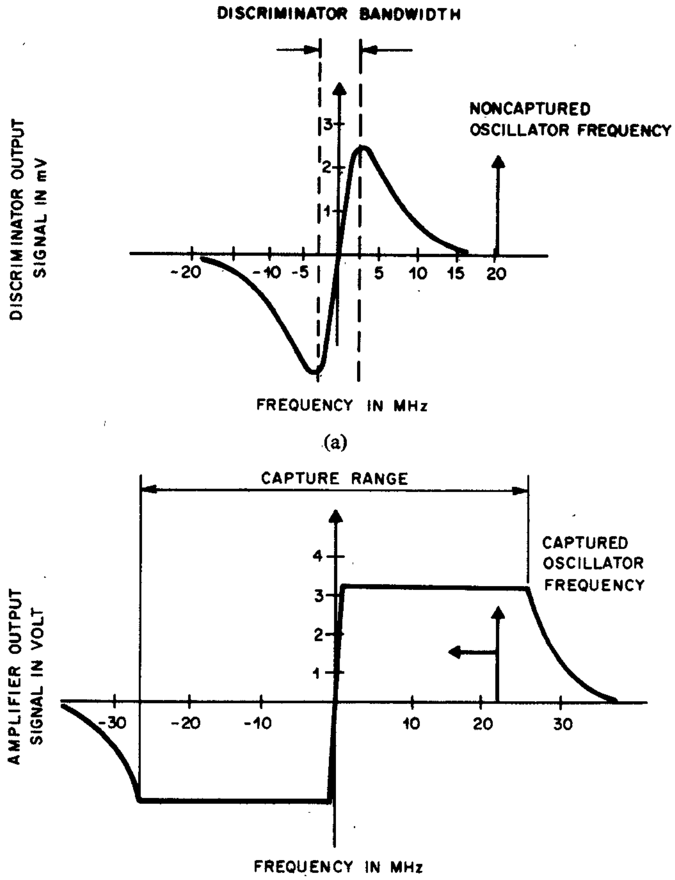

Fig. 5. (a) Discriminator output voltage. (b) Discriminator output voltage after amplification by the saturated IC amplifier.

inator, an IC amplifier, and a varactor, can be integrated with the oscillator circuit on a single dielectric substrate resulting in a compact device. This device can be mass-produced with high reliability at relatively low cost by using thin-film techniques and can be readily scaled to high frequencies. The stabilization circuit also provides self-capture of the oscillator frequency over a large frequency range.

References

[1] W. W. Snell, Jr., Patent 3 796 970.
[2] B. Glance and W. W. Snell, Jr., "Low noise integrated millimeter-wave receiver," in *Proceeding of the 1973 European Microwave Conference*, Brussels, Sept. 1973.
[3] J. R. Ashley, "Reduction of FM noise in microwave diode oscillators by cavity and injection stabilization," in *Proceeding of the 1971 International Microwave Symposium*, pp. 94 and 95.

Posttuning Drift of a Transferred-Electron-Device Voltage-Controlled Oscillator

KENNETH M. JOHNSON, MEMBER, IEEE

***Abstract*—An experimental study was done on the parameters affecting the frequency drift of a transferred-electron-device (TED) voltage-controlled oscillator (VCO) after tuning from one frequency to another within its tuning band. For the VCO measured, reduction in frequency drift was observed and measured when: 1) the output power was reduced either by decoupling the load or by using a lower power TED; 2) the voltage swing was restricted so as to draw less than 10 μA forward or reverse current; and 3) when a varactor with a high Q was used. With 5-mW output the TED VCO had a frequency drift less than 2.5 MHz from 1 μs to 100 ms when step tuning the frequency anywhere in the frequency band from 6.8 to 9.1 GHz.**

Manuscript received April 21, 1975; revised April 5, 1976.
The author is with the Government Electronics Division, Motorola Inc., Scottsdale, AZ 85257.

Introduction

For many systems it is becoming increasingly important to control accurately the frequency of a voltage-controlled oscillator (VCO) after a step change in tuning voltage has been applied to its tuning port. This is particularly true with the use of digital signals to control the frequency of the VCO. After a step change in tuning voltage the frequency does not immediately reach its "steady state" or final value, but there is some frequency drift, called the posttuning drift (PTD) which may continue for many hours. Often the PTD is distinguished from the settling time of the VCO; the settling time being the time required for the VCO to reach within a fixed percentage error of some fixed-frequency value after the application of a step change in tuning voltage. The settling time usually has reference to time periods shorter than 10 μs, whereas the PTD is applied to time periods after 1 μs. Usually, PTD is separated into long- or short-term drift depending on the time from which the drift is measured. For long-term drift the time is typically 100 ms.

This short paper will concern itself with short-term PTD as measured from 1 μs to 100 ms. The reason for choosing 100 ms is that by this time most thermal effects associated with the active devices have reached a very small value. The circuit studied is a VCO using a transferred-electron device (TED) with a single varactor as the tuning element. There are many factors affecting PTD arising from the driver amplifier, varactor diode, the active TED device, and circuit characteristics including harmonic termination. These effects are quite difficult to separate except those involving the driver amplifier. The approach in this short paper is to first discuss the measurement technique used. Then the experimental measurements will be described which were performed to study the effects of various parameters on the VCO PTD. Finally, some general comments will be made suggesting the causes of the various PTD drifts observed.

Measurement Technique

There are several possible methods for measuring the settling time of a microwave VCO. These include: 1) use of a stable oscillator with a precisely known frequency whose output is mixed with the VCO output to get a beat note that may be observed on a scope; 2) use of a microwave discriminator to convert the frequency to a voltage which again may be measured on an oscilloscope; and 3) use of a spectrum analyzer to observe the shift in spectral line as pulsewidth is increased.

These techniques have various advantages and disadvantages which will not be described here. The technique used in this study is the spectrum-analyzer approach, which possesses the advantage of being able to "see" the entire drift across a pulse that appears as a broadening of the spectral line. Fig. 1 shows a block diagram for this approach.

The pulse is passed through a diode-clamp circuit in order to obtain an accurate drive voltage for the VCO. Any droop or overshoot in the driver pulse will result in a frequency error in the VCO output. Two typical diode-clamp circuits are shown in Fig. 2. Combinations of series and shunt diodes may be used to more precisely fix the frequency. The output from the diode clamp passes into the VCO driver amplifier to develop the necessary 40 V to tune the VCO through its entire range. Usually, some

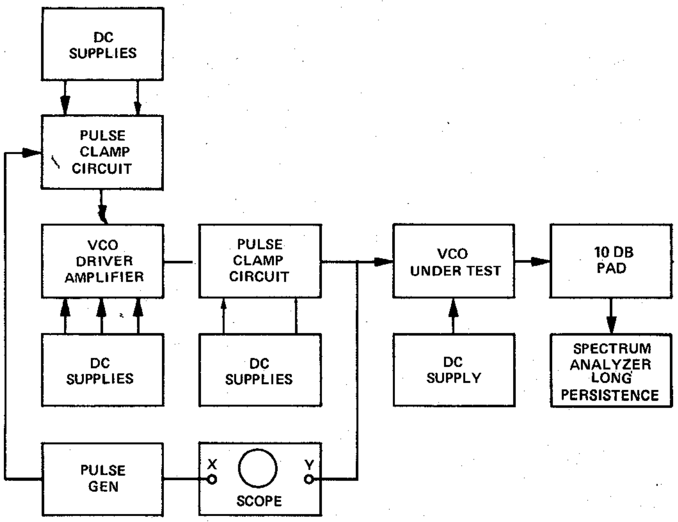

Fig. 1. VCO settling-time measurement setup.

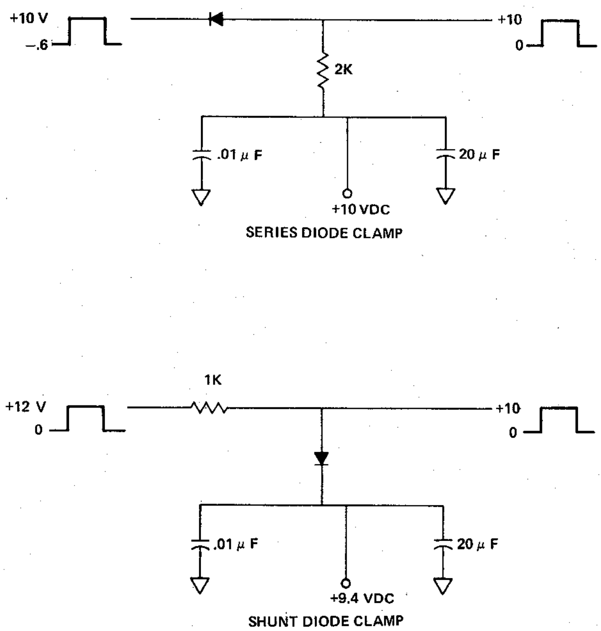

Fig. 2. Pulse-diode-clamp circuits.

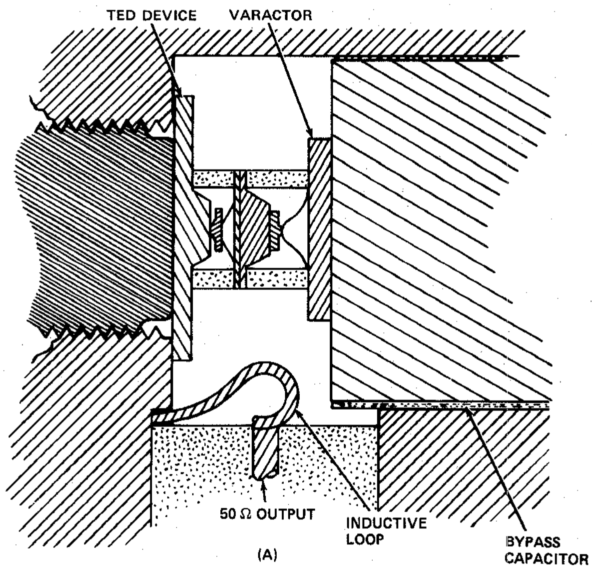

Fig. 3. TED VCO coaxial circuit with loop coupling for output.

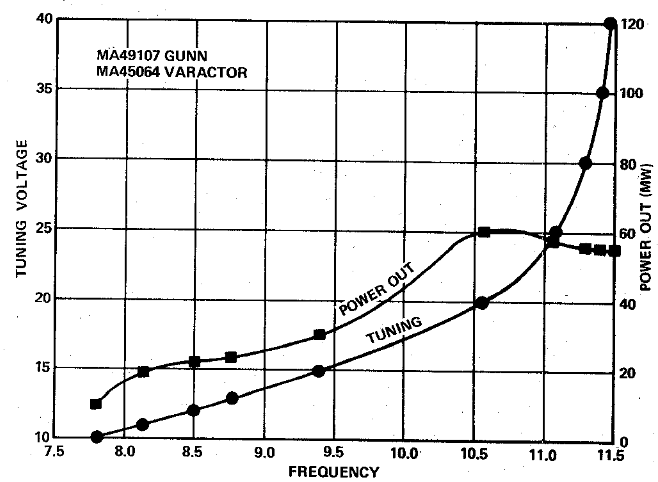

Fig. 4. TED VCO tuning voltage and power output vs. frequency.

type of operational amplifier is used. This amplifier must have a settling time of less than 1 μs with a voltage accuracy of 0.01 percent or better if measurements are to be made at 1 μs. In addition, it may be desired that this amplifier have output-pulse rise and fall times of 100 ns, or better in order to drive a second clamp circuit between the driver amplifier and VCO. This clamp circuit may be used, if desired, to effectively eliminate any drift that might be caused by the driver amplifier. The pulse voltage now available to the VCO is constant within 0.01 percent of the voltage step. The VCO output is fed into a spectrum analyzer, and the spectrum observed at either end of the tuning range where the pulse is clamped. As the pulsewidth (usually a square wave) driving the VCO is widened from 1 μs, for example, the spectrum will widen corresponding to a shift in frequency. This technique has been compared with other approaches with identical results. Notice that a scope is also used to monitor pulsewidth when making the measurements.

Having defined the measurement technique a brief description of the VCO used in the tests is given next.

Transferred-Electron-Device Voltage-Controlled Oscillator

The design of the TED or Gunn VCO has been discussed by a number of authors [1], [2] and will not be repeated here. The approach in this study uses a coaxial circuit in which the TED and varactor are in series with inductive loop coupling provided for output-power coupling. This is shown in Fig. 3. Although not shown, bias was provided through a "filtercon" and a fine No. 34 wire, $\lambda/4$ long, to the point where the active devices meet. Using this circuit, the tuning and power output characteristics shown in Fig. 4 were obtained. Notice that a bandwidth of nearly 4 GHz is achieved with a minimum output power of 10 mW. The VCO was quite stable with no observed spectral breakup of "jumps" when tuning. The PTD measurements may now be described.

Posttuning-Drift Measurements

It was observed when testing the VCO that forward rectified current greater than 1 mA was measured through the varactor when the VCO was tuned to 7.8 GHz, and it did not drop to less than 100 μA until the frequency reached 8.5 GHz. This current is due to peak RF swing entering forward conduction in the

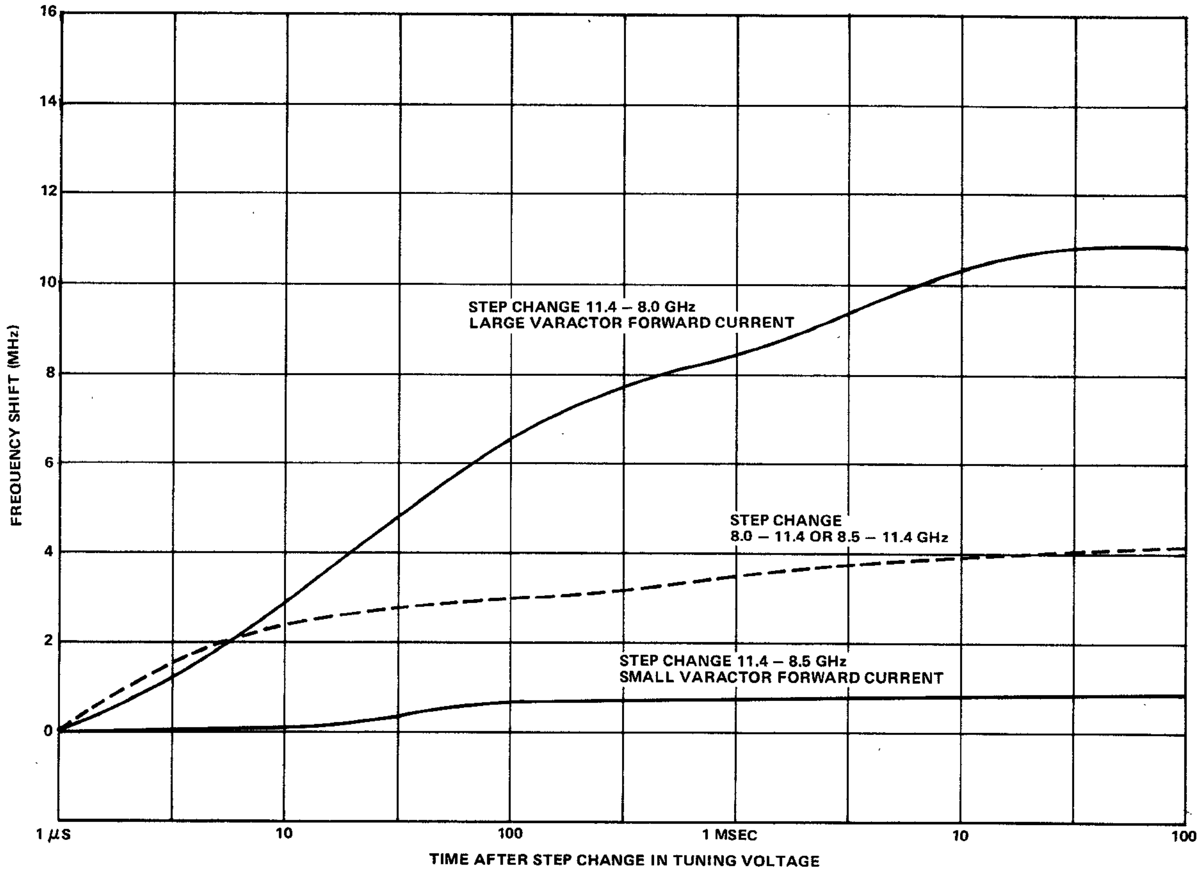

Fig. 5. Frequency deviation of TED VCO from 1-μs value after step change in tuning voltage as a function of time for large and small varactor forward current.

varactor and producing dc current. Likewise, reverse current in excess of 100 μA was measured at about 10.6 GHz. Consequently, measurements were made to determine the effect of forward and reverse diode current on settling time. The VCO was set up as previously described and measurements made from 1 μs to 100 ms in two different frequency steps: one down to a minimum frequency of 8 GHz, and one down to 8.5 GHz from the maximum frequency of 11.4 GHz. Measurements were then made from the two lower frequencies of 8 and 8.5 GHz up to the upper frequency at 11.4 GHz. The results are shown in Fig. 5. A reduction in frequency drift of nearly an order of magnitude was measured when no forward current was drawn over that obtained when switching from maximum to minimum frequency. However, negligible change was seen when switching from minimum to maximum frequency. This clearly indicates that for low PTD the tuning band must be restricted to prevent forward current from being drawn. However, it was also found in some cases that an improvement in PTD could be measured when the VCO was driven hard into forward conduction with greater than 10-mA forward current. The effect was that the varactor itself behaved as a sort of diode clamp to fix the frequency. Consequently, care must be taken in making PTD measurements to be sure that the varactor is not driven hard into forward conduction to avoid misleading results.

Similar tests were conducted to determine the effects of reverse current. The VCO for these tests tuned from 6.8 to 9.7 GHz. Reverse current in excess of 100 μA was measured at 9.1 GHz. The VCO was step tuned from 6.8 to 9.7 GHz, and conversely, and likewise from 6.8 to 9.05 GHz. Results are shown in Fig. 6. A substantial reduction in PTD took place when no reverse current was drawn.

The third parameter tested was the effect of diode Q. For this purpose two varactors were used. One had a Q of 2400 at 50 MHz, and the second a Q in excess of 4000. Results are shown in Fig. 7. The high-Q varactor gave a much lower PTD than the low-Q varactor. Since some forward and reverse current was being drawn at both ends of the step, the VCO was stepped over shorter steps to be sure that reverse- and forward-current effects were not entirely responsible for the improved PTD. The results were consistently the same: the higher the varactor Q the lower the PTD. Improvement could, in part, be explained by the fact that the low-Q varactor dissipated nearly twice the power that the high-Q varactor dissipated. Consequently, if a lower power TED were used, the PTD should improve substantially.

A comparison was made of the PTD of a VCO using a low-power TED (50-mW oscillator power output) with a high-power TED (250-mW output). Results are shown in Fig. 8. As expected, the PTD for a low-power TED is substantially better than for a high-power TED. The low power had slightly worse PTD below 30 μs. This may be attributed to effects in the TED rather than in the varactor, which will be discussed later. The final tests to be described are those made on output-power coupling. In these tests the load was decoupled and hence the output power was reduced. This has the effect of increasing the circuit loaded Q without decreasing the power dissipated in the varactor. In fact, the power dissipated in the varactor increased slightly. The fact that the RF power in the varactor increases when the load is decoupled was substantiated by the following test. A TED VCO was set up using a GaAs tuning varactor and with the rectified dc current in the varactor diode measured as the load was decoupled by removing it physically from the TED cavity. These tests were done with the varactor biased to near 0 V, so that the

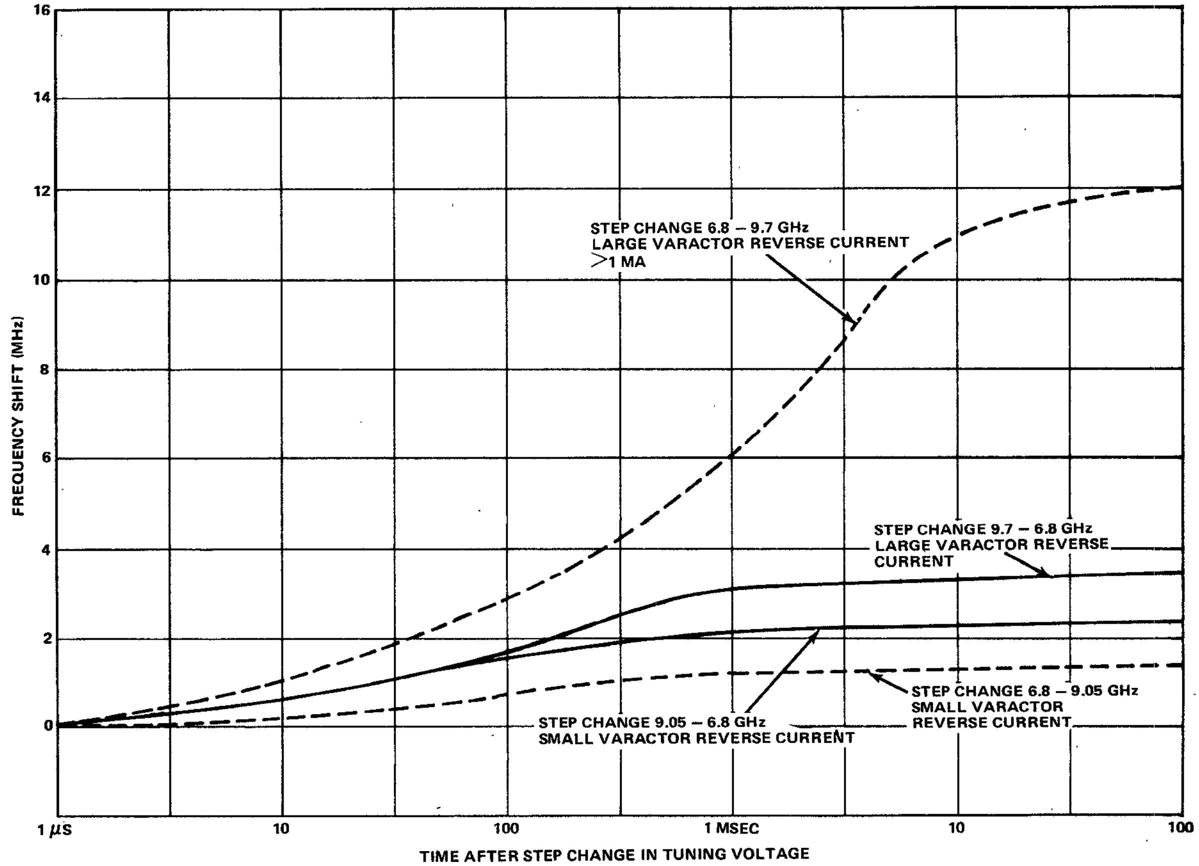

Fig. 6. Frequency deviation of a TED VCO from 1-μs value after step change in tuning voltage as a function of time for large and small varactor reverse current.

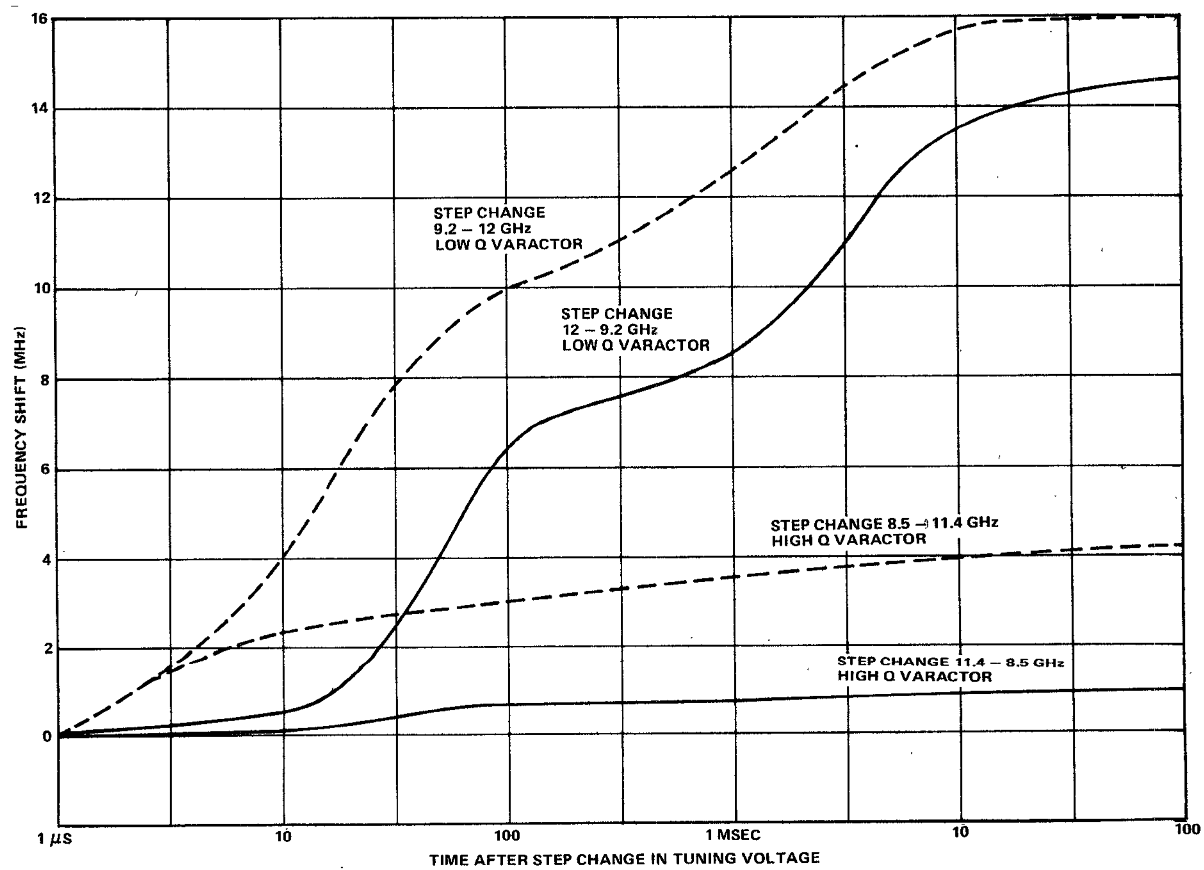

Fig. 7. Frequency deviation of a TED VCO from 1-μs value after step change in tuning voltage as a function of time for low- and high-Q tuning varactors.

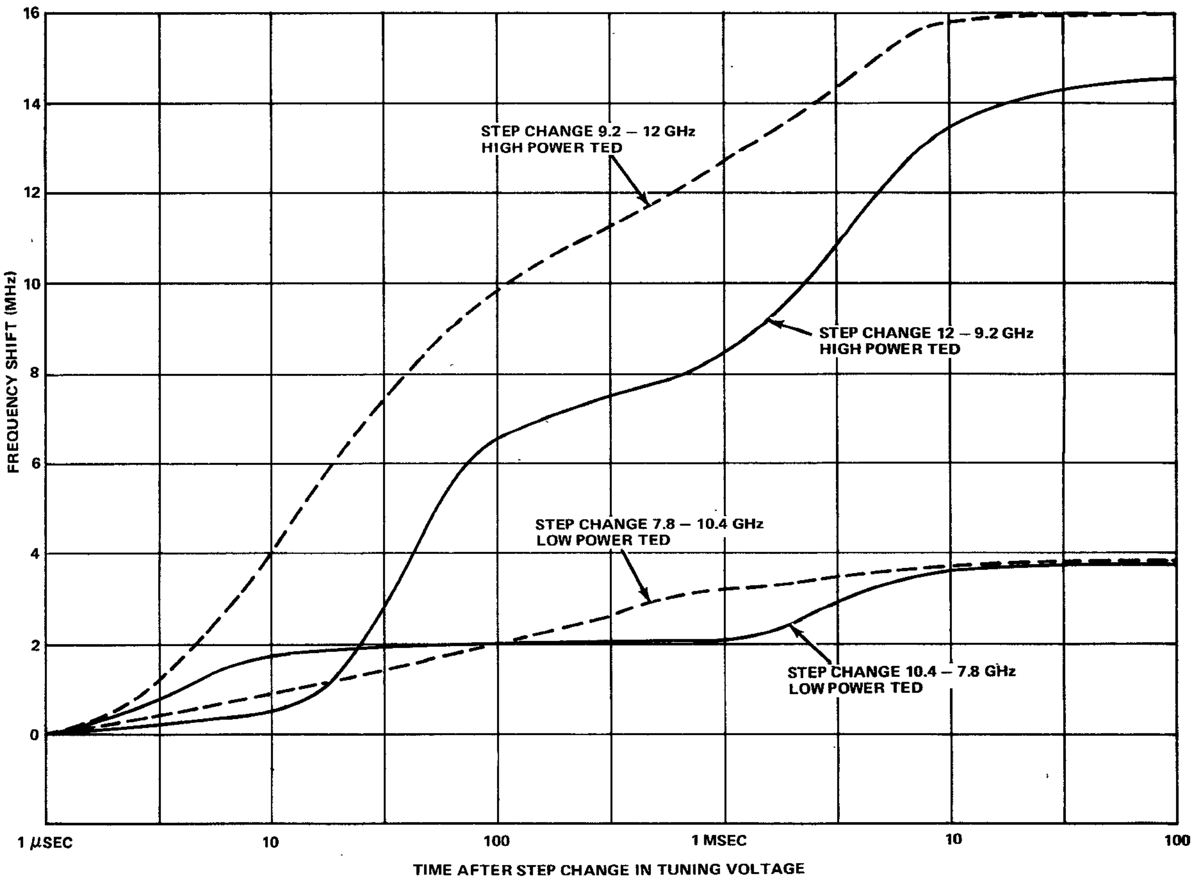

Fig. 8. Frequency deviation of a TED VCO from 1-μs value after step change in tuning voltage as a function of time for high- and low-power TED's.

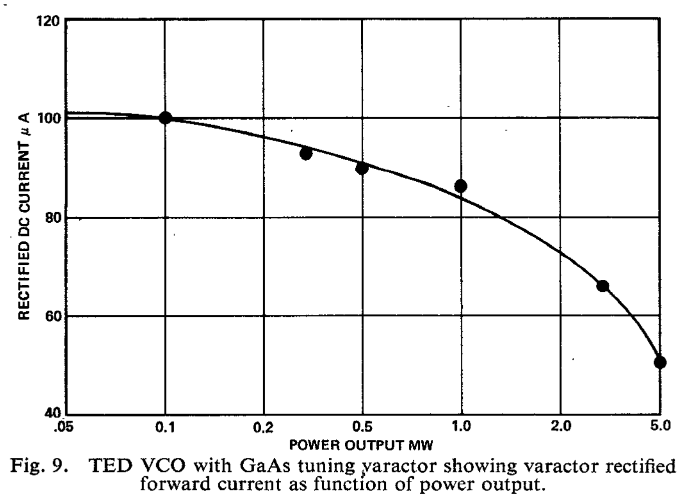

Fig. 9. TED VCO with GaAs tuning varactor showing varactor rectified forward current as function of power output.

50 μA of forward current was drawn with no output power. The results are shown in Fig. 9. Here it is seen that as the load is decoupled so that the output power drops from 5 to 0.05 mW, the dc rectified current increases from 50 to 100 μA. The tests on power output coupling were performed using two different types of circuits. One, which used parallel bars for output coupling, could easily be adjusted to vary the output. The second circuit type, the one shown in Fig. 3, required some disassembly and reassembly to remove the coupling loop and thereby vary the output power. (Later it was found sufficient merely to unscrew the load from the output connector to reduce power.) The behavior of both VCO types was the same. Results for the VCO of Fig. 3 are shown in Fig. 10. Decoupling the load reduced the PTD by nearly a factor of 3. This shows that circuit Q is an important parameter to consider; in this case, even masking the effect of increased varactor dissipation when the load is decoupled. In the tests of Fig. 10, the power output is too low for most applications. With the power output adjusted for a minimum of 5 mW, the PTD was a maximum of 2.5 MHz regardless of where the VCO was step tuned in the band from 6.8 to 9.1 GHz. While better results could be obtained with carefully adjusted steps, they do not accurately represent the worst case VCO behavior.

Thermal Effects

As discussed previously, the varactor may be required to dissipate a significant amount of RF power. When tuning from one end of the band to the other, this dissipated power will change considerably. There are several reasons for this. First, the conductance due to the diode's series resistance transforms to the TED terminals with considerable variation over the frequency band. Second, the series resistance of the varactor diodes near forward conduction is a function of voltage. Third, the power dissipated in the varactor is a function of the load, which may vary with frequency, and the TED negative conductance, which does vary with frequency. Fourth, if the varactor draws significant forward or reverse current, this will result in power dissipation equal to the applied voltage across the junction times the dc current being drawn through the device. The

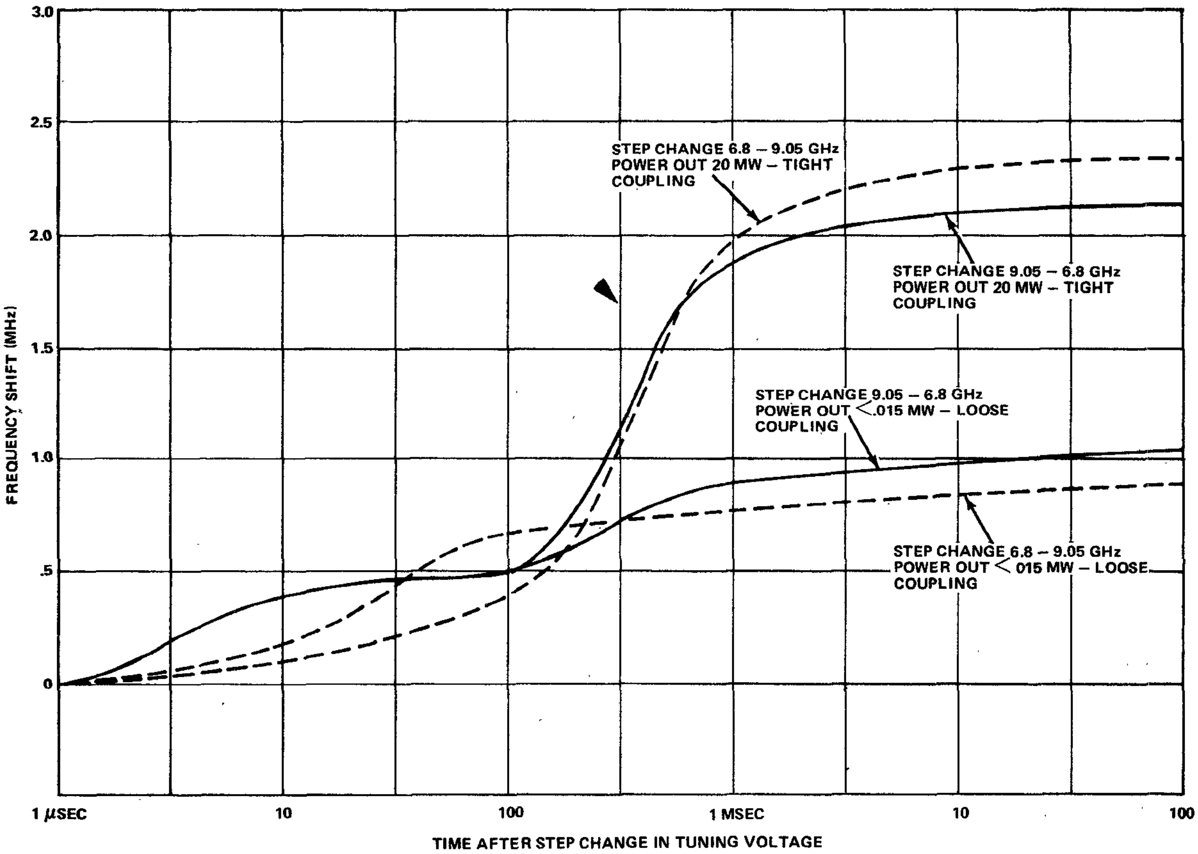

Fig. 10. Frequency deviation of a TED VCO from 1-μs value after step change in tuning voltage as a function of time for tight and loose output-power coupling.

change in power causes a temperature change at the varactor junction. This temperature change is reflected by a capacitance change due to the temperature coefficient of the capacitance of the varactor. This effect can produce changes in frequency of several megahertz.

A simplified thermal model for the varactor or TED devices may be constructed, consisting first of the heat source applied to the device junction. Heat is stored in the thermal capacitance of the device chip, then passed through the thermal resistance of the chip-to-pedestal-block interface. It is then stored in the thermal capacitance of the pedestal block, and finally passed through the thermal resistance of the pedestal-block-to-"infinite"-heat-sink interface to the infinite heat sink. The transient response of this circuit may be computed and is found to have two time constants, one associated with the device chip and the second with the pedestal block. The only significant effect is that due to the device-chip capacitance and thermal resistance.

The effect of the varactor-diode-chip temperature change in a worst case may be estimated as follows. In a worst case, the total change in dissipated power in the varactor switching anywhere in the band may be estimated to be $W_1 = 30.0$ mW. From the data on the diode chip to heat sink thermal resistance R_c, given by the manufacturer to be 130°C/W, the temperature rise in the junction is computed to be 3.84°C. The temperature coefficient of the diode capacitance varies with voltage with a zero value of $C_T = 0.115$ percent/°C. At -4 V this reduces to 0.03 percent/°C, and by -20 V it has reduced to 0.0265 percent/°C. At 0 V, this temperature rise corresponds to a frequency shift of 22 MHz, while at -20 V the frequency shift would be at most 5.1 MHz. These, of course, are worst case estimates. The actual frequency shift would be less, depending on how the varactor couples to the circuit and the amount of varactor forward or reverse current.

The change in frequency Δf with time may be computed in the usual manner as

$$\Delta f = \Delta f_{\max}(1 - \exp(-t/\tau)) \tag{1}$$

where $\tau = R_c C_c$ is the thermal time constant with C_c given as the product of the specific heat of silicon ($C_V = 0.176$ cal/g · °C) and the chip mass. In this case $C_C = 44.6 \times 10^{-6}$ W · s/°C and $\tau = 5.79$ ms. Thus the temperature effect in the varactor is expected to be significant from about 200 μs to 25 ms. This behavior shows up most clearly in Fig. 7 where the tests were made on a low-Q varactor which would show the largest frequency shift. Notice, however, that when drawing reverse current the frequency shift is much larger than can be accounted for by changes in varactor capacitance with temperature. That is, a change in power of 100 mW would be required to get the 12-MHz drift measured as shown in Fig. 6.

Three possible explanations may be offered to account for the increased PTD when drawing reverse current. First, it may be observed that when the varactor RF swings into breakdown, this has the effect of introducing a small voltage in series with the junction. This voltage is temperature dependent following the temperature coefficient of the breakdown voltage which is 0.024 V/°C. For the 3.84°C maximum rise, previously estimated, this is a voltage change of 93 mV. The tuning sensitivity of the VCO at the maximum frequency is about 50 MHz/V, which thereby gives a frequency shift of 4.65 MHz. This still seems somewhat low to account for all of the observed shift. A second possible explanation is that, as the varactor enters breakdown, an ionization-mode space charge is set up which introduces a reactance into the circuit, thereby altering frequency with time, depending on the temperature coefficient of the reactance. Third, the varactor Q may be reduced when drawing reverse current. Possibly all three are responsible. In this connection it is possible that reduction in varactor Q could be responsible for the degradation in PTD measured when the varactor draws forward current. In the forward-conduction region the varactor Q does drop significantly due to both increased capacitance and increased series resistance.

In a similar manner, the effect of changes in dissipated power in the TED may be studied. The changes in dissipated power are due to two factors. First, the negative conductance changes with frequency, and second, the dynamic operating point of the TED changes due to changes in saturation level and hence small changes in rectified dc current. The total magnitude of these power changes is uncertain, but for purposes of estimation will be taken to be on the order of 20 mW.

In this case, the thermal resistance of the mounted TED chip is much smaller since it is mounted "flip chip," i.e., the active portion of the semiconductor is mounted closest to the package pedestal. For the TED used, the chip-to-pedestal thermal resistance is $R_c = 15°C/W$ and hence the temperature rise is much smaller than the varactor; i.e., $\Delta T = 0.3°C$. However, the temperature coefficient of the frequency shift at resonance is larger. This has been measured to be $C_{TF} = 0.086$ percent/°C, corresponding to a frequency shift of 2.7 MHz at 10 GHz. To calculate the time constant, the mass times the specific heat ($C_V = 0.086$ cal/g · °C) gives the thermal capacitance $C_C = 21.8 \times 10 \times 10^{-6}$ W · s/°C; then $\tau = R_c C_c = 327$ μs. Thus the thermal effects of the TED are expected to be significant from 10 μs to 1 ms. This probably is important at time periods somewhat less than 10 μs, since the thermal resistances and capacitances are actually composites of the three regions that make up the TED chip, along with the interfaces of the metal semiconductor ohmic contacts. These regions and their interfaces have differing thermal resistances and capacitances due to their differing doping levels. This would also apply to the varactor-tuning diodes. It is seen that time periods of thermal drift for the two semiconductor devices overlap somewhat, which makes isolation of these effects difficult. Measurements on a number of VCO's indicate that these thermal effects are settled out by about 50 ms.

For the time period prior to a few microseconds, effects other than thermal may be responsible for frequency drifts. This requires further study.

Conclusion

The experimental study has shown that if the previously discussed techniques are employed, the PTD of an X-band TED VCO can be significantly reduced for the time period from 1 μs to 100 ms. To do this, the oscillator power and the output power must be kept low and the bandwidth restricted to prevent the varactor from drawing large forward or reverse current. Finally, a high-Q varactor-tuning diode must be used.

References

[1] D. Large, "Octave band varactor-tuned GUNN diode sources," *Microwave Journal*, vol. 13, pp. 49–51, Oct. 1970.
[2] K. M. Johnson, "Widebandwidth IMPATT and GUNN voltage tuned oscillator," in *1972 IEEE-GMTT International Microwave Symposium Digest of Tech. Papers*, pp. 185–186, May 1972.

Absolute Load Detection with Microwave Gunn Oscillators

LARRY D. PARTAIN, MEMBER, IEEE, W. ALLAN COOK, HUA-FENG HUANG, MEMBER, IEEE, AND LEWIS C. GOODRICH, MEMBER, IEEE

***Abstract*—The change in the dc *I–V* properties of Gunn-flange microwave oscillators with a change in microwave load are shown to provide a method of measuring the physical properties of dielectric samples that only requires the measurement of dc voltages and currents. A phenomenological equivalent-circuit model has been developed that predicts a dependence of detection sensitivity on bias resistance that agrees closely with experiment and that explains the restrictions on such a bias resistor's maximum allowed value. Properties of a prototype system capable of measuring sample size with 8-percent accuracy are presented.**

I. Introduction

It is well known that microwave oscillators are quite sensitive to the load impedance they see. A high degree of isolation between a source and a variable load is usually necessary to achieve stable operation. However, if close coupling is used, oscillator power and frequency variations with load changes can be used to measure load properties. Such measurements can be accomplished rapidly, continuously, and with no mechanical contact with the load.

The strong dependence of Gunn microwave oscillators on load has been studied [1] as a means of tuning and of obtaining maximum output power. Utilizing this effect, Nagano and Akaiwa [2] have described a method for measuring the relative velocity of a moving-target load by monitoring oscillatory changes in the dc bias current flowing through a Gunn device. Application of their results leads to extremely simple Doppler radar systems. Generalized treatments have been presented by Takayama [3] and by Nygren and Sjolund [4] for such Doppler-signal detection with various negative-resistance-device oscillators such as Gunn oscillators.

In this short paper we discuss extensions of these earlier results that allow absolute values of load parameters to be measured by monitoring just the dc voltage and current of Gunn-device oscillators. In Section II we describe the measurement apparatus and the experimental results. In Section III a phenomenological equivalent-circuit model is developed that predicts a given dependence of detection sensitivity on dc bias resistance that is verified by experiment.

II. Experimental Results

The dc current-voltage measurements were made using two 4½-digit digital multimeters and a variable-voltage power supply. The Gunn devices were mounted in X-band Gunn flanges whose construction and operating characteristics have been described elsewhere [5]–[7]. The dc properties of these flanges are to serve as low-pass filters so that no microwave or high-frequency signals generated by the Gunn device reach the flange terminals. In order to demonstrate that the experiments were characteristic of a range of systems, measurements were taken on two different flanges attached to one of three different cavities. One flange was a Frequency West G0(X)-104 flange with the Gunn device permanently mounted. The other was a Frequency West G0(X)-107 flange slightly modified to accept a Microwave Associates

Manuscript received May 27, 1975; revised March 5, 1976.
L. D. Partain is with the Department of Electrical Engineering, University of Delaware, Newark, DE 19711.
W. A. Cook, H. F. Huang, and L. C. Goodrich are with the Engineering Physics Laboratory, E. I. DuPont Company, Wilmington, DE 19898.