LOW NOISE OSCILLATORS

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ABSTRACT

A large number of papers have been published on low noise oscillators however they are usually very specific to the particular application. This paper will describe a set of design rules which are general and can be used to produce oscillators with very low noise performance where both additive (thermal noise) and Flicker noise are considered. Linear theories will be described which accurately describe the noise performance of resonator type oscillators. The limits set by varactor diodes on the noise performance will be described and the noise degradation caused by open loop phase error will be shown. Seven design examples will be demonstrated which show close correlation with the theory.

OSCILLATOR NOISE THEORIES

The model chosen to analyse an oscillator is extremely important because the conclusions drawn can often vary. For this reason two models are presented here. Each model can produce different results as well as improving the understanding of the basic model. Both an equivalent circuit model and a block diagram model will be described. We will start with the equivalent circuit model originally used by the author to design oscillators with the potential for high efficiency and easy analysis. These models are used to describe the effects of thermal noise. Flicker noise effects are described later.

EQUIVALENT CIRCUIT MODEL

The model is shown in Figure 1 and consists of an amplifier with two inputs with equal input impedance, one for noise (Vin_2) and one as part of the feedback resonator.

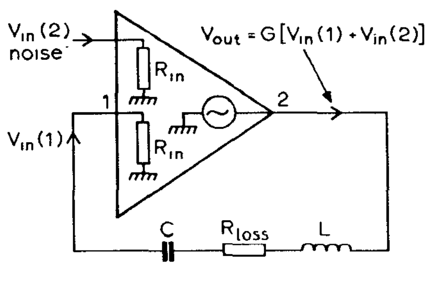

Figure 1 Equivalent circuit model of oscillator

The feedback resonator is modelled as a series inductor capacitor circuit with an equivalent loss resistance Rs which defines the unloaded Q (Q_o) of the resonator as ωl/rs.

The operation of the oscillator can best be understood by injecting white noise at the input Vin_1 and calculating the transfer function while incorporating the usual boundary condition of $GB_o = 1$ where G is the limited gain of the amplifier when the loop is closed and B_o is the feedback coefficient at resonance where $fo = 1/2\pi\sqrt{LC}$.

The amplifier model had zero output impedance a known input impedance and a resonant positive feedback network. The zero output impedance of the amplifier was used because the design of highly power efficient oscillators was of interest. This also reduced the pulling effect of the load. The zero (low) output impedance of the amplifier is achieved by using a switching output stage. In fact the same theory and conclusions can be obtained for oscillators with a finite output impedance so for convenience this will be left at zero.

Input Vin_2 is used at the input of the amplifier to model the effect of noise. In a practical circuit the noise would come from the amplifier. The noise voltage Vin2 is assumed to be added at the input of the amplifier and was dependent on the input impedance of the amplifier, the source resistance presented to the input of the amplifier and the noise figure of the amplifier. In this analysis, the noise figure under operating conditions, which takes into account all these parameters, was defined as F.

The circuit configuration is very similar to an operational amplifier feedback circuit and therefore the voltage transfer characteristic can be derived in a similar way.

If the total RF power dissipated in the complete oscillator is limited (this is useful if minimum sideband noise is required for minimum DC input power) then the following equation can be derived.

$$L_{fm} = \frac{GFkT}{8 Q_L^2 P_{rf}} (f_o/\Delta f)^2$$

where G is the voltage gain, F is the noise Figure, k is Boltzmans constant, T is the operating temperature, Q_L is the loaded Q of the resonator Q_o is the unloaded Q of the resonator and P_{rf} is the power dissipated in the resonator and amplifier.

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Eqn. 1 shows that L_{fm} is inversely proportional to P_{rf} and that better noise is thus obtained for higher feedback power. This is because the absolute value for the sideband power does not vary with the total feedback power. It should be noted that P_{rf} is the total power in the system excluding

the losses in the amplifier, from which:

P_{rf} = (DC input power to the system) x efficiency

For minimum noise the noise figure (F), and the value of G/Q^2_L should be as small as possible. It should be noted, however, that F, G and Q_L are directly related to each other and thus cannot be varied independently.

OPTIMISATION FOR MINIMUM PHASE NOISE

The first equations is now examined to see which parameters are interrelated so that the equation can be optimised for minimum phase noise. At resonance the gain of the amplifier is $1/\beta_o$ and Δf is 0, then as $G = 1/(1-Q_L/Q_o)$

$$L_{fm} = \frac{FkT}{8Q_o^2 (Q_L/Q_o)^2 (1-Q_L/Q_o) P_{rf}} (f_o/\Delta f)^2$$

This noise equation is minimum when

$$\frac{dL_{fm}}{dQ_L/Q_o} = 0$$

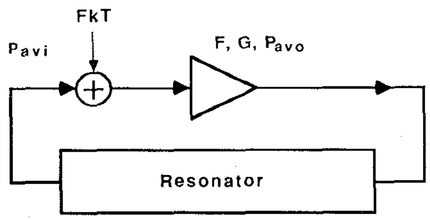
Minimum noise therefore occurs when $Q_L/Q_o = 2/3$. To satisfy $Q_L/Q_o = 2/3$, the voltage insertion loss of the resonator is 1/3 which sets the amplifier voltage gain to 3.

It is extremely important to use the correct definition of power (P), as this affects the values of the parameters required to obtain optimum noise performance.

If the power is defined as the power available at the input of the amplifier P_{avi} then the gain (G) will disappear from the equation. At first glance it would appear that minimum noise occurs when Q_L is made large and hence tends to Q_o. However this would require that the amplifier gain and output power both tend to infinity.

If we take the general oscillator model shown in figure 2 where we now define the limited output power as the power available at the output then the following equation can be derived where it is assumed that $G = (1-Q_L/Q_o)^2$ where G is now the transducer power gain.

$$L_{fm} = \frac{FkT}{8Q_o^2 (Q_L/Q_o)^2 (1-Q_L/Q_o)^2 P_{avo}} (f_o/\Delta f)^2$$



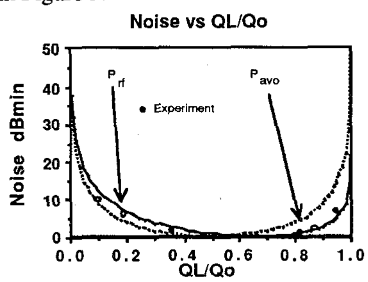
The minimum of the equation occurs when Q_L/Q_o=1/2. It

should be noted that P_{avo} is constant and not related to Q_L/Q_o. The power available at the output of the amplifier is different from the power dissipated in the oscillator, but by chance is close to it. Parker has shown a similar optimum for SAW oscillators.This should be compared with the high efficiency model where the power is defined as the total power dissipated in the resonator and the impedances of the amplifier (P_{rf}) which is useful if highly efficient oscillators are required, then the optimum condition occurs at $Q_L/Q_o = 2/3$. The equation for the noise performance becomes:

$$L_{(fm)} = \frac{FkT}{8Q_o^2 (Q_L/Q_o)^2 (1-Q_L/Q_o) P_{rf}} (f_o/\Delta f)^2$$

where the last term in the denominator has now changed from $(1-Q_L/Q_o)^2$ to $(1-Q_L/Q_o)$.

These results are most easily compared graphically as shown in Figure 3. Measurements of noise variation with Q_L/Q_o have been demonstrated using a low frequency high efficiency oscillator where the power is defined as P_{rf} and these are also included in Figure 3.



The difference in the noise performance and the optimum operating point predicted by the different definitions of power is small. However care needs to be taken in the P_{avo} definition if it is necessary to know the optimum value of the source and load impedance. For example if P_{avo} is fixed it would appear that optimum noise performance would occur when $R_{out} = R_{in}$ because P_{avo} tends to be very large when R_{out} tends to zero. This is not the case when P_{rf} is fixed.

It should be noted that the noise factor is dependent on the source impedance presented to the amplifier and that this will change the optimum operating point dependent on the type of active device used. If the variation of noise performance with source impedance is known then this can be incorporated to slightly shift the optimum value of Q_L/Q_o

FLICKER NOISE TRANSPOSITION

The theory and optima described earlier applied in the region where thermal (additive) noise is the major noise noise source. This is where the noise in the oscillator falls off at a $1/\Delta f^2$ rate In fact for Flicker noise (modulation noise) it is often the case that Q_L should be as high as possible. However the modulation mechanisms are not well understood and therefore this group have made measurements, presented at the MTT conference on June 1990, on the cross correlation between the baseband noise on the drain and the AM and PM components transposed onto the carrier. Further measurements have now been made which include noise measurements on the gate as well. These measurements have shown why low frequency feedback does not greatly improve the Flicker noise

The system is shown in Figure 4. A low noise reference signal is passed through the amplifier and both the gate and drain noise are measured. The AM and PM components of the noise are measured using the delay line discriminator either in phase for AM or at 90 degrees for PM. By measuring the direct noise at points on the device and the demodulated noise simultaneously on a digitising oscilloscope the cross correlation functions can be derived. Further by making multiple measurement at different and therefore the noise sources isolated. From these measurements new techniques for Flicker noise reduction can be devised. The reason why a number of the techniques do not work very well can also be explained. A typical cross correlation function is shown in Figure 5.

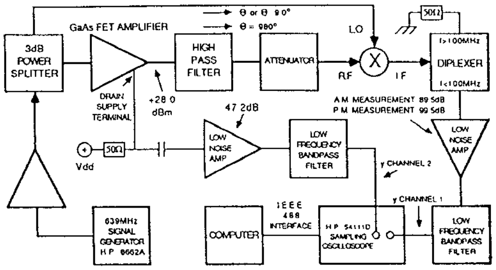


Figure 4. Flicker Noise Measurement System

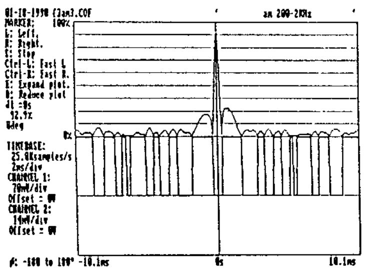


Figure 5. AM/Drain Cross Correlation Function

Oscillator Designs

A number of low noise oscillators have been built and these are shown in the following figures.

A 150 MHz Inductor Capacitor oscillator is shown in Figure 6 where part of the series inductor and the shunt capacitors are used to provide impedance matching to produce the correct value of Q_L/Q_O. The unloaded Q of the resonator was 300 and the noise performance at 25 KHz offset was 136 dBc.

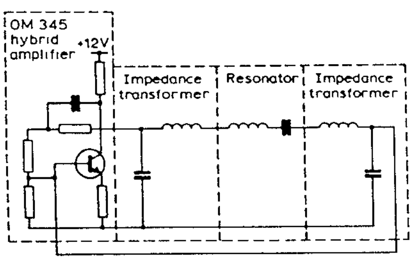


Figure 6.

A 262 MHz SAW oscillator using an STC resonator with an unloaded Q of 15,000 was built by Curley and Everard in 1987. This oscillator was built using low cost components and the noise performance was measured to be better than -130dBc at 1 Khz where the Flicker noise corner was around 1KHz. The oscillator consists of a resonator with an unloaded Q of 15,000, impedance transforming and phase shift networks and a hybrid amplifier as shown in Figure 7.The phase shift networks are designed to ensure that the circuit oscillates on the peak of the amplitude response of the resonator and hence at the maximum in the phase slope ($d\Phi/dw$). The oscillator will always oscillate at phase shifts of N*360 degrees where N is an integer but if this is not on the peak of the resonator characteristic, the noise performance will degrade with a Cos^4 relationship.

This noise performance was in fact limited by the measurement system and new measurements of identical oscillators are about to be made. Parker has demonstrated some excellent 500 Mhz SAW oscillator designs where he has reduced the Flicker noise in the resonators and operated at high power to obtain -140 dBc at 1Khz offset however the noise appears to be Flicker noise limited over the whole band.

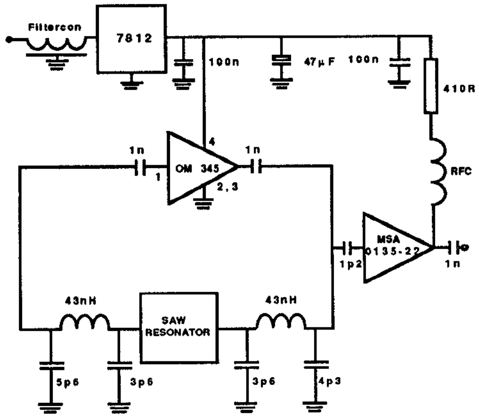


Figure 7 Low noise 262 MHz SAW oscillator

A transmission line oscillator is shown in Figure 8. Here the resonator operation is similar to that of an optical Fabry Perot and the shunt capacitors act as mirrors. The value of the capacitors are adjusted to obtain the correct insertion loss and Q_L/Q_o calculated from the loss of the transmission line.

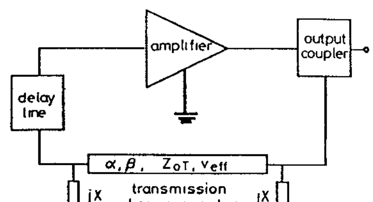


Figure 8. Transmission line oscillator

The resonator consists of a low-loss transmission line (length L) and two shunt reactances of normalised susceptance jX. If the shunt element is a capacitor of value C then $X = 2\pi fCZo$. The value of X should be the effective susceptance of the capacitor as the parasitic series inductance is usually significant. These reactances can also be inductors, an inductor and capacitor, or shunt stubs. If ZoT = Zo, where ZoT is the resonator line impedance and Zo is the terminating impedance, then S21 is given by the following equations

$$S21 = 4\,\Gamma/\{(1+jX) - X2(1-\Gamma 2)\}$$

where $\Gamma = exp\{-(\alpha + j\beta)L\}$

α is the attenuation coefficient of the line β is the phase constant of the line. For small αL (< 0.05) and $\Delta f/fo << 1$, the following properties can be derived for the first resonant peak (fo) of the resonator where df = f - fo,

$$fo = (veff/2L)\{\ 1 + (1/\pi)tan{-1}(2/X)\},$$

$$S21(\delta f) = S21(0)/\{1 + j2QL(\Delta f/fo)\}$$

$$S21(0) = (1-QL/Qo) = 1/\{1 + (\alpha L/2)X2\}$$

$$QL = \pi S21(0)X2/4$$

$$Qo = \pi/2\alpha L$$

From these equations it can be seen that the insertion loss and the loaded Q factor of the resonator are interrelated. In fact as the shunt capacitors (assumed to be lossless) are increased the insertion loss approaches infinity and QL increases to a limiting value of $\pi/2\alpha L$ which we have defined as Qo. It is interesting to note that when S21 = 1/2, QL = Qo/2.

A similar design using a copper L band Helical resonator with an unloaded Q of 600 is shown in Figure 9. The helix produces both the central line and the shunt inductors; where the shunt inductors are formed by placing taps 1mm away from the end to achieve the correct Q_L/Q_0.The equations which describe this resonator are identical to those used for the 'Fabry Perot resonator describedearlier except for the fact that X now becomes $-Z_0/2\pi fl$ where l is the inductance and L is the effective length of the transmission line. As the Q becomes larger the value of the shunt l becomes smaller eventually becoming rather difficult to realise.

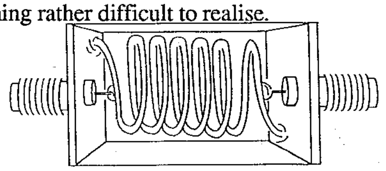


Figure 9.

POWER LIMITATIONS CAUSED BY THE VARACTORS: The noise performance of a broad tuning range oscillator is usually limited by the Q and the voltage handling capability of the varactor as has been described by Underhill. However this has not been applied to the oscillator under optimum operating conditions. If it is assumed that the varactor diode limits the unloaded Q of the total circuit, then it is possible to obtain useful information from a simple power calculation. If the varactor is assumed to be a voltage controlled capacitor in series with a loss resistor (rs). The power dissipated in the varactor is:
$P = (Vrs)^2/rs$ The voltage across the capacitor V_c in a resonator is $Vc = QVrs$. Therefore the power dissipated in the varactor is $Pv = Vc^2/Q^2 rs$.
The noise power in oscillators is proportional to $1/PQ_0^2$. Therefore the figure of merit (Vc^2/rs) should be as high as possible and thus the varactor should have large voltage handling characteristics and small series resistances. However the definition of P and the ratio of loaded to unloaded Q are important and these will alter the effect of the varactor on the noise performance. If we set the value of Q_L/Q_0 to the optimum value where again the varactor defines the unloaded Q of the resonator then the noise performance of such an oscillator can be calculated directly from the voltage handling and series resistance of the varactor. If the value of QL/Qo is put in as 2/3 then:

$$Lfm = \frac{9FkT.rs}{16Vc^2}(fo/\Delta f)2$$

If we take a varactor with a series resistance of 1 ohm which can handle an rf voltage of 0.25 volts rms at a frequency of 1 GHz, then the noise performance at 25 Kilohertz offset can be no better than -97 dBc for a noise figure of 3dB. This can only be improved by reducing the tuning range by coupling the varactor into the tuned circuit more lightly, or by switching in tuning capacitors using PIN diodes, or by improving the varactor. The voltage handling capability can be improved by using two back to back varactors although care needs to be taken to avoid bias line currents.

Two tunable resonators are shown in Figures 10 and 11. Because the transmission line has a low impedance in the middle of the line at the operating frequency the bias resistor can be made low impedance. This means that the low frequency noise caused by the bias resistor ($en^2 = 4kTBr_b$) which could cause unwanted modulation noise via the varactors can be kept low.

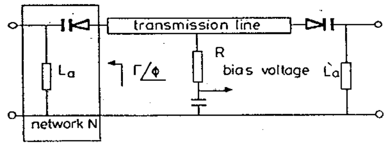


Fig.10. 3-6 Ghz transmission line resonator with constant Q_L/Q_0

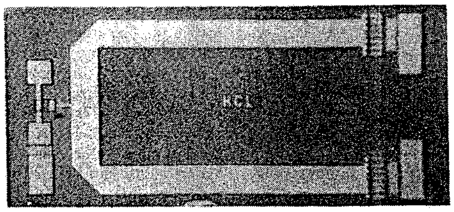


Fig.11. 8-10 Ghz MMIC resonator

DEGRADATION OF PHASE NOISE WITH OPEN LOOP PHASE ERROR

If the open loop phase error of an oscillator is not close to Nx360 degrees the effective Q is greatly reduced. It can be shown theoretically and experimentally that the noise performance degrades both in the additive noise region and the Flicker noise region by $Cos^4\theta$ where θ is the open loop phase error. For high Q dielectric resonators with for example a Q of 4,000; an offset frequency of 1 MHz at 10 GHz can produce a noise degradation of over 10 dB. A typical plot of the noise degradation with phase error for a high Q oscillator is shown in Figure 12. This shows results for oscillators using both Silicon and GaAs as the active devices. The black dots are for the GaAs oscillator and the white dots are for silicon oscillator.

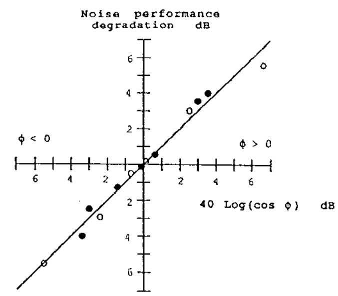


Figure 12.

CONCLUSIONS:

The theory required to design low noise oscillators using most of the commonly used resonators has been described and a number of experimental circuits presented. The major parameters which degrade the noise performance have been described.

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