A New Class of Oscillators

Randy Rhea

The Colpitts is perhaps the best known class of oscillators and is realized in many forms using a variety of sustaining stage active devices. Colpitts’ original triode high-vacuum tube oscillator, with the capacitors grounded at their common tap point, bears little resemblance to the field-effect transistor (FET) and bipolar form of today, with one end of the inductor grounded. It is the topology of the resonator that uniquely identifies the oscillator class, and the familiar feature of the Colpitts resonator is the capacitive tap. The capacitive tap requires a sustaining stage with dissimilar input and output impedance. This article describes a new class of oscillator that uses a resonator with three transmission zeros at dc formed by a series capacitor, a shunt inductor and a series capacitor. This structure is suitable for sustaining stages with either equal or dissimilar input and output impedance. The new oscillator class is economic and offers alternative features to complement the classic Colpitts.

Background

Consider the Colpitts resonator in Figure 1(a). The values selected resonate near 500 MHz, and the ratio of values in the capacitive tap transforms a 50-Ω source up to 1,000 Ω. The components are assumed ideal except that the inductor has an unloaded Q of 100. The transmission amplitude and phase of this resonator is shown in Figure 1(b). The finite unloaded Q of the inductor results in approximately 0.6 dB of insertion loss at resonance. Notice the transmission phase is not exactly 0° at the transmission amplitude peak. As the loaded Q (QL) of the resonator is increased, the phase zero crossing approaches 0° at resonance.

The capacitive tap of the Colpitts resonator dictates the use of a sustaining stage with unequal input and output impedances. Also, because the transmission phase of the Colpitts resonator is near 0° at resonance, the sustaining stage must be noninverting. To satisfy both the match and phase constraints, the Colpitts employs a common-collector (CC) or common-base (CB) bipolar device or a common-drain (CD) or common-gate (CG) FET device. Common-emitter and common-source amplifiers typically have relatively similar input and output impedances, and they invert, so they are unsuitable for Colpitts oscillators.

Figure 1(c) shows a basic Colpitts oscillator using a common-collector bipolar transistor. Open-loop analysis of the initial small-signal starting conditions for this oscillator is accomplished by opening the feedback path between the emitter and the capacitive tap. Figure 1(d) shows an open-loop, small-signal simulation of the transmission amplitude, transmission phase, and QL. The phase shift of the sustaining stage is not exactly 0°, so there is a small shift in the alignment of the amplitude peak and the phase zero crossing. In an ideal oscillator, the active device has no influence on the oscillation frequency. When the final oscillator loop is closed, provided the gain is higher than 0 dB, oscillation builds at the frequency of the phase zero crossing until nonlinear behavior establishes a steady-state condition of 0-dB gain at the phase zero crossing. The ideal sustaining stage has low amplitude modulation to phase modulation (AM to PM) conversion so that compression has little affect on the phase zero crossing.
The common emitter (CE) and common source (CS) configurations provide the best amplifier stability. To improve the stability of CC, CD, CB and CG configurations, series base resistance is generally required. In Figure 1(c), a 470-Ω resistor was placed in series with the base of the CC amplifier. Because the input impedance of the CC amplifier is high, a modest value of resistance has a minor impact on the gain.

The New Oscillator Class

In a recent search for an oscillator configuration with maximum economy and small values of capacitance, I explored a new resonator topology [1]. This topology exhibits other desirable attributes, as will be shown. Typical oscillators use blocking capacitors to isolate the active device dc voltage from the resonator, either to avoid shorting the bias voltage or to isolate the dc voltage from a tuning varactor. Economy would be served if the series blocking capacitors were the capacitive elements of the resonator. To achieve resonance, an inductive element is required. These conditions are satisfied with a third-order highpass structure with a series capacitor, shunt inductor and series capacitor. An example 100 MHz resonator is given in Figure 2(a). Even though all transmission zeros are placed at dc, if the reactance of the capacitors and the inductor is high, the transfer function approximates a bandpass, as shown in Figure 2(b) for the example resonator inserted in a 50-Ω system. When the sustaining stage has similar input and output impedances, C1 and C2 are equally valued. When a sustaining stage with dissimilar impedances is used, unequal values of C1 and C2 are employed.

A 100-MHz version of this new oscillator using an MSA0386 silicon monolithic microwave integrated circuit (MMIC) is given in Figure 3. The MSA0386, originally manufactured by Avantek/Hewlett Packard, is also the Mini-Circuits MAR3 device. Inverting gain blocks available from various suppliers offer matched input and output impedances, stable gain and phase, and require few external components. The inverting transmission phase of the new resonator complements the inversion of these MMICs. The cascade forms an economic oscillator.
The small-signal, open-loop transmission amplitude of the cascade in Figure 3 is plotted in red and the transmission phase is plotted in blue in Figure 4(a). The phase zero crossing is slightly below the gain peak at 100 MHz. The oscillator $Q_l$, a critical parameter for estimating the phase noise, is plotted in green and is just over ten at the phase zero crossing. The impact of sustaining stage nonlinear behavior on the small signal estimation of the oscillation frequency is considered later.

Figure 4(b) is the input and output impedances plotted on a Smith chart. The excellent match to the 50-Ω measurement impedance is a direct result of the design of the MSA0386 MMIC, which employs resistive feedback to stabilize the gain and improve the match.

**Design Formula for the New Oscillator**

Consider the schematic in Figure 5 of a resonator driven by a sustaining stage with output impedance $Z_{out}$ into the sustaining stage input impedance $Z_{in}$:

$$X_b = X_{in} + X_{C2}, \quad (1)$$

and

$$X_a = X_{out} + X_{C1}. \quad (2)$$

The net series reactance transforms the terminations into conductance and susceptance in parallel with the shunt inductor. The sum of the termination susceptances must be capacitive for resonance to occur. For the cascade to be matched, the conductance transformed in parallel with the inductor must be equal for the input and the output. The conductance and the shunt inductor define the $Q_l$ for the resonator. To simply the algebra, the mean loading resistance of the two termination resistances is used:

$$R_m = \sqrt{R_{in}R_{out}}. \quad (3)$$

The mean coupling reactance is then

$$X_m = Q_l R_m, \quad (4)$$

where $Q_l$ is the desired loaded $Q$. The shunt inductor must resonate with the total susceptance transformed from the terminations, which is twice the susceptance of a single termination. Therefore, the susceptance of the shunt inductor is

$$B_L \cong -\frac{2Q_l R_m}{R_m + X_m^2}. \quad (5)$$

The susceptance is approximate because the mean resistance was used, which is an approximation to the individual values. From $Q_l$ and inductive reactance, the conductance from each termination is half the total conductance

$$G = -\frac{B_L}{2Q_l}. \quad (6)$$

And then using the parallel to series impedance transform at the resonant frequency, the reactance of the series capacitors is found...
\[ X_{C1} = -X_{\text{out}} - \sqrt{\frac{R_{\text{out}} - GR_{\text{out}}^2}{G}}. \]  
\[ (7) \]

and

\[ X_{C2} = -X_{\text{in}} - \sqrt{\frac{R_{\text{in}} - GR_{\text{in}}^2}{G}}. \]  
\[ (8) \]

If desired, the exact inductor value may be found by recomputing the susceptance transformed from the terminations individually rather than using the original mean susceptance:

\[ B_a = \frac{X_a}{R_{\text{out}}^2 + X_a^2}, \]  
\[ (9) \]

\[ B_b = \frac{X_b}{R_{\text{in}}^2 + X_b^2}, \]  
\[ (10) \]

and

\[ B_L = B_a + B_b. \]  
\[ (11) \]

Recomputing the inductor introduces a small error in \( Q_l \). The actual resonator \( Q_l \) will be higher than the design value from the formula. Example calculations for a 300-MHz resonator are given in Table 1.

**Extracting Output Power**

You may have noticed that the previous oscillator schematics have no signal output port. Power may be extracted from an oscillator at almost any node. When properly done, the extraction has a minor influence on \( Q_l \). For an in-depth discussion of signal coupling please refer to [1].

Consider Figure 6, the schematic of a 10-MHz oscillator that uses a 2N2222 bipolar transistor sustaining stage [2]: the transistor was modeled using a nonlinear model published by Infineon. The GENESYS simulator [3] was used to linearize this model and compute the small-signal, open-loop responses in Figure 7. Capacitor \( C4 \) in Figure 6 is required for the open-loop simulation to avoid shorting the transistor dc bias network by the 50-Ω source of the simulator. In the final oscillator, Port 2 is connected to Port 1, thus closing the loop. In this case, \( C2 \) provides the dc blocking, and \( C4 \) is not required.

Notice in Figure 7 that the phase zero crossing occurs at 9.94 MHz, and the open-loop gain at that frequency is 2.43 dB. Also notice the misalignment of the phase zero crossing and the peak in the gain and \( Q_l \) caused by a 40° down shift.

<table>
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<tr>
<th>Table 1. Example resonator design for a 300-MHz resonator.</th>
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<td>( Z_{\text{out}} ) (MSA0386 @ 300 MHz)</td>
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<td>Recomputed ( B_L(L) )</td>
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<td>Actual ( Q_l )</td>
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of the transmission phase plotted in blue. This shift is caused by the HF characteristic of the 2N2222 transistor amplifier. The resulting $Q_l$ at the operating frequency is approximately five, but the resonator is capable of delivering a $Q_l$ of ten, except for this misalignment. A higher frequency transistor or a method described later corrects this problem.

The test resonator used a toroid inductor with an unloaded $Q$ of 50. Power was extracted by placing $C_3$ in series with the inductor and coupling signal across this capacitor. Notice that $C_3$ is significantly larger than $C_1$ and $C_2$. The low reactance of $C_3$ partially decouples the load from the resonator and reduces load pulling.

Plotted in red in Figure 8 is the output spectrum predicted by the Harbec nonlinear simulation module of GENESYS [3]. Plotted in green is the measured output spectrum captured using the Testlink module. Notice the exceptional harmonic performance of the oscillator. The second harmonic is 40 dB below the carrier. This results from coupling the output signal directly from the resonator that is acting as a signal filter. This oscillator also has exceptionally fast starting because of the small RF capacitor values that must be charged by the supply when power is first applied. Starting waveforms are given in [2]. Figure 9 is a photograph of the oscillator used to take measured data.

Additional Implementations of the New Oscillator Class

Figure 10(a) is the schematic of a variation on the 10-MHz oscillator that includes back to back varactors in parallel with the resonator inductor. The output coupling capacitor is replaced with a tap of the resonator inductor to provide a dc return for the top varactor. This voltage-controlled oscillator (VCO) tunes about 10% with a varactor capacitance range of 1–8 pF. Wider tuning requires smaller inductor and series capacitor values and larger values of varactor capacitance. Wide tuning is not an advantage of this oscillator class.

This implementation also includes an inductor in the collector of the sustaining stage. The value of this inductor is chosen to tune out the transistor collector capacitance. This reduces the phase shift of the amplifier. Notice in Figure 10(b) that the phase zero crossing is now aligned with the peak of the gain and $Q_l$, thus improving the $Q_l$ of the oscillator.

Replacing the output coupling capacitor with a tap of the resonator inductor has increased the second harmonic to 26 dB below the carrier, as shown in the harmonic balance simulation of the output spectrum in Figure 10(c). This is still outstanding harmonic performance.

Figure 11(a) is the schematic of a 1,152-MHz oscillator that uses no resistors, capacitors, or inductors. The entire oscillator consists of distributed elements printed on a 62 mil thick
polytetrafluoroethylene (PTFE) substrate and an MSA0386 MMIC amplifier.

The 20 mil wide by 240 mil long transmission line Ld connects the supply to the output of the MSA0386. Because this line is narrow and electrically short, it acts much like an inductor. Its length was chosen to align the phase zero crossing with the gain peak, as was done with the 1,500-nH inductor in the 10-MHz VCO.

A layout of this oscillator is given in Figure 11(b). The shunt inductor of the new resonator is achieved with the 40 mil wide line that is grounded with the same viahole used to ground the MSA0386. This 40 mil wide line runs downward and splits left and right into two wide lines that run back up. The wide lines are 220 mils wide by 580 mils long. This structure serves as a stepped-impedance resonator (SIR) which is physically shorter than a constant-width resonator. A standard quarter-wave line at 1,152 MHz would be about 2 in long. SIRs are covered in [4]. Splitting and folding the wide line around the narrow line further conserves space. The coupling length is less than 45°, the narrow line is high current and develops a magnetic field, and the open-end wide lines develop electrostatic fields, so minor coupling exists between these lines.

The series capacitors of the new resonator are realized as capacitive fingers penetrating the open end of the wide lines. The oscillator was designed with two outputs, Ports 1 and 2, in the layout. The load lines are coupled electrostatically to the wide sections of the shunt inductor. The resonator was designed using the EMPOWER electromagnetic simulator module of GENESYS. The open-loop transmission amplitude, phase, and Ql are plotted in Figure 11(c).

Despite extreme simplicity and economy, the Ql of this oscillator is over 50, resulting in good stability and phase noise. This oscillator is an illustration of how reverting to first principles

Figure 8. Plotted in red is a nonlinear harmonic balance simulation of the oscillator in Figure 6 with the loop closed. Plotted in green is measured spectrum for the oscillator captured by the Testlink module.

Figure 9. Photograph of the completed test oscillator used to take the measured data.

Figure 10. Variation of the 10-MHz oscillator with varactor tuning and optimized to align the phase with the gain and Ql peak: (a) schematic, (b) open-loop responses, and (c) simulated output spectrum.
and zero-base designing achieves performance without complexity.

**Summary**

A new class of oscillator is introduced that utilizes a high-Q, high-pass resonator with series-coupling capacitors, thus eliminating the need for blocking capacitors. This economic structure is capable of excellent performance without resorting to complexity. A few example implementations were provided along with measured performance data for a test oscillator.

**Acknowledgment**

I would like to thank Bill Clausen of Eagleware for constructing the 10-MHz oscillator and taking the measurement data.

**References**


