Tracking Phaselock Loop Characteristics with a VCO Using a Barium Strontium Titanate (BST) Thin-Film Varactor

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***Abstract* — Barium Strontium Titanate (BST) varactors utilized in a VCO demonstrate unique tuning characteristics compared to junction varactors. A 2.7 GHz microstrip line VCO operates in a tracking phaselock loop configured as an X4 frequency multiplier. BST oscillator tuning and the effect on loop dynamics is observed by the intentional design of an under damped system. Oscillator noise correction and peaking is noted particularly at low tune voltages. Loop dynamics are addressed through proper design of the loop compensator filter. The differences between BST varactor and junction varactor operation in frequency control are noted.**

***Index Terms* — BST, VCO, PLL, loop filter, tracking loop, ferroelectric, resonator.**

I. INTRODUCTION

Oscillators embedded in a phaselock loop (PLL) benefit from improved phase noise performance if the characteristics of the loop are properly designed. Stringent phase noise requirements are a challenge for high power oscillator designs, which emphasize DC conversion efficiency, output power, tune range, and harmonic content. Circuit phase noise, sensitivity to vibration and microphonics, and power supply noise sensitivity are suppressed through the error feedback function associated with the loop gain and dynamics [1].

This work addresses the influence of oscillator tune characteristics on loop performance. The tune characteristic of a BST varactor displays a unique C–V curve, which unlike a junction varactor, presents a non-monotonic tune response. This tune response coupled with the reduction in oscillator tune gain at zero volts is discussed. The effect on operating performance in a tracking phase loop configured as frequency multiplier, a method of readily observing the loop dynamics and required design modifications are presented.

II. BST VS. JUNCTION VARACTOR PROPERTIES

The capacitance variation of a BST varactor was compared with that of a semiconductor junction varactor. BST varactor exhibits symmetrical tuning about 0 V (see Fig. 2) while the junction varactor increases monotonically with decreasing reverse bias down to forward bias of 570 mV. Beyond that the junction shows a positive reactance due to parasitic effects and high forward conduction (see Fig. 1). This is due to the fundamental difference in the tuning mechanism for the BST and the junction varactor. The C–V curve for a junction varactor is asymmetrical about zero volts since the tuning mechanism relies on existence of a depletion region in the presence of reverse bias and the change of depletion width with voltage. Capacitive tuning in a BST varactor is due to distortion of the unit cell as a function of applied electric field magnitude and hence is independent of the sign of the applied voltage [2].

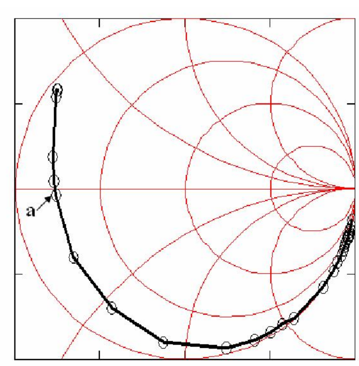

Fig. 1: Junction varactor reflection coefficient versus swept bias from -30V to +1V. Point (a) on the chart represents 570 mV forward bias. The measurement frequency is held constant at 50 MHz.

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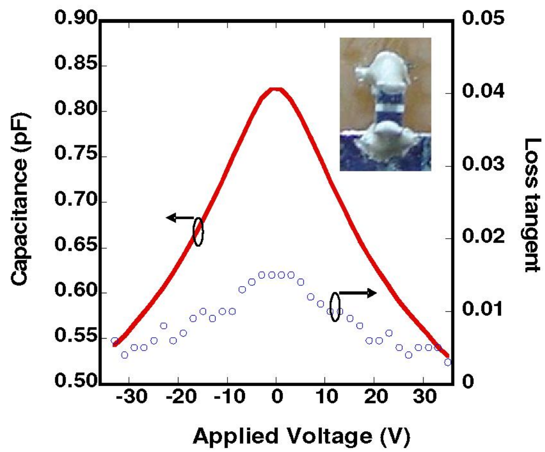

Fig. 2: Representative C-V characteristics of a BST interdigital varactor measured after dicing at 1 MHz.

III. BST VCO Operation

The VCO in this work operates at 2.7 GHz and is shown in Fig. 4. The inductive portion of the resonator consists of a 60 degree shorted 30-ohm microstrip line fabricated on Roger 4350 material. The coupling elements to the active device and varactor network are also mounted on the same board. This implementation uses a single BST varactor located in the upper right corner of the card. The design card provides for additional series stacked back-to-back BST varactors as well as additional tune bias RF choke components [3]. This configuration will reduce the RF voltage across each varactor by a factor of N, where N is the total number of varactors in the stack up, and hence improve linearity [4]. The average oscillator tune gain, K_o for 0V << $V_{TUNE} \leq$ 30V (30V limit imposed by the active loop compensator) is 1 MHz/V. Open loop measurements to 60 V also demonstrates good linearity. Although the breakdown voltage of the BST varactor is 200 V, measurements are limited to 60 V due to on-chip ceramic on glass (COG) chip capacitor breakdown. The adjustment of the capacitive coupling between the varactor, transmission line, and active device, will set K_o and linearity.

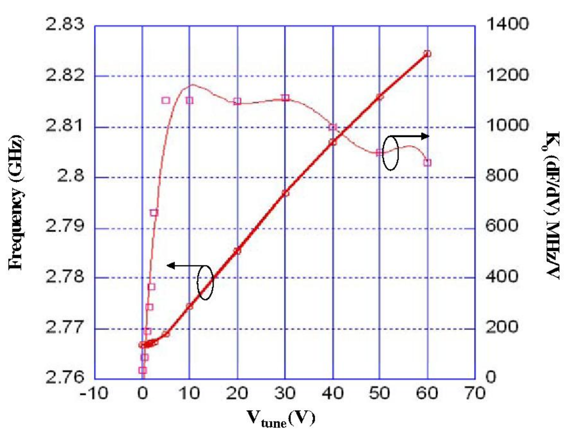

Fig. 3: Measurement of tune frequency and K_o versus tuning voltage

The incremental K_o, evaluated by measuring a small deviation in tune voltage about a nominal value is 400–550 KHz/V below a tune voltage of 10 V. As the tune voltage approaches zero this gain is reduced, see Fig. 3. Depending on resonator topology, varactor index of nonlinearity (γ for junction varactors or power law of the junction), and the interface to the active device, it is possible to extend the tune voltage [5] to lower values while maintaining K_o.

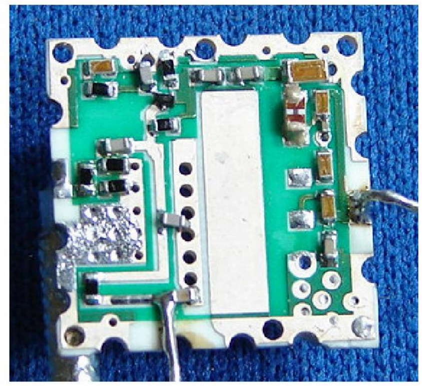

Fig. 4: Microstrip VCO shown with tune line input on the right output power obtained from matching transmission line on left. The BST varactor is located in upper right corner.

Evidence of the reduction in tune gain near zero volts for the VCO presented in this work is clear by inspection of loop behavior discussed next.

IV. Tracking Loop Framework

The property of oscillator performance and correction by a tracking loop is documented in a number of references [1, 6, 7]. The loop bandwidth is a function of the VCO gain, K_o and the loop control elements, which include the phase detector gain, K_d, loop compensator, F(s), and counter, N. If the counter is absent, the loop acts as tracking filter, the filter response set by the loop compensator. The VCO will track a reference source, f_{REF}, which must be at the same frequency as the VCO. Otherwise if a counter is present with a count value of N, the VCO will operate at Nf_{REF}. If the loop has sufficient open loop gain given by the product of $K_oK_dF(s)/sN$ then the phase noise of the reference is transferred to the controlled VCO. Although other sources of noise in the loop can dominate, for this analysis they are neglected. The transfer function assuming small signal analysis, a stable locked loop and N=1 is given in terms of the Laplace variable $s=j\omega$ as,

$$\theta_o(s)=\frac{K_oK_dF(s)}{s+K_oK_dF(s)}\phi_i(s) \tag{1}$$

The phase response of the output VCO, $\theta_o(s)$, tracks the response of the input reference, $\phi_i(s)$ and is low pass function. The loop compensator $F(s)$ assists in shaping the noise response and compensates for a varying K_o. However, this adjustment impacts the stability of the loop, which therefore limits its effectiveness as a compensator. The transfer function of the corrected VCO phase is the error response of the loop. This function is given by

$$\theta_o(s)=\frac{s}{s+K_oK_dF(s)}\phi_V(s) \qquad (2)$$

The high pass characteristic permits the VCO uncorrected phase $\phi_V(s)$ to be suppressed by the loop gain. The adjustment of this low pass and high pass frequency response permits noise correction of the VCO. The ability to track a reference source accurately is dependent on the available open loop gain.

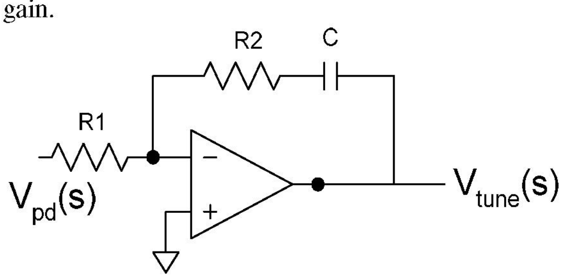

Fig. 5: Active loop compensator produces a second order type II PLL used in this study.

The response is shaped by adjustments in F(s) and will control noise peaking and permit optimum loop bandwidth for noise correction [8]. One form of loop compensator consists of a zero and pole combination, referred to as a RRC filter [9] and permits independent control of loop dampening and bandwidth. Consider F(s) of the type shown in Fig. 5. The phase detector voltage and the VCO tune voltage are denoted as V_{pd} (s) and V_{tune} (s). The loop transfer function will be second order-type II and the response is

$$\theta_o(s)=\frac{K_dK_o(s\tau_2+1)/\tau_1}{s^2+sK_dK_o\tau_2/\tau_1+K_dK_o/\tau_1}\phi_i(s) \qquad (3)$$

Time constants τ_1 and τ_2 are given by R_1C and R_2C products respectively. This 2nd order transfer function is written in terms of control feedback notation, which identifies a natural frequency ω_n and a dampening factor δ. Using this notation (3) is re written as

$$\theta_o(s)=\frac{2\delta\omega_n s+\omega_n^2}{s^2+2\delta\omega_n s+\omega_n^2}\phi_i(s) \qquad (4)$$

Expression for the output power spectral density (PSD) follows from (1) through (4) after taking the magnitude squared of the respective transfer functions. The final values of ω_n and δ used in the loop in this work permit small variations in the VCO gain to be readily measured. This is particularly true for tuning voltages in the vicinity of 0 V. If δ is set to unity, which is a moderate over damped, no peaking in the loop response occurs. Although significant noise correction is apparent from the power spectrum measurement, it is difficult to reconcile the actual loop bandwidth from this measurement alone. In addition, loop gain variations from K_o reduction are difficult to detect unless large.

V. Measurement Results

The uncorrected oscillator, see Fig. 6(a) and corrected, see 6(b) power spectrum is shown. The loop bandwidth is 220 KHz and near carrier to noise ratio for the locked loop is improved. The loop compensator is then reset for a dampening factor much less than 0.7 in Fig. 7 and 10 dB noise peaking near the loop bandwidth is shown.

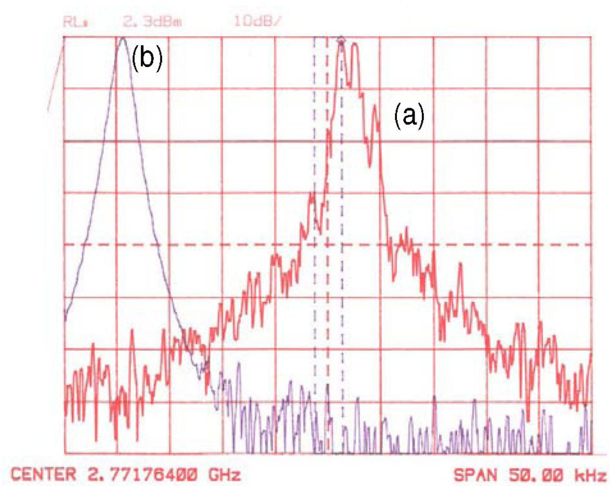

Fig. 6: The uncorrected noise (a) is suppressed in the tracking wide loop when properly damped for the loop gain and has no noise peaking (b).

Adjusting the reference frequency forces the VCO to track. The comparison frequency is decremented down to 690 MHz. As the tune line voltage decreases, the BST varactor operates on a lower sensitivity portion of the C-V curve. The VCO and loop gain drop. Because the loop transfer function is 2nd order, the reduction in loop gain will force the dampening factor to decrease. Severe reduction in loop gain will eventually allow the open loop gain to cross unity gain at a 12dB/octave rate resulting in loop instability [10].

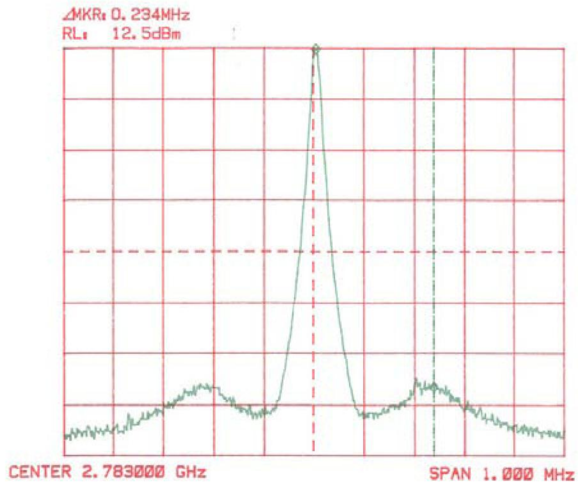

Fig. 7: Adjustment in the loop compensator permits noise peaking and assist in monitoring the loop gain variation by monitoring the power spectrum.

Prior to instability, noise peaking occurs; see Fig. 8(a). The peaking occurs at reduced offset frequency, less than 100 kHz. Further reductions in tune line voltage causes instability. The loop is still locked, tracking the reference. But incidental FM modulation is evident at less than 50 KHz see Fig. 8(b).

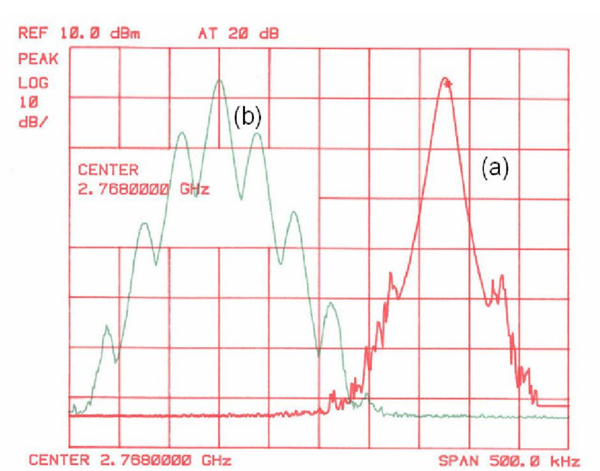

Fig. 8: Significant drop in loop gain due to K_o reduction causes larger noise peaking (a) and eventually instability (b).

In the loop studied here, the tune line voltage could not be reduced below 850 mV. The VCO using a BST varactor exhibits non-monotonic tuning. As the tune line voltage is decreased, incidental FM, coupled with further reduction in loop dampening, will force the sign of the VCO K_o to reverse. As seen in Fig. (2), the total decrease in tune voltage below zero volts supports a reduction in varactor capacitance. Now a slight reduction in feedback control voltage will cause the loop to hang and the VCO is forced to a negative rail voltage if it exists. The conventional phase frequency comparator will not permit acquisition without additional frequency discrimination circuits. To eliminate this problem, the output of the compensator should be clamped or the supply voltage $-V_b$ not used.

VI. Conclusion

A tracking phaselock system is used to study the unique tune characteristics of a VCO utilizing a BST varactor. An under damped phaselock loop configured as a tracking loop, frequency multiplier, facilitates the evaluation. Although the BST varactor can provide suitable tuning gain characteristics, care must be exercised as the tune voltage approaches the inflection point near zero volts. Phase noise peaking will occur as the loop gain is reduced. Since the BST based VCO tune curve is not monotonic, PLL hang is likely if not accounted for in design.

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