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BROAD TUNING MICROWAVE OSCILLATORS UTILISING MULTILAYER TECHNOLOGY AND SiGe DEVICES

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Abstract - **This paper describes the design of a low phase noise 3.27 – 5.27GHz VCO demonstrating phase noise performance varying from –96 to –102dBc/Hz at 100kHz offset over the whole band. The VCO incorporates a silicon germanium transistor as the active device. The resonator consists of a printed electrically short transmission line with printed shunt inductors at either end. Two varactor diodes are coupled in parallel to the centre of the resonator. The amplifier is placed on the other side of a 3-layer board and coupled to the resonator using capacitive vias. The oscillator operates at V_{CE} = 2V and I_C = 30mA. The design of a broadband (3 – 6GHz) negative group delay equaliser [4] capable of cancelling the phase shift in the amplifier is also described although this was not used in the final design due to the high insertion loss.**

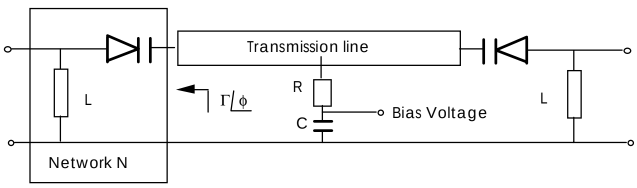
Keywords - **Voltage controlled oscillators, VCO, low phase noise.**

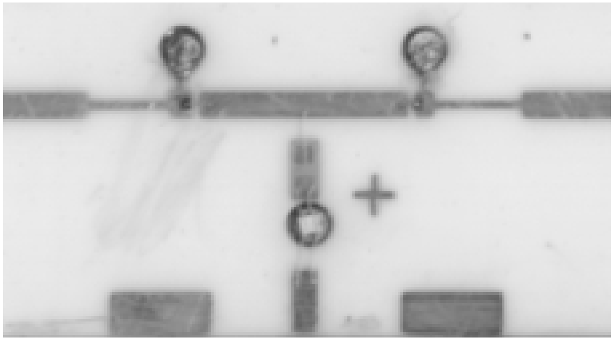
I INTRODUCTION

Broad tuning VCOs are important in most communications systems. As the frequency goes up it becomes more difficult to maintain the optimum operating conditions for oscillation with low phase noise over a broad frequency range. These parameters include: amplifier gain and noise figure; resonator loaded and unloaded Q and hence insertion loss ($S_{21} = 1-Q_L/Q_0$); the closed loop phase shift [3] and the maintenance of the correct phase shift and gain over the whole operating frequency.

II RESONATORS

For minimum phase noise it is important to set the insertion loss of the resonator to –6dB and hence Q_L/Q_0 to ½ [1] [2]. In the late 1980s this group developed some broad tuning resonators in both hybrid and MMIC topologies [5] [6] as illustrated in figures 1 and 2. These resonators were optimised to maintain near constant Q_L/Q_0 over a large frequency range.





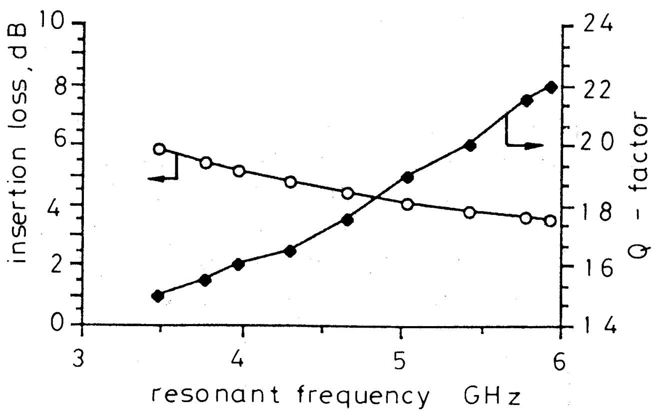
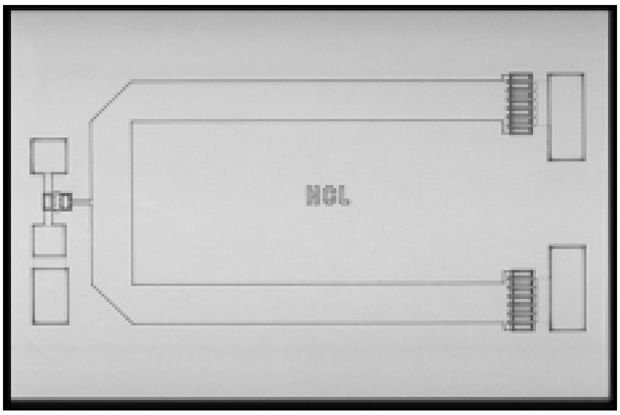


Figure 1 Octave Tuning resonator on alumina



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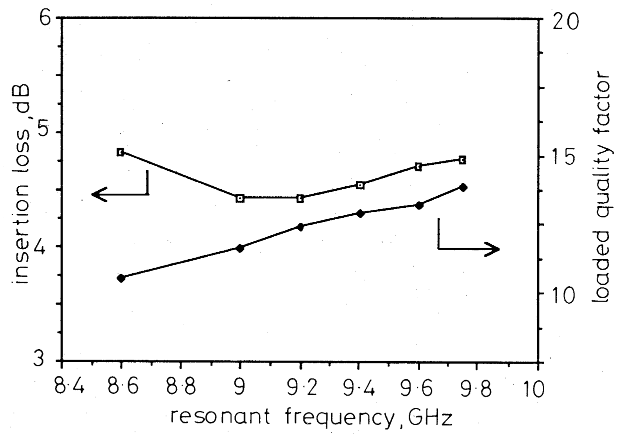


Figure 2 MMIC X band resonator

For broad tuning oscillators it is important to obtain low phase shift (or compensating phase shift) over broad frequency ranges. Therefore a wide variety of general resonators were evaluated and the amplitude, phase and field response were investigated. Some of the results are presented here. The general model for the resonator is shown in Figure 3.

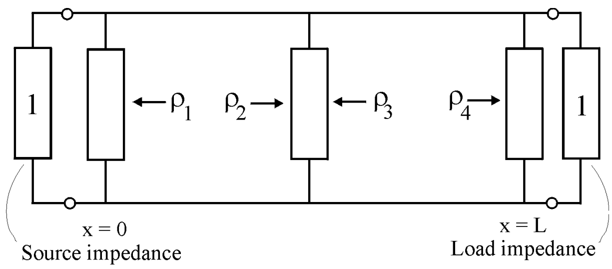


Figure 3. Model of a general resonator

This resonator incorporates 3 shunt reactances which can be inductive or capacitive. Only the results for the LCL resonator will be presented here where the centre component is a tunable capacitor.

The LCL bode plot, standing wave voltage, standing wave current and the magnitude and phase of S_{21} versus frequency are plotted in Figure 4, 5, 6, 7, and 8. This particular combination has low phase shift variation with frequency as well as decreasing S_{21} with increasing frequency. All of these make it attractive as a resonator for a broad tuning oscillator.

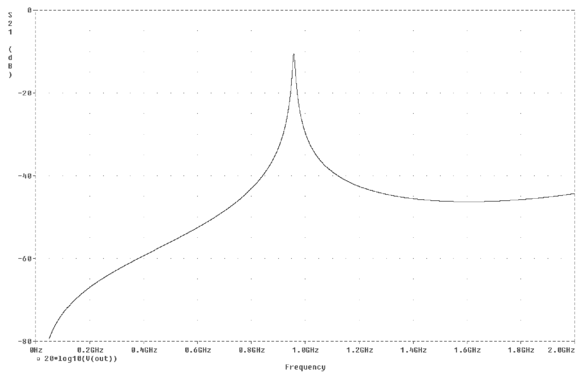


Figure 4 Frequency response

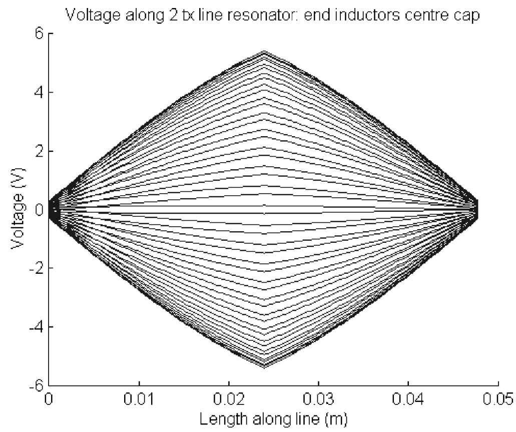


Figure 5 Voltage standing wave

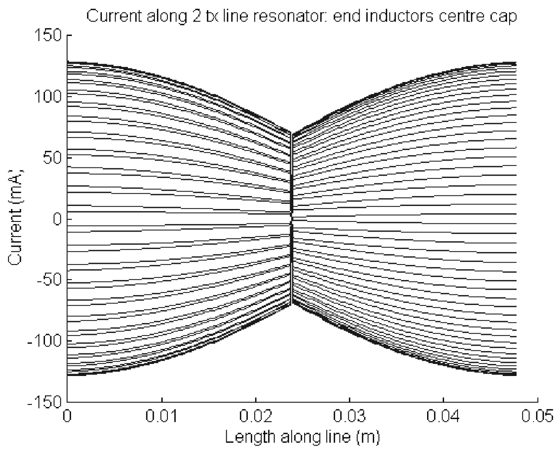


Figure 6 Current standing wave

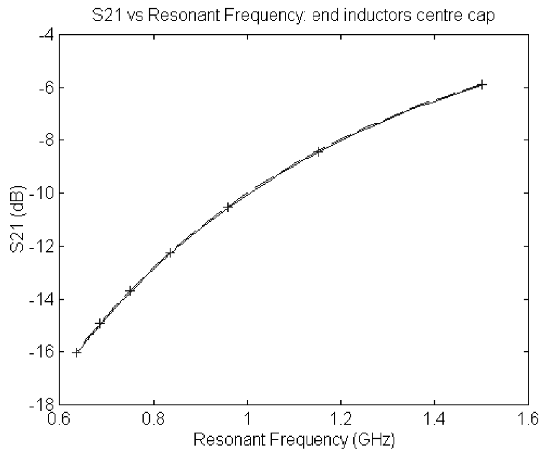


Figure 7 S_{21} (minimum loss) vs frequency

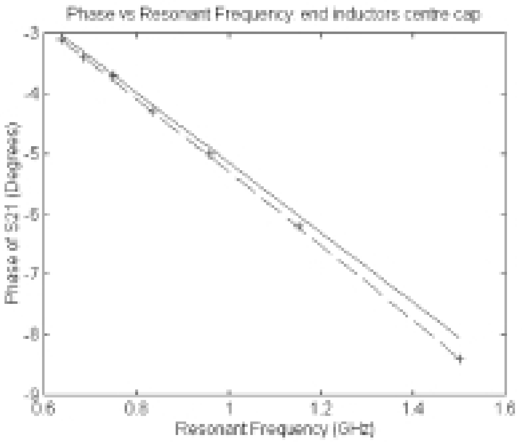


Figure 8 Phase shift variation at resonance vs frequency

III OSCILLATORS

Using these parameters an oscillator was built in triplate using a structure of the form shown in Figure 9. The resonator is placed on microstrip directly below the amplifier. Coupling between them is achieved using capacitive vias. The amplifier is biased with an inductive bias T. The bias T, the capacitive vias and the interconnecting lines are adjusted to achieve maximum tuning range through optimisation of the phase shift and the gain.

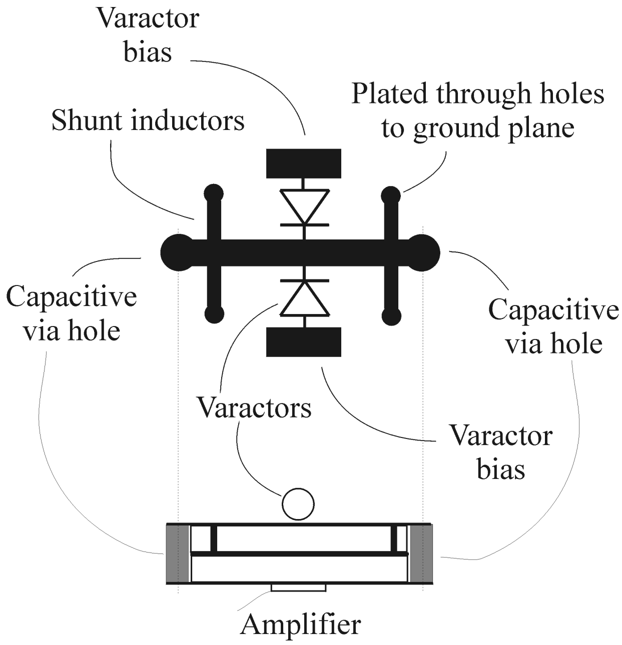


Figure 9 Triplate multilayer oscillator

Photo-graphs of the oscillator are shown in Figures 10 and 11 where it can be seen that the resonator is an electrically short LCL printed resonator with two varactors placed in parallel in the middle. The output power is coupled via a quarter wave high impedance (100Ω) line presenting 200Ω to the oscillator.

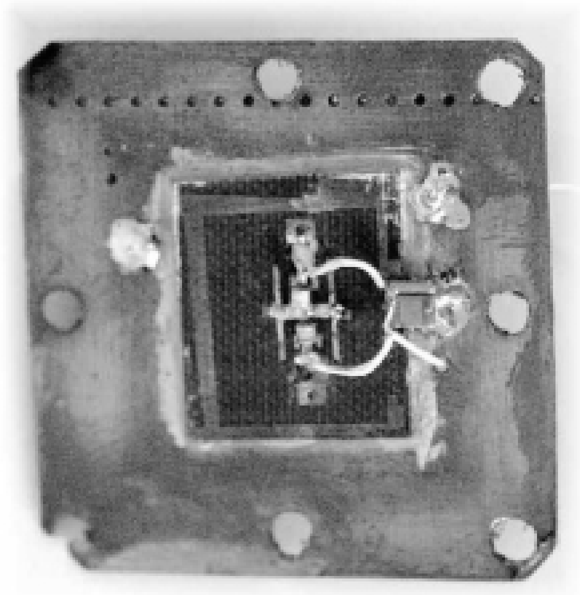


Figure 10 VCO rear view showing resonator with shunt inductors and varactors. Resonator length is 3.4mm.

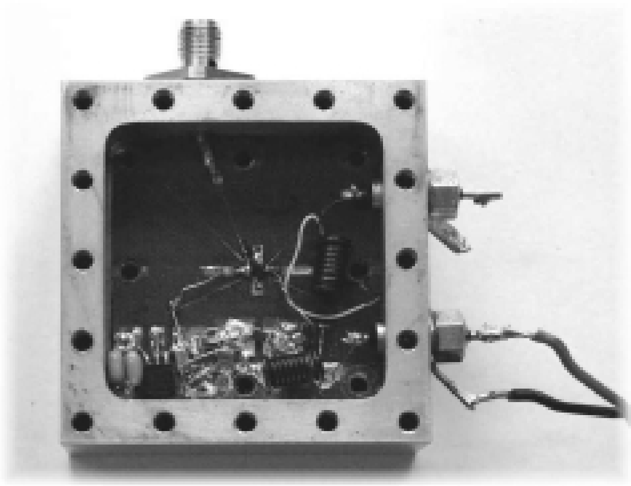


Figure 11 VCO front view with amplifier and output coupler and disconnected negative group delay circuit.

In figure 11 it can be seen that there are a number of additional transmission lines which are not connected. This was part of a broadband negative group delay circuit [4] which was not used in the final design.

IV BROADBAND NEGATIVE GROUP DELAY

Negative group delay can be obtained using a lossy shunt series LCR circuit or a parallel in line parallel resonant LCR circuit. To obtain broadband negative group delay in the smallest possible space, the shunt series tuned circuit is preferable as it is possible to place a number of these in parallel at a single point (Figure 12). Further it is fairly easy to implement these using resistive loaded transmission lines.

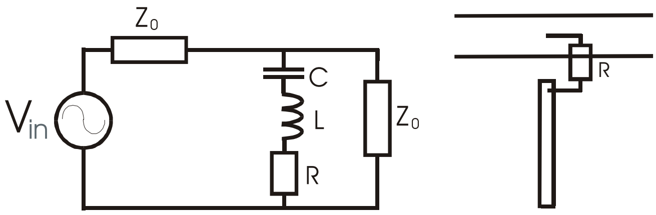


Figure 12 (a) Lossy series tuned shunt circuit. (b) Printed transmission line Version

The associated group delay of a shunt series tuned network can be expressed as:

$$\tau_{peak} = \tau\big|_{\omega_0} = -\frac{\partial \angle S_{21}}{\partial \omega}\bigg|_{\omega_0} = -\frac{\left(4Z_0R + 2{Z_0}^2\right)L}{R\left({Z_0}^2 + 4R^2 + 4Z_0R\right)}$$

As R increases the magnitude of the negative group delay decreases. This indicates that at any given frequency the magnitude of negative group delay required can be set simply by adjusting R. The peak insertion loss is:

$$S_{21} = \frac{2R}{2R + Z_0}$$

This shows increasing loss with increasing negative group delay, as R is lowered. It should also be noted that as L is increased, the magnitude of negative group delay is also increased. This can be directly related to increasing the Q of the structure. However this also has the effect of decreasing the bandwidth of the negative group delay region.

Broadband operation

It appears that optimum broadband performance occurs by directly coupling these circuits together in parallel. Two techniques were considered, a) Maintain constant resonator Q, constant R and vary the resonant frequency spacing b) Maintain constant frequency spacing, fix R and vary resonator Q. In this example the 2nd technique (b) is used. The frequency spacing is set by designing the next resonant frequency to be that of the previous resonators 3dB point. The 3dB bandwidth is made identical for all the resonant networks. This in turn is set depending on the number of coupled resonators and the total group delay bandwidth required.

The bandwidth covered is 3-6GHz and a 4 resonator coupled network was chosen. This forces resonant frequencies at 3.2, 4.05, 4.9 and 5.75GHz and a single resonator 3dB bandwidth of 1.7GHz. Therefore L, C and R can be calculated for each tuned circuit using simple circuit theory. At resonance an LC circuit can be modelled by a $\lambda/4$ transmission line.

A typical broadband circuit is shown in Figure 13. These structures were built adjacent to an identical reference line to enable calibration for zero delay.

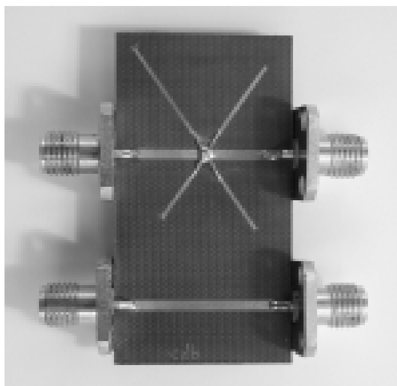


Figure 13 Photograph of a negative group delay network realized on microstrip.

Figure 14 displays the measured response of the coupled resonator negative group delay network. The lower trace shows the group delay centred around -50ps, the upper trace shows the insertion loss centred around 10dB.

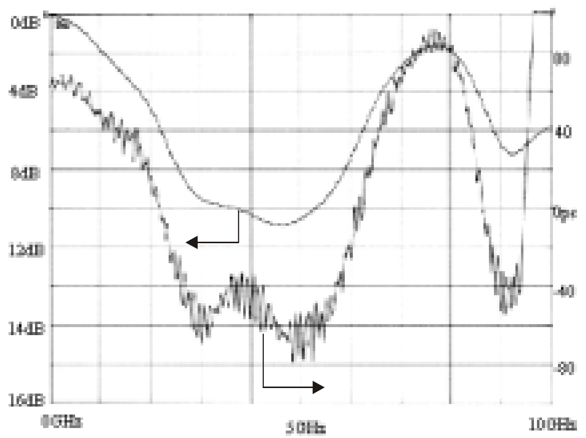


Figure 14 Measurement of a 4 coupled resonator negative group delay network with resonances at 3.2GHz (L=17mm), 4.05GHz (L=13mm), 4.9GHz (L=11mm) and 5.75GHz (L=9mm), Z = 85 ohms (w=0.5mm), R = 22 ohms.

V PHASE NOISE THEORY AND MEASUREMENT RESULTS

A. Phase Noise Calculations

It is important to develop a simple model to calculate and predict the noise performance of an oscillator [1]. A suitable model is shown in figure 15. This consists of an amplifier with two inputs which are added together. These represent the same input but are separated to enable one to be used for the noise input and the other for feedback. The resonator is represented as an LCR circuit where any impedance transformation is achieved by varying the component values. This circuit operates as a Q multiplication filter but also contains the additional constraint that the AM noise is removed. The model is put in this form to highlight all the effects, which often do not show up in a block diagram model.

A general equation for the phase noise can be derived as shown in equation V.1 which allows for a number operating conditions including power and the output and input impedances [1].

Equation V.1:

$$L_{FM} = A \cdot \frac{FkT}{8\left(Q_0\right)^2 \left(Q_L/Q_0\right)^2 \left(1 - Q_L/Q_0\right)^N P}\left(\frac{f_0}{\Delta f}\right)^2$$

where:

1. N = 1 and A = 1 if *P* is defined as P_{RF} and R_{OUT} = zero.
2. N = 1 and A = 2 if *P* is defined as P_{RF} and $R_{OUT} = R_{IN}$.
3. N = 2 and A = 1 if *P* is defined as P_{AVO} and $R_{OUT} = R_{IN}$.

P_{RF} is the total power dissipated in the output, input and loss resistances and P_{AVO} is the power available at the output of the amplifier.

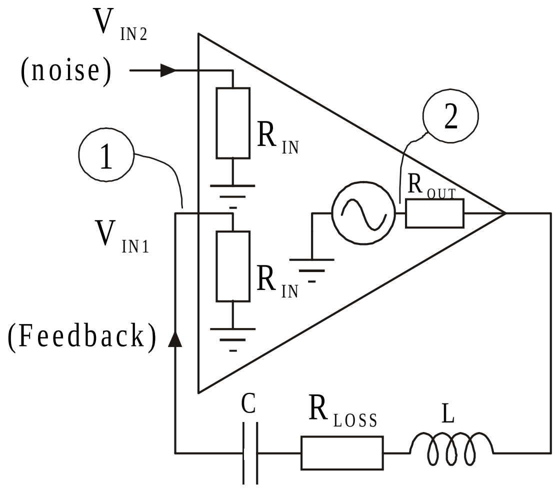


Figure 15 Oscillator model

If we take expansion 3 and show the full equation we obtain equation V.2 [1].

Equation V.2:

$$L_{FM} = \frac{FkT}{32Q_0^{\ 2}\left(Q_L/Q_0\right)^2\left(1-Q_L/Q_0\right)^2 P_{AVO}}\left(\frac{\left(R_{OUT}+R_{IN}\right)}{R_{OUT}\,.R_{IN}}\right)\left(\frac{f_0}{\Delta f}\right)^2$$

$$\text{Under optimum conditions } \rightarrow \frac{2FkT}{Q_0^{\ 2}P_{AVO}}\left(\frac{f_0}{\Delta f}\right)^2$$

This is minimum when $R_{OUT} = R_{IN}$ and $Q_L/Q_0 = ½$ and hence when the insertion loss of the resonator is –6dB. This minimum occurs because the amplifier gain is set by the insertion loss of the resonator which is $S_{21} = (1-Q_L/Q_0)$. This optimum value for Q_L/Q_0 is also described by Parker in a paper on SAW oscillators [2].

B. Phase Noise measurements

The oscillator tuned from 3.27 to 5.27GHz with an output power around 0dBm. The phase noise at 100kHz offset was measured to be –102dBc/Hz at 3.27GHz and –96dBc/Hz at 5.27GHz.

A full equation for the phase noise including the effect of the flicker noise corner, the far out noise and the output coupler is shown in Figure 16. The parameters are shown in the figure. The plot of the phase noise shows close correlation with the theory.

Phase Noise Plots

Phase Noise plots of oscillators in terms of Q0, QL, Power, Noise figure, Flicker Noise corner, Carrier frequency, Offset frequency and output coupler ratio.

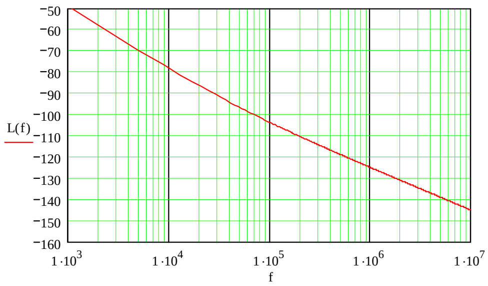
Q0 is the unloaded Q, QL is the loaded Q
F is the noise figure, k is boltzmanns constant, T is the temperature
fo is the carrier frequency, f is the offset frequency, fc is the flicker corner
C0 is is the coupler ratio with respect to the output of the coupler

$Q0 := 27.66 \quad QL := 8.3 \quad F := 4 \quad T := 293 \quad k := 1.38\,10^{-23}$

$f0 := 4.5\,10^{9} \quad fc := 40000 \quad P := .004 \quad C0 := 2 \quad ORIGIN := 0$

$f := 1000, 5000 .. 10000000$

$$L(f) := 10\cdot\log\left[\frac{C0\cdot F\cdot k\cdot T}{P} + \frac{F\cdot k\cdot T}{2\cdot P}\cdot\left[\frac{1}{\left(1-\frac{QL}{Q0}\right)^2}\right] + \frac{F\cdot k\cdot T\cdot\left(1+\frac{fc}{f}\right)}{8\cdot(Q0)^2\cdot\left(\frac{QL}{Q0}\right)^2\cdot\left(1-\frac{QL}{Q0}\right)^2\cdot P}\cdot\left(\frac{f0}{f}\right)^2\right]$$



VI. CONCLUSIONS

A broad tuning VCO has been demonstrated illustrating the requirements for optimum resonator and amplifier design. The oscillator tuned from 3.27 to 5.27 GHz with an output power around 0dBm. The phase noise at 100kHz offset was measured to be –102dBc/Hz at 3.27 GHz and –96dBc/Hz at 5.27GHz. Broadband negative group delay circuits were discussed.

VII. ACKNOWLEDGEMENTS

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