LINCOLN LABORATORY
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A complete practical and mathematical treatment of Microwave LC-Based 0scillators using Bipolar, MESFETs and CMOS Transistors

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160 GHz Oscillator – Measured Phase Noise Plot

OUTLINE

- **Introduction**
- **What is an oscillator?**
- **Introduction to Microwave Oscillators and Their Mathematical Treatment**
- **Transistor Types**
- **Transistor models and Noise Contributions**
- **Linear Approach to the Calculation of Oscillator Phase Noise**
- **Typical Microwave Oscillator (Colpitts Oscillator)**
- **Designing an Oscillator**
- **Tuning Diode Noise Contribution**
- **Designing an Oscillator based on Linear S-Parameters**
- **Large signal Operation of Oscillators (Colpitts/YIG)**
- **Classical Linear Two-Port Oscillator Design**
- **Microwave Resonators**
- **A Novel Tunable Active Spiral Inductor**
- **The Active Inductor Using a Gyrator**

- **INTRODUCTION: Active Inductor Oscillator**
- **Synthesized Inductor Circuits**
- **Phase Noise Contribution of the Various Parts of the Oscillator Using an Active Inductor**
- **The Modern Time-Domain Behavior of an Oscillator**
- **Test Case: Design Example of a 100MHz crystal oscillator**
- **Phase Noise Analysis Based on the Negative Resistance Model**
- **Oscillator Phase Noise**
- **Design Examples of Oscillators for Best Phase Noise and Good Output Power**
- **Phase Noise of the integrated CMOS oscillator**
- **Appendix: A Solution for Calculating Phase Noise in Oscillators**
  - o **Oscillator Noise Models**
 - o **Noise Generation in Oscillators**
- **Key References**

➢ DEFINITION OF AN OSCILLATOR

➢ An Oscillator is an Electronic Circuit that converts DC power to RF power, this can range from a few Hz to Tera Hz and higher

➢ An oscillator consists of an active device acting as an amplifier, a resonator, and a feedback circuit

➢ A small amount of energy feedback is needed to sustained oscillation and the majority of available energy appears at the output terminals

➢ Resonators can be LC based circuits, transmission line based, crystal, ceramic, dielectric resonator ,YIG (Yttrium Garnet) based, and others

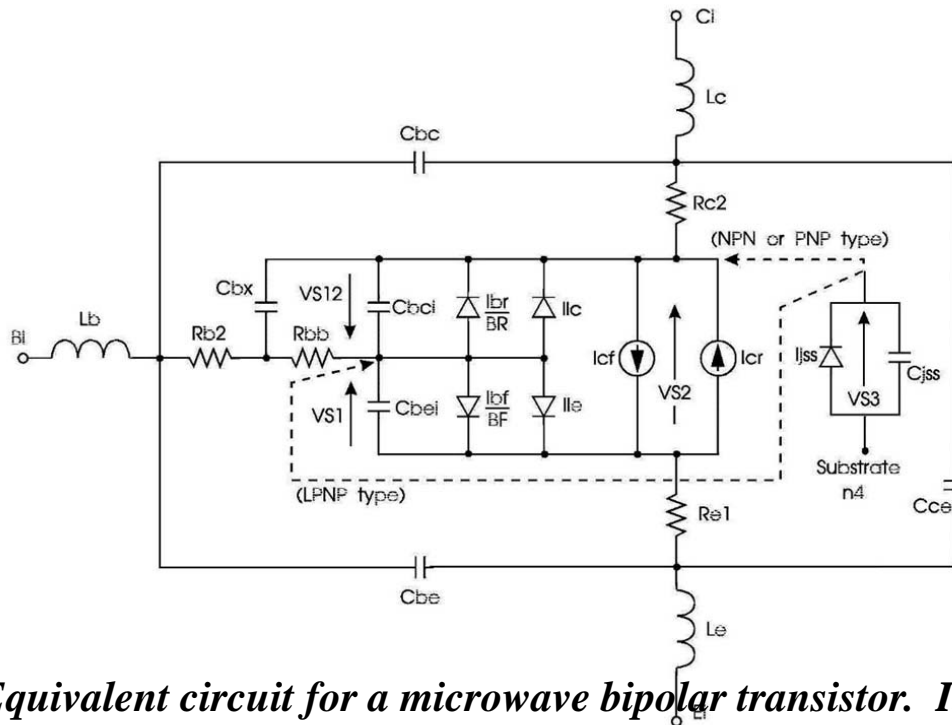**For RF application, the most relevant features besides size are:**

- Output power
- Harmonic content
- Phase Noise
- Power consumptions, to name a few

OSCILLATOR BASICS
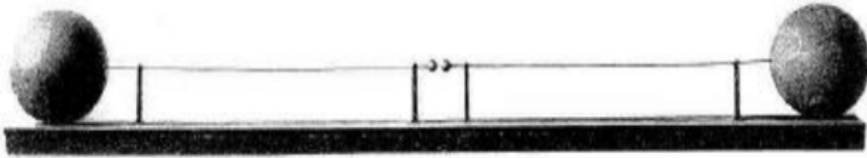
Original dipole made by Heinrich Hertz in 1887 using balls at the end to form a capacitive load (Deutsches Museum, Munich)
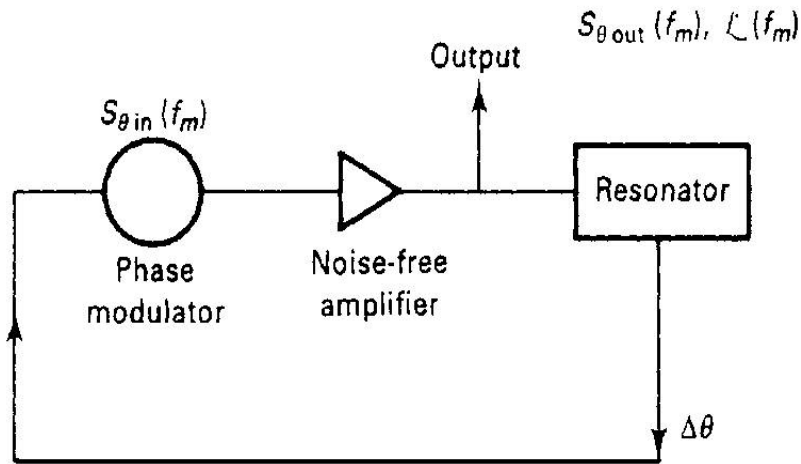
Basic Oscillator Configurations

Equivalent circuit for a microwave bipolar transistor. It deviates from the SPICE implementation by having two base-spreading resistors
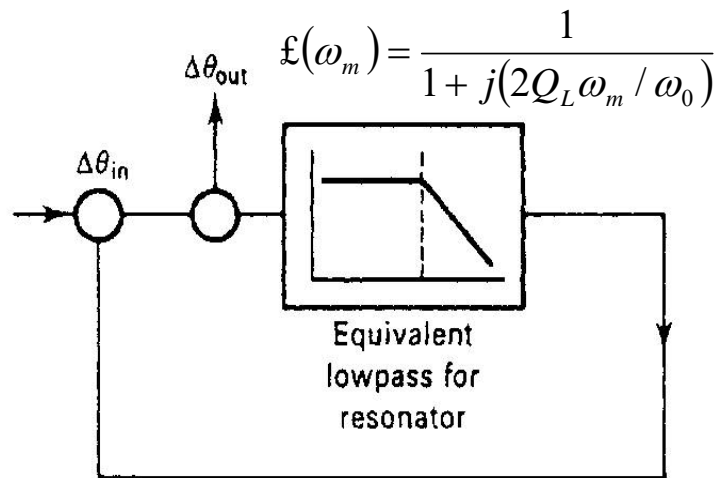
➢ OSCILLATOR BASICS

A typical linear oscillator phase noise model (block diagram) Leeson Model

$$\pounds(\omega_m) = \frac{1}{1 + j(2Q_L\omega_m/\omega_0)}$$

The resulting phase noise in linear terms can be calculated as

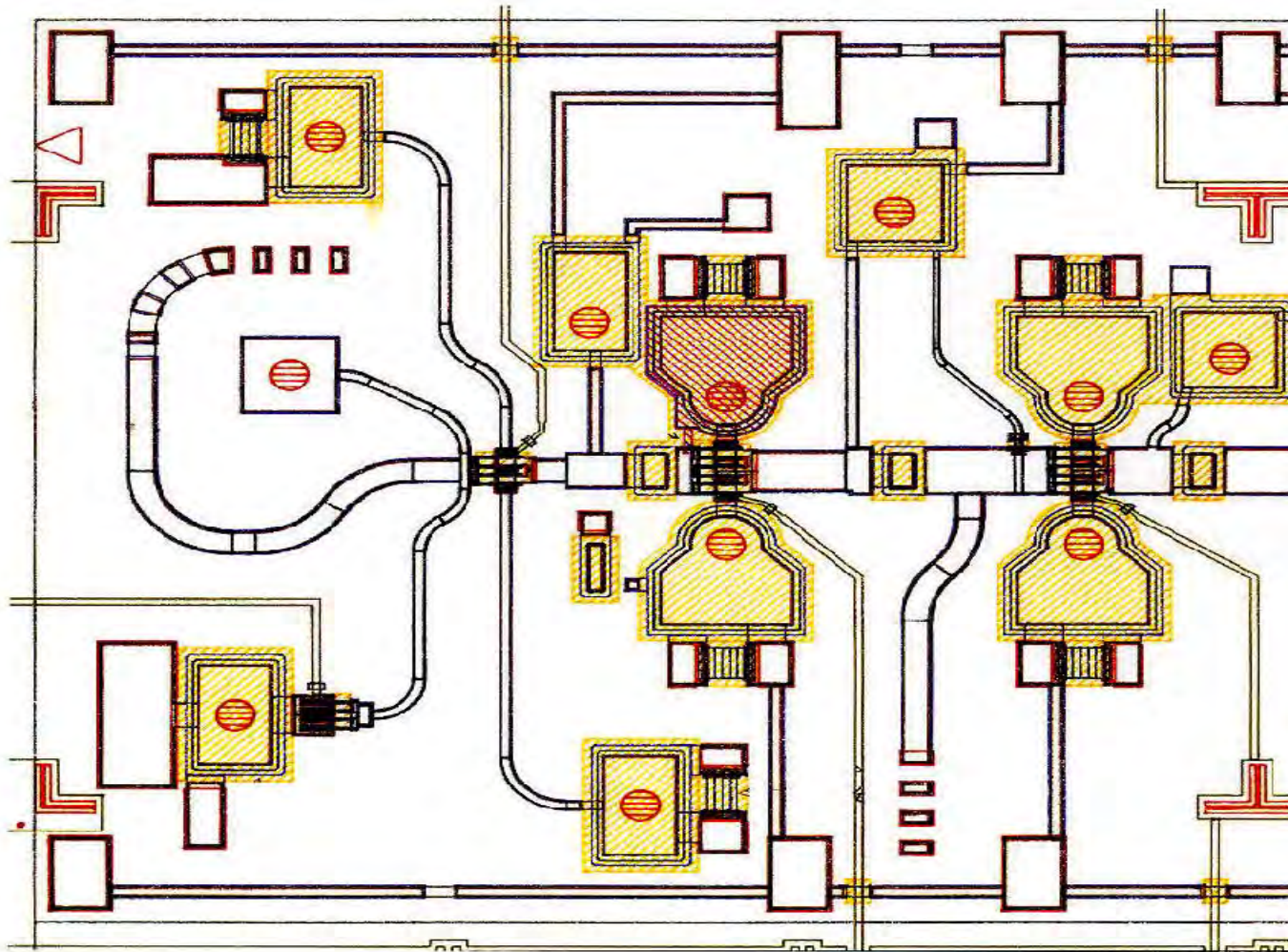$$\mathcal{L}(f_m) = \frac{1}{2}\left[1 + \frac{\omega_o^2}{4\omega_m^2}\left(\frac{P_{in}}{\omega_o W_e} + \frac{1}{Q_{unl}} + \frac{P_{sig}}{\omega_o W_e}\right)^2\right]\left(1 + \frac{\omega_c}{\omega_m}\right)\frac{FkT_o}{P_{sav}}$$

input power over reactive power

phase perturbation

resonator Q

flicker effect

signal power over reactive power

This equation is the linear Leeson equation, with the pushing effect omitted and the flicker term added by Dieter Scherer (Hewlett Packard, about 1975).
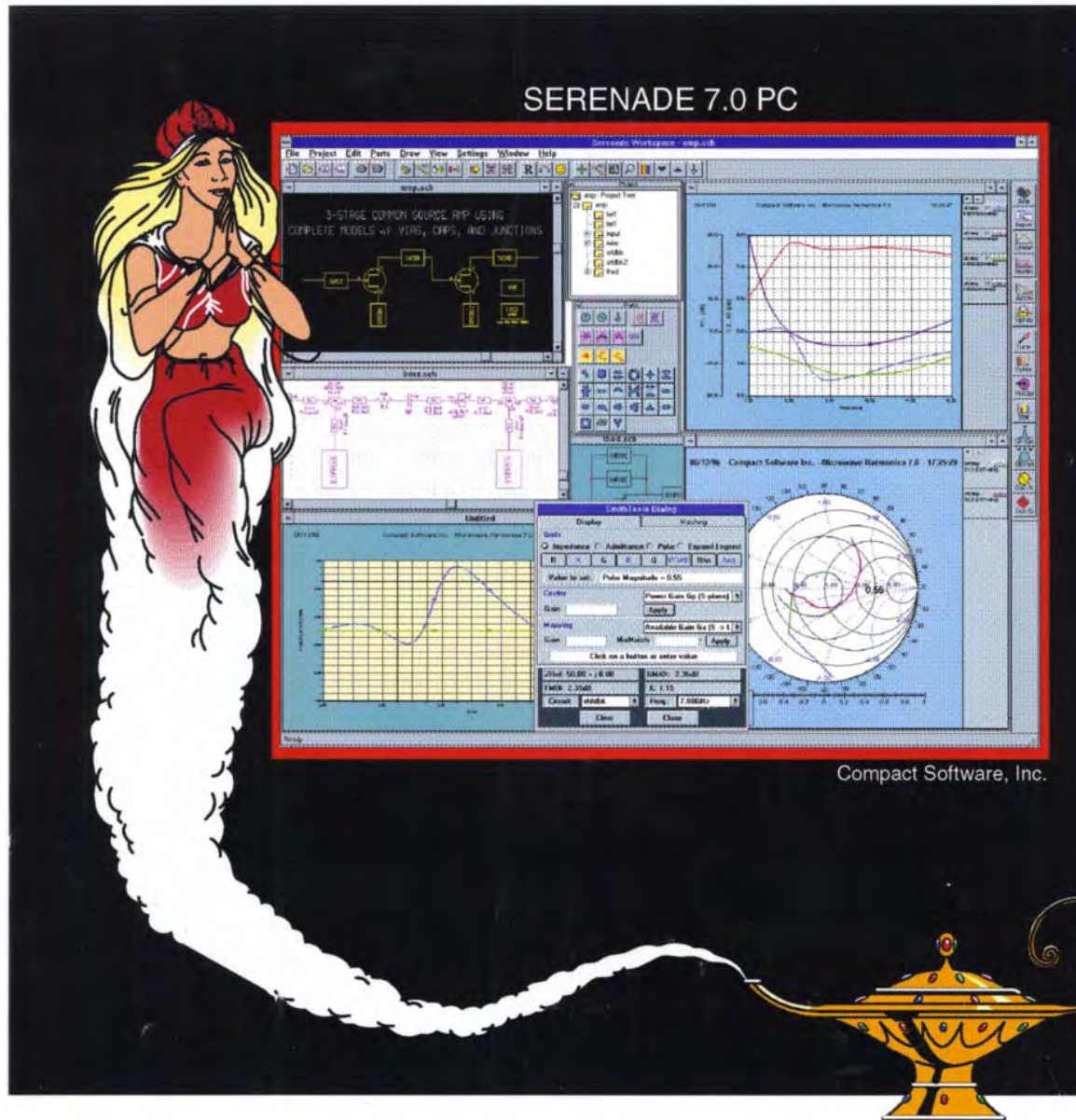
Phase noise is a dimensionless number, and expressed in dBc/Hz, measured at an offset of Δf (f_m) from the carrier relative to the RF output power. At 0 dBm output, the ideal phase noise level far off the carrier is -174dB (T_0= 300 Kelvin)

Two Stage FET Based Microwave Tunable Oscillator Built On GaAs Substrate

Eliot Cohen, "The MIMIC Program - A Retrospective", *Microwave Magazine*, June 2012, pp. 77-88

SERENADE 7.0 PC

3-STAGE COMMON SOURCE AMP USING
COMPLETE MODELS of VIAS, CAPS, AND JUNCTIONS

Compact Software, Inc.

This CAD tool was developed under the MIMIC Program

A detailed mathematical analysis of the time-domain behavior of oscillators, intended as a stand-alone mathematical derivation was developed between Prof. Rizzoli and team from the University of Bologna, Rowan Gilmore and Prof. Fred Rosenbaum from the University of Washington, St. Louise, and my team at Compact Software, where we introduced the world's first harmonic balance mathematics based simulator, that could handle nonlinear noise in amplifiers, frequency doublers, mixers and finally oscillators.

The DARPA program manager was Dr. Eliot Cohen. In 2012 Eliot Cohen wrote an article on the MIMIC Program for IEEE MTT-S Microwave Magazine which is a good resource if you want to learn about this historic program. Here are the details:

Eliot Cohen, "The MIMIC Program - A Retrospective", *Microwave Magazine*, June 2012, pp. 77-88
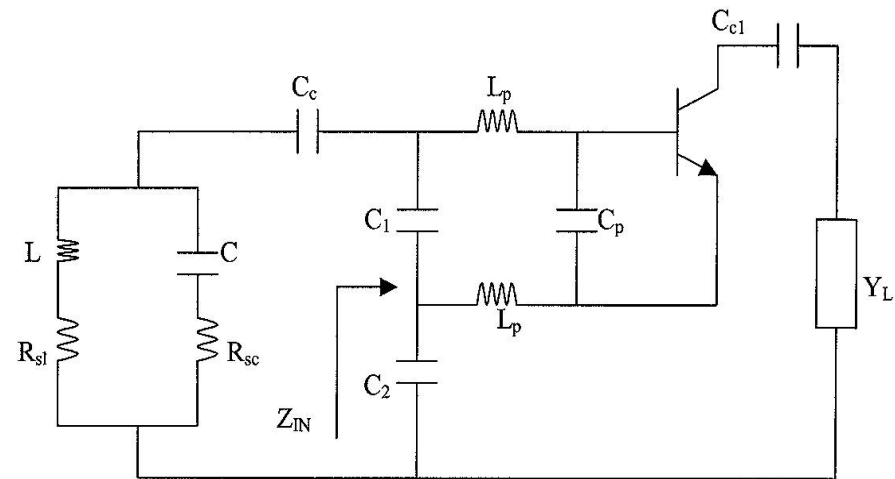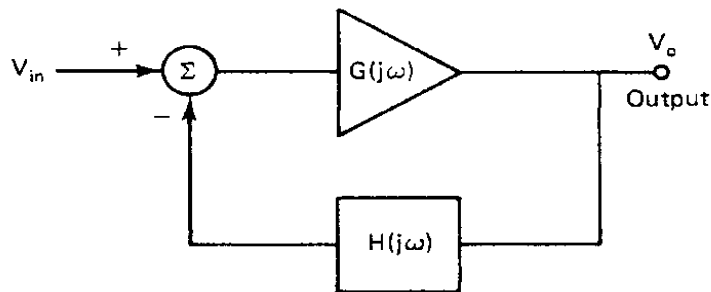
OUTLINE

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- **Microwave oscillators are based on the negative resistance principle to compensate for the loses.**

- **Maximum frequency of oscillation can be determined from linear analysis for start-up conditions, but not necessarily for sustaining oscillation (large signal condition will reduce the gain and shift the frequency).**

- **Linear analysis is unreliable to determine resonance frequency and other dynamic parameters, beware of parasitics.**

A typical block diagram of feedback oscillator circuit

$$Z_{IN}\big|_{pacakage} = -\left[\frac{Y_{21}^*}{\omega^2(C_1+C_p)C_2}\frac{1}{(1+\omega^2 Y_{21}^{*2}L_p^2)}\right] - j\left[\frac{(C_1+C_P+C_2)}{\omega(C_1+C_P)C_2} - \frac{\omega Y_{21}^* L_P}{(1+\omega^2 Y_{21}^{*2}L_p^2)}\frac{Y_{21}^*}{\omega(C_1+C_P)C_2}\right]$$

Y_{21}^* **= Large signal value of g$_m$=Y$_{21DC}$**

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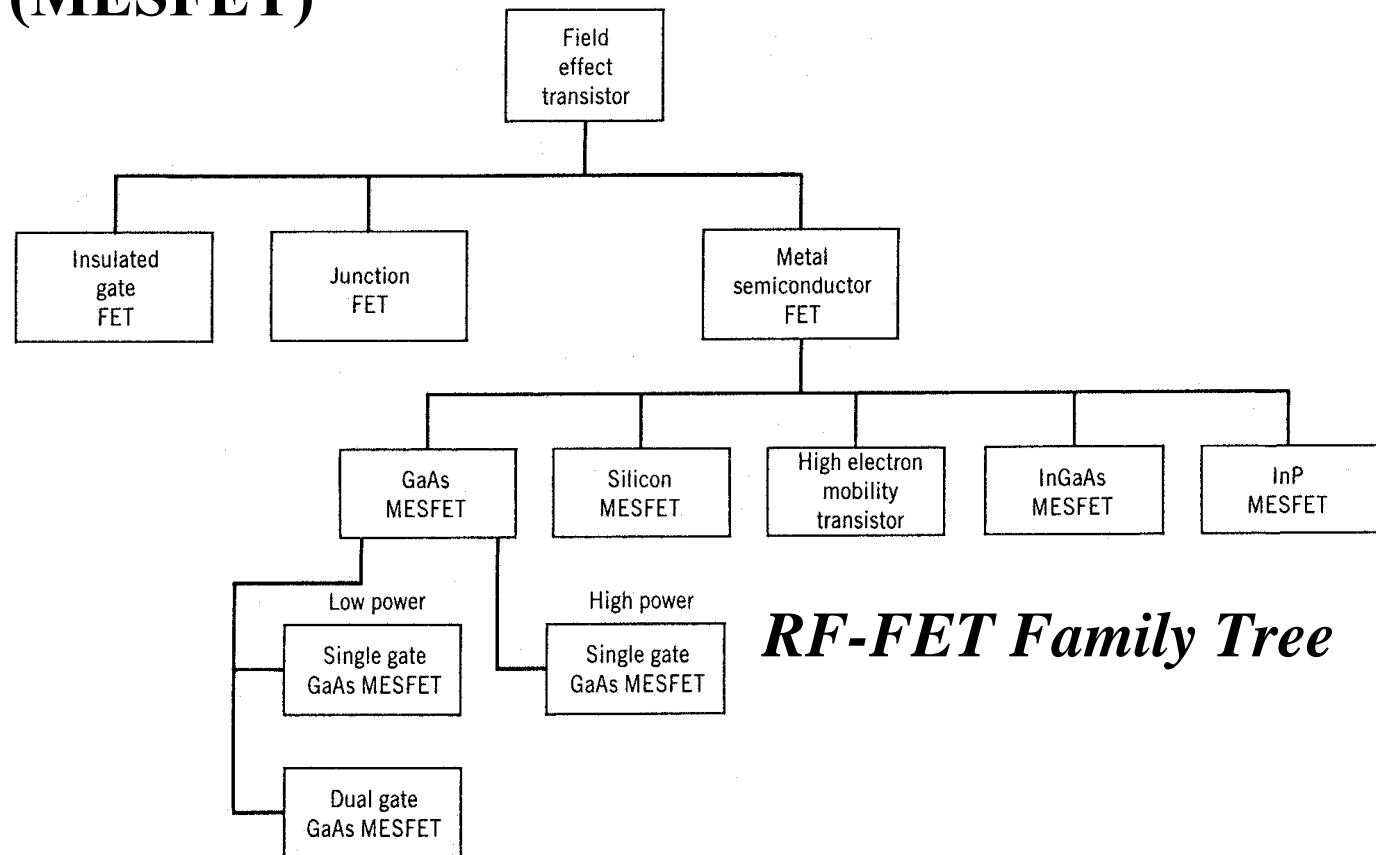 - Noise Generation in Oscillators
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➤ TRANSISTOR TYPES

Bipolar Transistors

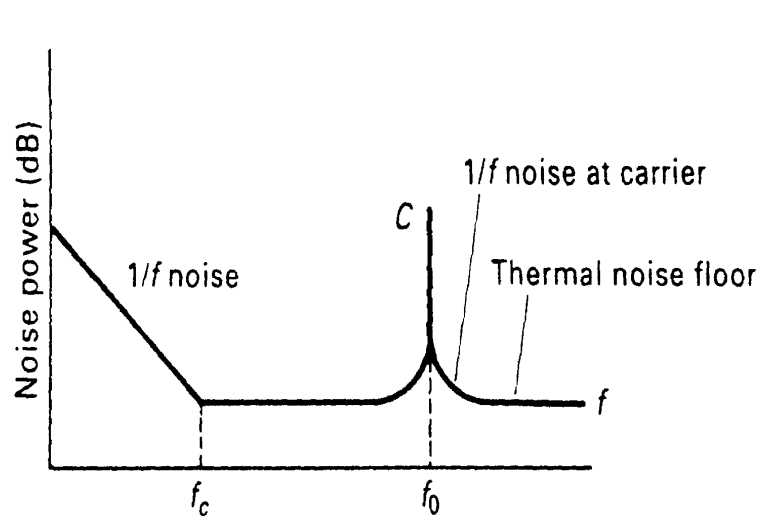**Field-Effect Transistors**

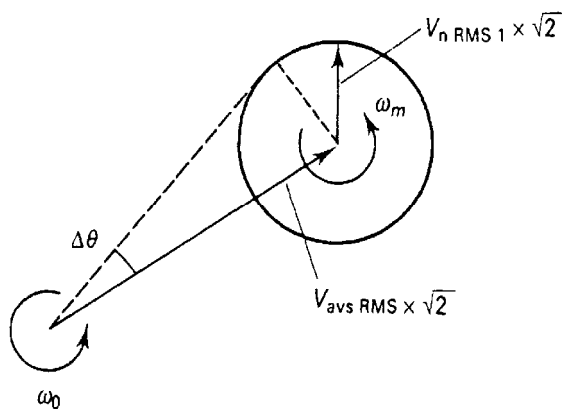**GaAs FETs (MESFET)**

Other….
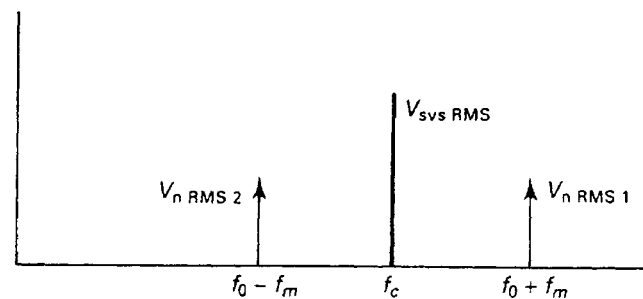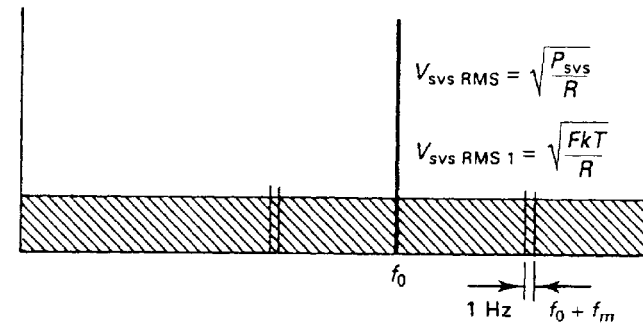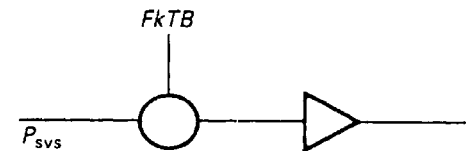

RF-FET Family Tree

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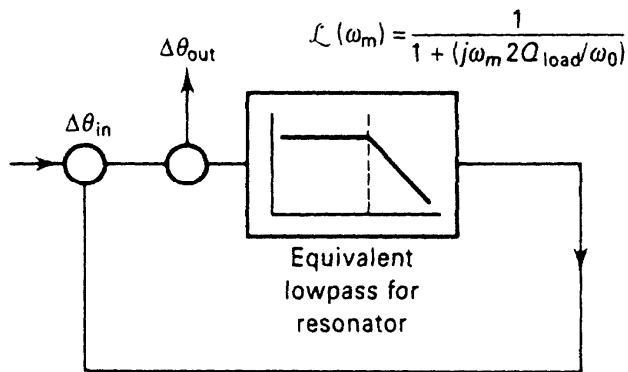- Key References

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Noise power versus frequency of a transistor amplifier with an input signal applied.

$$V_{svs\,RMS} = \sqrt{\frac{P_{svs}}{R}}$$

$$V_{svs\,RMS\,1} = \sqrt{\frac{FkT}{R}}$$

Phase noise added to carrier

$$\mathcal{L}(\omega_m) = \frac{1}{1 + (j\omega_m 2Q_{load}/\omega_0)}$$
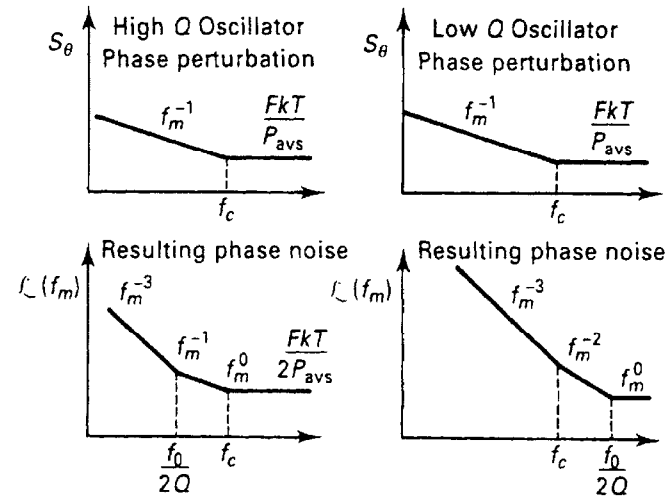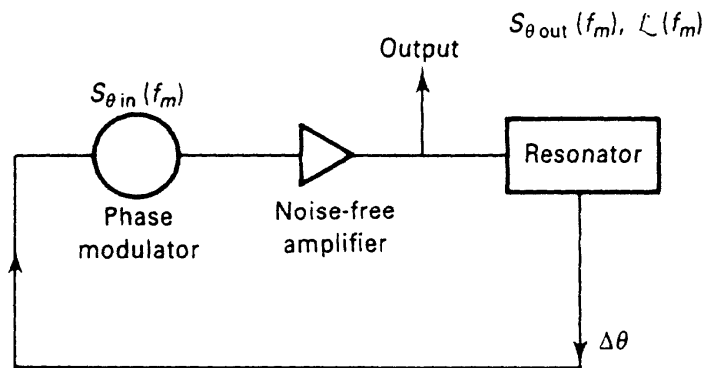
Equivalent feedback models of oscillator phase noise
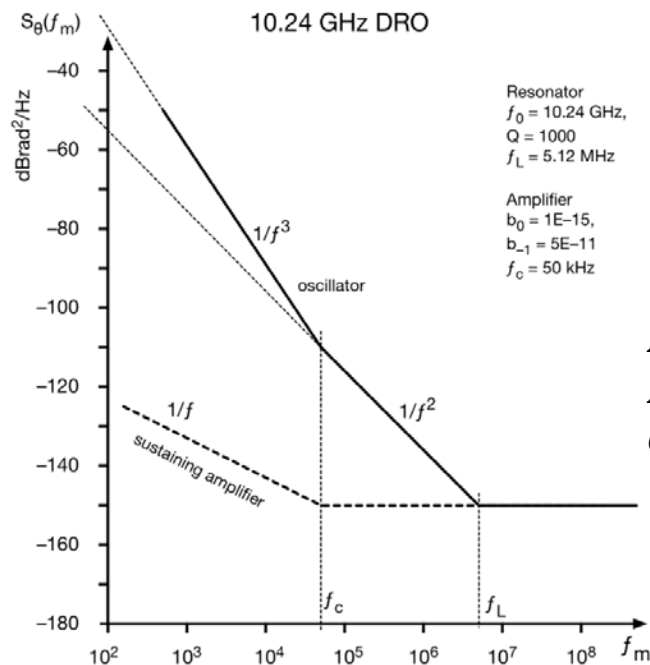
Phase Noise PSD of the 10.24 GHz DRO

OUTLINE

$$\mathcal{L}(f_m) = \frac{1}{2}\left[1 + \frac{\omega_o^2}{4\omega_m^2}\left(\frac{P_{in}}{\omega_o W_e} + \frac{1}{Q_{unl}} + \frac{P_{sig}}{\omega_o W_e}\right)^2\right]\left(1 + \frac{\omega_c}{\omega_m}\right)\frac{FkT_o}{P_{sav}}$$

input power over reactive power

phase perturbation

resonator Q

flicker effect

signal power over reactive power

Oscillator as a feedback model and as a one port producing a loss compensation with the electronically generated negative impedance

Colpitts oscillator with base-lead inductances and package capacitance. C_C is neglected

$$Z_{IN}\big|_{pacakage} = -\left[\frac{Y_{21}^*}{\omega^2(C_1 + C_p)C_2}\frac{1}{(1 + \omega^2 Y_{21}^{*2}L_p^2)}\right] - j\left[\frac{(C_1 + C_p + C_2)}{\omega(C_1 + C_p)C_2} - \frac{\omega Y_{21}^* L_P}{(1 + \omega^2 Y_{21}^{*2}L_p^2)}\frac{Y_{21}^*}{\omega(C_1 + C_p)C_2}\right]$$
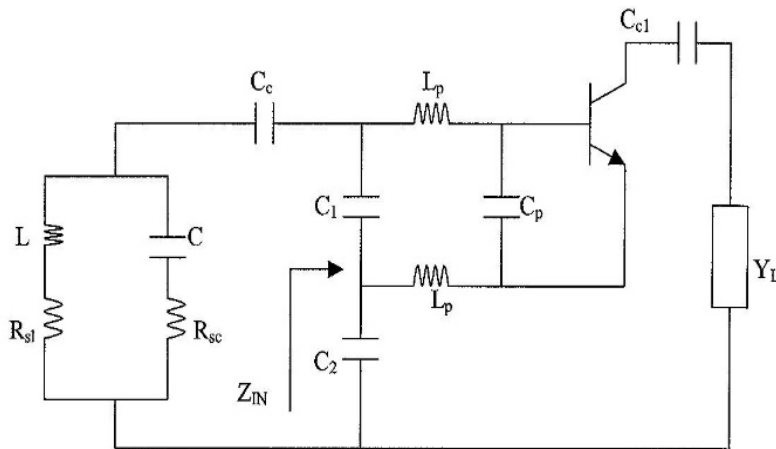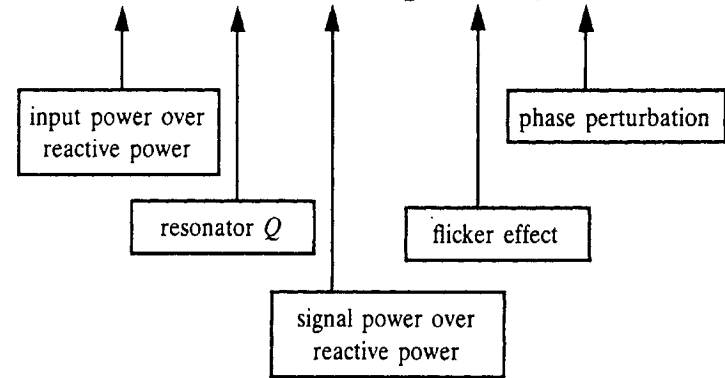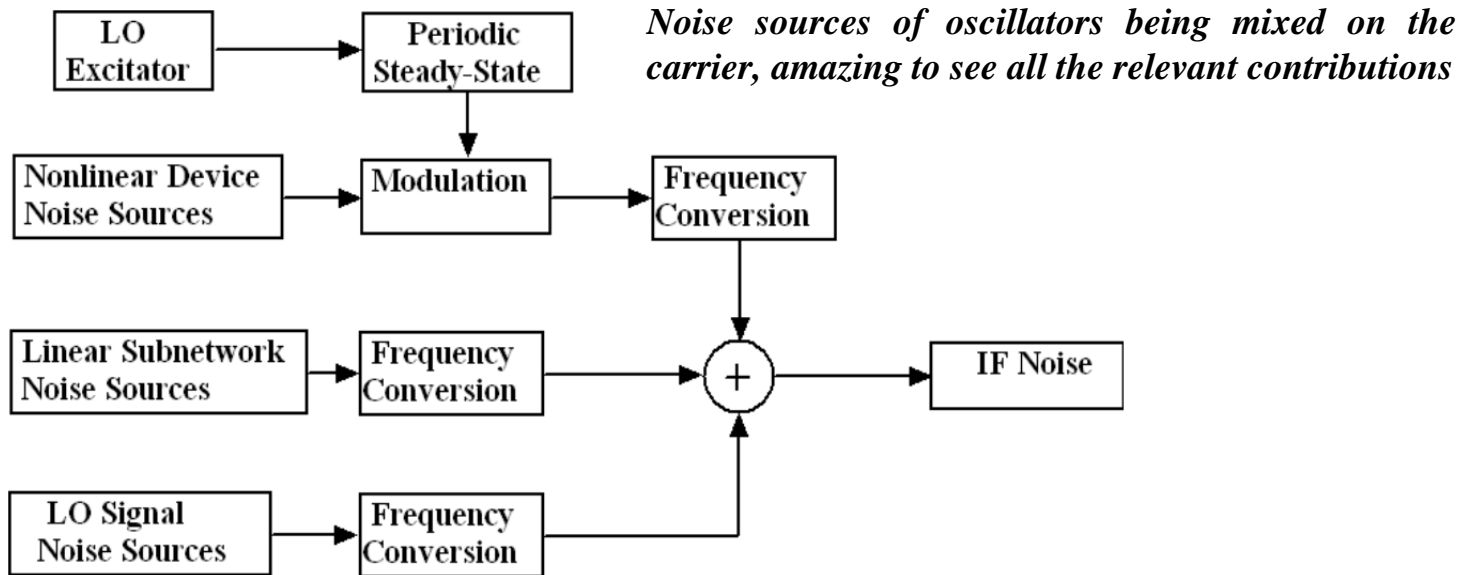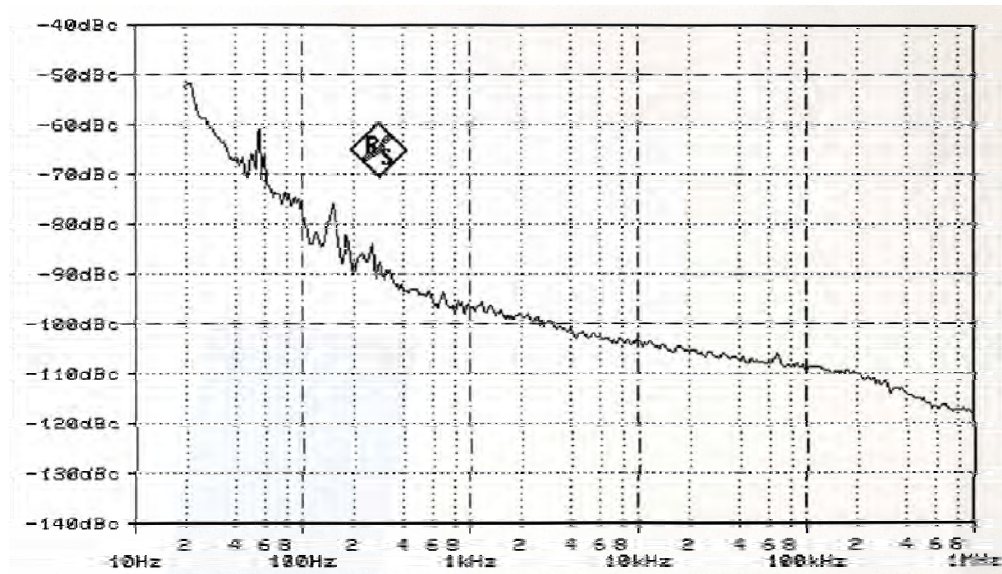
$$Y_{21}^* = Large\ signal\ value\ of\ gm = Y21$$

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➢ DESIGNING AN OSCILLATOR

Noise sources of oscillators being mixed on the carrier, amazing to see all the relevant contributions

Typical phase noise of a PLL
50 GHz PLL synthesizer

➢ DESIGNING AN OSCILLATOR

Single sideband phase noise as a function of the normalized drive level x
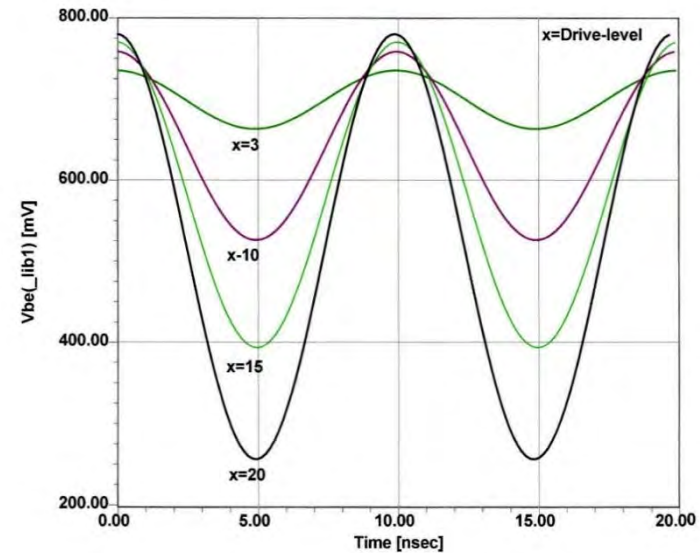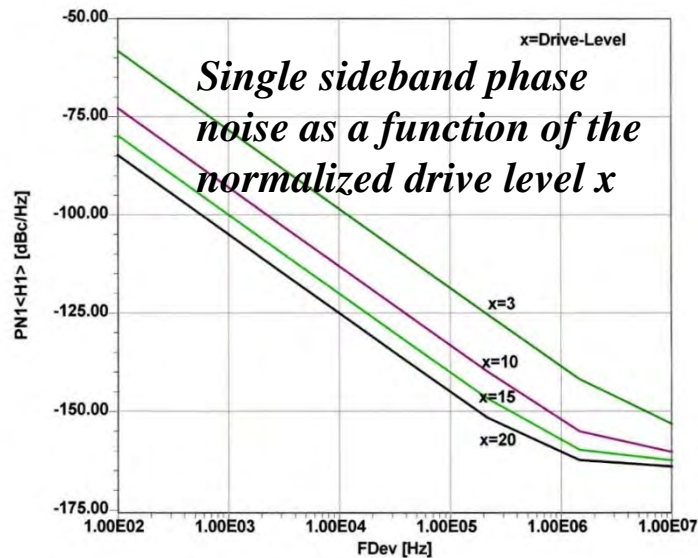
RF voltage Vbe across the base-emitter junction as a function of the normalized drive level x

collector current pulses of the oscillator
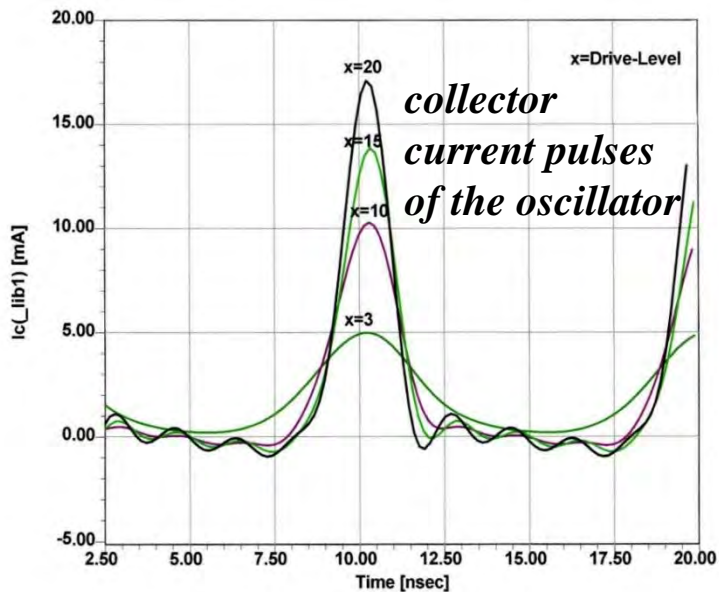
As the driven voltage $V_1(cos(\omega t))$ increases and develops enough amplitude across the base-emitter junction, the resulting current is a periodic series of pulses whose amplitude depends on the nonlinear characteristics of the device

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➢ LC OSCILLATOR PHASE NOISE

VARACTOR-TUNED LOW Q-FACTOR 1.9 GHz LC OSCILLATOR

PHASE NOISE OF VARACTOR-TUNED 1.9 GHz LC OSCILLATOR @ 10 kHz=-55 dBc/Hz

PN DEGRADATION=15 dB @ 10 kHz Offset

PHASE NOISE OF 1.9 GHz LC OSCILLATOR @ 10 kHz=-71 dBc/Hz
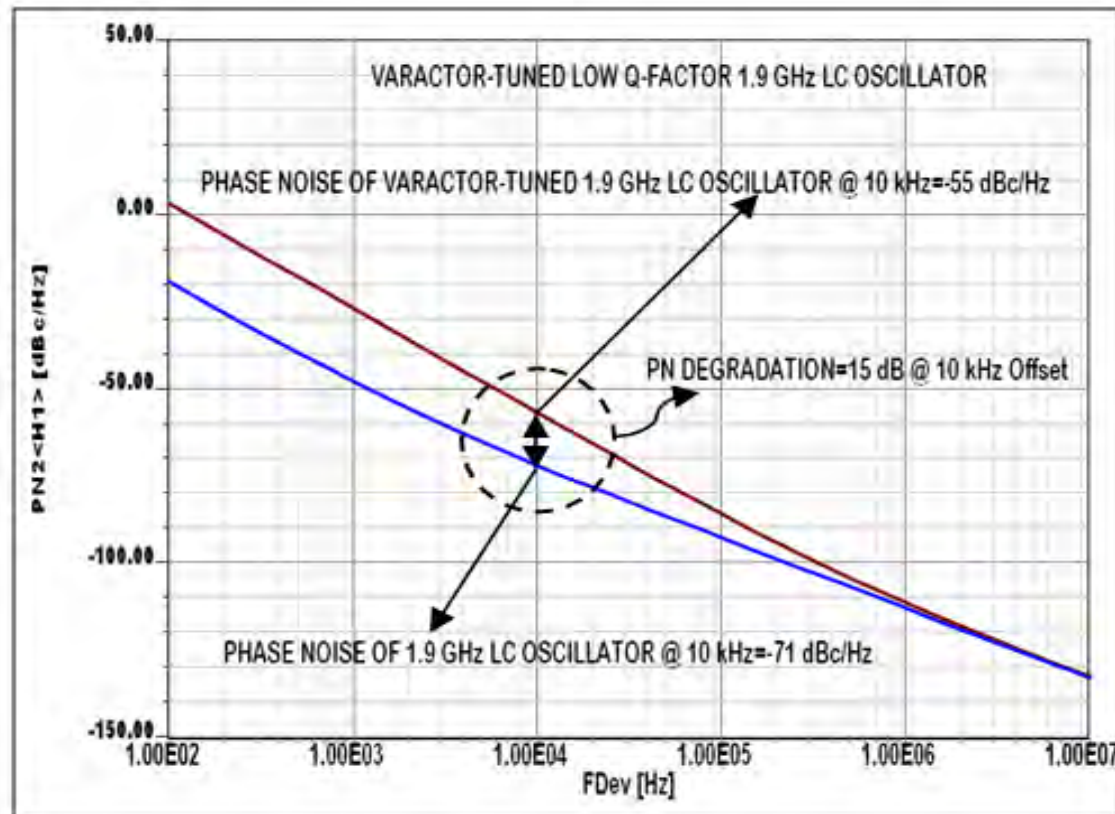
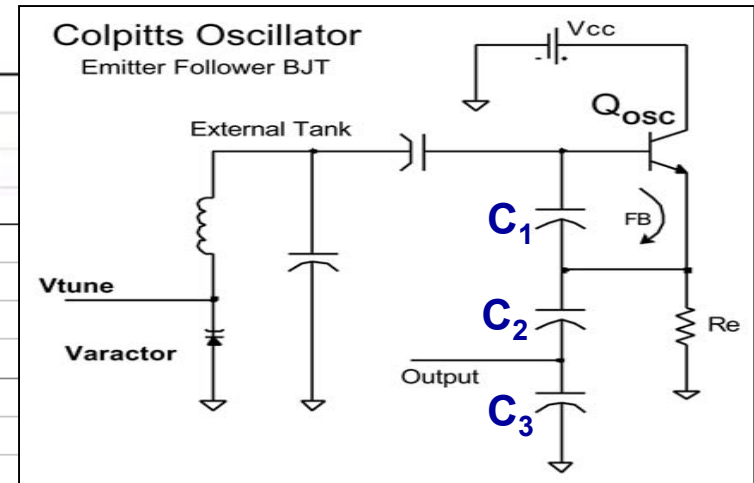**The Leeson phase noise equation is modified to accommodate the tuning diode noise contribution**

$$\pounds(f_m) = 10 \; \log \left\{ \left[1 + \frac{f_0^2}{(2 f_m Q_0)^2 \, m^2 \, (1-m)^2} \right] \left(1 + \frac{f_c}{f_m} \right) \frac{FkT}{2 P_0} + \frac{2\,kTRK_0^2}{f_m^2} \right\} \qquad m = \frac{Q_L}{Q_O}$$

The Equation above explain the phase noise degradation (as compared to the fixed frequency LC oscillator due to the oscillator voltage gain K_0 associated with the tuning diode network as described by Rohde). The reason for noise degradation is due to the increased tuning sensitivity of the varactor diode tuning network.

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Real and imaginary values for Z_{11}

For oscillation, $R_n = -\dfrac{Re(S_{21})}{\omega^2 \times C_1 \times C_2}$

$C_3 \gg C_2$

Colpitts Oscillator
Emitter Follower BJT

External Tank

Vtune

Varactor

Output

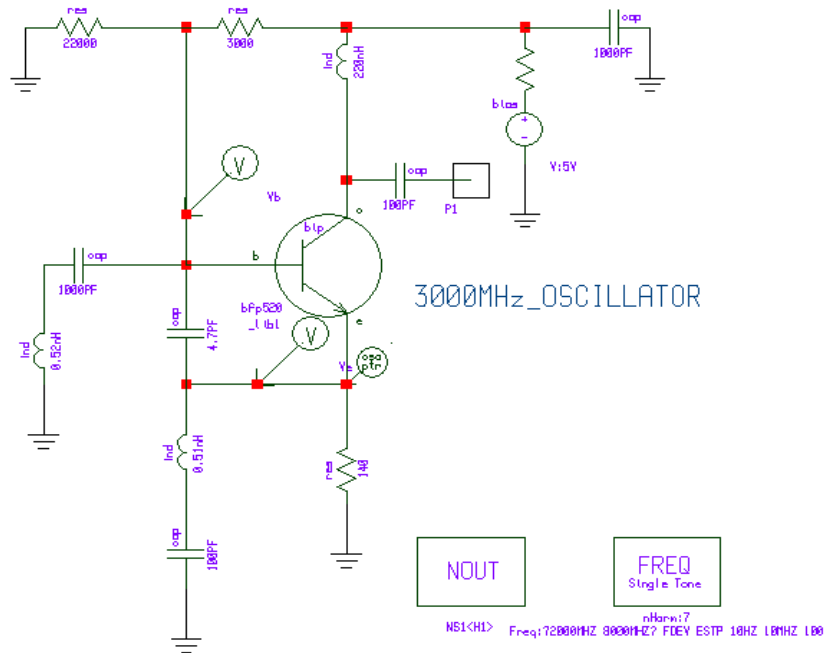V_{cc}

Q_{osc}

C_1

C_2

C_3

FB

Re
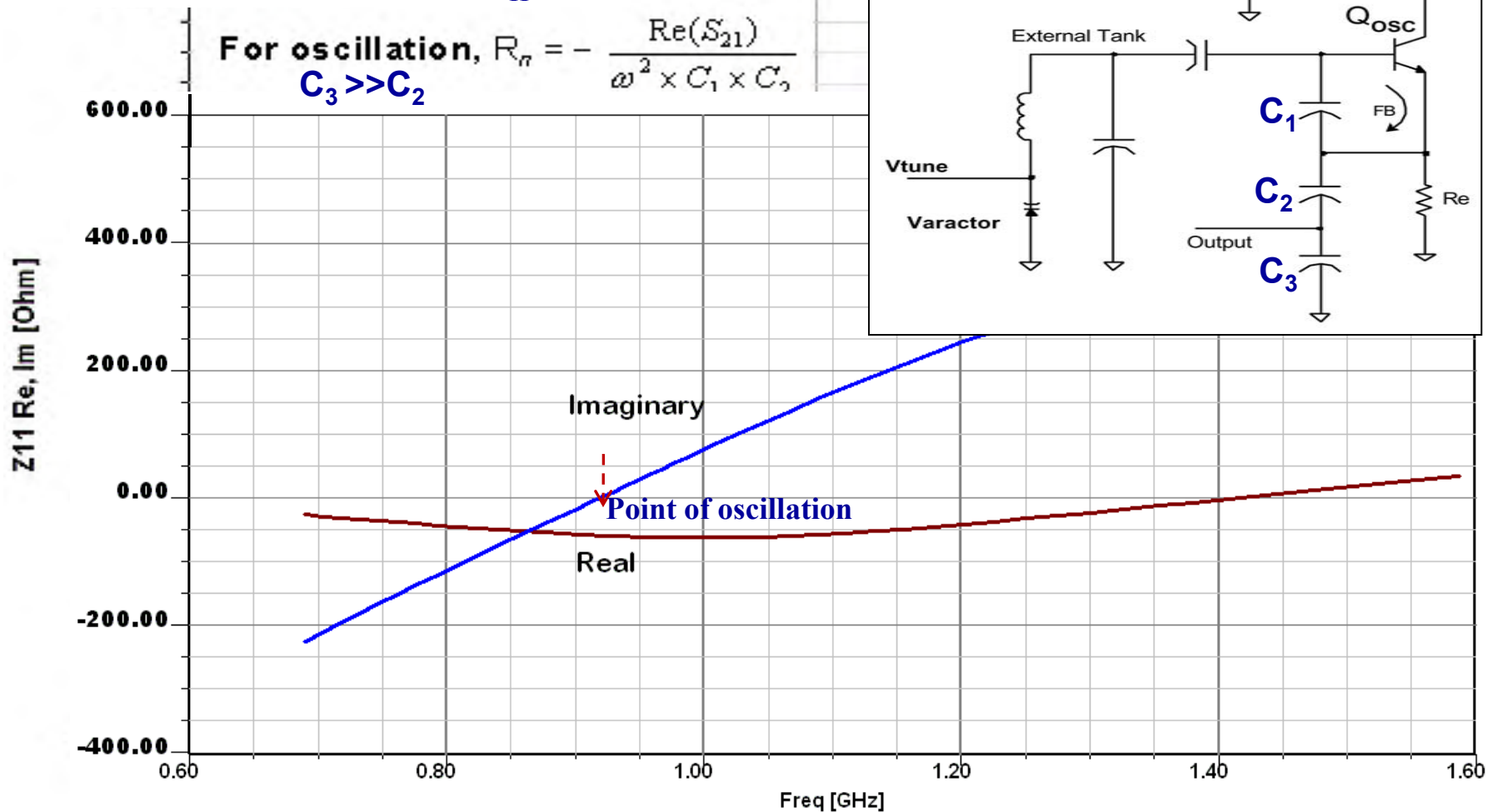
Real (Z_{11}) must be slightly more negative than the loss resistance in the circuit for oscillation to start. The resulting dc shift in the transistor will then provide the amplitude stabilization as g_m will be reduced.
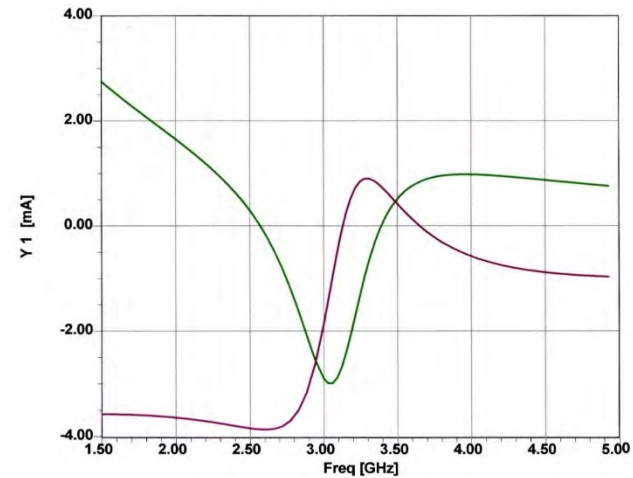
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➢ DESIGNING AN OSCILLATOR BASED ON LINEAR S-PARAMETERS

3000MHz_OSCILLATOR

Shows the real and imaginary currents for oscillation. The reactive current crosses the zero line at 3120 MHz. This is close, but not exactly at the point of most negative resistance current. The reason for the shift of 120 MHz is due to the use of small-signal analysis rather than the large-signal analysis
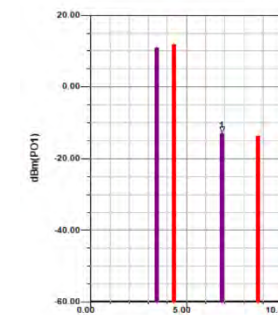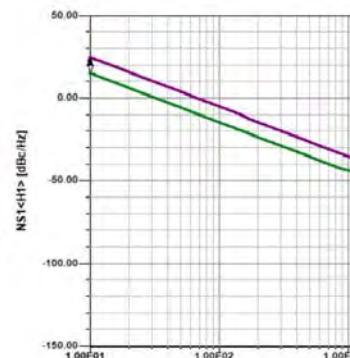
A 3000 MHz oscillator using a BFP520 transistor operating at 2V and 20mA. In this case, the capacitor C_2 needs to be replaced by an inductor L_3 which tunes out the collector emitter capacitance to achieve the optimum value. The 1nF on the left is a DC separation capacitor. This case is optimized for output power


Increasing the output power strongly reduces the phase noise

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➢ LARGE SIGNAL OPERATION

- **Definition: RF voltages/currents are of similar magnitude as the DC values. Test points were V_c = 2V, I_c = 20mA.**

- **The transistor behaves differently under large signal conditions.**

- **Large signal parameters can be obtained from simulation using SPICE parameters, calculating the Bessel functions of the currents of the intrinsic transistor and adding the parasitics and measurements.**

**This Figure shows the R&S VNA and the test fixture for the transistor of choice**
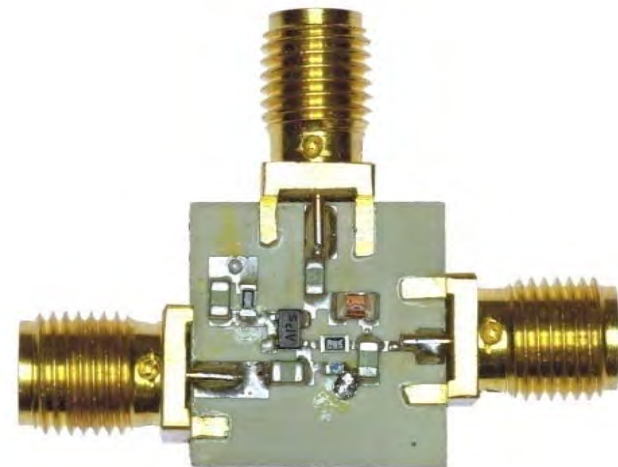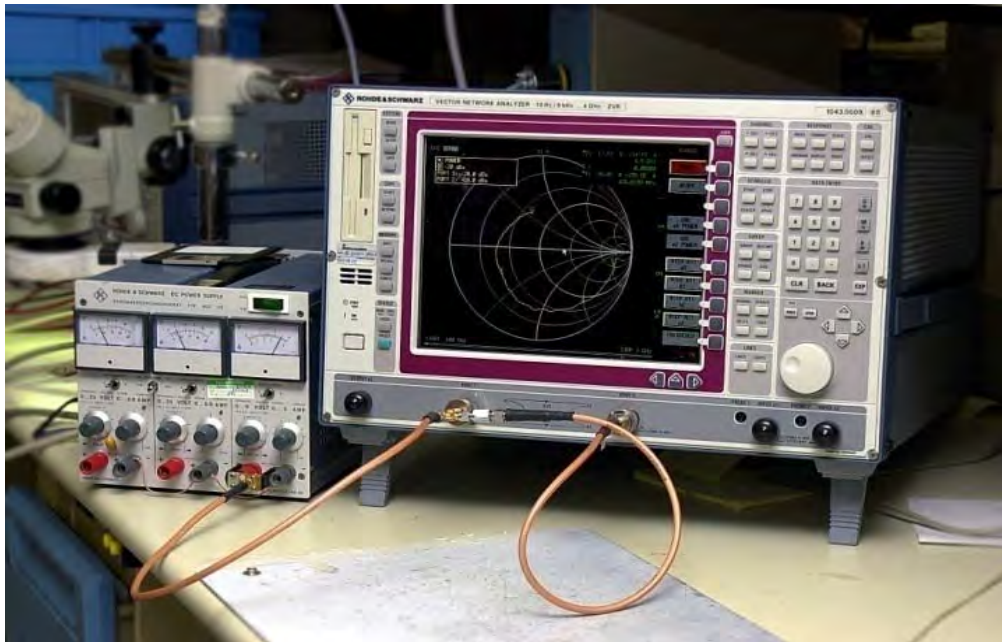
Typical measurement setup for evaluation of large signal parameters (R&S vector analyzer and the test fixture for the transistor of choice), Agilent now calls this X Parameters
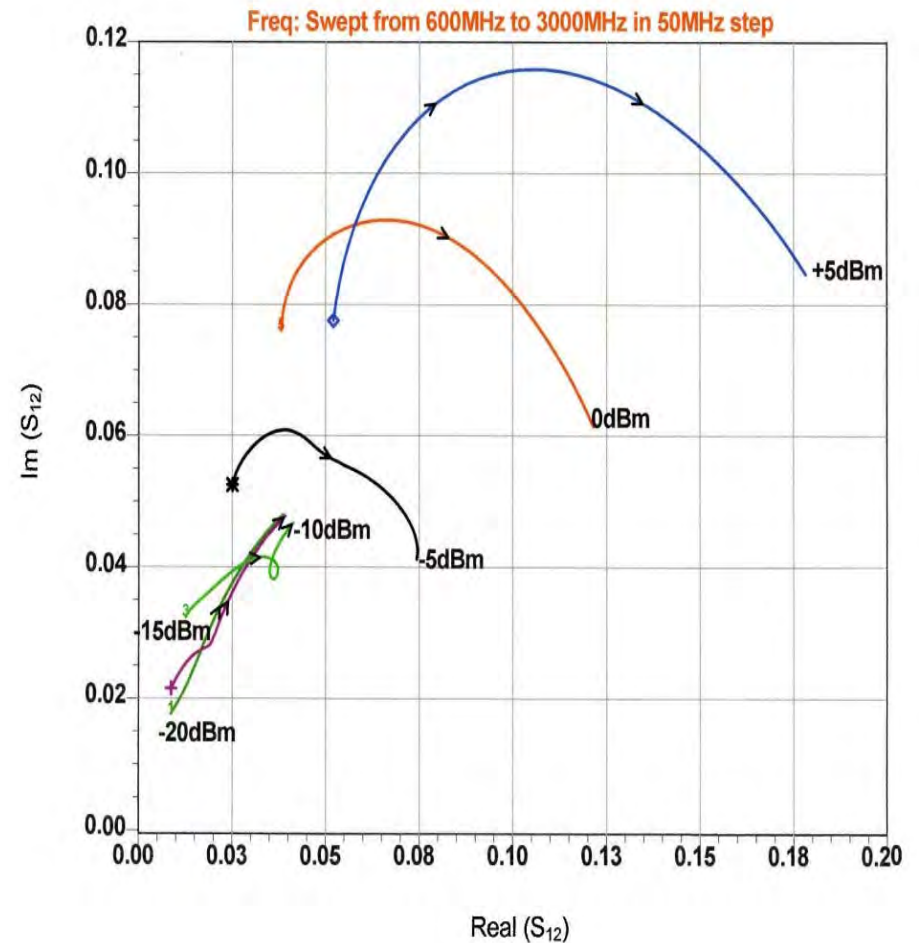
The bias, drive level, and frequency dependent S parameters are then obtained for practical use
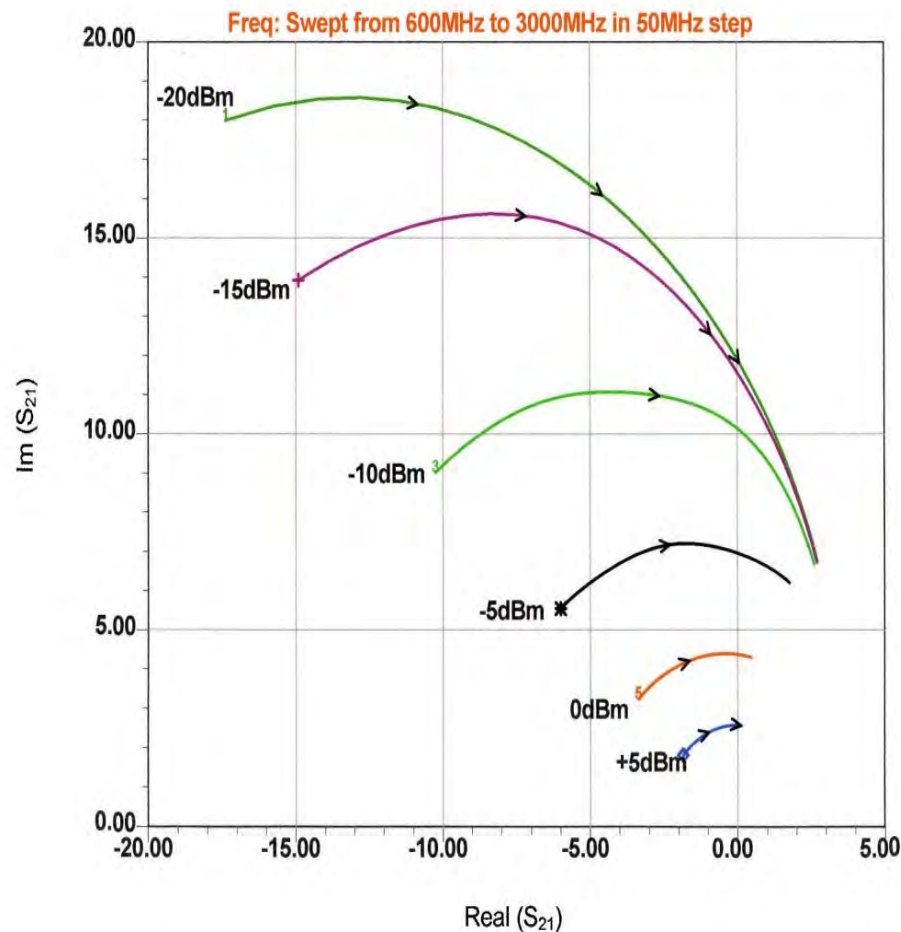
Measured large-signal S_{11} of the BFP520
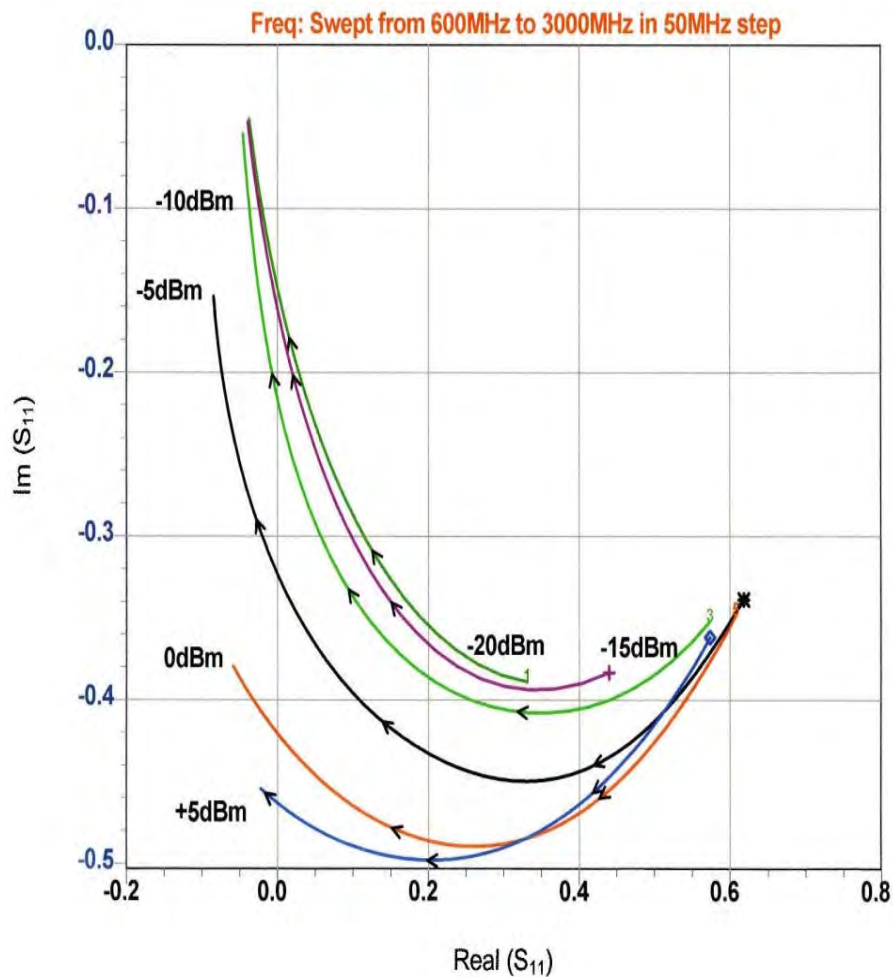
Measured large-signal S_{12} of the BFP520

The bias, drive level, and frequency dependent S parameters are then obtained for practical use

Measured large-signal S_{21} of the BFP520
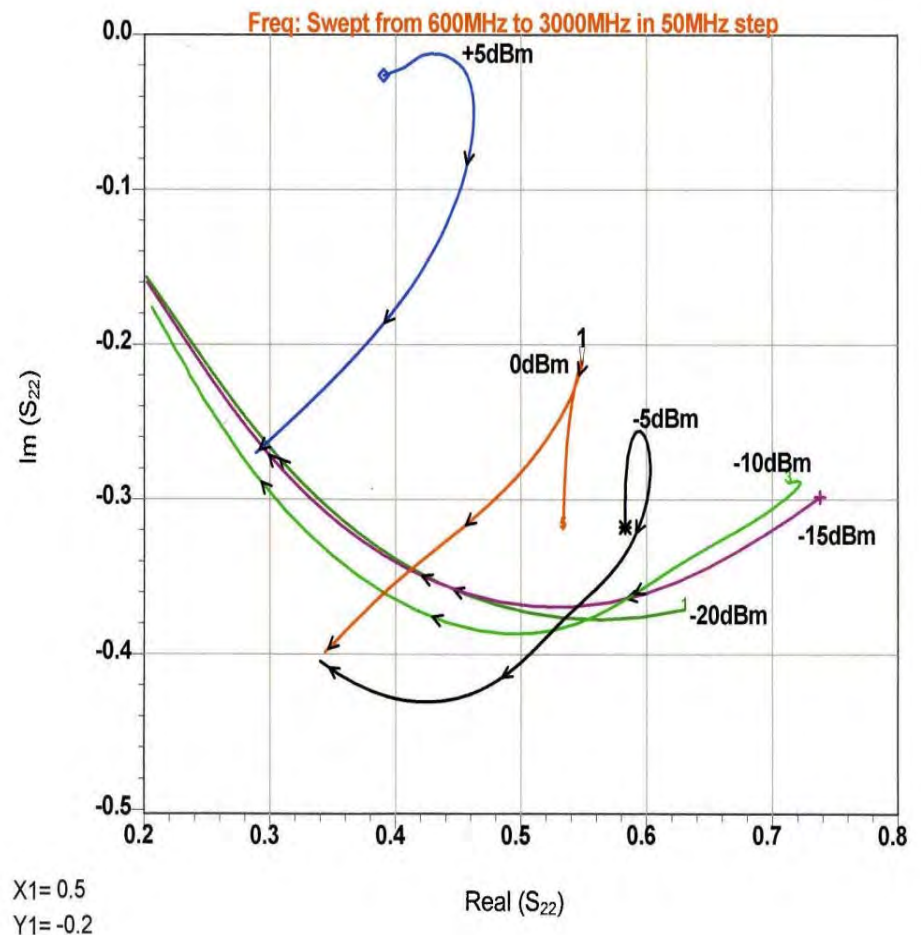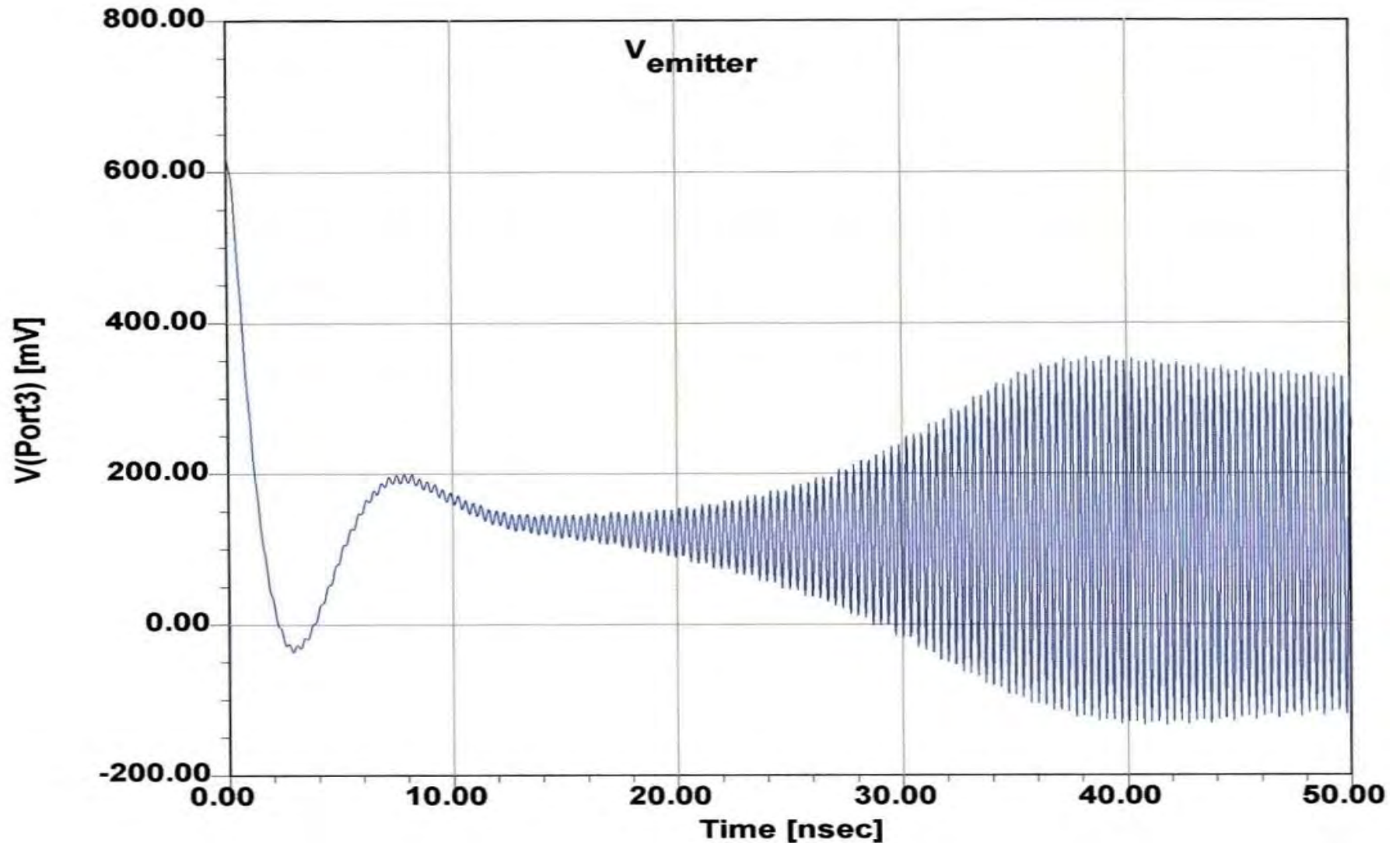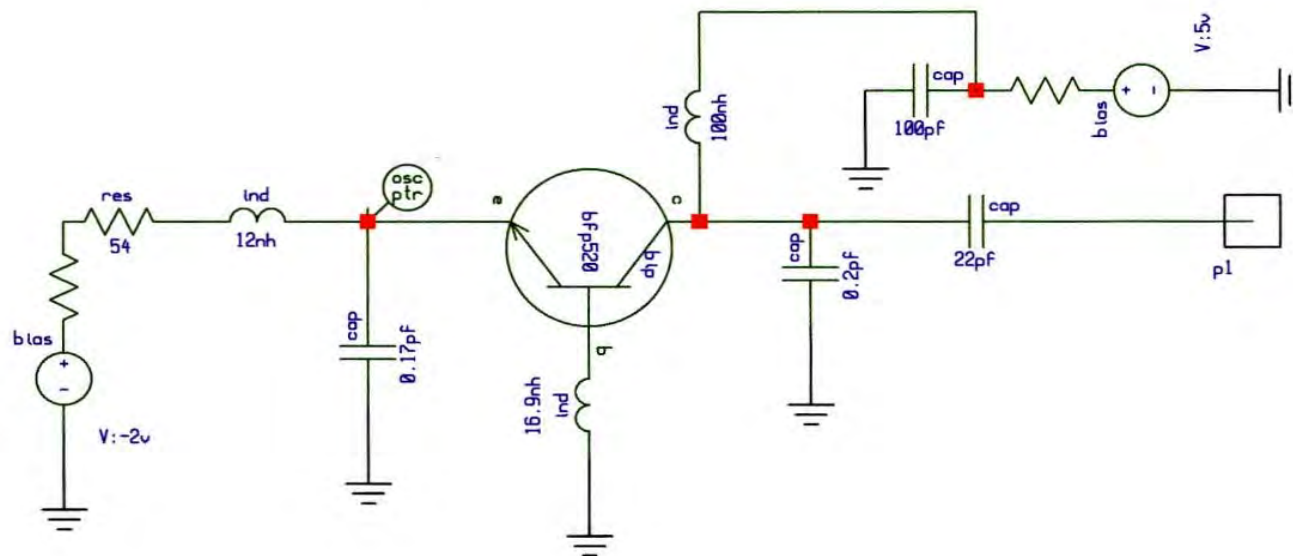
Measured large-signal S_{22} of the BFP520

Typical transient simulation of a ceramic resonator-based high-Q, 1GHz oscillator (node of the voltage for display is taken from the emitter)

Example: 3000 MHz YIG Oscillator


Series-Feedback Oscillator

Series feedback oscillator: For oscillation condition, the base to ground inductance and the emitter to ground capacitance is required. The 12nH inductor acts a choke. The output is tuned and terminated into 50Ω
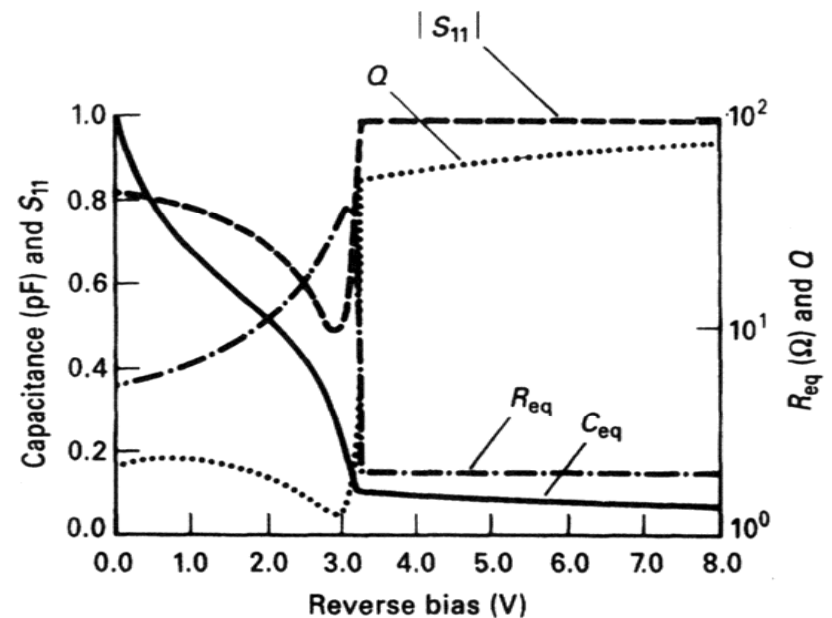
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➢ TUNING DIODE NOISE CONTRIBUTION
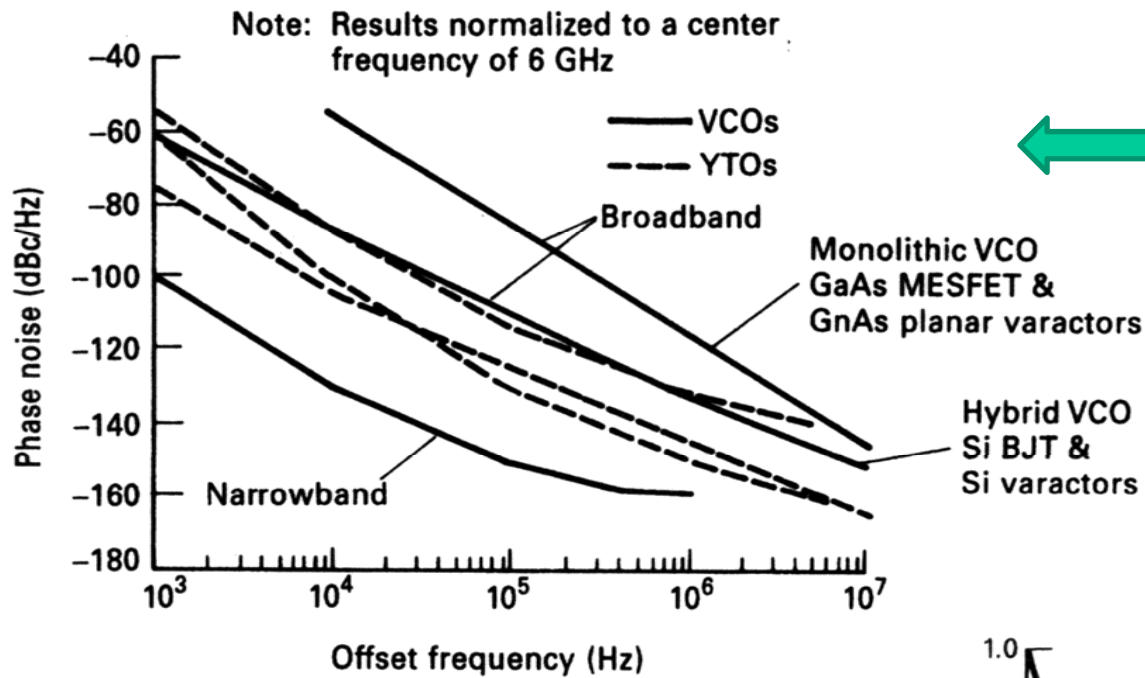
Phase noise comparison of different YIG and varactor tuned oscillators

Varactor parameters: capacitance, equivalent resistor, and Q, as well as the magnitude of S_{11}, as a function of reverse voltage
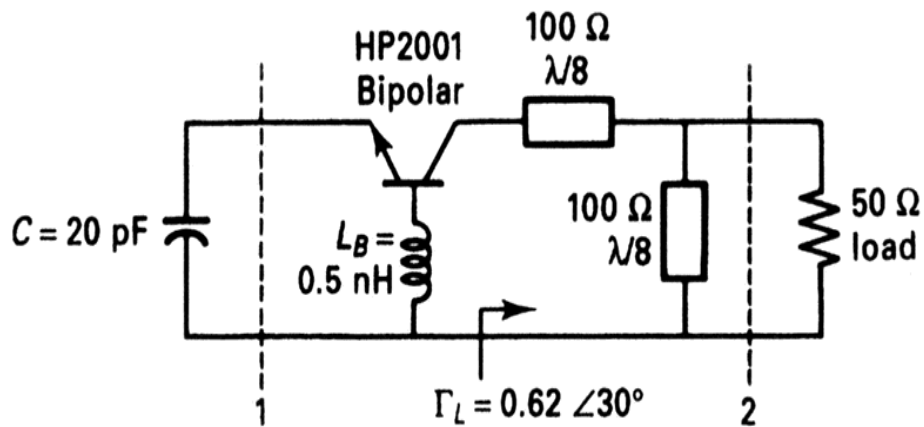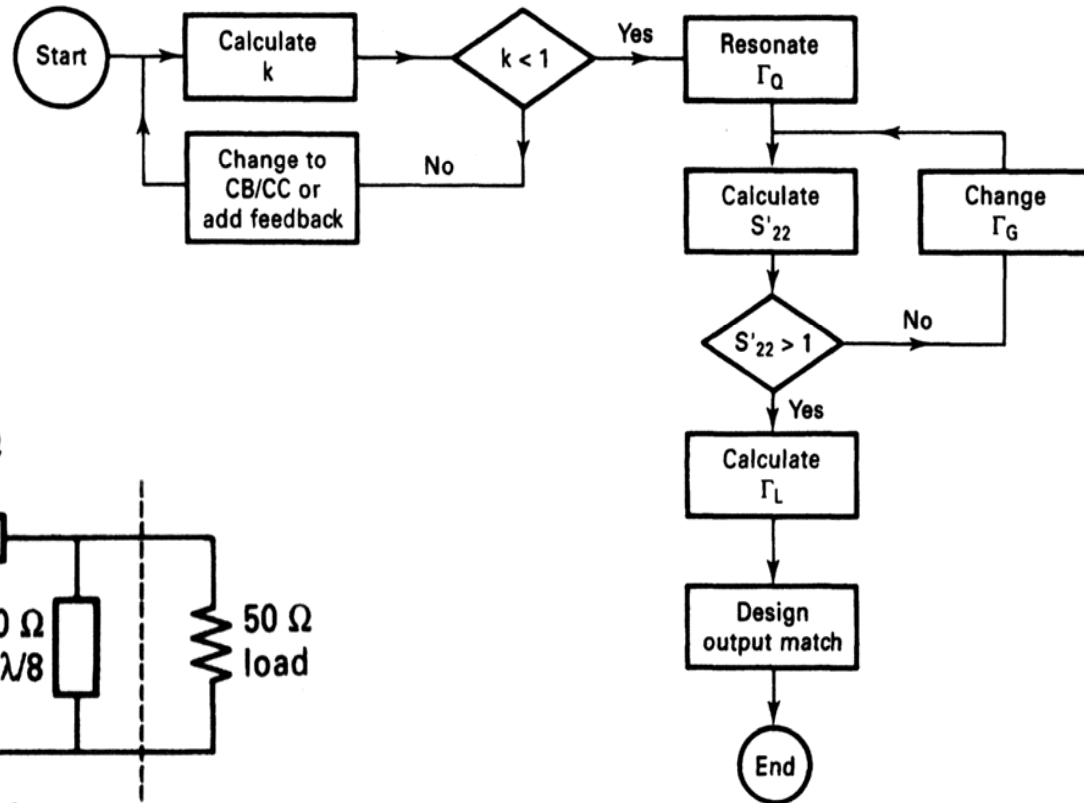
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➢ CLASSICAL LINEAR TWO-PORT OSCILLATOR DESIGN

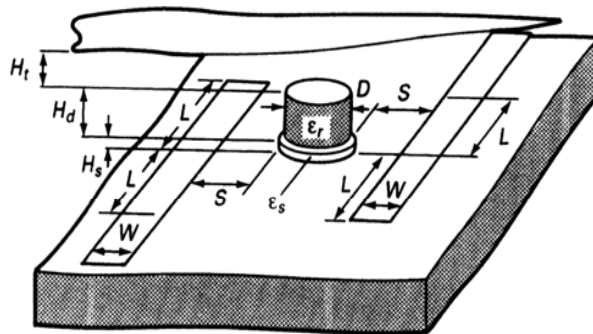*Oscillator design flowchart.*

Two Port Oscillator example at 2 GHz

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➢ MICROWAVE RESONATORS

SAW Surface Acoustic Wave Oscillators
DRO Dielectric Resonator-based Oscillators
YIG Oscillators

DRO on microstrip as Bandpass filter

Usually $L = \dfrac{\lambda_g}{4}$
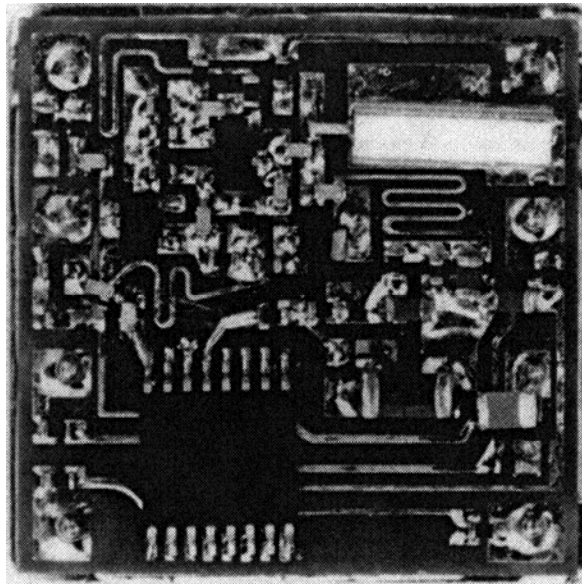
Ceramic Resonator Oscillator

Mounting and heater
Power input feedthroughs
Coupling loop
YIG sphere
Soft iron plate
GaAsFET "Negative resistance" oscillator
Output connector (through base plate)
Surface mount and chip components on ceramic "thick film" circuit

Fine tuning "FM" coil
Main coil
Main coil
Oscillator power
Heater power
dc power
RF output
Overlapping Mu – metal cans
YIG sphere
Cross section

The yttrium–iron–garnet (YIG) sphere serves as the resonator in the sweep oscillators used in many spectrum analyzers
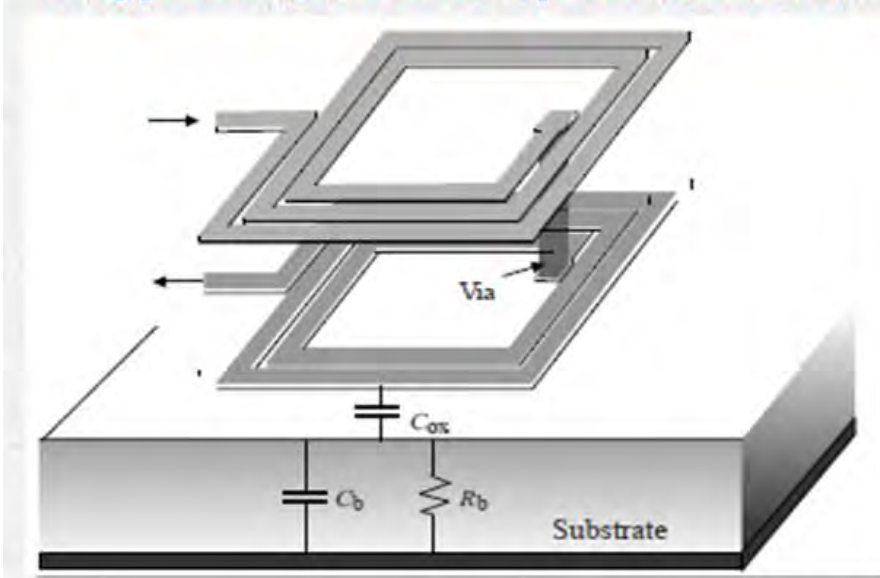
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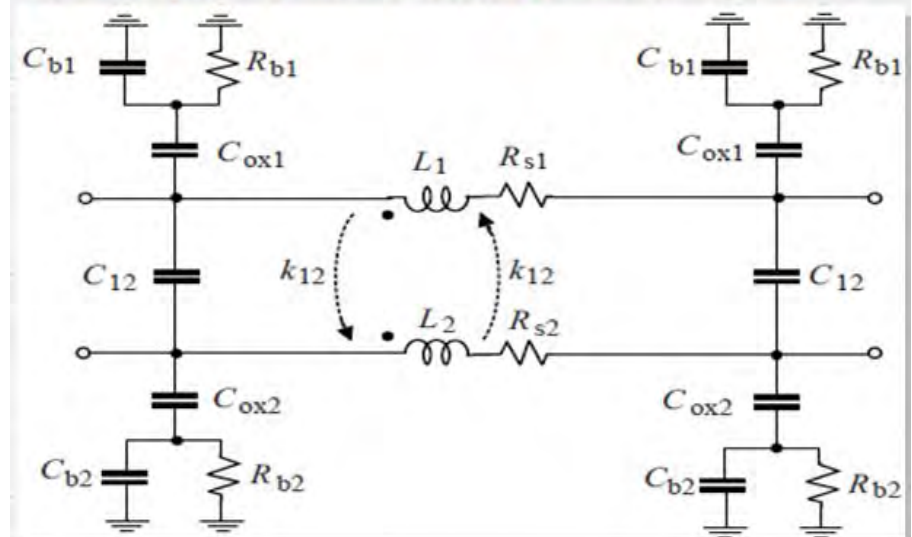
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➢ **A NOVEL TUNABLE ACTIVE SPIRAL INDUCTOR**

A typical sketch of spiral inductor

A lumped model of spiral inductor

On-Chip Spiral Inductor
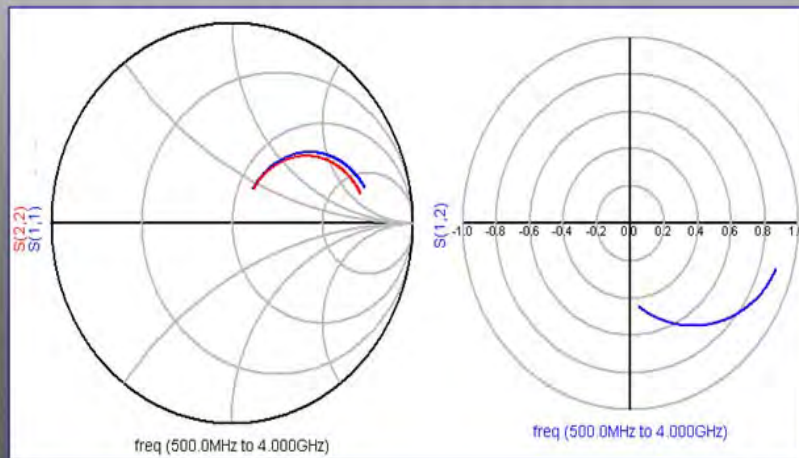
N= 3 turns
W=15 um
S= 20 um,
Ri=200 um
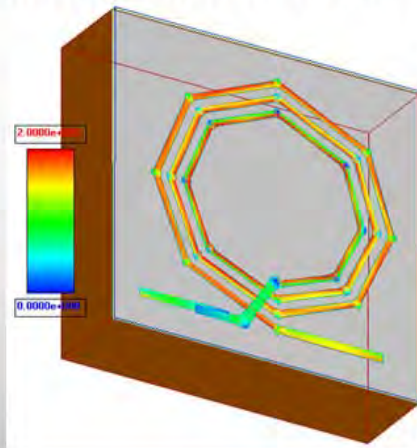
On-Chip Inductor
requires large dies
area :   Not a
Cost-Effective
Solutions !!!

EM Simulated Current

EM Simulated S-Parameters

freq (500.0MHz to 4.000GHz)

freq (500.0MHz to 4.000GHz)

The agreement between the measured S-parameter based and EM-Simulated values is extremely good. However, the physical size of the chip inductor requires a large die area, thus not a cost-effective solution

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➢ THE ACTIVE INDUCTOR USING A GYRATOR

B. D. H. Tellegen of Philips Research Laboratory proposed a new 2-port network element, a Gyrator in 1948, which exhibits a immittance conversion property, needed to generate an synthesized active inductor using transistors.
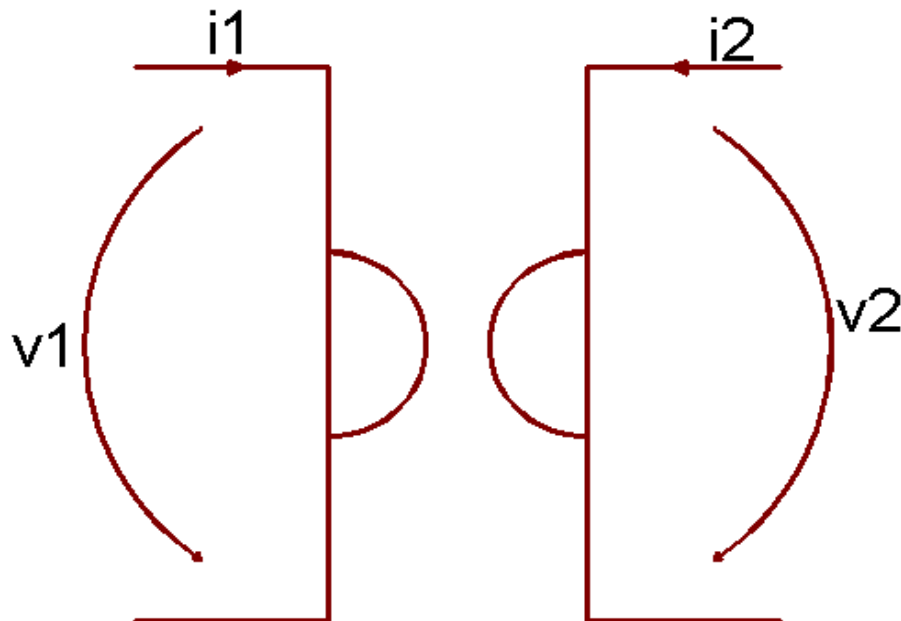
$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} 0 & g \\ -g & 0 \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \end{bmatrix}$$

Where 'g' is called gyration capacitance

An admittance Y connected to the secondary terminals is converted to its dual g^2/Y, this phenomena is called immittance conversion, C transforms into L, parallel tuned circuit into series tuned circuit

Tunable Active Inductor (TAI)

✓ **Integrable and Compact**

✓ **Cost-Effective**

✓ **Power-Efficient Solutions**

TAI: Design Challenges

✓ **High Power Consumption**

✓ **Noise Figure & Instability**

✓ **Low Dynamic Ranges**

TAI using Gyrator
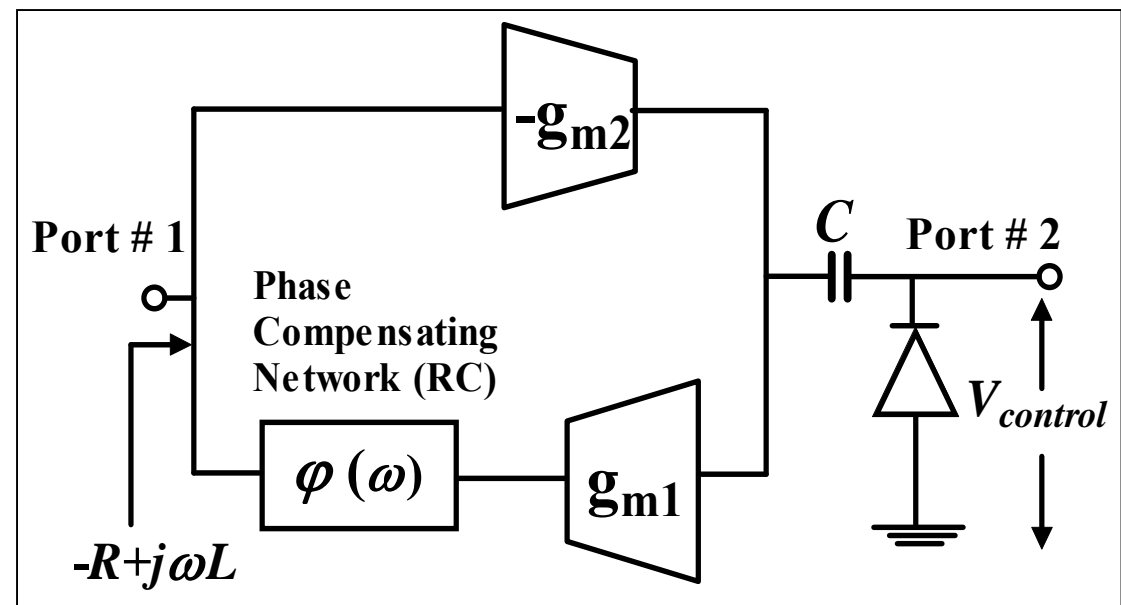
Phase shift network $\varphi(\omega)$ is required in TAI topology for suppressing the higher order modes and self oscillation

OUTLINE

This Figure shows the typical Active Inductor Oscillator (AIO), includes a stable active inductor within a conventional integrated LC oscillator
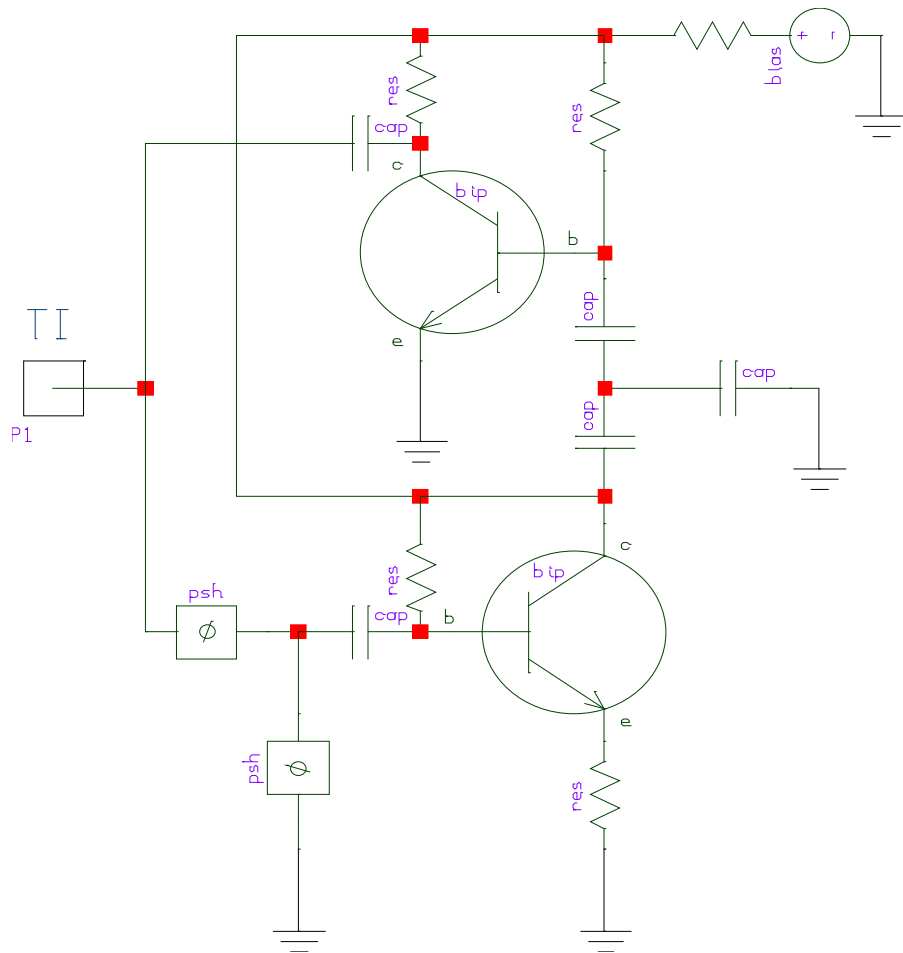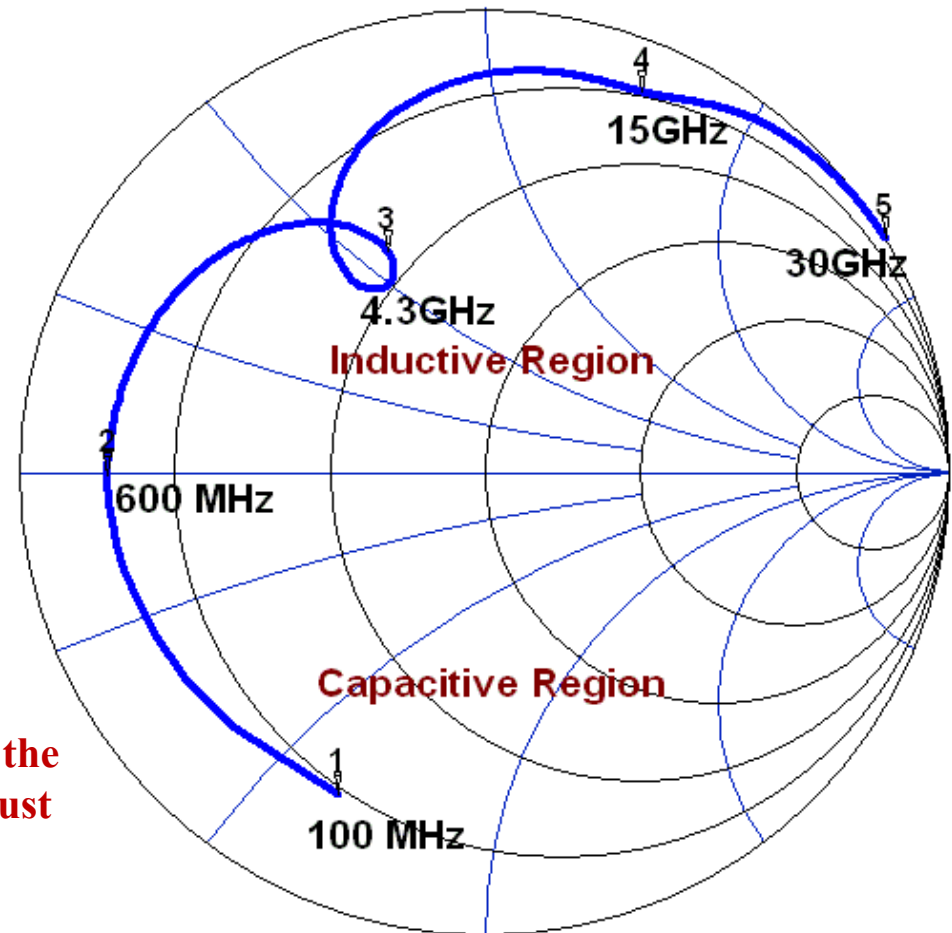
OUTLINE

➤ SYNTHESIZED INDUCTOR BEHAVIOUR

Typical Schematic of Synthesized Inductor

Impedance Plot

Impedance plot reveals the inductive behavior of the circuit from 600MHz (#2) to 30GHz (#5). Care must be taken to avoid the encircling and crossing at 4.3GHz (#3), which limits the applications.

➢ **Why use an Active Inductor instead of a Spiral Inductor?**

L_1

Layout

L_2
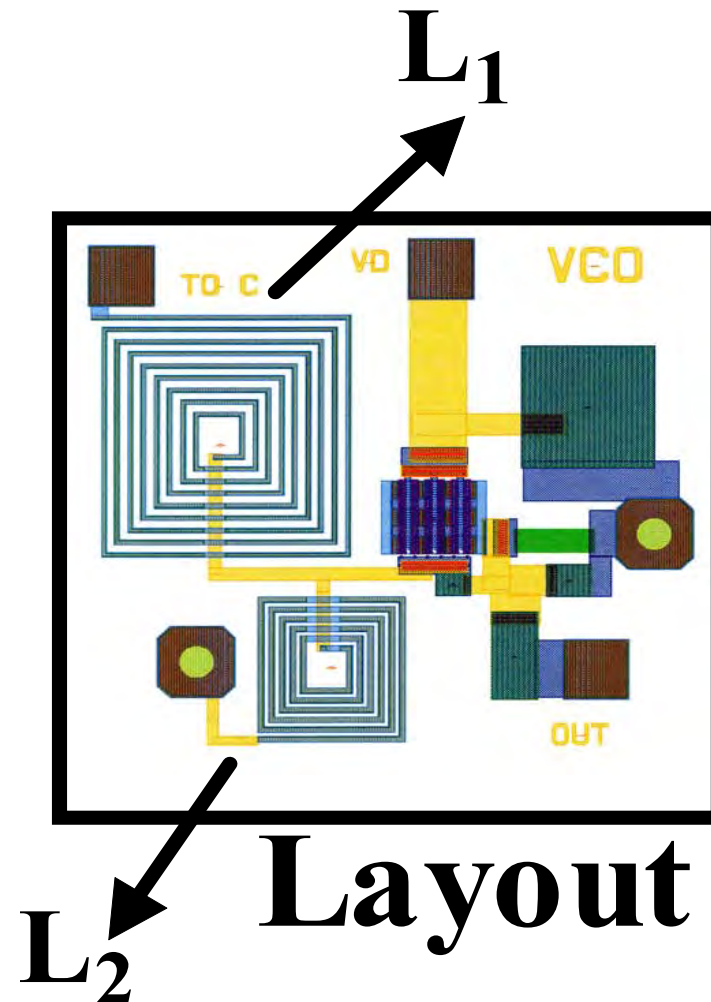
Physical size of spiral inductor

Physical size of spiral inductor

1.1nH Inductor

Bond Pad

1pF Capacitor

RF BJT
(20μm x 0.4μm x 10)

Minimum
Geometry
BJT
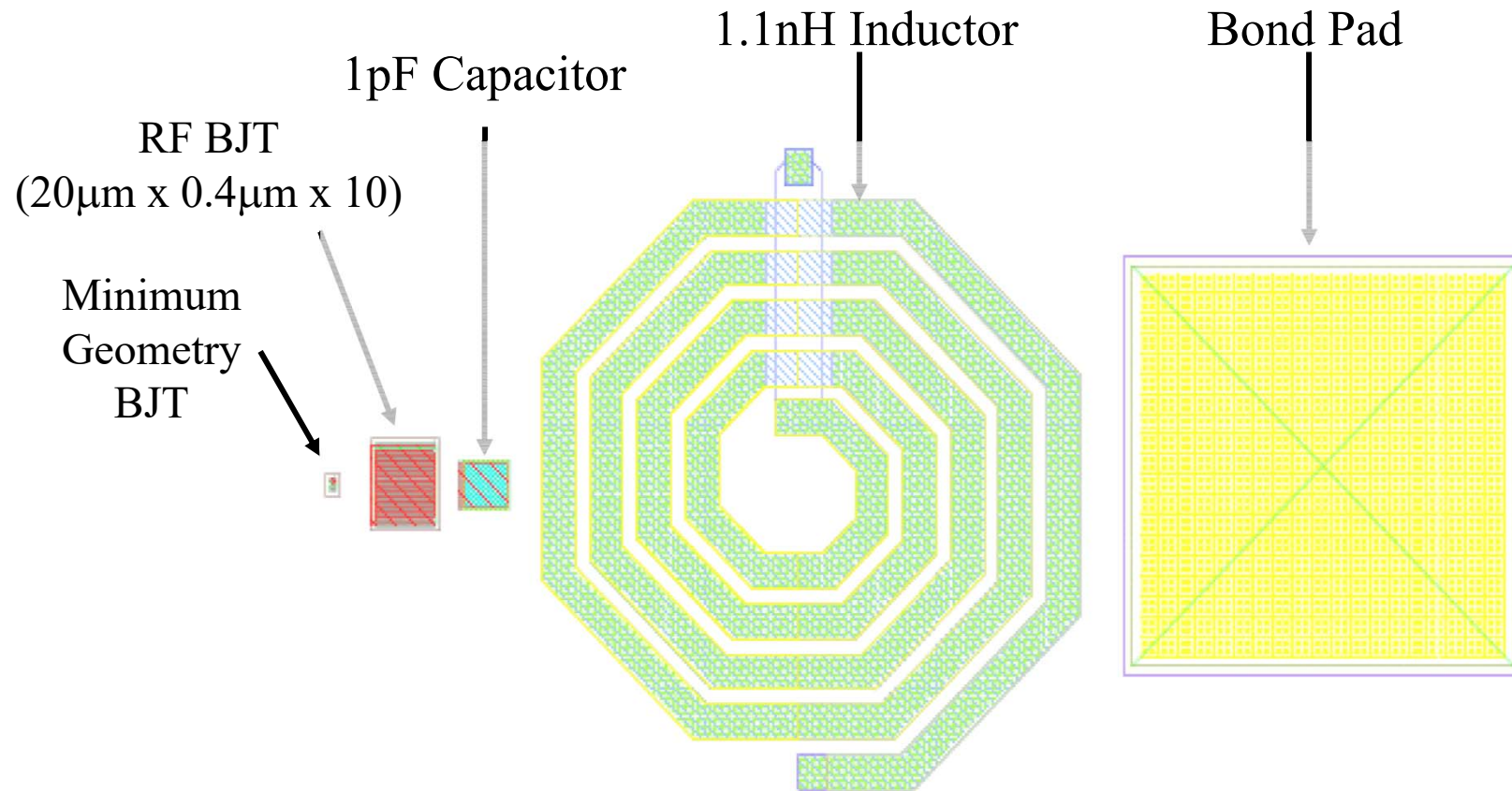
Why we avoid the use of on-chip inductors!

Phase Noise Plots

Surface of phase noise @ 1MHz offset from the carrier
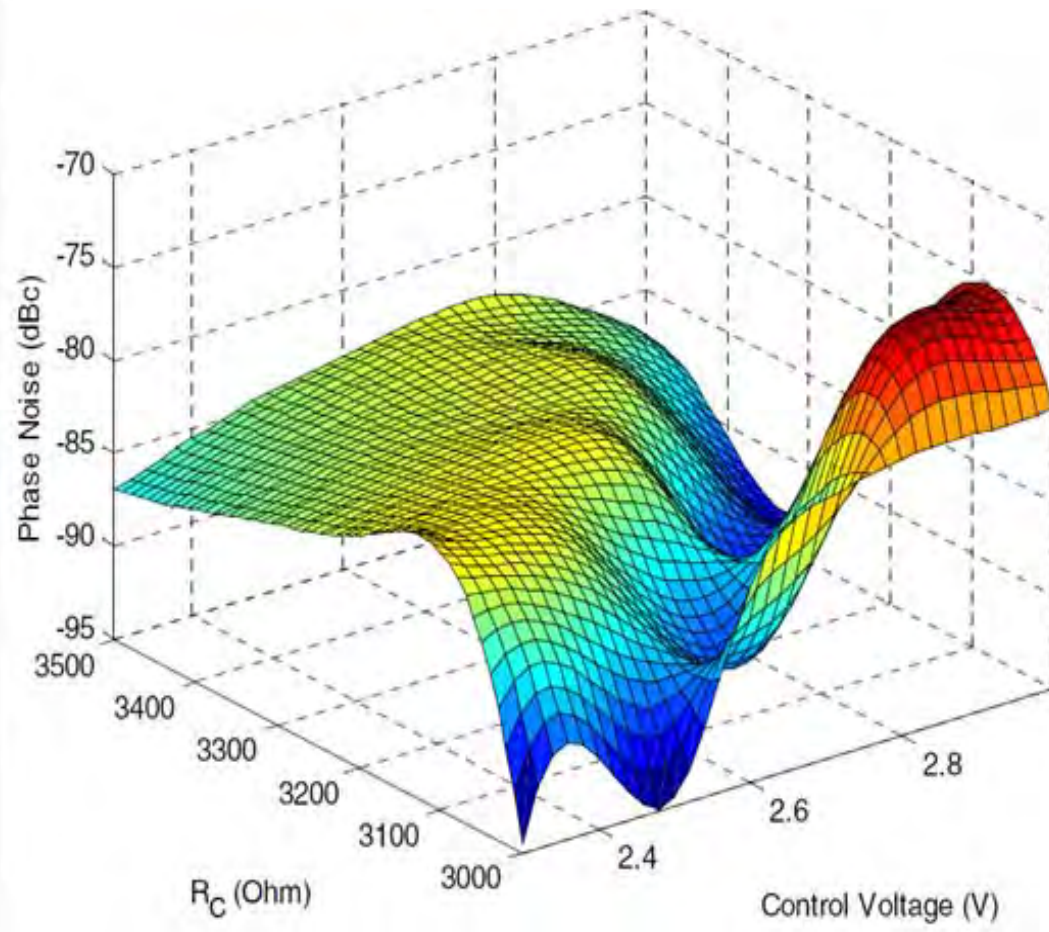
Advantages:
1. Broadband VCO Operation (Tuned L)
2. Multi-Band VCO Operation (Switched L)

Drawbacks:
1. DC Power Consumption
2. Poor Phase Noise Performances
3. Limited Large Signal Performances

➢ **Phase Noise Contribution of The Various Parts of the Oscillator Using an Active Inductor**

SIMULATED TUNBALE INDUCTOR USING SiGe HBT DEVICE
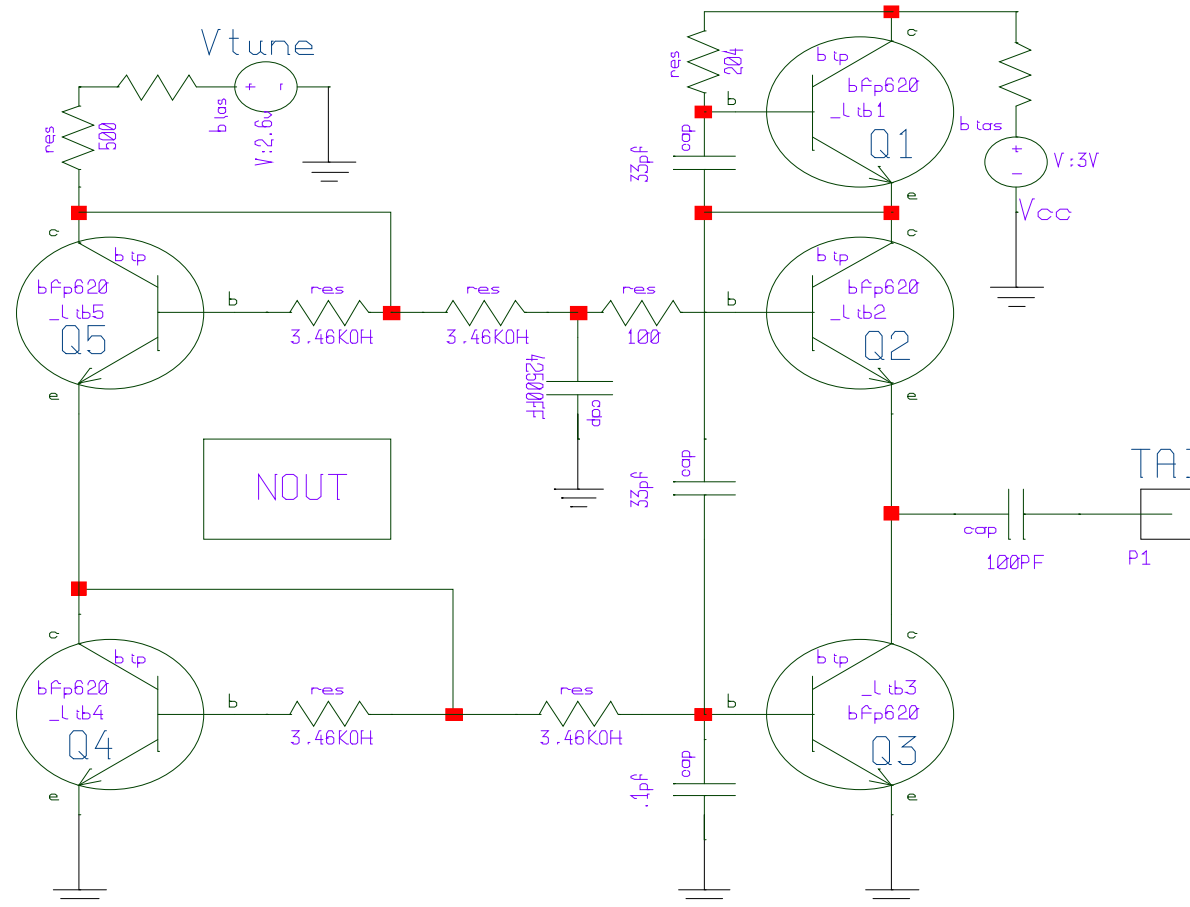
This Figure shows a schematic of a transistorized inductor using SiGe HBT (BFP 620) from Infineon . The reason for using a high cut-off frequency (f_t=75 GHz) SiGe HBT transistor is to minimize the package parasitic effects and allow comparative evaluations of the 1.9 GHz varactor-tuned and synthesized inductor-tuned LC oscillator using discrete components for experimental validations.

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➢ COMPARISION: PASSIVE &ACTIVE INDUCTOR

Passive and Active Inductors

| Performances	Passive Inductor (Spiral)	Active Inductor (Simulated: Active Device
Q-factor	Low Q: Q-factor can be improved by differential method but added cost	High Q: active inductor offers higher Q than the passive spiral inductor
Tunability	Fixed/Limited	Large tuning range
Die-Area	Large die-area	Small die-area
Power-Consumption	Zero	Significant: consumes power for generating active inductance, resulting to high Q that may offsets the power consumptions
Linearity	Good Linearity $Z = j\omega L$	Poor Linearity: driven under large-signal condition, causing shift in operating point, distortions, and impedance fluctuations
Noise	Superior: good phase noise performance	Poor : poor phase noise performance
EMI	Significant: Due to EM coupling in spiral inductors	EMI insensitive
Floor-Planning	Poor: large die- area makes difficult floor-planning	Not required

Typical Schematic of Switched Mode and VCO Circuit
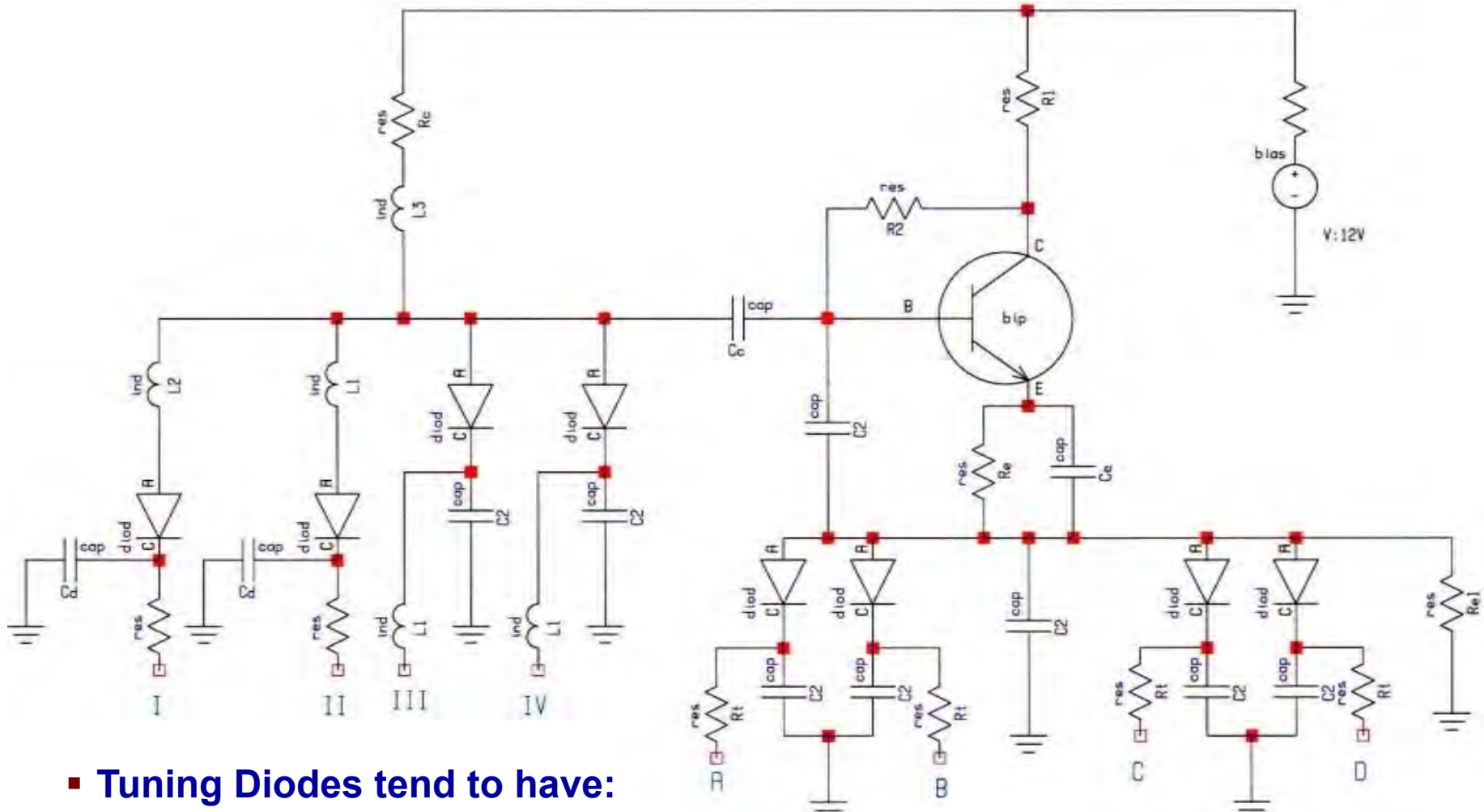
- **Tuning Diodes tend to have:**
  - ➢ **Low Q at RF & MW frequencies**
  - ➢ **Limits Tunability Due to Package Parasitic**
  - ➢ **Changes characteristics due to low SRF**
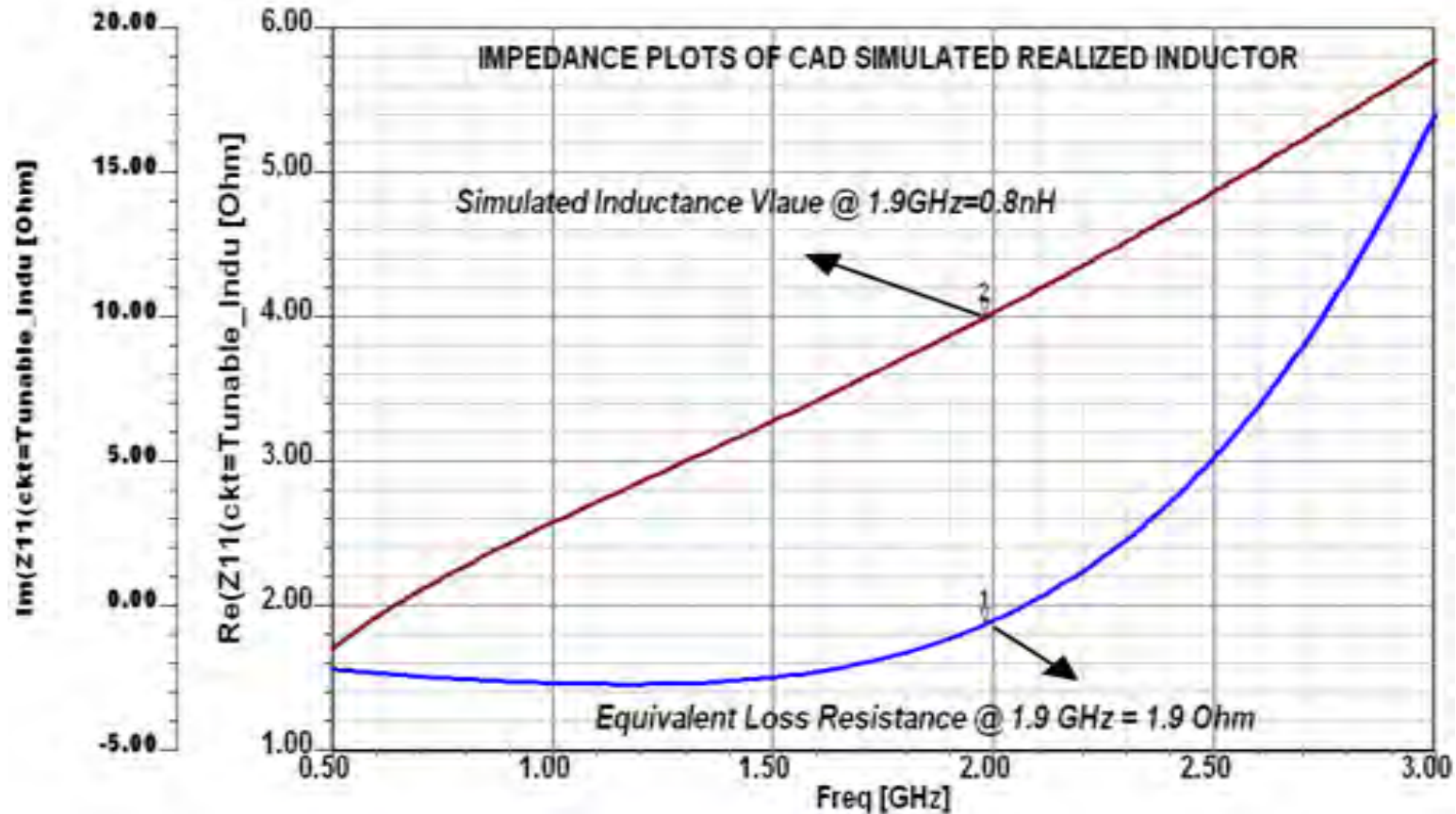- **Active Tunable Inductor ATIs can over come the above difficulties**

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**Noise sources of oscillators being mixed on the carrier**
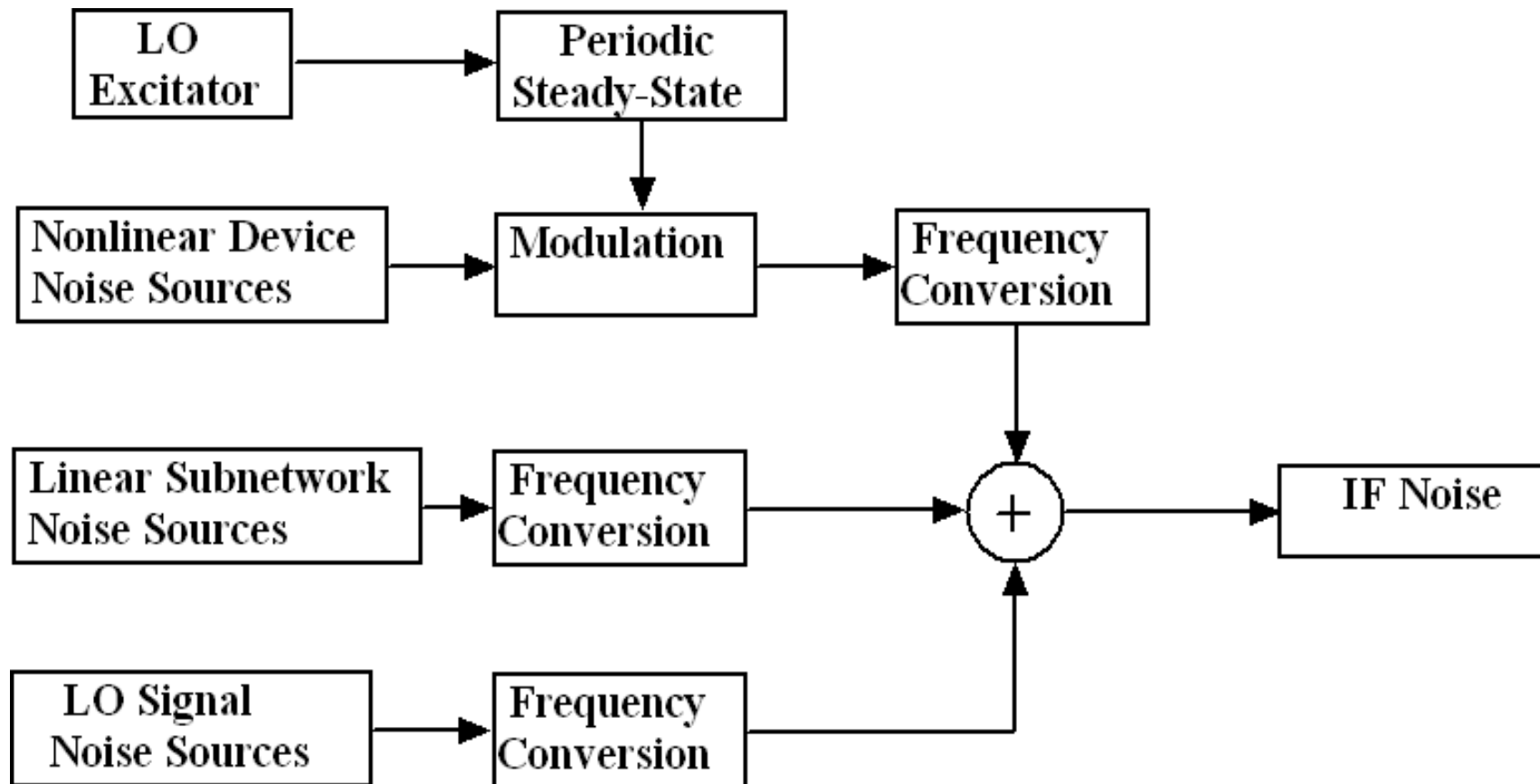
This Figure shows the typical plot of reactance and equivalent loss resistance of the synthesized inductor using high cut-off frequency (f_t=75 GHz) SiGe HBTs.

IMPEDANCE PLOTS OF CAD SIMULATED REALIZED INDUCTOR

Simulated Inductance Vlaue @ 1.9GHz=0.8nH

Equivalent Loss Resistance @ 1.9 GHz = 1.9 Ohm

As shown in Figure, the value of the realized inductance and associated equivalent loss resistance are 0.8nH and 1.9Ω at 1.9 GHz for the operating DC bias condition (3V, 1.8mA) and V_{tune} (2.5V). The operating DC bias and V_{tune} are adjusted in such a way that realized equivalent noise resistance must be positive to avoid the multi-mode oscillations caused by the regenerative effect (if the simulated loss resistance associated with realized inductor is negative in value).

Realized Inductor-Tuned Osillator
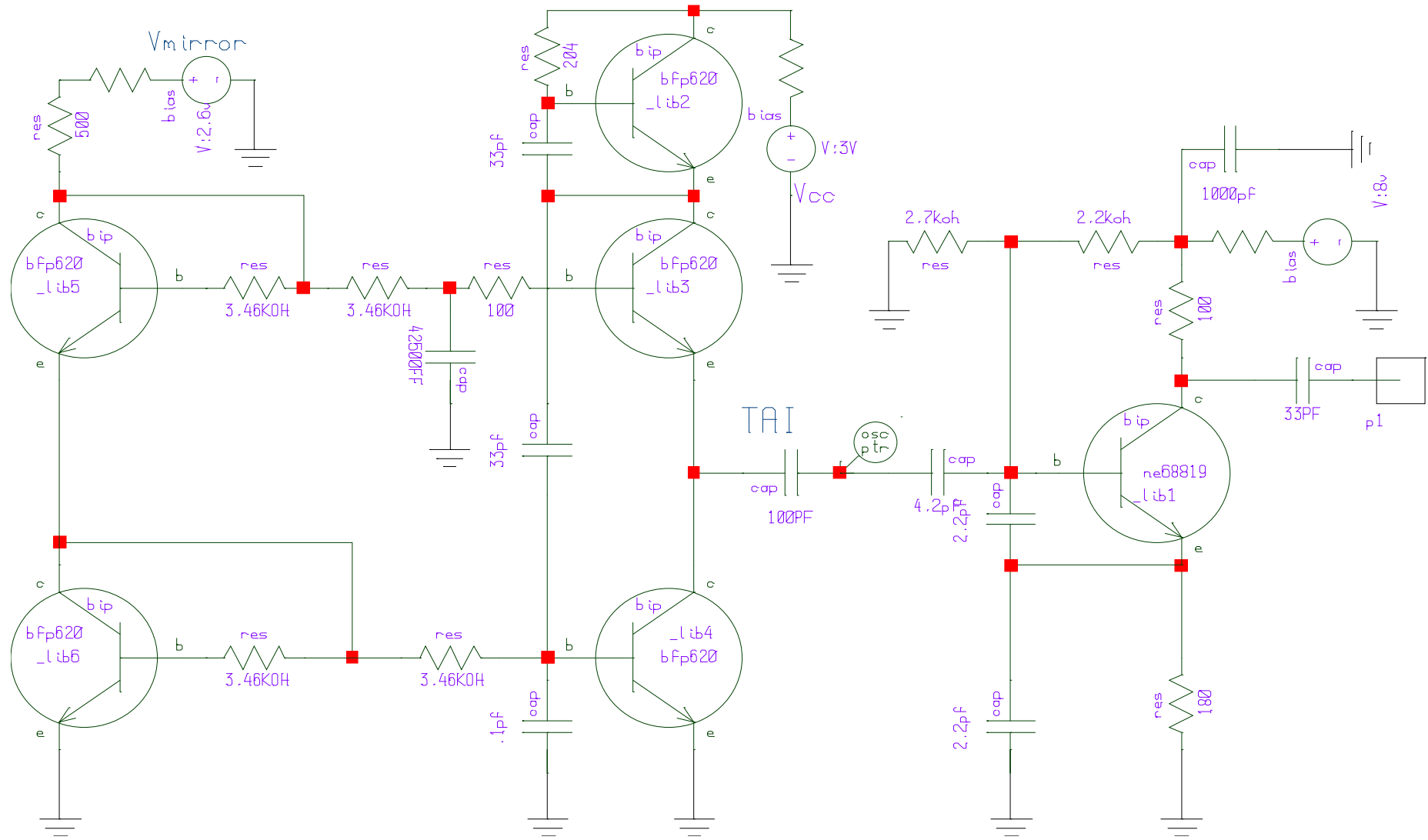
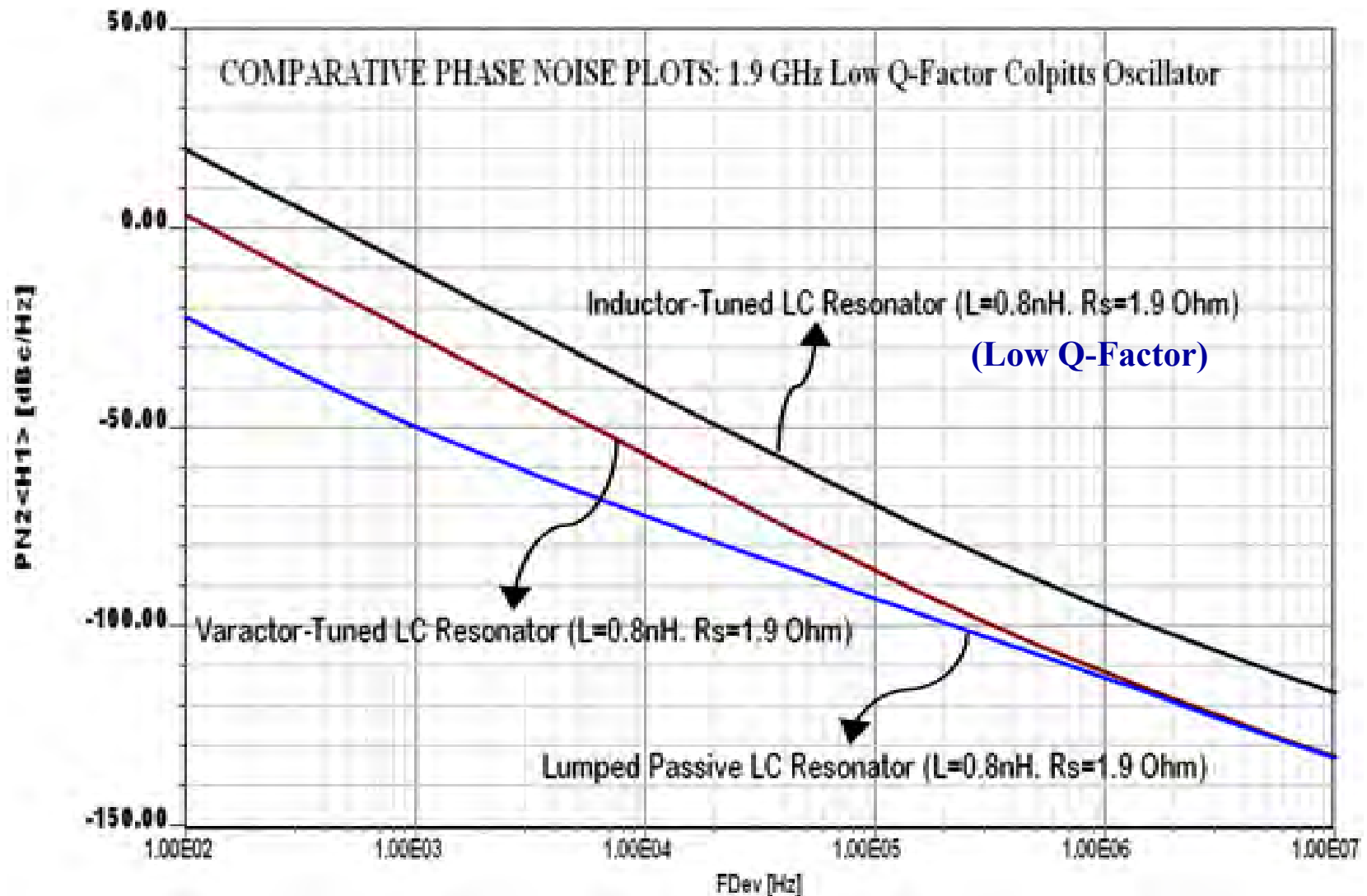**Schematic of the Colpitts oscillator circuits using CAD simulated inductor (0.8 nH, 1.9 Ω)**

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This plot shows the comparative phase noise plots for the LC Colpitts oscillator using the lumped LC resonator, the varactor-tuned lumped LC resonator and the synthesized inductor-tuned resonator network for identical inductance value and loss resistance (0.8 nH with series loss resistance 1.9 Ω).

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Capacitor loaded gyrator based active inductor resonator with noise source
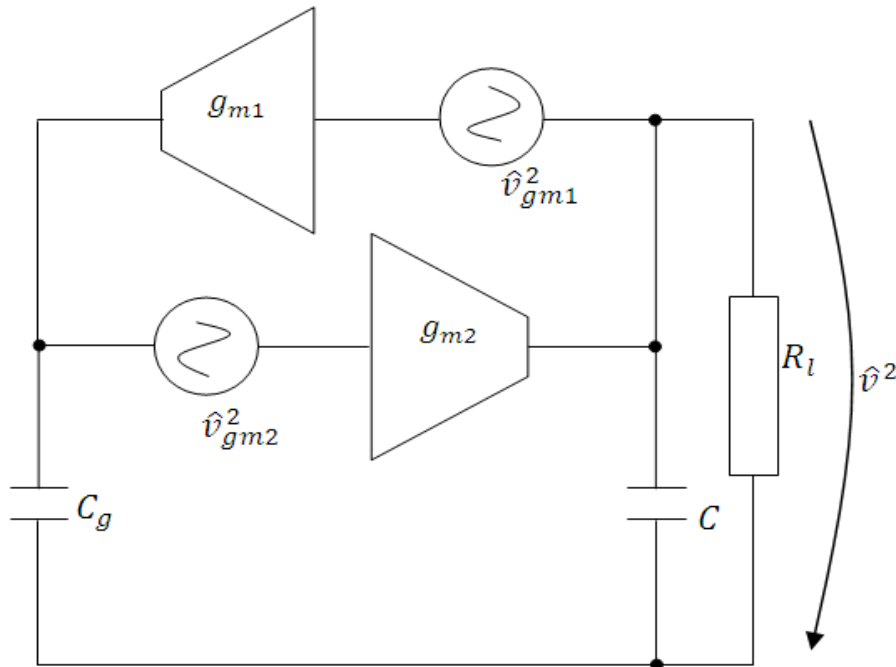
A simplified circuit of active inductor resonator with noise sources
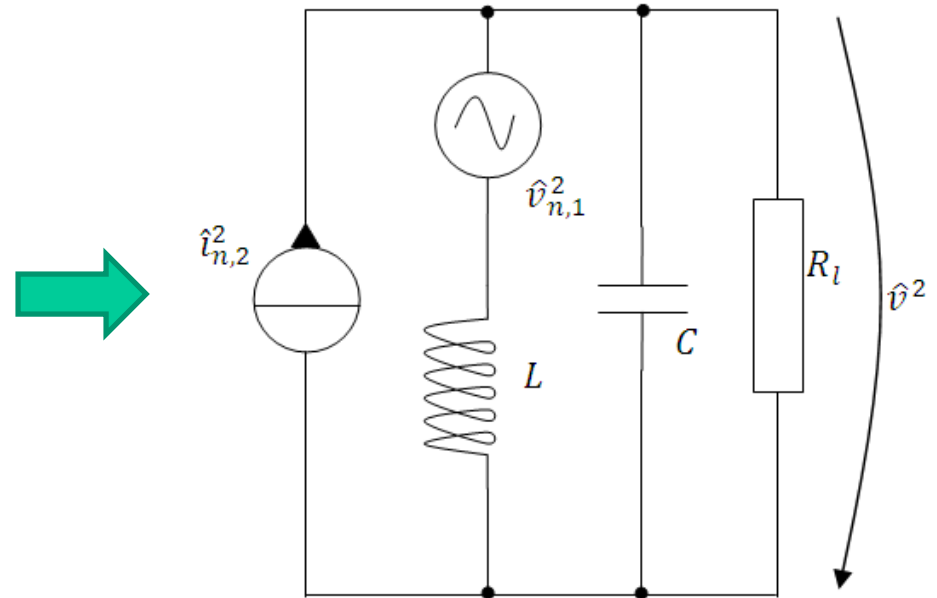
$$\hat{v}_{n,1}^2 = \hat{v}_{gm1}^2 \cong \frac{4kT\lambda}{g_m}$$

$$\hat{i}_{n,2}^2 = \hat{g}_{gm2}^2 \hat{v}_{gm2}^2 = 4kT\lambda g_m$$

V_{gm1} and V_{gm2} are the equivalent noises from the transconductances of the Gyrators

Since

$$\hat{v}_n^2 \cong 4kTr_n|R|, \quad \hat{i}_n^2 \cong 4kTg_n|G|$$

where _R_ and _G_ are the negative resistance and conductance values, and the coefficients r_n and g_n are frequency dependent relative noise resistance and conductance (these give a comparative value of how much noise the active negative resistor produces compared to a passive resistor of the same value).
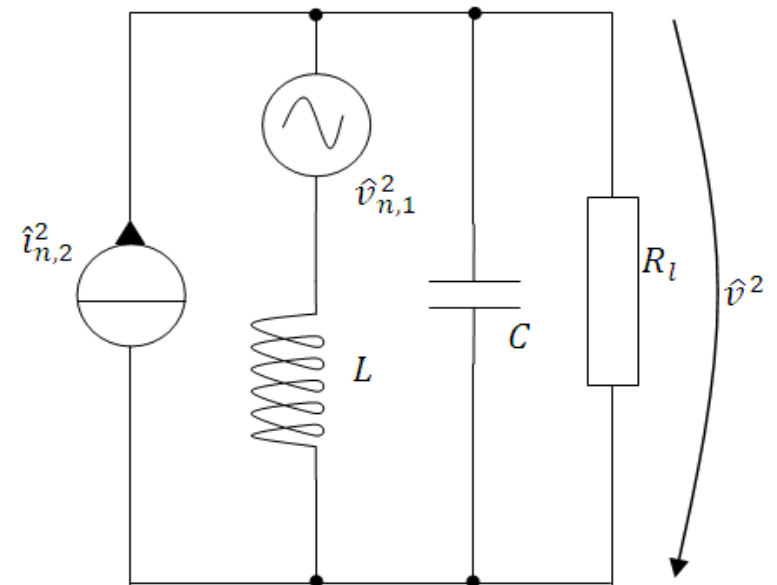
The total noise voltage spectral density of the active inductor resonator is
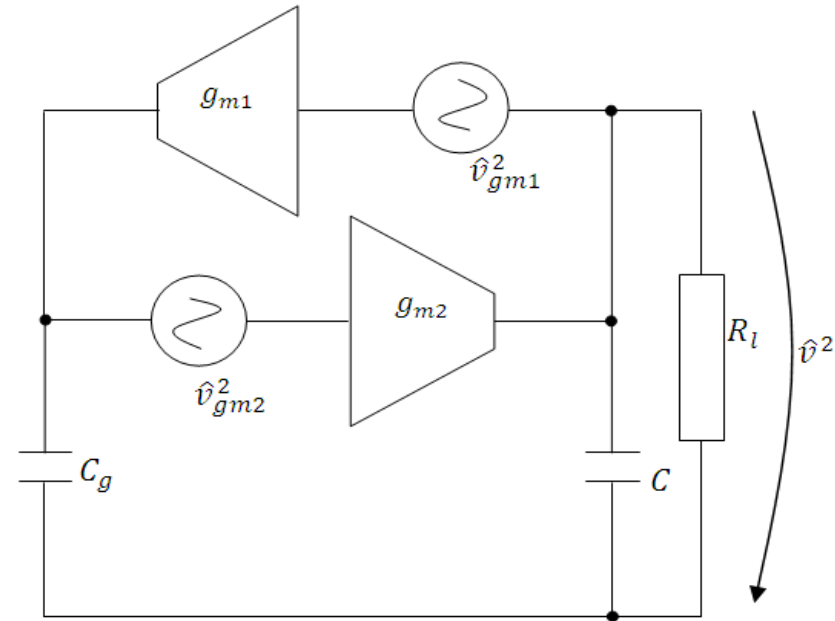
$$\hat{v}^2 = \frac{\hat{v}_{gm1}^2 + \hat{v}_{gm2}^2 g_{m2}^2 \omega^2 L^2}{\omega^2 L^2 G_l^2 + (\omega^2 LC - 1)^2} \Longrightarrow \hat{v}^2|_{\omega \to \omega_0} = Q_l^2 \left(\hat{v}_{gm1}^2 + g_{m2}^2 \hat{v}_{gm2}^2 \omega^2 L^2 \right) = 4kT\lambda Q_l^2 \left(\frac{1}{g_{m1}} + \frac{g_{m2}}{\omega_0^2 C^2} \right)($$

The time average Q-factor of active inductor is $$\frac{\omega_0(1 + r_n, g_n)}{\lambda g_{m2}\left(\frac{1}{C} + \frac{1}{C_g}\right)}$$

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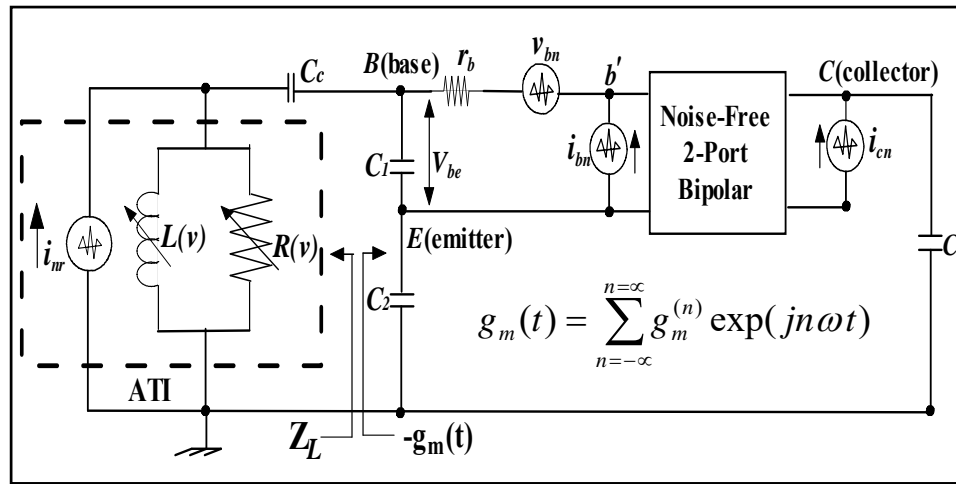# ➤ ACTIVE INDUCTOR NOISE, cont'd.

The time average normalized noise power of an active inductor resonator can be determined by

$$\overline{v^2} = \int_0^\infty \hat{v}^2 df = \frac{\hat{v}^2_{gm1}}{2\pi} \omega_o \int_0^\infty \frac{d\omega}{[(\omega^2 LC - 1)^2 + \omega^2 L^2 G_l^2]} + \frac{\hat{g}^2_{gm2} \hat{v}^2_{gm2}}{2\pi} \int_0^\infty \frac{\omega^2 L^2 d\omega}{[(\omega^2 LC - 1)^2 + \omega^2 L^2 G_l^2]}$$

$$\overline{v^2} \approx \frac{kT\lambda}{C} Q_l \left(\frac{g_{m2}}{g_{m1}}\right)^{1/2} \left[\left(\frac{C_g}{C}\right)^{1/2} + \left(\frac{C}{C_g}\right)^{1/2}\right] = \frac{kT\lambda}{C} Q_l \frac{\omega_o}{g_{m1}}(C + C_g)$$

The total noise voltage power within 1 Hz bandwidth can be described by

$$\left|\overline{e_n^2(\omega)}\right|_{\omega=\omega_0} = \overline{[e_n^2(\omega_0)]}_{-gm1} + \overline{[e_n^2(\omega_0)]}_{-gm2}$$

The first term is related to the active inductor noise due to the active inductor and the second term is related to negative resistance generative active device.

After some lengthy calculations and approximations, adding shot noise, flicker noise and the loss resistor, the equivalent expression of the phase noise is given by

$$\left|\overline{e_{gm2}^2(\omega)}\right|_{\omega=\omega_0} = \left[4kTR\right] + \left[\frac{4qI_{c0}g_{m2}^2 + \dfrac{4K_f I_b^{AF}}{\omega}g_{m2}^2}{\omega_0^2 C_1^2[\omega_0^2(\beta^+)^2 C_2^2 + g_{m2}^2 \dfrac{C_2^2}{C_1^2}]}\right]$$

$$g_{m2} = \left[Y_{21}^+\right]\left[\frac{C_1}{C_2}\right]^q$$

$$\beta^+ = \left[\frac{Y_{21}^+}{Y_{11}^+}\right]\left[\frac{C_1}{C_2}\right]^p$$

$$g_m(t) = \sum_{n=-\infty}^{n=\infty} g_m^{(n)} \exp(jn\omega t)$$

The values of p and q depend upon the drive level.

After lengthy calculations and approximations, adding shot noise, flicker noise and the loss resistor, the equivalent expression of the phase noise is given by

$$L(\omega) = 10 \times \log \left[\left[k_0 + \left(\frac{k^3 k_1 \left[\frac{Y_{21}^+}{Y_{11}^+} \right]^2 [y]^{2p}}{[Y_{21}^+]^3 [y]^{3q}} \right) \left(\frac{1}{(y^2 + k)} \right) \right] \left[\frac{[1+y]^2}{y^2} \right] \right]$$

$$k_0 = \frac{kTR}{\omega^2 \, \omega_0^2 C_2^2 L_{active-inductor}^2 V_{cc}^2} \qquad k_1 = \frac{qI_c g_{m2}^2 + \frac{K_f I_b^{AF}}{4\omega} g_{m2}^2}{\omega^2 \, \omega_0^4 L_{active-inductor}^2 V_{cc}^2} \qquad k_2 = \omega_0^4 (\beta^+)^2 \qquad k_3 = \omega_0^2 g_{m2}^2$$

$$k = \frac{k_3}{k_2 C_2^2} \qquad L_{active-inductor} = \frac{C_1 C_2}{[C_1 + C_2] g_{m1}^2} \qquad g_m(t) = \sum_{n=-\infty}^{n=\infty} g_m^{(n)} \exp(jn\omega t)$$

$$g_{m2} = [Y_{21}^+] \left[\frac{C_1}{C_2} \right]^q \qquad \beta^+ = \left[\frac{Y_{21}^+}{Y_{11}^+} \right] \left[\frac{C_1}{C_2} \right]^p \qquad y = \frac{C_1}{C_2} \qquad$$ **The values of p and q depend upon the drive level.**

## Self-Injection-Locked 1.9 GHz Tunable Inductor Osillator
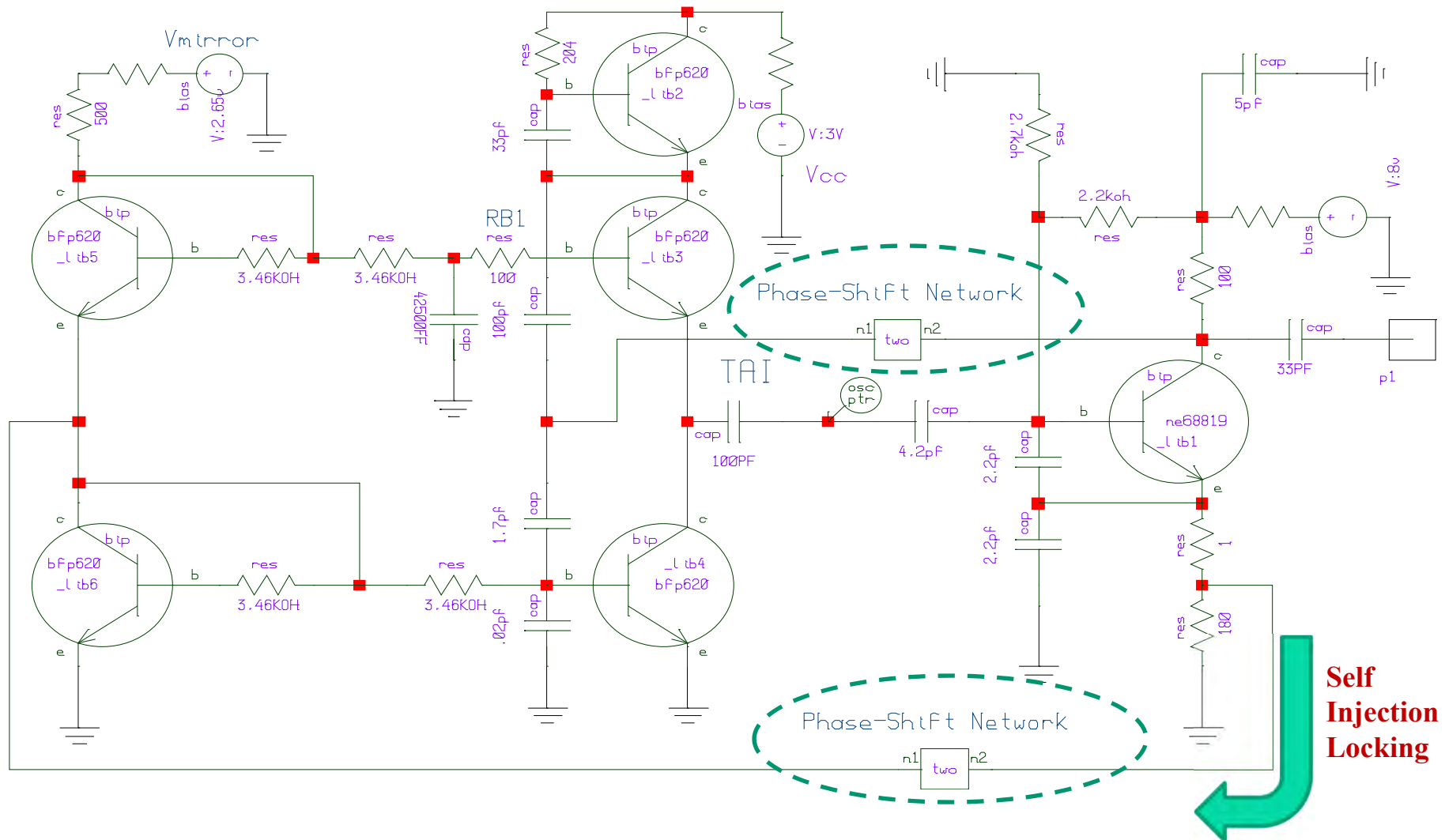
This Figure shows the schematic of self-injection-locked inductor-tuned Colpitts oscillator realized by incorporating phase shifter network in the feedback path, which improves the 1/f noise, including linearization of the large signal drive-level characteristics of the synthesized inductor circuits.

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CAD simulated phase noise plot
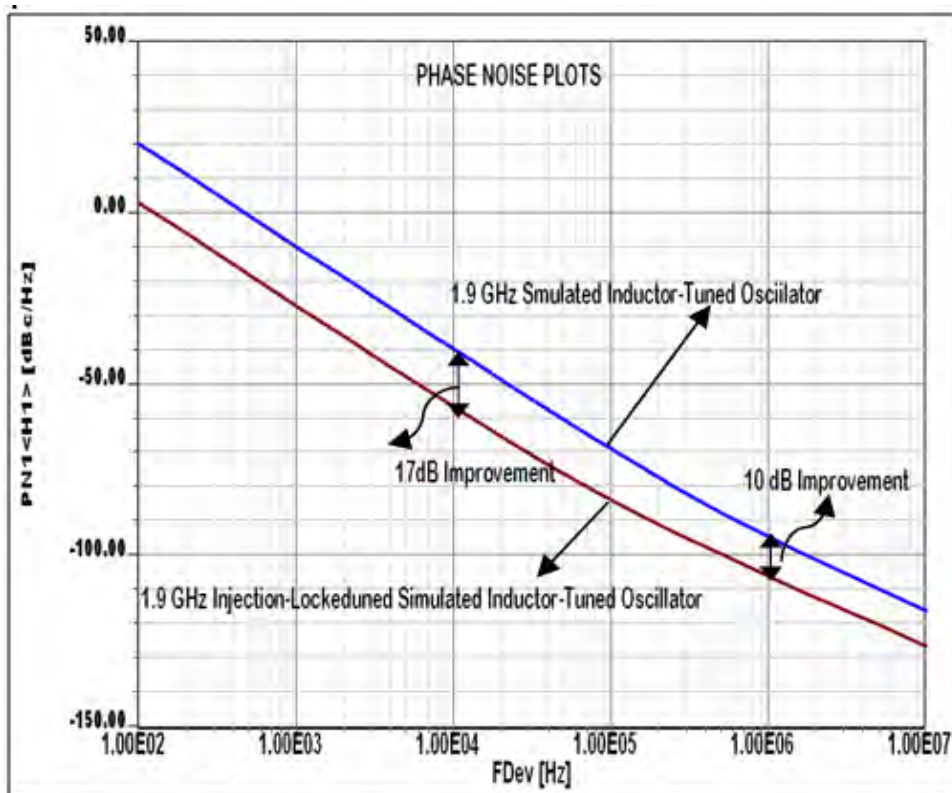
Measured phase noise plot (Injection-Locked)
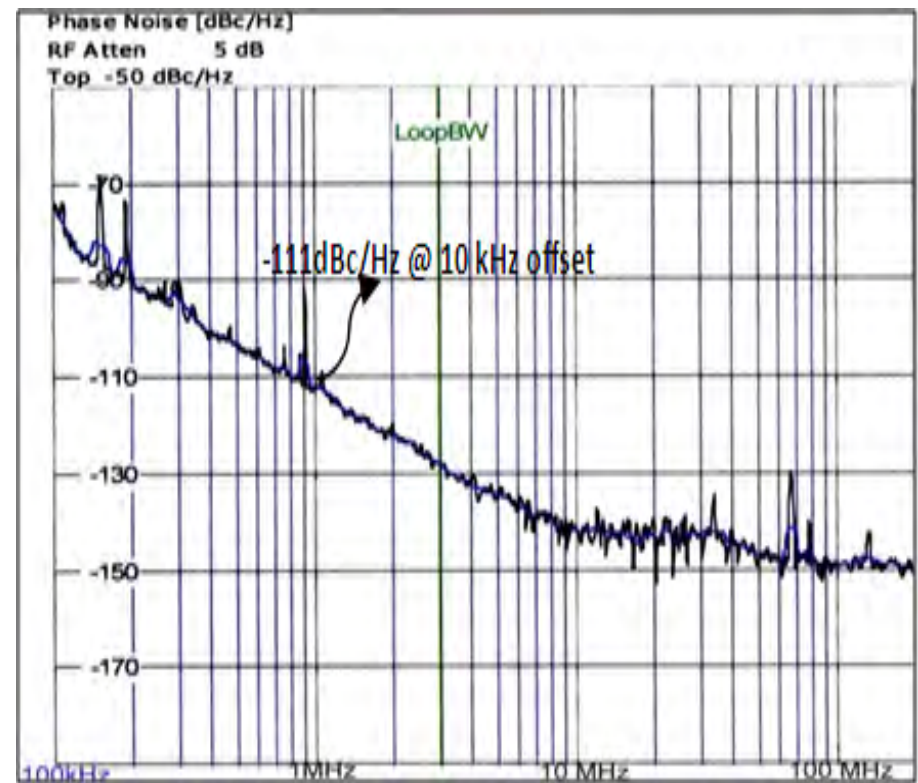
Figures show the CAD simulated (WITH AND WITHOUT INJECTION LOCKING) and measured phase noise plot of injection locked 1.9 GHz TAI. As shown in Figure, injection-locking improves the phase noise performance by 8-10 dB, including the locking range.

TUNABLE ACTIVE INDUCTOR - RECAP

- ✓ **This research work demonstrates the state-of-the-art in designing tunable inductor based VCO (voltage controlled oscillator) circuits**

- ✓ **Use of TAI (Tunable Active Inductor) resonator is relatively new and its application to replace tuning diodes in VCO (voltage controlled oscillator) have recently begun to be explored**

- ✓ **Closed form noise models for TAI VCOs involved complex mathematical treatment due to the convergence problems at large drive-level**

- ✓ **Limitation in the dynamic range may restricts the applications in high performance tunable filters, nevertheless by incorporating my novel techniques one can improve the dynamic range up to an accepted limit**

TAI - RECAP

✓ **The behavior of the TAI (Tunable Active Inductor) oscillator was studied and verified with practical examples.**

✓ **Intensive studies were conducted to find the optimum configuration for the improvement in the phase noise over the tuning range, and a US Patent application was filed.**

✓ **The extension of the research work is to increase the tuning range and dynamic range by employing injection mode coupling and noise cancellation techniques in monolithic IC technology.**

✓ **I expect to see continued research in this field and the use of TAI (Tunable Active Inductor) components as a cost-effective alternative of tuning diodes (Varactor) as a tuning element in filter, resonator, antenna matching network and phase shifter for the applications in concurrent and configurable RF & MW modules/systems.**
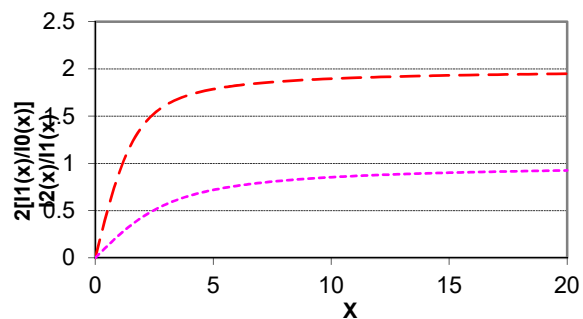
OUTLINE

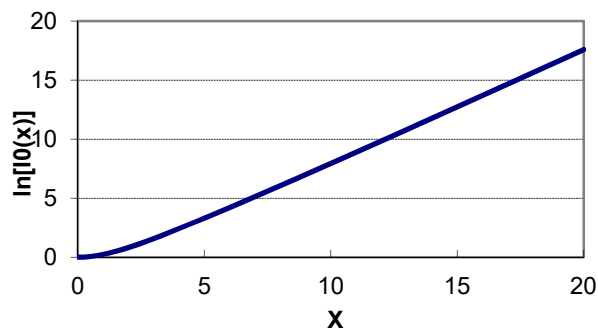# ➢ THE MODERN TIME-DOMAIN BEHAVIOR OF AN OSCILLATOR

Equivalent oscillator circuit for the analysis of the transformed conductance seen by the current source
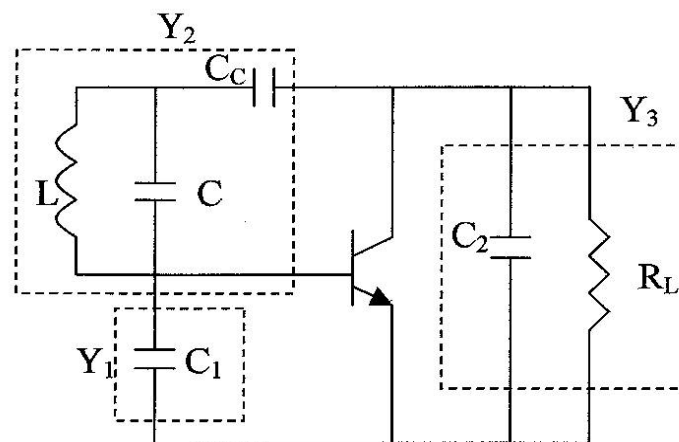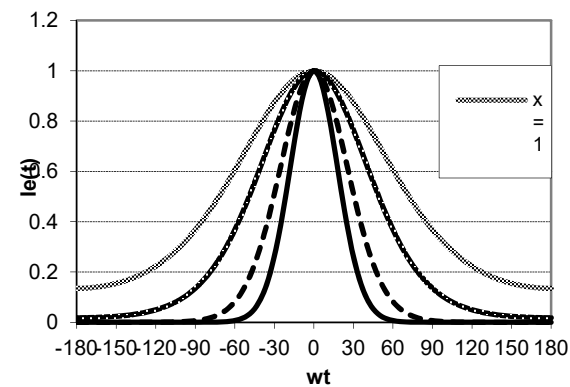

Oscillator circuit with the passive components Y_1, Y_2, and Y_3

Plot of the normalized fundamental current $2I_1(x)/I_0(x)$ and second harmonic $I_2(x)/I_1(x)$ with respect to the drive level x
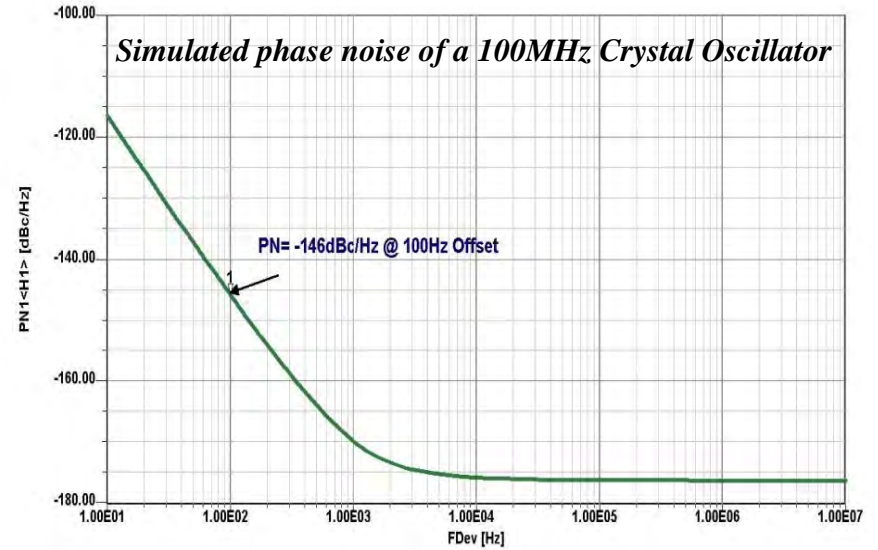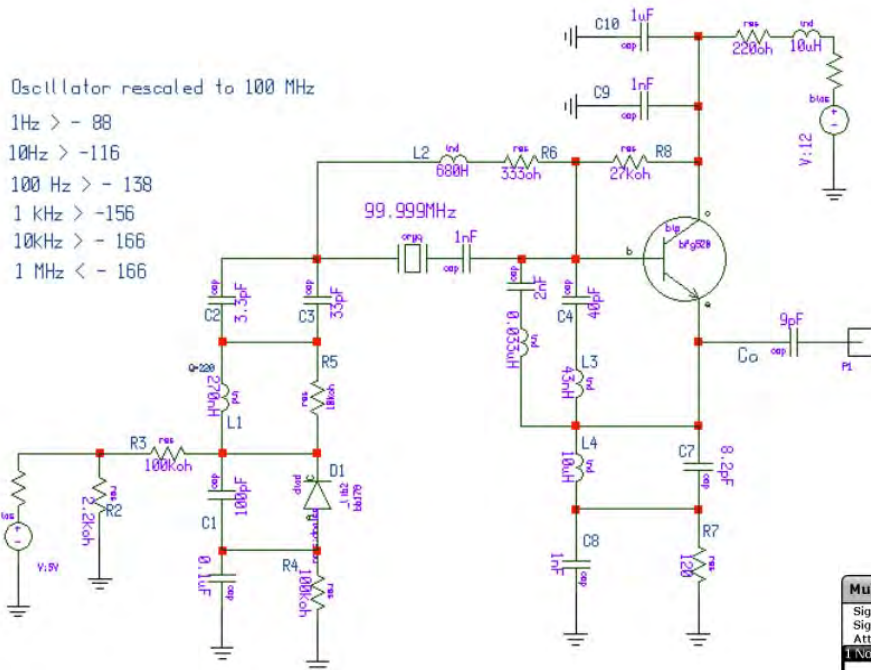

Plot of [ln I0 (x)] Vs drive level X

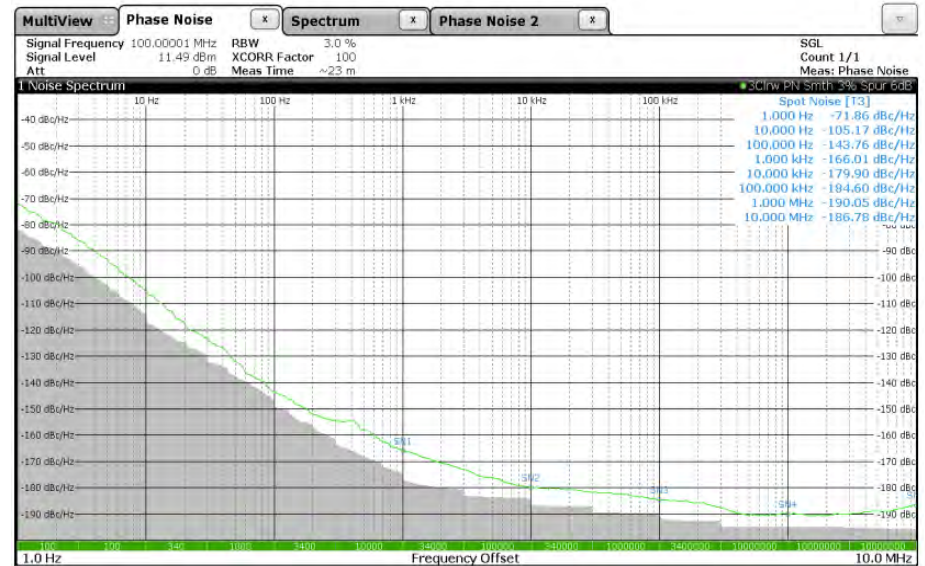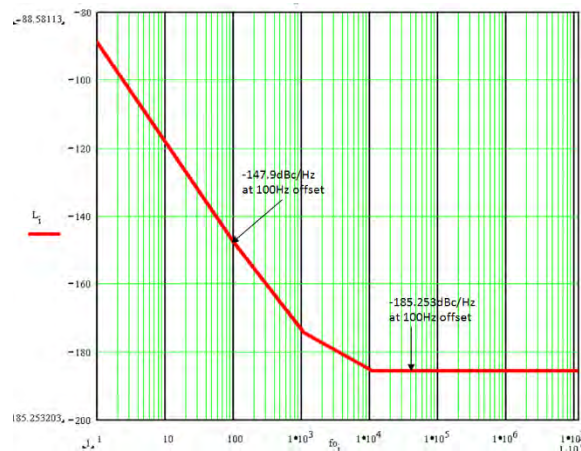*Plot of current with respect to conduction angle- (wt) and drive level X*

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- Design Examples of Oscillators for Best Phase Noise and Good Output Power

- Phase Noise of the integrated CMOS oscillator

- Appendix: A Solution for Calculating Phase Noise in Oscillators

   o Oscillator Noise Models

   o Noise Generation in Oscillators

- Key References

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Crystal Oscillator simulated schematic scaled to 100 MHz based on the 10MHz HP10811A Circuit
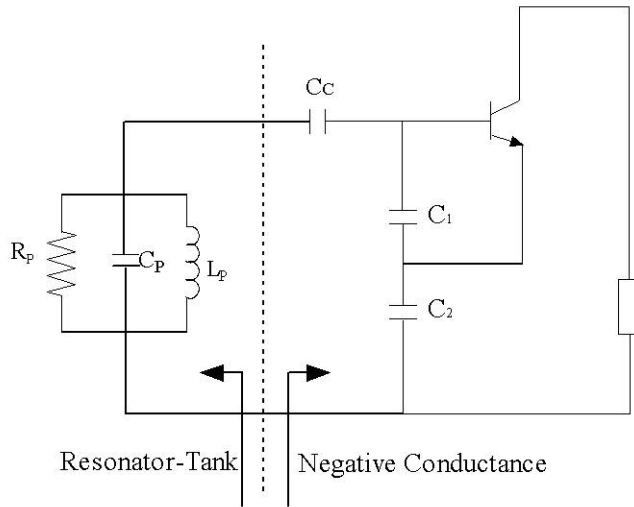
Simulated phase noise of a 100MHz Crystal Oscillator

PN= -146dBc/Hz @ 100Hz Offset

Calculated phase noise for a 100MHz Crystal Oscillator

-147.9dBc/Hz at 100Hz offset

-185.253dBc/Hz at 100Hz offset

Measured phase noise of a 100MHz Crystal Oscillator

OUTLINE

Colpitts Oscillator configuration for the intrinsic case, no parasitics assumed, and an ideal transistor considered

Equivalent representation of the oscillator circuit in presence of noise

Plot of negative resistance
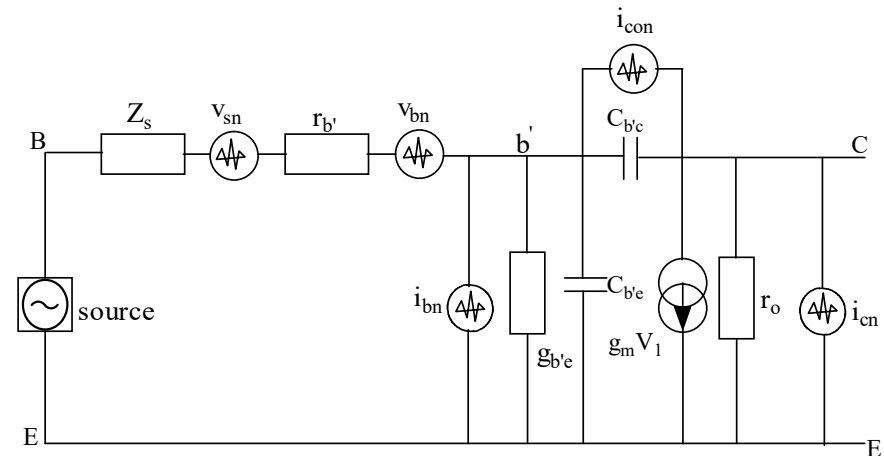
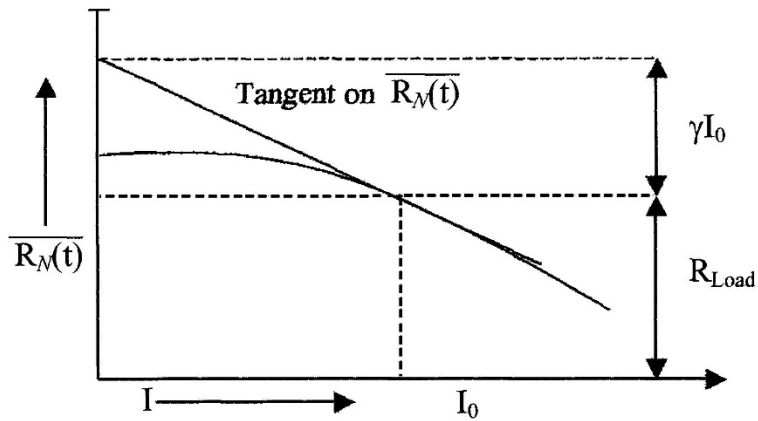*Hybrid-π configuration of the grounded bipolar transistor with noise sources*

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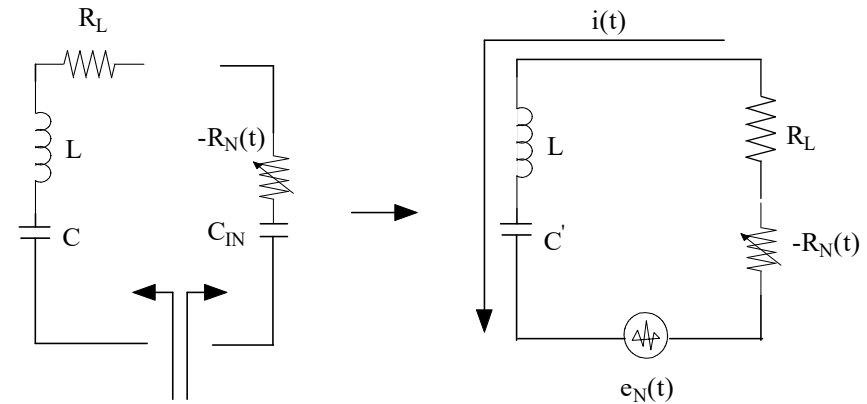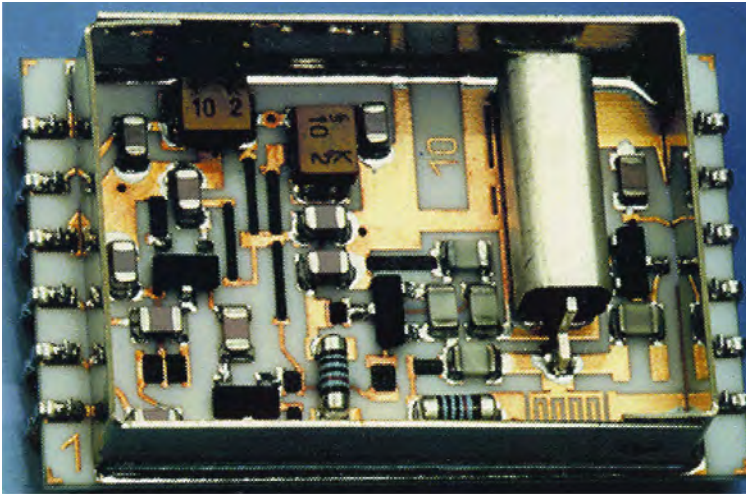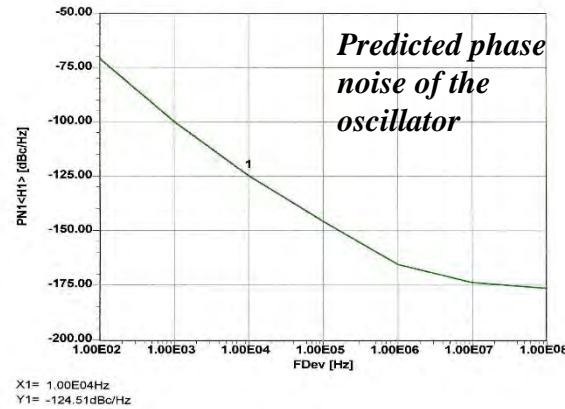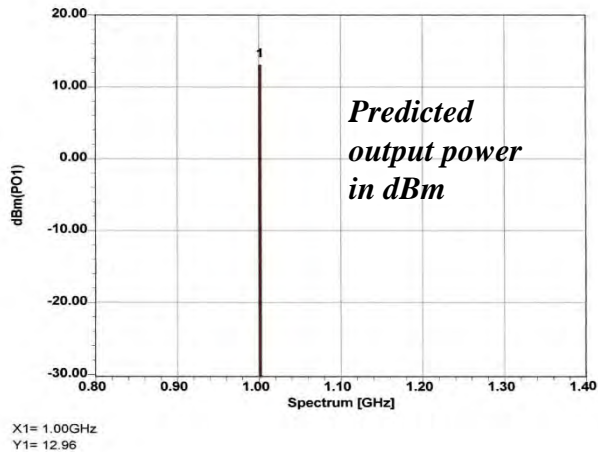 - Noise Generation in Oscillators

- Key References

➢ DESIGN EXAMPLES OF OSCILLATORS FOR BEST PHASE NOISE AND GOOD OUTPUT POWER

Parallel-tuned Colpitts oscillator using a ceramic resonator circuit, built and measured

1000MHz_Parallel-Tuned_Resonator_Oscillator

Predicted output power in dBm

X1= 1.00GHz
Y1= 12.96

Predicted phase noise of the oscillator

X1= 1.00E04Hz
Y1= -124.51dBc/Hz
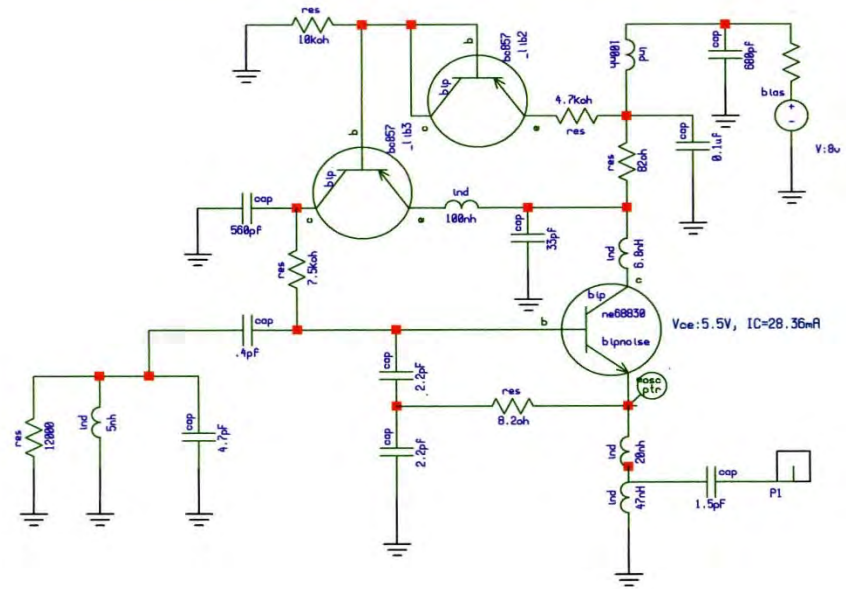
Current distribution in the oscillator circuit
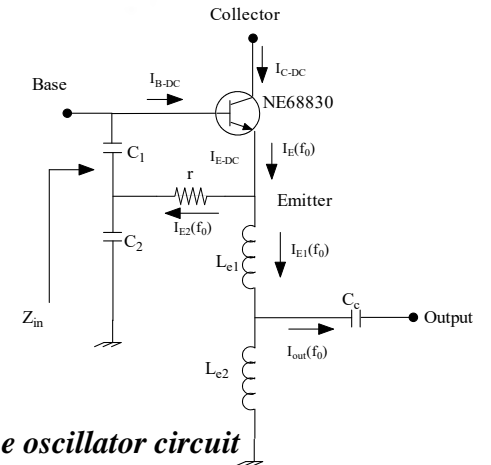
X1= 1.00E05Hz   X2= 1.00E06Hz   X3= 1.00E07Hz
Y1= -98.04dBc/Hz   Y2= -122.22dBc/Hz   Y3= -142.68dBc/Hz

Circuit schematic the 2 GHz GaAs FET oscillator built on GaAs substrate with very fine resolution of the components

Predicted phase noise of the oscillator.   The measured values were 100 dBc/Hz at 100 kHz and 120 dBc/Hz at 1 MHz. There is a deviation of about 2 dB compared to simulation

DC-IV and the load line for the GaAs FET oscillator
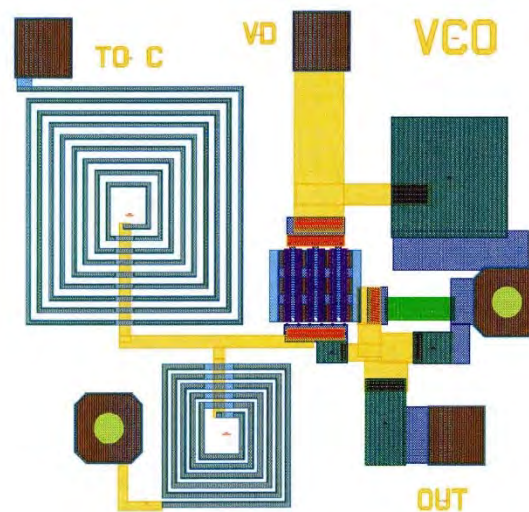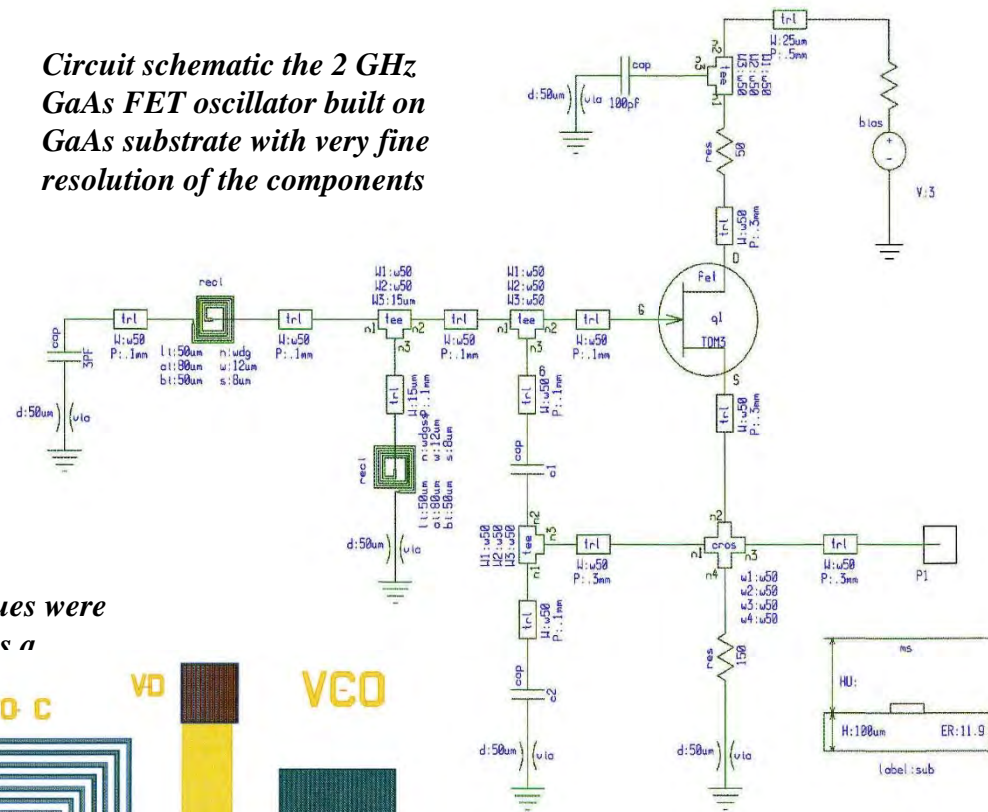
Layout of the 2 GHz GaAs FET oscillator
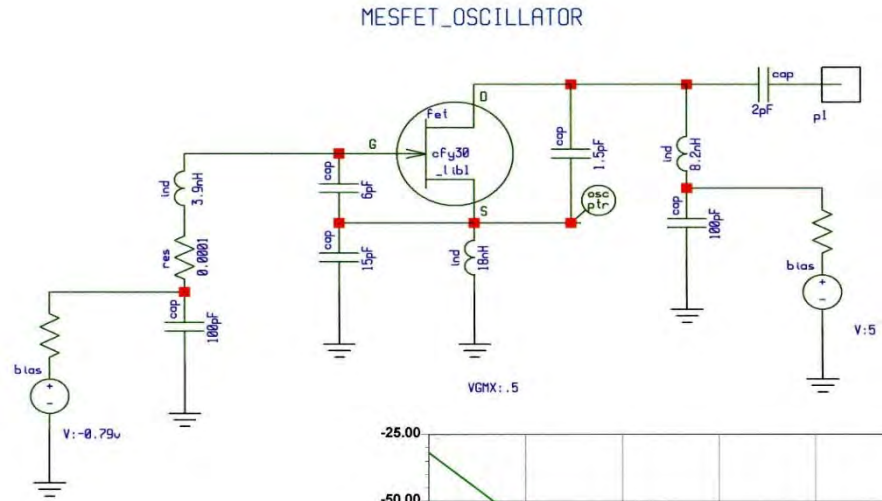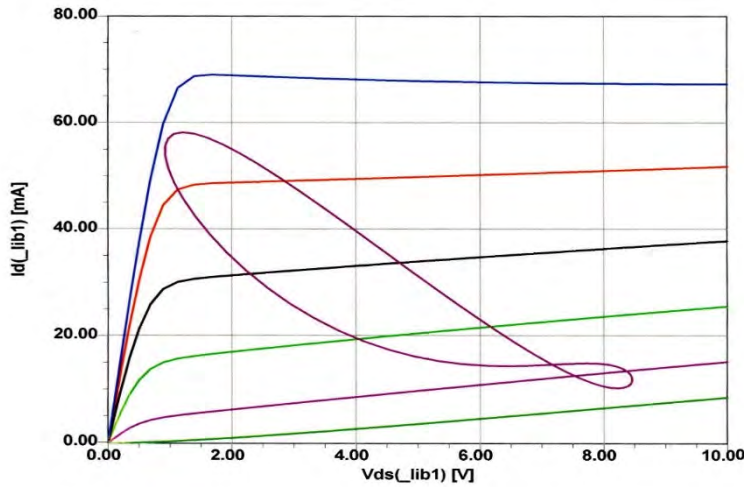
Load line of the MESFET oscillator - Because the load is a tuned circuit, the "load line" is a curve and not a straight line

Simulated output power of the oscillator in dBm
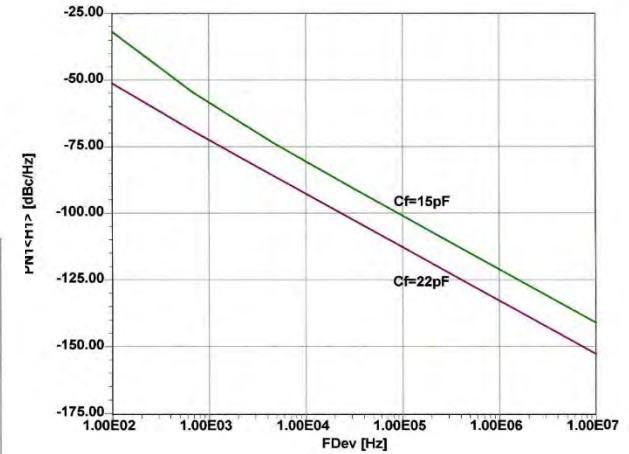
Simulated noise figure - An increase of the feedback capacitor from 15 to 22pF improves the phase noise
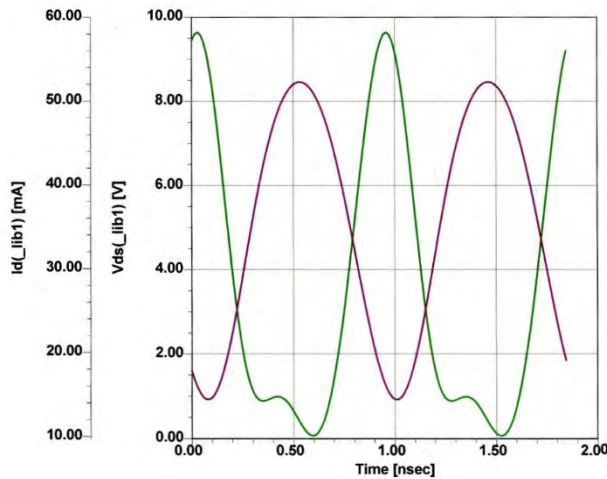
Plot of drain current and drain source voltage as a function of time

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2400 MHz MOSFET-Based Push-Pull Oscillator
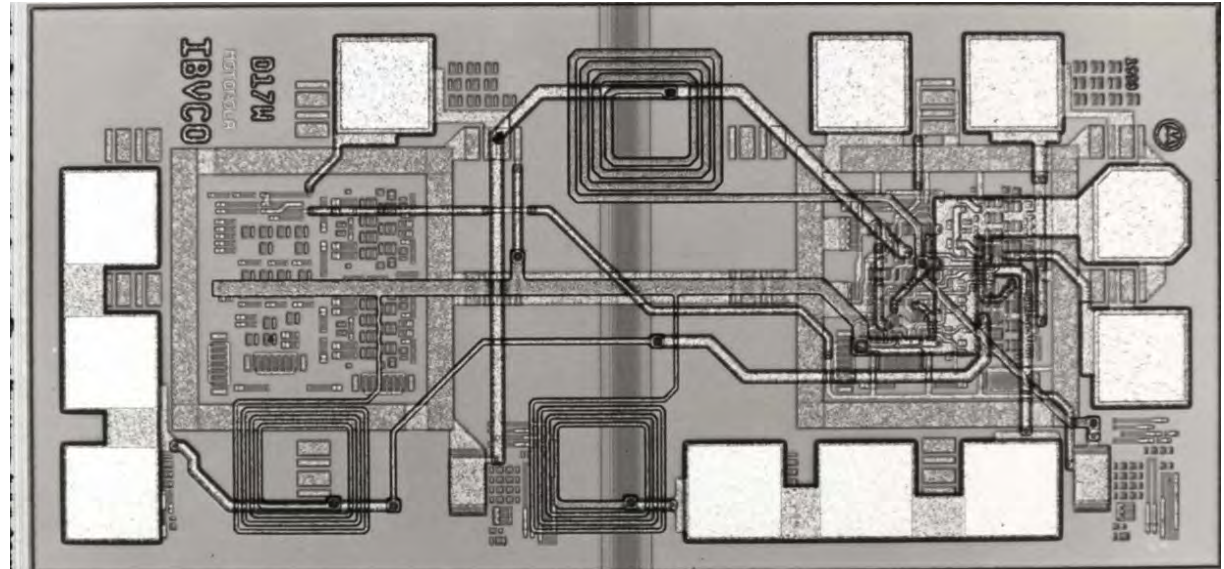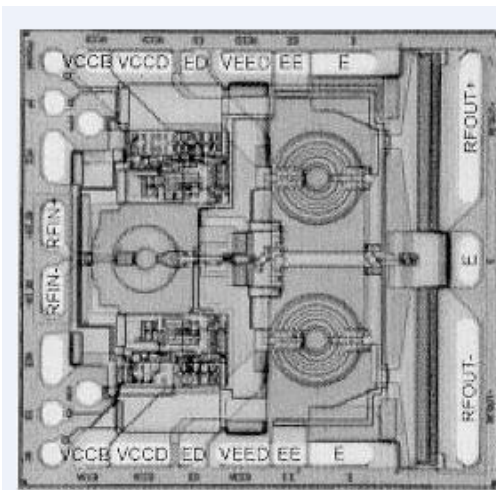
A Symmetric LC Oscillator

Adjust ratios to fine tune symmetry
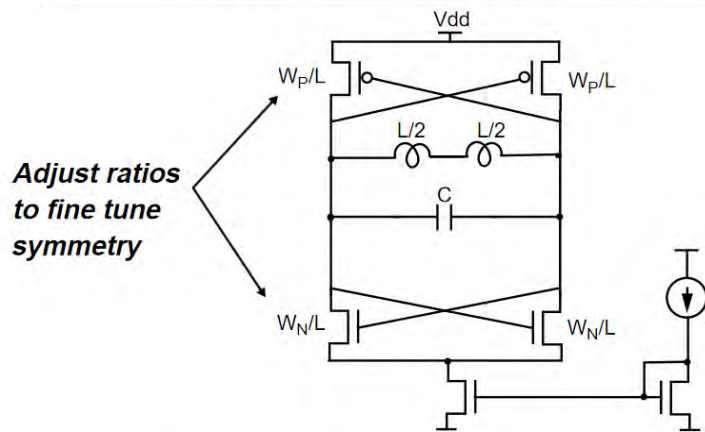
Uses the same current twice for high transconductance.

[Also appears in: J.Craninckx, *et al*, Proceedings of CICC 97.]

Most mixers in RFIC's are built on the principle of differential amplifiers (Gilbert cell) and require a phase and out of phase signal (symmetrical drive). For these symmetrical requirements, this is best achieved using a push-pull technology with two outputs. The design choices are SiGe transistors or BiCMOS transistors, and the critical phase noise is determined by the Q of the inductor and other elements of the resonator and by the flicker noise from the device.

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- Key References

➢ PHASE NOISE OF THE INTEGRATED CMOS OSCILLATOR

Symmetrical NMOS/PMOS Osc
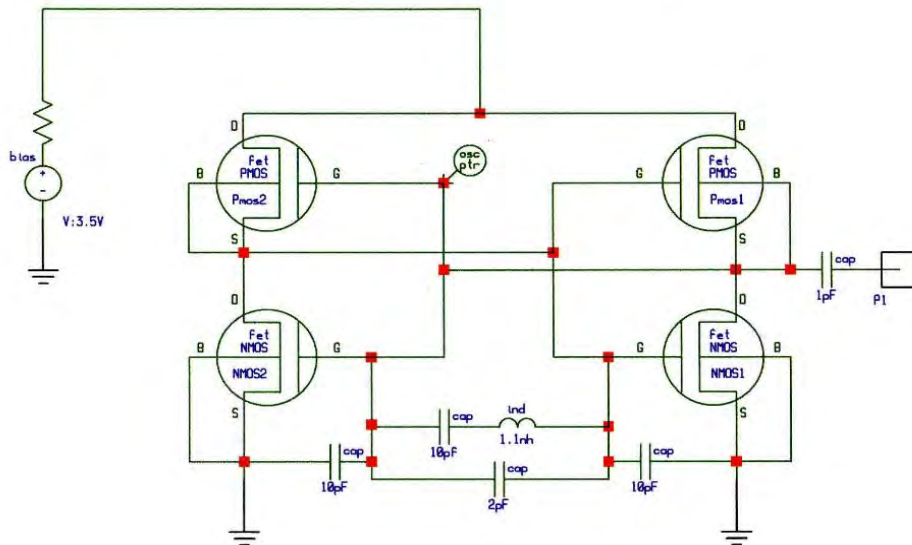
Circuit of the 2400 MHz integrated CMOS oscillator
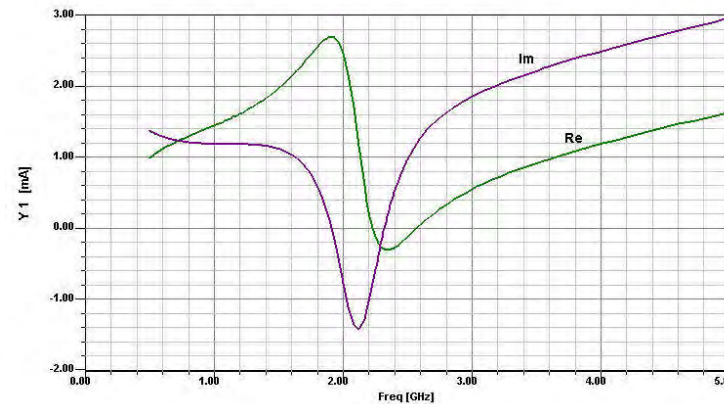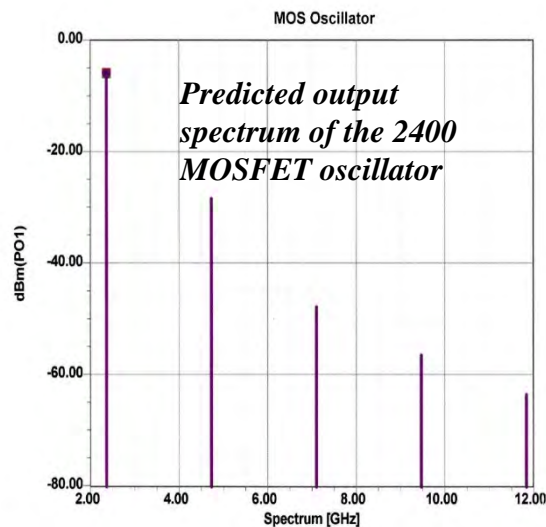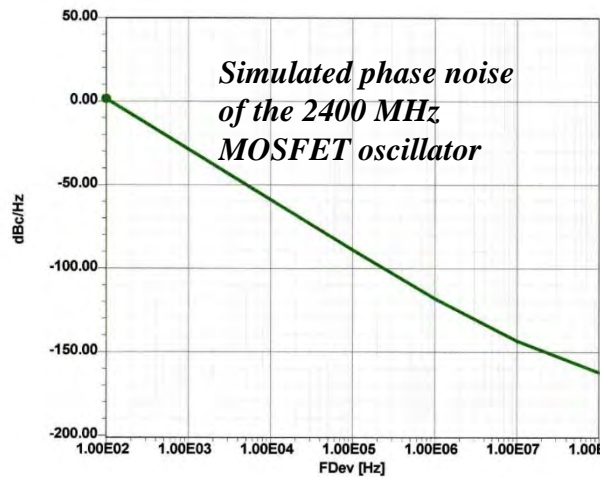
The real and imaginary currents which causes the negative resistance for oscillation

Simulated phase noise of the 2400 MHz MOSFET oscillator

MOS Oscillator

Predicted output spectrum of the 2400 MOSFET oscillator

Determining the transconductance of the differential circuit of the cross-coupled PMOS and NMOS pair

➤ PHASE NOISE OF THE INTEGRATED CMOS OSCILLATOR

The phase noise of CMOS oscillators has been subject to endless discussions. The main contributors still are the resonant circuit with a low Q and the flicker frequency contribution from the device.

$$PN_{imf}(\omega_0 + \omega) = \frac{4KT}{R_P}[NFT_{brr}(\omega_0)]^2 = \frac{4KT}{R_P}\left\{\frac{1}{2}\left[\frac{1}{2\omega_0 C_{eft}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2$$
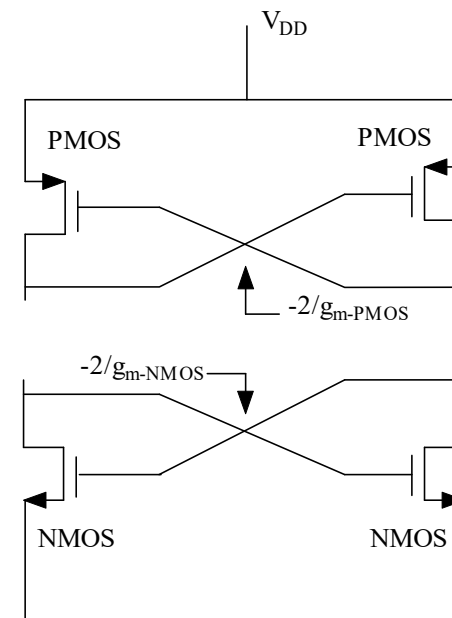
→ phase noise contribution from the resonator.

$$PN_{V_{gn}}(\omega_0 + \omega) = 4KT_g\left[NFT_{V_{gm}}(\omega_0)\right]^2$$

$$= 4KTr_g\left\{\frac{1}{2}\left[\frac{C_1 + C_2}{C_2}\right]\left[\frac{1}{2Q_0}\right]\frac{\omega_0}{\omega}\right\}^2$$

→ phase noise contribution from the gate resistance.

$$PN_{bgr}(\omega_0 + \omega) = 2qI_g\left[NFT_{kg}(\omega_0)\right]^2 = 2qI_R\left\{\frac{1}{2}\left[\frac{C_2}{C_1 + C_2}\right]\left[\frac{1}{\omega_0 C_{cft}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2$$

→ phase noise contribution from the gate current.

$$PN_{sm}(\omega_0 + \omega)$$

$$= \left(\frac{8kTgm}{3} + \frac{K_f I_z^{*F}}{f_m^F}\right)\left|NF_{y_k}(\omega_0)\right|^2 = \left(\frac{8kTgm}{3} + \frac{K_f I_\varepsilon^{NF}}{\omega}\right)\left\{\frac{1}{2}\left[\frac{C_2}{C_1 + C_2}\left[\frac{1}{2\omega_0 Q_0 C_{es}}\right]\left[\frac{\omega_0}{\omega}\right]\right]\right\}^2$$

→ phase noise contribution from the flicker noise of the transistor.

The predicted output spectrum of the CMOS oscillator
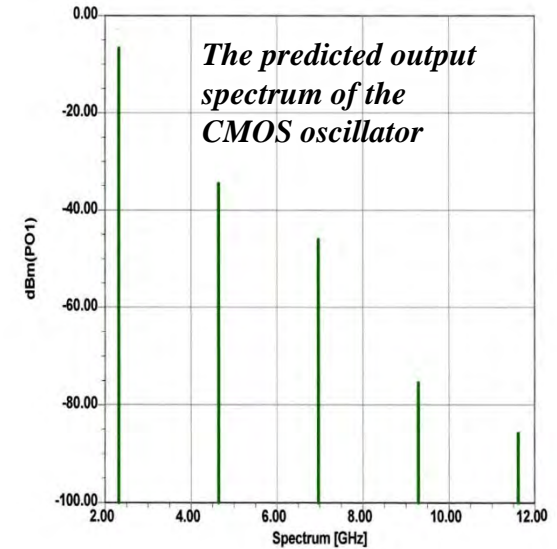
➢ PHASE NOISE OF THE INTEGRATED CMOS OSCILLATOR

The total effect of all the four noise sources can be expressed as

$$PN(\omega_0 + \omega) = [PN_{inr}(\omega_0 + \omega)] + [PN_{vgn}(\omega_0 + \omega)] + [PN_{tgn}(\omega_0 + \omega)] + [PN_{tjn}(\omega_0 + \omega) + [PN_{tdn}(\omega_0 + \omega)]$$

$$PN(\omega_0 + \omega) = \frac{4KT}{R_F}\left\{\frac{1}{2}\left[\frac{1}{2\omega_0 C_{ef}}\left[\frac{\omega_0}{\omega}\right]\right]\right\}^2 + 4KT_g\left\{\frac{1}{2}\left[\frac{C_1 + C_2}{C_2}\right]\left[\frac{1}{2Q}\right]\frac{\omega_0}{\omega}\right\}^2$$

$$+ \left[2qI_g + \frac{8kTgm}{3} + \frac{2\pi K_f I_g^{AF}}{\omega}\right]\left\{\frac{1}{2}\left[\frac{C_2}{C_1 + C_2}\right]\left[\frac{1}{2Q\omega_0 C_{eff}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2$$

$$+ 2qI_d\left\{\frac{1}{2}\left[\frac{C_3}{C_1 + C_2}\right]\left[\frac{1}{2\omega_0 Q C_{v\sigma}}\right]\left[\frac{\omega_0}{\omega}\right]\right\}^2$$

where

K_f = Flicker noise constant,

AF = Flicker noise exponent.

$$C_{eff} = C + \frac{C_1 C_2}{C_1 + C_2}$$

The following values were used for the calculation of the phase noise.

$R_p = 190\Omega$

$f_0 = 2.4$ GHz

$L = 1.1$nH

$C_O = 2$pF

$C_1 = C_2 = 0.2$pF

$n = 2$

$I_g = 100\mu A$

$I_d = 14$mA

$AF = 2$

$KF = 5E\text{-}5$

$q = 1.6E\text{-}19$

$T = 290°$ K

and the following contributions were obtained at 1 MHz offset

PN1 = -117.78 dBc/Hz

PN2 = -146.37 dBc/Hz

PN3 = -123.4 dBc/Hz

PN4 = -140.9 dBc/Hz

These calculations show that the phase noise contribution from the tuned circuit dominates and sets the value at –117.78 dBc/Hz.

The circuit was then analyzed using Microwave Harmonica/Ansoft Designer, using a lossy circuit with a Q_0 of 10 and using the SPICE-type parameters which were obtained from the manufacturer

OUTLINE

➢ OSCILLATOR NOISE MODELS &

➢ NOISE GENERATION IN OSCILLATORS

Rohde's Modified Leeson's Phase-Noise Equation

$$(f_m) = 10\log\left\{\left[1+\frac{f_0^2}{(2f_mQ_L)^2(1-\frac{Q_L}{Q_0})^2}\right]\left(1+\frac{f_c}{f_m}\right)\frac{FkT}{2P_{sav}}+\frac{2kTRK_0^2}{f_m^2}\right\}$$

Leeson has introduced a linear approach for calculation of oscillator phase noise and his noise formulae was extended by Rohde by adding the term $\dfrac{2kTRK_0^2}{f_m^2}$

responsible for the modulation noise

L (f_m) = SSB noise power spectral density defined as ratio of sideband power in 1-Hz bandwidth at f_m to total power in dB, unit is dBc/Hz.

f_m = frequency offset

f_0 = center frequency

f_c = flicker frequency- region between $1/f^3$-$1/f^2$

Q_L = loaded Q of the tuned circuit

Q_0 = unloaded Q of the tuned circuit

F = noise figure of the oscillator

kT = 4.1 × 10⁻²¹ at 300 K$_0$ (room temperature)

P_{sav} = average power at oscillator output

R=equivalent noise resistance of tuning diode

K_0= oscillator voltage gain

Shortcomings of the Modified Leeson Noise Equation

F is empirical, apriori, and difficult to calculate due to Linear Time Variant (LTV) characteristics of the noise.

Phase noise in $1/f^3$ region is an empirical expression with fitting parameters

Lee and Hajimiri Noise Model

This very nice model, unfortunately gives no insight in the contribution of the semiconductor used and only the flicker noise is discussed and the other noise sources for either the bipolar transistor or the variation of the FET are not considered

The Nonlinear Approach for the Computation of the Noise Analysis of Oscillator Circuits

Traditional approaches relying on frequency conversion analysis are not sufficient to describe the complex physical behavior of a noisy oscillator. The accuracy of this nonlinear approach is based on the dynamic range of the harmonic-balance simulator and the quality of the parameter extraction for the active device

Ulrich L. Rohde
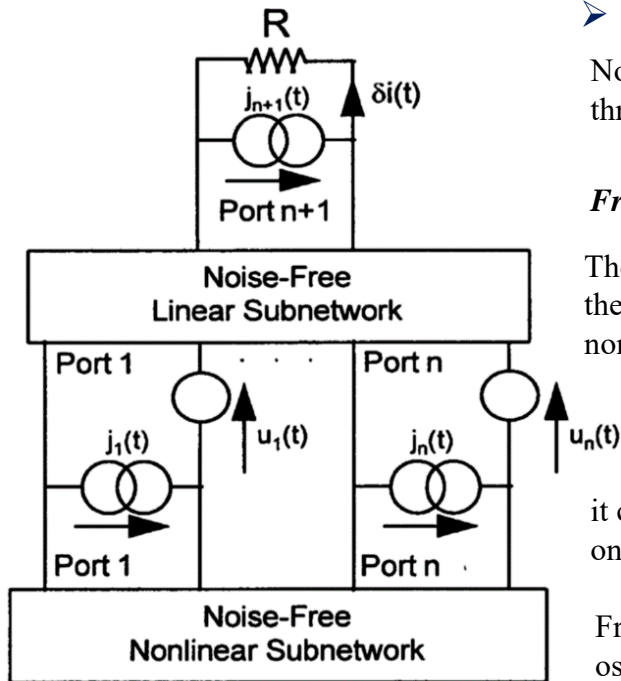
A general noisy nonlinear network

Oscillator noise components

➢ **Noise Generation in Oscillators**

Noise components at high frequency deviations result in phase modulation of the carrier through mean-square phase fluctuation proportional to the available noise power

Frequency Conversion Approach

The noise performance of the circuit is determined by the exchange of the power between the sidebands of the unperturbed steady state through frequency conversion in the nonlinear subnetwork. From the expression

$$M_{BB}\partial X_B = J_B(\omega)$$

it can be seen that the oscillator noise is essentially an additive noise that is superimposed on each harmonic of a lower and upper sideband at the same frequency offset

Frequency conversion analysis correctly predicts the far-carrier noise behavior of an oscillator, but the oscillator noise floor does not provide results consistent with the physical observations at low frequency deviations from the carrier, which can be removed by adding the modulation noise analysis

Modulation Noise Analysis

PM noise due to noise modulation, AM noise due to noise modulation, and PM-AM correlation coefficient due to noise modulation can be expressed in terms of a simple algebraic combination of equations.

Ulrich L. Rohde

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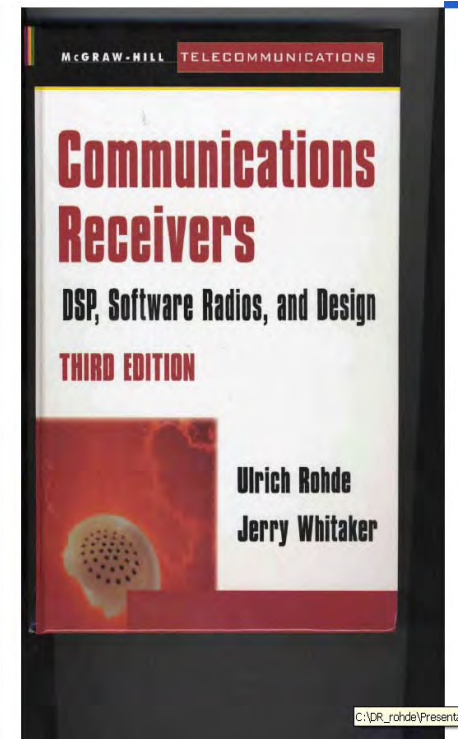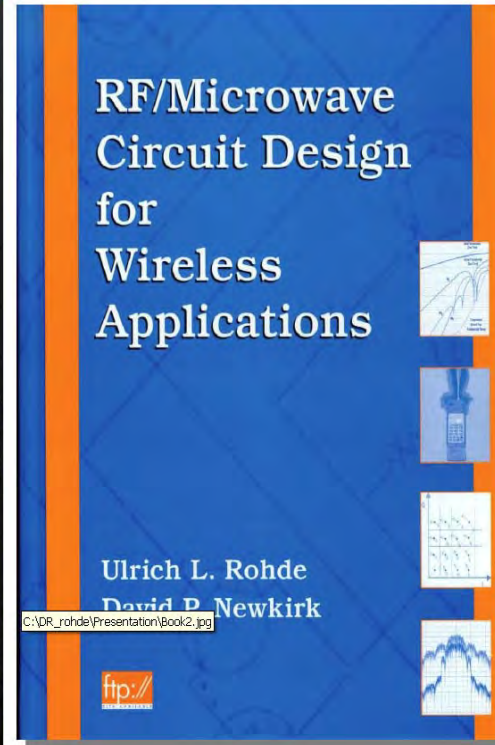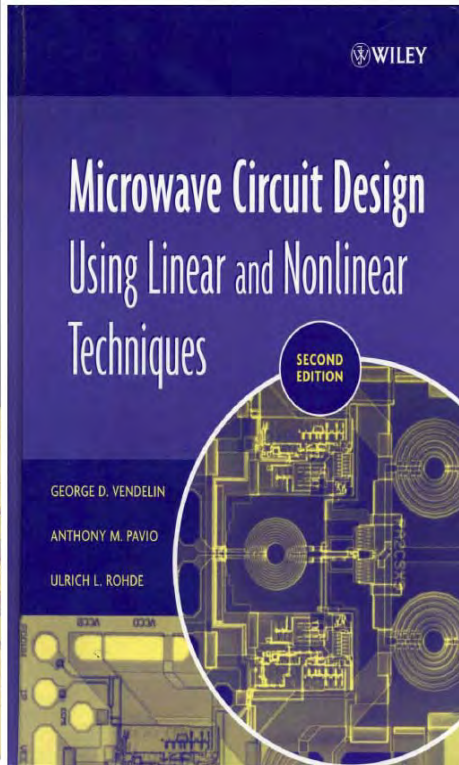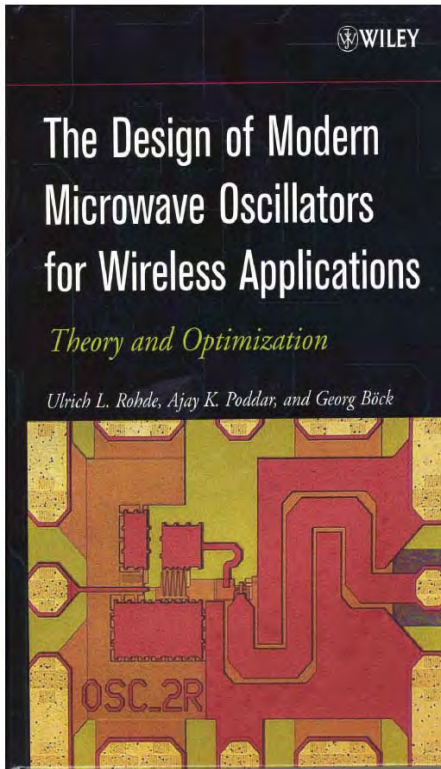
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