not the dominant cause!

Summary

The design of low noise oscillators is no longer such a mystical task. When I finally got my own R&S FSUP 8 with optimized internal signal sources, I went through the task of measuring my oscillators built 40 years ago, as well as some commercial devices. A good example was the older HP 8640B, and the famous HP 10544A 10 MHz crystal oscillators.

HP products typically were better than promised, something I could not claim for all of my designs, but I was not that far off — and yes some were better than published.

Sadly I found that many VHF crystal oscillators around in the past did not perform as well as we know today, and the same applies to signal generators.

This paper also lists a large number of references and I recommend *RF/Microwave Circuit Design for Wireless Applications* (see Note 2), *Microwave and Wireless Synthesizers: Theory and Design* (see Note 7), and *The Design of Modern Microwave Oscillators for Wireless Applications* (see Note 17) for text books for any readers interested in learning more about synthesizers and oscillators.

Microwave and Wireless Synthesizers: Theory and Design gives a detailed insight into PLL design, but companies now sell

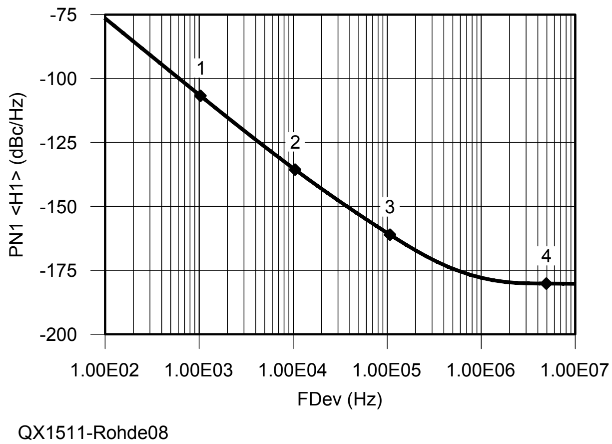

Figure 8 — This graph is the predicted phase noise of the Colin Horrabin oscillator, based on the simulation of Figure 7.

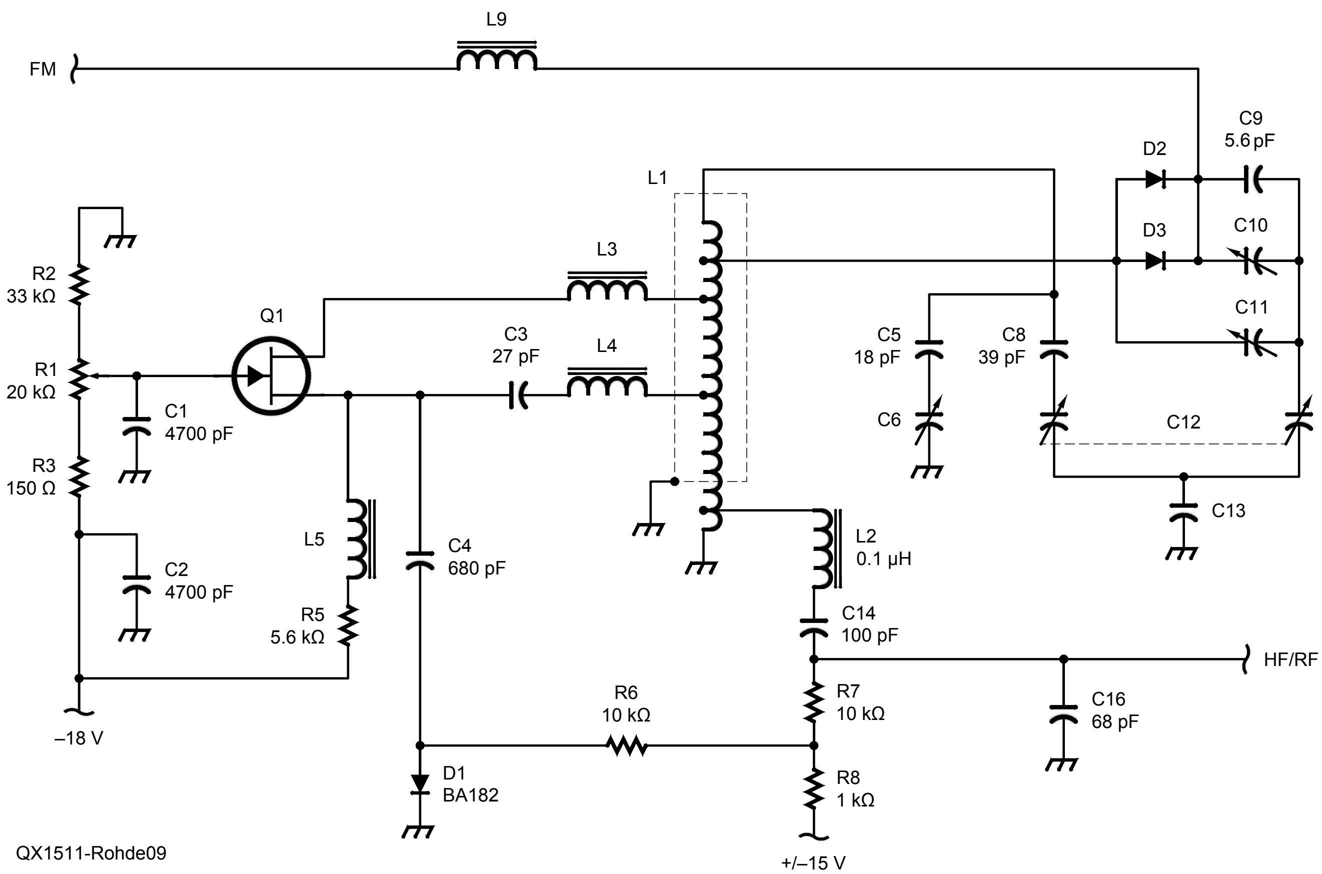

Figure 9 — This schematic diagram shows a 60 MHz VCO optimized for phase noise. It uses the 2N4416 FET and a ±15 V source, which switches the oscillator on and off. L1 is a helical resonator. R&S 1975 Model SMDU radio tester.

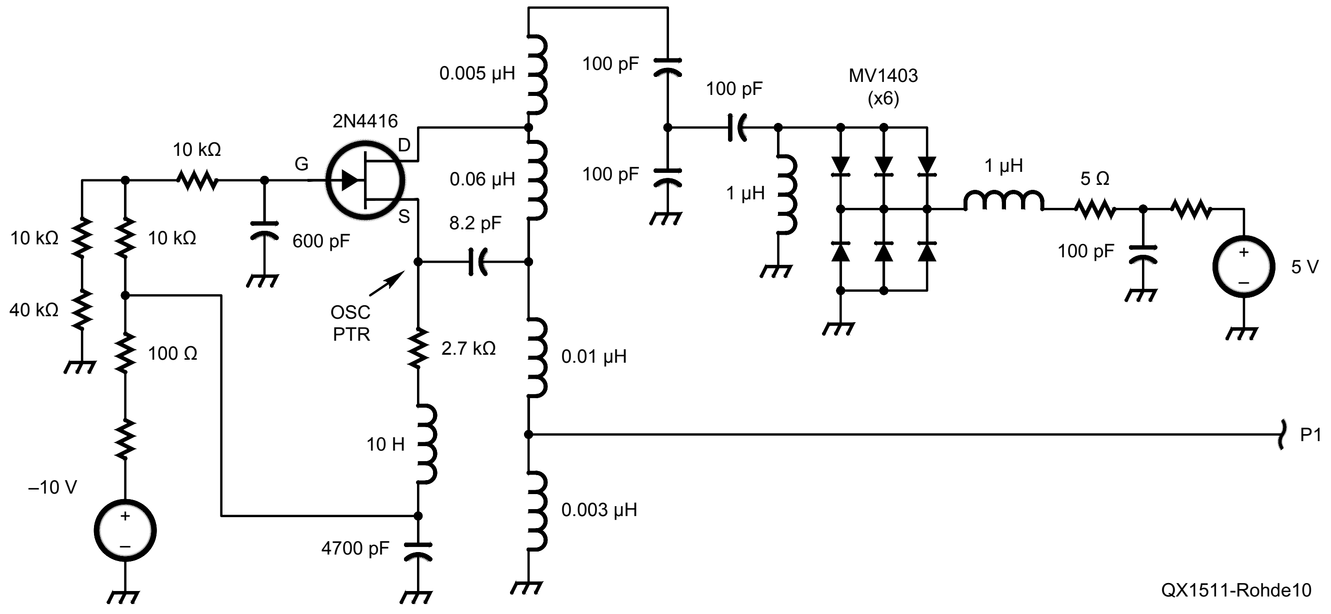

Figure 10 — Here is a typical circuit diagram of the 144 MHz low noise VCO using a 2N4416 FET. Note the six diodes for a wider tuning range of the oscillator. [R&S SMDU]

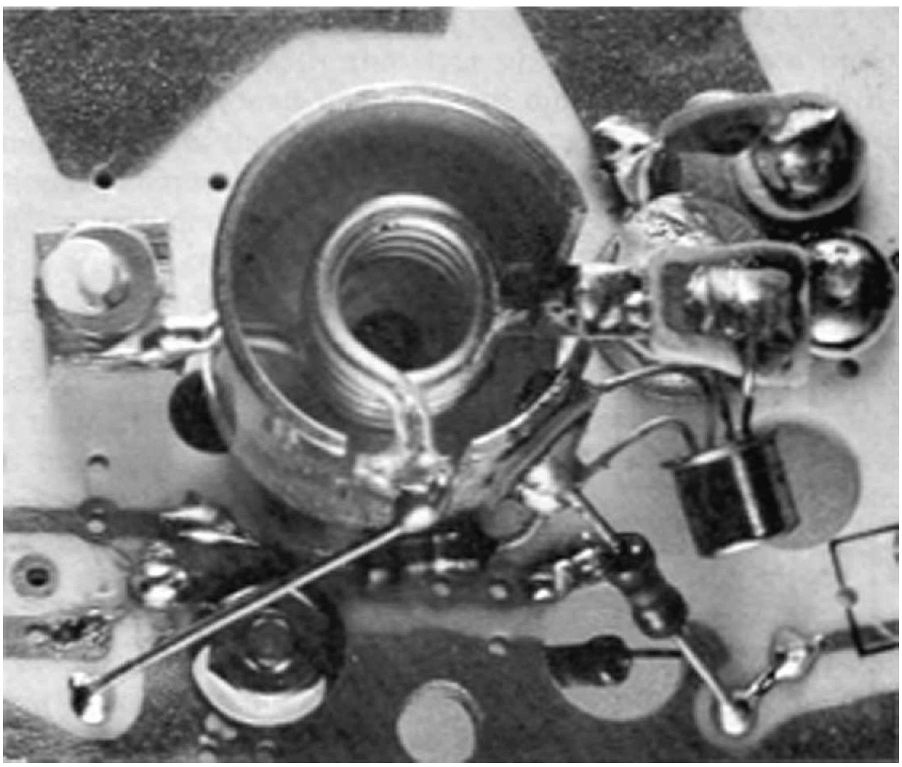

Figure 11 — This photo shows the helical resonator as part of the actual implementation of the oscilltor of Figure 9. Now the phase noise will be interesting. In practice, such an oscillator will have a buffer stage. The buffer stage will make the far off noise worse, so the result will be limited to about –165 dBc/Hz. This oscillator has an output level of 10 dBm (10 mW). The theoretical noise limit is 177 dB + 10 dB ≧ 187 dBc/Hz. The difference is due to the large signal noise figure of the transistor. [R&S SMDU]

complete PLL chips, so the individual designs disappear. Also the crystal chapter written by Roger Clark, then from Vectron, gives very valuable insight into this topic.

RF/Microwave Circuit Design for Wireless Applications, second edition, is a complete desk reference book, which also covers CMOS designs, and spends many pages on oscillators and CAD use.

The Design of Modern Microwave Oscillators for Wireless Applications addresses the very latest of wideband VCO design and push-push oscillators, and provides all the interesting phase noise calculations and design rules.

Based on the mathematics and design rules shown above, and good test equipment to validate the data, the design has become much easier.

As to the Horrabin oscillator, in one of his e-mails he mentioned a *Q* of 70 and the simulation supports that.

The improved oscillator above (no PLL!) at 3 kHz has a phase noise of –135 dBc/Hz while the Horrabin PLL design sits at –120 dBc/Hz. At higher frequencies the measured data published by Colin Horrabin supports a well-designed PLL based oscillator, but *not* any advantage of a symmetrical design. The practical designs above for a 144 MHz bipolar transistor based oscillator and this VCO gives some insight in good designs, both from a mathematical point and from a practical point.

Ulrich L. Rohde, N1UL, studied electrical engineering and radio communications at the Universities of Munich and Darmstadt, Germany. He holds a PhD in electrical engineering (1978) and a ScD (Honorary, 1979) in radio communications, a Dr-Ing (2004), University of Berlin, Germany in oscillator circuits and several honorary doctorates. In 2011 he earned a Dr-Ing Habil. Degree from the University of Cottbus, Germany.

He is President of Communications Consulting Corporation; Chairman of Synergy Microwave Corporation, Paterson, New Jersey; and a partner of Rohde & Schwartz, Munich, Germany. Previously he was President of Compact Software, Inc, Paterson, New Jersey; and Business Area Director for Radio Systems of RCA, Government Systems Division, Camden, New Jersey. He is a Professor of RF Microwave Circuit Design at Cottbus and has held Visiting Professorships at several universities in the United States and Europe.

Dr Rohde holds 25 patents and has published more than 200 scientific papers and has written or contributed to many books.

Dr Rohde is an ARRL Life Member, and is a Fellow of the IEEE, with positions on many IEEE Committees and Societies. In addition to his US call sign, he has held German call signs (DJ2LR/DL1R) since 1956 as well as Swiss call sign HB9AWE.

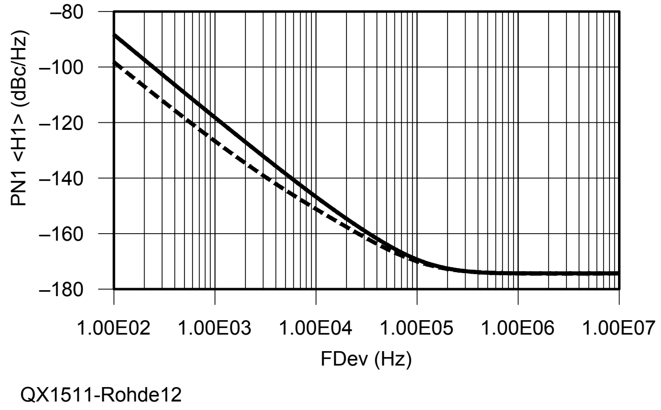

Figure 12 — Here is the phase noise simulation of the 60 MHz oscillator, using a helical resonator and tuning diodes.

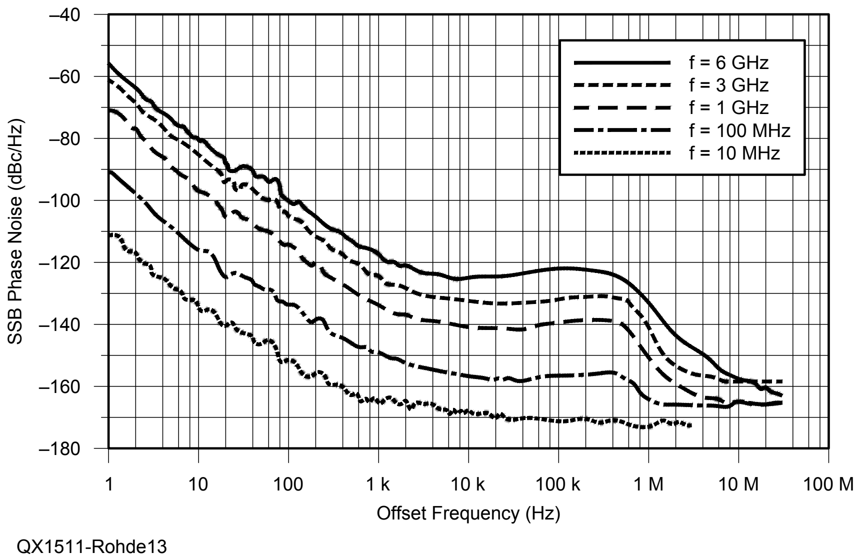

Figure 13 — Here is the measured phase noise of the oscillator of Figure 10, imbedded in a PLL system and multiplied up. For 60 MHz, the result would be between the lowest and second measured curve.

Notes

[1]Colin Horrabin, G3SBI, "The Development of the Low Phase Noise Double Tank Oscillator," Nov/Dec 2014 *QEX*, pp 35 – 43.

[2]Ulrich L. Rohde, N1UL, and Matthias Rudolph, *RF/Microwave Circuit Design for Wireless Applications*, Second Edition, John Wiley & Sons, 2013, ISBN 978-0-4-470-90181.

[3]Ulrich L. Rohde, N1UL, Juergen Schoepf, and Ajay K. Poddar, "Low-Noise VCOs Conquer Wide Bands", *Microwaves & RF*, pp. 98-106, June 2004.

[4]Ulrich L. Rohde, N1UL, and Ajay K. Poddar, "Noise Analysis of Systems of Coupled Oscillators," Integrated Nonlinear Microwave and Millimeter wave Circuits (INMMIC) workshop, Monte Porzo, Cantone, Italy, November 15-16, 2004.

[5]Ulrich L. Rohde, N1UL, and Ajay K. Poddar, "Ultra Low Noise Low Cost Multi Octave Band VCO," 2005 IEEE Sarnoff Symposium on Advances in Wired and Wireless Communication, 18–19 April 2005, Princeton, USA, pp 05 – 08.

[6]Ulrich L. Rohde, N1UL, and Ajay K. Poddar, "Novel Multi-Coupled Line Resonators Replace Traditional Ceramic Resonators in Oscillators/VCOs," 2006 IEEE International Frequency Control Symposium and Exposition, 5-7 June 2006, Florida, USA, pp 432 – 442.

[7]Ulrich L. Rohde, N1UL, *Microwave and Wireless Synthesizers: Theory and Design*, New York: John Wiley & Sons, 1997, ISBN 0-471-52019-5.

[8]Ulrich L. Rohde, N1UL, and Jerry Whitaker, *Communications Receivers*, Third Edition, McGraw Hill, New York, NY, January 2001,ISBN 0-07-13621-9.

[9]Ulrich L. Rohde, KA2WEU, "All About Phase Noise in Oscillators," Part 1, Dec 1993 *QEX*, pp 3 – 6, Part 2, Jan 1994, pp 9 – 16, Part 3, Feb 1994, pp 15 – 24.

[10]Ulrich L. Rohde, KA2WEU, "Designing Low-Phase-Noise Oscillators," Oct 1004 *QEX*, pp 3 – 12. This article is available on the ARRL website at: **www.arrl.org/files/file/Technology/ard/rohde94.pdf**.

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