The Virtual Ground in Oscillator Analysis — A Practical Example

This analysis technique helps us understand oscillator behavior for the design of reliable circuits

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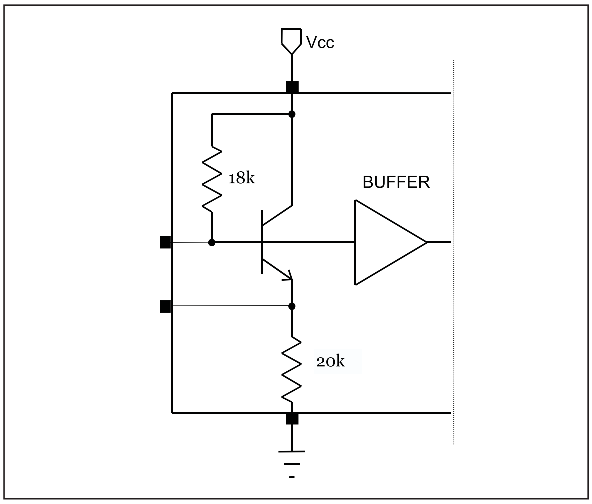
In a 1998 article, I presented the virtual ground concept for oscillator analysis [1]. Since then, I have received many comments, sometimes interesting, sometimes amazing. Recently, a paper mentioned my article approvingly, then proceeded with an almost contradictory approach. I decided that a practical and simple explanation of the virtual ground concept was needed.

My earlier explanations dealt with rather well-defined oscillator examples, based on discrete microwave transistors strictly described by their *s*-parameters. Yet, this is not always the case. Sometimes a designer has little data about the active component with which an oscillator must be built, especially when dealing with specialized integrated circuits. These circuits include embedded oscillator parts for which a few passive components are needed to yield oscillation. Usually, such an IC has two pins for that task, typically the base and emitter of an internally biased bipolar transistor with the output available via a buffer. The NE602 mixer/oscillator from Philips serves as a good example. Its oscillator part is extremely simple as shown in Figure 1.

Let us say we want to make a 100 MHz oscillator, knowing that it is capable of operation at 200 MHz. Is it possible to design a reasonable circuit with so little information? There is an application note [3] on the NE602 oscillator, with examples, but it is not very helpful for analysis. Obviously, for a somewhat experienced engineer it is not a problem to build such an oscillator, even without any analysis. Simple frequency scaling may be sufficient based on any given application, and some trimming may accomplish the matter if needed.



▲ **Figure 1. The oscillator portion of the NE602 mixer/ oscillator.**

But here the problem is different; we want to get a complete insight into the process. I will show that application of the virtual ground concept, then proceeding with transmission analysis, yields optimal results.

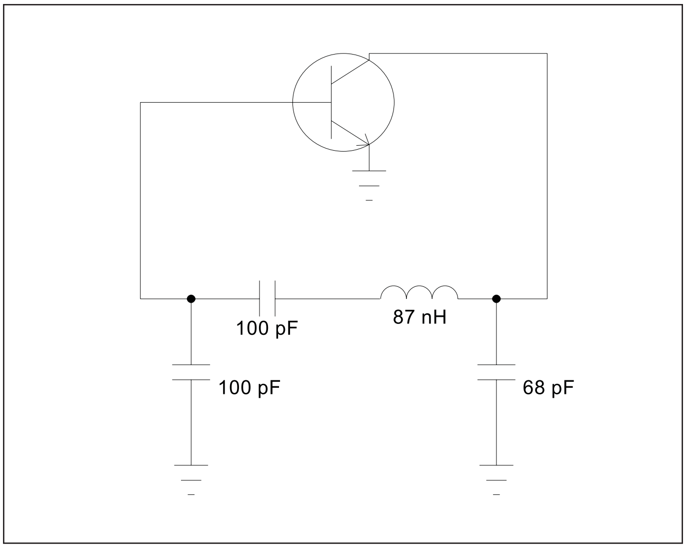
The NE602 active circuit is simple for oscillator applications. The output buffer, aided by the internal transistor bias, enables design simplification by neglecting the output coupling. The high values of the biasing resistors makes it possible to ignore them in the AC circuit. The transistor alone becomes the whole amplifier portion, only needing a resonator to complete the circuit we wish to analyze.

In the NE602, the base and emitter pins are at our disposal and the collector at AC ground. Does this make any difference in the way we analyze the oscillator?

The virtual ground concept in oscillator analysis states that essentially, it does not matter where the physical ground is set. Contrary to other transistor circuits where some external signal is applied to the transistor in a prescribed way, in an oscillator noise is the signal origin, and noise is everywhere. We need to take the proper point of view, plainly showing the oscillation starting process. This is done by establishing a virtual ground for the sake of analysis only. Later, the physical ground can be placed in any convenient point, considering such constraints as power coupling, DC supply and impact on parasitics.

The following is a short analysis based extensively on the considerations given in [1]. In the case of the bipolar transistor alone, without any feedback elements, the virtual ground would normally be placed at the emitter electrode. This may not always the case; for example, if there were a series feedback resistor in the emitter branch, the virtual ground would not be placed directly at the emitter, but after the resistor.

Having established what the active part of the oscillator is like, as well as how to take it into account, we now have to choose a resonator. The resonator of first choice is the shunt-C coupled series LC resonator, analyzed in more detail in [1]. Connecting it to the transistor to create a feedback loop, the simplest oscillator configuration is obtained, as shown in Figure 2.



▲ **Figure 2. The initial oscillator configuration with the virtual ground.**

Obtaining transistor data

To evaluate the resonator alone, we should know some information about the transistor. Yet, in this case there is no transistor data. Do not get discouraged; let us attempt some detective work. In the application note mentioned earlier, we can find some clues. First, the device is fabricated with a 6 GHz process, and second, the transistor has an extremely low 0.25 mA bias current, although the schematic resistors suggest a little lower value of ~0.2 mA. Searching Philips' wideband transistors for a device with similar characteristics, we cannot find such a low current device in a 6 GHz process group, but there is one in the lower group (BFT25) and in a higher group (BFT25A) for currents ranging from 0.1 to 1 mA — as we want. There is no question about which device's s-parameters to use at 100 MHz; they are almost the same. The BFT25 has data given for the 1 V/0.2 mA bias point, which is suitable for our design.

We are justified in our assumption that this transistor is similar to the one in the NE602, as BJT transistor properties are rather repetitive for similar processes, f_T and bias. The device housing is not very significant at 100 MHz and, as we will see later, the design will not be very sensitive to the particular device parameters because it is sufficiently determined by some generic properties.

Let us asses input and output transistor impedances first. Calculating them from s-parameter data for 100 MHz gives 9 kohm in parallel with 0.8 pF at the input and 100 kohms in parallel with 0.7 pF at the output. There are terrible resistances for RF circuitry, however, they should be expected since transistors with f_T in the GHz region have large impedances in the MHz region. When biased with small currents, the resistances can become very large. Looking at the chip schematic, we notice that the emitter resistor (20 kohm) is (for AC) connected between the collector and the emitter of the transistor. Thus it can be included in the transistor description, dropping its output resistance to a more bearable 17 kohm value. The s_{12} parameter will be neglected in the analysis with provisions provided later. As Figure 2 indicates, the transistor capacitances will be included with the resonator shunt capacitances, and the analysis will only include the remaining transistor resistances. This suggests that an adequate gain parameter for our analysis is the maximum unilateral power gain G_{UM}, conveniently provided in the data sheet. In this case, G_{UM} at 100 MHz is 40 dB, but after including the emitter bias resistor as above, it drops to 32 dB. The s_{21} phase of 173° is another value of interest; it means the resonator must provide about the same phase shift in the negative direction to meet oscillation conditions.

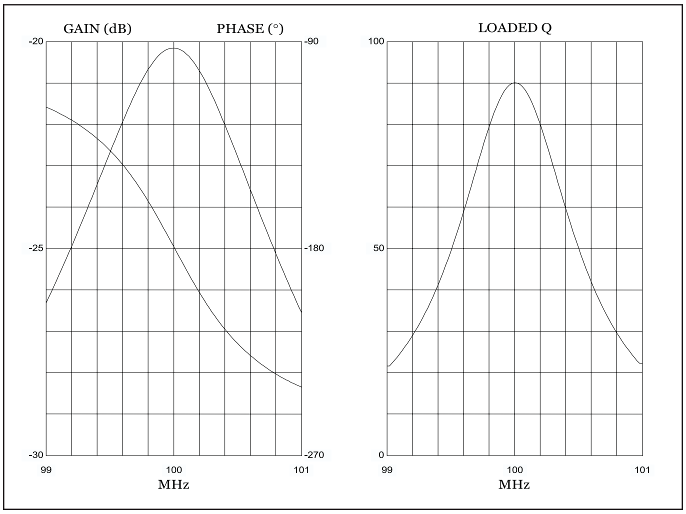
These few numbers are all that is needed, which is different from the microwave examples analyzed in [1]. These parameters may seem inconvenient. In the microwave example we had reasonable impedances in the vicinity of 50 to 200 ohms, allowing suitable resonator analysis and application. Now they are enormously high and strongly capacitive. The phase requirement for the resonator is also very tight, close to the –180° asymptote. The gain is also reasonable in the microwave example, allowing for a sufficient gain margin of a few dB in the oscillator loop for losses. Now the accessible gain seems too high to leave it in the loop, preventing hard limiting and uncontrolled operation. Fortunately, we will see that such constraints are not restrictive in practice. Working together, they allow for an efficient design.

It would be interesting to first calculate a resonator with reasonable input/output reference impedances of 100/200 ohms and a sufficient loaded Q of 10, and then to embed it into 1 k/2 k impedances and calculate this higher order. Following the procedure given in [1], starting with the series inductor reactance of 100 ohms, one gets a resonating circuit with the phase shift of the order of –150° at resonance. This is rather small as it is reasonable to keep phase shift error below 10°. Yet the resonator topology is quite universal, allowing us to achieve various phase shifts while retaining the resonant frequency (f_0) and loaded Q (Q_L). The higher Q_L is reached by raising series reactances or lowering shunt reactances. When they are raised, the resultant phase shift gets closer to –180°. The task can be accomplished easily by tuning within a circuit simulator. Lowering all the reactances in order to keep the f_0 and Q_L unaffected (with possible rounding off) and using standard capacitor values, a proper phase of –163° is achieved for the elements (as in Figure 2). The resonator transforms impedances with a factor of 2, so this makes the shunt capacitances proportional to √2. The inductor loss alone is enough to take this into account, and for a good air coil with Q=100, the resonator presents a –0.9 dB loss when analyzed with 100/200 ohm analysis characteristic impedances. Switching reference impedances to 1 k/2 k, we observe that the f_0 is only slightly disturbed with phase shift near –180° (as desired), with a 6dB loss and Q_L (measured as group delay × π × f_0) ≈ 50. Taking it an order higher to 9 k/17 k actual impedances, the loss amounts to –20 dB with the phase close to –180° and a Q_L of 90, as illustrated in Figure 3 (with the inductor value corrected to 87.9 nH to keep f_0=100 MHz).

The results are for a real inductor Q of 100. When analyzing a lossless circuit, the Q_L increase would be the same as that of the reference impedances, and the overall loss would be 0 dB. Notice that all the changes are an advantage. The real inductor Q now has a much greater impact on the resonator loss, yet it only cancels excessive gain. With an actual modified transistor G_{UM} of 32 dB, the overall loss could reach 22 dB, still leaving ~10 dB to sustain oscillation. Therefore, we can afford high loss not only caused by the component's limited Q, but also from mismatch in the loop.

The transistor in/out capacitances are consistently ignored, but notice that they are added to much larger resonator shunt capacitors and can be included with them. When excluded from the transistor, the phase is closer to +180°, but the resonator phase is now also close to –180° so they correspond well to one another.

The s_{12} parameter could also be simply neglected. By analyzing the transistor model, one can see that the parameter is determined mainly by the base-collector capacitance on the order of 0.3 pF. It does actually mean that the second feedback loop is parallel with the first loop through the resonator. For ideal transmission analysis there should be only one feedback loop and one resonator block that includes all reactive elements. The task can be readily accomplished by artificially excluding the base-collector capacitance from the transistor and including it across the resonator. As Figure 2 depicts, such a transposition does not influence the oscillation conditions, yet it enables a more strict analysis. Now, the effects of that undesired transistor capacitance can be examined simply by analyzing the resonator alone with transistor capacitances attached across it.



▲ **Figure 3. The analysis results of the resonator.**

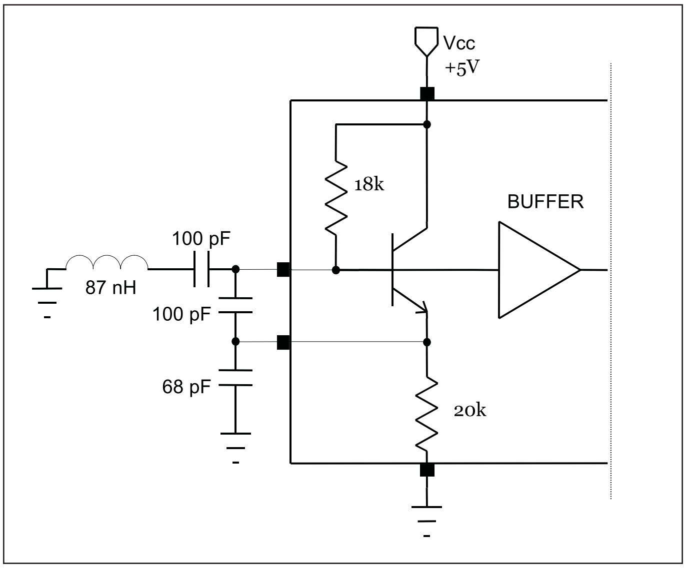
The results show those effects to be negligible. Although the oscillator analysis is associated with very high transistor impedances, the impact of the base-collector capacitance is small because the resonator shunt capacitances establish a low reactance value. It is also worth noticing that the s-parameter measurement process, conducted in the 50 ohm system, measures the base-collector capacitance as both input and output capacitance. So the input/output capacitances calculated earlier from the s-parameters may be as much as two times lower when the base-collector capacitance is excluded, making it more evident that they can be neglected when added to much higher resonator shunt capacitances.

The practical circuit

We can proceed with the practical design, transforming it from the virtual to the physical configuration in which the collector branch must be at ground, as required by the NE602 chip. In Figure 2, the ground marks are canceled and their branches are connected into one emitter branch. At the same time, the collector branch is grounded; the collector pin and the relevant capacitor and inductor pins are shorted to ground. The transformed circuit, using the NE602 diagram, is shown in Figure 4. The circuit was built and measured, with the oscillation frequency close to 100 MHz.

Notice that the resonator inductor is wound as an air coil with a wire diameter of 0.5 mm to assure a high Q of ~100. However, one may also use an SMD inductor, available in wide varieties. The circuit was checked with an 82 nH wire-wound SMD 0805-size inductor from Coilcraft (with Q=50) and oscillation was just at the limit. This makes it possible to determine the actual gain margin in the loop. When checking the inductor Q influence in the resonator analysis, note that it increases the loss from –20 dB to –26 dB as the Q changes from 100 to 50. So the real gain margin is about 6 dB — a sufficient value. Checking the oscillation with the inductor Q at half the original value seems to be a practical way to check for sufficient gain margin. The capacitor's Q should be above 1000 at RF frequencies, so the inductor alone determines the resonator loss.

At this point, notice the discrepancy between the calculated gain margin of 12 dB and the estimated actual margin of 6 dB. Most of this difference may be caused by the output buffer, which may present several kohm resistance with a part of pF capacitance at its input. Note that this impedance is actually connected (for AC) directly between the base and the collector electrodes of the oscillator transistor. In parallel with the base bias resistor, its effect of ~6 dB is quite possible. However, it is not known how much the actual transistor implemented in the chip differs from the BFT25. Nevertheless, the above estimation of gain margin is sufficient, even if its calculation is only approximate.



▲ **Figure 4. The oscillator part of NE602 after transformation from the virtual to the physical ground.**

Nonlinear circuit behavior

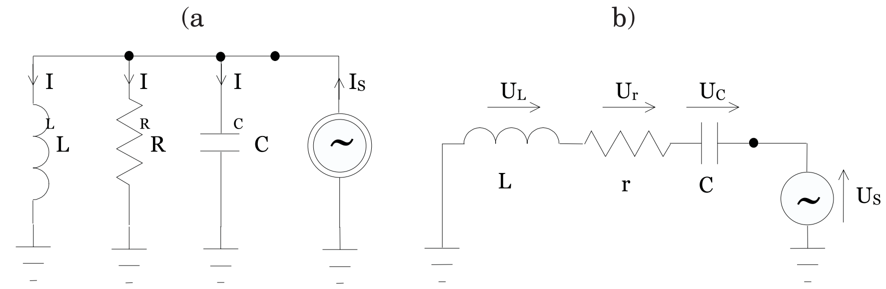
All the considerations mentioned thus far have proceeded according to the linear conditions of rising oscillation. However, the final conditions may be greatly nonlinear as the signal level reaches its limit, as determined by the actual transistor bias. By definition, the gain margin is associated with rising oscillation conditions, so there is no need to deal with it further. The phase shift balance in the oscillator loop (revealed only with the virtual ground) is proved throughout a wide range of possible impedances. The proper oscillation frequency also confirms the range of impedances, so the phase balance is not disturbed by nonlinear characteristics. Thus, only the change in loaded Q in the nonlinear state should be examined. One can expect a significant drop [2] as it depends directly on resonator terminations.

The common nonlinear software tool, SPICE, will be employed to research the oscillator final state. The schematic from Figure 4 (without the buffer and with inductor loss for Q=100) was drawn in SPICE to conduct a transient analysis. To initiate oscillation startup, an initial pulse is needed. It should be a narrow (5 ns, ~T/2) 10 V pulse source delivered through a high (100 k) resistor to the resonator inductor. Analyzing the results well into the settled state (within 25 ns after 1 µs), 100 MHz sinusoidal voltage waveforms can be observed at the base and emitter with values of 1 V_{p-p} and 0.63 V_{p-p} respectively. In contrast with these sine waves, the transistor currents exhibit highly asymmetric pulse shapes.

It was shown earlier that the transistor capacitances can be artificially included in the resonator for the sake of analysis and even neglected because of high resonator shunt capacitors. Now it is interesting to see what change will happen when neglecting them in a nonlinear analysis. Simply nullifying C_{JC} and C_{JE} in the SPICE transistor model, the oscillation frequency change is less than 1 MHz and the voltage waveforms remain almost the same. Thus the modification is justified, giving in effect very regular base and emitter currents as narrow (~2 ns) pulses of 30 µA and –1.5 mA peak values. The advantage is that the operation is now more evident and can be plainly described. In accordance with AC voltage on the B-E junction, the narrow pulse currents indicate that the transistor is switched off through the majority of the cycle and switched into active state for a small amount of time when it injects a high current (compared to the initial transistor current) into the resonator.

Here, one could remark that the intervention with transistor capacitances is much too rough to get any regular results. Let me stress that it is not the case. An oscillator is a very special circuit with the initializing signal (noise) existing throughout each element. Any reactance in the amplifying element behaves as part of the resonator and, if the analysis is to be clear, it should always be included in the resonator. For clarity and conciseness, we can neglect the transistor capacitances (and their nonlinearities).

Oscillator analysis is usually limited to linear conditions, which is frequently justified, especially when dealing with reasonably low impedances or high currents at high frequencies. Here we have a clearly different case with high nonlinear conditions within a pulse circuit. The behavior is really amazing; the resonator is cut off from the transistor except during short moments of charging it. Therefore it is hard to retain a linear Q_L approach. Rather, one should define and determine the loaded Q in nonlinear conditions, referred to as Q_{NL}. Examining currents in the circuit will help here. Notice a high sinusoidal current across the resonator components of 27 $mA_{p\text{-}p}$, compared to the small, narrow emitter current as given above. It is intuitively understandable that the real Q should be proportional to the first and inversely proportional to the second, as shown by the basic equivalent schematics in Figure 5.



$$I_L \cdot X_L = I_R \cdot R = I_C \cdot X_C$$
$$I_L = -I_C \qquad I_R = I_S$$
$$X_L = -X_C$$
$$Q = \frac{I_L}{I_S} = \frac{-I_C}{I_S} = \frac{I_L}{I_R} = \frac{I_C}{I_R} = \frac{R}{X_L} = \frac{R}{X_C}$$

$$\frac{U_L}{X_L} = \frac{U_R}{r} = \frac{U_C}{X_C}$$
$$U_L = -U_C \qquad U_r = U_S$$
$$X_L = -X_C$$
$$Q = \frac{U_L}{U_S} = \frac{U_C}{U_S} = \frac{U_L}{U_r} = \frac{U_C}{U_r} = \frac{X_L}{r} = \frac{X_C}{r}$$

Q relations with magnitudes at resonance.

▲ **Figure 5. Two basic resonator forms with ideal current and voltage excitations.**

Simple equations in Figure 5, along with these basic topologies, show ways for determining Q. The second of the pictures illustrates the scheme of the first Q-meter. Here, the unknown inductor is placed in a series with a low resistance voltage source and the voltage across the capacitor is measured using a high impedance voltmeter. The ratio of the capacitor voltage to the driving voltage gives the circuit Q.

In our case, the first of the pictures, with the application of an excitation source, is the current one. However, the transition from the schematic of Figure 4 to Figure 5a may be not be readily evident. Notice that the sinusoidal voltages in the circuit are determined exclusively by the resonator sinusoidal current circulating in the loop consisting of the inductor and the three capacitors. Any resemblance to the emitter follower action of the transistor would be misleading. In contrast, the B-E junction is freely controlled by the resonator-induced voltage, causing the transistor to act as a current source connected in parallel with the 68 pF capacitor. Q's determination of the small base current can even be neglected, so let the equivalent uncontrolled current source in parallel with 68 pF capacitor represent all the transistor action. The remaining resonator inductor and 100 pF capacitors can be represented in parallel form, giving in effect a simple parallel resonator excited by a current as in Figure 5a, with overall loss represented by one resistor R.

The explanations made so far should assure that the Q in the circuit of Figure 4 is simply determined by comparing the inductor current with the transistor current. And according to Figure 5, that would simply be the proportion of their amplitudes if they were both sinusoidal. But the actual situation is different, because the transistor behaves as a charge pump, delivering some amount of charge during a short part of the cycle to sustain the charge circulating in the resonator. With this in mind, it is straightforward to extend the Q formula over nonlinear conditions and any current shapes. If the shapes are different and non-sinusoidal, one should take into account their integrals over the cycle — the charge injected versus the charge circulating in the resonator. Referring to injected currents and charges as I_i and q_i and to circulating currents and charges such as I_c and q_c, one can describe the formula for loaded Q in nonlinear conditions as:

$$Q_{NL} = \frac{\int_O^T \left|I_c(t)\right| dt}{\int_O^T \left|I_i(t)\right| dt} = \left[\frac{q_c}{q_i}\right]_{t=T}$$

The integrals can be obtained with help of built-in SPICE functions while the q_c can be calculated from the $I_c(t)$ sine waveform. Remember that the charge in both positive and negative halves should be taken into account, shown as:

$$2 \cdot \int_O^{T/2} I_c(t)dt$$

The Q_{NL} obtained this way is ~40. Although significantly less than the value calculated for linear conditions, it is still relatively high as it represents the plain LC resonator for which typical values are 10 to 30.

Conclusions

A practical oscillator design has been discussed in detail, showing that is it possible for efficient design with little data. It is based on rough, but reasonable assumptions along with an effective analysis. Although a particular integrated circuit was used as an example, the procedure covers a wide range of applications. The idea extensively presented in [1] is confirmed, and it should cause us to abandon the negative resistance approach and avoid resolving circuit equations. Instead, the initial task is to find and define the oscillator positive feedback loop, which requires applying the virtual ground concept. Then it is possible to identify and isolate the active stage and the resonator into their basic forms, enabling a separate analysis. This is the way to accomplish a complete transmission analysis, providing full insight into oscillator behavior as well as determining all of its parameters.

Nonlinear conditions in the oscillator settled state were also examined, along with their influence on the oscillator parameters. This analysis also confirmed the validity of the technique as the oscillator loaded Q in nonlinear conditions were defined and calculated. ■

References

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