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Noise Reduction in Transistor Oscillators: Part 1—Resonant Circuits

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Noise in oscillators can be controlled by proper choice of the resonator, supported by circuit design techniques for coupling, filtering and feedback

This series of articles discusses oscillator phase noise reduction techniques and gives specific resonant circuit solutions using lumped and distributed parameters for both frequency stabilization and phase noise reduction. These topics are covered in Part 1.

Phase noise improvement can also be achieved by appropriate low-frequency loading and feedback circuitry optimization. The feedback system incorporated into the oscillator bias circuit can provide the significant phase noise reduction over a wide frequency range from the high frequencies up to microwaves. Particular discrete implementations of a bipolar oscillator with collector and emitter noise feedback circuits will be described in Part 2. A filtering technique based on a passive LC filter to lower the phase noise in the differential oscillator will also be presented, with several topologies of fully integrated CMOS voltage-controlled oscillators using the filtering technique shown and discussed.

Part 3 concludes the series, and includes a novel noise-shifting differential VCO based on a single-ended classical three-point circuit configuration with common base, which can improve the phase noise performance by a proper circuit realization. An optimal design technique using an active element based on a tandem connection of a common source FET device and a common base bipolar transistor with optimum coupling of the active element to the resonant circuit is also presented. The phase noise in microwave oscillators can also be reduced using negative resistance compensation increasing the loaded quality factor of the oscillator resonant circuit. Finally, a new approach utilizing a nonlinear feedback loop for phase noise suppression in microwave oscillators is discussed.

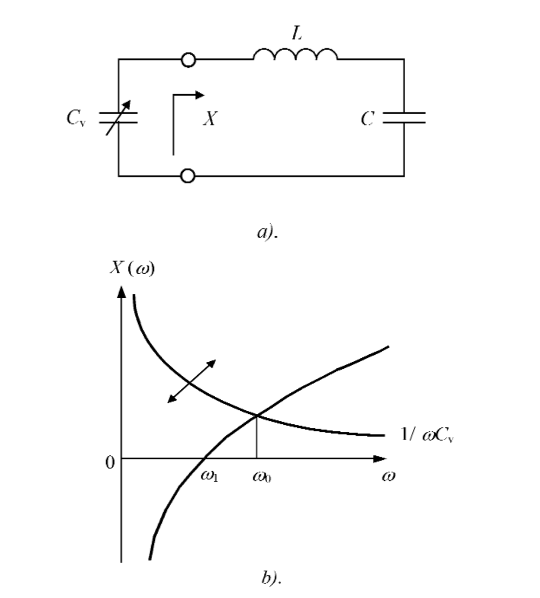

Figure 1 · An oscillator series resonant circuit with variable capacitance and its frequency performance.

Resonant Circuit Design Techniques

Microwave transistor oscillators have received much attention in recent years due to significant progress in device technology, which enables the development of the local oscillator and VCO components with reduced size, power consumption and improved overall electrical performance. For example, by using HBT devices, excellent low-noise characteristics are realized in monolithic oscillator applications up to about 20 GHz. Despite the high level of the low-frequency noise, the MESFET and HEMT devices extend the oscillator operating frequency to more than to 100 GHz. JFET devices continue to be a good choice for achieving low-noise oscillator performance at UHF band.

One way to improve the oscillator noise characteristics is to use the optimal bias condition; another way is to increase the loaded quality factor (Q) of the oscillator resonant circuit by the appropriate choice of the circuit topology. An increase in the quality factor also implies a decrease in the frequency sensitivity of the resonant circuit to the variation of its parameters.

Oscillation Systems with Lumped Elements

Figure 1(a) shows the oscillation system with a series resonant circuit representing as a simple series connection of the variable capacitance C_v and circuit inductance L and capacitance C, where C_v is considered the equivalent active device nonlinear capacitance varying over temperature, RF signal amplitude or supply voltage. The frequency behavior of such a resonant circuit is shown in Figure 1, where $\omega = 0$ and $\omega = \omega_1 = 1/\sqrt{LC}$ are pole and zero of the circuit reactance X, respectively. The oscillation frequency ω_0 of such a series resonant circuit can be found from

$$\omega_0 = \sqrt{\frac{C + C_v}{LCC_v}} \tag{1}$$

The relative sensitivity of the series resonant circuit to the variation of the variable capacitance C_v can be obtained from Eq. (1) in the form of

$$\frac{d\omega}{\omega_0} / \frac{dC_v}{C_v} = -\frac{1}{2}\frac{C}{C + C_v} \tag{2}$$

From Eq. (2) it follows that maximum sensitivity is realized for the series resonant circuit representing the only inductance and variable capacitance when $C \to \infty$ being equal to

$$\frac{d\omega}{\omega_0} / \frac{dC_v}{C_v} = -\frac{1}{2} \tag{3}$$

At the same time, minimum sensitivity of the series resonant circuit to the variation of capacitance C_v is realized under extreme condition when $C \to 0$.

Figure 2(a) shows the oscillation system with parallel resonant circuit representing a simple parallel connection of the variable capacitance C_v and circuit inductance L and capacitance C. The frequency behavior of such a resonant circuit is given by Figure 2(b), where $\omega = 0$ and $\omega = \omega_1 = 1/\sqrt{LC}$ are zero and pole of the reactance X, respectively. The oscillation frequency ω_0 of such a parallel resonant circuit can be found from

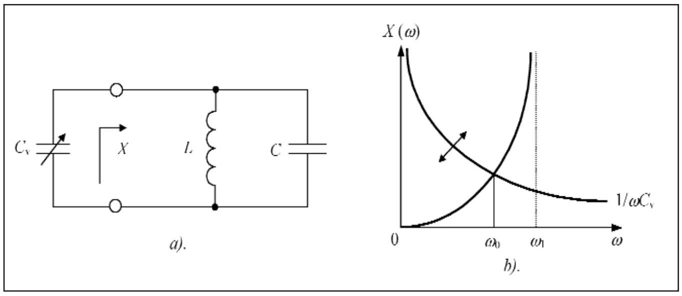

Figure 2 · Oscillator parallel resonant circuit with variable capacitance and its frequency performance.

$$\omega_0 = \frac{1}{\sqrt{L(C + C_v)}} \tag{4}$$

The relative frequency sensitivity of the parallel resonant circuit to the variation of the variable capacitance C_v can be obtained from Eq. (4) in the form of

$$\frac{d\omega}{\omega_0} / \frac{dC_v}{C_v} = -\frac{1}{2}\frac{C}{C + C_v} \tag{5}$$

From Figure 2(b) it follows that the tuning sensitivity of the parallel circuit can be easily reduced by moving pole of its reactance $X(\omega)$ to the left side, i.e., increasing the value of the resonant circuit capacitance C. Further reduction of the circuit sensitivity can be achieved by moving zero of its reactance X to the right side—towards its pole. Such a possibility can be realized for the series-parallel resonant circuit with a variable capacitance C_v shown in Figure 3(a).

The frequency behavior of such a series-parallel resonant circuit is illustrated by Figure 3(b), where $\omega = \omega_1 = 1/\sqrt{L(C + C_1)}$ and $\omega = \omega_2 = 1/\sqrt{LC}$ are zero and pole of the

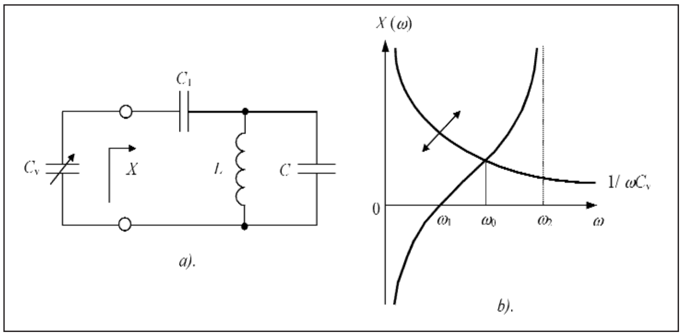

Figure 3 · Oscillator series-parallel resonant circuit with variable capacitance and its frequency performance.

reactance X, respectively. The oscillation frequency ω_0 of the series-parallel resonant circuit can be found from

$$\omega_0 = \frac{1}{\sqrt{L\left(C + \dfrac{C_v C_1}{C_1 + C_v}\right)}} \tag{6}$$

The relative frequency sensitivity of the series-parallel resonant circuit to the variation of the variable capacitance C_v can be obtained from Eq. (6) in the form of

$$\frac{d\omega}{\omega_0} \Big/ \frac{dC_v}{C_v} = -\frac{1}{2}\,\frac{C}{C + C_v}\,\frac{C_v C_1}{C(C_v + C_1) + C_v C_1} \tag{7}$$

From Figure 3(b) it follows that the frequency tuning sensitivity of the series-parallel oscillator resonant circuit can be significantly reduced by the simultaneous approaching of zero and pole by its reactance X. For the same resonant frequency ω_0, reducing the series capacitance C_1 and increasing the shunt capacitance C can do it. This contributes to the oscillator frequency stabilization by minimizing the effect of the variation of the equivalent active device nonlinear capacitance on the oscillator resonant frequency.

Oscillators with Transmission Line Resonators

It is known that the frequency sensitivity of the resonant circuit containing a short-circuited transmission line can be reduced, if instead of the uniform transmission line a non-uniform one is used, the spectrum of natural frequencies of which is non-equidistant. Figure 4 shows the resonant circuits with (a) uniform and (b) two-section transmission lines, where θ_1 is the electrical length of the low-impedanced section with the characteristic impedance Z_1, and θ_2 is the electrical length of the high-impedance section with the characteristic impedance Z_2. The main difference in the frequency properties between the oscillation system with (a) short-circuited uniform transmission line with an equidistant spectrum and (b) short-circuited two-section transmission line with a non-equidistant spectrum in the case of $Z_2 \gg Z_1$ is illustrated by Figure 5. The graphic results show that it is advisable to use a two-section line for frequency stabilization at odd resonant frequencies, including the fundamental, in the case of a low-impedance section adjacent to the variable capacitance C_v.

The sensitivity to capacitance variation of the oscillation system with a two-section line can be qualitatively evaluated according to two-by-two rapprochement, or remoteness of zeros and poles of function $Z_{in}(\omega)$ along a frequency axis. The input impedance of short-circuited two-section transmission line provided can be written by

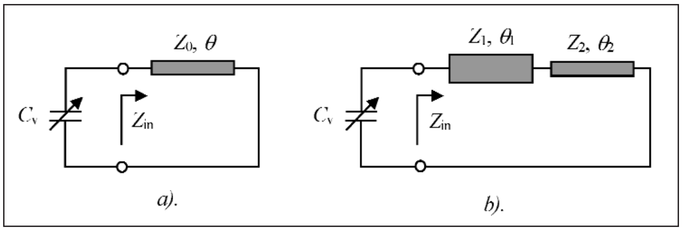

Figure 4 · Oscillation systems with (a) uniform and (b) two-section transmission line.

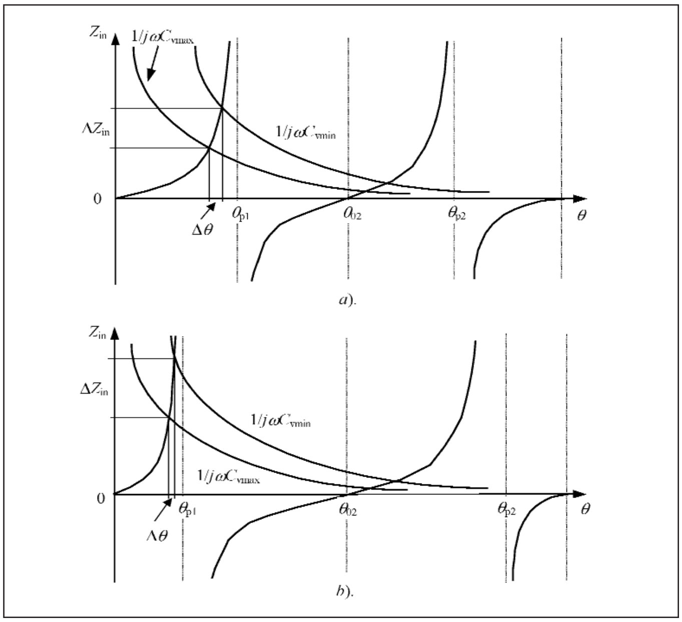

Figure 5 · Frequency dependencies of an oscillation system with a two-section transmission line.

$$Z_{in} = jZ_1 \frac{1 + M}{1 - M\tan^2\dfrac{\theta}{2}} \tan\frac{\theta}{2} \tag{8}$$

where $M = Z_2/Z_1$ which is the characteristic impedance ratio. From Eq. (8) it follows that the two-section transmission line has its zeros at $\theta = k\pi$ and poles at $\theta = k\pi - 2\tan^{-1}(1/\sqrt{M})$, $k = 0, \pm1, \pm2$, etc. The first pole of the function $Z_{in}(\theta)$ for the fundamental resonant frequency operation at $M = 20$ is displaced substantially to the left and becomes equal to $\theta_{p1} = 25.2°$. In this case, the same frequency deviation corresponds to significantly larger capacitance variation in comparison with the resonant circuit with uniform transmission line. This effect contributes to the increase of the sensitivity of the oscillator phase characteristic and loaded quality factor resulting in the reduction of the oscillator phase noise.

As it follows from an analysis of the parallel feedback or negative resistance oscillator noise spectrum, the oscil-

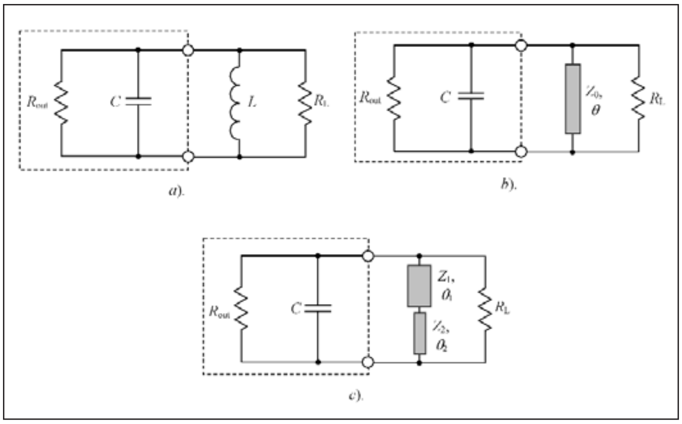

Figure 6 · Equivalent parallel oscillator resonant circuits with (a) lumped parameters and (b, c) distributed parameters.

lator phase noise is inversely proportional to the square of the loaded quality factor, Q_L of the oscillator resonant circuit. Consequently, for a given load conductance G_L and output power P_{out}, the value of Q_L has a significant effect upon the oscillator phase noise level. Generally, the loaded quality factor Q_L of the oscillator parallel resonant circuit can be defined by

$$Q_L = \frac{\omega_0 S_0}{2G_L} \qquad (9)$$

where

$$S_0 = \left.\frac{\partial(B_{out} + B_L)}{\partial\omega}\right|_{\omega=\omega_0}$$

is the susceptance slope parameter or susceptance sensitivity of the oscillator parallel resonant circuit at the resonant frequency ω_0, B_{out} and B_L are the equivalent output active device and load susceptances, respectively [1]. In Eq. (8) it is assumed that the real part of the equivalent output active device and load admittances are frequency independent in the vicinity of the steady-state oscillation conditions. Such a simplification has usually been accepted as a basis for oscillation analysis. The similar expression can be written through the impedance parameters for the series resonant circuit. Hence, in order to improve the oscillator loaded quality factor, it is necessary to increase the susceptance sensitivity S_0.

In order to compare the oscillators with lumped and distributed parameters for the purpose of the best phase noise characteristics, it is sufficient to determine the susceptance sensitivity S_0 for each oscillator circuit with equal capacitances. For a simple parallel lumped resonant circuit shown in Figure 6(a),

$$S_0 = \frac{\partial}{\partial\omega}\left(\omega C - \frac{1}{\omega L}\right) = C\left.\left(1 + \frac{1}{\omega^2 LC}\right)\right|_{\omega=\omega_0} = 2C \qquad (10)$$

The oscillator parallel resonant circuit with uniform transmission line is shown in Figure 6(b). Since the ratio between the transmission line electrical length θ and the frequency ω is defined as

$$\theta = \omega l \sqrt{\varepsilon_r} / c \qquad (11)$$

where l is the length of the transmission line, ε_r is the dielectric constant, c is the free-space velocity of light, the following condition exists:

$$\frac{\partial\theta}{\partial\omega} = \frac{\theta}{\omega} \qquad (12)$$

Then, by taking into account that $C = 1/\omega_0 Z_0 \tan\theta$ at the resonant frequency, the sensitivity S_0 for the parallel resonant circuit with a lossless uniform transmission line is obtained by

$$S_0 = \frac{\partial}{\partial\omega}\left(\omega C - \frac{1}{Z_0 \tan\theta}\right) = C\left.\left(1 + \frac{1}{\omega C Z_0 \sin^2\theta}\right)\right|_{\omega=\omega_0} = C\left(1 + \frac{2\theta}{\sin 2\theta}\right) \qquad (13)$$

Similarly, the susceptance sensitivity S_0 for the parallel resonant circuit with lossless two-section transmission line of equal section lengths $\theta_1 = \theta_2 = \theta/2$ shown in Figure 6(c) can be written as

$$S_0 = \frac{\partial}{\partial\omega}\left.\left(\omega C - \frac{1}{Z_1 \tan\frac{\theta}{2}} \frac{1 - M\tan^2\frac{\theta}{2}}{1+M}\right)\right|_{\omega=\omega_0} = C\left(1 + \frac{\theta}{\sin\theta}\frac{1 + M\tan^2\frac{\theta}{2}}{1 - M\tan^2\frac{\theta}{2}}\right) \qquad (14)$$

Figure 7 shows the frequency dependencies of the normalized sensitivity S_0/C for different types of the oscillator circuits. For a lumped resonant circuit, $S_0/C = 2$ that means that it is frequency independent. For the resonant circuits with transmission lines, the value of sensitivity S_0 for a given capacitance C can be increased significantly with the appropriate increase of Q_L. If $\theta = 40°$, $S_0/C = 2.42$ for the resonant circuit with uniform transmission line, whereas the use of the two-section transmission line with $M = 5$ results in $S_0/C = 6.35$. In the case of $\theta = 24°$, by using the two-section transmission line with $M = 20$, it is possible to increase the value of S_0/C by more than ten times. For the same values of S_0/C, the use of the two-sec-

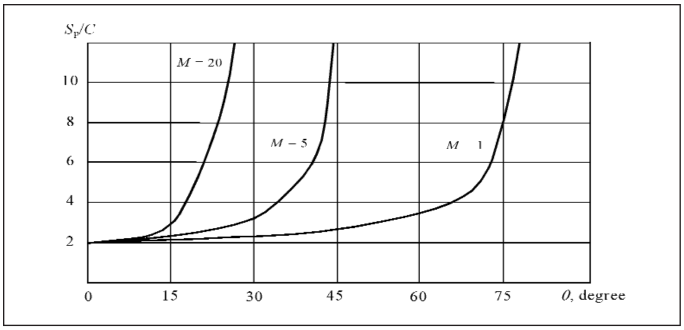

Figure 7 · Frequency dependencies of the normalized susceptance sensitivity S_0/C for different characteristic impedance ratio *M*.

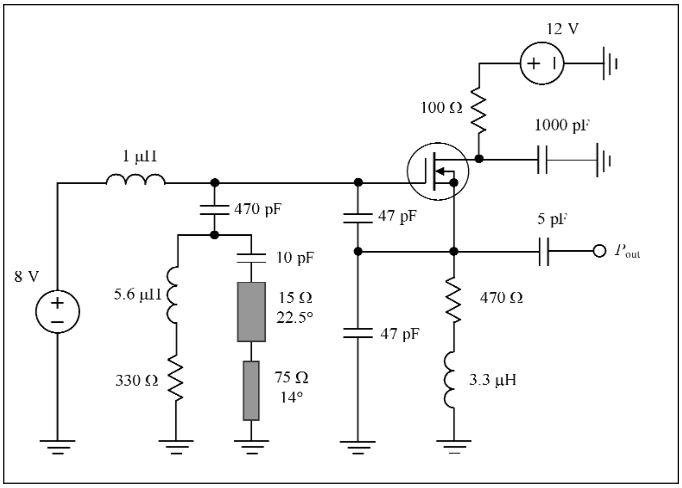

Figure 8 · Equivalent circuit of the MOSFET oscillator with two-section line.

tion transmission line with high value of *M* enables a reduction the total electrical length of the transmission line by three-fourths.

The verification of the theoretical assumptions regarding the improvement of the oscillator noise characteristics in the case of the resonant circuit with transmission lines was made on the basis of the MOSFET microwave oscillator included as an example in the circuit simulator Serenade 7.5 [2]. The equivalent circuit of the MOSFET oscillator with two-section transmission line connected instead of lumped inductance into the gate circuit is shown in Figure 8. The oscillation frequency is 400 MHz, and the output power is 11 dBm.

Figure 9 shows the results of computer simulation of the oscillator phase noise characteristics, where the phase noise level of the oscillator with ideal lumped inductance L = 22 nH is the same as those of the oscillator using ideal uniform transmission line with the characteristic impedance Z_0 = 75 ohms and the electrical length θ = 37° (solid line). The phase noise characteristics can be improved by decreasing the characteristic impedance of the transmission line according to the theory predictions. So, the phase noise level reduces by approximately 7 to 8 dB up to the frequency offset f_m = 10 kHz for the value of Z_0 = 15 ohms (dotted line). With such a low characteristic impedance of the uniform transmission line, to maintain the same value of the input impedance of its short-circuited configuration, the value of its electrical length must be increased to θ = 74°. However, the same improved phase noise level can be obtained by using a two-section transmission line with half the total electrical length: Z_1 = 15 ohms, θ = 23° and Z_2 = 75 ohms, θ = 14°. In practice, it is difficult to realize an inductor with a high value of Q, especially at microwaves, which leads to the substantial deterioration of the oscillator noise characteristics. Therefore, it is advisable to use the transmission lines with as short as possible electrical length for monolithic microwave integrated circuit oscillators.

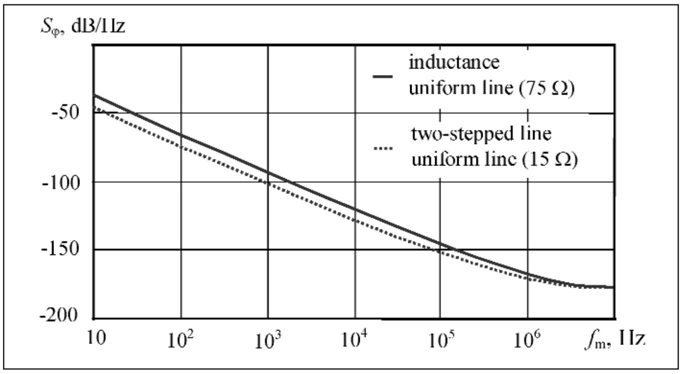

Figure 9 · MOSFET oscillator phase noise characteristics.

From Eqs. (11) to (13) it follows that, with the significant increase in the length of the transmission line, the sensitivity S_0 and, consequently, loaded quality factor Q_L can also be increased significantly. Figure 10 shows a schematic diagram of a microwave push-pull oscillator containing the open-circuit transmission line with the characteristic impedance Z_r and geometrical length l_r as a resonator [3]. The microwave oscillator consists of two identical FET devices whose gates are interconnected by a transmission line with the characteristic impedance Z_0 and electrical length θ. The length of this transmission line is chosen such that the two FET devices resonate with each other [4]. A negative conductance at the desired

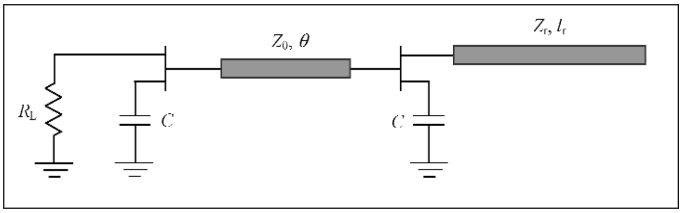

Figure 10 · Block diagram of microwave oscillator with open-circuited transmission line.

oscillation frequency is generated by a capacitive source positive feedback. The oscillator outputs are taken from the drains of two FETs for the load resistance R_L and transmission-line resonator, respectively, providing the required output port isolation. Stable oscillations at around 20 GHz with output power of 0 dBm were observed at gate and drain bias voltages of 0 and 3 V, respectively.

Three coaxial cables of different lengths were used as the open-circuited resonators. Spurious oscillations at unwanted signal frequencies were not observed in any of these cases. Table 1 shows a significant 10 dB reduction in phase noise with the connection of a transmission-line resonator of the 10 cm length compared with open-circuited output port. Further improvement in phase noise can be achieved by increasing the resonator length. The similar results were observed when short-circuited transmission lines were used as the resonators instead of the open-circuited transmission lines [3].

To minimize the sideband noise in the oscillators, it is necessary to provide an optimum ratio of the oscillator loaded quality factor Q_L to its unloaded quality factor Q_0 [5, 6]. Figure 11 shows a general equivalent oscillator model. It is assumed that the thermal noise can be modeled as a single noise source FkT at the input of the active device (operating as a current source in a common case), where F is the noise figure (generally dependent on the input impedance of the active device). The feedback resonator represents a series LCR resonant circuit with an equivalent loss resistance R. The active device has zero output admittance, which can be achieved by using a switched-mode operation mode, and known input impedance. It should be noted that such an approach could be applied only when the thermal (additive) noise is the major noise source. Therefore, the obtained results cannot be valid for small frequency offsets where low frequency flicker noise dominates.

The ratio of a single sideband power of the phase noise in a 1 Hz bandwidth at the offset frequency fm to the total signal power at the carrier frequency f_0 is defined by

$$L(f_m) = \frac{FkT}{8Q_0^2 (Q_L/Q_0)^2 \left(1-(Q_L/Q_0)\right)^2} \left(\frac{f_0}{f_m}\right)^2 \qquad (15)$$

Cable length, cm	Number of wavelengths, N	Cable insertion loss, dB	Phase noise improvement, dB
10	23	0.35	10
50	89	1.7	18
100	171	3.2	21

Table 1 · Measured parameters of the oscillator with open-circuited cable resonators (3).

where $Q_0 = 2pf_0L/R$ is the quality factor of the LCR resonant circuit and $Q_L = 2pf_0L/(R + R_{in})$ is the loaded quality factor, where R_{in} is the active device input resistance.

Equation (15) has minimum being differentiated with respect to Q_L/Q_0 for constant Q_0 and F when

$$\frac{\partial L(f_m)}{\partial (Q_L/Q_0)} = 0 \qquad (16)$$

As a result, a minimum single sideband phase noise power occurs when $Q_L/Q_0 = 1/2$ [5]. Therefore, Eq. (15) simplifies to

$$L(f_m) = \frac{2FkT}{Q_0^2} \left(\frac{f_0}{f_m}\right)^2 \qquad (17)$$

If the oscillator operates in a switched-mode high-efficiency operation with zero low output admittance (the device output is represented by a lossless switch), a minimum single sideband phase noise power occurs when $Q_L/Q_0 = 2/3$ [6]. Consequently, Eq. (15) can be re-written by

$$L(f_m) = \frac{27FkT}{32Q_0^2} \left(\frac{f_0}{f_m}\right)^2 \qquad (18)$$

The same results obtained by Eqs. (17) and (18) for minimum phase noise level are achieved for the transmission line oscillator shown in Figure 11(b), where Z_0 is the characteristic impedance, a is the attenuation coefficient, b is the phase constant and l is the length of the

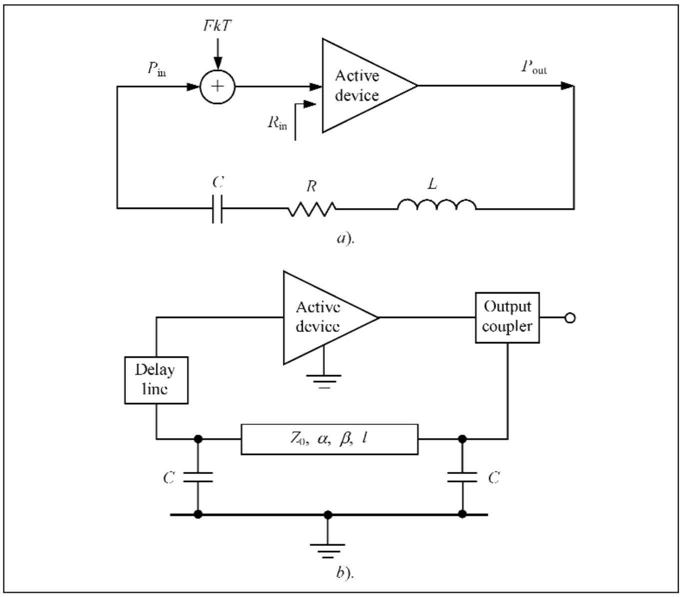

Figure 11 · Equivalent circuits of LC and transmission line oscillators.

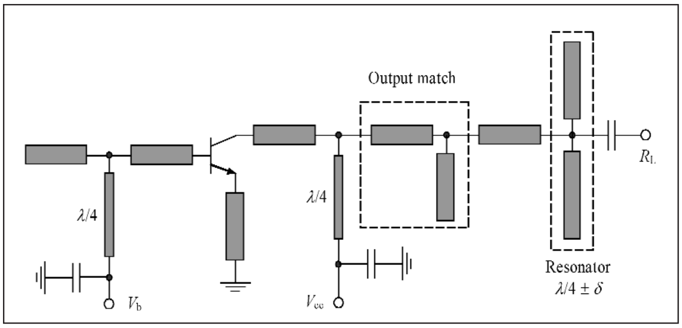

Figure 12 · Circuit schematic of bipolar oscillator with parallel resonator (9).

transmission line, respectively [7, 8]. It is assumed that the active device and transmission line resonator have both inputs and outputs matched to Z_0. The delay line is necessary to provide the phase compensation. In this case, for the small values of αl <0.05 and $f_m/f_0 \ll 1$, the loaded quality factor is defined by

$$Q_L = \frac{\pi}{1 + 2\alpha l\left(\pi f_0 C\right)^2}\left(\pi f_0 C\right)^2 \tag{19}$$

From Eq. (19) it can be seen that as a value of the shunt capacitance C is increased, the loaded quality factor Q_L increases to a limiting factor of

$$Q_0 = \frac{\pi}{2\alpha l} \tag{20}$$

which is defined as an unloaded quality factor.

A microstrip oscillator with oscillation frequency of 1.49 GHz based on a bipolar transistor NE68135 was fabricated using a RT Duroid substrate with dielectric permittivity of $\varepsilon_r = 10.2$. To deliver power to the external load, a 3 dB Wilkinson divider was used. Phase compensation is accomplished by means of a short length of transmission line which is fine tuned using a trimming capacitor. For a transmission line with $\alpha l = 0.019$ resulting in the unloaded quality factor $Q_0 = 83$ and noise figure $F = 3$ dB, Eq. (17) gives the theoretical limit in terms of sideband noise power of approximately –102.6 dBm/Hz at 10 kHz offset, which is only 1.7 dB greater than the measured phase noise power level. An increase in the noise level was observed if the optimum condition was not met.

To improve the oscillator phase noise characteristic, the parallel circuit resonator based on the open-circuited transmission lines can be added to the oscillator resonant circuit at its output, as shown in Figure 12 [9, 10]. In this case, the resonator is simply a combination of two open-ended stubs having lengths of $l_\pm = \lambda/4(1 + \Delta\omega/\omega_0) \pm \delta$, where λ is the wavelength corresponding to the oscillator resonant frequency ω_0. The input impedance of the lossless resonator with one end terminated by the load resistance R_L can be written as

$$Z_r = R_L \frac{1}{1 + j\left(\tan\theta_- + \tan\theta_+\right)} \tag{21}$$

where $\theta_\pm$ are the electrical lengths of the open-ended stubs corresponding to their lengths $l_\pm$, respectively. The two tangent functions in Eq. (21) cancel each other at the resonant frequency ω_0 when $\Delta\omega = 0$ due to their odd properties. However, when $\Delta\omega \neq 0$, the electrical lengths $\theta_\pm$ are not symmetrical relative to 90°. As a result, when $\Delta\omega$ increases in a forward positive direction, both tangent functions demonstrate a sharp increase simultaneously. According to Eq. (21), the susceptance sensitivity of such a parallel resonant circuit will increase rapidly when $\Delta\omega$ is sufficiently small, thus resulting in a significant increase of the resonator loaded quality factor. The unloaded quality factor of the resonant circuit depends on the attenuation coefficient of the stubs in accordance with Eq. (20), where $l = l_+ + l_-$, and can reach a value of 205 using a 10 mil alumina substrate or a value of 77 using an 80 mm GaAs substrate.

The hybrid 18 GHz oscillator, using AlGaAs/InGaAs HBT active device with a resonator fabricated on alumina substrate with $\delta = \lambda/30$ and a loaded quality factor of 45, exhibited an output power of 10.3 dBm and an efficiency of 19.3% at a supply voltage of 2 V with a phase noise level of –120 dBc/Hz at 1 MHz offset [9]. The effect of the resonator resulted in about 5 dB phase noise reduction. In addition, the oscillator pulling factor defined as $(f_{max} - f_{min})/f_0$, where f_{max} and f_{min} are the maximum and minimum oscillation frequencies respectively, is improved by factor of 6 when the load phase angle is changed over the range of 2π. The monolithic 38 GHz oscillator, using the similar HBT technology with a resonator fabricated on GaAs substrate with $\delta = \lambda/36$ and a loaded quality factor of 34, achieved an output power of 11.9 dBm and an efficiency of 10% at a supply voltage of 3.2 V with a phase noise level of –114 dBc/Hz at 1 MHz offset [10]. The effect of the resonator was about 5 dB phase noise reduction.

Coming in Parts 2 and 3

This series of articles will continue in the next two issues of *High Frequency Electronics*. Part 2 will discuss low frequency loading and feedback optimization to minimize 1/*f* noise, and present filtering techniques for noise reduction. Part 3 will conclude the series with discussions of specific noise management techniques, including noise shifting using bias and circuit topology, impedance matching for noise, and a nonlinear feedback loop technique for cancellation of noise components.

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