USING MICROSTRIP RESONATORS TO SELECT HIGH OVERTONE MODES IN UHF CRYSTAL OSCILLATORS

Ken Hartman & Dr Darrell Newell

Electrical Engineering Technology / Still Gym Rm 205
Northern Illinois University  DeKalb, IL  60115

ABSTRACT

An oscillator, fabricated from microstrip elements, has been combined with a conventional quartz crystal resonator. The resulting circuit demonstrates markedly improved characteristics over the free running microstrip oscillator. A design approach for fabricating this type of circuit is presented along with a discussion of elemental considerations and prototyping short cuts.

A continuous expansion of communication requirements has placed an upward pressure on the operating frequency of of the stable signal sources that are the heart of these communications systems. The quartz crystal resonator continues to be the most reliable, consistent frequency controlling element for this type of equipment.

There are a variety of circuit techniques available at UHF frequencies. Oscillator designs based on these techniques are numerous. Lumped element designs can be extended from lower frequencies into the UHF region by simple scaling of the elements. The limitation of this approach centers on the transformation of the simple Ls, Cs, and Rs into networks of these elements at UHF frequencies. Simple resistors evolve into resistor/inductor networks at frequencies as low as 300 mHz. Conventional capacitors develop unacceptable loss characteristics in this frequency range. Ceramic and mica capacitors display low loss characteristics but take on other undesirable characteristics such as self-resonance, because of their physical dimensions and package configurations. Air wound inductors are generally acceptable and display a good 0 below 1000 mHz. The volumetric limitations of most designs discourages the use of these elements in UHF design.

Distributed parameter circuits such as strip-line or microstrip line have the necessary versatility at UHF but lack the intrinsic stability of alternative designs. Elemental O's are limited to values not greater than about 200. Cavity and helical cavity designs display a higher operational O but require comparatively large volumes and are generally expensive to fabricate because of exacting machining requirements.

Historically, the use of frequency multipliers has filled the requirements for stable signal sources in the UHF range. The complexity and associated noise of these multipliers makes for a less than optimum solution to the problem. Recently, techniques have been developed to extend the range of low temperature coefficient quartz resonator designs into the 100 mHz to 500 mHz region. Unfortunately, at 450 mHz, these resonators are almost unusably fragile and are prone to failure.

There is a need for a quartz crystal oscillator that uses the fabrication ease of microstrip techniques. A design approach has been developed using the high overtone modes of a quartz resonator. This oscillator design employs a 25 mHz quartz crystal element operating on the 17th overtone. A fundamental mode microstrip oscillator was fabricated at 440 mHz. A crystal was added to the circuit at a point that resulted in the tuning out of the Co part of the crystal. The resulting oscillator displays spectral characteristics of a crystal stabilized oscillator. The incorporation of the crystal was made with allowance for the lead and package parasitic elements.

Two general design approaches can be used to generate microstrip oscillator circuits. The first is the reflection type oscillator. The second is the feedback oscillator type. The design of UHF feedback oscillators begins with the selection of a suitable transistor. A multiplier can be designed that will provide a conjugate match to the transistor input and output impedances. Since the transistor is a bilateral device, the input impedance as seen looking into the transistor is affected by the value of the load impedance. In a similar fashion the output impedance is affected by the value of the input impedance. Simultaneous conjugate matching of the input and output can be accomplished using the expression

\[ K_s = \left( S_{11} + S_{12}S_2K_r \right)^* \]

where \[ |K_r| = \frac{8 + \sqrt{p_2^2 - 4|C_2|^2}}{2|C_2|} \]

217
and

\[ B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |S_{12}|^2 \]

\[ C_2 = S_{22} - (D_{11})^* \]

\[ D_2 = S_{11}S_{22} - S_{12}S_{21} \]

Since the "load" and "source" impedances are actually the feedback network, the output load must be incorporated as a portion of the matching circuit.

Sections of microstrip line of a specific impedance and length, open circuit and short circuit microstrip line elements form the basic building blocks of the distributed parameter circuit design approach. There are some limitations to these elements that the designer must be aware of. In the case of the open circuit stub we encounter the problem of fringing capacitance at the ends of the open circuit line. Although this phenomena is also present along the length of the strip, it is most pronounced at the end of the strip due to the sharp corners. These corners tend to concentrate the electric field in this region. The result of the fringing capacitance is a reduction in the actual guide wavelength as compared to the theoretical guide wavelength. The magnitude of this differential is a function of the geometry of the open circuit stub. It is difficult to precisely predict the electrical "end" of an open circuit stub.

The short circuit stub has a more subtle but equally vexing problem. The "short circuit" description mandates the establishment of an rf short at the end of the line. In most cases this is accomplished by physically shorting the line to the groundplane. In certain instances this physical grounding may not be possible and the short is accomplished using a lumped capacitor. In either case, the quality of the ground has a profound effect on the line characteristics. In the case of the physical ground, a hole (generally a slot the width of the conductor) is made in the substrate. The conductor is continued through the hole to the groundplane. A certain amount of care must be exercised to insure that the substrate edge is straight and that the conductor contact to the groundplane is made across the entire width of the line. For screened conductors on alumina this means that the substrate must be slotted, generally in the fired state, and the screening process must insure that the ink is uniform through the slot to the groundplane. From an ease of implementation point of view an open circuit stub is preferred over a short circuit stub.

The advantages of microstrip architecture are based on the ease of fabrication. The open structure planar geometry allows for the easy attachment of lumped components and active devices. Connections to the outside can easily be made using standard rf connectors or fly wire connections. The current emphasis on surface mount components is resulting in an ever increasing selection of high quality components for use at high frequencies.

The main disadvantage of the microstrip structure lies in the open construction technique. Radiation at bends and corners becomes a significant lose mechanism at very high frequencies. The guide wavelength of a particular frequency is not reduced by the square root of the dielectric constant as is the case for striplines, but is somewhat less. This is due to the fact that the dielectric surrounding the conductor is a combination of the dielectric of the substrate and air ($\epsilon_r=1.0$). The effective dielectric constant can be determined by examining the cross section area of the microstrip and assigning a "filling factor" to account for the reduction in electric field intensity due to air. Because the circuit is open in a sense to the ambient environment, it is susceptible to environmental factors such as condensation and contamination. Conformal coatings have been used in an attempt to minimize these effects. The problem with conformal coatings is due to the fact that the electric field in the microstrip is not fully contained in the substrate region. The conformal coating represents a dielectric layer over the conductor. The change in filling factor and resulting change in effective dielectric constant is manifest as a change in guide wavelength.

Two open transmission lines that are in close proximity to one another will interact to exchange energy across the gap separating them. It would be convenient to describe the gap coupling phenomena in terms of a coupling coefficient $C$. In order to do this we must first examine the coupled line structure in terms of the field vectors associated with the energy contained in the lines. Figure 4 illustrates the electric and magnetic field geometries in the coupled line case. Note how the symmetrical nature of the structure gives rise to an even an odd mode field distribution. The effective dielectric constant derived from the "filling factor" described earlier is different for the even and the odd mode. The analysis for coupled lines is based on even and odd mode velocities of propagation and, most importantly, an even mode impedance and an odd mode impedance.

A derivation of the coupling coefficient in terms of the even and odd mode propagation can be found in Edwards. The resulting expression

\[ C = \frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}} \]

where $Z_{oe}$ is the even mode impedance. $Z_{oo}$ is the odd mode impedance. which can be expressed in decibels as

\[ C_{dB} = 20 \log \left| \frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}} \right| \]

The even and odd mode characteristics can be computed from the physical dimensions of the coupled lines. This coupling coefficient is
meaningful as long as the transmission lines are terminated in their characteristic impedances.

An expression for the "equivalent" characteristic impedance is given by

$$Z_0 = Z_{oe} Z_{oo}$$

which is valid for coupling coefficients less than -3 dB.

Frequency selective networks, filters, can be generated using cascaded coupled resonators. An easily comprehensible model of the equivalent transmission line network of one such network is given below.

For narrow line widths (w/h<3.3)

$$Z_0 = \frac{119.9}{2\sqrt{\varepsilon}} \left[ \ln \left\{ \frac{4}{w} \sqrt{\frac{\varepsilon}{\varepsilon_{oo} + 1}} \right\} + \frac{1}{2\varepsilon} \ln \left\{ \frac{\varepsilon_{oo} + 1}{\varepsilon_{oo}} \right\} \right]^{-1}$$

for wide line widths.

Coupled line calculations are much more complex. There are two programs that are being used for the calculation of even and odd mode impedances. The first program was written by T. G. Bryant and J. A. Weiss and is documented in various references. This program yields an extremely accurate estimate of the coupled microstrip characteristic impedance, phase velocity, and effective dielectric constant for the even and the odd modes of a traveling TEM wave. The program is written in FORTRAN and is a computational intensive algorithm. Most current investigation into coupled microstrip lines reference this standard program.

A recent program by G.B. Tait is a much simpler algorithm and is accurate to within a few percent of Bryant & Weiss. Both programs use the relative dielectric constant of the substrate, the width to height ratio, and the strip space to height ratio as the basis for the calculations. The Tait program also uses the thickness of the conductor as an input. Computer aided engineering programs such as COMPACT and EEsoft have been developed specifically for distributed parameter design and analysis. Super-Compact and the Compact family is probably the most popular of the microwave CAD/CAE programs.

Quartz crystal elements used in overtone operating modes are well understood. The bandwidth of the crystal is given by

$$BW = f_s - f_n$$

for the fundamental mode and as

$$BW = f_{s(n)}$$

for the overtone mode n

$$2(C_m/C_n(n))$$

since

$$C_m = C_m f / \pi^2$$

we can say

$$BW = \frac{BW}{n^2}$$
A crystal operating on the 17th overtone will have a reduction of bandwidth by a factor of approximately \(2891\). Table 1 shows the overtone number, the bandwidth reduction factor, and the resulting bandwidth for a nominal 20,000 Hz bandwidth of the fundamental mode.

<table>
<thead>
<tr>
<th>MODE</th>
<th>BW factor</th>
<th>BANDWIDTH</th>
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<tbody>
<tr>
<td>1</td>
<td>1</td>
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<tr>
<td>3</td>
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<td>5</td>
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<td>800</td>
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<td>7</td>
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<td>11</td>
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<td>13</td>
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<td>25</td>
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</table>

The dramatic bandwidth reduction of the high overtone operation mode would suggest that the final plating of the crystal be done in situ in order to place the oscillator at the desired frequency. In contrast to this limitation, since the frequency sensitivity to load capacitance goes down as the overtone mode goes up, the resulting oscillator should be somewhat insensitive to parametric changes in the circuit reactance.

RESULTS

An oscillator was fabricated according to the above procedure. The active device is an MMBR920 bipolar, silicon, SOT package, high frequency transistor. Small signal S parameter measurements for the device were obtained from the manufacturer. The matching network is a combination of microstrip elements and lumped capacitors. The spectrum of the oscillator shows energy at the fundamental and the second and third harmonics. The close in spectrum of the oscillator gives an indication of the stability of the microstrip oscillator. A 25.713647 mHz polished crystal was added to the circuit at a point such that the Co term of the crystal could be neutralized.

The resulting oscillator can be operated at frequencies close to the crystal 17th overtone (437.132 MHz). Adjustment of the single tuning capacitor will result in the oscillator locking to the crystal frequency. The resulting spectrum at the frequency of operation clearly shows the effect of the crystal element. The oscillator operates in a single mode over a supply voltage range of approximately 1.5 to 5 volts. Crystal operation can be maintained over a 2.0 to 4.5 volt range. Output power varies from about -2 dBm to +8 dBm. No attempt was made to measure the noise spectrum of the oscillator due to a lack of equipment for this type of measurement.

The strong second harmonic component of the prototype oscillator offers a convenient way to extend the range of operation for this type of signal source.

REFERENCES


T.G. Bryant and J.A. Weiss, IEEE Trans MTT-16, 1021, (December 1968)


