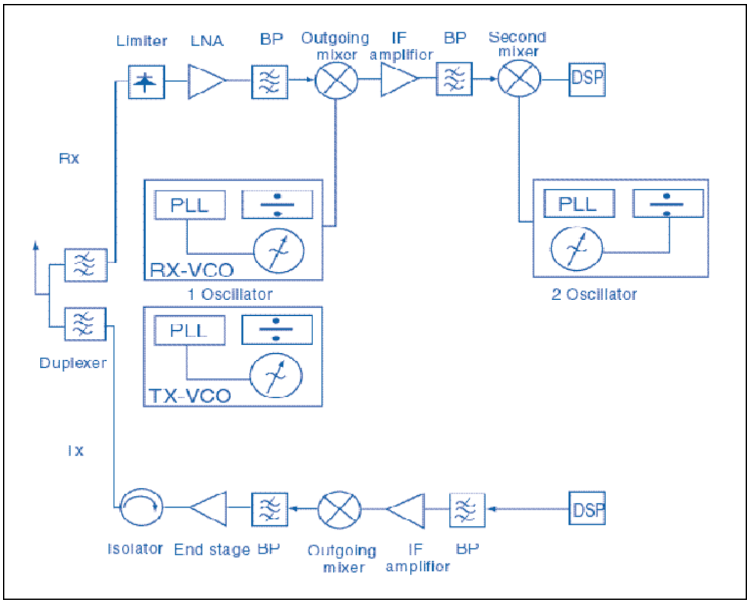
Low-Noise VCOs: Key Components for Base Stations

High performance communications systems require clean signal sources

By Frank Baberg
Tekelec Temex

The great economic success of modern mobile radio systems such as GSM and DCS means even greater utilization of the capacity of existing channels. It is therefore immensely important to exactly adhere to the GSM specifications.

In GSM systems, the available frequency range is divided according to the FDMA procedure] (frequency division multiple access) into radio channels of 200 kHz each [1], [2]. Each radio channel is further divided into eight traffic channels through a time multiplexing TDMA procedure (time division multiple access) [1], [2]. These channels contain the information (voice and data signals) in "bursts." In the case of a channel width of 200 kHz, this results in the typical GSM system channel number of 124 channels with bandwidths of 25 MHz (the first channel is not normally used). For DCS, there are 372 channels with a bandwidth of 75 MHz.



▲ **Figure 1. Block diagram of the RF portion of a typical wireless base station, showing frequency sources.**

The block diagram in Figure 1 shows the frequency generating scheme in a base station. In the transmitter part (TX), the working signal must be converted into a RF signal. In the reception path (RX), the radio frequency signal received is converted into one (or two) fixed intermediate frequencies.

Each of the two conversion processes requires a local oscillator (LO). As a base station works in full duplex mode, the RX and TX paths are seprate and must have their own local oscillators. With mobile phones, a common local oscillator is sufficient because it they use half-duplex operation due to the time slots (TDMA).

Obviously, the specific scheme used in a particular radio can differ significantly from Figure 1. For example, the RX path can also be constructed using only one intermediate frequency.

In modern communication systems, a synthesizer is normally used. An oscillator is typically synchronized with a reference via a phase-locked loop. There are a number of different ways to create a precise reference, such as deriving it from the fixed network clock or by synchronization via GPS.

A VCO is used as an oscillator, as its frequency is dependent on an applied voltage, so that the VCO can be tuned (switched) to the channel frequencies relatively simply and quickly.

Structure of the VCOs

Microwave oscillators are usually analyzed using the concept of "negative resistance" (e.g. [3]). In designing oscillators, various basic switching operations can be found in the literature, such as the Hartley-Meissner or Colpitts switching operations.

The so-called Clapp switching operation has proven itself in VCOs in particular. The Clapp switching operation is very similar to the Colpitts switching operation, only here the inductor is replaced by a resonant circuit (Figure 2a).

In Figure 2b, the Clapp oscillator in 2a is shown, includign only the RF components that are important for the operation. Figure 2c shows the equivalent circuit diagram, extremely simplified, as we are only interested in the principle here. The following applies to the impedance in the place of the serial resonant circuit:

$$Z = \frac{V}{i} = \frac{V_{be} + V_{c2}}{i} = \frac{\frac{i}{J\omega_{c1}} + \frac{i}{J\omega_{c2}} + \frac{Gn}{J\omega_{c3}}}{i} = \frac{L}{J\omega_{c1}} + \frac{L}{J\omega_{c2}} - \frac{Gn}{J\omega_{c3}} \quad (1)$$

Thus from Equation (1), the oscillation condition to produce a negative resistance is as follows:

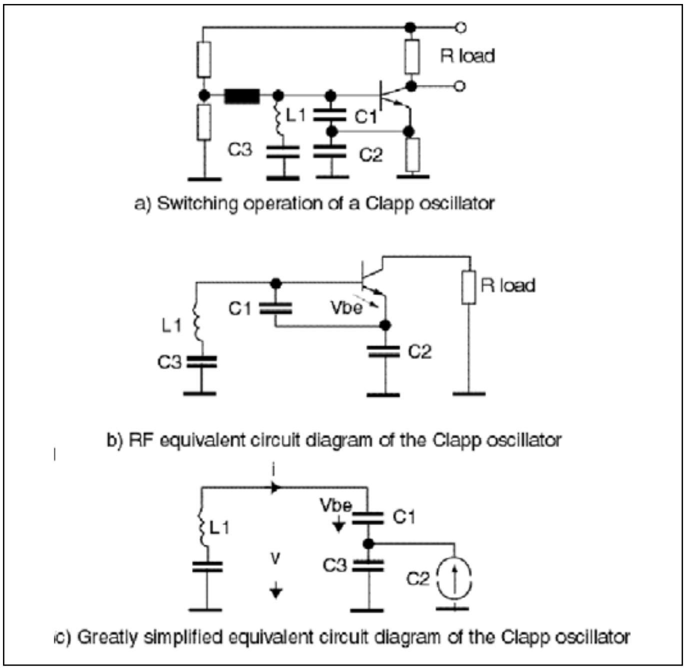
$$\frac{1}{J\omega_{c1}} + \frac{1}{J\omega_{c2}} \leq \frac{Gn}{\omega^2{}_{c1c2}} \quad (2)$$

The frequency is determined by the series connection of the three capacitors:

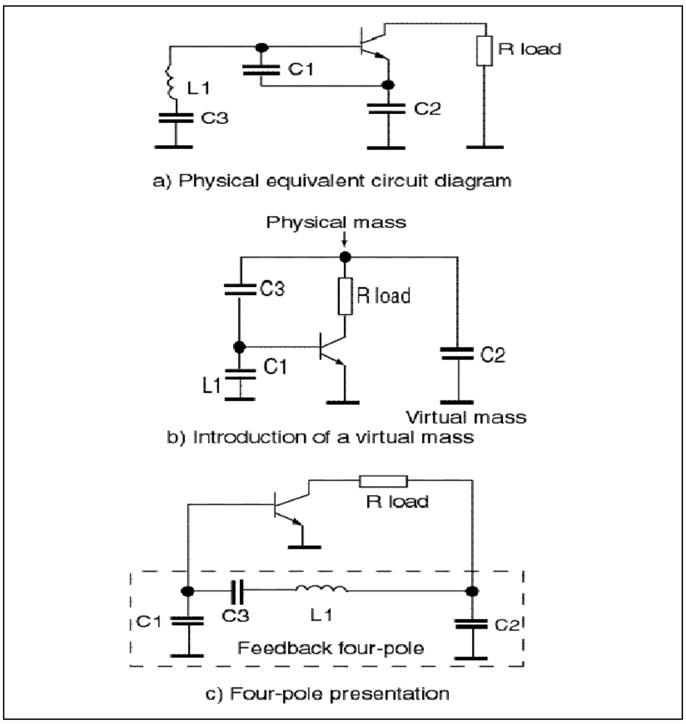
$$\omega^2 = \frac{1}{L_1}\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right) \quad (3)$$

In radio frequency technology, the oscillator is often shown as a four-pole amplifier with amplification V, and its output voltage is fed back to the input via a feedback network. If one draws the Clapp oscillator in accordance with Figure 3, introducing a "virtual mass," then one will get the known four-pole switching operation.

In order to determine the frequency of the oscillator when tuned by a voltage, one of the capacitors is replaced by a varactor. This component exploits the junction capacitance of a diode operated in the junction direction, which is dependent on the reverse voltage applied.

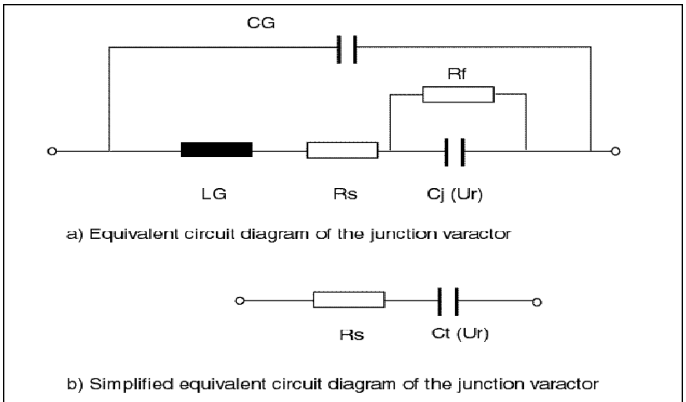


▲ **Figure 2. Clapp oscillator operation and equivalent circuit diagrams.**



▲ **Figure 3. Clapp oscillator feedback repsented as a four-pole circuit.**

Analogous to the capacitor, the junction capacitance of a PN junction is dependent on the cross-section surface and the width of the junction. A theoretical analysis produces the following relationship for the voltage dependence of the junction capacitance [2]:



▲ **Figure 4. Equivalent circuit diagram of the junction varactor in the package.**

$$C_j = \frac{C_{j0}}{\left(1+\frac{V_r}{V_d}\right)^m} \approx C_{j0}\left(1+\frac{V_r}{V_d}\right)^m \text{ for } V_d << V_r \qquad (4)$$

In this equation, C_1 is the junction capacitance, where $V = V_r$, C_{j0} at $V_r = 0$ V, V_d is the diffusion potential (approx. 0.65 V for silicon) and V_r is the reverse voltage.

The exponent m depends on the course of doping and is decisive for the voltage dependency of the junction capacity. In a diffused junction, the junction of the acceptor density N_A (P area) is linear to the donor density N_D (N area); in this case $m = 0.33$. With an abrupt junction, the transfer is carried out suddenly; in this case $m = 0.5$.

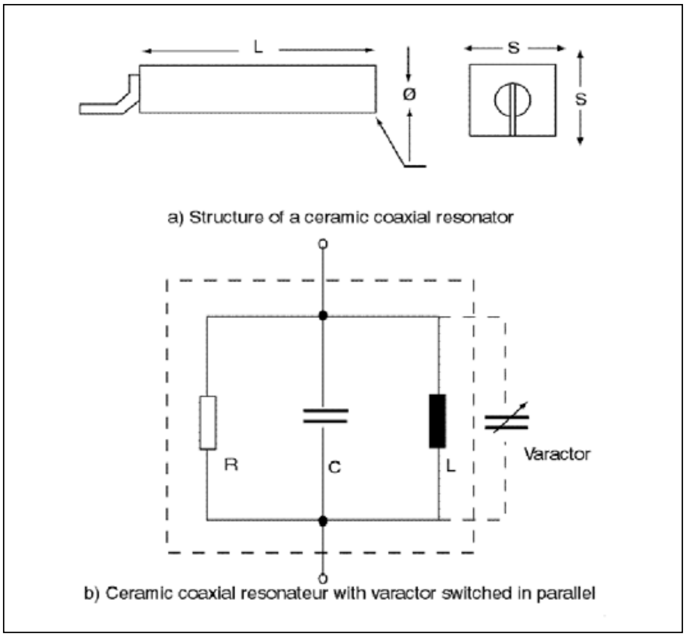
If one requires a particularly strong dependence of the junction capacitance on the voltage, m must be >0.5. In this case, doping density must again fall after the abrupt junction. Such doping profiles are called hyper-abrupt.

Figure 4 shows the small signal equivalent circuit diagram of a junction varactor. The resistance R_S takes into consideration the reverse current of the diode and should be as large as possible, in the interest of low noise (shot noise). In the case of higher frequencies, the bulk resistances above all become noticeable. The influence of the package is described by the line inductivity L_G as well as the package capacitance C_G.

In addition to the capacitance relationship C_1/C_{j0}, the quality factor Q is a decisive characteristic. Analogous to the capacitor, the quality of the varactor is also defined as the relationship between the reactive and the active performance. From the eqivalent circuit diagram 4b, ($C_T = C_j + C_G$), thus results:

$$Q = \frac{1}{\omega C_t(V_r)R_b} \qquad (5)$$

Basically, the quality of abrupt junction crossings are significantly better than those of hyper-abrupt crossings; however, very high reverse voltages of up to 90 volts are required to achieve sufficiently large capacitance variations.



▲ **Figure 5. Ceramic coaxial resonator as used in a high quality oscillating circuit.**

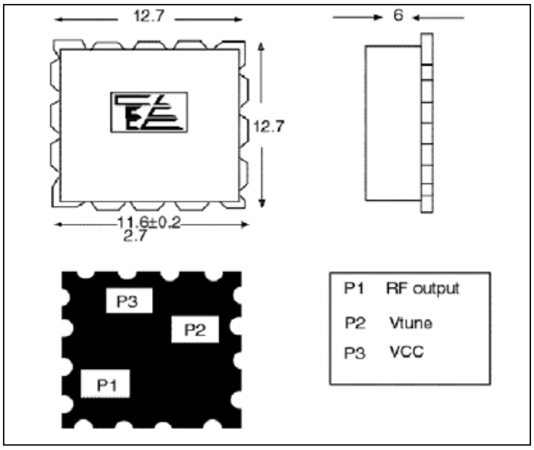
These high reverse voltages are also required because, in the case of voltages which lie significantly below the breakdown voltage, the reverse zone is only partially purged; while the ohmic resistance, which is in series with the capacitance, still lies within the non-purged zone.

For these reasons, hyper-abrupt junction varactors are most commonly used as oscillating circuit capacitances. In selecting a suitable varactor, it is also important that the relationship between the reverse voltage and the junction capacitance is defined over as wide a range as possible.

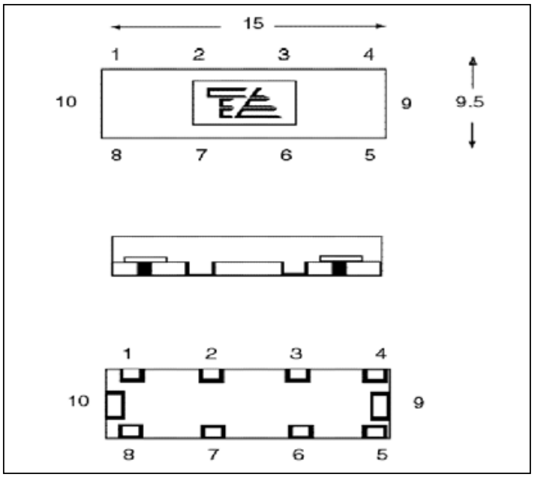
Q and its effect on phase noise

In the next section, we will state why the quality of the oscillating circuit is important for the phase noise. Since modern high Q capacitors offer excellent quality [4], [5], the selection of the inductor [L], as well as the varactor, determines the phase noise to a significant extent. With low demands on phase noise, one can create inductance through a coil printed onto the PCB. Better results are obtained in using air-core reactors in SMD packages.

The best characteristics are obtained by the use of coaxial resonators (Figure 5). Ceramic resonators are shaped as cuboids with a coaxial bore. The inner and outer surfaces are metallized. The capacitance, inductance and the resistance of the metallization create a resonant circuit which oscillates in TEM mode. Particularly space-saving are the one-quarter wave-



▲ **Figure 6. Approximately square VCO package, model SM1.**



▲ **Figure 7. Rectangular package, model SM4.**

length (λ/4) types. The additional metallization of an end face creates the required short-circuit.

The resonant frequency is obtained from the relative permittivity counter and the length of the resonator. Basically, the context for λ/4 resonators is as follows:

$$L = \frac{\lambda_0}{4} \times \frac{1}{\sqrt{\varepsilon_r}} \qquad (6)$$

Dielectric values of ε_1 = 20 to 78 are available.

The Q is almost exclusively determined by the final conductivity of the metallization to a value of Q <800. Where higher quality is required, a special silver metallization is recommended; in the case of price-sensitive applications, a copper-plated metallization is preferred. The no-load operation quality Q_0 is defined as the quotient of the resonance frequency and the 3 dB bandwidth of the resonance curve:

$$Q_0 = \frac{f_r}{B3_{db}} \qquad (7)$$

Q_0 increases in the first approximation by $\sqrt{f}$. Higher Q values can be achieved using larger cross-section measurements, where it is then critical to integrate the resonator into the normal very small VCO package.

Characteristics of the VCOs

Basically, VCOs are customer-specific modules; each user design is different and thus the VCO modules must be adapted to customer specifications. For example, typical values are stated in Table 1 [6]. Such details can only be used as guide values for sample deliveries; for series, customer-specific values are generally stated. In this section, an additional statement is made on the most important parameters of a VCO specification.

Phase noise— Phase noise is the most critical parameter in designing a VCO and must be specified with particular care. In the case of sensitive preamplifiers, normally only the amplitude noise is taken into consideration as it characterizes the sensitivity of the amplifier. With oscillators, the amplitude noise plays only a subordinate role. Decisive here are the stochastic changes in the zero transits of the sinusoidal oscillation created by the oscillator. The phase noise characteristic thus describes the relationship of the carrier magnitude to the noise magnitude in the region near the carrier frequency. This relationship is described by the function $\xi = F(f_m)$, dependent on the carrier offset.

The most obvious significance of the phase noise can be found in the case where phase noise creates interference in the neighboring channel. A typical VCO specification, therefore, states certain values depending on the carrier offset (see Table 1).

The phase noise of a VCO has been observed in numerous theoretical experiments, and without going into detail [7, 8], Equation (8) is determined:

$$\zeta(f_m) = 10\log\left\{\left[\left[1 + \frac{f_0^{\,2}}{(2 f_m Q_{load})^2}\right]\left(1 + \frac{f_c}{f_m}\right)\frac{FkT}{f_m} + \frac{2kTRV_0^{\,2}}{f_m^{\,2}}\right]\right\} \qquad (8)$$

In this equation, the meanings are as follows:

- $\xi(f_m)$ Relationship of the magnitude of the phase noise at 1 Hz bandwidth to the common output magnitude of the VCO, stated in dBc/Hz
- f_m Offset from the carrier frequency
- f_0 Carrier frequency
- f_c Noise corner of the flicker or $1/f$ noise of the active oscillator
- Q_{load} Quality of the loaded resonator (resonance circuit with active load and parasitic elements)
- F Noise figure of active oscillator four-pole
- k Boltzmann constant ($k = 1.38 \times 10^{-23}$ J/K)
- T Temperature in Kelvins
- P_{av} Output magnitude of the oscillator
- R Equivalent noise resistance of the varactor
- V_0 Voltage amplification of the oscillator

Even if this relationship is based on idealized values, one can derive some important parameters for the design of VCOs.

Type	Frequency range [MHz]	Output level [dBm]	Tuning voltage [V]	Phase noise @ 10 kHz	Phase noise @ 100 kHz	Phase noise @ 800 kHz [dBc/Hz]	2nd harmonic [dBc] [dBc/Hz]	Power supply [V/mA] [dBc/Hz]
VLA195	195-220	8	1 - 14	–110	–130		–8	12/20
VLA255	255-320	8	1 - 9	–100	–120		–8	12/20
VLA380	380-430	0	0 - 5	–112	–132		–8	12/20
VLA809	809-845	5	0.5 - 5	–115	–135	–152	–12	5/25
VLA925	925-960	3	1.5 - 6.5	–115	–135		–12	5/25
VLA950	950-986	3	1 - 6	–115	–135		–12	5/25
VLA1250	1250-1350	3	1 - 8	–100	–120		–12	8/25
VLA1450	1450-1550	3	1 - 8	–105	–125	–145	–12	5/25
VLA1500	1500-1650	1	1 - 8	–97	–117	–137	–12	5/25
VLA1594	1594-1669	3	1 - 6.5	–105	–125	–145	–15	5/25
VLA1750	1750-1900	1	1 - 8	–97	–117	–187	–12	8/16
VLW1800	1800-2700	1	0 - 19	–85	–105		–8	12/25
VLA2650	2650-2850	2	0 - 12	–90	–110		–12	8/20

▲ **Table 1. Technical characteristics of a selection of VCO models.**

1) The loaded Q of the resonator directly affects the phase noise; for this reason, coaxial resonators must be used in the case of very high performance requirements.

2) Low-noise oscillators require components with a low corner frequency of the flicker ($1/f$) noise. Bipolar transistors are normally used in VCOs instead of FETs. GaAs devices are not suitable, as they have a significantly higher noise corner.

3) The noise figure of the oscillator, which is internal to the switching, depends not only on the noise figure of the active component but also on the switching configuration. The setting of the capacity of the oscillator signal also influences the noise; in this, however, the current consumption must not be neglected.

One very important point, which is not taken into consideration in equation (8), is the voltage supply. Significant fluctuations can occur in the voltage supply. Unwanted modulation side bands, which lie outside the loop of the PLL are produced from these fluctuations in the bias of the VCO.

Tuning sensitivity

Tuning sensitivity describes the tuning frequency range, depending on the tuning voltage at the varactor input. The tuning sensitivity depends on the available capacity variation and is inversely proportional to the loaded quality of the resonance circuit.

The frequency dependence of the tuning sensitivity here must also be borne in mind. If this is too great, then the performance of the synthesizers is adversely affected.

Load pulling

Load pulling gives the sensitivity of the free-running VCOs compared to the load fluctuations at the VCO output. This load pulling is specified for a mismatched load with a defined VSWR (e.g. at VSWR = 2.0), where the phase angle can lie between 0° and 360°. At its simplest, this requirement may be achieved using an additional buffer amplifier. Such a buffer amplifier also improves the drive level of the VCO, which must also supply RF to the prescaler of the PLL synthesizer in addition to a mixer stage. However, a buffer amplifier increases the current required by the VCO. Because of the load associated with power amplifiers, the load pulling performance of the transmitter branch VCO can be of particular importance.

Packaging

Obviously, the design of the VCO must be such that it can be processed in modern manufacturing installations that assemble large quantities of products using SMD technology. In practice, two basic packages have been successful: the approximately square package of Figure 6 and the rectangular package shown in Figure 7.

Summary

In this article, I have attempted to present the fundamentals for the design and the use of VCOs. From what has been said, it is clear that the VCO, together with the PLL, represents an elementary unit that makes an important contribution to the design of a base station. It would therefore be sensible if the manufacturers of VCOs were also involved in the manufacture of suitable PLL components. Therefore, a later presentation is planned for PLLs. ■

References

1. H. Lobensommer, *Die Technik der modernen Mobilkommunikation*, (The Technology of Modern Mobile Communications), Franzis-Verlag.
2. H. Preibisch, *GSM Mobilfunk: Übertragungstechnik* (GSM Mobile Radio Transmission Technology), Schiele & Schön, 1994.
3. W. Boyles: “The Oscillator as a Reflection Amplifier: an Intuitive Approach to Oscillator Design,” *Microwave Journal*, June 1986.
4. F. Baberg, “Kapazitive Bauelemente für Hochfrequenzapplikationen” (Capacitive Components for Radio Frequency Applications), *HF Report*, May 1999.
5. “Microwave & RF Devices: Ceramics and Capacitors,” *Time and Frequency Products Catalogue 2000*, issued by TEKELEC TEMEX S. A., Montreuil.
6. D. R. Leeson: “A Simple Model of Feedback Oscillator Noise Spectrum,” *Proceedings of the IEEE*, 1966.
7. Ulrich Rohde, Frank Hagemeyer: “Feedback Technique Improves Oscillator Phase Noise” *Microwaves & RF*, Nov. 1998.
8. W. Schleifer, *Hochfrequenz und Mikrowellen-Meßtechnik in der Praxis* (Radio Frequency and Microwave Measurement Techniques in Practice), Hüthig.

Author information

Frank Baberg can be reached by mail at Tekelec Temex, Kapuzinerstrasse 39, 80337 Munich, Germany, by telephone at +49 8951640 or by fax at +49 8951764110.