THE CERN ACCELERATOR SCHOOL

BASICS OF RF ELECTRONICS II

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SUMMARY

➢ *A. Gallo, Basics of RF Electronics*

- Attenuators
- (signal) Amplifiers
- RF Transformers
- Power Splitters/ /Combiners
- Hybrid junctions/ /Directional Couplers
- Circulators/Isolators
- Filters
- Modulation Transfer Functions

- Frequency Mixers
- Phase Detectors (I&Q)
- Bi-phase attenuators/ /I&Q modulators
- Peak Detectors
- Step Recovery diodes
- PIN diode Switches/ /Attenuators
- Phase shifters
- VCOs
- PLLs

2nd Lecture

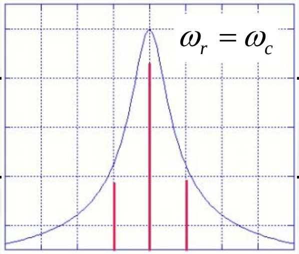
MODULATION TRANSFER FUNCTIONS

➢ *A. Gallo, Basics of RF Electronics*

LLRF servo-loops and feedback loops often need to **apply AM** and **PM modulation** to the **RF drive** signal. The response of a **resonant cavity** to AM and PM excitations depends on its **bandwidth** and **tuning** relative to the carrier:

$$v_i(t) = A_i[1 + a_i(t)]\cos(\omega_c t) \quad \longrightarrow \quad v_o(t) = A_o[1 + a_o(t)]\cos(\omega_c t)$$

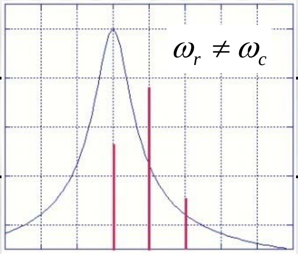
$$v_i(t) = A_i\cos[\omega_c t + \phi_i(t)] \quad \longrightarrow \quad v_o(t) = A_o\cos[\omega_c t + \Delta\phi_o + \phi_o(t)]$$



$$x(t) \overset{L-transform}{\Rightarrow} \hat{x}(s) \qquad G(s) = \frac{\hat{a}_o(s)}{\hat{a}_i(s)} = \frac{\hat{\phi}_o(s)}{\hat{\phi}_i(s)} = \frac{1}{1 + s/\sigma} \quad with \quad \sigma = \frac{\omega_r}{2Q_L}$$

$$v_i(t) = A_i[1 + a_i(t)]\cos(\omega_c t) \quad \longrightarrow \quad v_o(t) = A_o[1 + a_{o,a}(t)]\cos[\omega_c t + \Delta\phi_o + \phi_{o,a}(t)]$$

$$v_i(t) = A_i\cos[\omega_c t + \phi_i(t)] \quad \longrightarrow \quad v_o(t) = A_o[1 + a_{o,p}(t)]\cos[\omega_c t + \Delta\phi_o + \phi_{o,p}(t)]$$



$$G_{aa}(s) = \frac{\hat{a}_{o,a}(s)}{\hat{a}_i(s)}; \quad G_{pp}(s) = \frac{\hat{\phi}_{o,p}(s)}{\hat{\phi}_i(s)}; \quad G_{ap}(s) = \frac{\hat{\phi}_{o,a}(s)}{\hat{a}_i(s)}; \quad G_{pa}(s) = \frac{\hat{a}_{o,p}(s)}{\hat{\phi}_i(s)}$$

MODULATION TRANSFER FUNCTION

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It may be demonstrated that **direct** and **cross** modulation transfer functions are given by:

$$G_{pp}(s) = G_{aa}(s) = \frac{1}{2}\left[\frac{A(s+j\omega_c)}{A(j\omega_c)} + \frac{A(s-j\omega_c)}{A(-j\omega_c)}\right]; \quad G_{ap}(s) = -G_{pa}(s) = \frac{1}{2j}\left[\frac{A(s+j\omega_c)}{A(j\omega_c)} - \frac{A(s-j\omega_c)}{A(-j\omega_c)}\right]$$

with A(s) = transfer function in Laplace domain of the filter applied to the modulated signal. If the signal is filtered by a resonant cavity, one has to consider $A(s)=A_{cav}(s)$ given by:

$$A_{cav}(s) = A_0 \frac{2\sigma s}{s^2 + 2\sigma s + \omega_r^2} \quad with \quad \omega_r \approx \omega_c + \sigma \tan\phi_z$$

where ϕ_z is the ***cavity tuning angle***, i.e. the phase of the cavity transfer function at the carrier frequency ω_c. Finally one gets:

$$G_{pp}(s) = G_{aa}(s) = \frac{\sigma s + \sigma^2\left(1+\tan^2\phi_z\right)}{s^2 + 2\sigma s + \sigma^2\left(1+\tan^2\phi_z\right)}; \quad G_{ap}(s) = -G_{pa}(s) = -\frac{\sigma \tan\phi_z \, s}{s^2 + 2\sigma s + \sigma^2\left(1+\tan^2\phi_z\right)}$$

MODULATION TRANSFER FUNCTION

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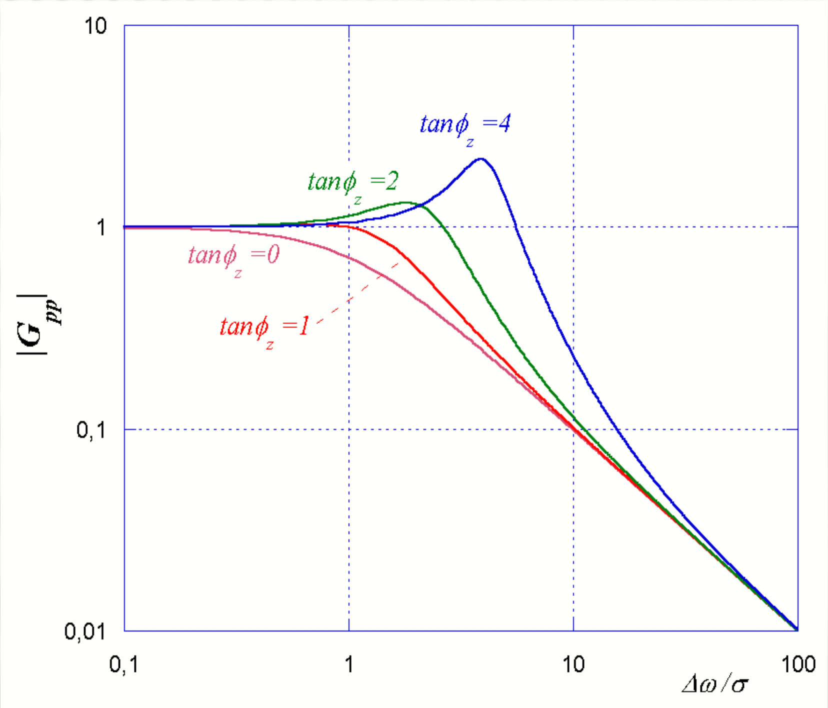
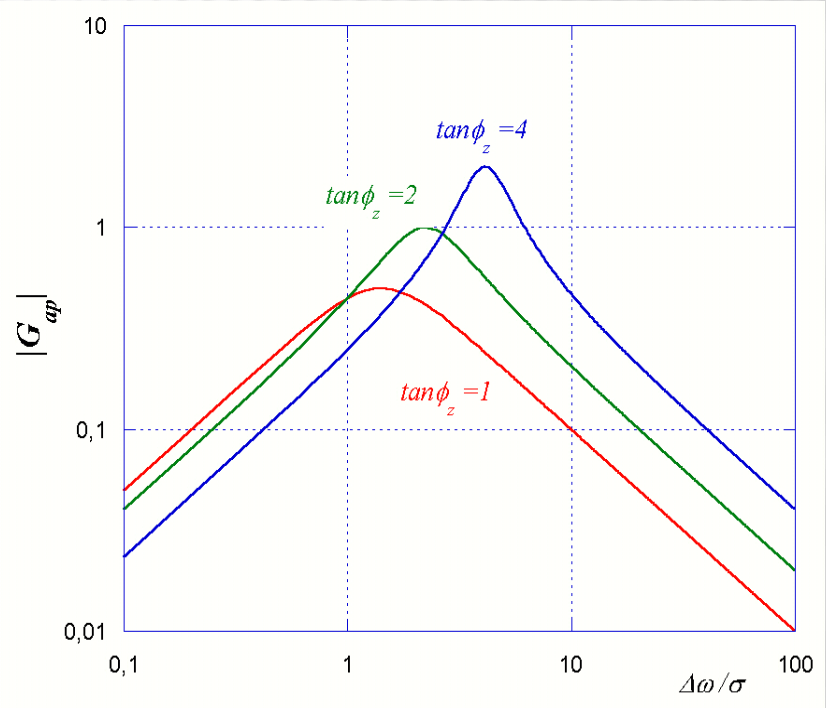
The general form of the modulation transfer functions features **2 poles** (possibly a complex conjugate pair) and **1 zero**, and degenerates to a **single pole LPF** response if the cavity is **perfectly tuned** (cross modulation terms vanish in this case).





MODULATION TRANSFER FUNCTION: THE PEDERSEN MODEL

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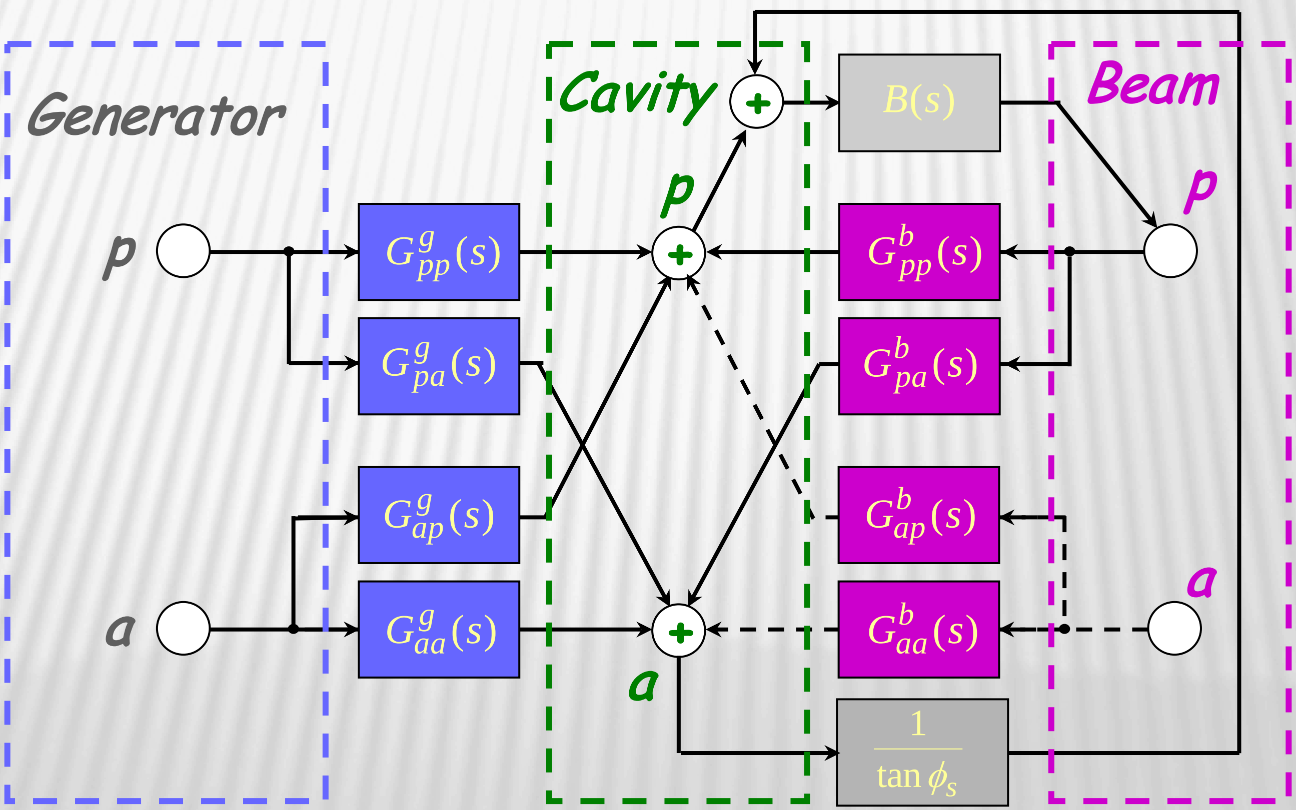
In **circular accelerators** the **beam phase** depends on the **cavity RF phase** through the **beam transfer function**, while the **cavity RF amplitude** and **phase** depend on the **beam phase** through the **beam loading mechanism**. The whole **generator-cavity-beam linear system** can be graphically represented in a **diagram** called **Pedersen Model**.
The modulation transfer functions vary with the stored current and definitely couple the servo-loops and the beam loops implemented around the system.



FREQUENCY MIXERS: f- TRANSLATION

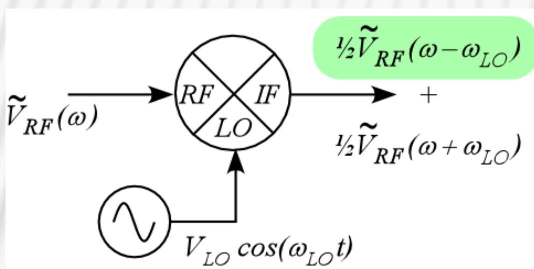
➢ *A. Gallo, Basics of RF Electronics*

Frequency mixers are **non-linear**, (generally) **passive** devices used in a **huge variety** of **RF applications**. Basically, a mixer is used to perform the **frequency translation** of the spectrum of an RF signal to be manipulated. The spectrum shift is obtained from an **analog multiplication** between the **RF signal** and a **Local Oscillator** (LO).

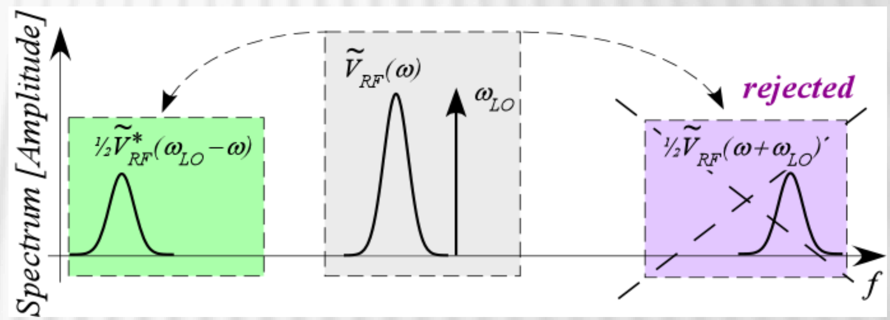
$$V_{IF}(t)=kV_{RF}(t)V_{LO}\cos(\omega_{LO}t)\;;\quad \tilde{V}_{RF}(\omega)=\tilde{F}[V_{RF}(t)]=\frac{1}{\sqrt{2\pi}}\int_{-\infty}^{+\infty}V_{RF}(t)e^{-j\omega t}\,dt$$

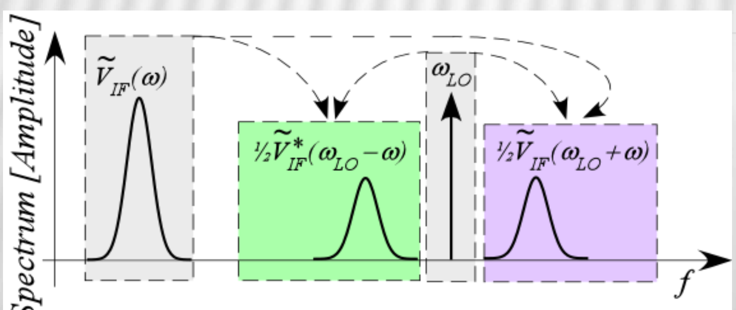
$$=\tilde{V}^{*}_{RF}(\omega_{LO}-\omega)$$

$$\tilde{V}_{IF}(\omega)=\frac{kV_{LO}}{2}\frac{1}{\sqrt{2\pi}}\int_{-\infty}^{+\infty}V_{RF}(t)e^{-j\omega t}\left(e^{j\omega_{LO}t}+e^{-j\omega_{LO}t}\right)dt=\frac{kV_{LO}}{2}\left[\tilde{V}_{RF}(\omega-\omega_{LO})+\tilde{V}_{RF}(\omega+\omega_{LO})\right]$$

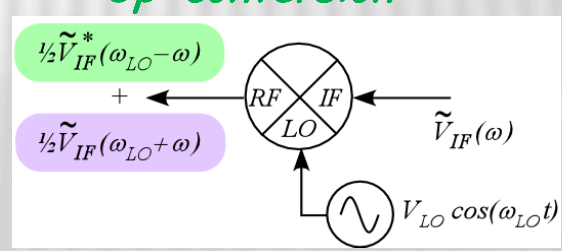


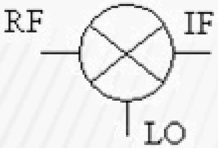
Down-conversion





Up-conversion





FREQUENCY MIXERS

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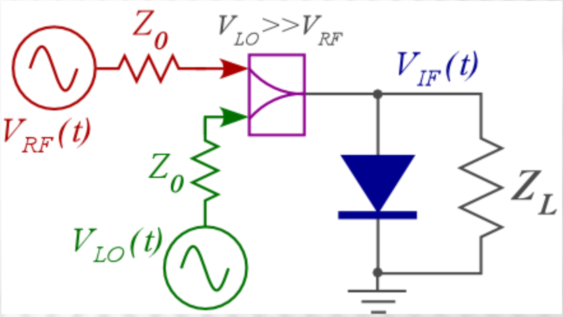
In principle **any non linear device** could produce the desired **frequency translation**. If the mixing RF and LO signals are fed into a single diode, under the assumption $V_{LO} >> V_{RF}$, the LO voltage turns the diode ON and OFF and the IF voltage is:



$$V_{IF}(t) = k\left[V_{LO}(t) + V_{RF}(t)\right] \cdot \left[1 - \mathrm{sgn}\left(V_{LO}(t)\right)\right]$$

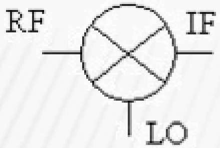
Being $V_{LO}(t)$ a sine wave, the function $1\text{-}sgn[V_{LO}(t)]$ is a **square wave** expressing the on-off modulation of the diode according to the polarity of the LO voltage. The **square wave** contains all the **odd harmonics** of f_{LO}, so that each frequency f_{RF} contained in the RF signal produces the output frequencies f_{IF}:

$$f_{IF} = n\, f_{LO} \pm f_{RF} \qquad n = any\ odd\ integer$$

Due to the frequency content of the square wave, the real mixer produces **many frequencies lines other than** the $f_{LO} \pm f_{RF}$ ones. These are called "**spurious intermodulation products**".

Actually, real diodes are not ideal switches and on-off commutations are smooth. This effect produces **more intermodulation products**, so that the frequencies present in the output spectrum are:

$$f_{IF} = n\, f_{LO} \pm m f_{RF} \qquad n, m = any\ integers$$



BALANCED MIXERS

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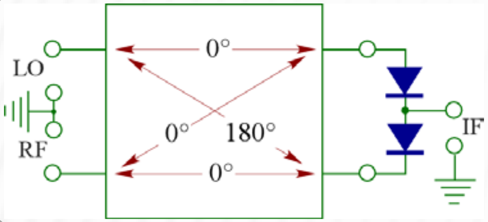
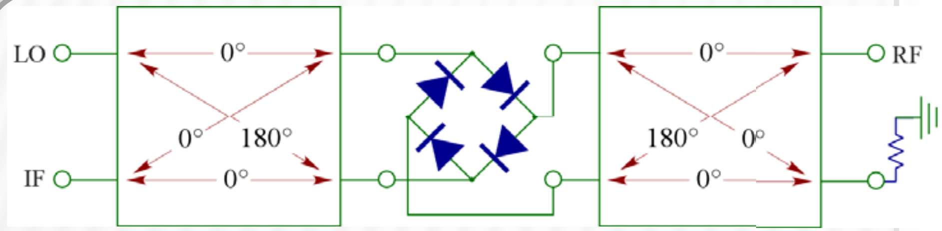
Single diode mixing provides **no inherent isolation** between ports. **Lack of isolation** results in a **large number** of **intermodulation products**, **poor conversion loss** (i.e. a large value of the ratio between the power of unconverted and converted signals) and various **interference** and **cross-talk** problems.

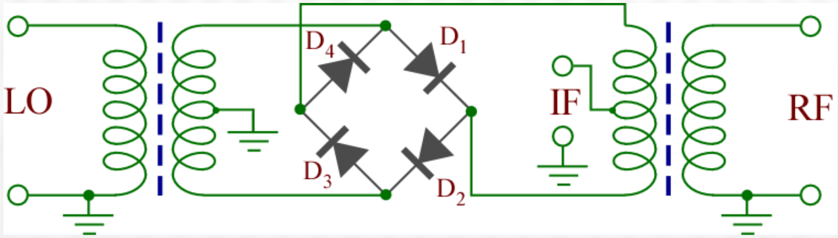
Port isolation is obtained by **exploiting symmetries** in the mixing network design.

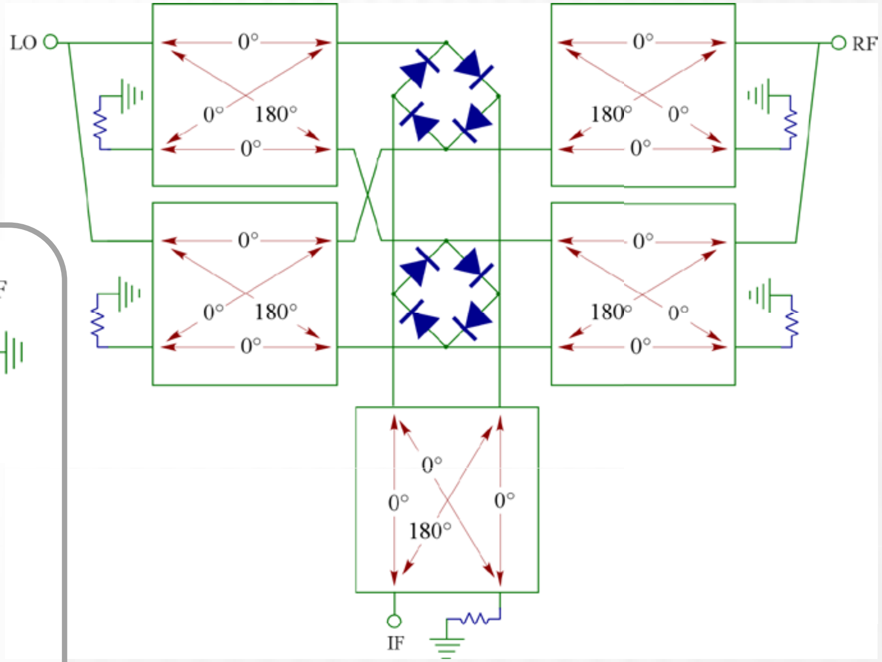


Single Balanced Mixer

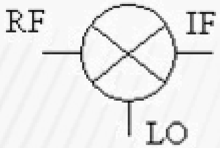


Double Balanced Mixers





Triple Balanced Mixer



DOUBLE BALANCED MIXERS

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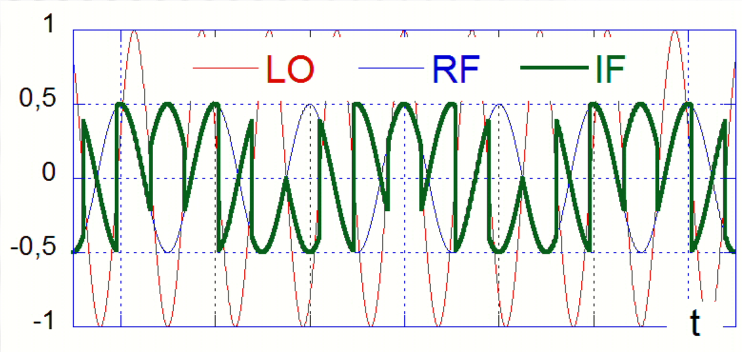
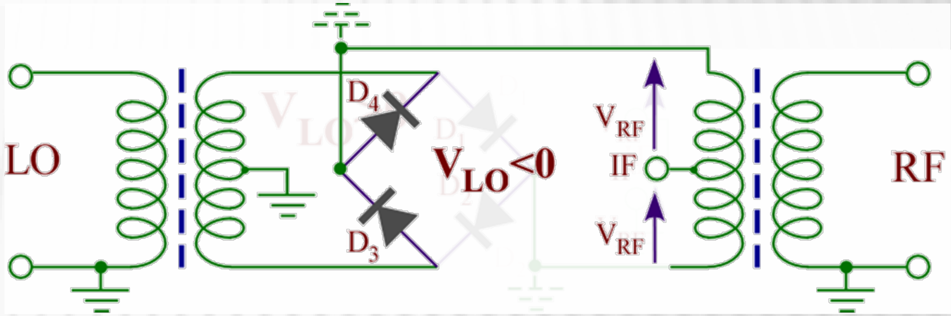
The **Double Balanced Mixer** is the **most diffused** type of frequency mixers, ensuring **good isolation** and **excellent conversion loss**. The LO voltage is differentially applied on the diode bridge switching on/off alternatively the D_1-D_2 and D_3-D_4 pairs, so that the IF voltage is given by:

$$V_{IF}(t) = V_{RF}(t) \cdot \mathrm{sgn}\left[V_{LO}(t)\right]$$





$$V_{RF}(t) = V_{RF} \cdot \cos\left(\omega_{RF} t\right); \quad V_{LO}(t) = V_{LO} \cdot \cos\left(\omega_{LO} t\right)$$

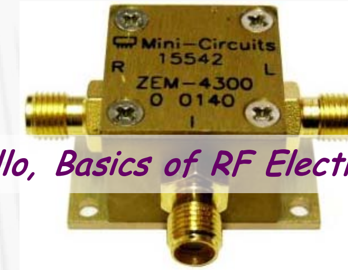
$$V_{RF} << V_{LO}$$

$$V_{IF}(t) = V_{RF} \cos(\omega_{RF} t) \cdot \mathrm{sgn}\left[\cos(\omega_{LO} t)\right] = V_{RF} \cos(\omega_{RF} t) \cdot \sum_{n=odds} \frac{4}{n\pi} \cos(n\omega_{LO} t) =$$

$$= V_{RF} \cdot \sum_{n=odds} \frac{2}{n\pi} \left[\cos((n\omega_{LO} - \omega_{RF})t) + \cos((n\omega_{LO} + \omega_{RF})t)\right] =$$

$$= \frac{2}{\pi} V_{RF} \left[\cos((\omega_{LO} - \omega_{RF})t) + \cos((\omega_{LO} + \omega_{RF})t) + intermod\ \ products\ \right]$$

DOUBLE BALANCED MIXERS: BASIC SPECFICATIONS



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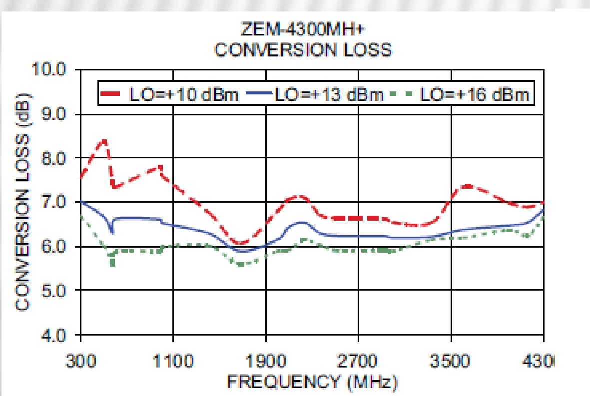
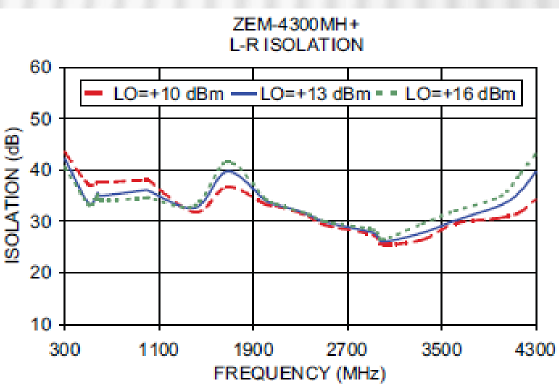
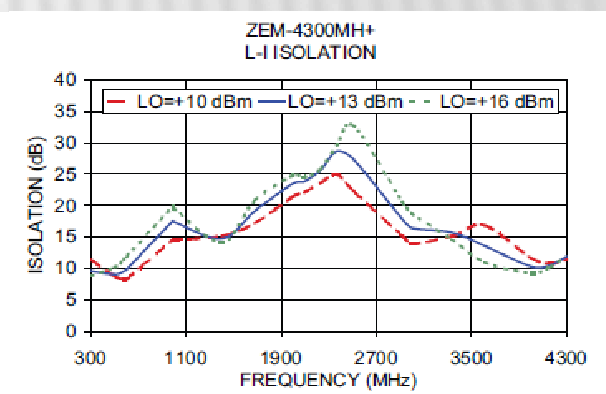
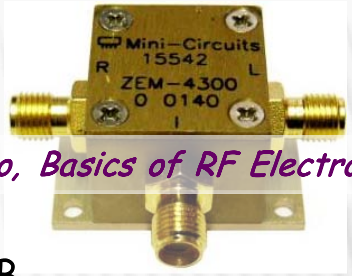
- **Frequency range (specific to each port)**
  from DC to > 10 GHz, multi-decades covered by a single device.
  normally IF band is < of RF, LO bands. IF may be DC or AC coupled.
- **Mixer level**;
  minimum level at LO to switch on/off the diodes. Typically +3 ÷ +23 dBm, depending on the diode barrier and the number of diodes in series in the bridge.
- **Conversion Loss**;
  ratio between the unconverted (RF) and converted (Single Sideband IF) signal levels . Theoretical minimum = 3.9 dB ($=20 \cdot Log(2/\pi)$); practical values in the 4.5 ÷ 9 dB range.
  Is an "integral" spec. Low CL means also good isolation (not necessarily vice-versa)
- **Isolation**;
  amount of direct signal leakage from one to another ports (reciprocal parameter). L-R critical for interference in the RF circuitry. Typical values 25 ÷ 35 dB.
  L-I critical for filtering when f_{IF} and f_{LO} are close. Typical values 20 ÷ 30 dB.
  R-I is usually not an issue ($P_{RF} \ll P_{LO}$).Typical values 25 ÷ 35 dB.





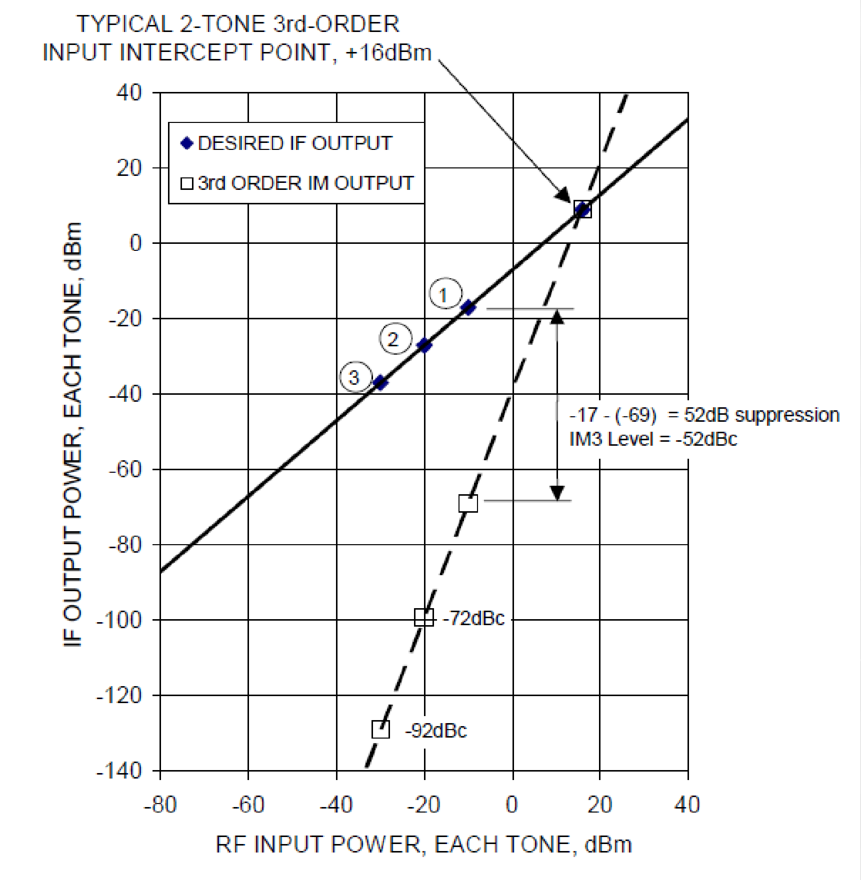


DOUBLE BALANCED MIXERS: BASIC SPECFICATIONS



➢ *A. Gallo, Basics of RF Electronics*

- **1 dB Compression**:
  is a figure of the mixer linearity. defined as the RF level showing a 1 dB increase of the conversion loss. typical values ≈4 dB below mixer specified LO level
- **Noise Figure**:
  $NF_{mix} \geq CL$, $NF_{mix} \approx CL$ (signal reduced, white noise unaffected by up/down conversion);
  3 dB worse for SSB wrt DSB signals (signals add coherently, noise quadratically);
  mixer + IF amps cascades have noise figures $NF = NF_{mix} + CL \cdot (NF_{IF} - 1)$. Being magnified by the mixer conversion loss the IF amp noise figure is crucial.
- **Single and Multi-tone Intermodulation Distortion / 2 tones 3rd order intercept**:
  output content of harmonics other than $f_{LO} \pm f_{RF}$. mixer non-linearity allows multitone RF signals (f_{RF1}, f_{RF2}, …) generating output harmonics at $m_1 f_{RF1} + m_2 f_{RF2}$. level 3rd order harmonics $2f_{RF1} - f_{RF2}$, $2f_{RF2} - f_{RF1}$ grows with 3rd power of RF signal, while fundamental $f_{LO} \pm f_{RF}$ tone level grows linearly. 2-tones 3rd order intercept defined as the RF level where the two output lines cross.

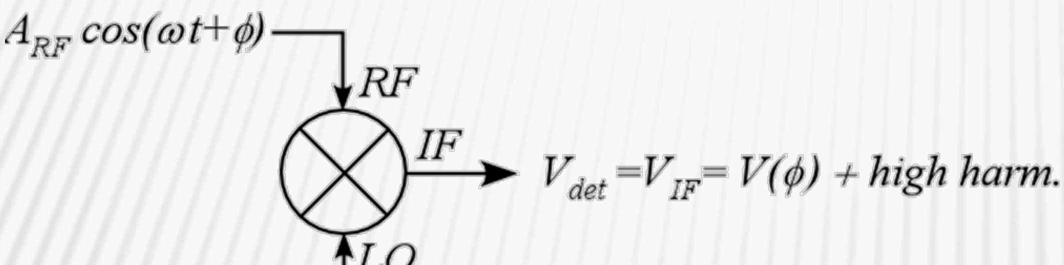


DOUBLE BALANCED MIXERS: PHASE DETECTORS

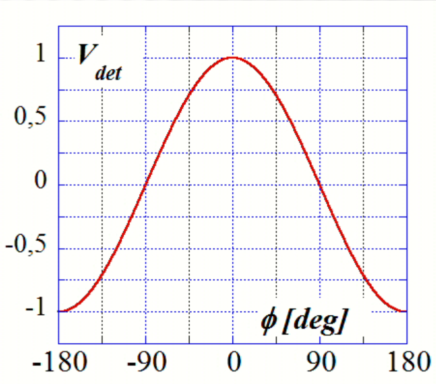
➢ *A. Gallo, Basics of RF Electronics*

If $f_{LO}=f_{RF}$ the IF signal has a DC component given by:

$$V_{IF}(t) = A_{RF}\cos(\omega t+\phi)\cdot \mathrm{sgn}\left[A_{LO}\cos(\omega t)\right] \Rightarrow V_{IF}\big|_{DC} = \langle V_{IF}(t)\rangle = k_{CL}A_{RF}\cos\phi$$



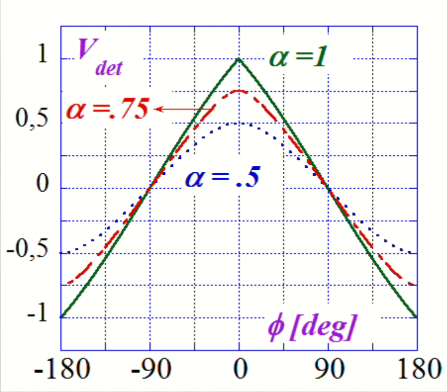
$$A_{RF} \ll A_{LO} \quad \Rightarrow \quad V_{det}(\phi) = k_{CL}A_{RF}\cos\phi$$



Mixers dedicated to phase detection can be **operated in saturation**, i.e. with **similar levels** at **both RF** and **LO** inputs. The diodes are turned on/off by either LO or RF signals, and a more linear phase detection results according to:

$$A_{RF} \approx A_{LO} \quad \Rightarrow \quad V_{IF}(t) = \begin{cases} V_{RF}(t)\cdot \mathrm{sgn}\left[V_{LO}(t)\right] & \textit{if} \quad |V_{RF}(t)| < |V_{LO}(t)| \\ V_{LO}(t)\cdot \mathrm{sgn}\left[V_{RF}(t)\right] & \textit{if} \quad |V_{RF}(t)| > |V_{LO}(t)| \end{cases}$$

$$V_{det}(\phi) = k_{CL}A_{RF}\frac{\sqrt{1+\alpha^2+2\alpha\cos\phi}-\sqrt{1+\alpha^2-2\alpha\cos\phi}}{2\alpha} \quad \textit{with } \alpha = \frac{A_{RF}}{A_{LO}}$$

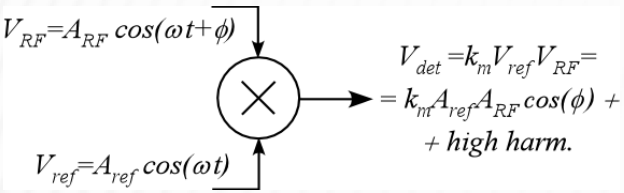


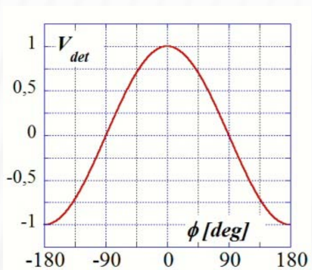
OTHER PHASE DETECTORS

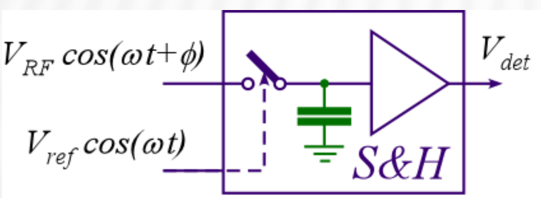
➢ *A. Gallo, Basics of RF Electronics*

The same operation can be accomplished by **analog multiplier circuits** or **digital comparators** (ex-OR circuits) with **larger sensitivity** and output **dynamic range**, **better linearity**, but far much **smaller bandwidths** (typically ≤ 500 MHz).

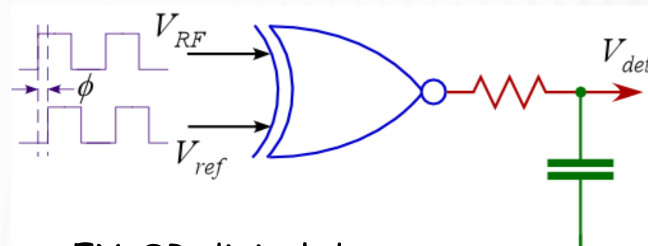
Analog multiplier detector



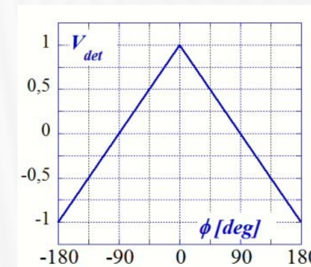




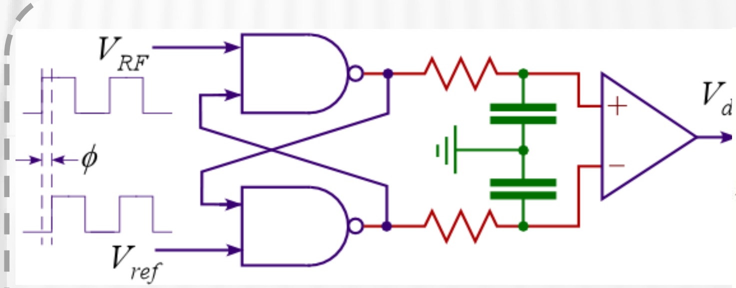
Sample & Hold detector

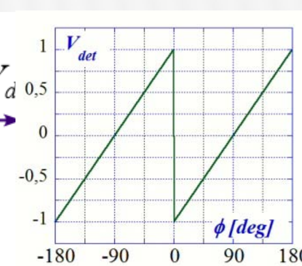


EX-OR digital detector

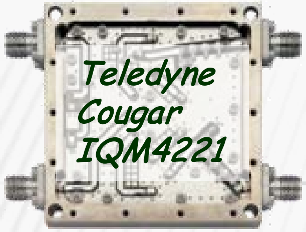


Digital detectors are more linear, but their use is mainly limited at already down-converted signals.





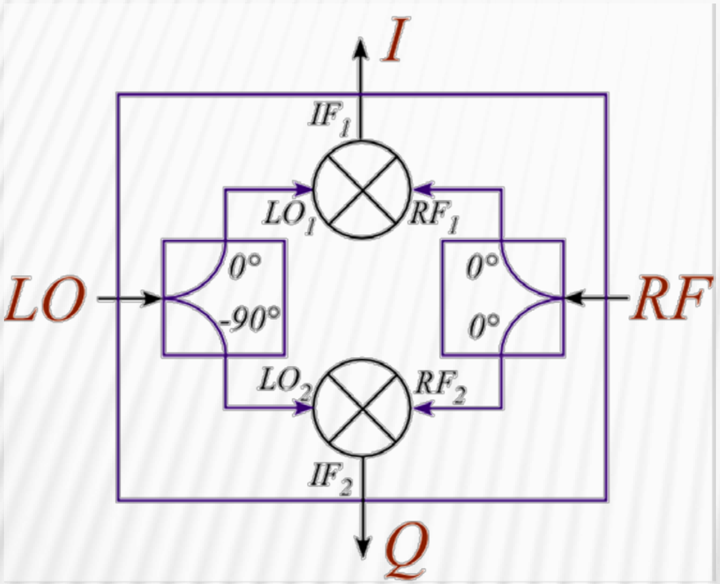
Flip-flop digital detector



DOUBLE BALANCED MIXERS: I&Q DETECTORS

➢ *A. Gallo, Basics of RF Electronics*

A circuit made of 2 mixers, 1 splitter and 1 quadrature hybrid allows extracting **in-phase** and **in-quadrature components** of the RF signal.



$$V_I(t) = k_{CL} A_{RF} \cos\left[(\omega_{LO} \pm \omega_{RF})t\right] = A_{IF} \cos(\omega_{IF} t)$$

$$V_Q(t) = k_{CL} A_{RF} \sin\left[(\omega_{LO} \pm \omega_{RF})t\right] = A_{IF} \sin(\omega_{IF} t)$$

if $\omega_{LO} = \omega_{RF}$

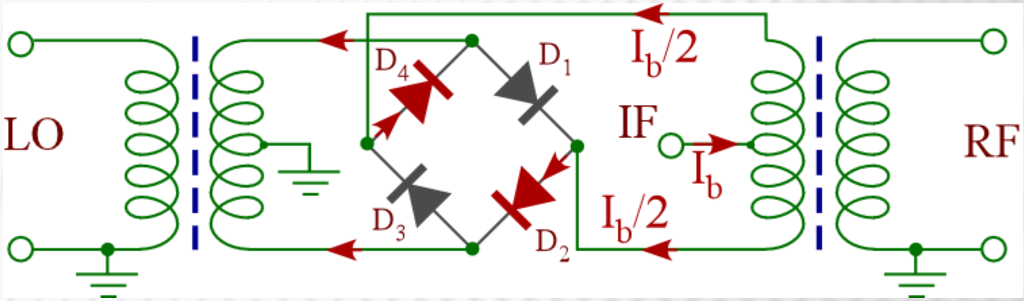
$$\begin{cases} V_I = k_{CL} A_{RF} \cos(\phi) + high\ harmonics \\ V_Q = k_{CL} A_{RF} \sin(\phi) + high\ harmonics \end{cases} \Rightarrow \begin{cases} A_{RF} \div \sqrt{V_I^2 + V_Q^2} \\ \phi = \arctan\left(V_Q / V_I\right) + \dfrac{\pi}{2}\left[1 - \mathrm{sgn}(V_I)\right] \end{cases}$$

DOUBLE BALANCED MIXERS: BI-PHASE AMPLITUDE MODULATORS

➢ *A. Gallo, Basics of RF Electronics*

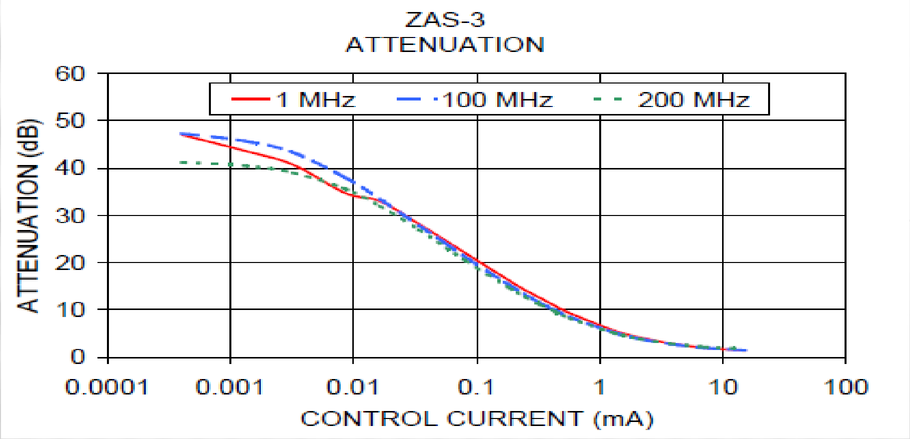
A **bi-phase amplitude modulator** can be obtained by **lowering** the **L-R isolation** of a double balanced mixer in a controlled way by injecting a **bias current** in the **IF port**.

Positive and negative bias IF currents I_b increase the transconductance of the diode pairs D_2-D_4 and D_1-D_3, respectively.



The bias current I_b controls the **value** and the **sign** of the L-R coupling coefficient ***k***.

$$V_{RF}(t) = V_{LO}(t) \cdot \left[\frac{Z_{D_3}}{Z_{D_3} + Z_{D_4}} - \frac{Z_{D_2}}{Z_{D_1} + Z_{D_2}} \right] = k\left(I_b\right) \cdot V_{LO}(t)$$



Characteristics:

- Passive
- Non-linear
- Bi-phase
- Broadband control port (IF port bandwidth: DC÷1 GHz)

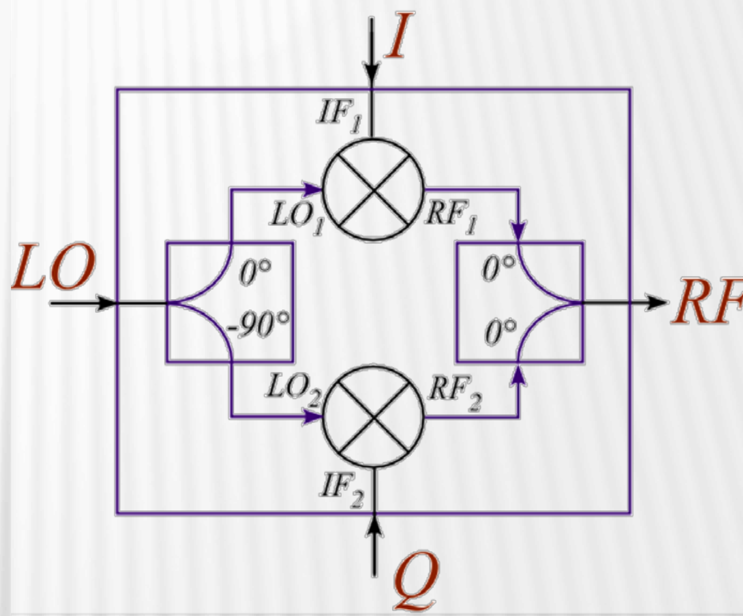
DOUBLE BALANCED MIXERS: I&Q MODULATORS

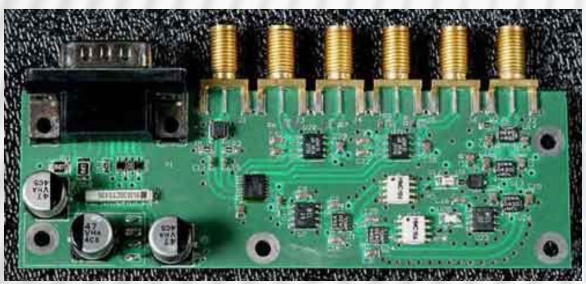
➢ *A. Gallo, Basics of RF Electronics*

Using mixers as bi-phase controlled attenuators, **vector (I&Q) modulators** can be obtained very similarly to I&Q detectors. I and Q copies of the input signal are obtained from a quadrature hybrid and re-combined by a vector combiner after being individually attenuated with independent control signals.

$$V_{RF}(t) = A_{LO}\left[k_I \cdot \cos(\omega_{LO}t) + k_Q \cdot \sin(\omega_{LO}t)\right] =$$

$$= A_{LO}\sqrt{k_I^2 + k_Q^2} \cdot \cos\left[\omega_{LO}t + \arctan\left(k_Q / k_I\right)\right]$$





Analog I/Q modulator

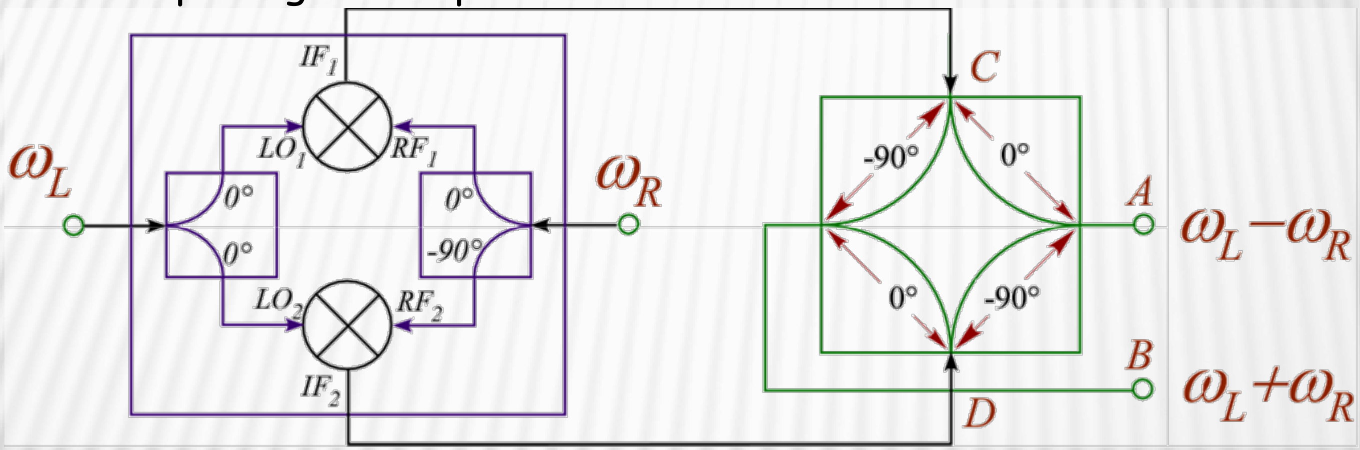
syndetix incorporated

The level of the output signal is controlled by moving k_I and k_Q proportionally, while unbalanced changes produce variations of the output signal phase.

DOUBLE BALANCED MIXERS: IMAGE REJECTION

➢ A. Gallo, Basics of RF Electronics

Filtering mixer image frequencies can be difficult and/or costly, especially in up-conversion processes where the image frequency bands are relatively closely spaced. By **cascading** an **I&Q mixer** and a **quadrature hybrid** an "**image rejection**" network is obtained, where the 2 image signals are **separately available** at **2** physically **different ports**. No narrowband filtering is then necessary to separate the 2 output signal components.



this config. only works if $\omega_L > \omega_R$

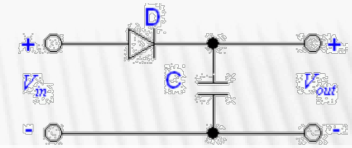
$$V_{IF_1}(t) = k_{CL} A_{RF} \left[\cos\left((\omega_L + \omega_R)t\right) + \cos\left((\omega_L - \omega_R)t\right)\right]$$

$$V_{IF_2}(t) = k_{CL} A_{RF} \left[\sin\left((\omega_L + \omega_R)t\right) - \sin\left((\omega_L - \omega_R)t\right)\right]$$

$$\Downarrow$$

$$V_A(t) = k\, A_{RF} \cos\left((\omega_L - \omega_R)t\right) \; ; \; V_B(t) = k\, A_{RF} \sin\left((\omega_L - \omega_R)t\right)$$

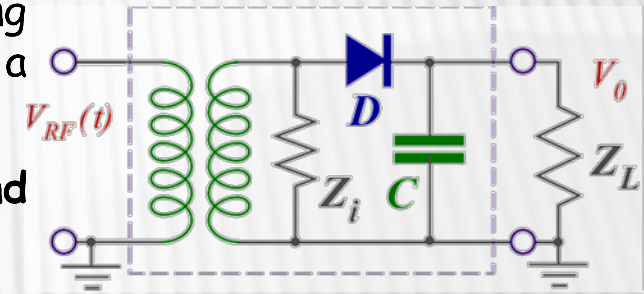
PEAK DETECTORS

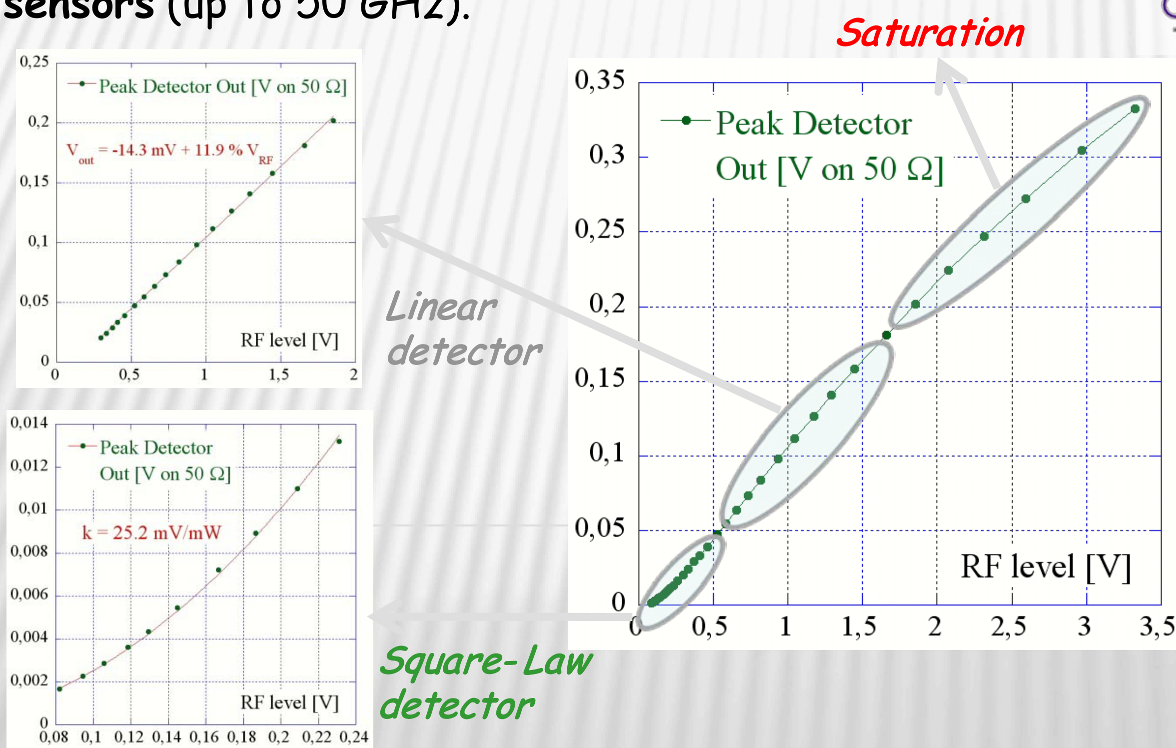


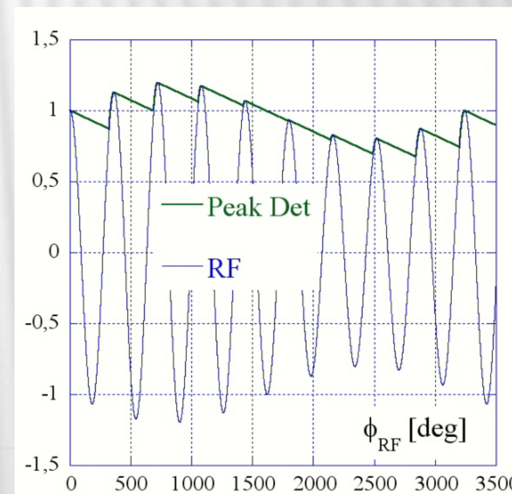
➢ *A. Gallo, Basics of RF Electronics*

Diode **peak detectors** are used to **sample** the amplitude of RF signals. They basically work as rectifiers, sampling the RF peak while charging a capacitance, and holding the peak voltage slowly discharging the capacitance on a load (typically 50 Ω to follow fast level variations).

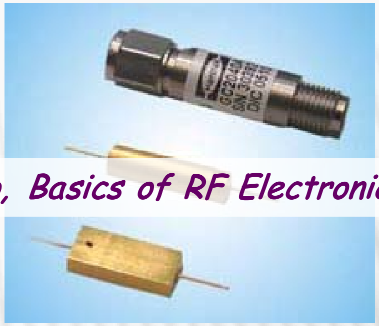
Schottky diodes are used for zero-bias, very **broadband sensors** (up to 50 GHz).







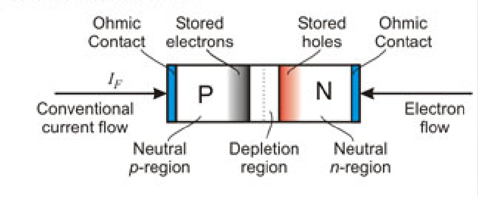
STEP RECOVERY DIODES (SRDS) COMB GENERATORS



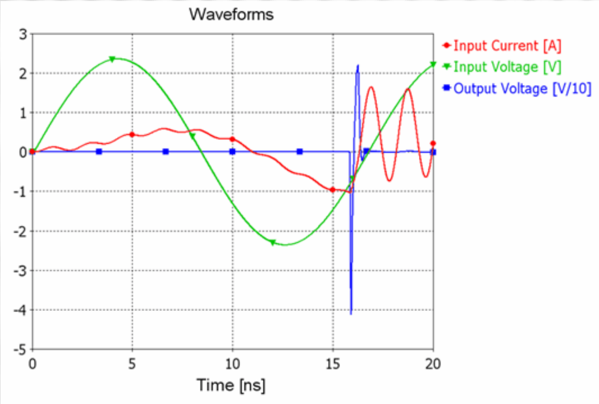
➢ *A. Gallo, Basics of RF Electronics*

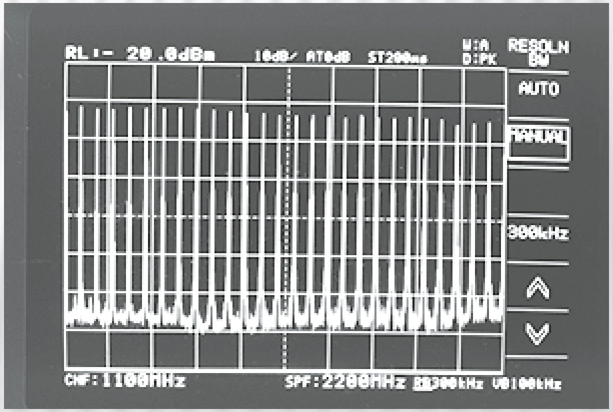
Diodes switching from forward to reverse polarization deliver in a certain time all the charge stored on both sides of the space-charge/depletion region. **SRDs** have a **PIN-like** structure with a **special doping profile** allowing the reverse current to circulate across the depletion region for a short time before **abruptly dropping to zero in few tens of ps**.

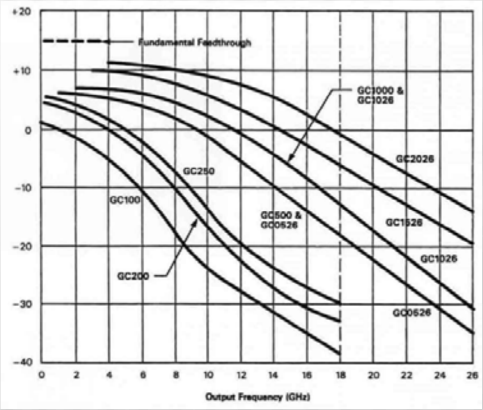
The sharp variation of the circuit current can be used to generate **short voltage pulses** on a load. The spectrum of the output signal is a **series of peaks** containing **all the harmonics** of the **input sine-wave** and may extends well beyond *10 GHz* (comb generator).



SRDs are used to generate very short pulses (time domain) or to extract any required harmonics of the input signal by properly filtering the output voltage (frequency multiplier).





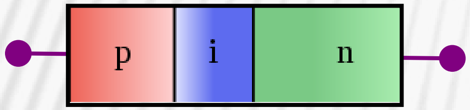


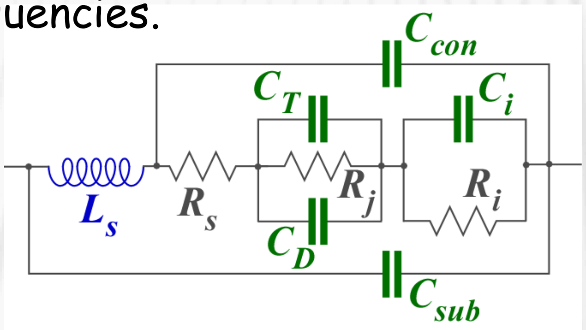
PIN DIODES
VARIABLE ATTENUATOR/SWITCH

➢ *A. Gallo, Basics of RF Electronics*

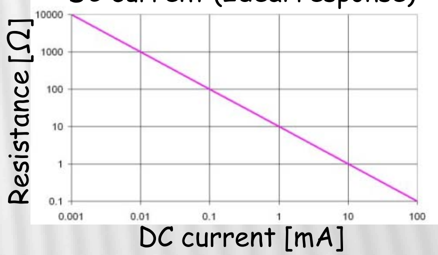
Dynamic impedance of biased **PIN diodes** offered to RF signals is **inversely proportional** to the **bias current**. This makes PIN diodes suitable **controlled resistors** to be put in T or Π configuration of resistive attenuators.

PIN technology is used since the **intrinsic layer resistance remains dominant** in forward bias, while in standard PN diodes the junction diffusion capacitance shorts the device at high frequencies.

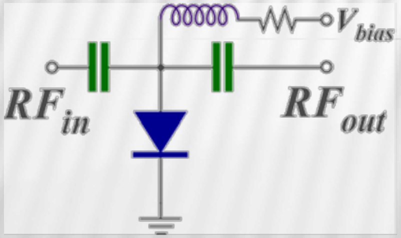


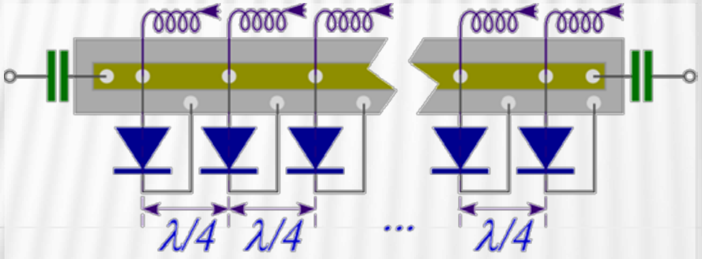


Resistance of a PIN diode versus DC current (Ideal response)



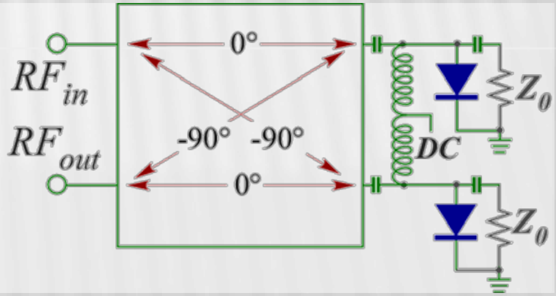
Single cell attenuators are mostly reflective





λ/4 cell cascades improve both dynamic range and matching. Best results obtained with unequal diode bias.

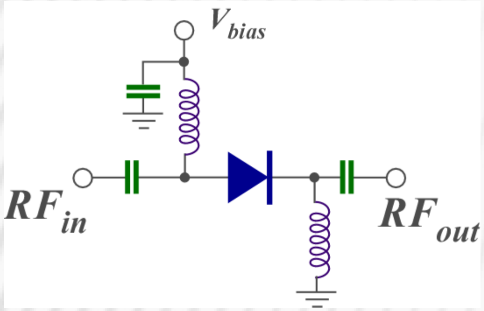
A quadrature hybrid with balanced mismatch offers low reflectivity.



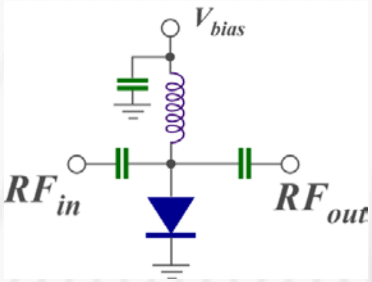
PIN DIODES
VARIABLE ATTENUATOR/SWITCH

➢ *A. Gallo, Basics of RF Electronics*

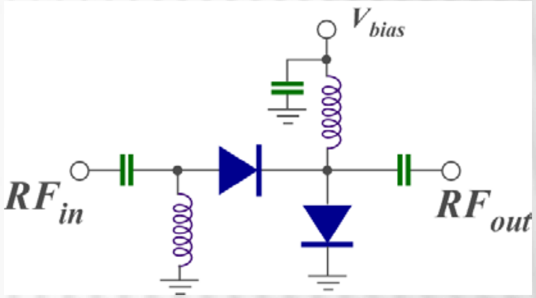
In the **extreme bias conditions** the diodes are on/off (=open/short) so that an RF signal can be **fully transmitted** or **fully stopped** and the device acts as a controlled **RF switch**.



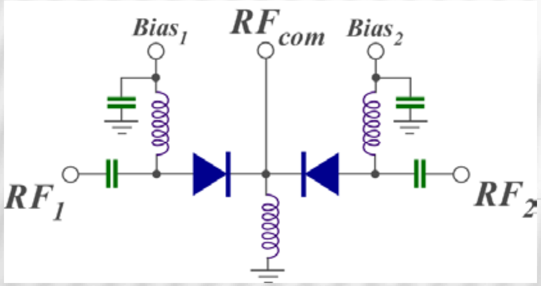
Single-Pole Single-Trough (SPST) series switch



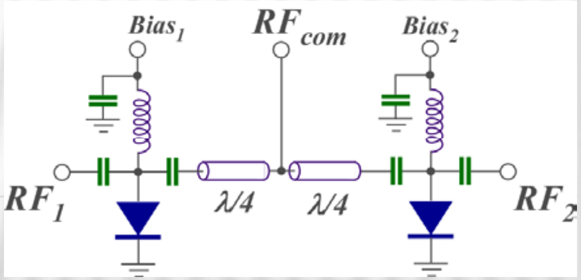
SPST shunt switch



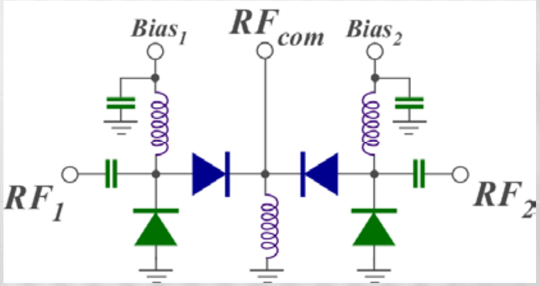
SPST compound series-shunt switch



Single-Pole Double-Trough (SPDT) series switch. Poor isolation.



SPDT shunt switch. Bandlimited.



SPDT compound series-shunt switch.

PIN DIODES
VARIABLE ATTENUATOR/SWITCH

➢ *A. Gallo, Basics of RF Electronics*

GLOSSARY

Attenuators

- **Frequency range**:
  from DC to > 10 GHz, multi-octaves
- **Level**:
  Maximum power at the input (typ. 10÷30 dBm)
- **Insertion Loss**:
  Minimum device attenuation (typ. 1÷6 dB)
- **Isolation**:
  Signal transmission at maximum attenuation (typ. 30÷80 dB)
- **Dynamic range**:
  Excursion of the available attenuation values (typ. 30÷80 dB)
- **Flatness**:
  Attenuation fluctuation over the frequency range at fixed control voltage (typ. 1÷3 dB)
- **Control bandwidth**;
  Modulation frequency producing a peak AM 3 dB lower compared to that produced by a low frequency voltage of the same value

Switches

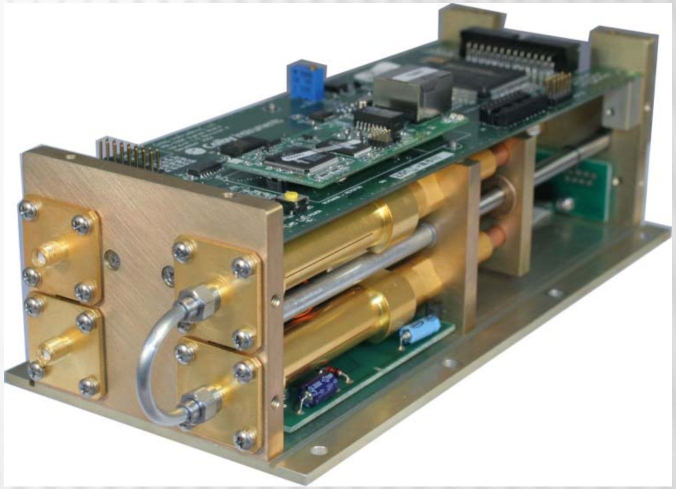
- **Frequency range**:
  from DC to > 10 GHz, multi-decades
- **Level**:
  Maximum power at the input (typ. 10÷30 dBm)
- **Insertion Loss**:
  Attenuation in the "ON" state (typ. 1÷3 dB)
- **Isolation**:
  Signal transmission in the "OFF" state to the output (SPST) or to the unselected port (SPDT) (typ. 25÷80 dB)
- **Switching time**:
  Minimum time required to turn ON/OFF the device (typ. > 5 ns)

PHASE SHIFTERS / STRETCHED DELAY LINES

➢ *A. Gallo, Basics of RF Electronics*

Phase shifters are devices ideally capable to transmit an RF signal **shifting its phase** to any desired value **without attenuation**.

Depending on the nature of the control mechanism phase shifters can be classified as **mechanical or electrical**, and **continuously or digitally** (i.e. in steps) **variable**.



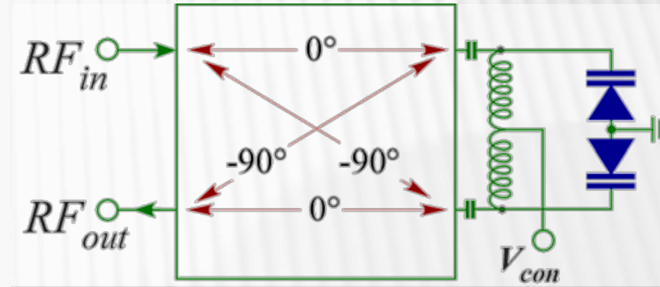
Manual and motorized trombones

Stretched delay lines (trombones) are mechanical, continuously variable, low attenuation and very broadband, and may be used whenever variation speed is not an issue. However, they are expensive and not compact compared to other solutions.

PHASE SHIFTERS / STRETCHED DELAY LINES

➢ *A. Gallo, Basics of RF Electronics*

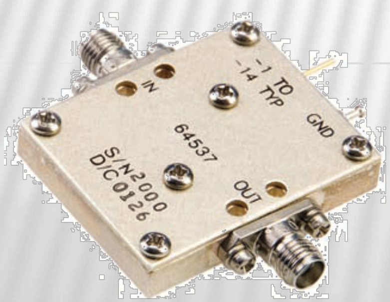
Much **cheaper** and **faster** phase shifter are based on a **90° hybrid** and **varactor** diodes integration. The RF signal sees the **transition capacitance** of the matched, inverse biased varactor diode pair which depends on the control voltage.

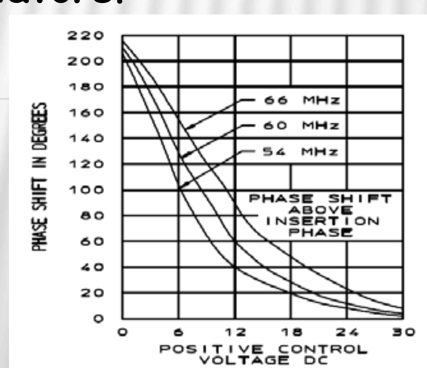


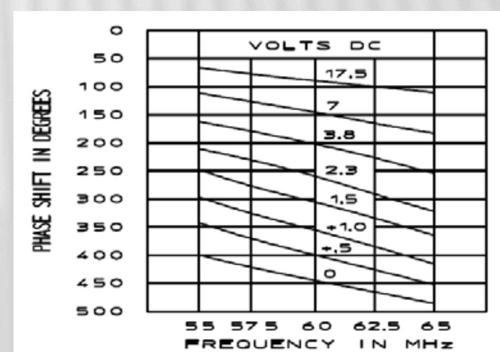
$$\frac{V_{ref_{in}}}{V_{fwd_{in}}} = \frac{1}{2}\left[\frac{1 - j\omega C_T(V_{con})Z_0}{1 + j\omega C_T(V_{con})Z_0} + (-j)^2 \frac{1 - j\omega C_T(V_{con})Z_0}{1 + j\omega C_T(V_{con})Z_0}\right] \approx 0; \qquad \frac{V_{fwd_{out}}}{V_{fwd_{in}}} = -j\frac{1 - j\omega C_T(V_{con})Z_0}{1 + j\omega C_T(V_{con})Z_0}$$

$$\Delta\phi_{out-in} = -\frac{\pi}{2} - \arctan[2\omega C_T(V_{con})Z_0] = \arctan[\omega_0(V_{con})/\omega] + \pi \quad with\ \omega_0(V_{con}) = 1/(2C_T Z_0)$$

Analog continuously variable phase are **typically narrowband** (f_{BW} of the order 10 % of f_0) but available in a wide frequency range extending to ≈ 10 GHz. The **control bandwidth** can extend **beyond 1 MHz**. Therefore this kind of shifters can be also used as **phase modulators**.







PHASE SHIFTERS / STRETCHED DELAY LINES

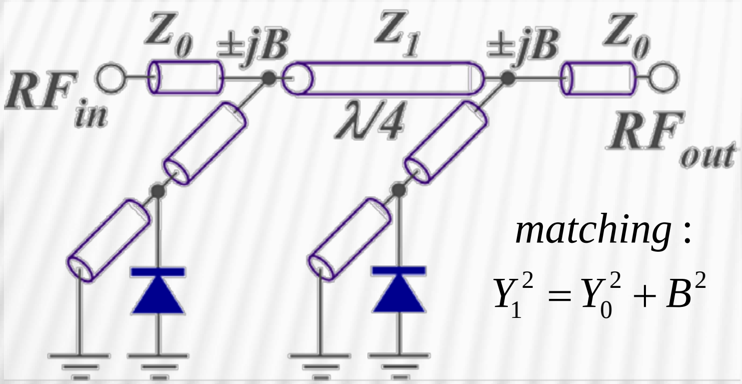
➢ *A. Gallo, Basics of RF Electronics*

Digital phase shifter are based on **PIN diodes arrays**. Depending on the structure of the basic cell they are of the following types:

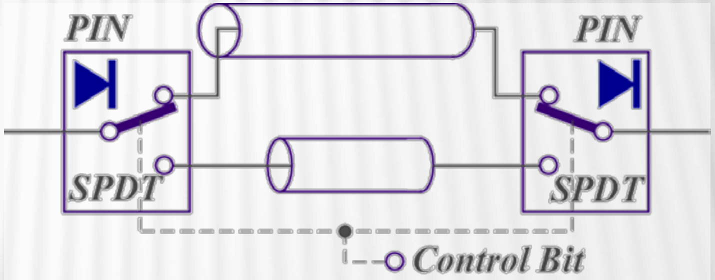
• Periodically loaded-line

• Switched -line

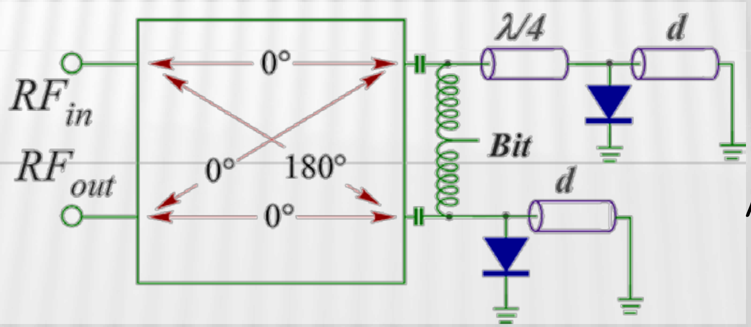
• Hybrid-coupled line



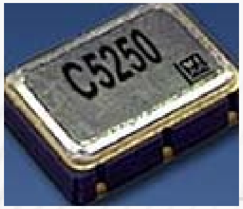
Periodically loaded line



Switched line



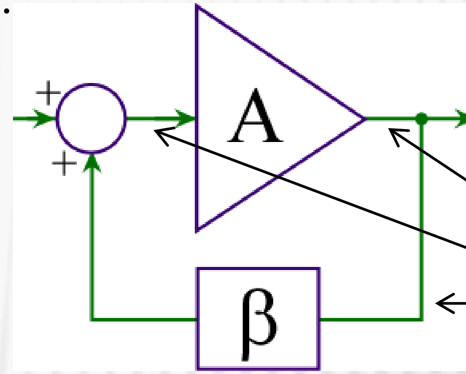
Hybrid coupled line



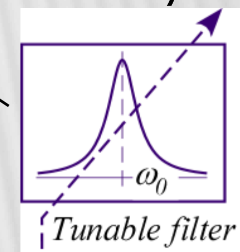
VOLTAGE CONTROLLED OSCILLATORS (VCOs)

➢ *A. Gallo, Basics of RF Electronics*

VCOs are **RF oscillators** whose actual output frequency can be **controlled** by the **voltage present** at a **control** (tuning) port.



$$A_V = \frac{1}{1 - A\beta}$$



Barkhausen Criterion:

Systems breaks into oscillations at frequencies where the loop gain G = Aβ is such that:

$$\left|G(j\omega)\right| = 1; \quad \angle G(j\omega) = 2n\pi$$

More realistically, oscillation occurs at frequencies where the small signal linear gain G_s is:

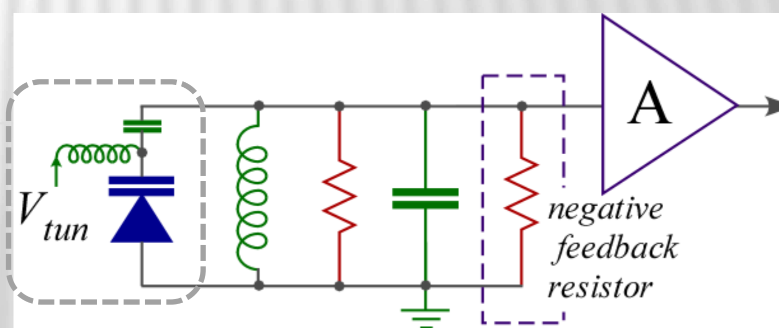
$$\left|G_s(j\omega)\right| > 1; \quad \angle G(j\omega) = 2n\pi$$

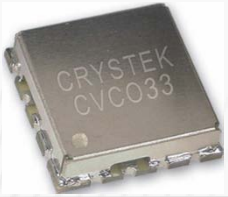
and are confined at amplitudes where non-linearity (compression) sets the large signal G_L at:

$$\left|G_L(j\omega)\right| \approx 1$$

Oscillation frequency is controlled by inserting **tunable filters** in the loop. **Positive feedback** can be modeled as a **negative resistor** compensating the losses of the filter, which behaves as a lossless resonator.

externally tunable element

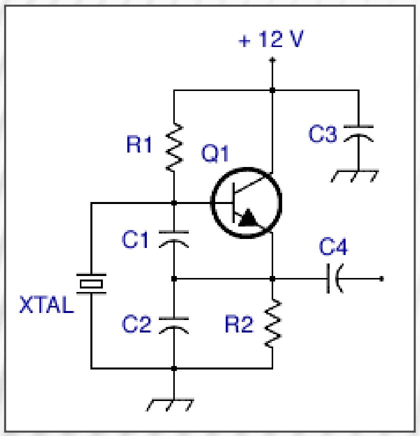




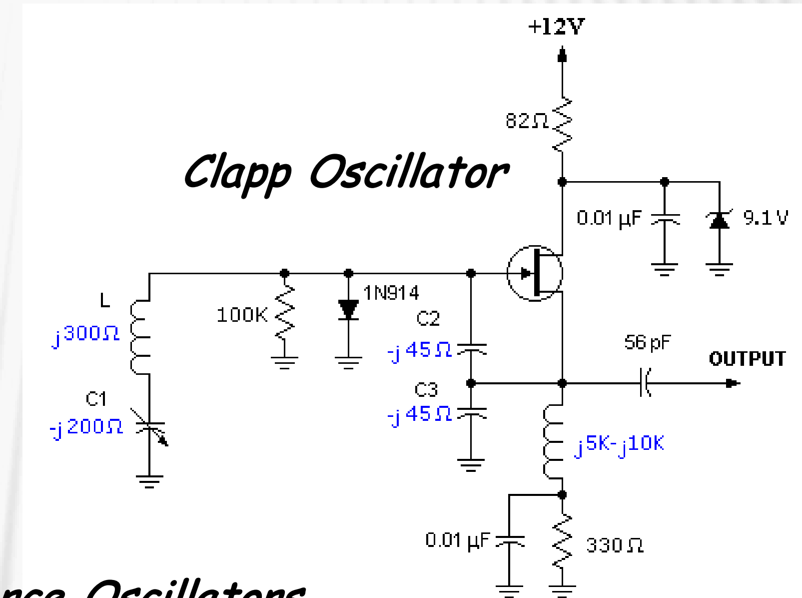
VOLTAGE CONTROLLED OSCILLATORS (VCOs)

➢ *A. Gallo, Basics of RF Electronics*

There are a number of possible oscillator architectures. VCOs use **varactors** as **tuning elements**. Resonant tuning filters can be **lumped**, **transmission line** based or **dielectric resonators** (DROs).

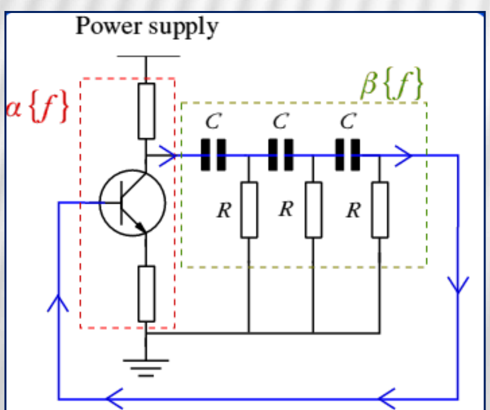


Colpitts Oscillator

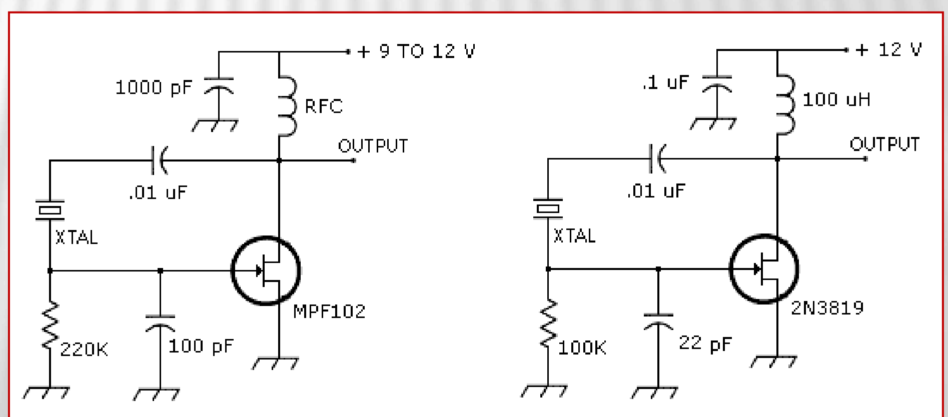


Clapp Oscillator

Phase Shift Oscillator



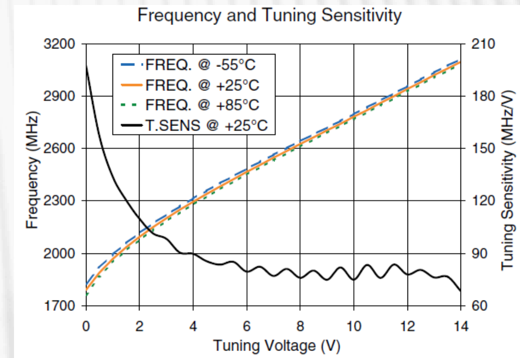
Pierce Oscillators

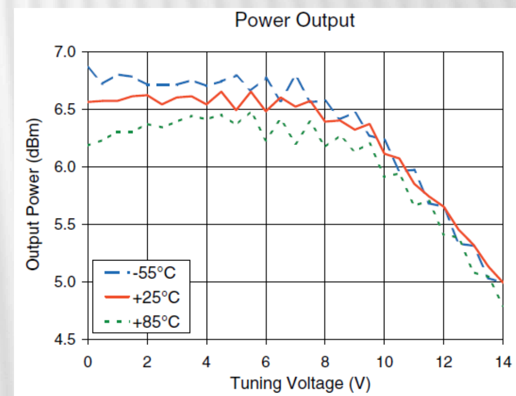


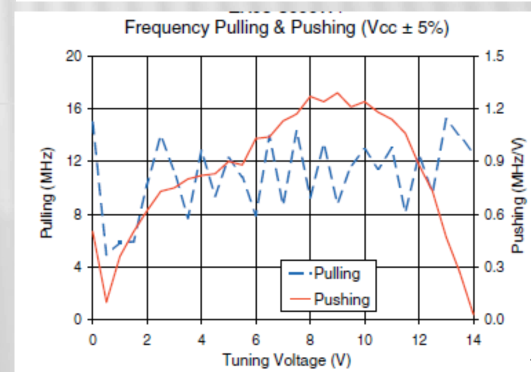
VCO GLOSSARY

➢ *A. Gallo, Basics of RF Electronics*

- **Tuning characteristics**;
  Frequency versus tuning voltage plot.
- **Tuning sensitivity**;
  Slope of the tuning characteristics, typically given in MHz/V. Is a local parameter in case the tuning characteristics is not linear over the entire range.
- **Temperature sensitivity**;
  Frequency variation with temperature at a fixed tuning voltage.
- **Modulation bandwidth / Tuning speed**;
  Modulation frequency producing a peak frequency deviation reduced by 3 dB compared to that produced by a dc voltage of the same value / Time required to settle the output frequency deviation to 90% of the regime value after application of a voltage step variation on the tuning port. The two parameters are obviously correlated.
- **Output power / Output power flatness**;
  Level of the oscillator output fundamental harmonic into a 50 Ω load / Variation of the output level over the specified VCO frequency range
- **Frequency pushing / Frequency pulling**;
  Variation of the VCO frequency with the supply voltage at fixed control voltage / Variation of the output frequency with the load mismatch (typically given as peak-to-peak value at 12 dB return loss, any phase)







VCO GLOSSARY

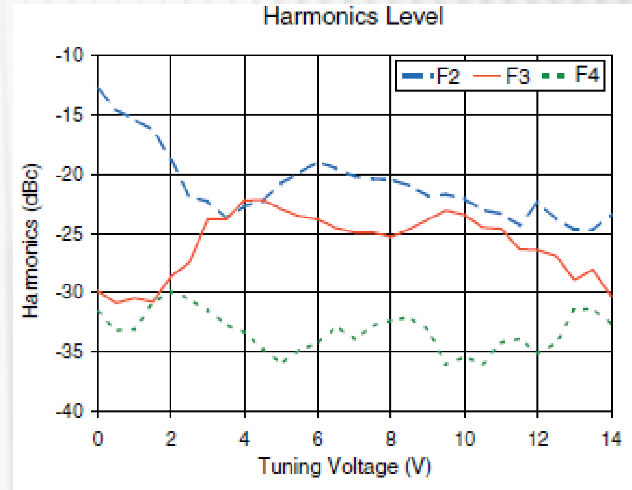
➢ *A. Gallo, Basics of RF Electronics*

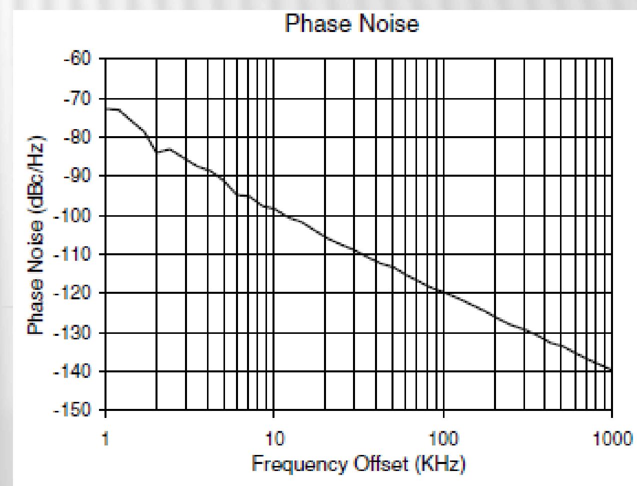
- **Harmonic suppression**;
  Level of the harmonics relative to the fundamental (typically given in dBc = dB below the carrier).
- **Spurious content**;
  Level of the spurious, non-harmonic output signals relative to the oscillator output (typically given in dBc).
- **SSB phase noise**;
  Single sideband phase noise in 1 Hz bandwidth as a function of the frequency offset from the carrier frequency, measured relative to the carrier power and given in dBc/Hz. Very important to evaluate the expected residual phase noise in Phase Locked Loops.
- **rms phase jitter**;
  rms value of the instantaneous phase deviation, which is given by the integral of the SSB power spectrum:

$$\varphi_{rms}^2 = \frac{1}{\Delta T}\int_{t_0}^{t_0+\Delta T}\varphi^2(t)\,dt = 2\int_{f_L}^{f_H}10^{-(SSB_{dBc}/10)}\,df$$

rms jitter expressed in terms of frequency deviation is known as "***residual FM***", defined as the SSB power spectrum integral between f_L=*50* Hz and f_H=3 kHz:

$$\Delta f_{rms}^2 = \frac{1}{\Delta T}\int_{t_0}^{t_0+\Delta T}\Delta f^2(t)\,dt = 2\int_{f_L}^{f_H}10^{-(SSB_{dBc}/10)}\,f^2 df$$



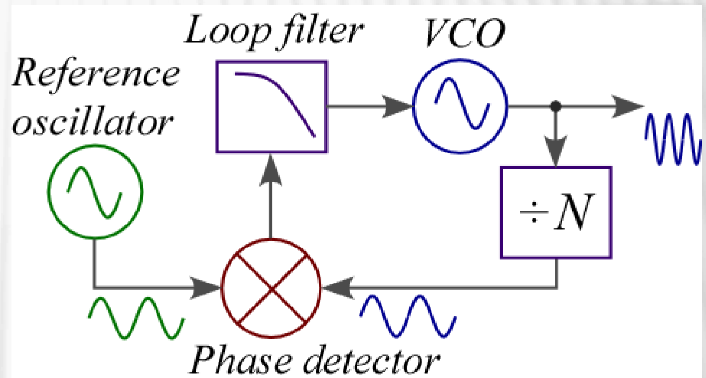


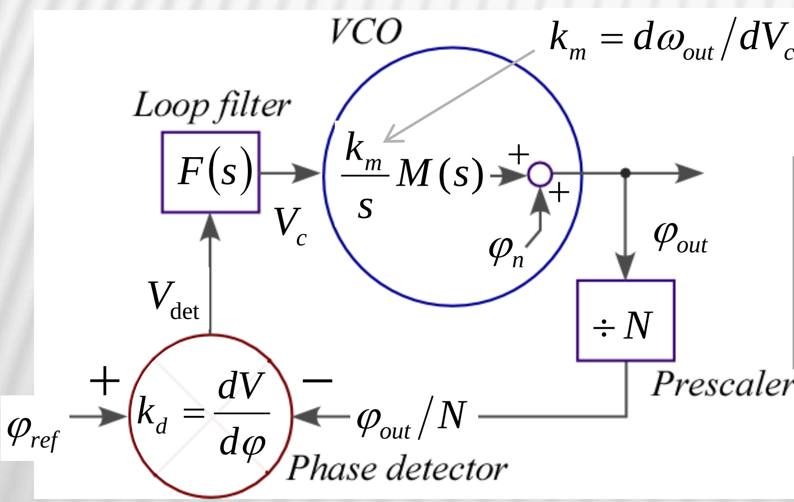
PHASE LOCKED LOOPS (PLLs)

➢ *A. Gallo, Basics of RF Electronics*

PLLs are a very **general subject** in RF electronics. They are used **to synchronize oscillators** to a **common reference** or to **extract the carrier** from a **modulated signal** (FM tuning). The PLL main components are:

- A VCO, whose frequency range includes Nf_{ref};
- A phase detector, to compare the scaled VCO phase to the reference;
- A loop filter, which sets the lock bandwidth;
- A prescalers (by-N frequency divider), which allows setting different output frequencies w.r.t. the reference one.





PLL linear model

PLL transfer function

VCO noise

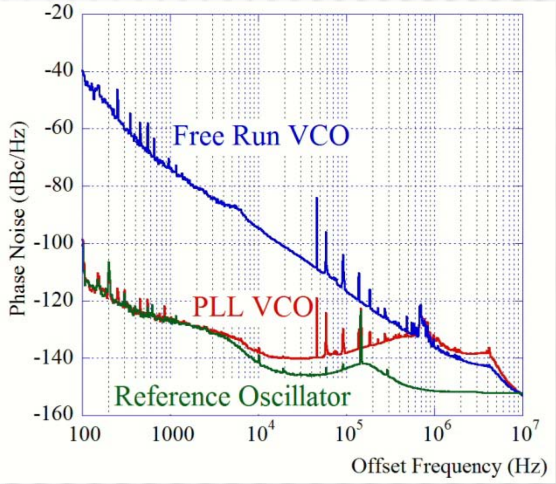
$$\varphi_{out}(s) = N\frac{H(s)}{1+H(s)}\varphi_{ref}(s) + \frac{1}{1+H(s)}\varphi_n(s)$$

$$\text{with } H(s) = \frac{k_d k_m}{N s}F(s)M(s)$$

freq-to-phase conversion — *loop filter* — *VCO mod. bandwidth*

PHASE LOCKED LOOPS (PLLs)

➢ *A. Gallo, Basics of RF Electronics*

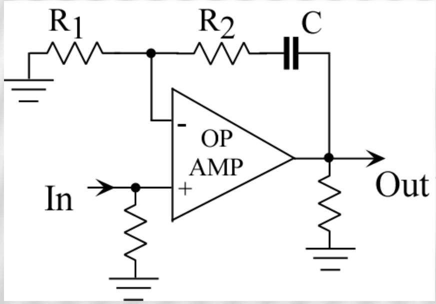


Loop filters provide **PLL stability**, tailoring the frequency response, and **set loop gain** and **cut-off frequency**.

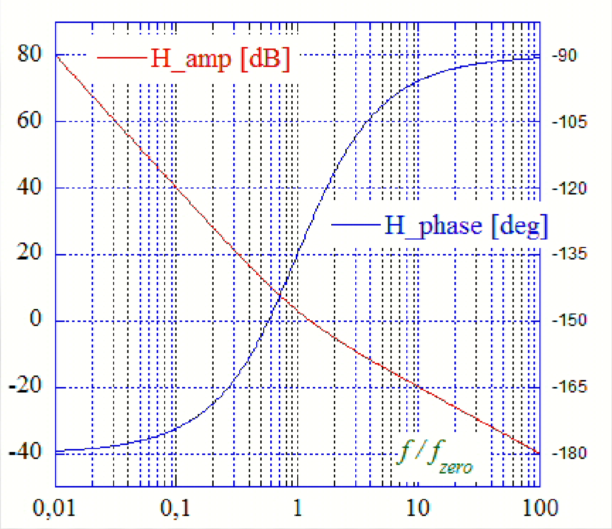
The **output phase** spectrum is **locked** to the **reference** one if **|H(jω)|>>1**, while it returns similar to the **free run VCO** if **|H(jω)|<1**.

A flat-frequency response loop filter gives already a pure integrator loop transfer function thanks to a pole in the origin (*f=0*) provided by the dc frequency control of the VCO.

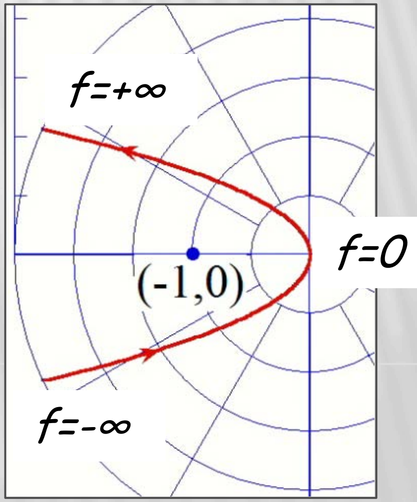
The low frequency gain can be further increased with a loop filter providing an extra pole in the origin and a compensating zero at some non-zero frequency ($f_{zero}=1/2\pi R_2 C$).



A very steep loop frequency response is obtained (slope = 40 dB/ decade) in stability conditions (see Nyquist plot).



Bode plot of the PLL loop gain



PLL loop gain: Nyquist locus

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THE END

Thank you for your attention