Lecture 15 - Microwave Oscillator Design

Microwave Active Circuit Analysis and Design

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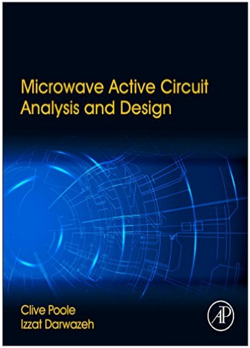

Intended Learning Outcomes

- Knowledge
  - Understand that any electronic oscillator can be understood by employing either a feedback oscillator model or a negative resistance model, and be aware of the equivalence between the two models.
  - Be familiar with the most common RF feedback oscillator topologies, such as Copitts, Hartley, Clapp/Gouriet and Pierce.
  - Be familiar with the advantages of the cross-coupled oscillator topology compared with other topologies.
  - Be familiar with the theory and application of various types of high stability resonator such as crystals, cavities, dielectric resonators and YIG resonators.
  - Understand how to generate negative resistance in a transistor using feedback.
  - Understand how varactors are used to implement an electronically tunable oscillator.
- Skills
  - Be able to carry out a simple fixed frequency negative resistance transistor oscillator design.
  - Be able to carry out a simple fixed frequency cross-coupled transistor oscillator design.

Simple conceptual oscillator

- The simplest conceptual oscillator can be modelled as a single capacitor and a single inductor connected together, as shown in figure 1. This is the simplest representation of a "tank" circuit, which is an essential subcircuit in any oscillator
- The switch S is included in figure 1 to define the exact instant t_o when current begins to flow (i.e. the instant when the switch is closed)
- We assume that all components in figure 1 are lossless, in other words there is zero resistance around the loop

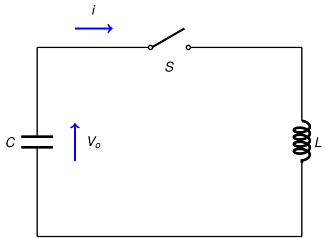

Figure 1 : Simplest possible electronic oscillator

Simple conceptual oscillator

- Assume the capacitor C is charged up to a static voltage V_o. At time $t = t_o$, we close the switch, S, thereby connecting the capacitor across the inductor, L. Current starts to flow from the capacitor into the inductor, starting from $i(t) = 0$ at time $t = t_o$ and building up to a maximum some time later.
- At the instant the inductor current reaches its maximum value, the voltage across the inductor will be zero, since the capacitor has been fully discharged, and the voltage across the capacitor must equal that across the inductor, in order to satisfy Kirchhoff's voltage law.
- To satisfy Kirchhoff's current law the same current that is flowing through the inductor at any instant, also has to flow through the capacitor. This means that the current flowing in the inductor will start to "recharge" the capacitor from $v_c(t) = 0$ back up towards $v_c(t) = V_o$, from whence the cycle will repeat.
- Since the circuit in figure 1 is lossless, the cycle will repeat indefinitely, with energy being exchanged back and forth between the capacitor and inductor, whilst the total energy in the circuit remains constant, as shown in figure 2.

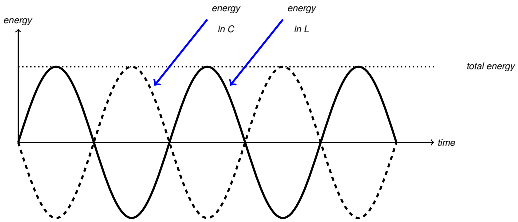

Figure 2 : Energy exchange in an oscillating LC circuit

Simple conceptual oscillator

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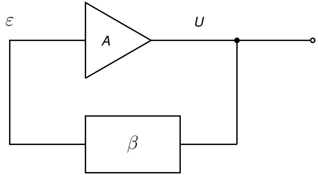
The feedback oscillator model

- Every oscillator must contain some form of active device to provide power gain.
- An oscillator can be considered as comprising an amplifier with gain A, coupled to a feedback network having a feedback factor β
- Both A and β are complex functions of frequency (ω).



$$\beta A = 1 \tag{1}$$

Where : A = Amplifier voltage gain
β = feedback fraction.

The feedback oscillator model

In order to build an oscillator, one needs a device which is capable of gain ($|A(j\omega_o)| > 1$) at the frequency of interest. Therefore any criteria which determine the amplification potential of a given active device also potentially determine its oscillation potential. In other words, and put simply; no gain means no oscillation.

We can, for example, implement the passive feedback network $\beta(j\omega)$ as a Π or T network of purely reactive elements. Consider a Π network shown in figure 3. If we make the reasonable assumption that the amplifier gain, $A(j\omega)$, is more or less flat over the frequency range of interest, then a simple analysis shows that to fulfil (**??**) , the reactance X_1, X_2 and X_3 need to satisfy the following condition:

$$X_3 = -(X_1 + X_2) \quad (2)$$

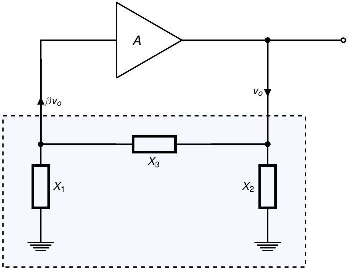


Figure 3 : Feedback oscillator with Π feedback network

The feedback oscillator model

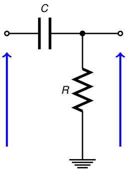
Equation (**??**) implies that, if X_3 is chosen to be an inductor, then X_1 and X_2 should be capacitors, and vice versa. We can now define two well known and widely used RF oscillators in table 1.

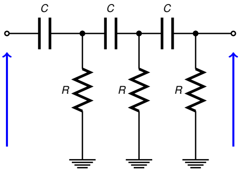
Table 1 : Classic feedback oscillators

Colpitts oscillator	$X_1 = Capacitor$ $X_2 = Inductor$ $X_3 = Inductor$
Hartley oscillator	$X_1 = Inductor$ $X_2 = Capacitor$ $X_3 = Capacitor$

Feedback Network Implementation

At low frequencies we can produce a phase shift of (180^o) using a 3 stage RC ladder network (each RC stage generates 60^o phase shift) :





The 3 stage RC network gives a phase shift of (180^o) when :

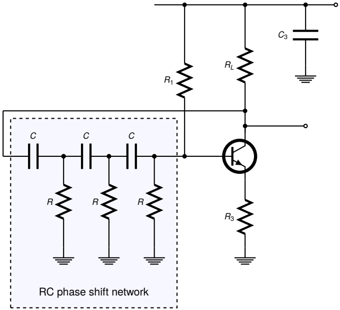
$$\omega = \frac{1}{RC\sqrt{6}} \tag{3}$$

at this frequency the feedback fraction is

$$\beta = \frac{v_o}{v_i} = \frac{1}{29} \tag{4}$$

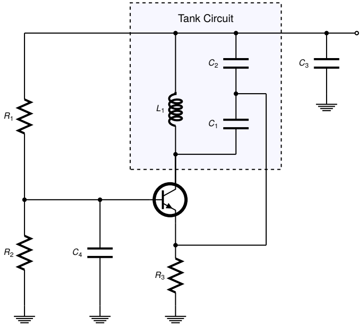
RC phase shift oscillator

- The gain of the CE inverting amplifier must exactly balance the loss of the feedback network. This means that the amplifier should have a voltage gain of 29.
- The phase shift of the amplifier (180^o) plus the phase shift of the feedback network (180^o) results in an overall phase shift of (360^o)
- The exact frequency of operation is determined by the values of R and C.



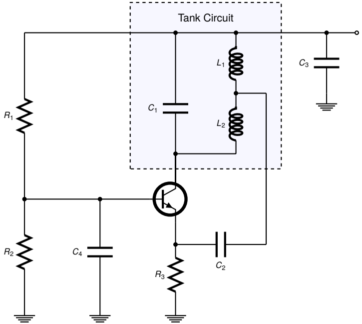
Colpitts Oscillator

- The Colpitts oscillator is an electronic oscillator circuit that uses an two capacitors in series and a parallel inductor in to determine the frequency.
- Power is delivered to the load, R
- Feedback is taken from the junction of L and C
- Adjustment of operataing frequency requires either a variable inductor or dual ganged variable capacitors.
- The Colpitts oscillator is the electrical dual of the Hartley oscillator.



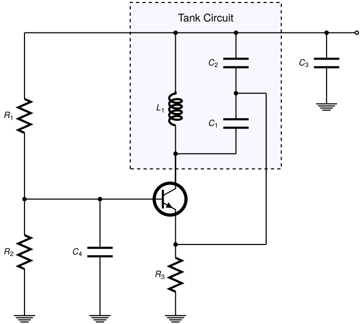
Hartley Oscillator

- The Hartley oscillator is an electronic oscillator circuit that uses an two inductors in series (or a centre-tapped inductor) and a parallel capacitor in to determine the frequency.
- Power is delivered to the load, R
- Feedback is taken from the junction of L and C
- The operating frequency may be adjusted using a single variable capacitor in place of C.



Clapp/Gouriet Oscillator

The Clapp/Gouriet oscillator topology is basically a Colpitts oscillator with an additional capacitor C_o added in series with the inductor L_1 in the tank circuit



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Cross-coupled oscillator

- Consider the single stage FET tuned amplifier shown in figure 4.
- The load of the FET in this cas e consists of a discrete RLC resonator.
- The susceptances of the *L* and *C* cancel at resonance, resulting in the resonator impedance reaching a maximum, purely real, value equal to *R*.

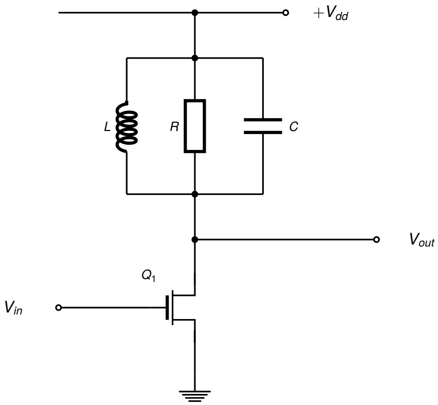

Figure 4 : Single stage FET tuned amplifier

Cross-coupled oscillator

- We can represent the FET in figure 4 by its absolute simplest small signal model, coupled to the equivalent load impedance of the tank circuit, Z_{tank}, as shown in figure 5.
- The small signal gain of the equivalent circuit in figure 5 is given by:

$$v_{out} = -g_m v_{gs} Z_{tank} \tag{5}$$

With the minus sign representing the 180^o degree phase shift between AC input voltage and output voltage, i.e. the circuit behaves as a signal *inverter*.

- The small signal gain, $A(j\omega)$ of the inverter in figure 4 is therefore :

$$A_1(j\omega) = \frac{v_{out}}{v_{gs}} = -g_m Z_{tank} \tag{6}$$

$$= \frac{-g_m}{\frac{1}{R} + j\left(\omega C - \frac{1}{\omega L}\right)} \tag{7}$$

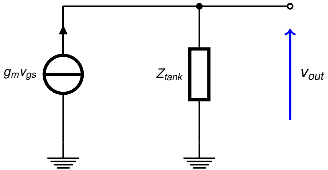

Figure 5 : MOSFET small signal model

Cross-coupled oscillator

- To make an oscillator using the tuned amplifier of figure 4, we need to add one more inversion to achieve an overall phase shift of 360^o around the loop.
- We can do this by adding a second inverter stage and a feedback path as shown in figure 6.

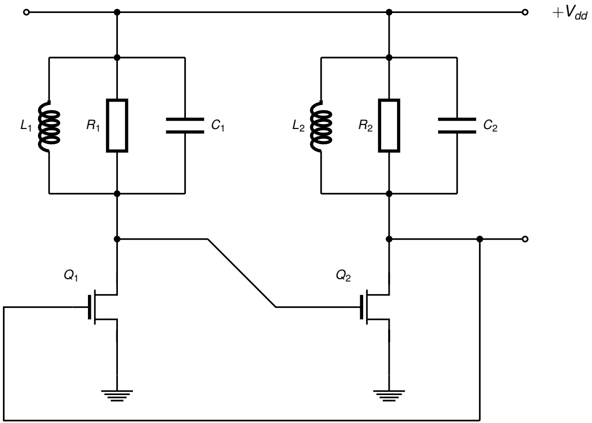

Figure 6 : Two stage FET tuned amplifier

Cross-coupled oscillator

- The circuit of figure 6 is more commonly drawn in the format shown in figure 7.
- If we set $R_1 = R_2 = R$, $C_1 = C_2 = C$ and $L_1 = L_2 = L$ then the gain of the second stage in figure 6 will be identical to the first. The overall loop gain of the circuit in figure 6 will therefore be :

$$A_2(j\omega) = \left[\frac{-g_m}{\frac{1}{R} + j\left(\omega C - \frac{1}{\omega L}\right)} \right]^2 \tag{8}$$

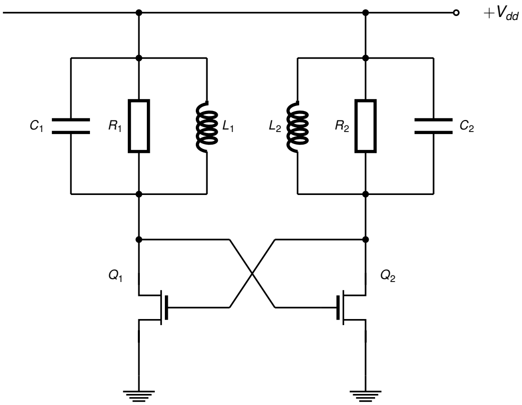

Figure 7 : Cross-coupled oscillator

Cross-coupled oscillator

Setting the imaginary part of (8) equal to zero gives the oscillation frequency of the circuits of figure 6 and figure 7 as:

$$f_o = \frac{1}{2\pi\sqrt{LC}} \tag{9}$$

And a loop gain, at resonance, of:

$$A_2(j\omega_o) = (g_m R)^2 \tag{10}$$

Since oscillation requires $A_2(j\omega_o) \geq 1$, we can write the condition for oscillation, from (10) as :

$$g_m R \geq 1 \tag{11}$$

Cross-coupled oscillator

Biasing the cross-coupled pair via a current source, as shown in figure 8(a), provides better amplitude control. With the current source inserted, the amplitude of the output will be $I_{SS}R$.

In many MMIC applications the two tank circuits of a cross coupled oscillator are merged, as shown in figure 8(b). In such cases, the equations for oscillation frequency and loop gain, (11) and (9), still apply.

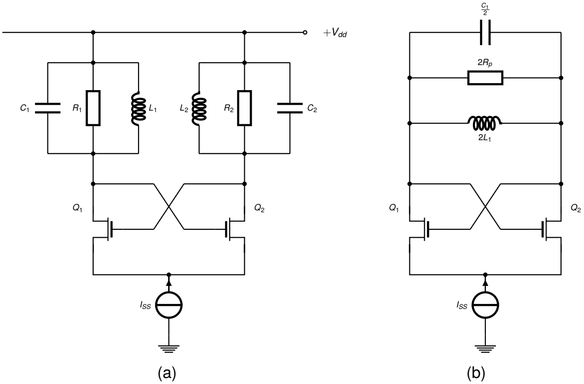

Figure 8 : MMIC implementation of the cross coupled oscillator : (a) current source bias, (b) load tanks merged

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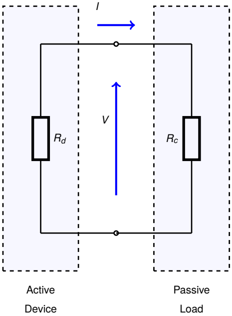
Negative Resistance

A simple one-port negative resistance circuit is shown in figure 9(a). If we observe the current flowing out of the one-port active device and into the passive load then the resistance of the active device must be:

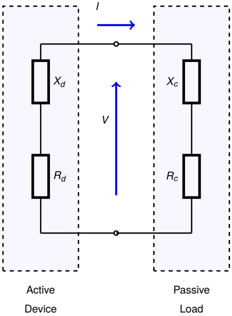
$$R_d = \frac{V}{-I} = -R_c \tag{12}$$

The resistance of the active device is therefore of equal magnitude but of opposite sign to that of an equivalent passive resistor. The condition for sustained current flow in the circuit of figure 9(a) is therefore:

$$R_d + R_c = 0 \tag{13}$$



(a) Pure Resistances



(b) Complex Impedances

Negative Resistance

Condition (13) is independent of frequency, suggesting that the circuit of figure 9(a) is capable of power generation over an infinite bandwidth. In the real world, both source and load will inevitably contain reactive elements which can be represented by X_d and X_c in figure 9(b). We therefore have the oscillation condition for the circuit of figure 9(b) as [7]:

$$Z_d + Z_c = 0 \tag{14}$$

In terms of resistive and reactive components (14) becomes:

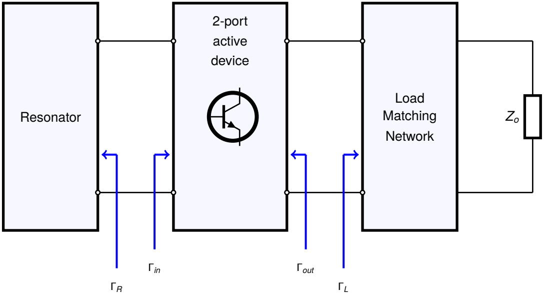
$$R_d = -R_c \tag{15}$$
$$X_d = -X_c \tag{16}$$

Finally, in terms of reflection coefficients, the oscillation condition or the circuit of figure 9(b) can be expressed as :

$$\Gamma_d \Gamma_c = 1 \tag{17}$$

Transistor Negative Resistance Oscillator

- Modern RF oscillators are mostly based on transistors which are characterised as 2-port active devices.
- The 1-port negative resistance design concept can be extended to transistor oscillator design by applying the oscillation conditions at each port.
- We generally connect the resonator to port 1 and the load to port 2.



N-port negative resistance oscillator

The one-port oscillation condition of (17) is really a special case of a more general oscillation condition for networks with an arbitrary number of ports. S-parameter analysis of n-port negative resistance oscillators is based on the oscillator representation shown in figure 10[6, 2]. The oscillator is considered as being divided into two-parts: (a) the *n*-port active device and (b) a passive *n*-port network which represents the resonant circuit(s), bias components output matching networks and any other such passive circuitry. In this conception, the passive network is explicitly taken to also include the load.

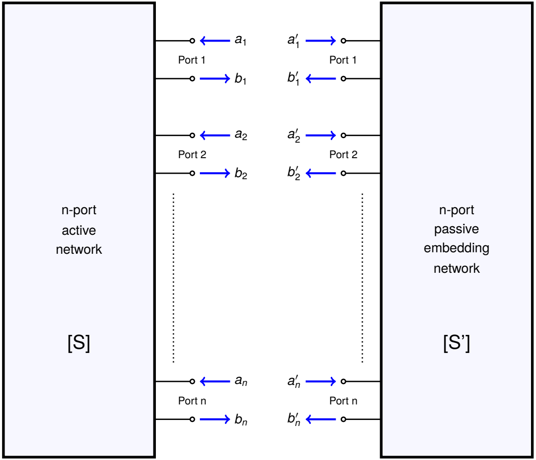


Figure 10 : Generalised n-port oscillator model

N-port negative resistance oscillator

The power wave relationship for the active device of figure 10 is as follows:

$$[b] = [S]\,[a] \tag{18}$$

Where $[S]$ is the nxn S-matrix of the active device. Similarly, the power wave relationship for the passive embedding network in figure 10 is:

$$\left[b'\right] = \left[S'\right]\left[a'\right] \tag{19}$$

Where $[S']$ is the nxn S-matrix of the embedding network. When the n-ports of figure 10 are connected together at each port the following relationships apply:

$$\left[b'\right] = [a] \tag{20}$$

$$[b] = \left[a'\right] \tag{21}$$

From equations (15) to (21), the following relationships can be deduced:

$$\left[a'\right] = [S]\left[S'\right]\left[a'\right]$$
$$\left[[S]\left[S'\right] - [I]\right]\left[a'\right] = 0 \tag{22}$$

Since in general $[a'] \neq 0$, equation (22) gives the following oscillation condition which applies to devices with any number of ports:

$$\left|[S]\left[S'\right] - [I]\right| = 0 \tag{23}$$

2-port negative resistance oscillator

Figure 11 shows a transistor in common emitter configuration and represented by its common emitter two-port S-matrix connected to an embedding network consisting of two passive impedances Γ_1 and Γ_2.

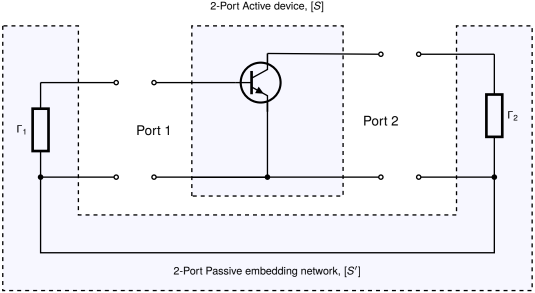


Figure 11 : 2-port oscillator model

The scattering matrix for the transistor in figure 11 is given by:

$$[S] = \begin{bmatrix} S_{11} & S_{21} \\ S_{12} & S_{22} \end{bmatrix} \quad (24)$$

The *S*-matrix for the embedding network in figure 11 can be written as.-

$$[S'] = \begin{bmatrix} \Gamma_1 & 0 \\ 0 & \Gamma_2 \end{bmatrix} \quad (25)$$

2-port negative resistance oscillator

Applying equation (23) to this case gives the following oscillation condition for the circuit of figure 11:

$$\begin{vmatrix} (S_{11}\Gamma_1 - 1) & S_{12}\Gamma_1 \\ S_{21}\Gamma_2 & (S_{22}\Gamma_2 - 1) \end{vmatrix} = 0 \tag{26}$$

Expanding the determinant in (26) results in the following two simultaneous oscillation conditions :

$$S_{11} + \frac{S_{12}S_{21}\Gamma_2}{1 - S_{22}\Gamma_2} = \frac{1}{\Gamma_1} \tag{27}$$

$$S_{22} + \frac{S_{12}S_{21}\Gamma_1}{1 - S_{11}\Gamma_1} = \frac{1}{\Gamma_2} \tag{28}$$

By comparison with equations (**??**) and (**??**) we can see that the left-hand sides of equations (27) and (28) correspond to Γ_{in} with port 2 terminated in Γ_2 and Γ_{out}, and with port 1 terminated in Γ_1 respectively. Equations (27) and (28) can therefore be re-written in simpler form as:

$$\Gamma_{in}\Gamma_1 = 1 \tag{29}$$

$$\Gamma_{out}\Gamma_2 = 1 \tag{30}$$

2-port negative resistance oscillator

If we arrange for condition (29) to be satisfied then we will find that (30) is automatically satisfied as well[1]. In other words, if the active two-port exhibits negative resistance at port 1 and Γ_1 is chosen such that condition (29) is satisfied, then Γ_{out} will assume a value such that:

$$\Gamma_{out} = \frac{1}{\Gamma_2} \tag{31}$$

If we consider a two-port oscillator with the output at port 2, then Γ_2 would be the matched load, i.e. $\Gamma_2 = 0$. Then, according to condition (31):

$$\Gamma_{out} = \rightarrow \infty \tag{32}$$

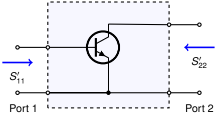
Which is consistent with port 2 being a source of signal power (i.e. $b_2 = 0$ when $a_2 = 0$).
The oscillation condition represented by (29) and (30) implies that, since Γ_1 and Γ_2 are both passive, $|S_{ii}|$ for the transistor should be greater than unity (where $i = 1, 2$). This is rarely the case for a transistor connected in the common emitter configuration without any feedback, as in figure 11.

Generating negative resistance in transistors

Transistors do not normally exhibit negative resistance behaviour. This normally has to be induced in some way by circuit configuration or design. There are basically two ways to induce negative resistance in a transistor :

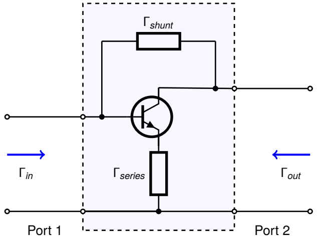
1. Change device configuration : Changing the configuration from the common emitter to common base or common collector sometimes results in $|S_{11}| > 1$ without any external components being required.
2. Apply feedback : adding a passive feedback element of a specific value

Sometimes, in order to get the required negative resistance we need to both change the configuration AND apply feedback.



Feedback Topologies

- There are basically two ways to place a feedback element around a 3-terminal active device :
- Series feedback : the feedback element is placed in series with the common terminal
- Shunt feedback : the feedback element is placed across the device between output port and input port.



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Crystal stabilisation

Q in Oscillators

$$Q = 2\pi \frac{Energy\ \ stored}{Energy\ \ lost\ \ per\ \ cycle} \tag{33}$$

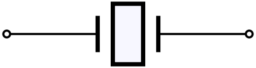
- Q is a function of the electrical losses in the components used to fabricate the resonator. Lower losses = Higher Q. High Q is desirable because :
- The higher the Q the greater the frequency accuracy and stability of the oscillator.
- The higher the Q the lower the phase noise of the oscillator In tunable oscillators there is a trade-off between high Q and tunability.

Quartz Crystals

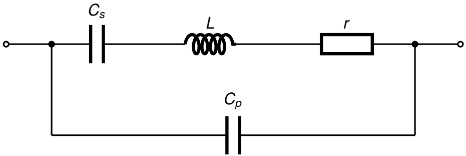
- Lumped components cannot provide very high Q values due to electrical losses and other parasitics.
- Lumped component values tend to drift with changing environmental conditions (temperature etc.).
- To obtain higher Q (greater than 160) and greater temperature stability we use Quartz Crystals which are electromechanical resonators which rely on the piezoelectric effect.
- Because Crystals are electromechanical devices, their upper resonant frequency is limited to around 250MHz.

Quartz Crystals

- The equivalent circuit for the quartz crystal is usually represented as an RLC series circuit, which represents the mechanical vibrations of the crystal, in parallel with a capacitance C_p, which represents the electrical connections to the crystal.
- Quartz crystal oscillators operate at "parallel resonance", and the equivalent impedance of the crystal has a series resonance where C_s resonates with inductance, L and a parallel resonance where L resonates with the series combination of C_s and C_p.



(a) Crystal circuit symbol



(b) Crystal equivalent circuit

Figure 12 : Crystal equivalent circuit

Quartz Crystal Resonance

- The slope of the reactance against frequency curve in figure 13 shows that the series reactance at frequency ω_p is inversely proportional to C_s because below ω_s and above ω_p the crystal appears capacitive
- Between frequencies ω_s and ω_p, the crystal appears inductive as the two parallel capacitances cancel out
- The point where the reactance values of the capacitances and inductance cancel each other out $X_c = X_L$ is the fundamental frequency of the crystal

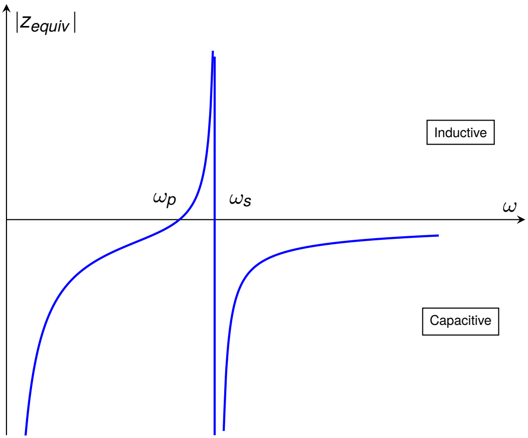


Figure 13 : Quartz crystal reactance characteristic

Pierce Crystal Oscillator

- One very common crystal oscillator topology is the Pierce circuit shown in figure 14. This circuit is essentially a common emitter amplifier with a feedback network, from collector to base, consisting of the π-network comprising C_1, C_2 and the crystal, X_1.
- The circuit oscillates just above the series resonant frequency of the crystal, i.e. the crystal is being used in it's inductive mode.

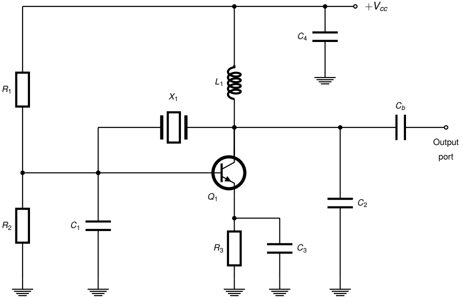


Figure 14 : BJT Pierce Crystal Oscillator

Pierce Crystal Oscillator

The load impedance seen by the transistor in figure 14, at the collector, can be written as :

$$Z_L = jX_2 \parallel \left[Z_{equiv} + jX_1 \parallel h_{ie}\right] \quad (34)$$

Where Z_{equiv} is the equivalent resonant impedance of the crystal, h_{ie} is the input impedance looking into the base of the transistor at that bias point and the symbol '$\parallel$' indicates a parallel connection. As we are operating the crystal close to ω_s, the resonant impedance will be inductive, i.e., $Z_{equiv} = r_e + j\omega L_e$. If $|X_1| << h_{ie}$ we can write :

$$Z_L = \frac{jX_2\left[r_e + j(X_1 + X_e)\right]}{r_e + j(X_1 + X_2 + X_e)} \quad (35)$$

If we assume that r_e is very small we can see that resonance, occurs when :

$$(X_1 + X_2 + X_e) = 0 \quad (36)$$

Solving (36) for ω_o results in :

$$\omega_o = \frac{1}{\sqrt{L_e\left(\frac{C_1 C_2}{C_1 + C_2}\right)}} \quad (37)$$

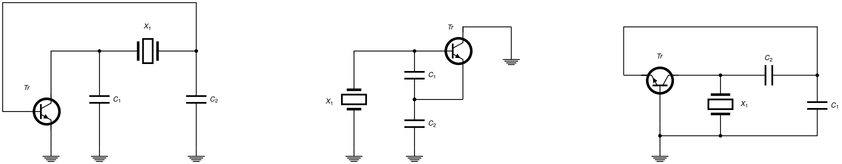
The gain condition for oscillation of the circuit in figure 14 is given by Gonzalez as [3]:

$$\frac{g_m}{\omega_o^2 R_e C_1 C_2} > 1 \quad (38)$$

Where g_m is the transconductance of Q_1

Crystal Oscillator Topologies

the Pierce, Colpitts and Clapp/Gouriet oscillator topologies are very closely related and only distinguished by which terminal of the transistor is grounded. This is illustrated schematically in figure 15, with biasing circuitry omitted for clarity.



(a) Pierce crystal oscillator (CE)

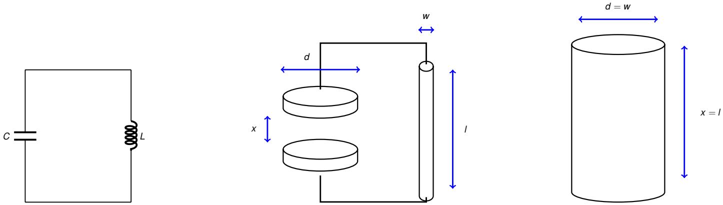
(b) Colpitts crystal oscillator (CC)

(c) Clapp/Gouriet crystal oscillator (CB)

Figure 15 : Crystal oscillator topologies

Evolution of a cylindrical cavity resonator

In figure 16(a) we see the conventional LC parallel tank circuit consisting of a single discrete capacitor and inductor. As frequency increases, the required value of both L and C are reduced. Eventually we reach a point where the capacitor is comprised of just two parallel circular plates, of diameter d, separated by a distance x. We also envisage the inductor as being comprised of a single hollow cylindrical conductor, of diameter w, and length l, as shown in figure 16(b). As the frequency increases further, the values of L and C reduce further, meaning that both the spacing between the capacitor plates and the diameter of the hollow cylindrical conductor increase. Ultimately, as frequency continues to increase, we reach a point where $d = w$ and $x = l$, so we can merge the two components together as shown in figure 16(c).



(a) LC circuit (b) LC intermediate stage (c) Cylindrical cavity

Figure 16 : Evolution of a cylindrical cavity resonator

Cavity stabilised oscillators

- Figure 17 shows three popular configurations used for microwave cavity stabilised oscillators [8]. The cavity can be coupled to the rest of the circuit via a small aperture, a wire probe or a loop.
- In the reaction stabilised circuit of figure 17(a), the cavity is situated between the active device and the load and is coupled to a transmission line connecting them. The cavity acts as a band reject filter and at the resonant frequency a small amount of power is reflected back towards the active device. This mode of operation provides high external Q but is susceptible to parasitic oscillation and mode jumping.
- The reflection stabilised circuit of figure 17(b) is well suited to situations where the active device can be configured as a two-port network. The cavity provides a passive termination at one-port of the active device which satisfies the oscillation condition, (29), at a particular frequency. This circuit has a slightly lower external Q than the reaction stabilised oscillator, but is relatively free from parasitic oscillations.
- In the transmission stabilised circuit (sometimes called the Kurokawa circuit) of figure 17(c), the cavity acts as a band-pass filter between the active device and the load. This configuration has a very high external Q and is practically free from parasitic oscillation, but the output power may be low due to coupling losses.

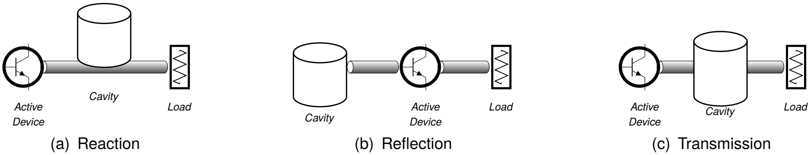
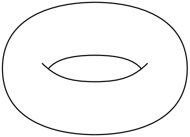


Figure 17 : Cavity stabilsation of a transistor oscillator: Three possible configurations

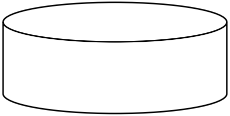
Dielectric resonator

The term dielectric resonator was first used by Richtmyer who postulated that if a cylindrical dielectric waveguide were bent round to form a torus, and joined at the ends, as shown in figure 18(a), then electromagnetic radiation would propagate inside the torus at certain distinct frequencies determined by the requirement that the fields be continuous at the junction[10].

The original torus shape proposed by Richtmyer is not widely used in practice as it is difficult to fabricate. The "puck" shape shown in figure 18(b) has very similar electrical properties to the torus and is much easier to fabricate. Most practical dielectric resonators therefore employ the puck format.



(a) Dielectric torus



(b) Dielectric puck

Figure 18 : Dielectric resonators

Dielectric Resonator resonant frequency

There are three types of resonant modes that can be excited in a dielectric resonator, namely the transverse electric (TE), transverse magnetic (TM) or hybrid electromagnetic (HEM) modes. Theoretically, there is an infinite number of modes in each of these three categories. In practice, it is the TE_{01n} mode that is most widely used in oscillator applications. The approximate resonant frequency, in GHz, of the TE_{01n} mode for an isolated cylindrical dielectric resonator, of the type shown in figure 18(b) is given by [4]:

$$f_{GHz} = \frac{34}{a\sqrt{\epsilon_r}}\left(\frac{a}{h} + 3.45\right) \tag{39}$$

Where a is the radius of the resonator puck and h is its height. Both a and h are in millimeters in equation (39), which is accurate to about $\pm 2\%$ in the range:

$$0.5 < \frac{a}{L} < 2$$

$$30 < \epsilon_r < 50$$

Dielectric Resonator resonant frequency

If the material of the dielectric resonator has a large dielectric constant (i.e. the majority of the field is confined to a region close to the resonator) then the Q of the resonator is approximately given by:

$$Q = \frac{1}{tan(\delta)} \tag{40}$$

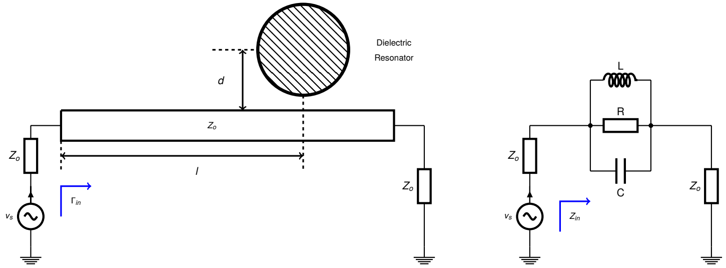
Where : δ = dielectric loss of the material.
The linear dimension of a dielectric resonator is in the order of $\lambda_o/\sqrt{\varepsilon_r}$, where λ_o is the free space wavelength and $\sqrt{\varepsilon_r}$ is the dielectric constant. For $\sqrt{\varepsilon_r} = 100$, the resonator is approximately one tenth of the size of an equivalent metallic cavity, making such resonators very attractive for use in practical oscillators.

Dielectric Resonator coupled to a microstrip line

The coupling between the resonator and a microstrip line is adjusted by varying the spacing between the resonator and the line, d. Figure 19(b) shows an equivalent circuit of figure 19(a) in which the dielectric resonator is modelled as an LCR resonant circuit. The normalised input impedance of the transmission line in figure 19(a) at a frequency $\Delta\omega$ away from the resonant frequency, ω_o, is given by [9, 11, 5]:

$$Z_{in} = 1 + \frac{\beta}{1 + jQ_o \frac{\Delta\omega}{\omega_o}} \tag{41}$$

Where β is the coupling coefficient of the resonator to the line and Q_o is the unloaded Q of the resonator. β can be varied by moving the resonator relative to the line. In this way, the impedance, Z_{in}, can be continuously varied and a position can be found where the oscillation condition is satisfied.



(a) Physical configuration (b) Equivalent circuit

Dielectric Resonator Oscillator Example

- Figure 20 shows a FET oscillator using a dielectric resonator as the stabilising element.
- The FET is made to present a negative resistance at the gate port by adding a series feedback reactance, jX_{fb} in the common source lead.

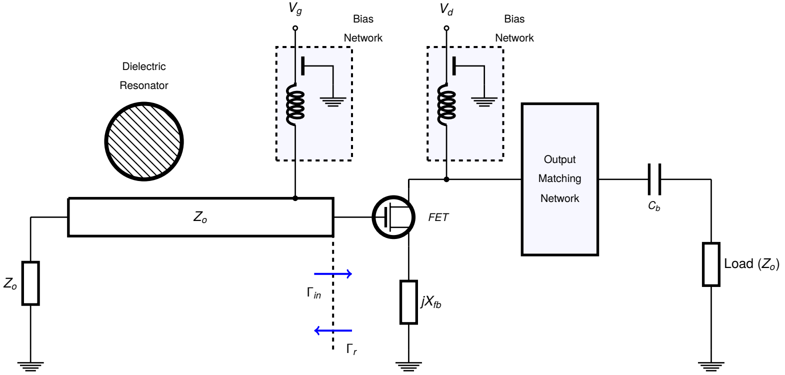


Figure 20 : FET based dielectric resonator oscillator

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BJT YIG tuned oscillator

- A sphere of Yttrium Iron Garnet (YIG) material, typically less than 1mm in diameter, can act as a microwave resonator that can be coupled to an active device via a wire loop.
- YIG resonators exhibit very high Q, similar to that of dielectric resonators, but have the added advantage that the resonant frequency can be tuned over a very wide range by varying an externally applied DC magnetic field.
- The frequency of resonance is a very linear function of the strength of the applied magnetic field and typically increases at a rate of 2.8 MHz per Gauss.
- A typical YIG oscillator uses a BJT in CB configuration with inductive series feedback to produce negative resistance at the input port (i.e. $|\Gamma_{in}| > 1$).

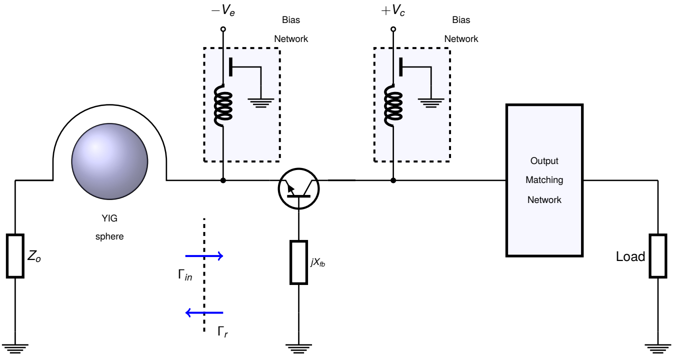

Figure 21 : Transistor YIG tuned oscillator

Varactor tuned oscillators

- A very common VCO configuration, especially in MMIC format, is based on the cross-coupled topology described previously.
- To convert the fixed frequency oscillator shown in figure 8 into a VCO we replace the tuning capacitors by varactor.
- The basic layout of a typical cross-coupled VCO topology is shown in figure 22.
- The capacitance of the two varactors VC_1 and VC_2 will be inversely proportional to the voltage, V_c, applied to their common connection point.
- The output frequency of the VCO will therefore be directly proportional to V_c.

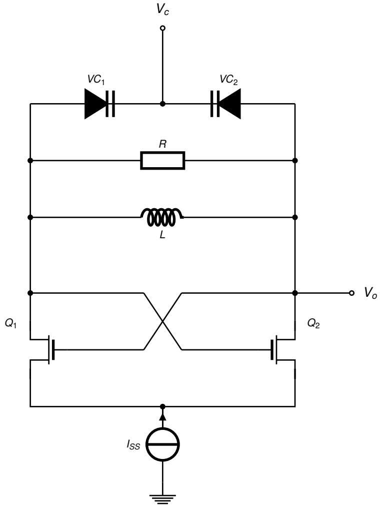

Figure 22 : Cross-coupled VCO using varactors

Varactor tuned vs YIG tuned oscillators

The advantages of varactor tuning are low power consumption, small size and high tuning speed. The price to be paid for these advantages are a relatively narrow tuning range of around 3% and a relatively low Q.

Table 2 gives a comparison of the current state-of-the-art for YIG and varactor tuning.

Parameter	YIG tuning	Varactor Tuning
Bandwidth	Very wide	Wide
Linearity	$<$ + 1	12
Slew rate	$<$ 1 MHz 1S	1 - 10 GHz 4S
Step response time	1 - 3 mS	$<$ 0.1 μSec
post tuning drift time constant	Seconds to minutes	μSeconds - milli-Seconds
Temperature stability	20 - 100 ppm/c	100 - 300 ppm/c
Power consumption	High	Low

Table 2 : Comparison of YIG and varactor tuning

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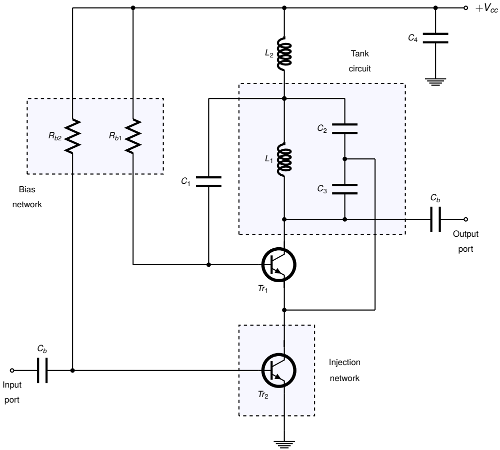
Synchronous Oscillator

Note the similarity between the synchronous oscillator in figure **??** with the Colpitts Oscillator, where L_1, C_2 and C_3 constitute the main tank circuit. An additional feedback path, C_1 is added to enhance regeneration.

- Tr_1 is the main oscillatory active device and the purpose of Tr_2 is to modulate the bias point of Tr_1 in accordance with the synchronising signal injected at the input port.
- R_{b1} and R_{b2} set the DC bias points of Tr_1 and Tr_2.
- The output signal is available from the collector of Tr_1 via the DC blocking capacitor C_b.



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