An Ultra Low Cost, Low Phase Noise VCO

A easy-to-manufacture capacitively-loaded microstrip resonating element is the key to this economical design

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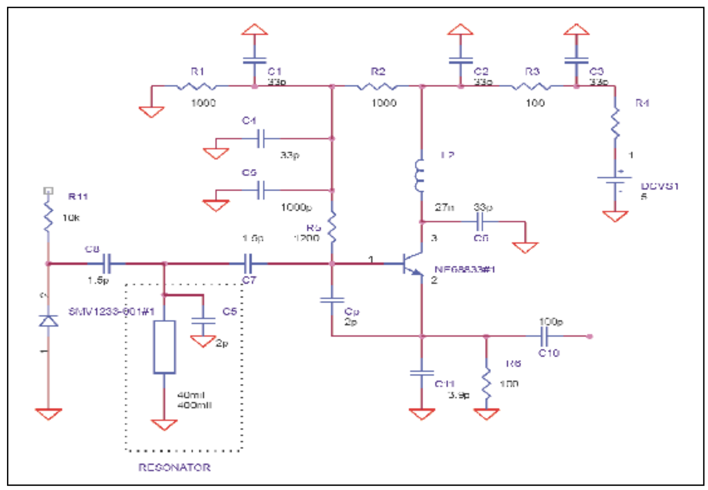
The ability to achieve good phase noise performance is paramount in most wireless design. Performance specifications that require adjacent channel rejection as well as transmitter signal purity are dependent on the phase noise of the receiver local oscillator and/or transmit local oscillator.

Oscillators capable of achieving these goals are becoming more available, but cost is still a factor when integrating off-the-shelf VCOs into any design. This article describes a method of achieving low phase noise at extremely low cost. The heart of the design is a resonating structure that can be employed directly on low cost printed circuit material and easily tuned to the desired frequency in a high volume production environment.



▲ **Figure 1. The resonating structure used in the oscillator keeps cost down while achieving low phase noise performance essential to wireless design.**

Both linear and nonlinear analysis of this oscillator has been accomplished using Xpedion's new simulator, Golden Gate. Using Golden Gate as a design tool, the theory of this design can be understood and the resulting performance ultimately verified with laboratory measurements.

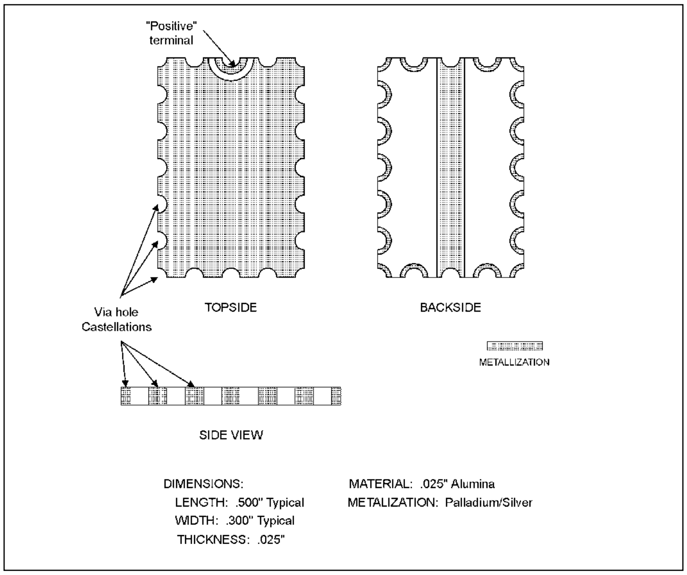
Basis for design

The basis for this design came from conventional VCO designs employing coaxial dielectric resonators, which offer advantages in terms of size, high unloaded Q and precise resonant frequency. However, these resonators are expensive compared to the other passive and active components in a VCO, making overall cost high. In addition, process variations from one VCO lot to another usually require some form of trimming of the resonator to achieve the proper center frequency. This process (grinding or laser trimming) can affect the resonator Q, lowering both phase noise and frequency stability.

Although this VCO uses essentially the same active circuit topology as a conventional VCO, it offers a significant cost advantage. In place of the coaxial resonator, an alternative resonating structure is used. This structure is quite simple, but when embedded properly can provide good performance in phase noise, tunability and frequency stability.

Basic oscillator topology

The basic circuit topology for this oscillator is shown in Figure 1. This configuration employs a grounded collector of the active device (NE68833), creating a negative resistance port



▲ **Figure 2. Illustration of the geometry of the resonator of the low-cost oscillator.**

at the base of Q1. The resonator (dotted area) sees this negative resistance and oscillations occur near its resonant frequency.

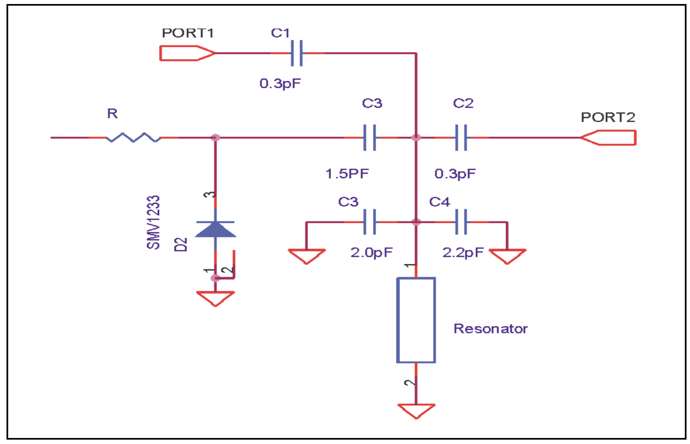
Resonating structure detail

The low cost resonating structure comprises three constituents:

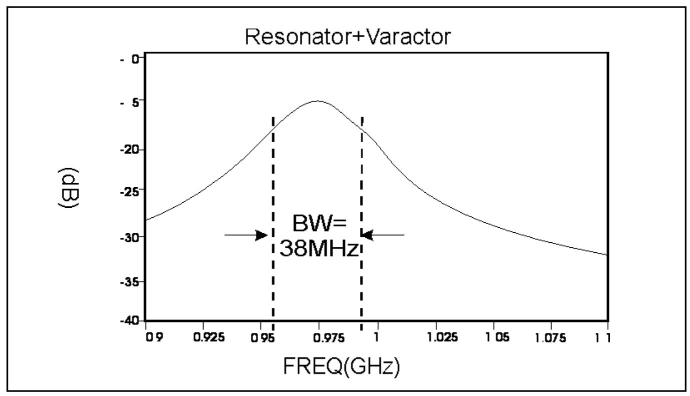
1. A short, low impedance transmission line (microstrip) on the PCB
2. A low cost thick film SMT alumina tuning structure covering the microstrip
3. A high-Q loading capacitor in shunt with the transmission line.

As noted by Rhea [4], the use of a short transmission line and high Q loading capacitor can produce a Q of several hundred, even on low cost FR-4. The unloaded Q of this structure was found to be around 60, the measurements for which will be described.

Figure 2 illustrates the geometry of the resonating structure. The transmission is a conventional microstrip line, in this case approximately 500 mils long and 60 mils wide. This line is fabricated on standard FR-4, in this case 62 mils in thickness. A unique part of the resonating structure is the alumina tuning structure, fabricated on low cost thick film alumina. The structure is designed so that it can be soldered directly on the transmission line to become a tunable element. The entire top of the alumina structure is grounded using a series of via holes along three of the four sides of the tuning slab. Two high Q loading capacitors are mounted directly adjacent to the "hot" end of the transmission line. In this case, a 2.0 pF and 2.2 pF capacitor were used.



▲ **Figure 3a. Loaded *Q* measurement setup of resonator and varactor.**



▲ **Figure 3b. Frequency response of the resonator and varactor using the test circuit shown in Figure 3a.**

The unloaded Q of the resonating structure was calculated using the relationship

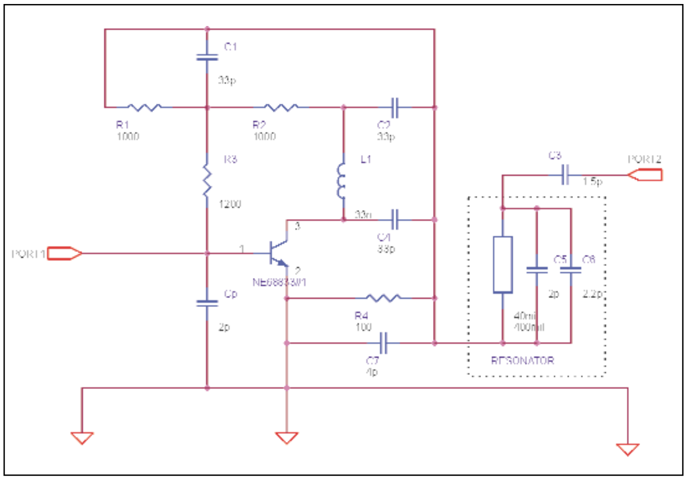
$$IL = -20\log\left(\frac{Q_u - Q_l}{Q_u}\right) \tag{1}$$

which is also from Rhea [4], as well as the fundamental relationship [3]

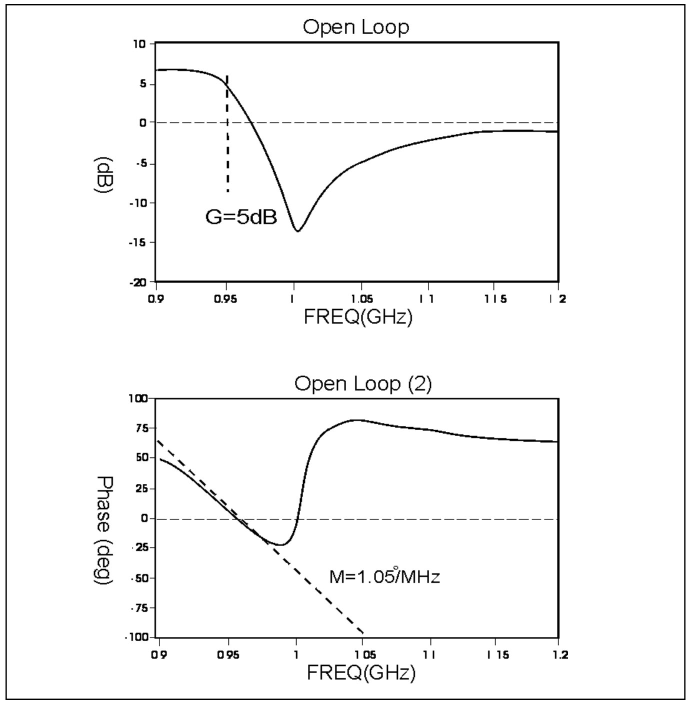
$$B = \frac{f_0}{Q} \tag{2}$$

The S-parameters of the resonator and the loading capacitors were first measured. The unloaded Q was then found by simulating the following test circuit containing the measured S-parameters of the resonator. By keeping C1 and C2 small, the loading effects of the 50 ohm test ports will be negligible. The unloaded Q will be

$$Q_u = \frac{Q_l}{1 - 10^{\frac{S_{21}}{20}}} \tag{3}$$



▲ **Figure 4. Analysis of circuit for gain and phase to determine potential for oscillation.**



▲ **Figure 5. Results of the analysis of the circuit in Figure 4.**

For the resonator alone, the center frequency was found to be 1103 MHz, the 3 dB bandwidth approximately 33 MHz, and the insertion loss 10 dB, resulting in an unloaded Q of 50.27.

With the addition of the tuning varactor (Alpha SMV1233-001) and series coupling cap (1.5 pF) some additional losses are introduced. At 0 volts bias, the center frequency dropped to 975 MHz and the insertion loss of the test network increased to 15 dB. This resulted in an unloaded Q value of approximately 31.2 for the resonator-varactor combination. Figures 3a and 3b show the S_{21} response of the resonator and varactor.

Open loop oscillator analysis

As well known from oscillator theory, two conditions are required to make a feedback system oscillate: the open loop gain must be greater than unity; and total phase shift must be 360 degrees.

Although not obvious, this circuit can be analyzed using this approach by the method of Harada [2]. This can be done by reconfiguring the original circuit to have a virtual ground at the emitter of Q_1, then breaking the connection at the base of Q_1 (the feedback path) to obtain a two port circuit for gain and phase analysis. The resulting circuit is shown in Figure 4. Using the linear S-parameter analysis, we can examine the oscillator's open loop transfer characteristic as a two port.

Figure 5 shows the results of this analysis (gain and phase). The phase crossover (frequency of oscillation) is approximately 960 MHz; at this same point, the open loop gain is approximately 4 dB, indicating potential for oscillation. The loaded Q of this circuit can be found by the relationship

$$Q_l = (\pi f_0/360) \times (d\phi/df) \qquad (4)$$

This can be estimated by taking the slope of the phase near the zero crossover, which is approximately 1.05 degrees per MHz at this point. Using Equation 4, the resulting loaded Q for the oscillator is approximately 9.94. It is important to note that the loaded Q for the oscillator is significantly lower than the unloaded Q for the resonator alone (due to losses in the active device). The loaded Q has an important bearing on phase noise, as will be discussed in the next section.

Making the loading capacitor larger will improve the loaded Q for this network, but reduce tunability. For this case, a 2.0 pF and 2.2 pF capacitor in parallel with the resonator was found to give good phase noise plus reasonable tuning bandwidth.

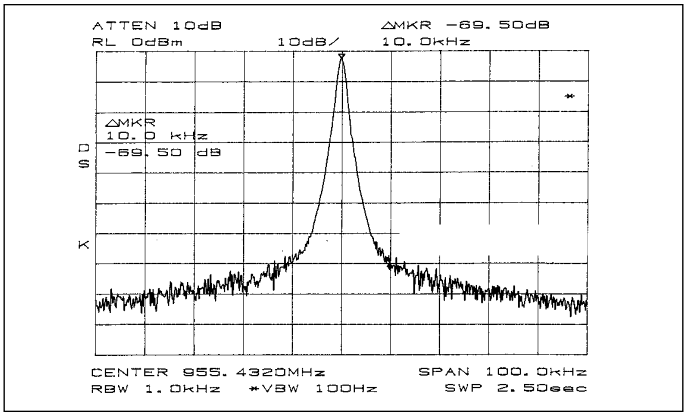
Phase noise prediction

Leeson's equation [6] describes the effect of loaded Q and other parameters on phase noise

$$L(f_m) = 10\log\left[\frac{1}{2}\left[\left(\frac{f_0}{2Q_l f_m}\right)^2 + 1\right]\left(\frac{f_c}{f_m}+1\right)\left(\frac{FkT}{P_s}\right)\right] \qquad (5)$$

$L(f_m)$ is a measure of the phase noise in dBm/Hz from the carrier. We see that $L(f)$ is proportional to $1/Q_l^2$. Other factors enter the equation, including

1. f_0 — the frequency of oscillation
2. f_c — the flicker noise corner
3. F — the numeric noise figure
4. P_s — the carrier power
5. kT — Boltzmann's constant × temperature (Kelvin)



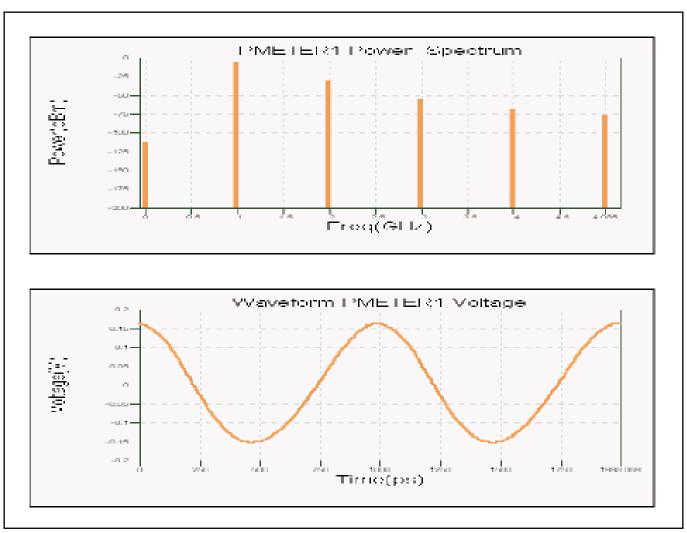
▲ **Figure 6. The design yields the expected oscillator phase noise response.**

Actual parameters used were

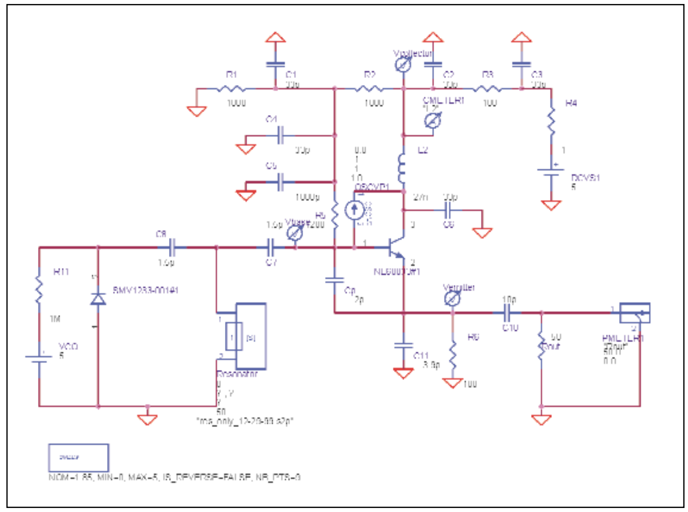
1. f_0 = 950 MHz
2. f_m = 10 kHz
3. Q_l = 9.94 (from above)
4. f_c (flicker noise corner) = 5.8 kHz (from datasheet)
5. F (noise figure) = 1.7 dB (from Xpedion analysis)
6. kT (Boltsmann's constant × temp) = 4.2×10^{-21}
7. P_s (available power) = 1 mW (estimated)

Flicker noise for the NE68833 was estimated using data from a similar device, the NE68819 [1], which has a flicker corner of approximately 5.8 kHz. The noise figure can be simulated using the open loop S-parameter analysis with noise parameters for the device. Using Xpedion's linear noise analysis, this gave a result of 1.75 dB at 950 MHz. P_s was estimated at 1 mW using measured data taken on the actual oscillator.

Inserting these values into Leeson's equation yields a resulting phase noise of –97.7 dBc/Hz at a 10 kHz offset.



▲ **Figure 8. Spectrum and voltage waveforms of oscillator response to linear noise analysis.**



▲ **Figure 7. Xpedion circuit model used for active device and varactor analysis.**

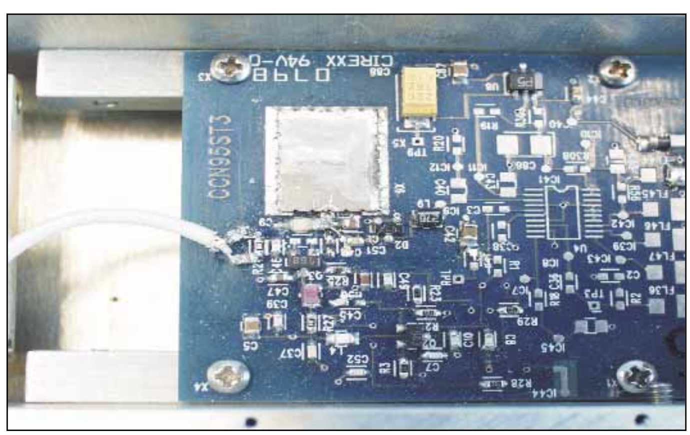
A graph of the actual oscillator response (Figure 6) indicates a phase noise of –98 dBc/Hz at 10 kHz offset, a very good correlation with the predicted result.

Nonlinear analysis

The active device (NE68833) was modeled using Xpedion's Gummel-Poon device. Input parameters for this model were obtained from manufacturer's device data. The varactor was modeled using Xpedion's SMV1233 (Alpha) varactor model operating at a reverse bias of 0 to 5 volts. The complete Xpedion circuit model is shown in Figure 7.

The spectrum and voltage waveform of the analyzed response is shown in Figure 8. Xpedion predicts an output power of –0.8 dBm, which is within 2 dB of the measured laboratory result. Xpedion predicts the first and second harmonics at 25 dBc and 45 dBc respectively. Results from laboratory measurements were 38 dBc and 59 dBc respectively, somewhat better than the simulation predicts.

Varying the varactor voltage in Xpedion produced a



▲ **Figure 9. Photograph of the oscillator as constructed.**

frequency tuning range of 27 MHz. Laboratory results indicated a tuning range of approximately 30 MHz over the same voltage range. In the actual VCO, the center frequency was adjusted by removing thick film metal from the top of the resonating structure using a grinding tool.

Conclusion

A very low cost VCO has been developed using a low cost resonating structure. The use of a short transmission line loaded with a high *Q* capacitor and thick-film tuning slab produces low phase noise at minimal cost. Xpedion's linear/nonlinear simulator has been used to analyze this device and predict power, harmonics, tuning bandwidth and phase noise. Good correlation between predicted and measured results for frequency, power, phase noise and tuning bandwidth has been obtained. By applying the proper design techniques, this simple structure can replace an expensive coaxial resonator for a multitude of wireless applications. ■

Acknowledgments

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