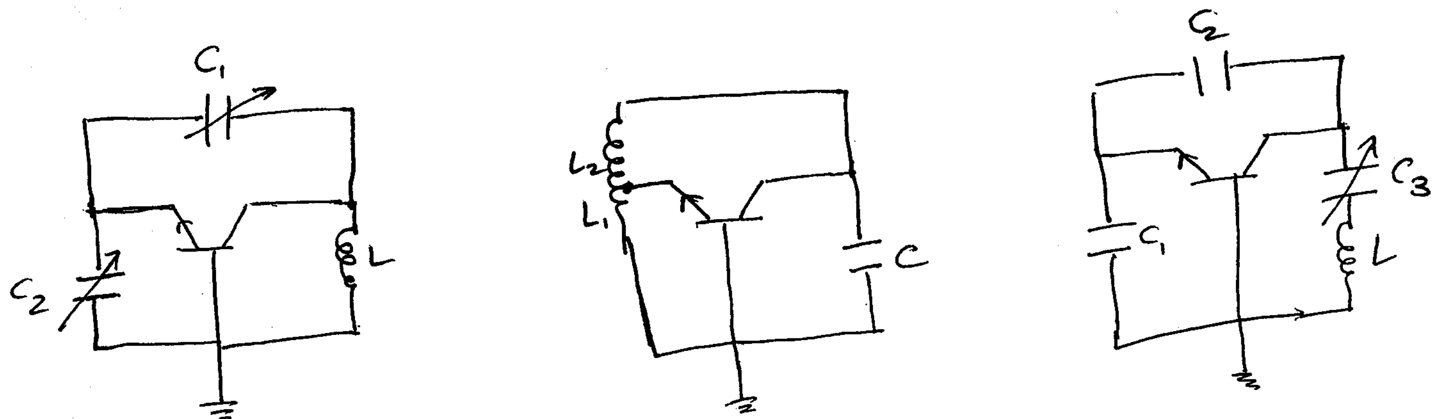
Oscillator Configurations

- At low Microwave frequencies lumped element oscillators are commonly used

- Three basic oscillator configurations used are:- Colpitts, Hartley, and Clapp oscillators

- Figure shows the Colpitts, Hartley, and Clapp oscillators



Colpitt Hartley Clapp

- Colpitt N/w uses a capacitor voltage divider in the tuned circuit to provide correct feedback

- Hartley N/w uses a tapped inductor tuned circuit

- Clapp N/w is similar to Colpitt N/w but with an extra capacitor in series with inductor to improve the frequency stability.

- High-Q tapped inductor required in the Hartley's oscillator is difficult to build. Therefore, the Colpitts and Clapp oscillators are preferred.

Design is presented using a feedback approach

$$A_{v_o} = \frac{1}{\beta_r(\omega)} \Rightarrow A_{v_o}\,\beta_r(\omega) = 1$$

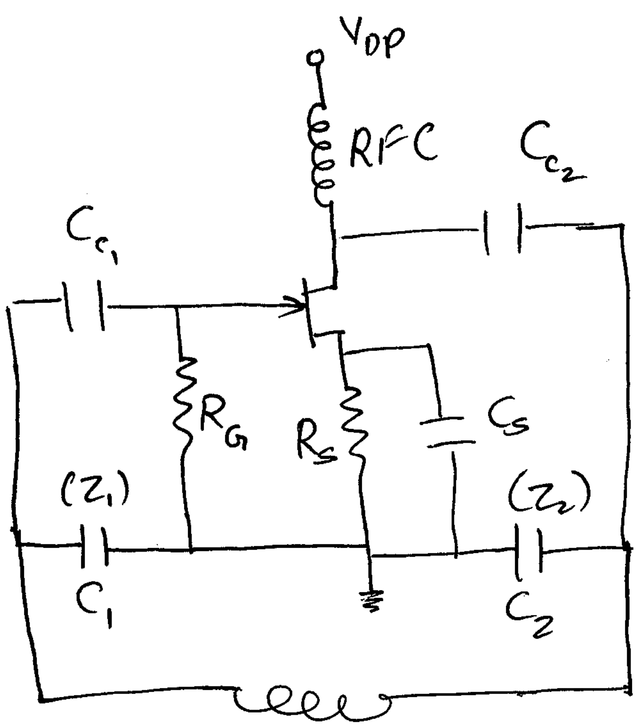
&

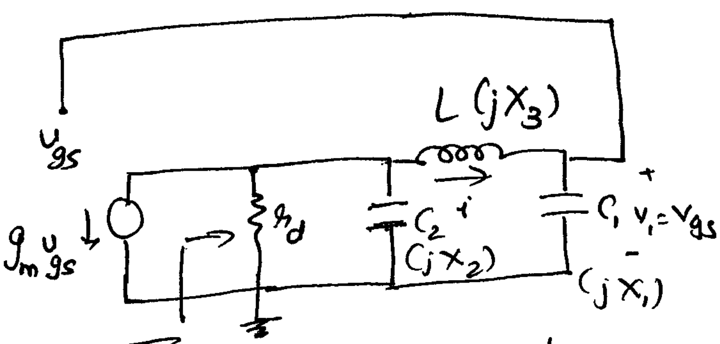
$$\beta_i(\omega) A_{v_o} = 0 \Rightarrow \beta_i(\omega) = 0$$

The oscillators presented here can also be analyzed using a negative-~~feedback~~ resistance approach.

For eg:-

Consider the FET Colpitts oscillator shown in figure below.

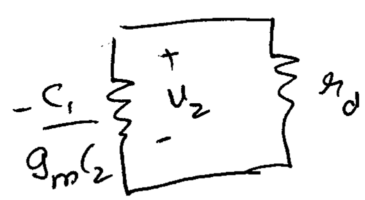




$$X_1 = -\frac{1}{\omega C_1}$$

$$X_2 = -\frac{1}{\omega C_2}$$

$$X_3 = \omega L$$



Using $\beta_1(\omega) A_{v_0} = 0$

$$X_1 + X_2 + X_3 = -\frac{1}{\omega C_1} - \frac{1}{\omega C_2} + \omega L = 0$$

$$\omega = \omega_0 = \frac{1}{\sqrt{L C_T}}$$

where

$$C_T = \frac{C_1 C_2}{C_1 + C_2}$$

$$\beta(\omega) = \frac{V_1}{V_2} = \frac{v_{gs}}{V_2} = \frac{X_1}{X_1 + X_3}$$

$$\beta(\omega) = \frac{v_{gs}}{v_2} = -\frac{X_1}{X_2} = -\frac{C_2}{C_1}$$

$$g_m v_{gs} = -g_m \frac{C_2}{C_1} v_2 \qquad \text{— (1)}$$

(1) shows that the source $g_m v_{gs}$ can be replaced by negative resistance given by $-\frac{C_1}{g_m C_2}$ and therefore at resonance the model in figure 'c' follows.

For oscillation to occur, the loop resistance must be zero

$$-\frac{C_1}{g_m C_2} + r_d = 0$$

$$g_m r_d = \frac{C_1}{C_2}$$

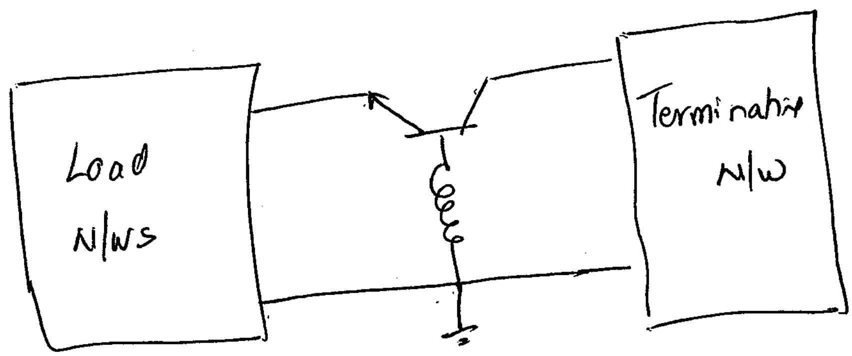
which is well known gain condition for the Colpitts oscillator

- At higher microwave frequencies the parasitic capacitances of the packaged transistors provide some or all the feedback needed for oscillation.
- In this range the negative-resistance design procedure is used, since the S parameters provide all the needed design information
- The negative-resistance design procedure basically consists of selecting a transistor in an oscillator topology that provides the required output power.
- The transistor in the configuration selected must be potentially unstable at the desired frequency of oscillation.
- Feedback can be added to increase the negative resistance associated with Γ_{IN} or Γ_{OUT}. The terminating & load matching N/ws must be designed to provide the proper resonance conditions.

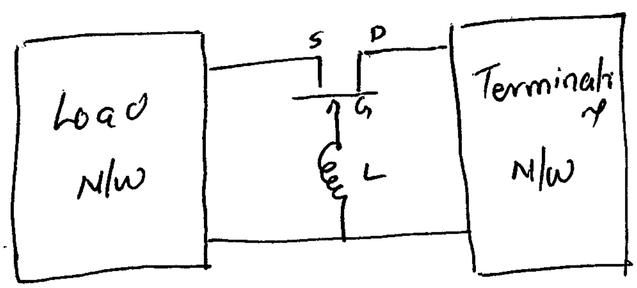
For BJT the most effective network topology is the common-base configuration.

This configuration is used in low-power oscillator circuits, and it is easy to tune. The inductor feedback element is used to increase $|\Gamma_{IN}|$

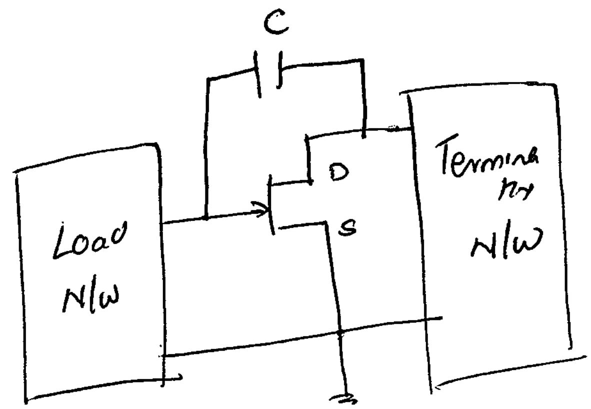
and $|\Gamma_{OUT}|$.



The two common N/w configurations for GaAs FETs are shown in figure below



Common-gate Configuration



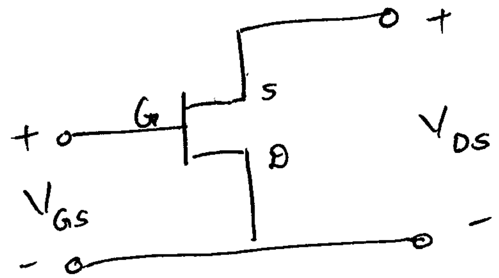
Common-Source Configuration

The Common-gate configuration is used in low-power oscillator ckts since it is easy to tune.

- A series inductive feedback is usually reqd to improve $|\Gamma_{IN}|$ and $|\Gamma_{OUT}|$. The Common source configuration is used for higher oscillator output power and the feedback N/w is usually a capacitor

- The Common-drain configuration is not popular because the oscillator implementation is difficult

A GaAs FET oscillator can also be built using the reverse channel configuration. A reverse-channel configuration uses a symmetrical GaAs FET with a negative voltage applied to the drain terminal.



- The transistor becomes a non inverting device, making the common lead inductance regenerative.
- The S parameters in the reverse-channel configuration show that $|S_{12}|$ rises markedly with frequency & $|S_{11}|$ is greater than unity in a large frequency range.

The load-tuning elements are not limited to lossless or RLC networks. They can be designed using dielectric resonators, YIG (yttrium iron garnet) resonators, varactor diodes etc.

Dielectric Resonator Oscillators.

These are of two types.
1) TEM mode 2) TE mode

TEM mode Dielectric Resonators

→ TEM-mode DR is similar to coaxial cable.

→ The main use of DR resonators is to implement high-Q inductors in oscillators to resonate with a capacitor or varactor.

→ DR resonators are available as $\frac{\lambda}{4}$ resonators with one end shorted, or as half-wave resonators with both ends open.

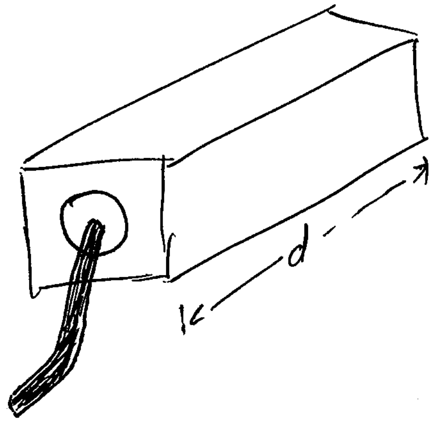
→ At resonance a short-circuited $\frac{\lambda}{4}$ transmission line appears as an open circuit, and at frequencies below resonance it behaves like an inductor.

→ These resonators are usually made with ceramic dielectric material and they are called TEM-mode ceramic resonators (TEM-mode CRs).

→ These particular resonators are found applications in many oscillators used in the wireless communication field.

→ They can be used in VCOs in the range of 300MHz to 5GHz.

→ Good performance is obtained by the use of a coaxial ~~reso~~ resonator, where the resonator is shaped in the form of a cube of length d with a coaxial bore and shorted at one end.



→ The inner and outer surfaces are plated with copper or with silver.

→ The conductivity of the platting affects the Q of the resonator.

→ Those platted with copper are cheaper than those plated with silver.

→ When high Q values are required, the silver-plated resonator is the best

→ The resonators are fabricated with a high dielectric constant in order to make the component small.

→ To be useful as an inductive component in microwave oscillators.

→ The high value of the relative dielectric constant ε_r is required for circuit miniaturization since the size of the component is inversely proportional to $\sqrt{\varepsilon_r}$

→ This is due to the fact that the wavelength in the dielectric material λ_d is given by

$$\lambda_d = \frac{\lambda_0}{\sqrt{\varepsilon_r}}$$

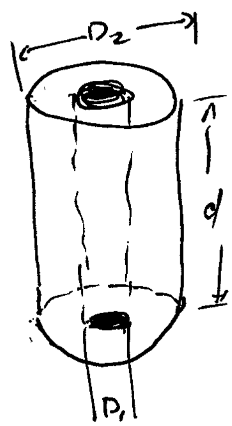
→ Typical values of ε_r are from 10 to 100 (eg $\varepsilon_r = 21, 38, 88$ and 92) with typical lengths of 3 to 14 mm

→ Typical values of the characteristic impedance are from 4Ω to 20Ω.

→ The Q of the DR describes the losses of the component, namely dielectric losses & conduction losses.

→ The temperature coefficient TC indicates the thermal stability of the DR. It describes how much the resonant frequency drifts as a function of temperature.

→ For a short-circuited coaxial-cable resonator as shown in the figure its capacitance per unit length is given by



$$C = \frac{2\pi \varepsilon_0 \varepsilon_r}{\ln\left(\frac{D_1}{D_2}\right)}$$

inductance per unit length is given by

$$L = \frac{\mu_0}{2\pi} \ln\left(\frac{D_1}{D_2}\right)$$

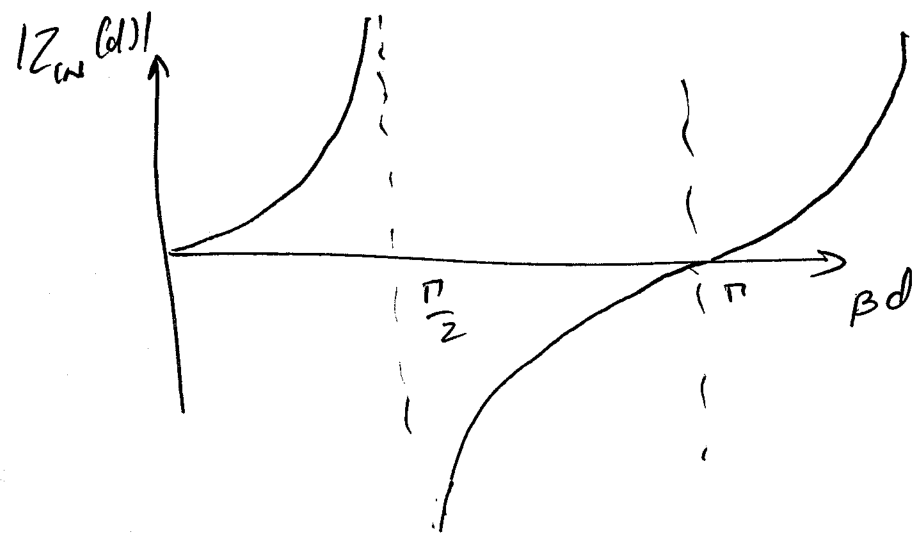
The characteristic impedance of the resonator is.

$$Z_0 = \sqrt{\frac{L}{C}} = \frac{60}{\sqrt{\varepsilon_r}} \ln\left(\frac{D_1}{D_2}\right)$$

→ a correction to the D_1/D_2 is required for the geometry of the resonator.

→ The DR is usually designed with a $\frac{\lambda}{4}$ length at a given frequency f_0. The impedance of a lossless short-circuited resonator is:

$$Z_{IN}(d) = j Z_0 \tan\beta d.$$



$0 < \beta d < \frac{\pi}{2}$ ('short circuited line is inductive')

$\frac{\pi}{2} < \beta d < \pi$ ('shorted line is capacitive')

- At a frequency lower than the resonant frequency of the resonator, the line is inductive.

Quarter-wave resonators behave like parallel-resonant circuits. To obtain an equivalent i/p parallel resonant ckt the i/p admittance of a $\frac{\lambda}{4}$ short-circuited lossy transmission line is considered

$$Y_{IN}(d) = Y_0 \coth(\alpha + j\beta) \qquad -(1)$$

$$= Y_0 \frac{1 + j \tan\beta d \tanh(\alpha d)}{\tanh(\alpha d) + j \tan(\beta d)}$$

For low losses (ie for $\alpha d \ll 1$)

$$\tanh \alpha d \approx \alpha d = \frac{\alpha\lambda}{4} \qquad -(2)$$

Since

$$\beta d = \frac{2\pi d}{\lambda} = \frac{\omega d}{v_p}$$

with $d = \frac{\lambda}{4}$ at $\omega = \omega_0$, and $\omega = \omega_0 + \Delta\omega$

$$\beta d = \frac{\omega_0 d}{v_p} + \frac{\Delta\omega d}{v_p} = \frac{\pi}{2} + \frac{\pi\Delta\omega}{2\omega_0}$$

$$\therefore \tan\beta d = \tan\left(\frac{\pi}{2} + \frac{\pi\Delta\omega}{2\omega_0}\right) = \frac{-1}{\tan\frac{\pi\Delta\omega}{2\omega_0}} \approx \frac{-2\omega_0}{\pi\Delta\omega} \qquad -(3)$$

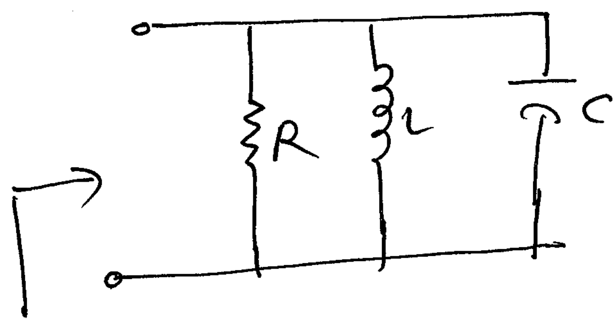
Substituting (2) & (3) in (1)

$$Y_{IN} = Y_0 \frac{1 - j\frac{2\omega_0}{\pi\Delta\omega}\frac{\alpha\lambda}{4}}{\frac{\alpha\lambda}{4} - j\frac{2\omega_0}{\pi\Delta\omega}} = Y_0 \frac{j\frac{\pi\Delta\omega}{2\omega_0} + \frac{\alpha\lambda}{4}}{j\frac{\alpha\lambda}{4}\frac{\pi\Delta\omega}{2\omega_0} + 1}$$

Since $\frac{\alpha\lambda}{4} \frac{\pi\Delta\omega}{2\omega_0} << 1$

$$Y_{IN} \approx Y_0\left(\frac{\alpha\lambda}{4} + j\frac{\pi\Delta\omega}{2\omega_0}\right) \qquad -(4)$$

From (4) we develop an equivalent ckt, we consider the i/p admittance of the parallel tuned ckt shown in figure below



Y_{IN}

$$Y_{IN} = G + \frac{1}{j\omega L} + j\omega C \qquad -(5)$$

$\omega = \omega_0 + \Delta\omega$ &

$$\frac{1}{\omega_0 + \Delta\omega_0} \approx \frac{1 - \frac{\Delta\omega_0}{\omega_0}}{\omega_0} \qquad -(6)$$

Substituting (6) in (5)

$$Y_{IN} = G + \frac{1 - \frac{\Delta\omega}{\omega_0}}{j\omega_0 L} + j\omega_0 C + j\Delta\omega C \qquad -(7)$$

Observing that

$$\frac{1}{j\omega_0 L} + j\omega_0 C = 0 \quad \text{or} \quad \omega_0 = \frac{1}{\sqrt{LC}}$$

(7) reduces to:

$$Y_{IN} = G + j\frac{\Delta\omega}{\omega_0^2 L} + j\Delta\omega C \qquad - (8)$$

$$\approx G + j2\Delta\omega C$$

Comparing (4) & (8) we get

$$R = \frac{4Z_0}{\alpha\lambda} \qquad - (9)$$

$$C = \frac{\pi}{4\omega_0 Z_0} \qquad - (10)$$

&

$$L = \frac{1}{\omega_0^2 C} \qquad - (12)$$

At resonance the input impedance is

$$Z_{IN}(\omega_0) = R$$

which, of course, is very high since α is small.

$$Q_u = \omega_0 RC = \frac{\pi}{\alpha\lambda} = \frac{\beta}{2\alpha}$$

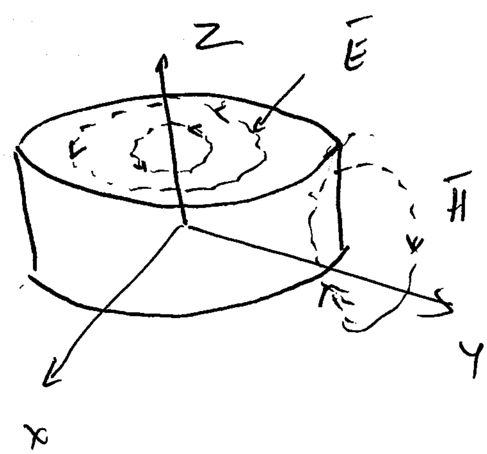
$$R = \frac{Q_y}{\omega_0 C} \quad -(13)$$

For a given resonator the palameters Z_0, ω_0 & it Q_u are known.

<u>TE mode DRs</u>

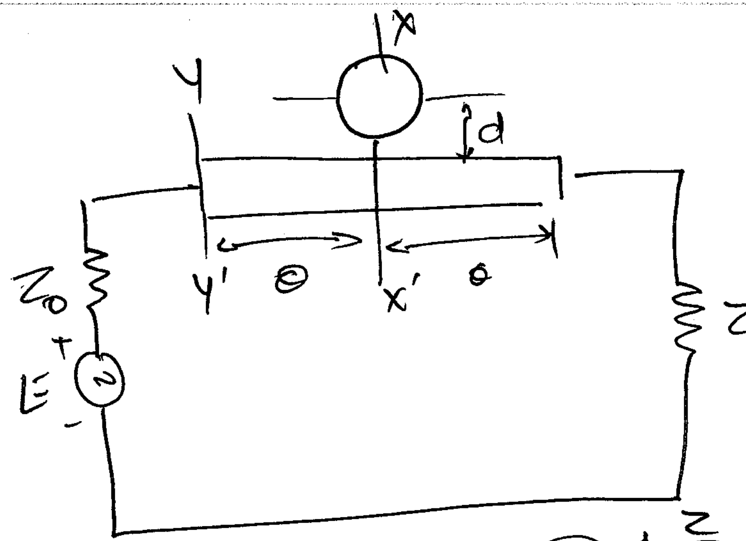
- A DR will resonate in several modes. The most commonly used mode in cylindrical resonators is a TE mode.

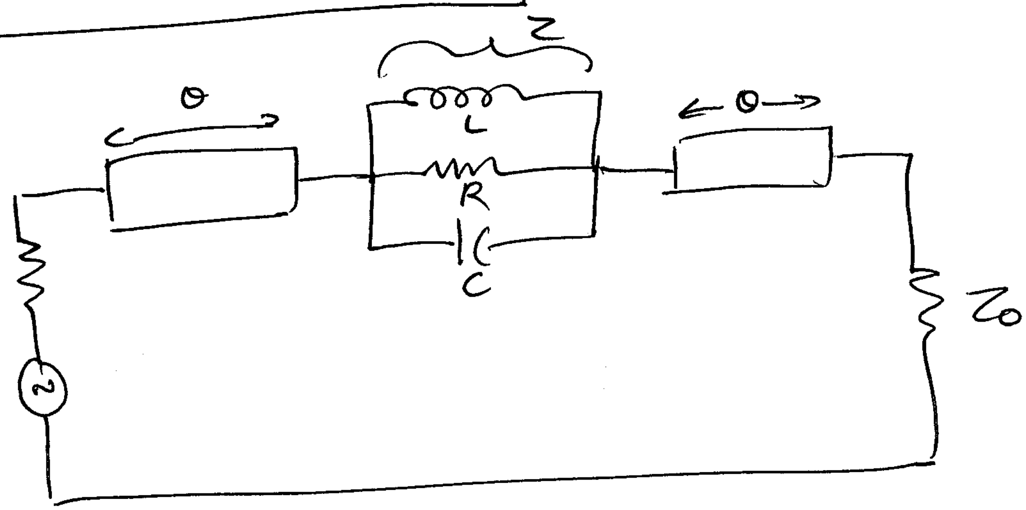
→ This mode can be easily coupled to a microstrip line. To understand the coupling of a dielectric resonator to a microstrip line, which shows the field distribution of the $TE_{01\delta}$ mode.



→ Figure shows the dielectric resonator coupled to a microstrip line

→ The dielectric resonator is placed on top of the substrate at a distance d from the microstrip line





- The ckt equivalent consists of a parallel tuned ckt placed in series, at position X X', with the transmission lines. The values of R, L, & C in the equivalent ckt depend on DR characteristics & distance d.

Z is given as :-

$$Z = \frac{1}{C} \frac{s}{s^2 + \frac{s}{RC} + \frac{1}{LC}} = \frac{1}{C} \frac{s}{s^2 + 2\alpha s + \omega_0^2} \quad -(1)$$

BW and resonant frequency are

$$BW = 2\alpha = \frac{1}{RC}$$

$$\omega_0 = \frac{1}{\sqrt{LC}}$$

Substituting $S = j\omega$ in (1) we get

$$Z = \frac{R}{1 + jQ_u \dfrac{(\omega^2 - \omega_0^2)}{\omega\omega_0}} \quad — (2)$$

$Q_u \rightarrow$ Unloaded Q of tuned ckt

$$Q_u = \frac{\omega_0}{2\alpha} = \omega_0 RC = \frac{R}{\omega_0 L} \quad — (3)$$

Since frequency of operation is very close to ω_0 (i.e. $\omega + \omega_0 = 2\omega$)

we can approximate (2) by

$$Z = \frac{R}{1 + j2Q_u\delta} \quad — (4)$$

where

$$\delta = \frac{\omega - \omega_0}{\omega_0}$$

At reference plane xx' the input impedance is

$$Z_{xx'} = Z + Z_0$$

$$z_{xx'} = \frac{Z_{xx'}}{Z_0} = \frac{R/Z_0}{1 + j2Q_u\delta} + 1$$

A coupling coefficient β is given by

$$z_{xx'} = \frac{2\beta}{1+j2Q_u\delta} + 1 \qquad -(5)$$

At $\omega = \omega_0$ $z_{xx'} = 2\beta + 1$.

The reflection coefficient at $\omega = \omega_0$ at xx' plane is

$$\Gamma_{xx'}(\omega_0) = \frac{z_{xx'} - 1}{z_{xx'} + 1} = \frac{\beta}{\beta+1} \qquad -(6)$$

In general

$$\Gamma_{xx'} = \frac{z_{xx'} - 1}{z_{xx'} + 1} = \frac{\beta}{\beta+1+j2Q_u\delta} \qquad -(7)$$

The reflection coefficient seen at i/p of line, denoted by $\Gamma_{yy'}$ is

$$\Gamma_{yy'} = \frac{\beta}{\sqrt{(\beta+1)^2 + (2Q_u\delta)^2}} e^{-j\left(2\theta + \tan^{-1}\frac{2Q_u\delta}{\beta+1}\right)} \qquad -(8)$$

At resonant frequency (8) reduces to

$$\Gamma_{yy'}(\omega_0) = \Gamma_{xx'}(\omega_0) e^{-j2\theta} = \frac{\beta}{\beta+1} e^{-j2\theta} \qquad -(9)$$

(9) shows that if β - constant & length of line is varied from $0°$ to $360°$ the values of $\Gamma_{YY'}$ lie on a ⊙ in the Smith Chart.

→ with appropriate selection of β the reflection coefficient $\Gamma_{YY'}$ can implement any passive impedance.

→ (9) is used to select the desired coupling and electrical length 'θ' of transmission line for a particular input impedance.

The parameters β, ω_o & Q_u describes the operations of DR.

The values of R, L and C can be calculated in terms of β, ω_o and Q_u. The values of R are calculated using

$$\beta = \frac{R}{2Z_0}$$

L & C are obtained as

$$Q_u = \omega_o RC = \frac{R}{\omega_o L}$$

At resonant frequency ω_o, the S parameters of the tuned ckt are

$$[S(\omega_o)] = \begin{bmatrix} \frac{\beta}{\beta+1} & \frac{1}{\beta+1} \\ \frac{1}{\beta+1} & \frac{\beta}{\beta+1} \end{bmatrix} \quad - (10)$$

β can be expressed in terms of $S_{11}(\omega_0)$ & $S_{21}(\omega_0)$.

$$\beta = \frac{S_{11}(\omega_0)}{1 - S_{11}(\omega_0)} = \frac{1 - S_{21}(\omega_0)}{S_{21}(\omega_0)}$$

For a closely coupled resonator, β are from 2 to 20

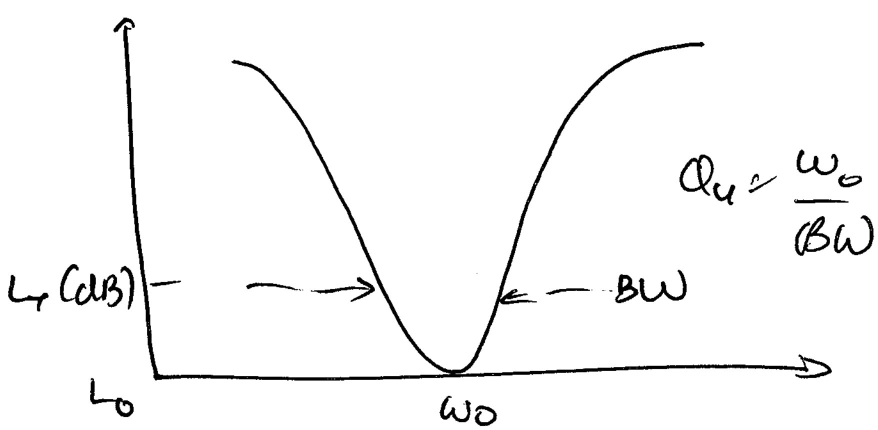
Q_u can be found by measuring S_{21}

insertion loss L_0 is given by

$$L_0\,(dB) = -20 \log |S_{21}(\omega_0)|$$

L_x is given by

$$L_x\,(dB) = L_0\,(dB) - 3 + 10 \log(1 + 10^{-0.1 L_0})$$



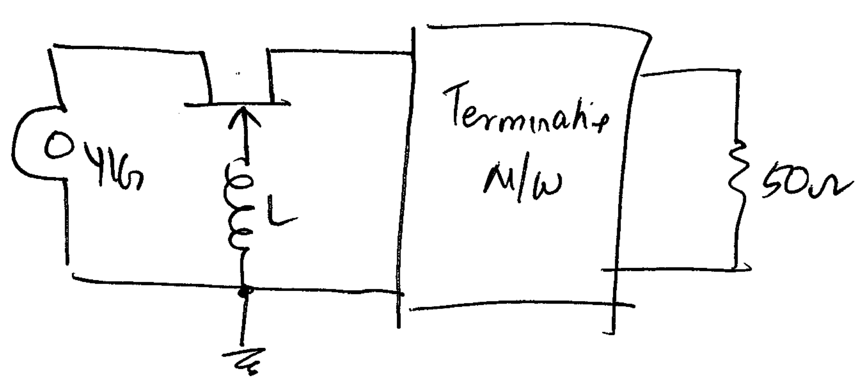
YIG Oscillators

→ YIG resonator consists of a ferrimagnetic material which can be modeled by a $\|^{el}$ RLC resonant ckt. The value of the elements depends on the magnetization, coupling & resonance linewidth of the YIG sphere & on the applied dc magnetic field.

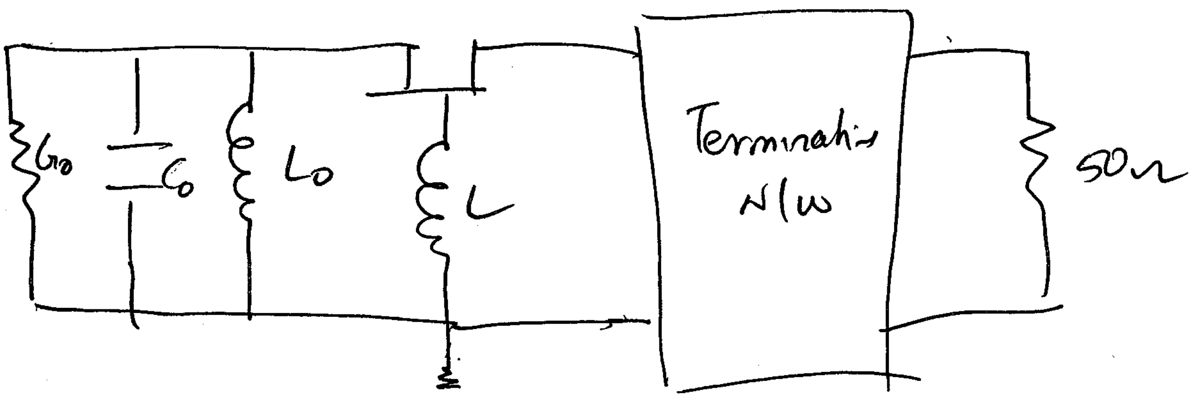
→ The uniform DC magnetic field is applied with an electromagnet with a single gap.

→ The gap design is important since a nonuniform dc magnetic field results in a tuning hysteresis & spurious responses.

→ A CG GaAs FET oscillator using a YIG resonator is shown in figure below.



Assuming YIG sphere is always magnetically saturated and the sphere diameter is $\ll \frac{\lambda}{4}$, the YIG device can be modeled as a $\parallel^{el}$ resonant ckt shown in figure below



The element values are given as

$$G_0 = \frac{d^2}{\mu_0 V \omega_m Q_u}$$

$$L_0 = \frac{\mu_0 V \omega_m}{\omega_0 d^2}$$

$$C_0 = \frac{1}{\omega_0^2 L_0}$$

$$\omega_m = \gamma\, 8\pi^2 M_s$$

$$Q_u = \frac{H_0 - 4\pi M_s/3}{\Delta H}$$

$4\pi M_s \rightarrow$ saturation magnetization

$V \rightarrow$ volume of sphere

$d \rightarrow$ coupling loop diameter

$\gamma \rightarrow$ gyromagnetic ratio (2.8 MHz / Oe)

$H_0 \rightarrow$ applied DC magnetic field.

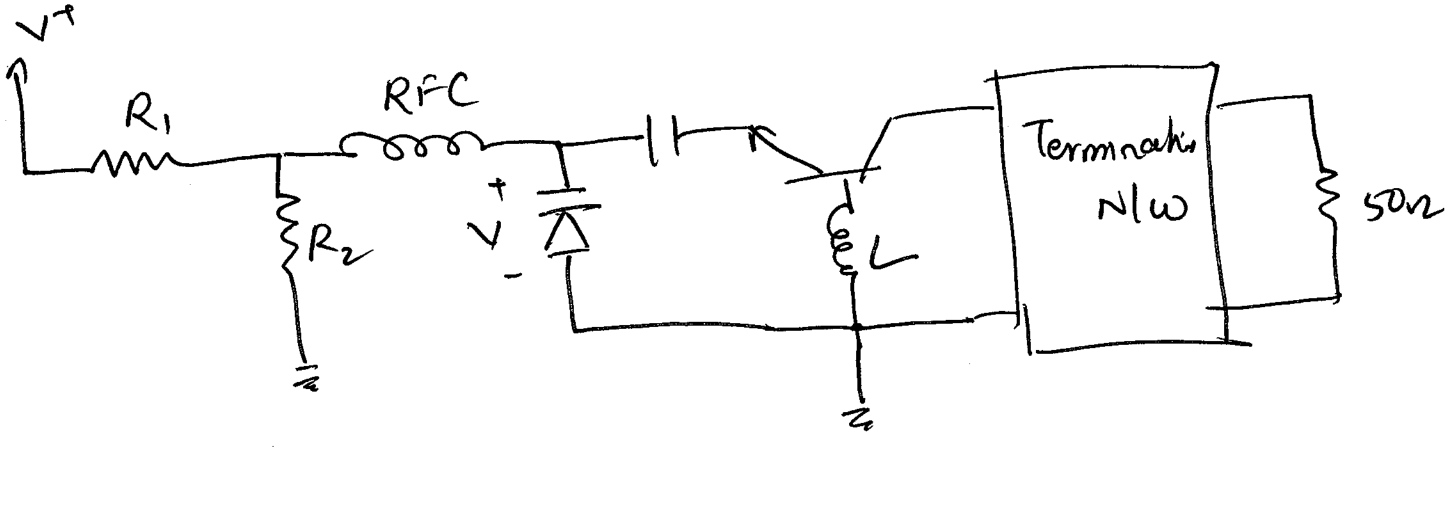
$Q_u \rightarrow$ unloaded Q.

ΔH – resonance line width

ω_0 – center frequency.

Varactor - Tuned Oscillators

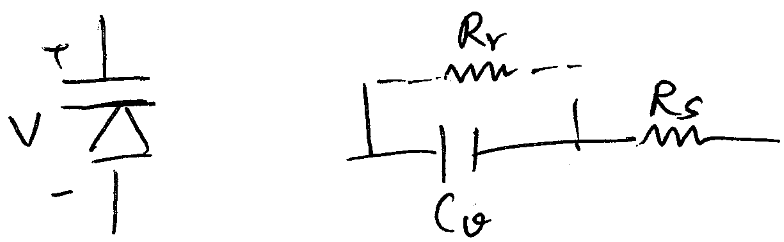
→ Uses voltage controlled capacitance of a varactor diode to accomplish the electronic tuning



The capacitance of varactor diode is determined by voltage V, which is set by V^+; R_1 and R_2

The varactor ckt shown below has varactor diode capacitance (C_v) for Schottky-type devices as.

$$C_v = \frac{C_0}{\left(1+\frac{V}{\phi}\right)^{1/2}}$$



C_0 – value of capacitance at zero voltage

V → reverse bias voltage

ϕ → Junction contact potential ($\phi \approx 0.7V$)

R_s → Series resistance of diode.

R_r → reverse diode resistance