

## APPLICATION NOTE

# APN1012: VCO Designs for Wireless Handset and CATV Set-Top Applications

## Introduction

Voltage Controlled Oscillators (VCOs) have come to the forefront of RF designs together with the first PLL circuits. In the era before the PLL, oscillators were mostly free running, and only in rare cases were varactors used for modulation or temperature compensation. Nowadays, we rarely see free running oscillators, instead they have become varactor-controlled oscillators. This is because most RF applications require band coverage, which can be realized through the PLL circuit requiring two sources of RF power. The reference source frequency is often a VCXO or TCXO, while the other frequency is controlled by the PLL phase detector.

Usually, both VCXO/TCXO and the RF VCO are voltage controlled oscillators. The difference between a reference oscillator and a tuned VCO is that the former usually has a very high Q resonator,

which allows for very stable oscillation, while the latter has a lower Q resonator, allowing a relatively high tuning range. In reference oscillators, varactors are used for fine tuning or temperature compensation (TCXO). In tunable oscillators, varactors are used to change (tune) the frequency. In some VCOs, varactors may be used also for modulation, for example in a DECT system where modulation is used to generate a constant envelope GMSK signal.

Although it is a small part of the RF design, the VCO is often a major headache for designers. The goal, in this application note, is to show how Skyworks products and services may help you to overcome VCO concerns and help to make your design among the best products on the market.

### VCOs in Digital Wireless Phones

Consider the hypothetical wireless handset phone. Today, the handset is a dual-band (cellular/PCS) and multimode system employing many VCO functions. There are many ways to realize these functions, making it virtually impossible to specify the frequency and tuning range for all designs. However, there are certain common features that are outlined in Figure 1.

In a typical receiver, dual conversion superheterodyne solutions are usually employed. They convert either 900 MHz (cellular) or 1.8 GHz (PCS) down to the SAW frequency range, which may be between 90–400 MHz. Further, this signal is either downcon-

verted or demodulated into a digital I/Q signal using a lower frequency IF VCO. The transmitter path is either directly modulated at 900 MHz or uses a dual conversion scheme requiring at least two VCOs.

When dual-band requirements are needed, up to eight or more VCOs may be required to satisfy specific frequency plans. This is often a technically and economically restrictive solution. Many designers try to solve this over-VCOed problem using both smart frequency planning and multiband VCOs, as shown in Figure 1.

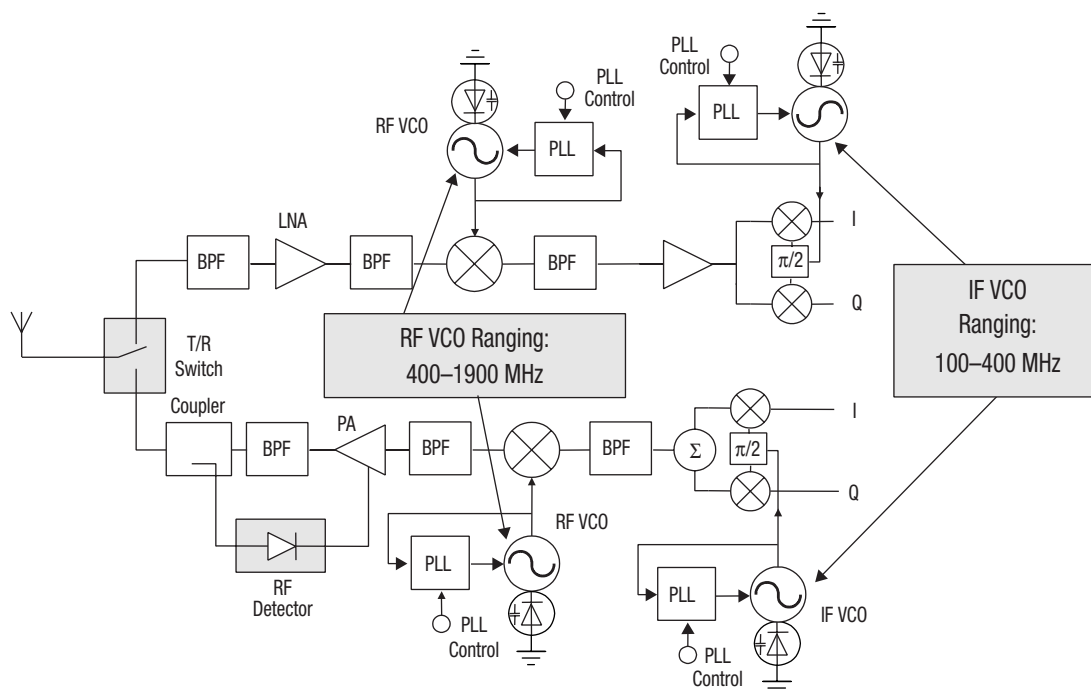


Figure 1. VCOs in a Digital Wireless Phone

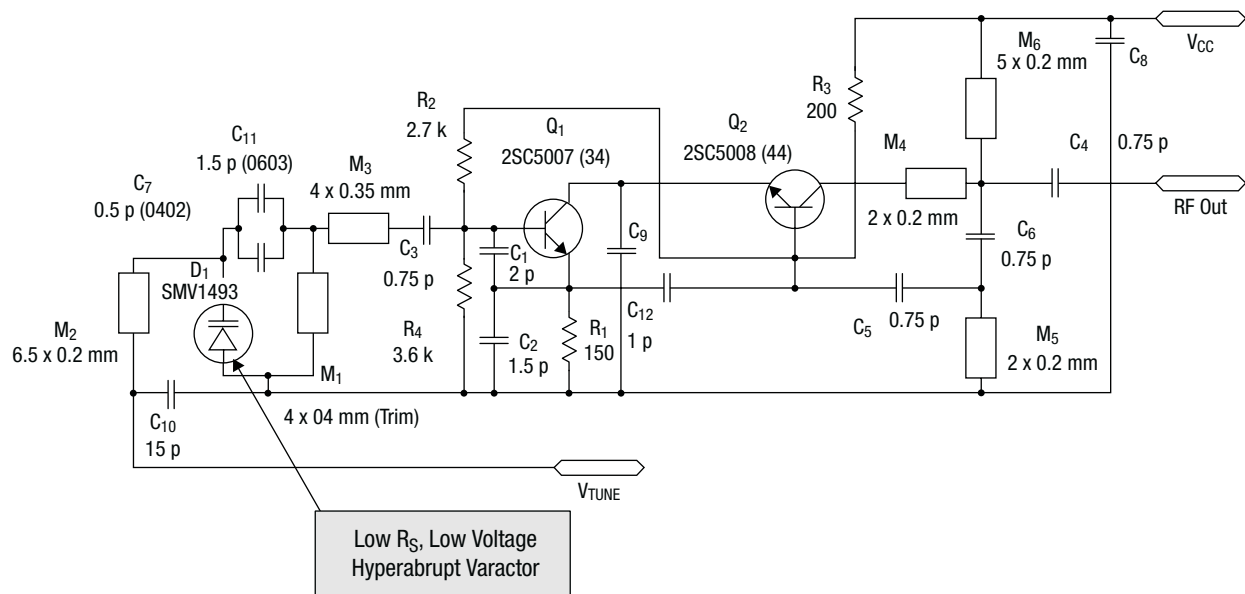
## Fundamental Low Noise Colpitts VCO

The characteristic feature of the Colpitts VCO is that it uses a capacitive divider for the feedback consisting of  $C_1$  and  $C_2$ , and an inductive branch including a parallel resonator and series capacitor  $C_3$ . The parallel resonator includes inductive element  $M_1$  (that may be a discrete inductor for lower frequencies or a length of microstrip line for RF) and a capacitive branch, consisting of a varactor and a series capacitor(s). The entire inductive branch should have inductive impedance at the frequency of oscillation, otherwise there will be no oscillation. This means that the resonant frequency should be higher than the oscillation frequency.

Note that the resonator current circulates through the varactor, series capacitor  $C_{11}$  and inductor  $M_1$  and is the largest current in the tank circuit. Because of this, losses introduced in this current path are the crucial ones with respect to phase noise.

Without delving deeply into phase noise theory, we note that phase noise is inversely proportional to the power bypassed through the feedback loop, and the loaded Q of the tank circuit. Thus, the more power lost on the way to the transistor base, the higher the noise. It is clear that varactor loss plays a crucial role in the phase noise property of the VCO. If phase noise is an issue, the varactor series resistance should be carefully considered.

There is an additional concern because phase noise is not only a function of varactor loss. The varactor capacitance voltage characteristic has a crucial impact on phase noise as well. With a higher capacitance ratio, the varactor's coupling to the resonator is reduced, resulting in lower resonator current. Therefore, a hyperabrupt varactor having higher series resistance is often a better choice than a lower capacitance ratio abrupt varactor having lower series resistance.



**Figure 2. Low Noise, High-Performance Colpitts VCO**

### Differential VCO for Integration with an RF IC

Designs based on RF IC solutions, with built-in VCOs, often employ a differential VCO configuration. One possible differential VCO configuration is shown in Figure 3. In this case, the tank circuit is formed by  $C_3$ ,  $C_4$  and a resonator  $C_8$ ,  $C_9$ ,  $D_1$ ,  $M_1$ . Here again, the resonator current plays a decisive role in phase noise definition. Thus, phase noise is strongly dependent on resonator loss. Capacitors  $C_3$  and  $C_4$  help establish the correct phase shift value in the feedback loop moving oscillations closer to the resonant frequency. This is the principal difference between a Colpitts

and a differential VCO. In the Colpitts case, the resonant frequency is always higher than the oscillation frequency; in the differential VCO the resonant and oscillation frequencies may coincide. Thus the loaded Q of the circuit becomes significantly higher, and the feedback loop losses are increased due to the higher resonant currents. When this happens, the differential VCO is more vulnerable to resonator loss than the Colpitts VCO and usually shows 5–10 dB higher noise if compared to an equivalent Colpitts case.

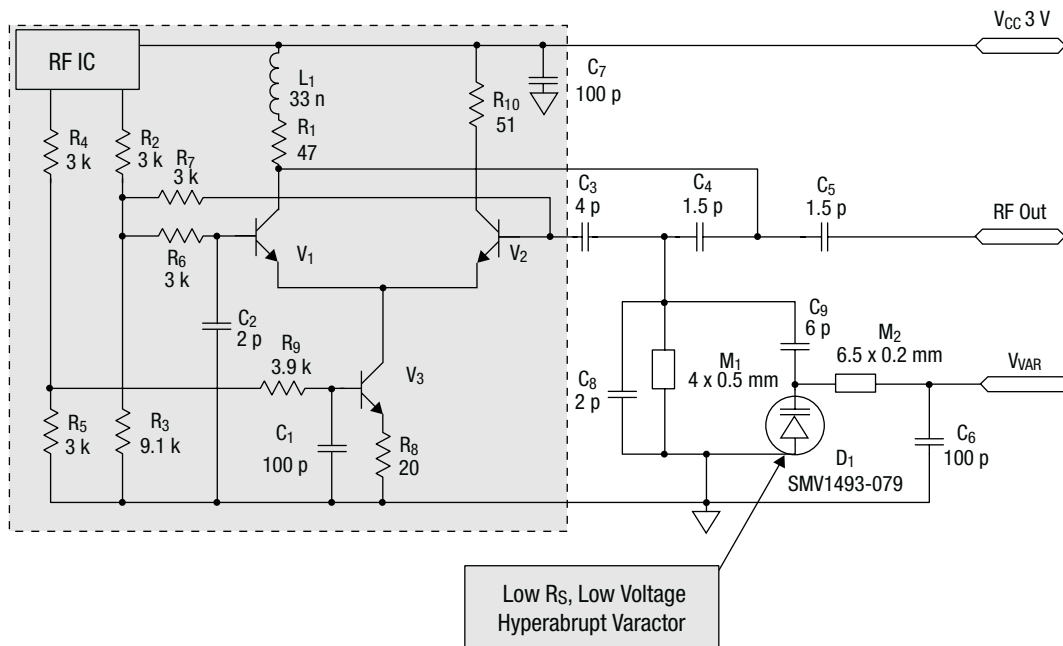


Figure 3. Differential VCO for the Integration with the RF IC

## Dual-Band Switchable VCO Schematic

One way to improve design economics in the multi-VCO requirement is to employ band switching in the VCO. If the frequency switching required isn't very large (say within 20%), it may usually be realized within the same tank circuit, by switching "on" or "off" an additional capacitor or inductor. However, if the required switching is more than 30%, it becomes very difficult to satisfy both wideband and low noise requirements in a single design. One possible solution is to use two separate tank resonator circuits switched with two PIN diodes. In this case, the feedback needs to be optimized to fit both band requirements at the same

time. Thus, a trick is used — connecting a capacitor  $C_{11}$  in parallel with  $C_6$  when a lower band resonator is selected. This provides significant improvement in phase noise since  $C_6$  may then be optimized for the best performance at high band, and  $C_{11}$  at the lower band.

Another important feature of this switching scheme is that the PIN diodes are not in the resonator current path. Because of this, phase noise is not sensitive to the PIN diode resistance. This is fortunate, since it means less forward current is needed. In addition, any noise on the PIN diode bias current (common for the noisy digital environment of today's phones) would not cause significant modulation noise.

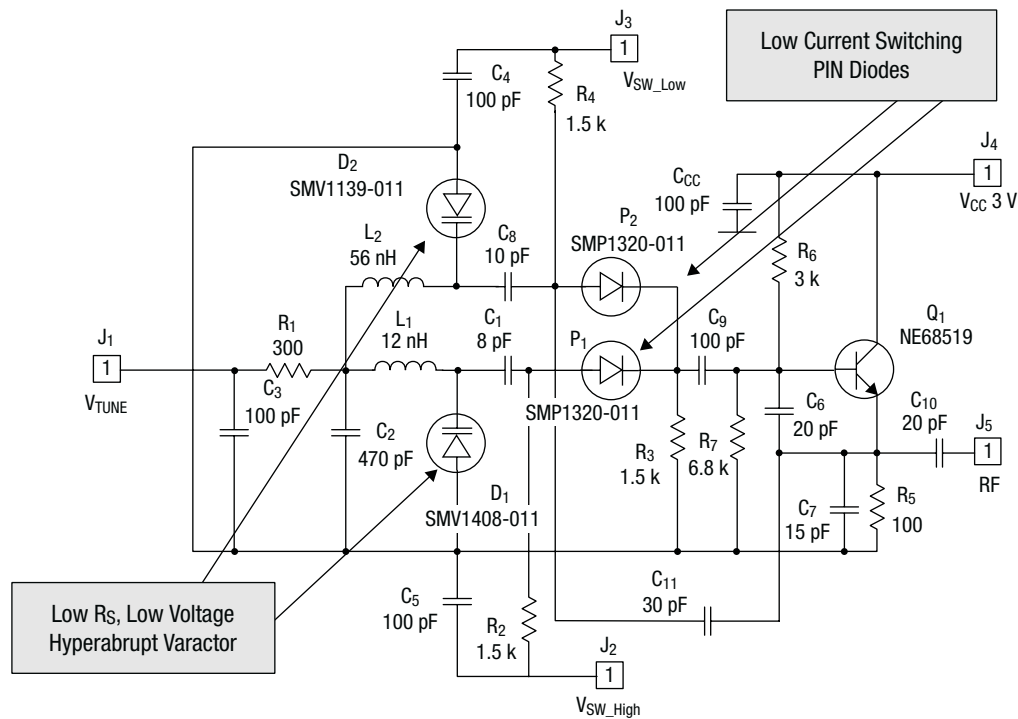
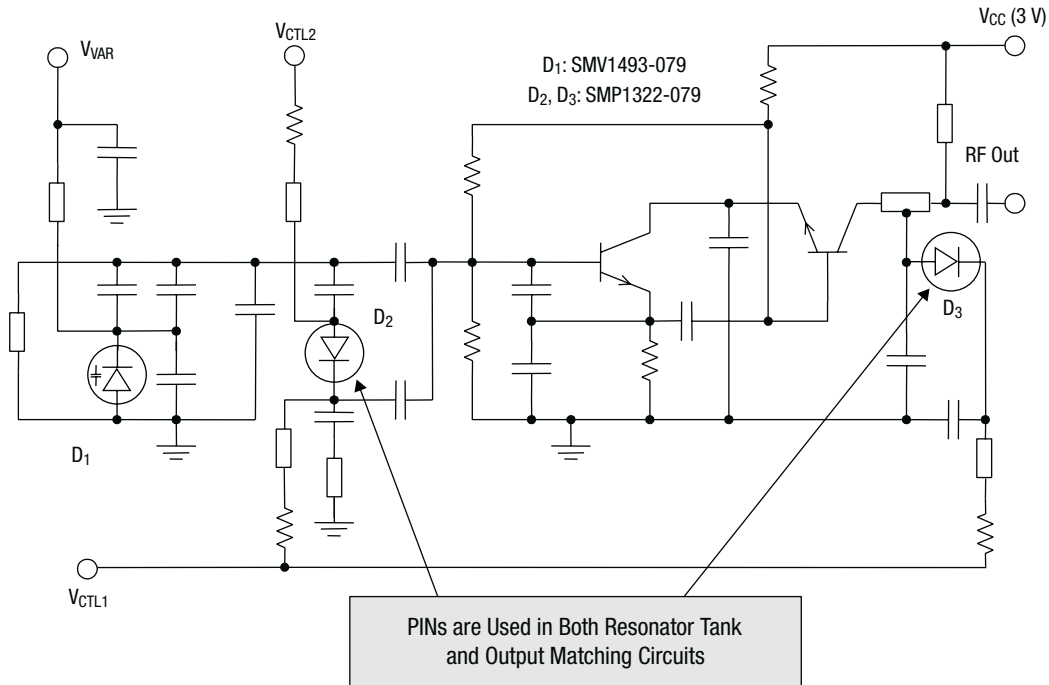


Figure 4. Dual-Band Switchable VCO Schematic

### Dual-Band Switchable RF VCO

As mentioned before, relatively small (less than 20%) frequency switching may be achieved inside the same tank circuit by connecting or disconnecting capacitors (and sometimes inductors). The PIN diode  $D_2$  performs a tricky task — it adds more capacitance in parallel with the existing parallel capacitance of the resonator, and also adds more capacitance in parallel with the

existing series capacitor. This technique is used to overcome the problem of increased resonator Q, when connecting additional parallel capacitance, by decreasing it with higher series capacitance. It allows  $D_2$  to keep phase noise near its optimum at both bands. Another PIN diode in the output matching circuit tunes the buffer to a frequency doubler mode when working in PCS band.



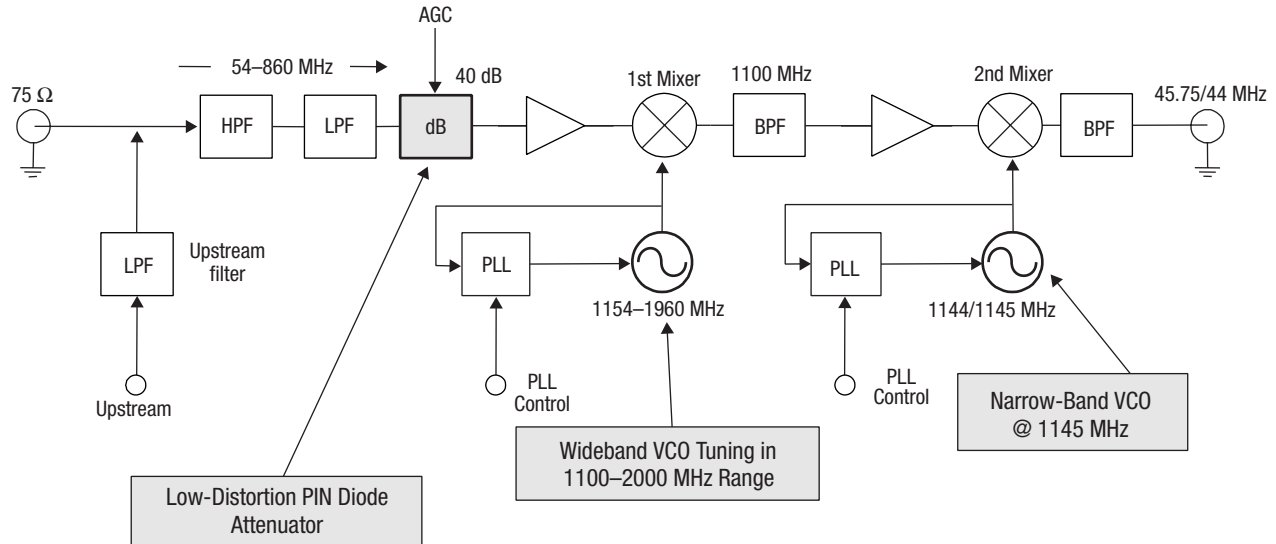
**Figure 5. Dual-Band Switchable RF VCO**

## VCOs in a Set-Top Cable Downconverter

The typical set-top downconverter is a dual-conversion receiver employing upconversion and downconversion techniques to overcome image problems in a wideband environment of 50–1000 MHz. In the dual up/downconversion scheme, the problem of image channel and input filtering virtually does not exist because there is no signal at the image channel. The image channel is always higher than the highest frequency of the cable signal.

Two RF VCOs are required for dual downconversion. The first is a wideband VCO tuned from 1100–2000 MHz with a control voltage from 1–20 V. The other is a narrow-band VCO, which may use a CDR, coaxial dielectric resonator, at 1144 MHz. In a digital system, the second IF signal may be further demodulated, requiring an additional 44 MHz VCO.

The specific action of the wideband VCO is its wideband tuning requirement. Let us consider some possible solutions for the wideband VCO.



**Figure 6. VCOs in a Set-Top Cable Down-Converter**

### Wideband Colpitts VCO Schematic

The unique action of the wideband Colpitts VCO is in its tank circuit design, which uses an inductor with a varactor connected in series and no parallel capacitor, in contrast to the low noise Colpitts VCO described in Figure 7. The feedback capacitors are optimized to the best power flatness over the entire frequency

band. Back-to-back varactors are often used to minimize parasitic mounting capacitance (between mounting pads and adjoining components). This circuit is usually designed to minimize any parasitic parallel capacitance that may be caused by component pads or transmission lines close to the inductive path.

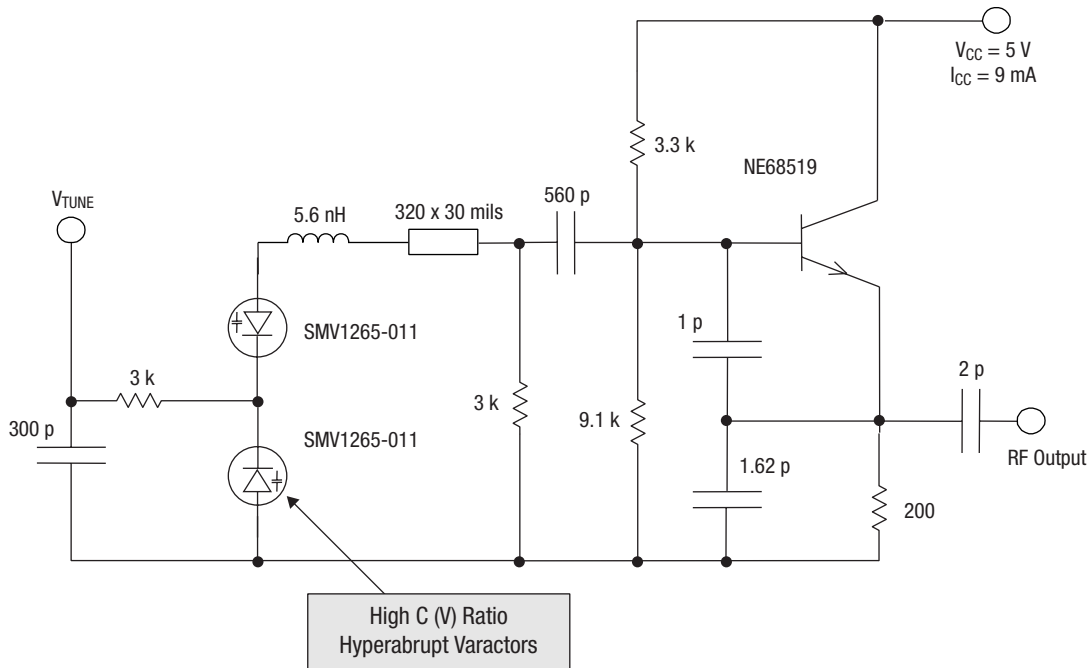


Figure 7. Wideband Colpitts VCO Schematic

A carefully designed layout with minimum parasitic capacitances may show large frequency coverage, for example 980–2120 MHz as the performance indicates. The varactor selection is a crucial

part of the design. Skyworks new SMV1265-011 varactor is specifically designed to fit this wideband application.

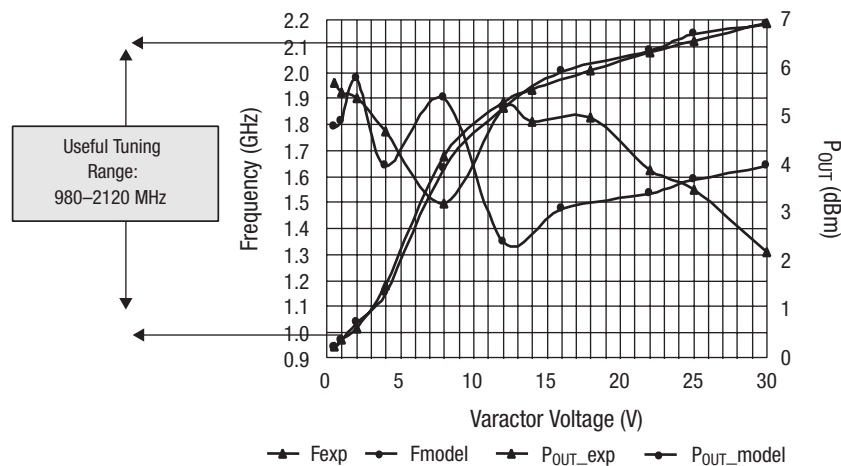


Figure 8. Wideband Colpitts VCO Performance

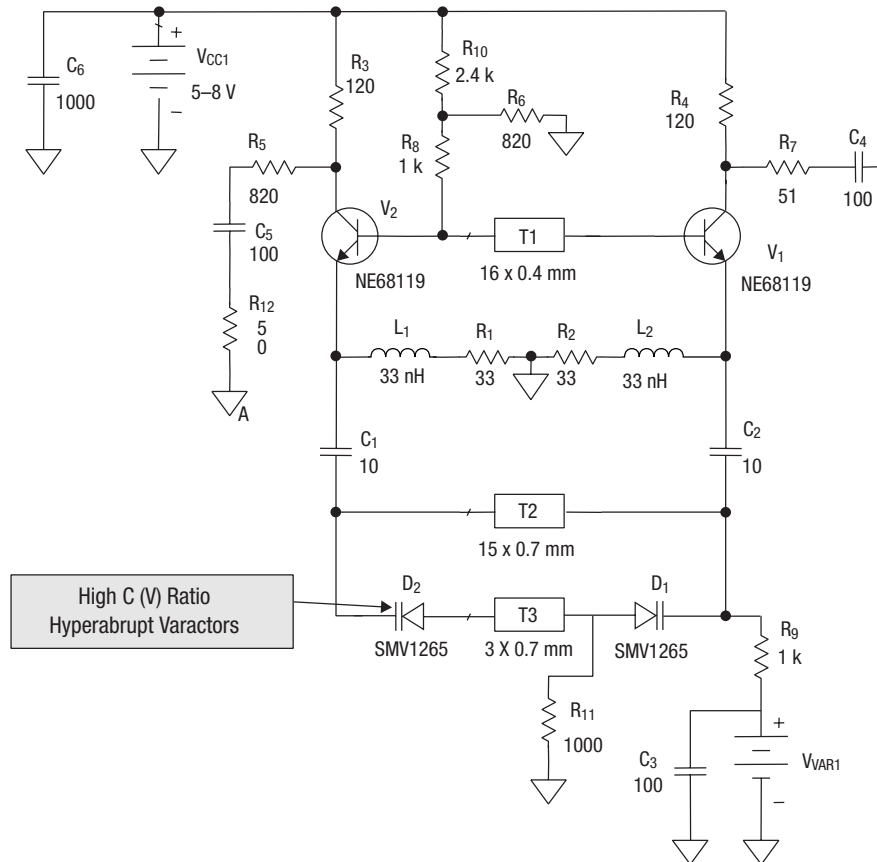


### Wideband Balanced VCO Schematic

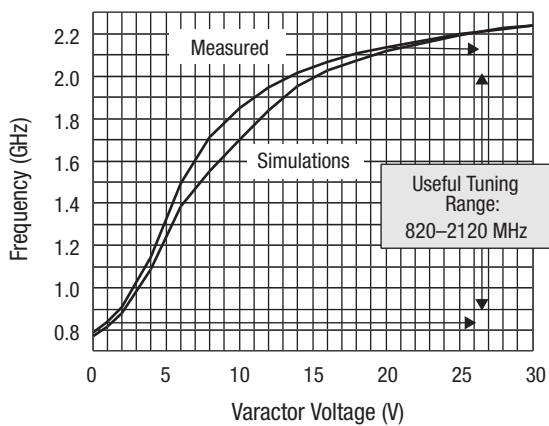
An even wider tuning range may be achieved with a balanced VCO configuration. The reason for its wider tuning performance is that the phase response of this VCO's active element is flatter

over the range of tuning compared to a Colpitts VCO. This allows the tank circuit more control over the oscillation frequency.

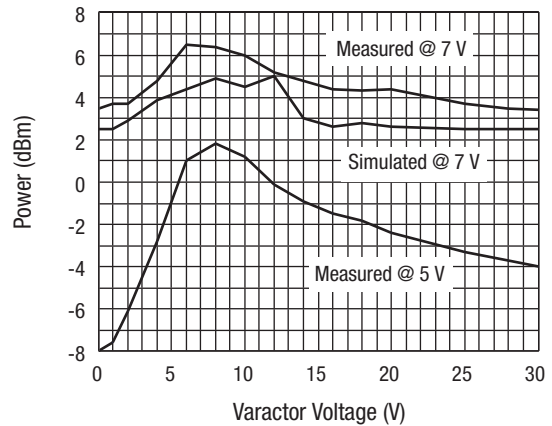
The best results are achieved with back-to-back connected SMV1265 varactors, where there is 820–2120 MHz coverage.



**Figure 9. Wideband Balanced VCO Schematic**



**Frequency Tuning**



**Power Response**

**Figure 10. Wideband Balanced VCO Performance**

### Varactor Fundamentals

Let us consider some fundamental properties of varactors. A varactor is a specially designed P-N junction diode, whose capacitance changes significantly in reverse bias mode. There are three important parameters characterizing varactors. The first is the capacitance ratio at two reverse voltages; this value characterizes the tuning ability of the varactor capacitance and is one of the most important parameters. The second is the value of capacitance at a given voltage. The third is the series resistance of the varactor.

The structure of the basic varactor, called an abrupt junction varactor, is shown in Figure 11. Generally, it is built as a P<sup>++</sup> - N - N<sup>+</sup> structure, using epitaxial N-growth on the N<sup>+</sup> substrate with a constant doping level in the N-region. The lower doped N-region is the active area where the electron concentration changes, depending on the reverse voltage applied between the anode and cathode of the varactor. There are certain limitations on the level of doping in the N-region, which is usually defined by the required capacitance ratio of the varactor. Because of this,

the conductance of the N-area is a major contributor to the varactor's series resistance. Note that as the reverse voltage is increased, the series resistance (due to the N-area) will decrease along with the capacitance.

The hyperabrupt junction varactor has a more complicated doping profile. Because of much higher doping on the P<sup>++</sup> border, the electron concentration changes much more abruptly compared to an abrupt junction. As a result, the capacitance of the hyperabrupt diode at zero bias is much higher than for the abrupt diode. Therefore, the capacitance change vs. reverse bias becomes significantly higher for hyperabrupt diodes. The trade-off for this better capacitance ratio is increased series resistance. The reason is that the doping level of the N-area has been reduced to keep average doping level over the N-region the same as the abrupt diode level. There are many ways to bring the series resistance in the hyperabrupt diode to as low a level as possible. Modern state-of-the-art hyperabrupt diodes for low noise VCOs have series resistance almost as low as discrete ceramic capacitors.

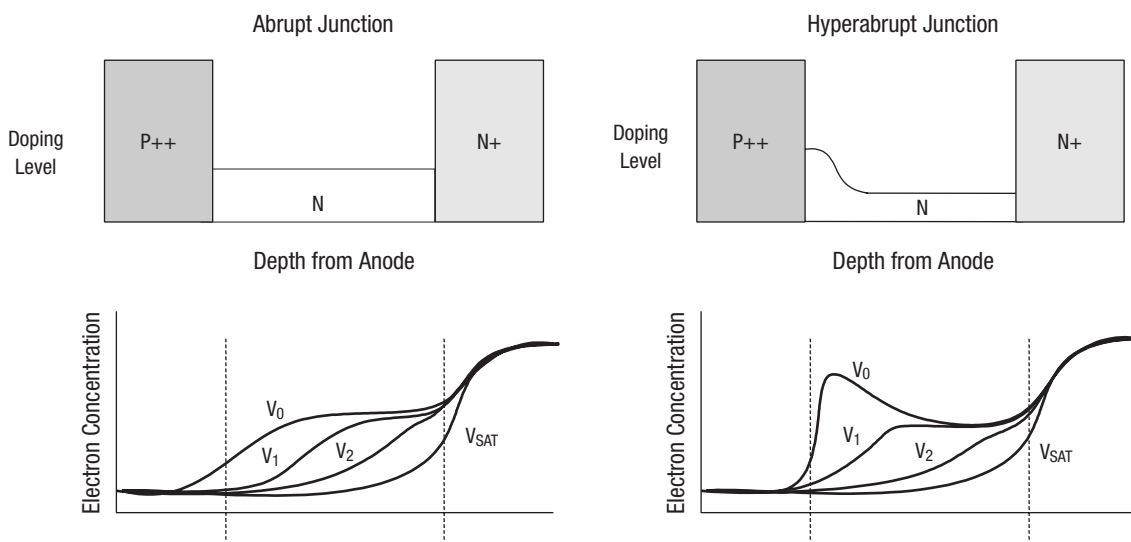
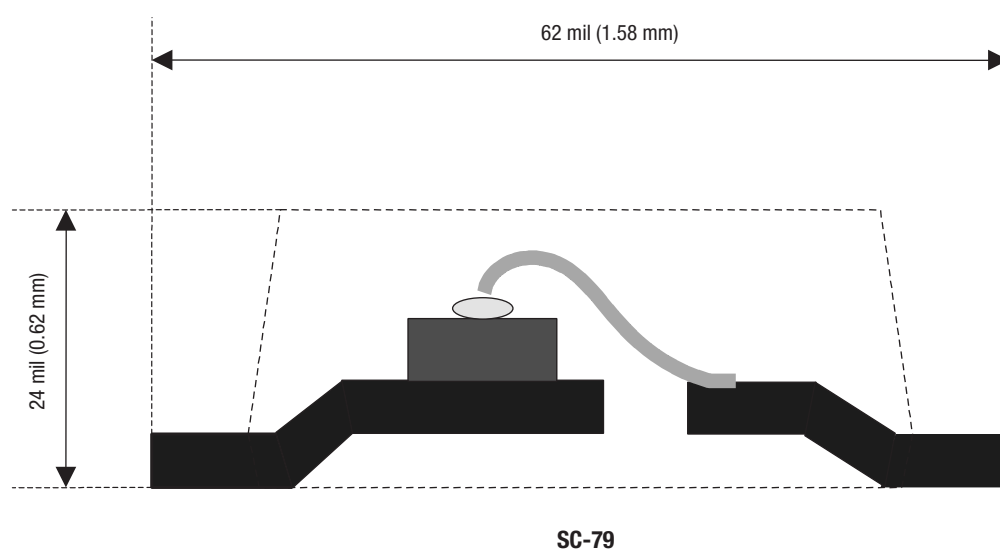


Figure 11. Varactor Fundamentals

## Varactor Packaging

Most high-volume discrete applications require varactors in low-cost, small surface mount plastic packages. Skyworks provides a

large variety of both plastic and ceramic packages. The recent, most advanced, miniature plastic package, SC-79, shown in Figure 12, is as small as 0402 discrete components.

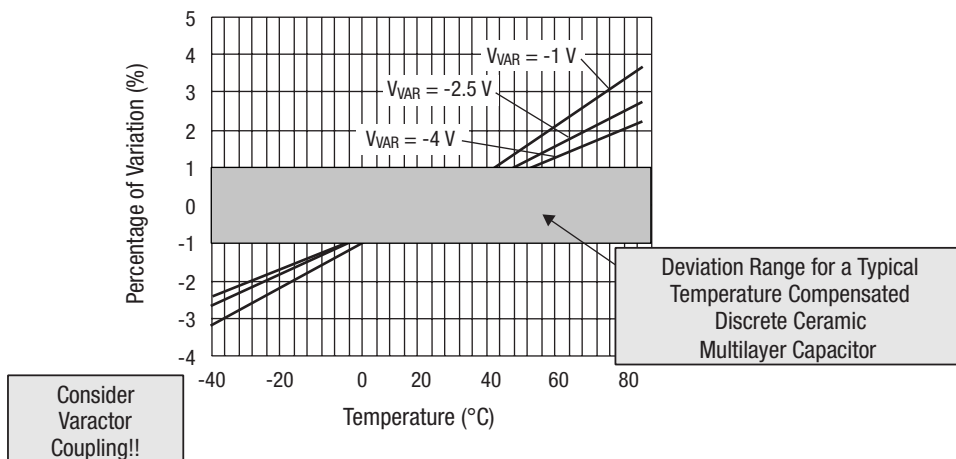


**Figure 12. Varactor Packaging**

### Relative Capacitance Change vs. Temperature

Figure 13 shows typical relative capacitance variations vs. temperature for different reverse voltages. It indicates a total capacitance change of 5–6% in the range of -40°C to +80°C. In comparison, a temperature compensated, ceramic capacitor

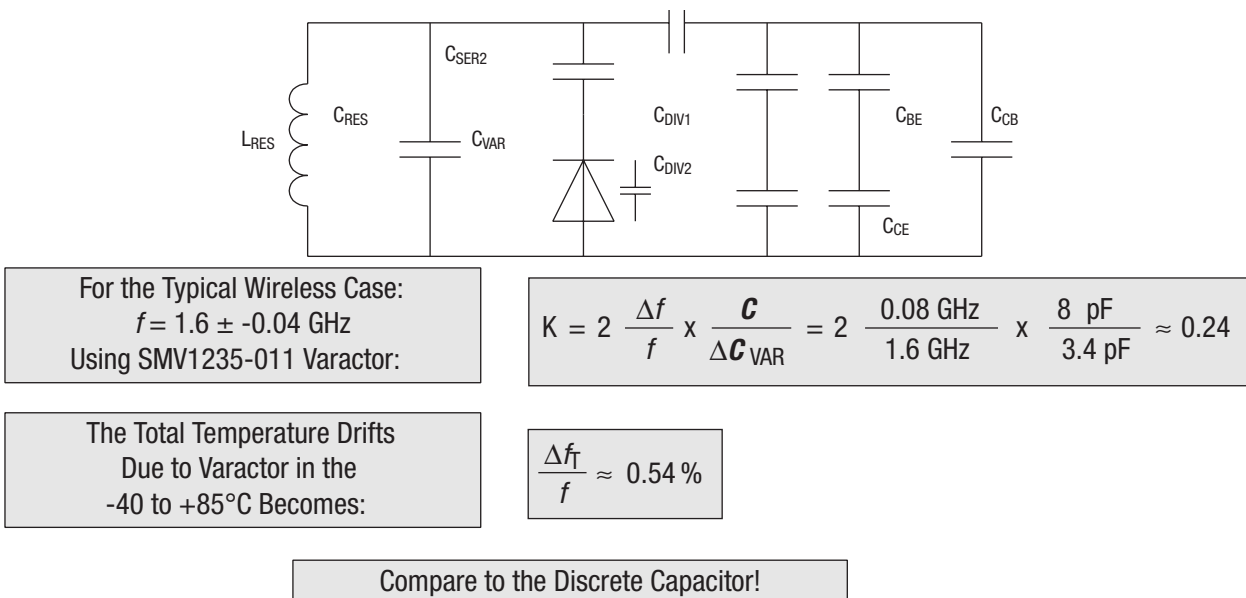
residual variation bar is shown for a typical ±100 ppm device. This has a possible total capacitance change of over 2%. When comparing the overall effect of temperature on varactors and ceramic capacitors, the coupling of the devices to the resonator circuit should be considered.



**Figure 13. Relative Capacitance Change vs. Temperature for Hyperabrupt Varactors**

The coupling coefficient may be derived from the known (or typical) values of the tuning frequency and varactor capacitance variation. Note that the total temperature drift in this case is about 0.5%, as compared to 1% maximum drift caused by tem-

perature compensated discrete ceramic capacitors. Even those numbers are extremely small when compared to the temperature drifts caused by a VCO transistor.



**Figure 14. Varactor Temperature Effect on the Oscillation Frequency**

### Varactor SPICE Model

To model a varactor in most commercial simulators, we recommend the available PN-junction diode SPICE model. We specify the barrier junction capacitance parameters  $C_{G0}$ ,  $V_J$ , and  $M$ , instead of the default parameters. In addition, we add a value of  $C_p$ , in parallel with the junction capacitor, which is not the

package capacitance. For ideal abrupt junction varactors, the parameters are constant and may be defined from physical theory. However, for actual abrupt or hyperabrupt varactors, these values are not constant. In these cases, we use the same equation, to fit its parameters, for the best compliance with measured capacitance vs. voltage response.

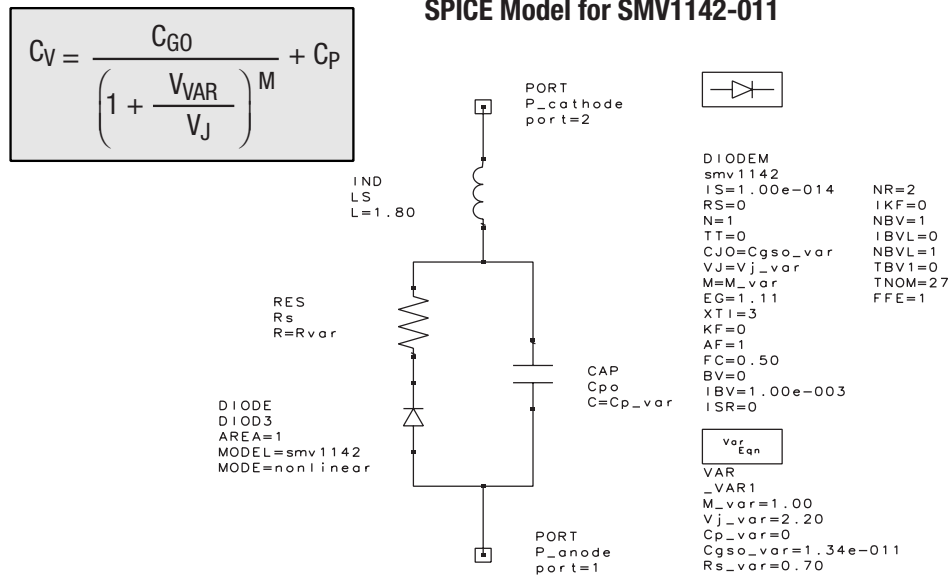


Figure 15. Typical Varactor SPICE Model

Because of formalization, parameters describing the junction capacitance of hyperabrupt varactors may be significantly different from the default values used in the SPICE simulators for the ideal silicon PN-junction.

For example, typical hyperabrupt varactor SMV1235 was fitted with  $M = 4$  as opposed to 0.5, which follows silicon PN diode

theory. Note that some SPICE simulators offer fixed default values of  $M = 0.5$  which can't be changed. In this case, a diode model may not be used, however, direct nonlinear capacitance may be used as defined in the given formula.

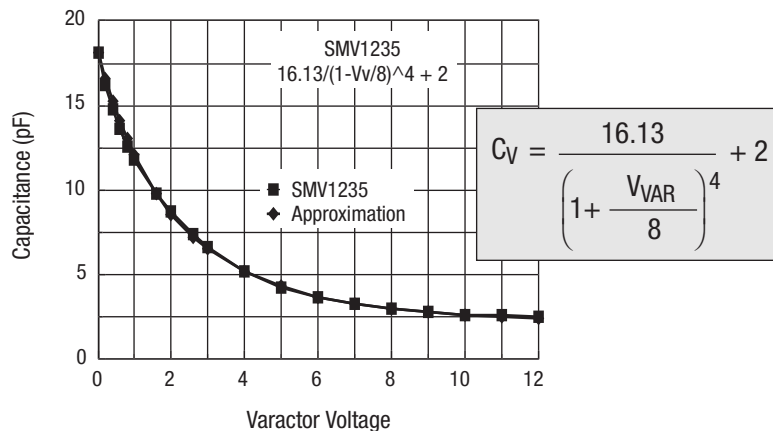
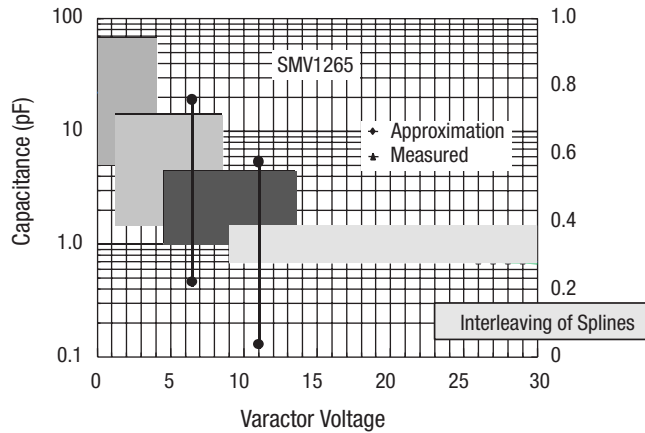


Figure 16. C (V) Curve Fitting for Typical Hyperabrupt Varactor

### Super-hyperabrupt Varactor Modeling

To overcome limitations of the “standard” PN-junction SPICE model for hyperabrupt and super-hyperabrupt devices, such as the SMV1265, an interleaving technique is used. In this technique, the entire capacitance reverse voltage range is broken into several subranges. These subranges are small enough for the

formula to provide good approximation, not only within a given subrange, but also for certain extensions beyond it. The extension margin is defined from previously estimated RF signal amplitude. Such interleaving ensures that the formula would work well, not only in terms of DC bias, but for large signal RF analysis as well.



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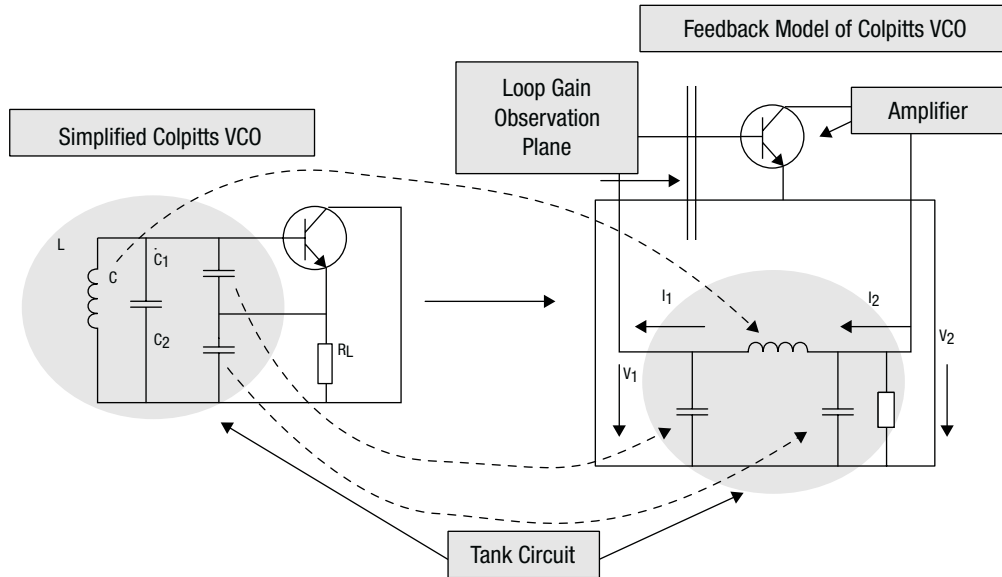
M = pwl (VVAR 0 2 2.5 2 2.500009 25 6.5 25 6.50009 7.3 11 7.3 11.0009 1.8 40 1.8)
VJ = pwl (VVAR 0 4 2.5 4 2.500009 68 6.5 68 6.50009 14 11 14 11.0009 1.85 40 1.85)
CP = pwl (VVAR 0 0 2.5 0 2.500009 0 6.5 0 6.50009 0.9 11 0.9 11.0009 0.56 40 0.56)
CJO = pwl (VVAR 0 22.5 2.5 22.5 2.500009 21 6.5 21 6.50009 20 11 20 11.0009 20 40 20)*10
RS = pwl (VVAR 0 2.4 3 2.4 4 2.3 5 2.2 6 2 7 1.85 8 1.76 9 1.7 10 1.65 11 1.61 12 1.5 40 1.5)
    
```

**Figure 17. Piece-Wise Curve Fitting for High C (V) Ratio Varactors**

### VCO Modeling Concept

For the purpose of modeling and analysis, a VCO design may be simulated as an amplifier with parallel feedback. This analysis involves measuring loop gain using a specific idealized direc-

tional coupler called OSCTEST in Libra IV. (For Harmonica users there is an application note showing how to implement OSCTEST function using S-parameters file. Refer to your Harmonica vendor for more information).



**Figure 18. VCO Modeling Concept**

The major goal of the large signal, open loop VCO analysis is to observe the magnitude (defined in dB) and the phase of the open loop voltage gain  $K_u$ , to identify particular features of the designed VCO.

First, we need to establish the optimum conditions for the oscillations in a given tuning range. Second, we need to find out

whether there are possibilities for parasitic oscillations both in the lower and higher frequency ranges. If there are parasitic oscillations, some preventive measures should be taken. Third, we need to find ways to make both  $Q_L$  and the loop power ( $P_{IN}$ ) as high as possible to facilitate phase noise performance. Finally, other features of the VCO need to be addressed, among them load pulling and  $V_{CC}$  pushing.

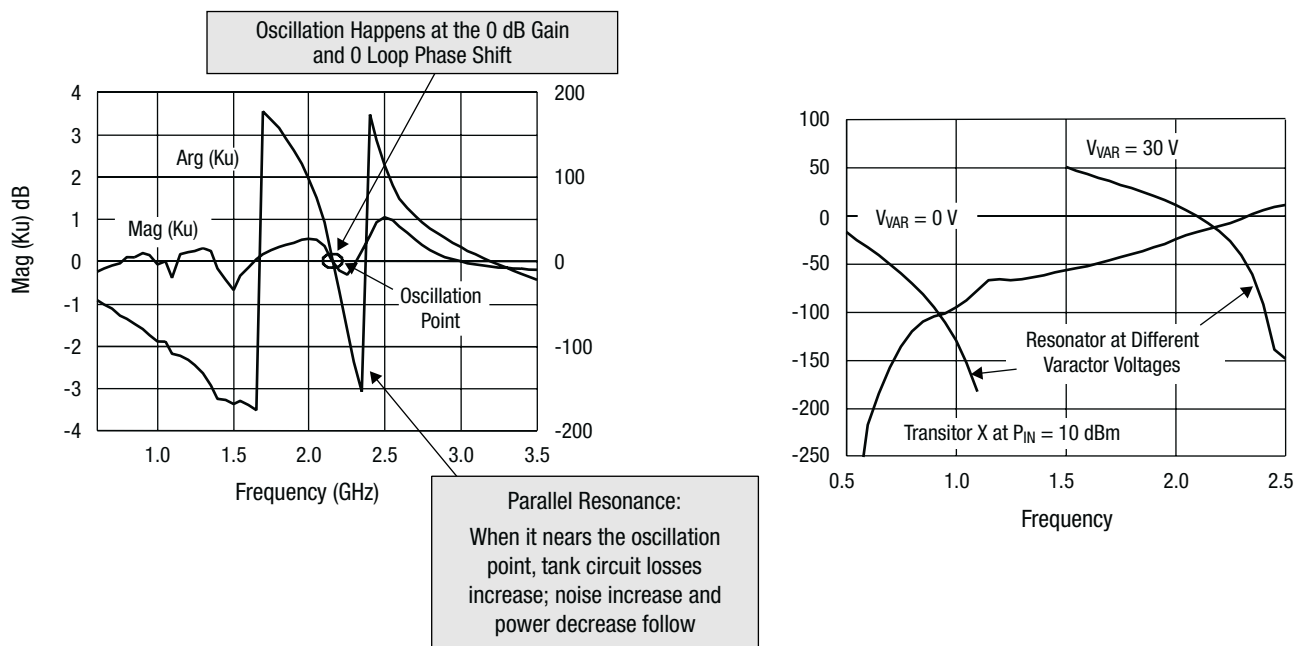


Figure 19. Typical Loop Gain Results for the Colpitts VCO



### Wideband Colpitts VCO Model

The OSCTEST component interrupts the oscillator feedback, allowing the designer to analyze the VCO as an ordinary two-port circuit (amplifier). To observe the loop response, we define the open loop voltage gain Ku. For more details, please refer to the

VCO application notes listed in the References section. The varactor model is defined as a PN-junction diode SPICE model for large signal, harmonic balance analysis. The transistor is described by the Gummel Poon SPICE model with parameters provided by the vendor.

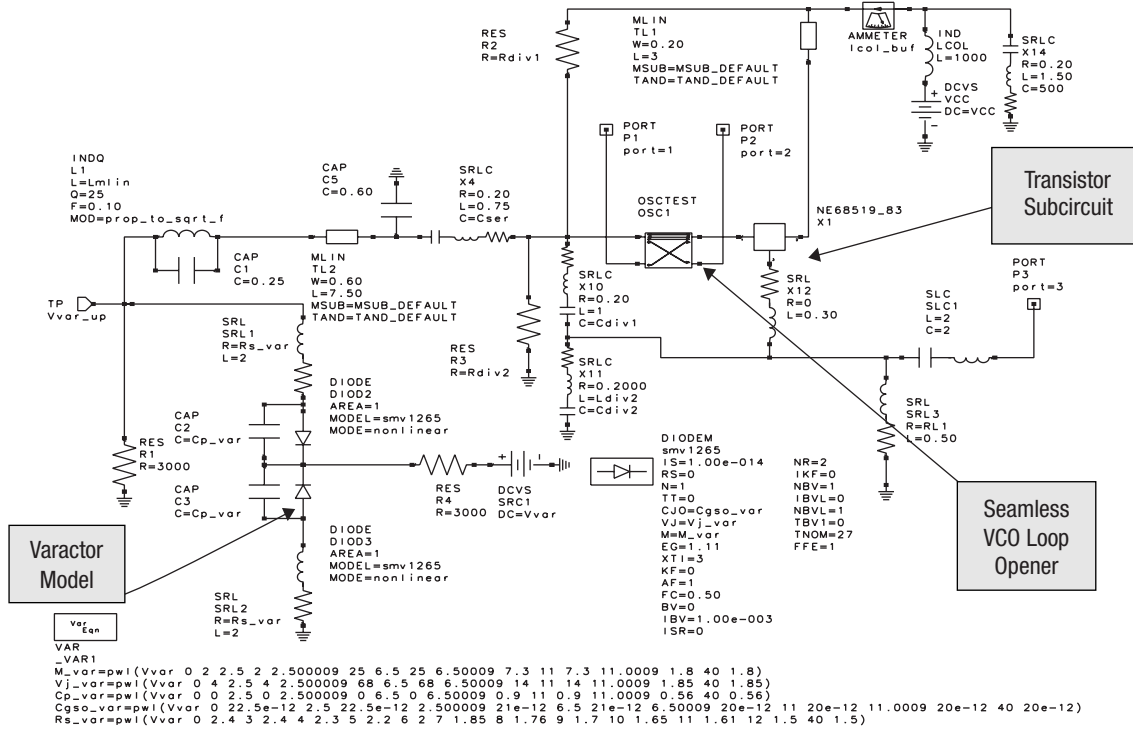
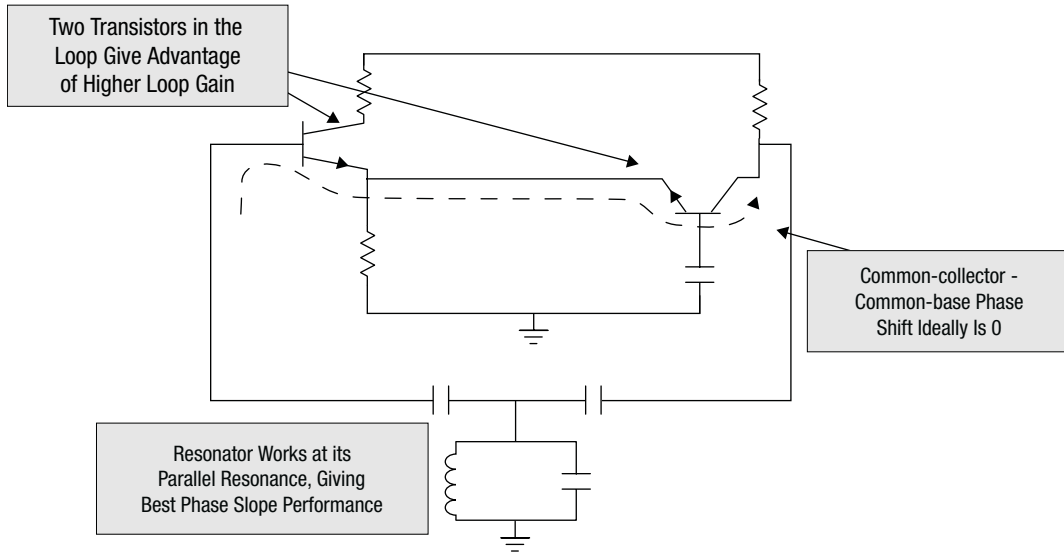


Figure 20. Wideband Colpitts VCO Model

### Differential VCO Fundamentals

The differential VCO utilizes paired transistors in common emitter and common base configurations. The phase balance condition for sustaining oscillations requires significantly lower phase shift in comparison to a Colpitts design (ideally 0 degrees vs. 180

degrees). This makes it possible to use a resonator tuned to the exact resonant frequency. However, the feedback losses may be higher because the higher resonating currents will cause increased ohmic losses.



**Figure 21. Concept of Differential VCO**

## Balanced VCO Fundamentals

Fundamental properties of the balanced VCO are more clearly understood using the simplified circuit diagram shown in Figure 22. The transistors are in common collector configuration. This is characterized by high input impedance, looking from the transmission line and referenced as  $L_B$ . Capacitor  $C_{BP}$  simulates the transmission lines and the grounding effect of the mounting pads. Coupling current  $I_{CPL}$  circulates between the transistor

bases to drive them with a  $180^\circ$  phase shift. The emitter current  $I_{FB}$  forms the feedback loop, carrying an amplified energy surplus that is needed to sustain resonant current  $I_{RES}$  and coupling current  $I_{CPL}$  through the emitter base path. Unlike a Colpitts VCO, this circuit does not require frequency dependent feedback to match the internal transistor, high frequency phase shifts. When properly compensated for wideband performance with interbase inductor,  $L_B$ , this circuit will be more broadband than a Colpitts VCO.

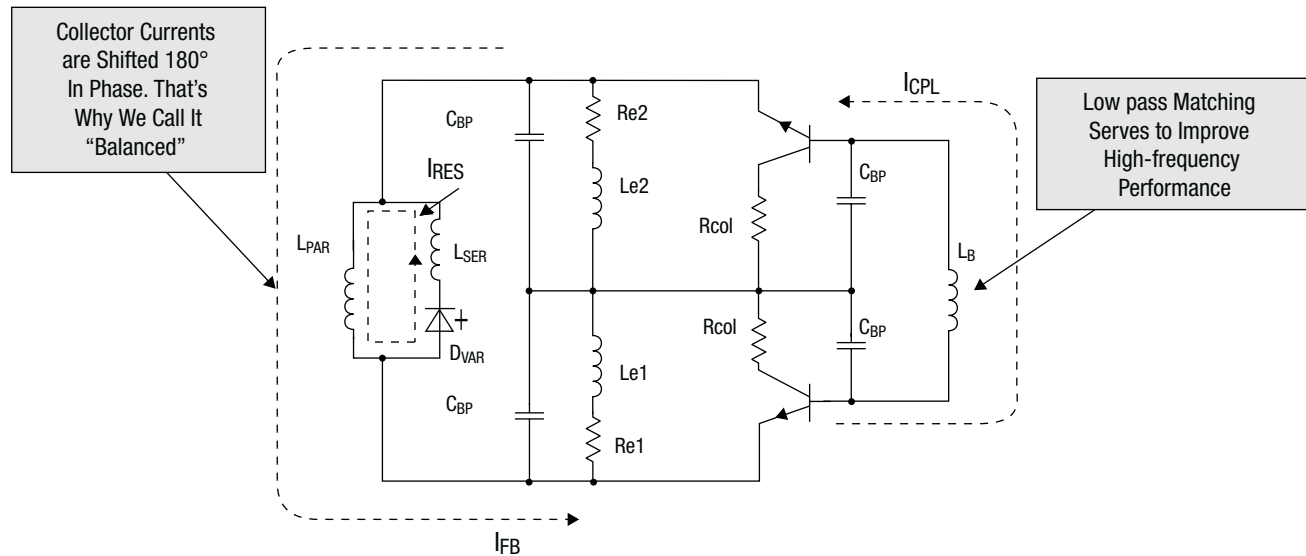


Figure 22. Balanced VCO Fundamentals

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"Wideband VCO for Set-Top Applications." Microwave Journal, April 1999.

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