VOLTAGE CONTROLLED OSCILLATORS

INTRODUCTION

In simple terms, an oscillator is an amplifier where sufficient energy is coupled back from the output to the input to become unstable and start oscillating. The output port then provides the wanted output power into the load and the overall circuit configuration determines the frequency stability and sensitivity to load changes. Figure 1 shows the feedback arrangement.

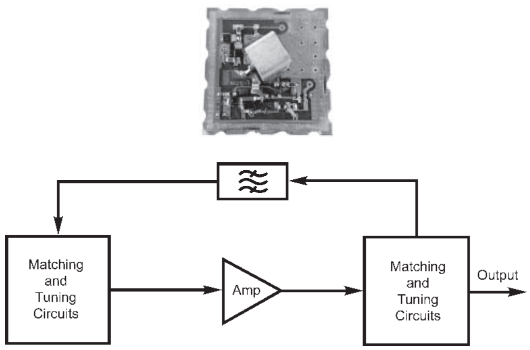

Figure 1 — An oscillator viewed as a feed-forward amplifier with positive feedback through the resonator. Start-up of oscillation requires that the gain of the amplifier exceed the loss of the resonator and that the total phase shift through the amplifier and resonator be a multiple of 360°. To sustain oscillation, the phase shift must remain the same and the amplifier gain must be equal to or greater than the resonator loss.

While there are several configurations possible, the most popular oscillator configuration up to very high frequencies is the Clapp oscillator circuit. Practically all high performance oscillators with wide tuning range follow this basic configuration. Energy can be coupled either from the collector or from the emitter of a BJT oscillator. The same basic concept applies to an oscillator using a silicon junction FET. The use of GaAsFETs is reserved for high microwave and millimeterwave applications, typically above 10 GHz. The principle of operation of this circuit is that the feedback loop generates a negative impedance (negative resistive part) which compensates for all the losses. The sum of all the resistive elements must still be slightly negative. This condition of the modified Barkhausen equation has to be met to start and maintain oscillation.

The following is an interpretation to calculate the conditions necessary for oscillation. It is based on the fact that an ideal tuned circuit (infinite Q), once excited, will oscillate infinitely because there is no resistance element present to dissipate the energy. In the actual case (where the inductor Q is finite), the oscillations die out because energy is dissipated in the resistance. It is the function of the amplifier to maintain oscillations by supplying an amount of energy equal to that dissipated. This source of energy can be interpreted as a negative resistor in series with the tuned circuit. If the total resistance is positive, the oscillations will die out, while the oscillation amplitude will increase if the total resistance is negative. To maintain oscillations, the two resistors must be of equal magnitude. To see how a negative resistance is realized, the input impedance of the circuit in Figure 2 will be derived. Figure 3 shows an equivalent small signal circuit of Figure 2.

The steady-state loop equations are

$$V_{in} = I_{in}(X_{C_1} + X_{C_2}) - I_b(X_{C_1} - \beta X_{C_2}) \quad (1)$$

$$0 = -I_{in}(X_{C_1}) + I_b(X_{C_1} + h_{ie}) \quad (2)$$

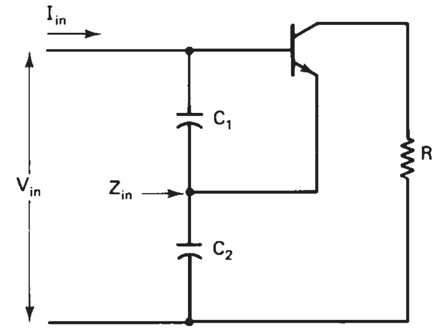

Figure 2 — Calculation of input impedance of the negative-resistance oscillator.

After I_b is eliminated from these two equations, Z_{in} is obtained as

$$Z_{in} = \frac{V_{in}}{I_{in}} = \frac{(1+\beta)X_{C_1}X_{C_2} + h_{ie}(X_{C_1} + X_{C_2})}{X_{C_1} + h_{ie}} \quad (3)$$

if $X_{C1} \ll h_{ie}$, the input impedance is approximately equal to

$$Z_{in} \approx \frac{1+\beta}{h_{ie}} X_{C_1}X_{C_2} + (X_{C_1} + X_{C_2}) \quad (4)$$

$$Z_{in} \approx \frac{-g_m}{\omega^2 C_1 C_2} + \frac{1}{j\omega[C_1C_2/(C_1+C_2)]} \quad (5)$$

That is, the input impedance of the circuit shown in Figure 3 is a negative resistor,

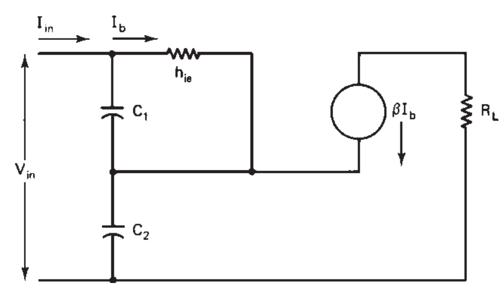

Fig. 3 - Equivalent small-signal circuit of Figure 2.

$$R = \frac{-g_m}{\omega^2 C_1 C_2} \quad (6)$$

in series with a capacitor,

$$C_{in} = \frac{C_1 C_2}{C_1 + C_2} \quad (7)$$

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which is the series combination of the two capacitors. With and inductor L (with the series resistance R_s) connected across the input, it is clear that the condition for sustained oscillation is

$$R_s = \frac{g_m}{\omega^2 C_1 C_2} \quad (8)$$

and the frequency of oscillation

$$f_o = \frac{1}{2\pi\sqrt{L[C_1 C_2 / (C_1 + C_2)]}} \quad (9)$$

This interpretation of the oscillator readily provides several guidelines which can be used in the design. First, C_1 should be as large as possible so that

$$X_{C_1} << h_{ie}$$

and C_2 is be large so that

$$X_{C_2} << \frac{1}{h_{oe}}$$

when these two capacitors are large, the transistor base-to-emitter and collector-to-emitter capacitances will have a negligible effect on the circuit's performance. However, Eq. (8) limits the maximum value of the capacitances since

$$r \le \frac{g_m}{\omega^2 C_1 C_2} \le \frac{G}{\omega^2 C_1 C_2} \quad (10)$$

where G is the maximum value of g_m. For a given product of C_1 and C_2, the series capacitances is a maximum when $C_1 = C_2 = C_m$. Thus Eq. (10) can be written

$$\frac{1}{\omega C_m} > \sqrt{\frac{r}{G}} \quad (11)$$

This equation is important in that it shows that for oscillations to be maintained, the minimum permissible reactance $(1/\omega C_m)$ is a function of the resistance of the inductor and the transistor's mutual conductance g_m.

An oscillator circuit known as the *Clapp* circuit or *Clapp-Gouriet* circuit is shown in Figure 4. This oscillator is equivalent to the one just discussed, but it has the practical advantage of being able to provide another degree of design freedom by making C_o much smaller than C_1 and C_2. It is possible to use C_1 and C_2 to satisfy the condition of Eq. (10) and then adjust C_o for the desired frequency of oscillation ω_o which is determined from

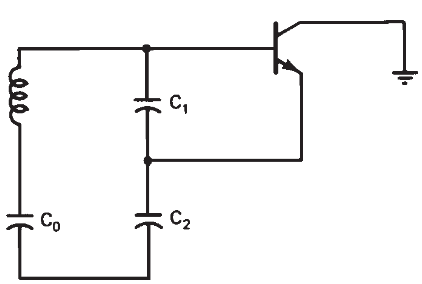

Fig. 4 — Circuit of Clapp oscillator.

For frequency applications above several hundred MHz, the discrete inductance is exchanged to be a transmission line, and for even higher frequencies (such as 800 MHz and higher), a ceramic resonator of high Q is the best choice. Also, one can use SAW oscillators (surface acoustic wave resonators) and dielectic-resonator-based oscillators.

$$\omega_o L - \frac{1}{\omega_o C_o} - \frac{1}{\omega_o C_1} - \frac{1}{\omega_o C_2} = 0 \quad (12)$$

These types of oscillators can be made voltage controlled oscillators by changing capacitor C_O from a fixed value to a voltage dependent capacitor, commonly referred to as tuning diode or varactor. The use of just a single diode is typically discouraged. For small DC voltages, the tuning diode becomes conductive in the positive half of the sine wave, and this reduces the Q and deteriorates the phase noise performance. As a minimum, a high-performance oscillator requires one set of anti-parallel diodes. A high-performance VCO is shown in Figure 5. A more detailed introduction to designing VCOs is found in References 1 and 2.

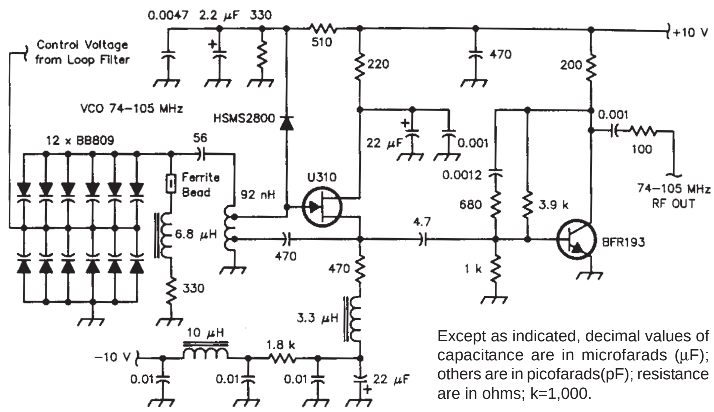

Fig. 5 — Improved phase-noise performance can be obtained by using a number of tuning diodes in the antiparallel arrangement shown here. The 92-nH coil is a 4-turn coil tapped at 1 and 2 turns from the ground end.

Wideband VCOs vs. Narrowband VCOs

The tuning range of an oscillator is determined by the amount of fixed capacitance versus available capacitance in the circuit. In many cases (such as in cellular telephone applications), the tuning range required is limited to 5% (for example, 50 MHz relative to 1000 MHz). Some applications require ranges of up to 2:1, however. In these cases, tuning diodes with a high capacitance range are required. These diodes are frequently referred to as hyperabrupt diodes. The drawback of hyperabrupt diodes is that their transfer characteristics, or change of capacitance as a function of voltage, is not very linear.

Figure 6 shows the capacitance voltage characteristic for three different types of diodes. The operating range of the capacitance diode or (its useful capacitance ratio) is limited by the requirement that the diode must not be driven into forward conduction or breakdown by the RF voltage superimposed on the tuning voltage. Otherwise, rectification would take place, shifting the bias of the diode and considerably affecting its figure of merit.

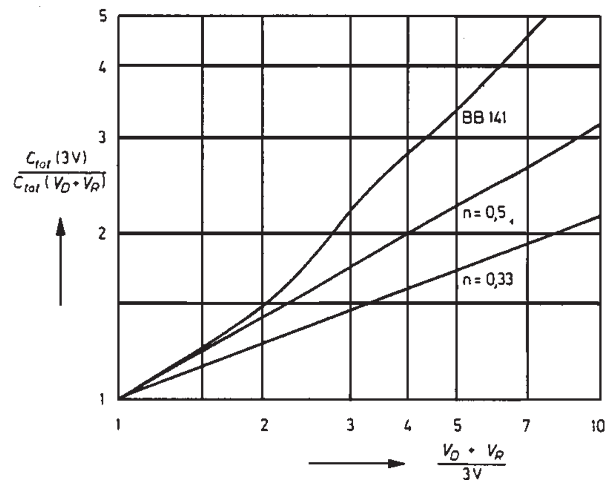

Fig. 6 — Capacitance/voltage characteristic for
a) an alloyed capacitance diode
b) a diffused capacitance diode
c) a wide-range tuner diode (BB141)

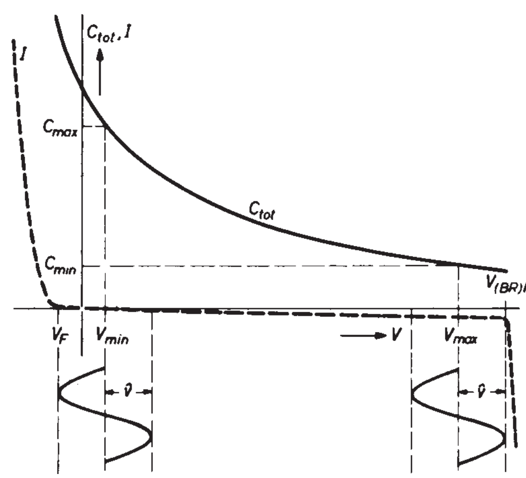

Fig. 7 — Basic current/voltage and capacitance/voltage characteristics of a tuning diode.

As a result of a high alternating voltage at the resonant circuit, capacitance contributed by tuning diodes may produce three undesirable effects:

1) The nonlinearity of the capacitance may generate more harmonics than the oscillator circuit produces by itself. The RF voltage across the diodes should therefore be kept small.

2) Even when a sinusoidal voltage is applied to the tuning diodes, their capacitance variation does not follow the sine law. Depending on the amplitude of the applied voltage, this may result in a change of the resonator frequency. Under certain conditions, this can even lead to bistable behavior, and the oscillator may show an effect called *squegging.*

3) Intermodulation, a disturbance caused by nonlinearities in the diode characteristics, is virtually independent of the oscillator amplitude. It can result in the transfer of an undesired signal (noise) to the oscillator output signal (carrier). The undesired signal is the noise internally generated by the oscillator transistor(s). This intermodulation, the mechanism of which is also referred to as *AM-to-PM conversion*, occurs in BJT PN and NP junctions and their FET equivalents.

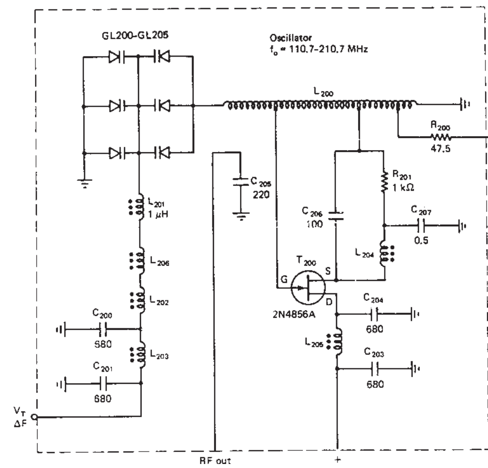

Fig. 8 — Practical circuit of a wideband VCO.

Phase Noise

One of the key characteristics of the oscillator is the oscillator noise frequently referred to as single *side-band phase noise.*

Oscillator Noise

The SSB phase noise of an oscillator is well described by the Leeson model and its enhancements. Leading engineers such as M. Driscoll of Westinghouse and Tom Parker of Raytheon have expanded the model to include the flicker-noise component even of passive components. The enhanced Leeson equation, as proposed by Parker and others, is shown in Figure 9.

$$
\begin{aligned}
S_\phi(f_m) = &\left[\alpha_R F_o^4 + \alpha_E \left(F_o / (2Q_L)\right)^2\right] \Big/ f_m^3 \\
&+ \left[(2GFKT / P_o)\left(F_o / (2Q_L)\right)^2\right] \Big/ f_m^2 \\
&+ (2\alpha_R Q_L F_o^3) \Big/ f_m^2 \\
&+ \alpha_E / f_m + 2GFKT / P_o
\end{aligned}
$$

Fig. 9 — The Leeson equation for oscillator phase noise. *G* is the compressed power gain of the loop amplifier, *F*, is the noise factor of the loop amplifier, *K* is Boltzmann's constant, *T* is the temperature in Kelvins, P_o is the carrier output power, in watts, at the output of the loop amplifier, F_o is the carrier frequency in hertz, f_m is the carrier offset frequency in hertz, Q_L is the loaded Q of the resonator in the feedback loop, and α_R and α_E are the flicker noise constants for the resonator and loop amplifier, respectively.

Numerous circuits have been developed to implement oscillators. These circuits have various advantages and disadvantages, depending on the frequency of operation and the resonator type. For circuits in the 400 to 2000 MHz range, modern oscillators tend to use transmission-line resonators and capacitive feedback of the Colpitts or Clapp type. At these frequencies, bipolar transistors are generally used, since few FETs have sufficient gain-bandwidth product for use in UHF oscillator circuits.

Requirements for Low-Noise Oscillators — The key elements that determine the phase noise of an oscillator are:

- the transistor's flicker-noise corner frequency, which depends on the device current;
- the loaded Q of the resonator, which depends on the coupling between the resonator and the transistor; and
- the ultimate signal-to-noise ratio, which depends on the RF output power of the oscillator and its large-signal noise figure.

Of these, the first two can be investigated using linear circuit analysis. But the active device's large-signal operation requires nonlinear analysis techniques, without which we can make only educated approximations of the ultimate signal-to-noise ratio.

The Linear Approach

The design goal of the linear approach is to achieve the maximum loaded Q of the resonator and to keep the bias (DC) device current to a minimum. A high Q helps restrict noise components to frequencies close to the frequency of oscillation, minimizing phase noise as we move away from that frequency. The requirement for minimum bias exists because the flicker, or 1/f, noise of the device is highly dependent on the current. Table 1 shows the flicker-noise corner frequency versus collector current for a typical bipolar transistor. JFETs have much less flicker noise than bipolar tansistors, while GaAsFETs have more.

Table 1 - Flicker Corner Frequency vs. Collector Current for a Typical Bipolar Transistor

I_c (mA)	F_c (kHz)
0.25	1
0.5	2.74
1	4.3
2	6.27
5	9.3

Oscillator Operation — At start-up, the oscillator's open-loop gain must be sufficient to begin oscillation. The circuit's amplitude stabilization mechanism is responsible for sustaining oscillation. We can view the oscillator as a two-terminal negative-resistance generator, as shown in Figure 10. Here, the total resistance—the sum of the resonator resistance and the resistance of the two-terminal oscillator—must be less than or equal to zero for oscillation. The net reactance will be zero at resonance.

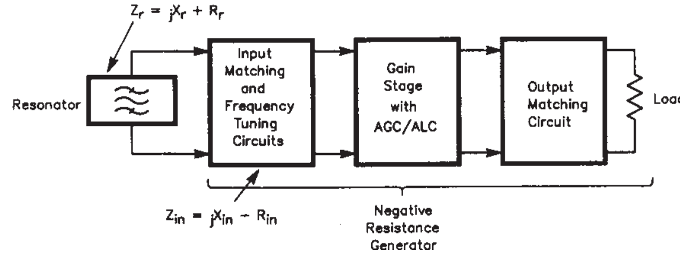

Fig. 10 - An oscillator viewed as a resonator and a negative resistance generator. At start-up, the resonator and oscillator reactances must be equal in value and opposite in sign, while the sum of the resonator and oscillator resistances must be less than 0. For sustained oscillation, the sum of the resistances must not become positive.

Although we can't precisely analyze the large-signal operation of a circuit using wholly linear techniques, we should recognize some effects that will impact our linear analysis. Chief among these is bias shift. The large signals present in the circuit in a bipolar oscillator will cause a shift in the bias current, because of the nonlinearity of the base-emitter junction. The device current may be about 10% different from the nominal (no-signal) current and may shift in either direction (more current or less). Since the flicker noise is bias-dependent, this effect is important to keep in mind.

The recommended approach to finding the bias-dependent loading of the resonator by the active device is to construct a linearized model of the device using its measured S-parameters at a particular bias point. This is especially important at higher frequencies. For simplicity, we have chosen not to do this in the example that follows, but to use a simple model.

Over a wide range of current, the device f_t remains constant. Since:

$$f_t = \frac{1}{2\pi R_d C_e} \qquad (13)$$

where R_d is the emitter diffusion resistance and C_e is the emitter capacitance, and since:

$$R_d = \frac{26\text{ mV}}{I_E} \qquad (14)$$

(at room temperature), where I_E is the emitter bias current, we can therefore adjust the R_d and C_e parameters of our device model to reflect the bias current we expect to use. This will allow our linear circuit model to reflect the bias dependency of the oscillator.

As mentioned before, flicker noise is dependent largely on the bias current. But the *effect* of flicker noise can be reduced. This noise contributes to the phase noise by modulating the oscillator's frequency via AM-to-PM conversion. We can reduce this modulation by use of negative feedback. A simplified noise model of a transistor is shown in Figure 11.

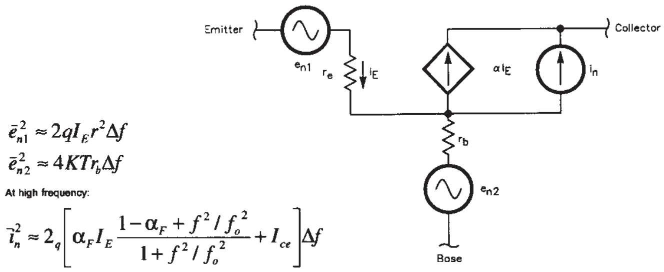

Figure 11 — A simplified bipolar transistor noise model with

The effect of applying negative feedback to reduce phase noise is shown in Figure 12.

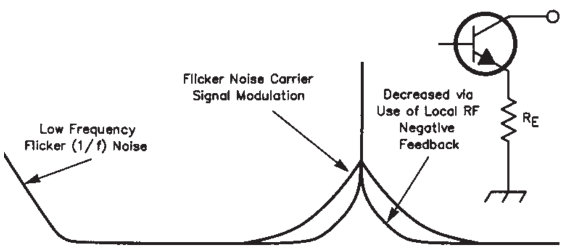

Fig. 12 — Adding negative feedback can reduce the amount of AM-to-PM modulation of the carrier by the transistor's flicker noise.

VCO Noise — So far, we have considered only the noise from the transistor. When we extend our design to become a VCO, by adding a tuning diode, we must also consider the phase noise introduced by that diode. Contrary to what you may have read elsewhere, this noise is not due solely to Q reduction from the added diode. The diode itself introduces noise that modulates the VCO frequency. The easiest way to analyze this noise is to treat the diode's noise contribution as that of an equivalent resistance, R. This resistor can then be considered to be generating the thermal noise voltage that any resistance exhibits:

$$V_n = \sqrt{4KT_0 R\Delta f} \tag{15}$$

where V_n is the open-circuit RMS thermal noise voltage across the diode, K is Boltzmann's constant, T_0 is the temperature in Kelvins, R is the equivalent noise resistance of the tuning diode, and Δf is the bandwidth we wish to consider. At room temperature (about 300 K), KT_0 is 4.2x10^{-21}.

Practical values of R for tuning diodes range from about 1 kΩ to 50 kΩ. For a value of 10 kΩ, for example, we would find a noise voltage from Eq. 15 of:

$$V_n = \sqrt{4\times 4.2\times 1-^{-21}\times 10{,}000}$$
$$= 1.265\times 10^{-8}\ \ \mathrm{V}/\sqrt{\mathrm{Hz}}$$

This noise voltage from the tuning diode modulates the frequency of the oscillator in proportion to the oscillator's VCO gain, K_0, (the frequency swing per volt of the tuning signal):

$$(\Delta f_{rms}) = K_0 \times (1.265\times 10^{-8}\,\mathrm{V}) \tag{16}$$

in a 1-Hz bandwidth. This can be related to the peak phase deviation, θ_d :

$$\theta_d = \frac{K_0\sqrt{2}}{f_m}(1.265\times 10^{-8}\,\mathrm{rad}) \tag{17}$$

in a 1-Hz bandwidth, where f_m is the frequency offset of the noise from the oscillator operating frequency. Applying a typical VCO gain of 100 kHz/V gives a typical peak phase deviation of:

$$\theta_d = \frac{0.00179}{f_m}\,\mathrm{rad} \tag{18}$$

in a 1-Hz bandwidth. For an offset of 25 kHz, as might be used to find the noise in an adjacent FM channel, this gives θ_d=7.17x10^{-8} rad in a 1-Hz bandwidth. Finally, we can convert this result into the SSB signal-to-noise ratio at the specified frequency offset:

$$L(f_m) = 20\log_{10}\frac{\theta_d}{2} = -149\,\mathrm{dBc/Hz} \tag{19}$$

It is also worth noting that the nonlinear capacitance vs. voltage characteristic of a tuning diode results in a tuning sensitivity—and thus a noise performance—that depends on the input tuning voltage. Modern CAD tools can help in the analysis of an oscillator.

Figure 13 shows the circuit of a typical 950 MHz ceramic resonator based oscillator.

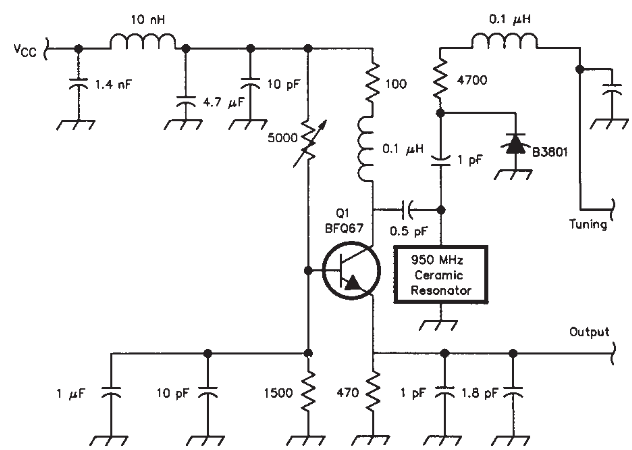

Fig. 13 — Circuit of typical 950-Mhz ceramic resonator based oscillator.

Figure 14a shows its measured performance while Figure 14b shows its CAD prediction. As can be seen from Figure 15, phase noise increases as the tuning sensitivity of the VCO is increased.

While oscillator design is done at Synergy Microwave, we have extensive use of CAD tools, such as Compact Software's Microwave Harmonica.

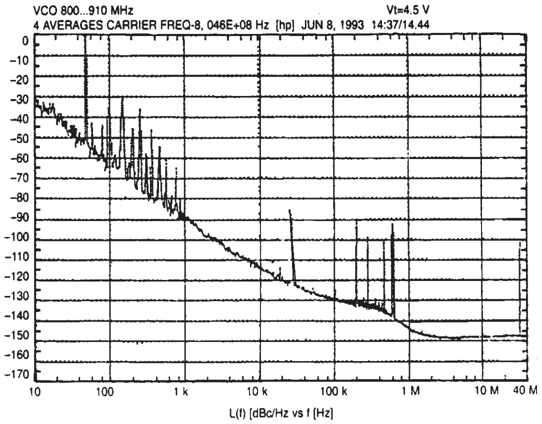

Fig. 14a — Measured phase noise of the oscillator of Figure 13.

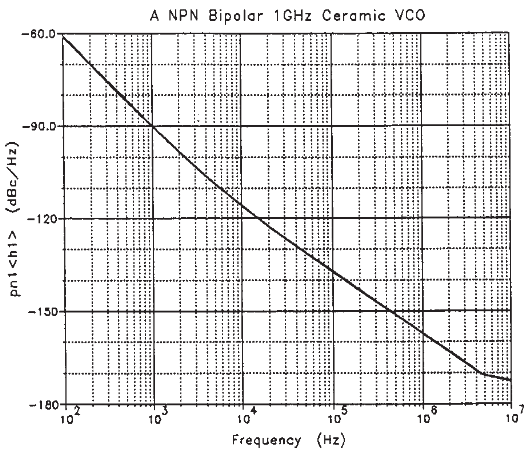

Fig. 14b — Predicted phase noise of the 1-GHz ceramic resonator VCO with the tuning diode attached. Note the good agreement between the measured and predicted phase noise.

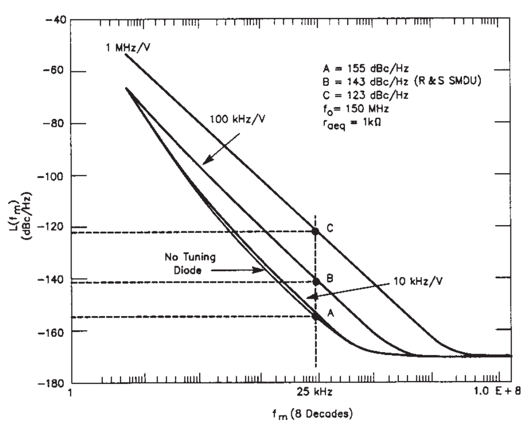

Fig. 15 — Phase noise increases as the tuning sensitivity of the VCO is increased, as shown here ($r_{aeq.}$ is the equivalent noise resistance of the tuning diode).

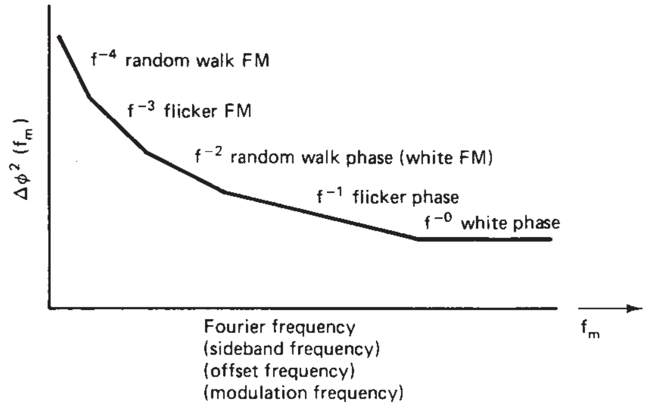

Fig. 16 — Characterization of noise sideband in the

VCO TERMINOLOGY

In specifying VCOs, the following list of definitions may help to get a better understanding of what to look for:

VCO (Voltage Controlled Oscillator): The output frequency of the oscillator is determined by a DC control voltage. This applied voltage tunes the oscillator over a specified range.

Phase noise: This term describes the "single sideband phase noise" of the oscillator. It is composed of noise close to the carrier (flicker noise) and noise measured at a spacing of 20 kHz and greater. The flicker noise measured in a VCO is generated only by the active devices, such as the transistor and the tuning diode. For silicon junction FET, the flicker corner frequency is typically at 50 Hz to 100 Hz. Microwave bipolar transistors have a flicker corner frequency of up to 5 kHz. More exotic devices, such as heterojunction bipolar transistors (HBTs), have flicker corner frequencies up to 50 kHz but also have f_T corner frequencies of up to 100 Hz. Finally, GaAsFETs have a corner frequency of several megahertz. The phase noise is measured at distances from 1 Hz off the carrier to several megahertz off the carrier in a 1-Hz bandwidth. Phase noise is the ratio of the output power divided by the noise power at a specified value and is expressed in dBc/Hz. Figure 16 shows a typical phase-noise plot. The area between the flicker corner frequency and the residual noise flow is proportional to $1/Q^2$ and is only determined by the quality of the resonator.

Output power: The output power of the oscillator, typically expresssed in dBm, is measured into a 50-Ω load. The output power is always combined with a specification for flatness or variation. A typical spec would be 0 dBm output power ±1 dB tolerance.

Output power as a function of temperature: All active circuits change in their behavior as a function of temperature. The output power over a temperature range should vary less than a specified value, such as 1 dB.

Harmonic output power: The harmonic content is measured relative to the output power. Typical values are 20 dB suppression or more relative to the fundamental. This suppression can be improved by additional filtering.

Spurious outputs: A VCO's spurious output specification, expressed in decibels, enumerates the strength of unwanted and nonharmonically related components relative to the oscillator fundamental. Because a stable, properly designed oscillator is inherently clean, such *spurs* are typically introduced only by external sources in the form of radiated or conducted interference. *See Harmonic output power*.

Frequency tuning characteristic: This shows the relationship, depicted as a graph, between a VCO's operating frequency and the tuning voltage applied. Ideally, the correspondence between operating frequency and tuning voltage is linear. *See Tuning linearity.*

Tuning linearity: For stable synthesizers, a constant deviation of frequency versus tuning voltage is desirable. It is also important to make sure that there are no breaks in the tuning range—for example, that the oscillator does not stop operating with a tuning voltage of 0. *See Frequency tuning characteristic***.**

Tuning performance: This datum, typically expressed in megahertz per volt (MHz/V), enumerates how much a VCO's frequency changes per unit of tuning-voltage change.

Tuning speed: This characteristic is defined as the time necessary for the VCO to reach 90% of its final frequency on the application of a tuning-voltage step. Tuning speed depends on the internal components between the input pin and tuning diode—including, among other things, the capacitance present at the input port. The input port's parasitic elements determine the VCO's maximum possible modulation bandwidth.

Post tuning drift: After a voltage step is applied to the tuning diode input, the oscillator frequency may continue to change until it settles to a final value. This post tuning drift is one of the parameters that limits the bandwidth of the VCO input.

Temperature drift: Although the synthesizer is responsible for locking and maintaining the oscillator's frequency, the VCO's frequency change as a function of temperature is a critical parameter and must be specified. Its value varies between 10 kHz/°C to several hundred kHz/°C depending on the center frequency and tuning range.

Sensitivity to load changes: This is called *frequency pulling* and the change of frequency resulting from a partially reactive load. Dividing from 50 Ω is used as a determination. Frequency pulling must be minimized, especially in cases where power stages are close to the VCO unit and short pulses may affect the output frequency. Such feedback may make locking impossible.

Frequency pushing: This is another case of the oscillator frequency being affected by external influences, typically by supply voltage. Taking the same example as above, a sudden current surge caused by the output amplifier of a two-way radio may produce a spike on the VCO's DC power supply and a consequent frequency jump.

References:

Ulrich L. Rohde, KA2WEU, "All About Phase Noise in Oscillators, Part III - Example oscillator circuits and their noise performance," QEX, *February 1994.*
Ulrich L. Rohde, KA2WEU, "Designing Low-Phase-Noise Oscillators," QEX, *October 1994.*
Ulrich L. Rohde, Microwave and Wireless Synthesizers: Theory and Design, *by John Wiley & Sons, New York, NY, ISBN: 0471-52019-5 , August, 1997 .*