Communication Circuits Design

Academic year 2018/2019 – Semester 2 – Week 2
Lecture 2.3: Oscillators

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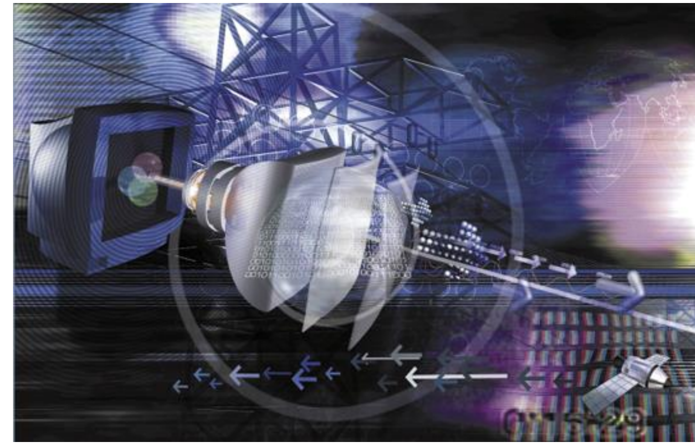
Outline



- **Criteria for oscillations**
- **Open loop design approaches**
- **Example of oscillators**
- **Modern crystal and voltage-controlled oscillators**
- **Metrics of quality and phase noise**

References:

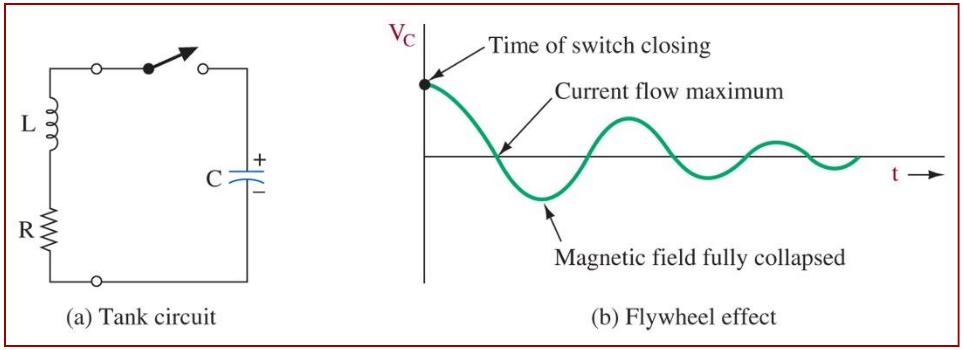
- *J. Beasley, G. Miller, "Modern Electronic Communication", Pearson, 9th ed. – Chapter 1.8*
- *B. Razavi, "RF Microelectronics", Prentice Hall, 2nd ed. – Chapter 8.7*
- *R. Sobot, "Wireless Communication Electronics", Springer, <u>at UoG Library online</u> – Chapter 8*

Simple LC oscillator

Oscillators are circuits that can generate a stable waveform in frequency and amplitude, often a controllable waveforms. They are essential in any communication circuit. Let us start with **sinusoidal oscillators**.

Let us start easy and take the circuit below. A “tank” where a capacitor has been charged to a voltage V_C and then we close the switch. At time zero there is no current and voltage is max, after a certain time the voltage is zero and there is max current, the magnetic field on the inductor will keep the current flowing to charge the capacitor in opposite polarity, hence opposite voltage.

This process of oscillation of the voltage continues until the losses dissipates the existing energy and the oscillations stop (damped oscillator). This exchange of energy between L and C is called “**flywheel effect**”.



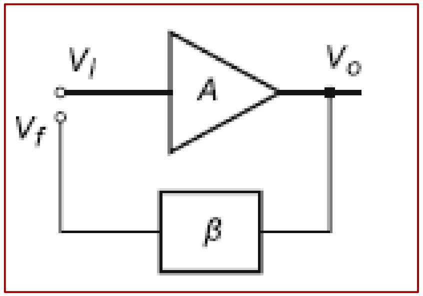
Simple LC oscillator

The process of energy conversion from L to C and vice versa happens at the resonant frequency of the LC parallel, $f_r = {}^{1}/_{2\pi\sqrt{LC}}$ which can be considered the frequency generated by the oscillator.

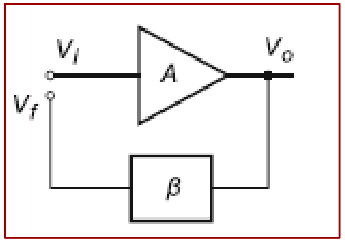
However, for the oscillator to create a stable signal at its output, we need some form of amplification.

But we have to be careful, if we amplify the signal too much, it may become unstable and uncontrolled, hence some form of feedback is needed.

So any oscillator can be intuitively modelled as a loop system with an amplifier of gain A and a feedback network with low gain β<1 to control the phase shift around the loop.



Criteria for oscillations



When the loop is open, we can derive the following:

$V_0 = AV_i$ and $V_f = \beta V_0 = \beta AV_i$

In other words any signal V_i entering the loop will feel a gain equal to Aβ – if this total gain is much larger than 1, the whole system can become too unstable when the loop is closed (signal becoming indefinitely large). To avoid this, the gain A of the amplification part has to be large enough to compensate for the losses of the feedback part β but not too large.

Control theory can offer (several) rigorous methods to solve the stability of this closed loop. We use the simplified **Barkhausen Stability Criterion** which states that for a feedback circuit to maintain oscillations:

-The net gain as to be no less than 1 $|A\beta| \geq 1$

-The net phase shift around the loop must be an integer multiple of 360° or 2π radians (N2π or Nx360°)

Criteria for oscillations

Barkhausen criterion give necessary but not sufficient conditions for maintaining oscillations. The practical conditions include:

-The initial loop gain has to be larger than 1 (so that oscillations can actually start without being damped)

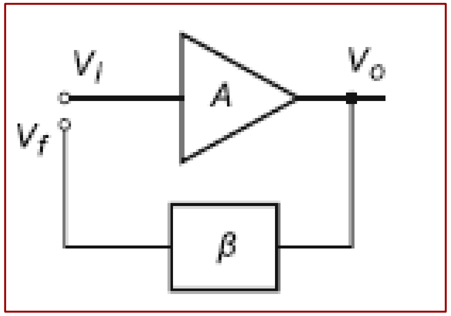
-An amplitude control mechanism must be in place to reduce the gain to avoid clipping/distortion of the oscillation

Also, note that both gains depend on frequency of the signal, A(f) and β(f)

If we close the loop and calculate the transfer function, we obtain:

$V_0 = AV_i + V_f = A(V_i + \beta V_0)$ -> $V_0/V_i = A/_{1-\beta A}$

If Aβ=1, the circuit is unstable (oscillations).



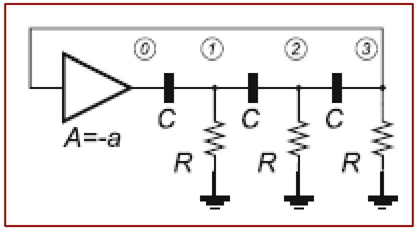
The practical implication is that any small voltage (internal noise or the DC surge from a switch) can start the oscillation.

Phase-shift oscillator

Simple example to understand Barkhausen criterion.

--An inverting amplifier of gain –a (so gain a and a phase shift of 180°)

--A cascade of CR/RC sections



If the amplifier is ideal (infinite input impedance, no current flowing into the amplifier), then the following relationships can be found. To impose that the total phase shift is N2π, the Imaginary Part has to be zero.

$$i_3 = \frac{v_3}{R} \qquad v_2 = v_3 + \frac{1}{j\omega C} i_3 = v_3 + \frac{v_3}{j\omega RC},$$

$$i_2 = \frac{v_2}{R} \qquad v_1 = v_2 + \frac{1}{j\omega C}(i_2 + i_3) = v_3 + \frac{3v_3}{j\omega RC} - \frac{v_3}{(\omega RC)^2},$$

$$i_1 = \frac{v_1}{R} \qquad v_0 = v_1 + \frac{1}{j\omega C}(i_1 + i_2 + i_3) = v_1 + \frac{(v_1 + v_2 + v_3)}{j\omega RC}$$

$$= v_3 + \frac{6v_3}{j\omega RC} - \frac{5v_3}{(\omega RC)^2} - \frac{v_3}{j(\omega RC)^3},$$

therefore,

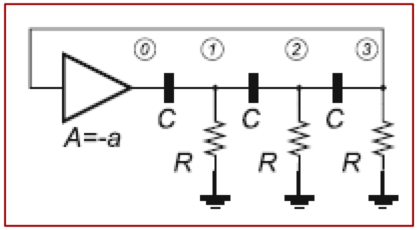
$$v_0 = \left[v_3 - \frac{5v_3}{(\omega RC)^2}\right] + j\left[\frac{v_3}{(\omega RC)^3} - \frac{6v_3}{\omega RC}\right] = \Re(v_0) + j\Im(v_0).$$

$$\frac{v_3}{(\omega RC)^3} - \frac{6v_3}{\omega RC} = 0 \qquad \therefore \qquad \omega_0 = \frac{1}{\sqrt{6}RC}$$

$$v_0 = v_3 - \frac{5v_3}{(1/\sqrt{6}RC)^2(RC)^2} \qquad \therefore \qquad \frac{v_3}{v_0} = -\frac{1}{29} = \beta.$$

Phase-shift oscillator

An interesting result. The gain of the feedback network β does not depend on the component values, is constant. To satisfy Barkhausen, the gain of the amplifier A has to be at least -29.



$$v_0 = v_3 - \frac{5v_3}{(1/\sqrt{6}RC)^2(RC)^2} \qquad \therefore \qquad \frac{v_3}{v_0} = -\frac{1}{29} = \beta.$$

Note that for real amplifiers their input impedance is not infinite, so the relationships become more complicated to solve.

Phase-shift oscillators are OK at audio frequencies, but not good at higher radio frequencies hence we need to study other models suitable at RF.

Let us first try to develop a design methodology to approach oscillators at RF frequencies.

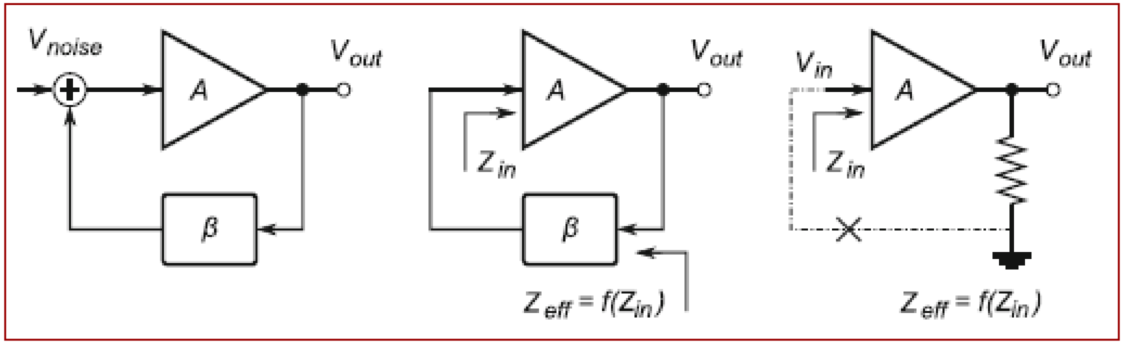
Open Loop design

So our design approach can first study the open-loop parameters, then build equivalent circuits for the close loop.

-First close the loop and assume that the internal noise can trigger the oscillation (*left*) – note that the system has only 1 port, the output of the amplifier, and there is no input

-Then consider (*centre*) that the amplifier perceives the feedback network as a load with impedance Zeff (equivalent to the input impedance of the feedback network)

-Finally consider that Zeff depends on the amplifier input impedance Zin, so in a way Zeff=f(Zin). Considering that dependency we can study the oscillator looking only at the input/output relationship at the amplifier



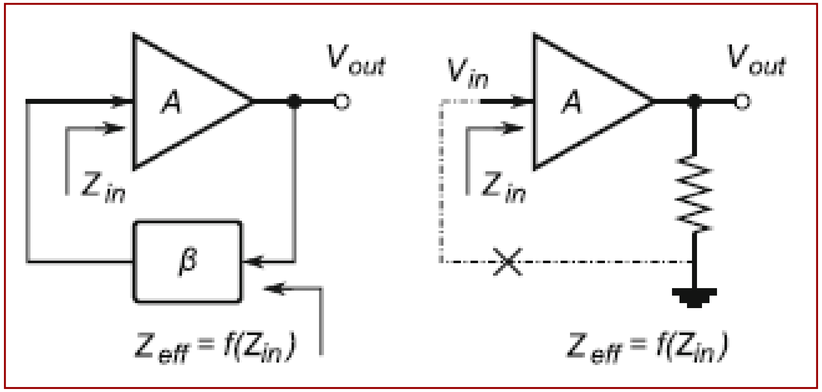
Open Loop design

So, an oscillator is made of two parts.
-An amplifier with gain A
-A feedback network

To characterise and design the feedback network of an oscillator we need to find three parameters.

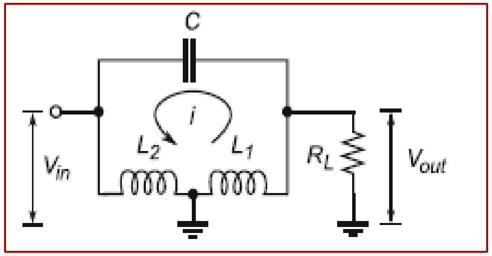
1. The gain of the feedback network **β**
2. The effective input impedance **Zeff** of the feedback network
3. The resonant (angular) frequency of the loop $\boldsymbol{\omega_0}$

Although there are infinite possibilities to design feedback networks, if we limit to using few components we can look at 4 main topologies (Sobot's book chapter 4).



Feedback networks

1) Tapped L Centre-Grounded Feedback Network



$\boldsymbol{\omega} = \sqrt{\frac{1}{(L_1+L_2)C}}$ assuming at resonance L1 and L2 in series

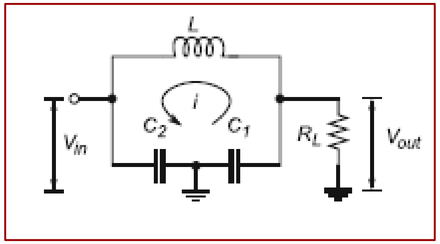
$\boldsymbol{\beta} = -\frac{L_1}{L_2}$ assuming current *i* at resonance stays within the LC loop

$\boldsymbol{R_{eff}} = R_L(\frac{L_2}{L_1})^2 // \frac{Q\omega L_2^2}{L_1+L_2}$ note that R_L (the load of the feedback network is the input impedance of the amplifier) – the calculation of this relation is complex and based on power balance so we omit it to concentrate on the final results. The idea is that the power entering the feedback network is distributed in parallel on the load R_L and on the internal loop.

The three parameters above are what is needed to characterise the feedback network of an oscillator.

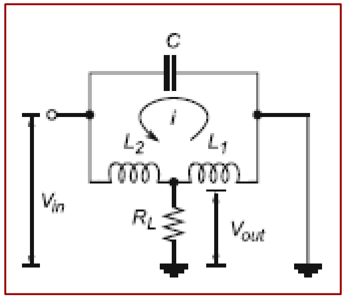
Feedback networks

2) Tapped C Centre-Grounded Feedback Network



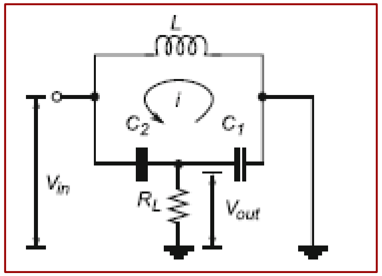
$$\boldsymbol{\omega} = \sqrt{\frac{C_1+C_2}{C_1C_2L}} \quad \boldsymbol{\beta} = -\frac{C_1}{C_2} \quad \boldsymbol{R_{eff}} = R_L\left(\frac{C_1}{C_2}\right)^2 //Q\omega L\left(\frac{C_1}{C_1+C_2}\right)^2$$

3) Tapped L Bottom-Grounded Feedback Network



$$\boldsymbol{\omega} = \sqrt{\frac{1}{(L_1+L_2)C}} \quad \boldsymbol{\beta} = \frac{L_1}{L_1+L_2} \quad \boldsymbol{R_{eff}} = R_L\left(\frac{L_2+L_1}{L_1}\right)^2 //Q\omega(L_1+L_2)$$

4) Tapped C Bottom-Grounded Feedback Network



$$\boldsymbol{\omega} = \sqrt{\frac{C_1+C_2}{C_1C_2L}} \quad \boldsymbol{\beta} = \frac{C_2}{C_1+C_2} \quad \boldsymbol{R_{eff}} = R_L\left(\frac{C_1+C_2}{C_2}\right)^2 //Q\omega L$$

Note that all these feedback networks are in a way or another an implementation of parallel LC (tank circuits)

Touch-base

Back to two key questions to understand how oscillators work.

1. If noise (with infinite frequency spectrum) starts the oscillation, why do we see only a single sinusoidal tone at the output?
2. If the loop gain is >1 how can the amplitude of the output be kept stable and finite in a real circuit?

1) **Frequency Selectivity**. The feedback networks are designed to be very selective band-pass filter (RLC components with high Q) so that only the desired (resonant) frequency can pass through the feedback and be presented at the output. We can say that the oscillator “locks” on a single tone.

2) **Amplitude Limitation**. The gain has to be >1 at the beginning to build up the oscillation, but then methods to control the amplitude are needed. The best but most complicated is *automatic gain control* (a part of the output is compared with the desired signal level and the resulting error signal is fed back to the oscillator to control its gain).

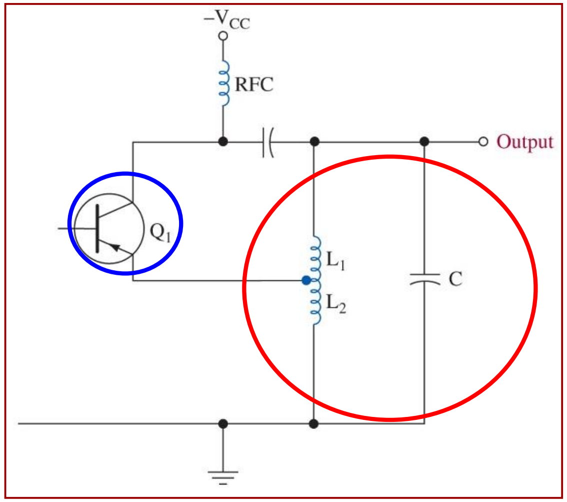
Oscillators types – Hartley

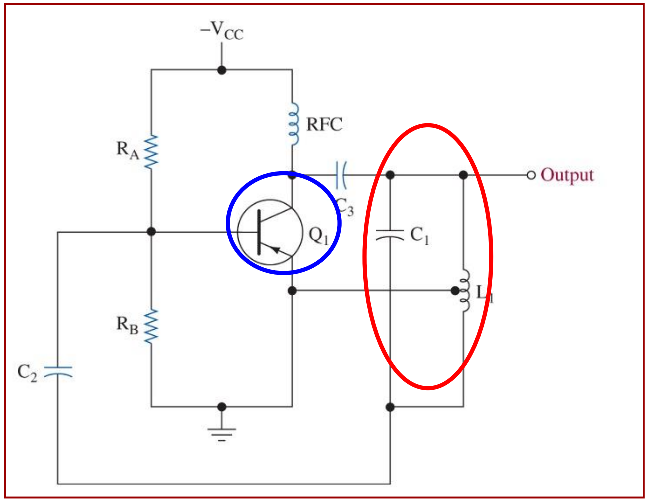
Simplified (left), practical (right) implementation

Amplifier (the **BJT** transistor in blue) –

Feedback (tapped L network made of **L1 L2** and **C** in red) hence the resonant frequency is $f \simeq \dfrac{1}{2\pi\sqrt{(L_1 + L_2)C}}$

Note that there are other components needed – **RA,RB** for amplifier bias – **C3** stops the DC from power supply to go into the tank – **C2** isolates the base of the BJT from the tank at DC – **RFC** (radio-frequency choke) is an inductor that allows DC but behaves as an open circuit at RF frequencies





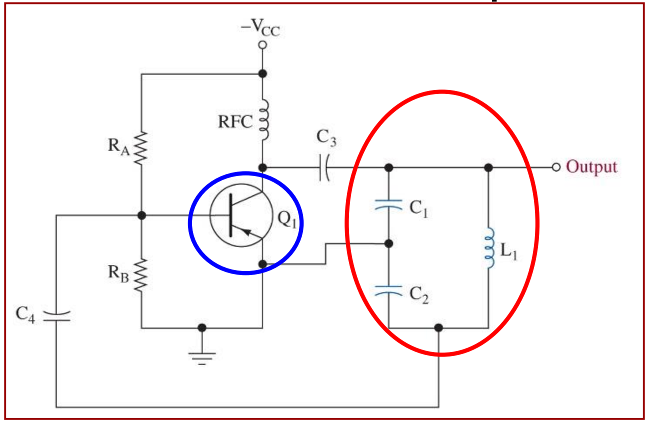
Oscillators types – Colpitts

Swaps the role of C and L in the feedback network (tank circuit)

Amplifier (the **BJT** transistor in blue) –

Feedback (tapped C network made of **C1 C2** and **L** in red) hence the resonant frequency is $f \simeq \dfrac{1}{2\pi\sqrt{[C_1C_2/(C_1 + C_2)]L_1}}$

Note that there are other components needed – **RA,RB** for amplifier bias – **C3** stops the DC from power supply to go into the tank – **C4** isolates the base of the BJT from the tank at DC – **RFC** (radio-frequency choke) is an inductor that allows DC but behaves as an open circuit at RF frequencies.



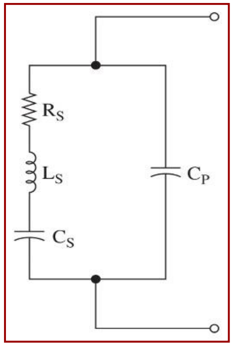
Both Hartley and Colpitts oscillators can be tuned (variable output frequency) by making either C or L in the feedback network variable.

Crystal controlled oscillators

Piezoelectric crystalline materials (quartz) have interesting properties: a mechanical deformation produces a voltage across them & a voltage across them produces mechanical deformation.

As mechanical deformation can be controlled very precisely by cutting the quartz into specific shapes/dimensions, these materials can be used to generate very stable oscillators. We call these **crystal oscillators CXO**.

The equivalent circuit of a crystal is given by an RLC series with another capacitance Cp in parallel. As the resonant frequencies of the series and parallel branches are very close, the **overall Q factor** is very high (20,000 is common vs ~1000 for top LC circuits. This means that the frequency component generated is very pure.



What about **frequency ranges**? A few kHz to tens of MHz (50-100) possible; beyond that the crystal may become too small to fabricate but higher order resonant-harmonics can be used.

Crystal controlled oscillators

More. How to measure frequency stability? We could say that CXOs typically have a frequency stability of ±0.001% - we can express that as ±10 parts per million (ppm).

Note that $0.001\% = 0.00001 = 10^{-5} = 10^{1}/10^{6}$

A simple CXOs can be further improved by adding Temperature Compensation circuitry (**TCXO**), digital microprocessor to improve the control mechanism of the oscillator (**DTCXO**), and finally an oven control (**OCXO**) which keeps the crystal at a (hot) constant temperature.

Performance increases (see below) but costs and power consumption also increase.

Table 5 Typical Performance Comparison for Crystal Oscillators

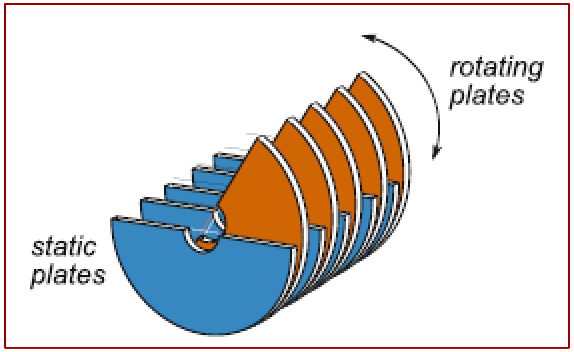
	Basic Crystal Oscillator (CXO)	Temperature Compensated (TCXO)	Digital TCXO (DTCXO)	Oven-Controlled CXO (OCXO)
Frequency stability from 0 to 70°C	100 ppm	1 ppm	0.5 ppm	0.05 ppm

Voltage controlled oscillators

In many practical communication applications we want to change the frequency of the oscillator output (tuning). This means changing the resonant frequency of the feedback network of the oscillator.

One could have a set of different circuits with fixed resonances and a switch to select the desired one, but only discrete values would be available in this way.

A better way is to use tuneable components (L or C) that can be changed mechanically in order to change their values, hence the resonant frequency. Manufacturing **tuneable C** is much easier than tuneable L, so for many years in (old) radio receivers variable capacitor were used.



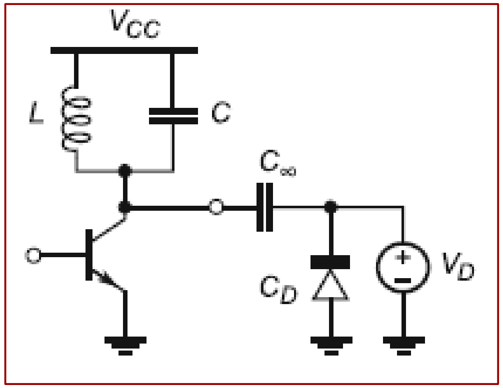
Practically, the rotation of a knob would change the overlapping area of the capacitors plates -> this changes C -> this changes resonant frequency generated

Voltage controlled oscillators

However these components are bulky, not easily miniaturised at high frequencies, hence unsuitable. In modern circuits they have been replaced by "*varicap diodes*" or *varactors*, diodes that exhibit a capacitance C_D which is a function of the voltage applied V_D.

These can be used to implement **Voltage Controlled Oscillators (VCO)**.

The simplified schematic below shows the varactors $\mathbf{C_D}$ with its biasing voltage $\mathbf{V_D}$ connected to the **LC** tank resonator through a decoupling capacitor $\mathbf{C_\infty}$ that stops the DC signal V_D to go into the tank.



At AC frequencies, V_{CC} is grounded and C_∞ is a short circuit, hence the varactor and the tank are in parallel, so the resonant frequency is

$$f_0 = \frac{1}{\sqrt{L(C + C_D(V_D))}}$$

It is practically a VCO whose resonant frequency depends on voltage V_D.

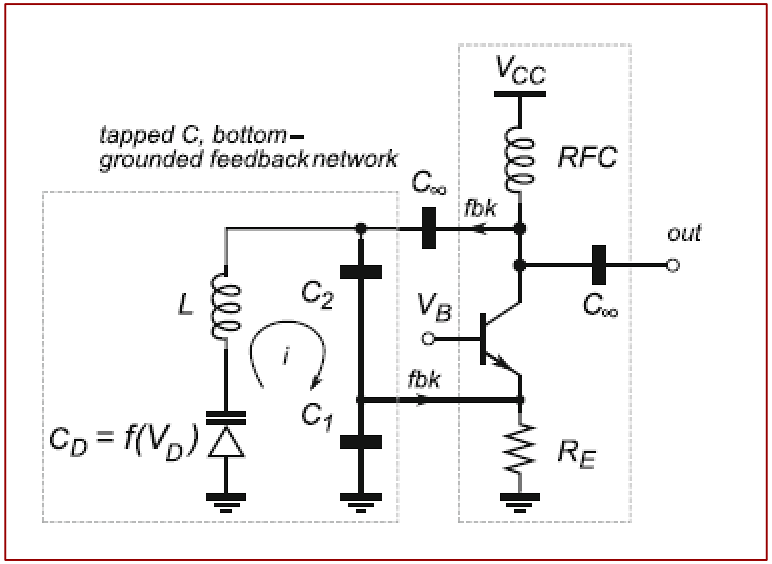
Note that the capacitance C_D of the varactor depends on its capacitance at 0V bias C_0 and applied voltage V_D

$$C_D \approx \frac{C_0}{\sqrt{1 + \frac{|V_D|}{0.5}}}$$

VCO – Clapp oscillator

A simplified diagram of an oscillator with voltage-control is shown below. The feedback network has a tapped C configuration with **C1 C2 L**, plus the varactor $\mathbf{C_D}$ in series (note that $\mathbf{V_D}$ circuit is not shown for simplicity). **RFC** provides DC connection for bias but blocks the RF signal; $\mathbf{C_\infty}$ provide DC decoupling (they stop the DC signal). The BJT in common base configuration is the amplifier.

In first approximation the total capacitance is the series of C1 and C2 $C_S = \frac{C_1 C_2}{C_1 + C_2}$.
With in series the varactor capacitance C_D so that $C = \frac{C_D C_S}{C_D + C_S}$
The final resonance angular frequency is $\omega_0 = \frac{1}{\sqrt{LC}}$.



This circuit is typically called a **Clapp** oscillator configuration (where feedback network is a tapped C, bottom-grounded). In this case we have also added voltage-control capability

Atomic and GPSDO clocks

Just to mention them.

For advanced applications there are two other classes of oscillators, **GPSDO** (GPS Disciplined Oscillators, where the signal from the GPS satellite is used to increase the performance of the crystal oscillator) and **Atomic Clocks** (where the atomic properties of elements such as Rubidium or Caesium are exploited for extremely stable oscillations).

These used to be rather expensive but recently have become more affordable (in relative terms).

At the forefront of the research, there are also **Quantum-based clocks**, that promise to deliver incredibly good performances in terms of the frequency stability of their output over long time.

Metrics for oscillators

Oscillators are always part of a larger communication system. They need to satisfy requirements in terms of their output, but also on how they interface with the other components.

Some parameters of quality (Razavi book chapter 8)

Output frequency range – This includes how well an oscillator can be tuned to generate frequencies across a more or less wide bandwidth (including variations for the temperature.

Output voltage swing – this takes into account how large is the signal generated by the oscillator with respect to its input; in a way it is a measure of efficiency of the oscillator

Power dissipation – the power drained by the oscillator and necessary for its good working

Output waveforms – typically sinusoidal, but can also be a square wave. For a sinusoidal oscillator, the oscillator can also generate undesired, discrete frequency components often called "*spurious*" components.

Phase Noise & jitter

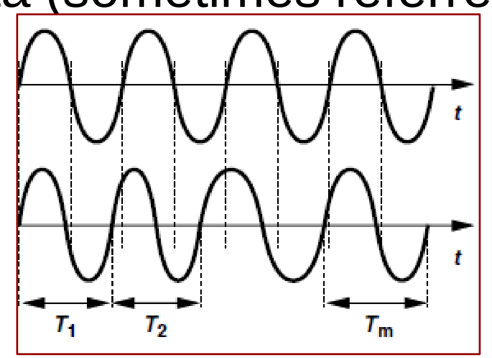
Phase noise – this accounts for how "pure" the output of an oscillator is and is related to the frequency stability (in ppm) we mentioned.

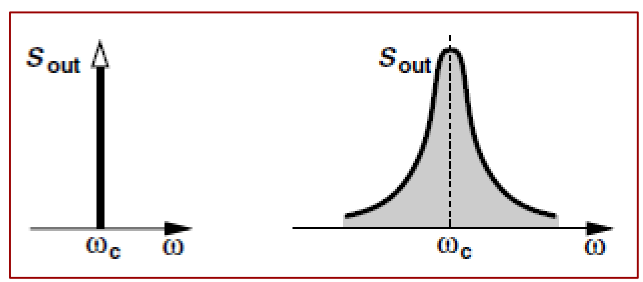
Ideally a sinusoidal oscillator should generate a signal $Acos(\omega_0 t)$

In practice, excluding the fluctuation on the amplitude A, the signal generated will look like $Acos(\omega_0 t + \varphi_n(t))$ where a **noise term** $\boldsymbol{\varphi_n(t)}$ has been added. This represents random fluctuations on the phase that perturbs the zero crossings of the waveform at each period T_0

Visually see the output for ideal (top) and real, noisy (bottom) sine waves in the time domain. The sequence of non-ideal zero-crossing is called "**jitter**".

If we look at the spectrum, assuming that the desired (angular) frequency is ω_c, then the ***effect of the phase noise*** is broadening the spectrum from the ideal delta (sometimes referred to as the "skirt").



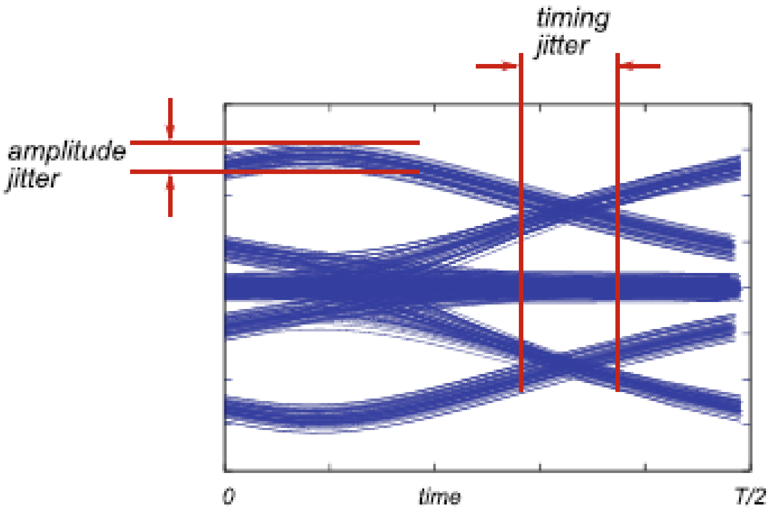


Phase Noise & jitter

How to characterise/measure phase noise or time jitter?

In the time domain you can build an **eye diagram** by cutting the signal into sections each long as half the (expected) period T, and then plot them one on top of each other (like a deck of playing cards).
If the time jitter is small, the eye will be open as all the half periods will tend to overlap; otherwise the time jitter can be measured around the cross-over point between raising and falling edges of the waveform. Commonly it is measured with respect to the period, like $t_{jit}=T/k$
Almost all modern oscilloscopes have built-in eye diagram options.

Fig. 8.16 Eye diagram of a long waveform showing time and amplitude jitter. The periodic waveform is split in sections each half a period long, i.e., $T/2$, and overlapped

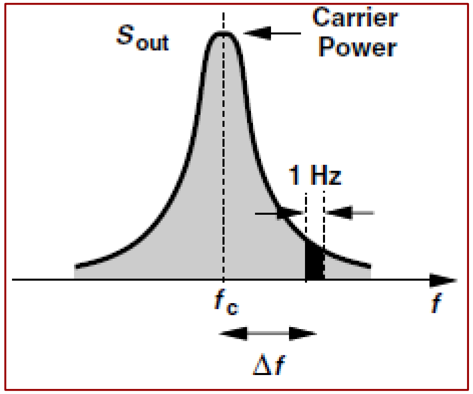


Phase Noise & jitter

How to characterise/measure phase noise or time jitter?

In the frequency domain, the phase noise can be observed on a **spectrum analyser** considering a unit (1Hz) of bandwidth at a certain offset Δf from the desired frequency (carrier), and calculating the power in such bandwidth normalised to the carrier power (peak). The phase noise is typically measured in [dBc/Hz], dB of power with respect to the carrier in 1 Hz bandwidth.

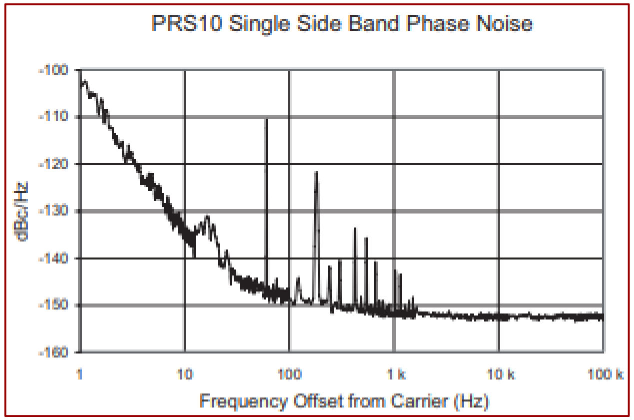
As the phase noise becomes almost flat and constant far from the carrier, good quality oscillators typically are compared on the close-to-carrier phase noise.



Noise Parameters

- Phase Noise on 10MHz RF output signal (dBm/Hz)

Offset	Typical	Max
10Hz	-118	-113
100Hz	-138	-133
1kHz	-148	-143
10kHz	-150	-145
100kHz	-150	-145
1MHz	-150	-150



Example for a OCXO (LFOCXO070939Bulk) left, and PRS10 Rubidium standard right

Summary

In this set of slides we have seen:

-Simple LC oscillators and more general criteria (Barkhausen) for sustaining oscillations

-Simple case of phase shift oscillator as a direct application of such criteria

-Open loop design of the oscillator defining three key parameters of the feedback network (**ω_0 β Reff**)

-Examples of common feedback networks with related formulae and examples of oscillators (Hartley, Colpitts)

-Crystal controlled oscillators

-Voltage controlled oscillators and the varactor

-Metrics for oscillators, in particular phase noise and its characterisation