HEWLETT® PACKARD

Voltage-Controlled Oscillators Evaluated for System Design

Application Note—M024

High tuning speed, small size, and low power consumption make VCOs important components in a number of microwave applications. The designer must be able to evaluate these devices accurately in order to write proper VCO specifications for critical applications.

With its unique combination of very high tuning speed (typical full-band tuning in less than 30 ns.), small size, and low power consumption, the varactor-tuned, voltage-controlled oscillator (VCO) is a vital component in electronic defense systems, frequency-hopping radar, frequency synthesizers, and many other microwave applications. Unfortunately, high-performance microwave VCOs are difficult to design, build, and optimize.

Most microwave system designers find it far more practical to consider VCOs or integrated oscillator subsystems as components, and purchase them from one of the specialized manufacturers. Available production "raw" VCOs and VCO assemblies, which offer a wide selection of optimized performance features, are readily available in the 300 MHz to 18 GHz range. This article is intended primarily to help system designers evaluate available VCOs, and to assist them in writing VCO specifications for critical applications.

A Look at Oscillator Fundamentals

In its simplest form, a sinusoidal or quasi-sinusoidal oscillator (as opposed to such square-wave or pulse sources as the blocking or relaxation oscillator) can be modeled as the combination of an amplifier with a positive feedback loop and a frequency-determining network. Figure 1 illustrates this concept. The general feedback formula is

$$A_f = \frac{A}{1-\beta_f A} \quad \text{where} \tag{1}$$

A = gain of the amplifier without feedback

A_f = gain of the amplifier with feedback

β_t = reverse transfer function of the feedback path

$$= \frac{V_f \text{ angle } \theta_1}{V_o \text{ angle } \theta_2} \tag{2}$$

(the difference between θ1 and θ2 represents the phase shift through the path)

When feedback is positive, β_f is a positive quantity and, according to the Barkhausen criterion, when $\beta_f A=1$, A_f becomes infinite and the feedback amplifier becomes an oscillator. Since both the amplifier and the feedback circuit contain capacitive or inductive energy-storage elements (even if only due to parasitics), $\beta_f A$ is complex. To satisfy the Barkhausen criterion, the real part of $\beta_f A=1$ and the imaginary part is 0; thus the real part is unity and the phase is 0. Clearly, under these conditions, once oscillation begins due to any small amount of noise at the input of the amplifier, the signal would build up until the output of the amplifier reaches its saturated limit.

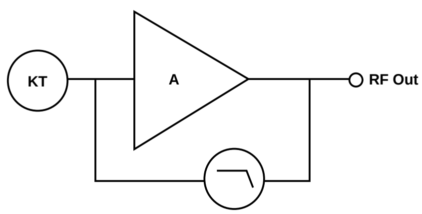


Figure 1. Block diagram of a microwave oscillator, modeled as a combination of an amplifier with a positive-feedback loop and a frequency-determining network. KT represents the noise input that starts oscillation.

The common-emitter amplifier circuit provides a nominal 180° phase shift, with the feedback network adding the additional 180° (or 540° . . . 360n + 180°). In the common-collector amplifier circuit, where the phase shift is nominally 0°, the feedback network must provide a full 360° phase shift.

The model in Figure 1 represents essentially a surface-acoustic-wave (SAW) oscillator in which the feedback path consists of a delay line and low-pass filter. Of course, if the transistor has a sufficiently high f_{max}, there is always a possibility that a simple delay will provide a total 360° phase shift at a fundamental frequency and at a number of harmonically related frequencies, or that other 360° feedback paths at non-harmonically related frequencies will exist since conductor lengths are a significant part of an electrical wavelength.

Actually, most microwave-oscillator designers use a different model, and analyze oscillators in terms of a negative-resistance generator and resonator. An example of a negative-resistance generator is shown in Figure 2. By first transforming it to its equivalent circuit (Figure 3), the input impedance of this circuit can be shown to generate a negative-resistance. It can be shown through nodal analysis that the input admittance of this circuit is:

$$Y_{in} = \frac{1}{R}\left\{\frac{1 - f_c^2}{f^2}\right\} + \frac{1}{j2\pi f1} \qquad (3)$$

if $f_c > f$ $\quad$ Re $|Y_{in}|$ <
$\quad$ Im $|Y_{in}|$ Inductive

$$\text{where } f_c = \frac{1}{2\pi\sqrt{LC}} \qquad (4)$$

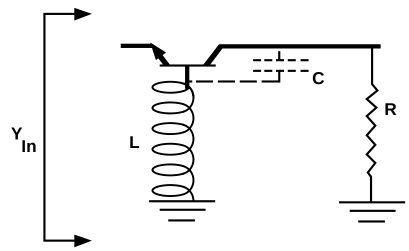


Figure 2. Example of a negative-resistance generator. Most designers model microwave oscillators as a negative-resistance generator and oscillator.

This equation demonstrates that a bipolar transistor with an inductor between base and ground becomes the equivalent of negative conductance in parallel with an inductive susceptance over a range of frequencies f_c>f. The inductor L is selected by design, whereas the value of C is provided by the collector-base capacitance of the transistor.

The common-base circuit is not the only topology that can be used to generate a negative resistance, and in practice a common-emitter configuration with the resonator in the base circuit is used as frequently as the former. The selection of an appropriate circuit topology is a complex task, dictated by such factors as the type of resonator being used, the characteristics of the transistor, and the frequency of the oscillator. The complete VCO is produced by adding a parallel or series resonant circuit (Figures 4 and 5).

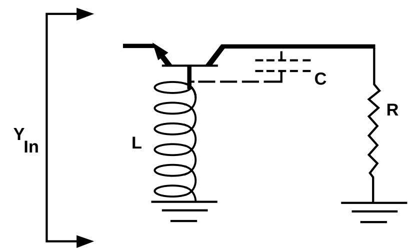


Figure 4. When the designer adds a parallel resonant circuit to the equivalent circuit shown in Figure 2, the result is a complete voltage-controlled oscillator.

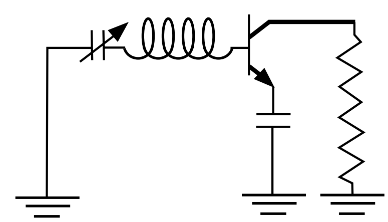


Figure 5. VCOs may also be constructed by adding a series resonant circuit to the negative-resistance generator. In this example, it is added to the base circuit in a common-emitter amplifier.

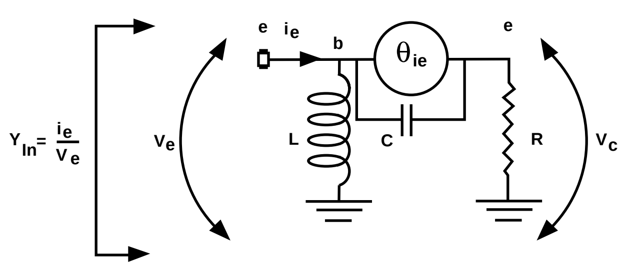


Figure 3. The equivalent circuit of the negative-resistance generator shown in Figure 2. Through nodal analysis, the input impedance of this circuit can be shown to generate a negative resistance.

Phase (FM) Noise — A Significant Criterion

The major concern of synthesizer designers is the phase stability (or phase noise). It is also critically important in EMC systems, frequency-agile radar systems, Doppler radar systems, radar warning receivers, and various communications applications. In such applications, an oscillator's phase-noise output may set the system's limits for dynamic range and reception sensitivity. The output of an ideal sine-wave oscillator can be described as:

$V(t) = V_0 \sin 2\pi\mu_0 t$

where V_0 is the nominal amplitude and μ_0 the nominal frequency. For an actual sine wave, the equation becomes:

$$V(t) = [V_0 + E(t)] \sin [2\pi\mu_0 t + \phi(t)], \quad (5)$$

where $E(t)$ is the magnitude of random variation in amplitude, and $\phi(t)$ that of phase.

The phase noise of an oscillator is best seen in the frequency domain, where special purity is determined by measuring noise power in sidebands about the output-signal center frequency (carrier). Note, though, that on a spectrum analyzer it is impossible to tell whether the power at different Fourier frequencies is a result of implitude or of phase fluctuations. Fortunately, since most oscillators operate in saturation, AM noise is limited. It is usually 20 dB lower than phase noise and can often be disregarded.

In practice, phase spectral density is measured by passing the oscillator signal through a phase discriminator, substantially amplifying the resulting discriminator-output spectrum, then displaying it on a low-phase-noise spectrum analyzer. Single-sideband phase noise is usually specified in dBc/HZ at a given frequency from the carrier. Figure 6 is an FM noise nomograph that converts between single-sideband noise power ratio and frequency deviation at any distance from the carrier.

Total FM noise can be expressed as a sideband-to-carrier power ratio by the following relationship:

$$\text{dB} = 20 \log \sqrt{2} \left\{ \frac{f_m}{\Delta f_{RMS}} \right\} \quad (6)$$

where
Fm = frequency from the carrier
Δf_{RMS} = deviation

When frequency deviation is known in a given bandwidth, the following equation can be used to normalize the frequency deviation to any reference bandwidth:

$$\Delta f_{2RMS} = \Delta f_{1RMS} \frac{B_2}{B_1} \quad (7)$$

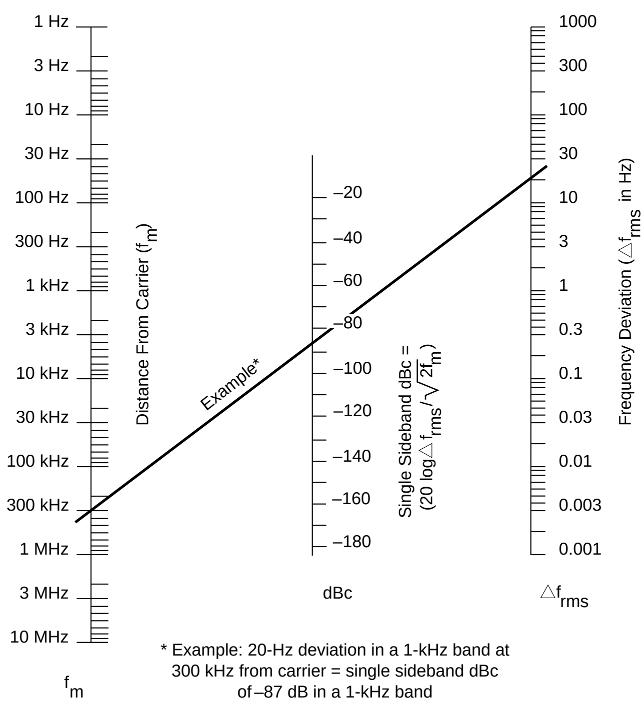


Figure 6. FM noise-conversion nomograph. By placing a straight edge on the frequency deviation in Hz (right column) and on the distance from the carrier at which that deviation occurs (left column), the single-sideband noise-power ratio in dBc is shown in the center column.

This relationship shows the importance of specifying the reference bandwidth when comparing the FM noise performance of various VCOs. The typical FM noise performance of four production VCOs is shown in Figure 7.

The phase noise generated by a VCO is determined primarily by the Q (quality factor–ratio of reactance to resistance) of the overall circuit and the Q of the varactor diode. The oscillator circuit itself is usually designed with a specific parameter in mind. In order to design a circuit with a very high Q, the tuning bandwidth must invariably suffer. Therefore, an oscillator circuit designed for optimum phase-noise performance will be ultimately a fairly narrow-band oscillator.

In most cases, choosing a varactor diode for low phase noise requires only that the highest available Q be selected for the operating frequency and tuning bandwidth required. The selection of transistors, however, is a more involved process. The transistor intended for use as a microwave oscillator must offer a high f_{max} to ensure reasonable efficiency, have a sufficiently large active area to provide adequate output power, and have a low-enough thermal resistance to ensure thermal stability.

The problem is that the f_{max} is higher for devices with smaller areas and, conversely, larger-area devices yield higher output power at lower frequencies. Thus, the transistor is selected by balancing output power vs. oscillator efficiency.

Once oscillation begins due to any small amount of noise at the amplifier's input, the signal will build up until the amplifier's output reaches its saturated limit.

The transistor with the largest device geometry or periphery that will operate at the design frequency is usually selected. For example, in designing a +10-dBm-power-output 10-GHz oscillator for low noise, a bipolar transistor with an F_t of 8 GHz and saturated output power of +20 dBm would be preferable to another device with an F_t of 12 GHz and the capability of only +13-dBm output power. As a general rule, silicon bipolar transistors are used rather than FETs in oscillators through Ku band when low noise is the most important consideration, since the phase-noise performance of silicon bipolar oscillators is typically 10 to 15 percent lower than that of FET oscillators operating under the same conditions.

The design of low-noise oscillators is complicated when frequency tuning is required. If only a narrow tuning range (< 2 percent) is needed, a cavity-stabilized or dielectric resonator oscillator with a frequency-pulling circuit — essentially an AFC with user access to the error voltage loop — is often the best choice. Thin-film VCOs can also be suitable with circuit optimization, and careful varactor and semiconductor selection.

When wider tuning bandwidths (10 to 15 percent) are required, VCO circuits using high-Q tuning diodes and low-noise silicon transistors, combined with special feedback techniques that reduce FM noise while retaining other critical VCO performance characteristics, are necessary. The best noise performance is obtained by tuning the VCO directly, without the use of a linearizer. A practical linearizer circuit will contribute AM noise to the control signal, which results in VCO phase noise at the oscillator output.

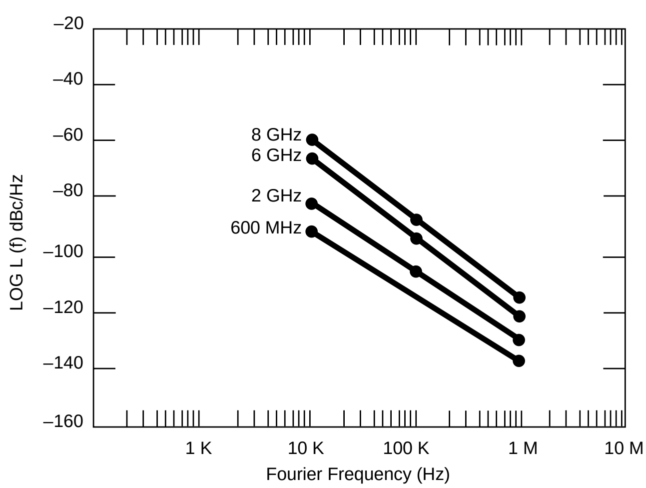


Figure 7. Single-sideband phase-noise comparison of four production VCOs.

The impedance of the tuning voltage source fed to a "raw" VCO should be very low and, in many cases, shielding may be required on interconnecting leads and wires to suppress stray pick-up. Noise specifications should include the required tuning video bandwidth, which should be as small as possible.

Power-supply voltage regulation is also a very important consideration when using low-noise VCOs. Fluctuations or noise on the bias voltage supplied to the oscillator will cause frequency pushing, which also appears as phase noise on the output-signal spectrum. Figure 8 shows the typical FM noise performance of a 600-MHz thin-film oscillator, and of an 8-GHz fundamental thin-film oscillator designed specifically for low-noise applications.

Settling Time and Post-Tuning Drift

For a VCO, settling time is defined as that interval between the time when the input-tuning-drive waveform reaches its final value and the time when the VCO frequency falls within a specified tolerance of a stated final value. In Figure 9, the input-drive waveform reaches its vinal value at time t_0. The VCO frequency reaches the lower edge of its specified tolerance band Df_{ST} at time t_{ST}. Settling time is the interval t_{ST} to t_0. The required final frequency is reached at time t_r.

Post-tuning drift is defined as the frequency drift that occurs between two arbitrarily defined times t_1 (which may be specified typically as 10 ms. to 1 sec. after the tuning step has been applied to the VCO) and t_2. For short-term PTD, time t_2 would generally be defined in the range of 10 ms. to 1 sec. after t_1; for long-term PTD, t_2 could range from 1 sec. to 1 hr. Drifting of bias voltages and thermal effects (i.e., changes in both the varactor and transistor junction temperatures) are the primary contributors to short-term PTD.

Bias circuit design is critical to PTD performance. The change in frequency due to a change in input bias is approximately 0.3 to 0.7 percent per Volt. This cannot be eliminated simply by using a well-regulated bias supply, since dramatic changes in PTD will occur due to changes in the transistor load present at the device end of the oscillator's internal bias circuitry. As the operating frequency of an oscillator is varied, the currents flowing in both the varactor and the transistor change (the reason why the frequency-vs.-output power curve is not perfectly flat), thus

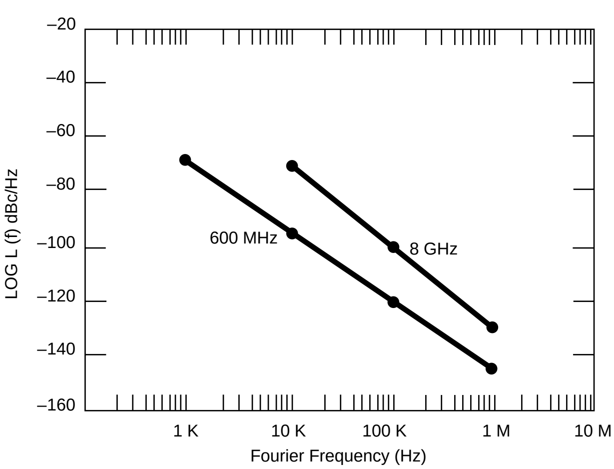


Figure 8. Low-noise VCOs typically have phase noise 10 to 15 dB lower than standard CVOs. These curves should be compared to those displayed in Figure 7.

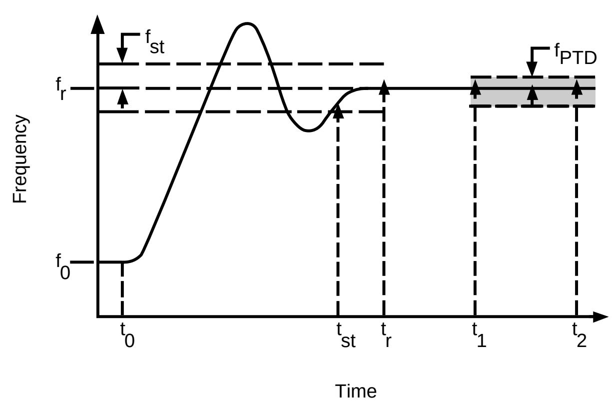


Figure 9. Frequency response of a VCO in response to a tuning-voltage step change. The input-drive waveform reaches its final value at time t_0. The VCO frequency reaches the lower edge of its specified tolerance band D_fST at time t_{ST}. Settling time is the interval t_{ST} to t_0. The required final frequency is reached at time t_r. The frequency drift that occurs between two arbitrarily defined times t_1 (which may be defined typically as 10 ms. to 1 sec. after the tuning step has been applied to the VCO) and t_2 (which is defined as "post-tuning drift").

varying the amoung of power dissipated (the efficiency of the oscillator). This means the load on the bias line varies with frequency, and the bias circuitry within the oscillator must compensate.

During the interval when a VCO is being tuned, the junction temperatures of both the transistor and varactor are also changing due to the changes in RF circuit efficiency and loading. This causes impedance changes, which result in frequency shift. The time interval during which this happens is dependent upon the thermal impedance of the devices.

Varactors used by Ku band are made typically from silicon. Above 12 GHz, GaAs varactors are used because of their higher Q. The GaAs devices have higher thermal resistance than silicon devices; this results in significantly higher short-term PTD factors.

Changes in the junction temperature of the transistor can be minimized through a reduction in input bias power. The trade-off is that reduced bias results in lower oscillator output power so that additional amplifier stages are required to bring the power output up to the required level.

Long-term PTD is affected mainly by the varactor-charging effect over a period of time. This reversible effect is caused by impurity ion buildup around the varactor junction over a long period of time. This causes a change in the capacitance of the varactor, resulting in a frequency change in the oscillator. Passivation of silicon varactors has been very successful in reducing this effect. Long-term PTD has been improved to yield less than 1 MHz frequency drift for one-hour periods. No such technique has yet been applied effectively to GaAs varactors, which display a significant charging effect.

Figure 10 is a comparison of fundamental transistor oscillators using a silicon varactor diode (VTO-8400) and a GaAs varactor diode (HTO-4000) at 4 GHz. Both settling time and post-tuning drift can also be affected by instability of the tuning signal.

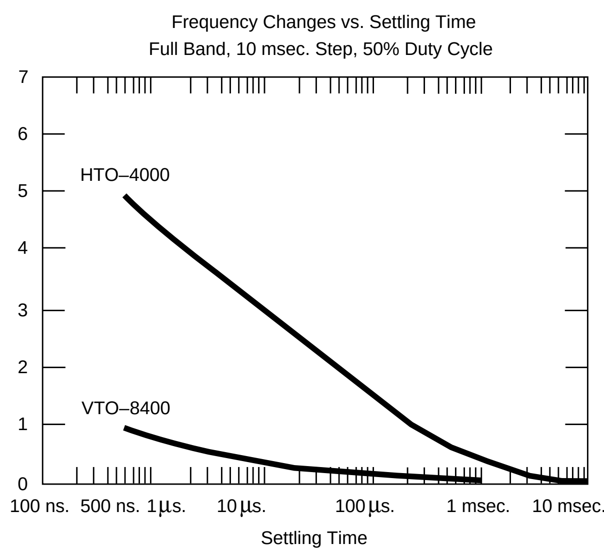


Figure 10. Comparison of the settling times of two standard VCOs at 4 GHz. Use of a silicon varactor results in substantially faster settling than does the use of a GaAs varactor.

Table I illustrates the various VCO parameters affected by the varactor. Care should be taken in selecting the correct type of VCO for a particular application, keeping in mind the trade-offs between tuning voltage limits, phase noise, settling time, and PTD. Figure 11 depicts a phase-noise comparison between a VCO using a silicon tuning diode versus one with a GaAs tuning diode, to highlight the phase-noise trade-offs.

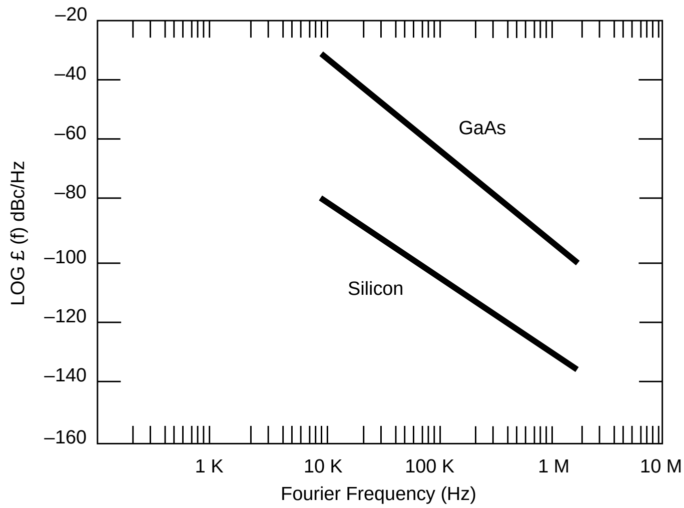


Figure 11. Single-sideband phase-noise comparison of a silicon and a GaAs varactor used in an 8-GHz VCO. The superiority of the silicon varactor in phase noise is readily apparent.

Tuning Linearity and Linearizers

Whether silicon or GaAs, there are two basic types of varactors: abrupt and hyperabrupt. The essential difference between the two is that the concentration of N-type dopant is nearly constant across the depletion region of an abrupt diode, but is nonlinear in a hyperabrupt diode. As the reverse bias is increased, the nonlinear doping profile causes a greater capacitance change in the hyperabrupt-junction diode than in the abrupt-junction diode. This results in a more nearly linear tuning curve, and a lower maximum tuning voltage.

The equation for junction capacitance vs. applied voltage for the abrupt varactor is approximated by:

$$C\,(V) = \frac{C\,(0)}{\left(1 + \frac{V}{\phi}\right)^{\gamma}} \qquad (8)$$

where
C(0) = junction capacitance at 0 V,
V = applied voltage
ϕ = contact potential
γ = a constant

γ is dependent on the doping of the device, but is usually equal to approximately 0.5 for an abrupt-junction diode. For $\gamma = 2$, which is approximately the case for a hyperabrupt-junction varactor, the tuning curve is nearly linear.

The abrupt-tuning diode will provide a very high Q with a continuous monotonic tuning curve, and will also operate over a very large range of tuning voltages (0 to 50 V). Because of its high Q, the abrupt diode offers the best available phase-noise performance. Both silicon and GaAs abrupt diodes are available, and both are used in VCOs.

Table I

Relative VCO Performance vs. Type of Varactor Diode

Diode	Linearity	Tuning Voltage	Harmonics	Residual FM	Phase Noise	Temperature Stability	Settling Time	PTD
Si-Abrupt	Fair	0 to 60	Good	Very Good	Very Good	Very Good	Excellent	Excellent
Si-Hyperabrupt	Good	0 to 20	Good	Good	Good	Good	Excellent	Excellent
GaAs-Abrupt	Fair	0 to 50	Good	Good	Good	Excellent	Good	Good
GaAs-Hyperabrupt	Good	0 to 20	Good	Fair	Fair	Fair	Fair	Fair

The hyperabrupt diode will provide a much more linear tuning response than the abrupt diode due to its linear tuning response than the abrupt diode due to its linear voltage-vs.-capacitance characteristics. This also enables it to cover a wider frequency range in a smaller tuning voltage (0 to 20 V). Its drawback is a much lower Q than the abrupt diode. This results in a phase noise typically about 3 dB higher than for an abrupt diode.

The trade-offs become obvious by examining the performance of two fundamental-output voltage-tuned oscillators covering 900 MHz to 1600 MHz (Figs. 12 to 14). The VTO-8090 employs an abrupt-tuning varactor while the VTO-9090 uses a hyperabrupt-tuning varactor.

The superior linearity of the VTO-9090 can be seen by comparing the two tuning-voltage characteristics (Fig. 12) and their modulation-sensitivity curves (Fig. 13). Note the power output difference of the two types of oscillators (Fig. 14). The abrupt-tuned VCO has higher power than its hyperabrupt-tuned counterpart. This is due to the higher Q of the abrupt varactor.

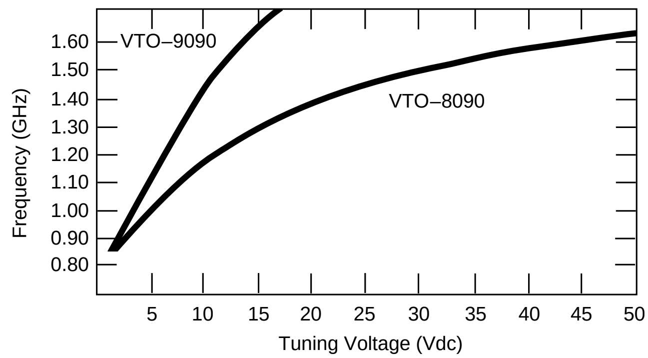


Figure 12. Tuning voltage vs. frequency for similar non-buffered VCO modules. One uses an abrupt-tuning varactor, the other a hyperabrupt-tuning varactor. The hyperabrupt varactor provides a significant improvement in tuning linearity with a significantly lower tuning-voltage range.

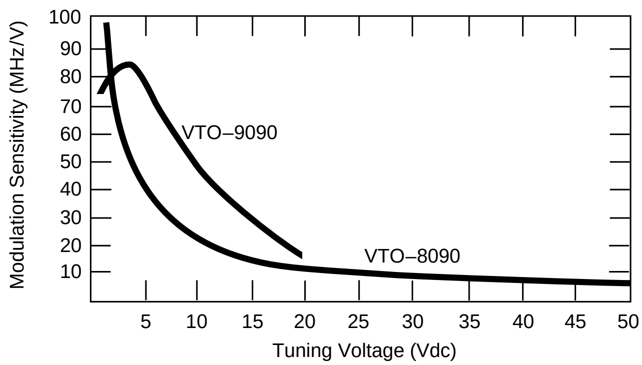


Figure 13. The superior linearity of the VTO-9090 can be seen by comparing the modulation-sensitivity curves displayed here, with the two tuning-voltage characteristics shown in Figure 12.

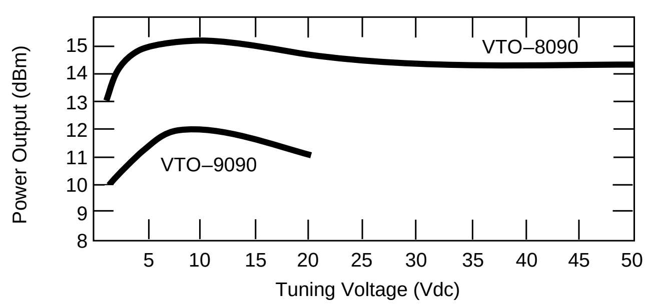


Figure 14. The disadvantage of the more-linear-tuning hyperabrupt varactor is that its lower Q tends to reduce the output power available from an oscillator.

A typical VCO tuning curve is shown in Figure 15. Where a straight voltage-vs.-frequency curve is necessary (e.g., in an open-loop system), a linearizer may also be needed.

Using a VCO linearizer can provide a very accurate tuning curve but creates its own set of problems. In addition to the tuning-curve correction, a linearizer can provide

1. a low tuning-voltage range, typically 0 to 10 Volts;
2. a constant tuning-port input impedance;
3. simple intervact to a digital-to-analog (D-A) converter.

On the negative side, a linearizer

1. requires additional input power;
2. may result in a higher MHz-per-Volt modulation sensitivity (which may not be desirable);
3. will almost invariably have a lower input impedance than the oscillator;
4. will reduce the modulation bandwidth.

Tuning curve of a linearized VCO will fall within a window about a straight line. Percent linearity is the termused to define the actual linearity achieved vs. the "best-fit straight line." This is graphically shown in Figure 16.

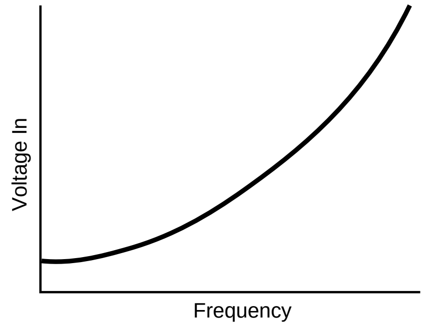


Figure 15. Typical varactor-tuned-oscillator voltage-vs.-frequency curve. Note that, although nonlinear, the curve is monotonic—an increase in tuning voltage always results in an increase in frequency. This curve can be linearized with additional circuitry at the expense of new problems.

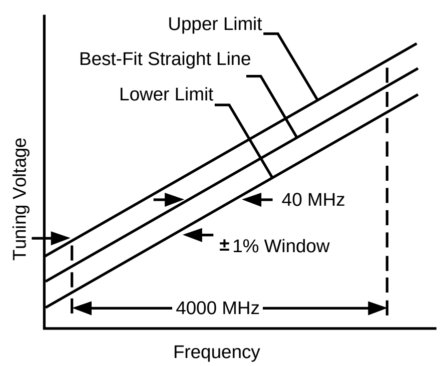


Figure 16. The tuning curve of a linearized VCO will fall within a window about a straight line. "Percent linearity" is the term used to define the actual linearity achieved vs. the "best-fit straight line." Within the window there will be typically some irregular variations in slope due to the breakpoints in the linearizer.

Calculating percent linearity is done by taking one-half the measured window, dividing this figure by the total tuning range, and expressing the result as a percentage. Using the values given in Figure 16, percent linearity is expressed as:

$$\text{Percent} = \pm \frac{80/2}{4000} \times 100 = 1\% \qquad (9)$$

Within the percentage window specified there will be typically some irregular variations in slope due to the breakpoints in the linearizer. The result is that the amount of frequency change per tuning-voltage change will vary from point to point along the curve. If modulation is applied to the tuning input, varying FM deviation would result over the tuning range. A slope-ratio limit should be specified where constant FM deviation is required.

A properly designed and "tweaked" linearizer can provide virtually any degree of linearity required for a particular application. Linearizer circuits may also incorporate theadditional function of shifting the tuning voltage provided by the system to one more appropriate for the oscillator itself.

Typically two types of linearization schemes are employed today: analog and digital. The use of an analog linearizer is desirable when the oscillator interfaces with an analog tuning voltage or when a linear modulation spectrum is desired at any point in the frequency range. A simple analog-linearizer circuit is shown in Figure 17. The primary use for a digital linearizer is an applications where the oscillator is to be tuned by a digital computer.

Referring to the analog-linearizer circuit shown in Figure 17:

$$V_o = V_i + (I_2R_F) + (I_1R_F)$$

$$I_2 = \frac{V}{R_A}$$

$$I_B = \frac{V_B - V_D - V_{TH}}{R_{TH}} \qquad (10)$$

$$I_1 = I_B + I_2$$

$$\text{Also } I_B = 0 \text{ if } (V_D + V_{TH}) \geq V_B \qquad (11)$$

$$\text{Therefore, } V_0 = V_i\left(1 + \frac{2R_F}{R_A}\right) + I_BR_F$$

From this it is easily seen that V_{TH} will determine where the increase in slope will occur and R_{TH} will determine the amount of the increase in the tuning slope.

To replace R_{TH} and V_{TH} with a simple resistive divider, Thevenin's theorem is used. By this method, the circuit shown in Figure 18 may be replaced by the circuit shown in Figure 19 where

$$V = \frac{V_S}{V_{TH}}$$

$$R_1 = \frac{R_{TH}}{V} \qquad (12)$$

$$R_2 = \frac{R_{TH}}{(V-1)}$$

Using this type of circuit gives the designer the capability of introducing almost any number of changes to the slope of the tuning curve, which may all be implemented in parallel, depending upon the degree of linearity required. This circuit will also provide good modulation response, which will only be restricted by the frequency response of the op amp itself.

One of the most efficient linearization techniques combines an analog-to-digital converter with a PROM and an op amp (Figure 20). Using this configuration and a small computer, the PROM may be programmed to provide linearity better than 0.5 percent across the full frequency spectrum of the VCO. The circuit will also provide extremely fast tuning-response time, primarily limited by the settling time of the op amp.

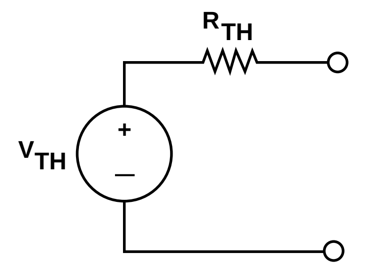


Figure 18. Simple analog-linearizer circuit. This equivalent model of V_{TH} and R_{TH} is derived from Figure 17.

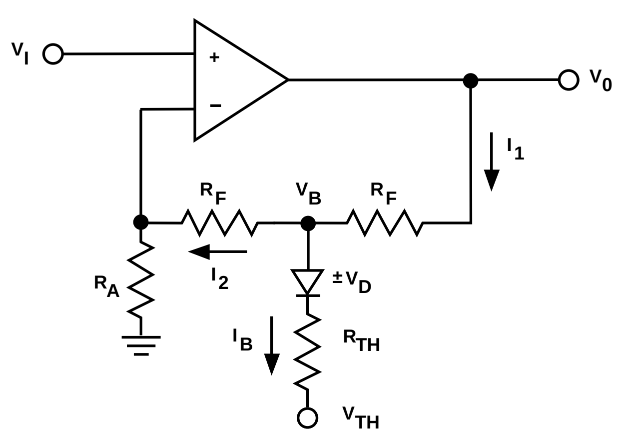


Figure 17. In this typical analog-linearizer circuit, V_{TH} will determine where an increase in the voltage-vs.-frequency slope will occur and R_{TH} will determine the amount of the increase in the slope that takes place.

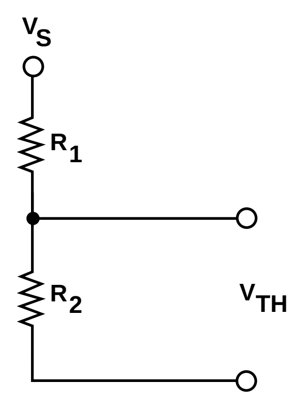


Figure 19. To replace R_{TH} and V_{TH} (Figs. 17 and 18) with a simple resistive divider, it is possible to calculate the required values using Thevenin's theorem. This provides the capability of introducing almost any number of changes to the slope of the tuning curve.

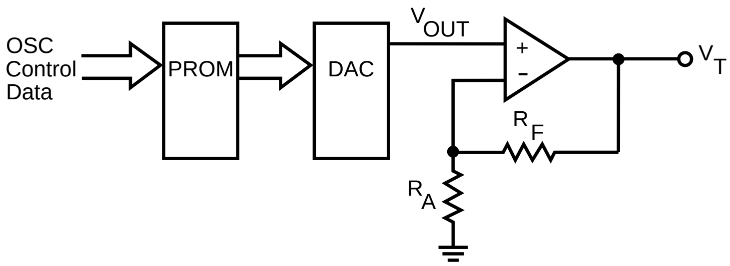


Figure 20. One of the most efficient linearization techniques combines an analog-to-digital converter with a PROM and an op amp. The PROM may be programmed to provide linearity better than 0.5% across the full frequency spectrum of the VCO and will also provide extremely fast tuning-response time, primarily limited by the settling time of the op amp.

Temperature Compensation and Stabilization

A reduction of VCO frequency variations with changes in temperature may be carried out using one or more of these three basic techniques: control of the oscillator temperature, tuning-voltage temperature compensation, or the use of a phase-locked loop.

The temperature of a small component such as a TO-packaged VCO is easily controlled either by a very small, low-power heater or by placing the component in a temperature-controlled chamber (oven). DC proportionally controlled heater assemblies specifically designed for use on TO-8-type cans are available commercially. It is also relatively simple to fabricate a heater by mounting the osciallator on a block (which provides thermal mass) that is temperature-controlled using a proportional heater (Figure 21).

A self-controlling heater that employs a material with a definitive temperature-vs.-resistance characteristic may be used. This material may be epoxied directly to the top of the TO-8 oscillator and then supplied with a bias voltage. The temperature-vs.-resistance characteristic of the material will make it act as a temperature-controlled heater that will provide very good temperature stability at a very low cost.

When any heater approach is used for temperature compensation, the temperature of the oscillator must be kept 5 to 10 degrees above the maximum expected system operating temperature. This will ensure that the oscillator will not be affected by the external temperature changes. The primary drawback to using the heater is the extra power required to keep the oscillator at a higher-than-ambient temperature.

The effect of temperature on the oscillator frequency may also be reduced indirectly by varying the tuning voltage in the proper direction to bring the oscillator back to the correct frequency. Temperature compensation of the tuning network requires using a negative-temperature-coefficient (NTC) or positive-

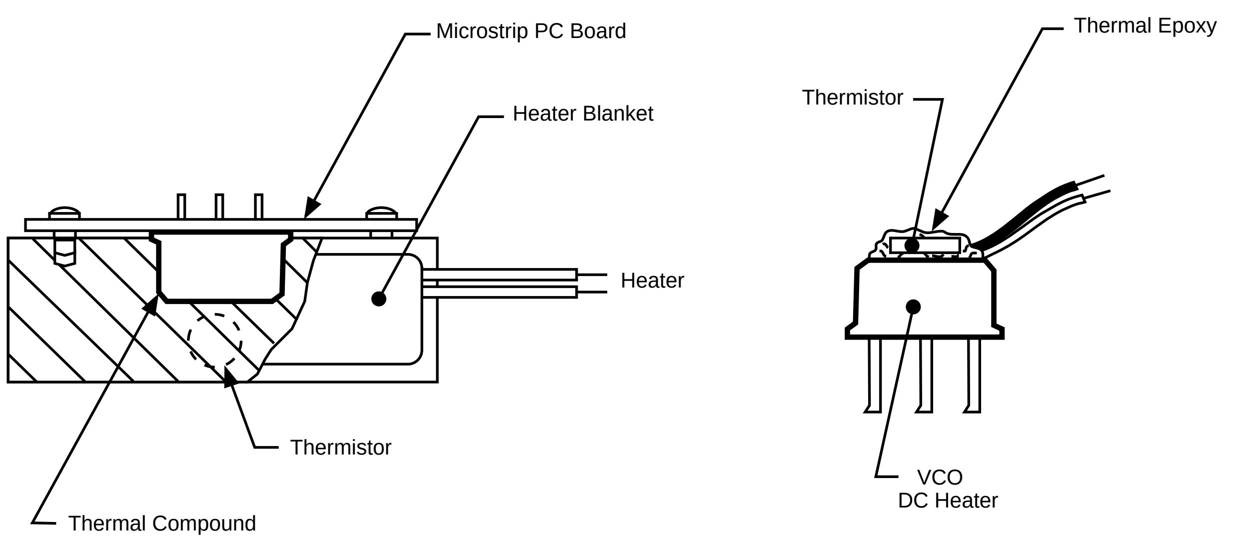


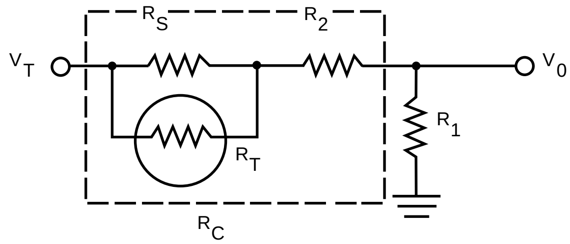
Figure 21. The temperature of a small VCO may be stabilized by mounting the oscillator on a block (which provides thermal mass) that is temperature-controlled by using a proportional heater, or with a small heater directly epoxied to the top of the TO-8 oscillator.

temperature-coefficient (PTC) thermistor or a network of thermistors, depending on the actual tuning circuitry used.

For example, VTO-8000 series oscillators will display typically a negative frequency-vs.-temperature drift coefficient. To compensate for this, a voltage-compensation network may be used. Two simple networks are shown in Figures 22 and 23.

Other types of resistance networks may be used in place of Rc. A suggestion for determining the resistor values for an effective compensation network is to hold the values of R1 fixed, and use a curve-fitting routine to determine the values of R_2 and R_S when the desired value of R_C is known for at least three different temperatures.

When temperature compensation of the tuning voltage is used, the temperature-sensing device should be mounted as close as possible to the oscillator itself. This will provide the shortest thermal time constant possible from the sensing device to the compensated oscillator.



PTC Network

$$V_0 = (V_T R_1) / [R_1 + R_2 + (R_S \,||\, R_T)]$$

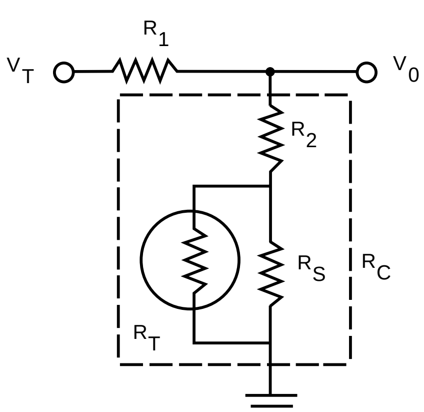
Figure 23. A VCO tuning-voltage-compensation network using a positive-temperature-coefficient (PTC) thermistor. Details of these terms are provided in Figure 22.

Phase-locked loops using VCOs are becoming much more common due to improvements in, and the greater availability of, divider techniques and SAW or crystal multiplied sources. Some of the more important requirements for an oscillator to be suitable for a phase-locked application are:

1. phase stability (spectral purity),
2. large electrical tuning range,
3. linearity of frequency vs. control voltage, and
4. (frequently) the capability of accepting wideband modulation.

From the information supplied so far it is clear that, if a very-low-noise oscillator is required, the best performance will be obtained when the bandwidth is kept as low as possible (<20%) and the tuning voltage as high as possible (>10 V).

The second network is that using an NTC thermistor.



$$V_0 = V_t\ (R_2 + R_s \,||\, R_T) / [R_1 + R_2 + (R_s \,||\, R_T)]$$

$$R_T = R_{25}\ (1 + A)\ \ (T - 25)$$

Where:
T = Temperature in °C
A = Temperature coefficient of thermistor % / °C @ 25°C
R_{25} = Thermistor resistance at 25°C

NTC Network

Figure 22. The effect of temperature on the frequency of a VCO may be reduced indirectly by varying the tuning voltage in the proper direction to bring the oscillator bakck to the correct frequency. This circuit, using a negative-temperature-coefficient (NTC) thermistor, is appropriate for use with oscillators that typically display a negative frequency-vs-temperature drift coefficient. Note that the current through the thermistor should be held to less than approximately 1 mA (depending on the thermistor's mass) to prevent self-heating. Figure 23 shows a similar circuit using a positive-temperature-coefficient (PTC) thermistor.

VCO Assemblies

Most VCO assemblies (such as the one shown in Figure 24) integrate the voltage-controlled oscillator with amplifiers, voltage regulators, linearizer, and heater circuitry. In some cases, multiple VCOs are used; this requires additional amplifiers and the incorporation of switch combiner networks. Typical specifications are shown in Table II.

To design a circuit with a very high Q, the tuning bandwidth must invariably suffer.

A complete VCO assembly is usually significantly smaller than an arrangement of separately packaged components. The thermal design is better, since it is done for the complete assembly. The integrated assembly minimizes the number of interfaces and connections; this helps improve reliability. It can also offer premium performance due to the complete control of all critical components, and of integrating and matching functions.

VCO assemblies tend to be custom designs for particular applications. The basis performance trade-offs for particular VCO types should also be considered when specifying VCO assemblies.

Table II

Typical Specifications for Integrated VCO Assemblies

Frequency range	1-2 GHz	8-12 GHz	12-18 GHz
Power output, min.	+15 dBm	+14 dBm	+14 dBm
Power flatness, max.	5 dB	3 dB	3 dB
Harmonics, min.	←	20 dBc	→
Spurious, min.	←	60 dBc	→
Residual FM (@ 3 dBc)	50 kHz	100 kHz	100 kHz
Load VSWR	←	1.5:1	→
Linearity	←	±2%	→
Mod. sense ratio	1.8:1	2.0:1	2.0:1
Operating temperature	←	–54° to +71°C	→
Tuning voltage	←	0 to +10 Vdc	→
Input power	←	+20 Vdc @ 600 mA	→
	←	20 Vdc @ 250 mA	→
	←	+28 Vdc @ 4.0 A (Heater)	→

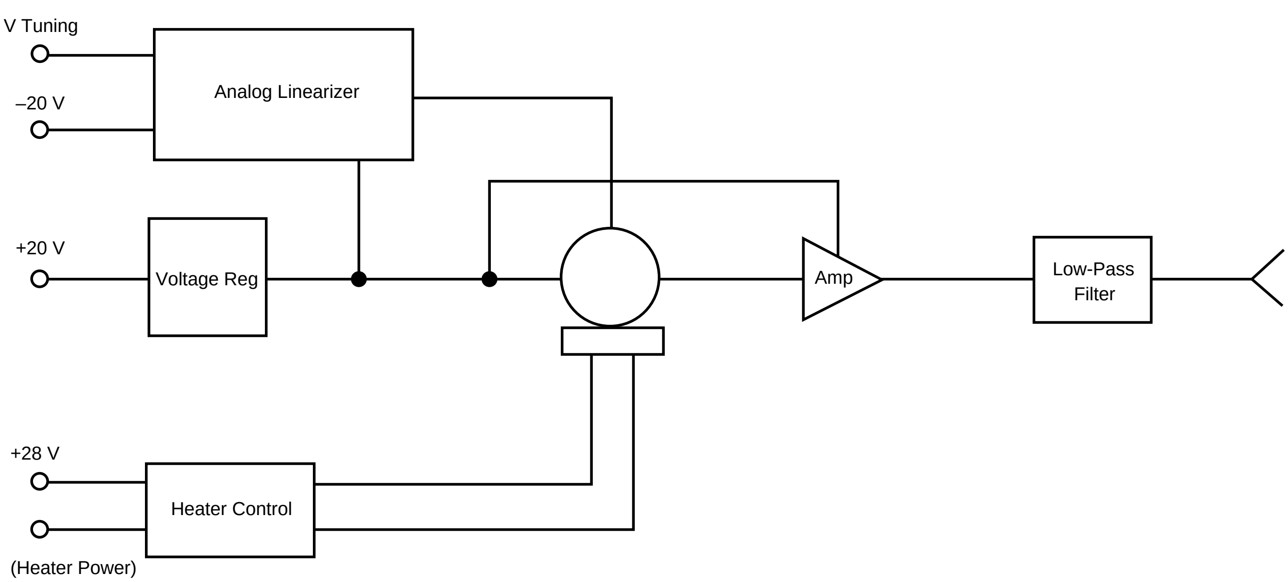


Figure 24. A typical VCO assembly integrates a voltage-controlled oscillator with amplifiers, voltage regulators, linearizer, and heater circuitry. Typical VCO assembly specifications are given in Table II.

HEWLETT® PACKARD

One Family of "Raw" VCOs

To examine some of the characteristics of off-the-shelf VCOs, it is useful to look at the product line. The VTO-8000 series commercial VCOs combine a silicon transistor chip with a silicon abrupt varactor diode. Maximum tuning voltages are between 40 and 50 Volts, which typically is acceptable for most commercial applications. This series of VCOs, mated with a user-supplied low-impedance driver, exhibits tuning speeds on the order of less than 1 ms.across the full band. The operating temperature range for these products is 0° to +65°C, and they are packaged in compact, lightweight TO-8 cans.

The VTO-9000 series oscillators use a silicon transistor chip and silicon hyperabrupt varactor diode. This produces more linear tuning curves than their abrupt-tuned VTO-8000 counterparts. Tuning voltages required are less than 25 Volts, making them very compatible with digital systems. Nanosecond running speeds are achievable with a low-impedance driver due simply to the lower tuning-voltage sqing required. The operating temperature range for these products is also 0° to +65°C.

Militarized MTO-8000 products exhibit the same basic performance characteristics as the VTO-8000 series. The major difference is that they have been designed and tested to meet performance specifications over the military temperature range of –54°C to +85°C.

HTO series militarized hyperabrupt VCOs are designed specifically for octave band coverage. They use silicon transistors and silicon hyperabrupt diodes for frequencies from 900 MHz to 2 GHz and use GaAs transistors from 2 to 18 GHz. Tuning speeds are on the order of 3 ms. The VCOs have been designed and tested for specification compliance from -54°C to +85°C, and are packaged in either hermetically sealed TO-8 cans or the Avanpak miniature microwave flatpak with field-replaceable coaxial connectors.

VTO-series buffered VCOs are designed specifically for good phase-noise and frequency settling-time characteristics. They can be tuned full band typically in less than 3 ms. while settling to within 1 MHz. Internal buffer amplifiers provide very good isolation from variations in load impedance while also minimizing frequency pulling. A customer-supplied heater is required to maintain the oscillator temperature at 80° (±5°C) for specified performance.

Table III

Voltage-Controlled Oscillators

Product Series	Tuning Voltage	Phase Noise	Bandwidth	Linearity BSFL	Harmonics	Case	Post-Tuning Drift
VTO-8000	0 to +60	Good	75% max.	Fair	–15 dBc	TO-8V	Good
VTO-9000	0 to +20	Good	75% max.	Excellent	–14 dB	TO-8V	Very good
MTO-8000	0 to +60	Good	75% max.	Fair	–10 dBc	TO-8V	Good
HTO-	0 to +20	Fair	Octave	Excellent	–12 dBc	TO-8V*	Fair
LNO-	0 to –20	Excellent	30% max	Good	–12 dBc	TO-8V	Excellent

(*Select models available in Avanpak miniature flatpack)

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