

A VARACTOR-TUNED OSCILLATOR WITH LINEAR TUNING CHARACTERISTIC

In certain areas of microwave applications such as FMCW radars, FM modulators and phase-locked loop (PLL) circuits, the system performance is enhanced when the frequency of a voltage-controlled oscillator (VCO) is linear with respect to its control voltage.^{1,2} Although YIG-tuned oscillators have good linearity as well as a wide tuning range, their low tuning speed restricts their usefulness.¹ Varactor-tuned oscillators (VTO), on the other hand, feature high speed tuning, although their linearity is inferior to that of YIG-tuned oscillators. Therefore, improving the linearity of VTOs has been the subject of much research. D. Peterson, for example, suggests a varactor doping profile³ which is capable of providing a wide band tuning linearity with series-tuned oscillator circuits. E. Marazzi, et al. use a CAD technique⁴ to optimize linearizing networks with predetermined topology. D. Kajfez⁵ utilized a pole-zero representation of lossless networks and a Taylor series technique in order to investigate the possibility of linear tuning. In some cases, external linearizers can be a solution, although

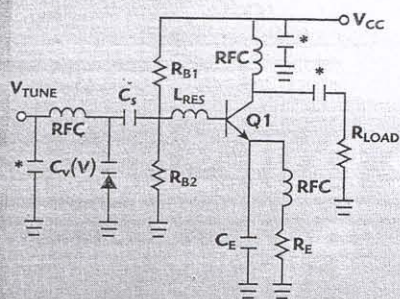
they introduce new problems, such as additional power consumption, lower input impedance and narrower tuning bandwidth.⁷

In this article, a practical approach to the design of linear VTOs is presented, which focuses on the optimum selection of a varactor and the design of its coupling network. Since the tuning linearity of a VTO is mainly determined by the voltage-to-capacitance relationship of a varactor as well as its coupling network, the choice of a varactor and the design of a coupling network are critical to achieve the desired tuning linearity. Measurement results of two 4.3 GHz VTOs are also included as an example.

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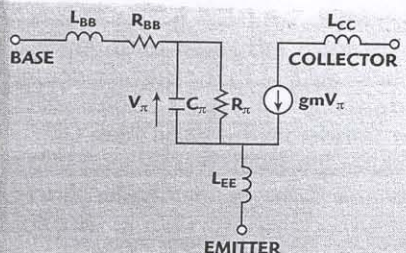
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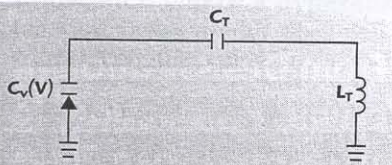
* $Z_C = 0$ AT OPERATING FREQUENCY

▲ Fig. 1 General varactor-tuned oscillator schematic.



▲ Fig. 2 Equivalent circuit of packaged microwave bipolar transistors.

▼ Fig. 3 Simplified VTO model.



DERIVATION OF A SIMPLIFIED VTO MODEL

The schematic of a popularly used VTO is shown in **Figure 1**.⁶ The capacitor at the emitter terminal, C_E ,

forms a positive feedback loop together with the base-emitter diffusion capacitance of Q1. The negative resistance generated by Q1, which is needed to sustain a stable oscillation, can be controlled by varying C_E .⁶ L_{RES} , C_S and $C_V(V)$ together act as an effective resonator whose reactance can be changed by controlling V_{TUNE} . The coupling capacitor, C_S , not only blocks the DC current flow between varactor and Q1, but also provides control over the tuning bandwidth.

Figure 2 is a general equivalent circuit of a packaged bipolar transistor, where the base-to-collector depletion capacitance is neglected for simplicity. If $R_\pi > 1/\omega C_\pi$ is assumed, which is valid at high frequency, R_π can also be neglected. After removing the bias circuitry and replacing Q1 by its equivalent circuit, a simple VTO model is obtained (as shown in **Figure 3**), where only the reactive elements that are responsible for the determination of the oscillation frequency are included. C_T and L_T are given by

$$C_T = (C_S^{-1} + C_\pi^{-1} + C_E^{-1})^{-1} \quad (1)$$

$$L_T = L_{RES} + L_{BB} + L_{EE} \quad (2)$$

Parasitic inductances, such as the via inductance and lead inductance of the varactor not considered here, may also be lumped into Equation 2, if their actual values are known or can be estimated. The varactor capacitance, $C_V(V)$, is a function of the con-

trol voltage, V , and can be expressed as

$$C(V) = \frac{C_{j0}}{\left(1 + \frac{V}{V_j}\right)^M} \quad (3)$$

where

C_{j0} = zero-bias junction capacitance
 V_j = built-in junction potential
 M = grading coefficient determined by the doping profile, generally ranging from 0.5 to 2

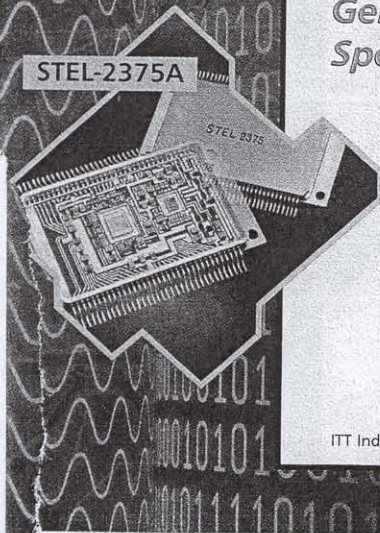
DESIGN PROCEDURE CONSIDERING TUNING LINEARITY

The resonant frequency of the simple VTO model is given as

$$\omega_0(V) = \frac{1}{\sqrt{L_T(C_V^{-1}(V) + C_T^{-1})}} = \frac{1}{\sqrt{L_T C_T} \sqrt{1 + \frac{C_T}{C_V(V)}}} = \frac{1}{\sqrt{L_T C_T} \sqrt{1 + \frac{C_T}{C_{j0} \left(1 + \frac{V}{V_j}\right)^M}}} \quad (4)$$

As shown in Equation 4, $\omega_0(V)$ is a nonlinear function of V , and therefore, in general, the tuning curve will not be a perfectly straight line. First, two extreme cases are considered:

[Continued on page 109]



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When

$$\frac{C_T}{C_V(V)} \gg 1$$

$\omega_0(V)$ is approximated by

$$\omega_0(V) \approx \frac{1}{\sqrt{L_T C_V(V)}} = \left(\frac{1}{\sqrt{L_T C_{j0}}} \right) \cdot \left(1 + \frac{V}{V_j} \right)^{\frac{M}{2}} \quad (5)$$

and thus the varactor capacitance solely determines the resonant frequency along with L_T . This would be the limiting case for a wideband VTO, and using a hyperabrupt junction varactor with $M=2$ will result in a straight line.

When

$$\frac{C_T}{C_V(V)} \ll 1$$

$\omega_0(V)$ is approximately given by Equation 6 where a hyperabrupt junction varactor with $M=1$ will linearize the tuning curve.

$$\omega_0(V) \approx \frac{1}{\sqrt{L_T C_T}} \left(1 + \frac{1}{2} \frac{C_T}{C_V(V)} \right) = \frac{1}{\sqrt{L_T C_T}} \left(1 + \frac{1}{2} \frac{C_T}{C_{j0}} \left(1 + \frac{V}{V_j} \right)^M \right) \quad (6)$$

In this case, the oscillation frequency is mainly determined by C_T , which is

much smaller than $C_V(V)$, resulting in a narrowband VTO.

In the general case between the two extremes just mentioned, the optimum grading coefficient, M , will be somewhere between 1 and 2, which means that a hyperabrupt junction varactor with $M \geq 1$ is essential in order to achieve the best linearity in a VTO configuration. Also, Equation 4 shows that the tuning curve is completely specified by four circuit parameters, $L_T C_T$, C_T/C_{j0} , V_j and M , as long as the VTO model is correct. The design goals — center frequency, bandwidth and tuning linearity — should be met by adjusting these parameters. Although the number of design parameters is greater than the number of design goals, the actual design procedure is less flexible than it seems to be, since a circuit designer generally does not have individual control over C_{j0} , V_j and M . These values become fixed once he chooses one from a list of available varactors.

$L_T C_T$ and C_T/C_{j0} can be expressed in terms of the center frequency f_c , the tuning bandwidth BW , V_j and M , by evaluating Equation 4 at the start and stop frequencies of oscillation. V_1 and V_2 are the start and stop tuning voltages, respectively

$$L_T C_T = 4 \frac{1 - K}{(2f_c + BW)^2 - K(2f_c - BW)^2}$$

$$\frac{C_T}{C_{j0}} =$$

$$\left(\frac{(2f_c - BW)^2 (1 - K)}{(2f_c + BW)^2 - K(2f_c - BW)^2} - 1 \right) \cdot \frac{1}{\left(1 + \frac{V_1}{V_j} \right)^M} \quad (7)$$

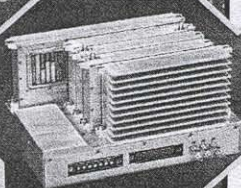
where

$$K = \left(\frac{1 + \frac{V_2}{V_j}}{1 + \frac{V_1}{V_j}} \right)^M$$

If the design requirement of f_c and BW are given, Equation 7 allows an entire tuning curve with a specific varactor to be obtained by evaluating Equation 4 at several control voltages. In order to calculate the best-fit straight line (BSL) of an actual tuning curve, a general-purpose spreadsheet program might be useful. Actual values of V_j and M from datasheets can be used, or V_j may be fixed at some representative value, namely 2, and the calculations repeated with different values of M to find the optimum grading coefficient. After a proper varactor is chosen, C_T and L_T can be

[Continued on page 111]

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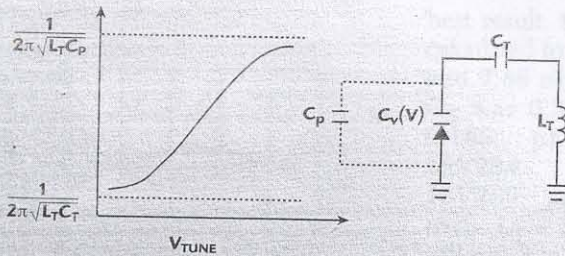
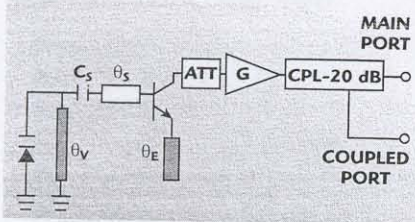


Fig. 4 The effect of varactor packaging capacitance on the tuning curve.

Fig. 5 Simplified RF equivalent circuit of a 4.3 GHz VTO.



found from Equation 7, thus completing the first-pass design.

Until now, the varactor package capacitance, C_p , has been neglected. It is connected in parallel with the varactor. Although this capacitance is small, typically ranging from 0.1 to 0.2 pF, depending on the structure, it sometimes has a non-negligible effect on the linearity. In the high frequency region of the tuning curve, the varactor capacitance approaches its minimum and the package capacitance of the varactor sets a lower limit of the varactor total capacitance. This will, in turn, impose an upper limit on the tuning curve. Figure 4 illustrates this situation, together with the effect of C_T , which establishes the lower limit of the oscillation frequency. The effect of C_T will actually be compensated if the optimum M is selected by repetitively evaluating Equation 4, simply because it is a function of C_T . The effect of C_p , on the other hand, can be compensated with an additional inductive element (θ_v) across the varactor, as shown in the overall RF schematic of Figure 5. This inductance will make the slope of varactor reactance steeper as the frequency approaches the resonant frequency, thus shaping the tuning curve concave upward, as will be seen in the next section.

DESIGN AND MEASUREMENT RESULTS

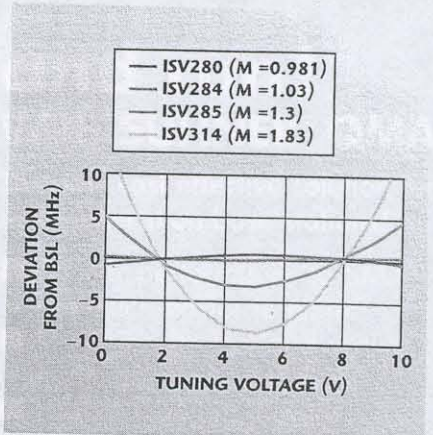
Based on the procedure outlined in the previous section, a 4.3 GHz VTO

was designed. The SPICE parameters of varactors under consideration are listed in Table 1. With the design requirement of $f_c = 4.3$ GHz and $BW = 160$ MHz, together with the listed varactor parameters, $L_T C_T$ and C_T/C_{j0} were calculated using Equation 7 for each varactor. Next, the tuning curves were calculated with Equation 4, and their deviation from corresponding BSLs are shown in Figure 6. Since this design belongs to a narrowband VTO whose fractional bandwidth is 3.7 percent, an optimum M close to 1 was expected. 1SV284 ($M = 1.03$) yields the smallest deviation from a BSL and 1SV280 ($M = 0.981$) gives the second

TABLE 1
SPICE PARAMETERS
OF FOUR VARACTORS

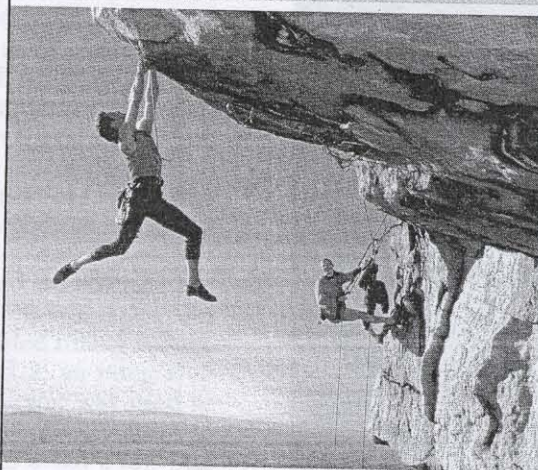
Part No.	I_s	BV	C_{j0} (pF)	V_j	M
1SV280	$5.38 \cdot 10^{-16}$	15	6.89	3.272	0.981
1SV284	$1.07 \cdot 10^{-15}$	10	23.0	2.18	1.03
1SV285	$4.17 \cdot 10^{-16}$	10	6.9	2.6	1.3
1SV314	$5.38 \cdot 10^{-16}$	10	10.4	2.567	1.83

Fig. 6 Calculated deviations from a BSL.



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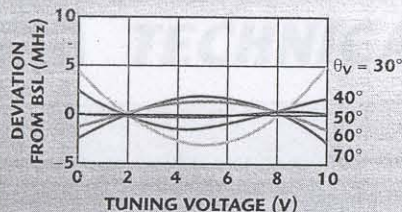
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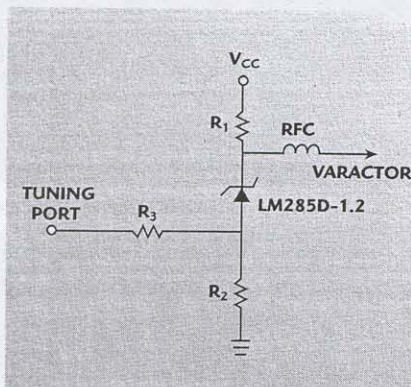


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▲ Fig. 7 The effect of θ_V on the tuning curve shape.

▼ Fig. 8 Tuning voltage shifter.



whose lengths are smaller than $1/4 \lambda$, and thus correspond to L_{RES} and C_E , respectively. C_S is realized by a series connection of two 1005-size capacitors. θ_V cancels the varactor packaging capacitance, and can be controlled to further shape the tuning curve, if needed. It can be seen, in **Figure 7**, that the curve becomes concave upward as θ_V becomes electrically shorter. The optimum θ_V is found to be approximately 50° . The AT-42086 bipolar transistor from Agilent was biased at 20 mA and its operating f_T was approximately 7 GHz. To increase both isolation and output power, a Minicircuits ERA-5SM amplifier was cascaded. Its S_{21} at 4.3 GHz is about 14 dB with 60 mA of bias current. The output power at the main port is designed to be greater than 10 dBm. The coupled port can be used to provide a feedback signal in a PLL, or a downconverting LO in FMCW radars.

Two VTOs were implemented based on virtually the same design. Both VTOs share the same schematic, but one of them additionally employs an added tuning voltage shifter shown in **Figure 8**. LM285D-1.2, an integrated zener diode from National Semiconductor, was used in this circuit. The shifter adds a constant offset voltage to the incoming tuning voltage in order to avoid a low bias region where junction rectification may occur due to the RF signal. This rectification is known to increase a post-tuning drift³, and thus is undesirable. Another function of this shifter is to scale down the range of tuning voltage by a voltage divider formed by R_3 and R_2 . C_S must be increased accordingly, since the actual range of tuning voltage applied to the varactor is reduced. The advantage with this simple shifter is that it offers another control over the tuning curve by carefully adjusting the amount of shifting and scaling.

[Continued on page 114]

best result. L_T was calculated to be 3.9 and 7.88 nH, and C_T was 0.37 and 0.185 pF for 1SV284 and 1SV280, respectively. $L_T = 3.9$ nH will result in a very small value of L_{RES} , since L_T is the sum of all the inductances along the resonating path as shown by Equation 2. Therefore, 1SV280 was finally selected despite its slightly worse linearity than 1SV284, since, in this case, $L_T = 7.88$ nH is a more desirable value.

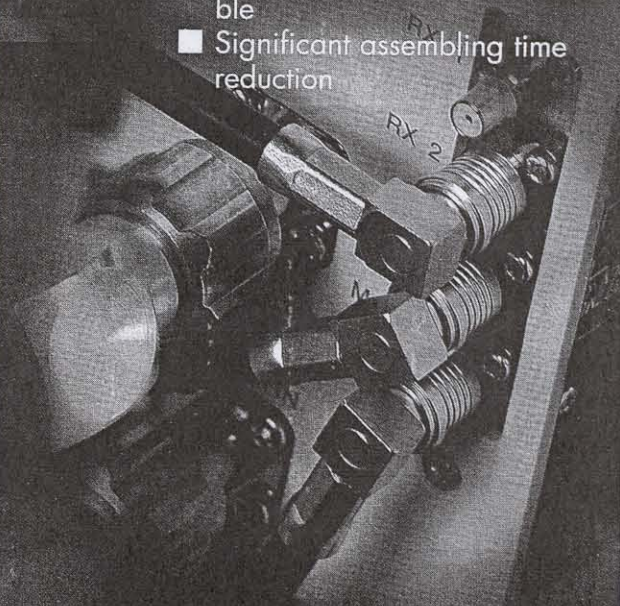
In the overall RF schematic, θ_S and θ_E are microstrip lines

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TABLE II

MEASURED ELECTRICAL CHARACTERISTICS OF THE TWO VTOs

	VTO without Shifter	VTO with Shifter
Power supply (V)	15	
Current consumption (mA)	20 for VCO only	
Tuning voltage (V)	0 to 10	
Center frequency (MHz)	4271.3	4292.5
Tuning bandwidth (MHz)	170.6	180.0
Output power @ center frequency (dBm)	10.5	10.2
Power variation across entire band (dB)	< 1.4	< 1.9
Maximum deviation from BSL (MHz)	3.83	1.32
Average tuning sensitivity (MHz/V)	16.45	17.84
Phase noise (100 kHz offset) (dBc/Hz)	-104.4	-97.9
Second harmonic (dBc)	-23.5	-28.5
Pushing factor (MHz/V)	4.7	1.8
Pulling factor (12 dB return loss) (MHz)	19.2	16.8

input impedance at the tuning port is lowered, and the tuning speed will be slower due to the response time of the zener diode and the RC delay from the biasing resistors and diode capacitance. This delay will have to be carefully checked if the VTO is to be used in fast-switching applications.

CONCLUSION

A design procedure has been proposed which emphasizes the tuning linearity of a VTO. If the center frequency and bandwidth requirements are

given, $L_T C_T$ and C_T/C_{j0} can be calculated using Equation 7 with the SPICE parameters of a specific varactor. Next, the entire tuning curve can be obtained with Equation 4, which gives the resonant frequency of the simplified VTO model. With the help of a spreadsheet program, a BSL and the deviation of an actual tuning curve from a BSL can be calculated. This procedure enables a designer to evaluate the tuning linearity prior to actual circuit design, and thus helps to select the optimum varactor. Based on this design procedure, two 4.3 GHz VTOs were designed, implemented and measured. The tuning linearity and ease of design were both considered in selecting the varactor. By using an inductive element parallel to the varactor, the shape of the tuning curve was further optimized. The maximum deviations from a BSL were measured to be 3.83 and 1.32 MHz for the VTOs with and without a tuning voltage shifter, respectively, verifying the usefulness of the proposed approach.

ACKNOWLEDGMENT

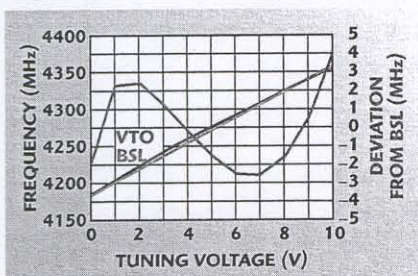
The authors would like to thank Mr. Jae-Hoon Lee for his help and discussion during the measurements. ■

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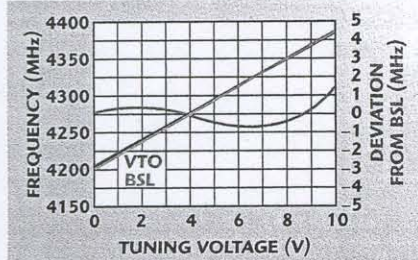
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▲ Fig. 9 Measured tuning curve of the VTO without tuning voltage shifter.

Fig. 10 Measured tuning curve of the VTO with tuning voltage shifter. ▼



The measured electrical characteristics of the two VTOs are summarized in Table 2. The shifter was adjusted to provide approximately 2.5 V of shifting and 0.7 of scaling down. A comparison of Figures 9 and 10 show that the VTO with a shifter has a better linearity with 1.32 MHz of maximum deviation from a BSL, compared with the VTO without a shifter, which has 3.83 MHz of maximum deviation. The price paid for this improvement of linearity is the higher phase noise due to the noise generated by the zener diode. Also, the



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