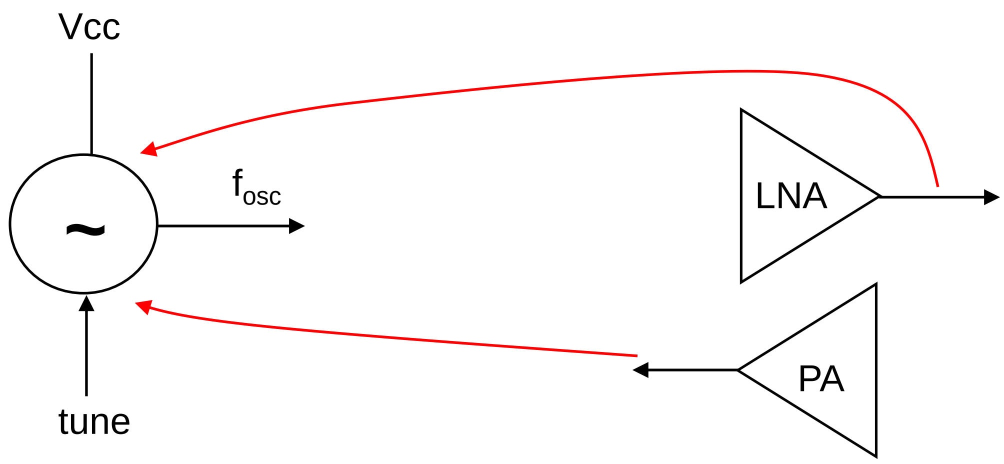
LO generation design

- **LO generation design**
  - The issue of VCO pulling
  - Integer VCO/LO generation scheme
  - Fractional VCO/LO generation scheme
  - Super-regenerative LO generation scheme
  - VCDL LO generation scheme

- **References**

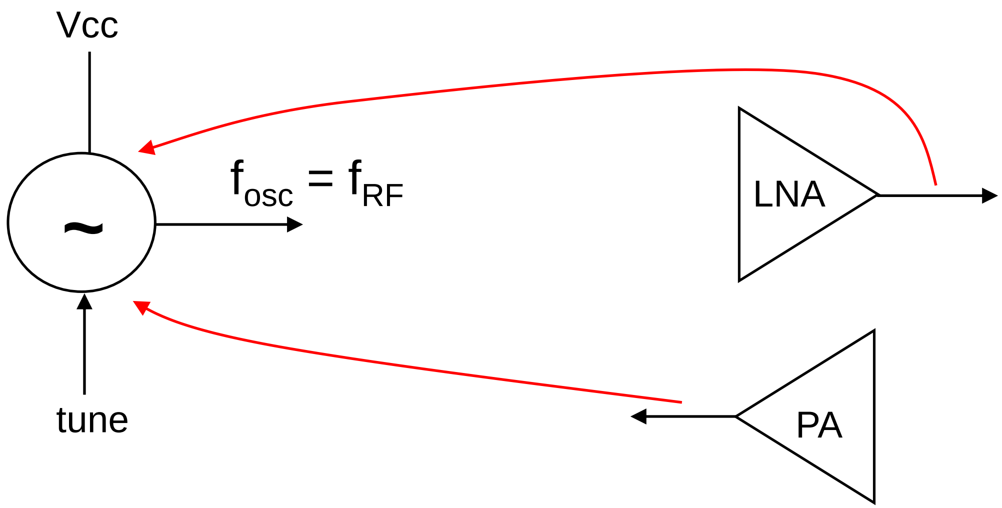
The issue of VCO pulling:



integrating the VCO with PA, and/or LNA makes it prone to pulling. This is due to finite isolation on-chip between these blocks. The VCO pulling is the disturbance of the oscillation frequency due to presence of large signals on-chip. The LNA/PA coupling to VCO can be via:

- magnetic due to spiral inductor coupling or bondwire coupling
- capacitive due to layout parasitics
- substrate
- thermal (PA increases substrate temperature causing VCO pulling due to finite varactor temperature-coefficient)

LO generation scheme of VCO running at RF frequency:



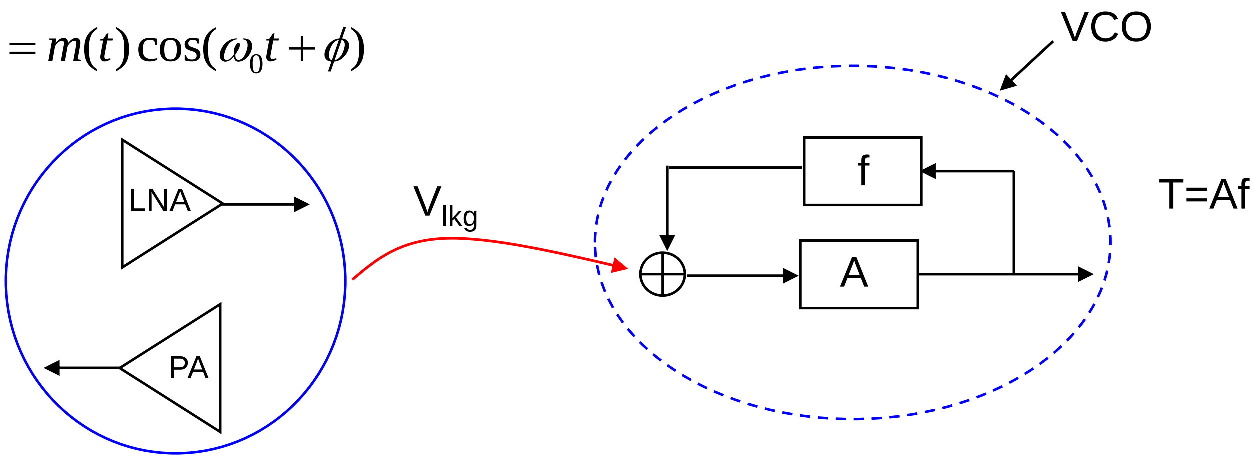
In this scheme, VCO, LNA and PA are all operating at the same frequency. Let us assume that the VCO loop (with both LNA and PA off) has a single frequency solution ($A\cos\omega_0 t$) that satisfies the criteria of oscillation. That is:

$$|H(s)| = \frac{A}{1+T} = 1$$

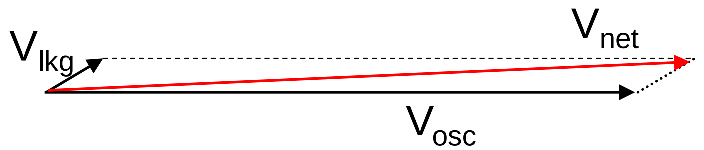
$$\Phi(H(s)) = 360^\circ$$

Let us assume there is an amount of “leakage” from PA or LNA to VCO. This leakage energy can be written as

$$V_{lkg}(t) = m(t)\cos(\omega_0 t + \phi)$$

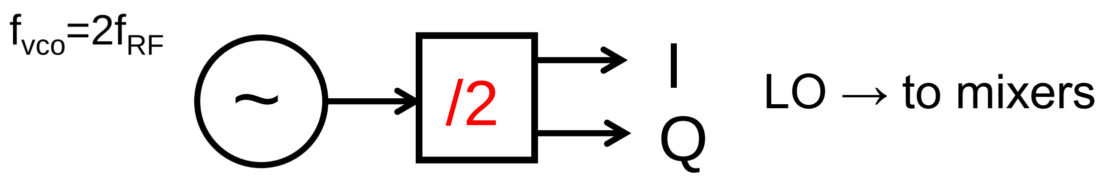


The leakage energy will disturb the equilibrium of the feedback circuit of the VCO. The disturbance can be viewed as modification of both magnitude and phase around the VCO loop. As a result, there will be a new “solution” of the loop that satisfies the criteria of oscillation. If the leakage energy is only a pure tone of the same frequency as of the VCO (m(t)=0 and ϕ is constant), then this results only in a phase shift of the oscillation signal.



If the leakage signal has AM, PM or both, the “solution” for the loop that satisfies the criteria of oscillation will change as the leakage signal amplitude or phase changes. As a result the output of the VCO will carry both AM and PM of the leakage signal. Note that if the leakage signal frequency is the same or close to the VCO pull-free oscillation frequency, the loop gain of the VCO (T) is still high enough to track the leakage signal due to the Q of the VCO tank, making VCO most vulnerable to pulling at or very close its resonance frequency. If the leakage signal frequency, however, is far from the VCO frequency, the VCO loop gain T is very small and provides rejection to the leakage signal (again due to VCO tank), making the VCO more immune to pulling. Therefore, the LO generation scheme where the VCO runs at the same frequency as RF (Rx or Tx) is the most scheme prone to VCO pulling and is rarely used.

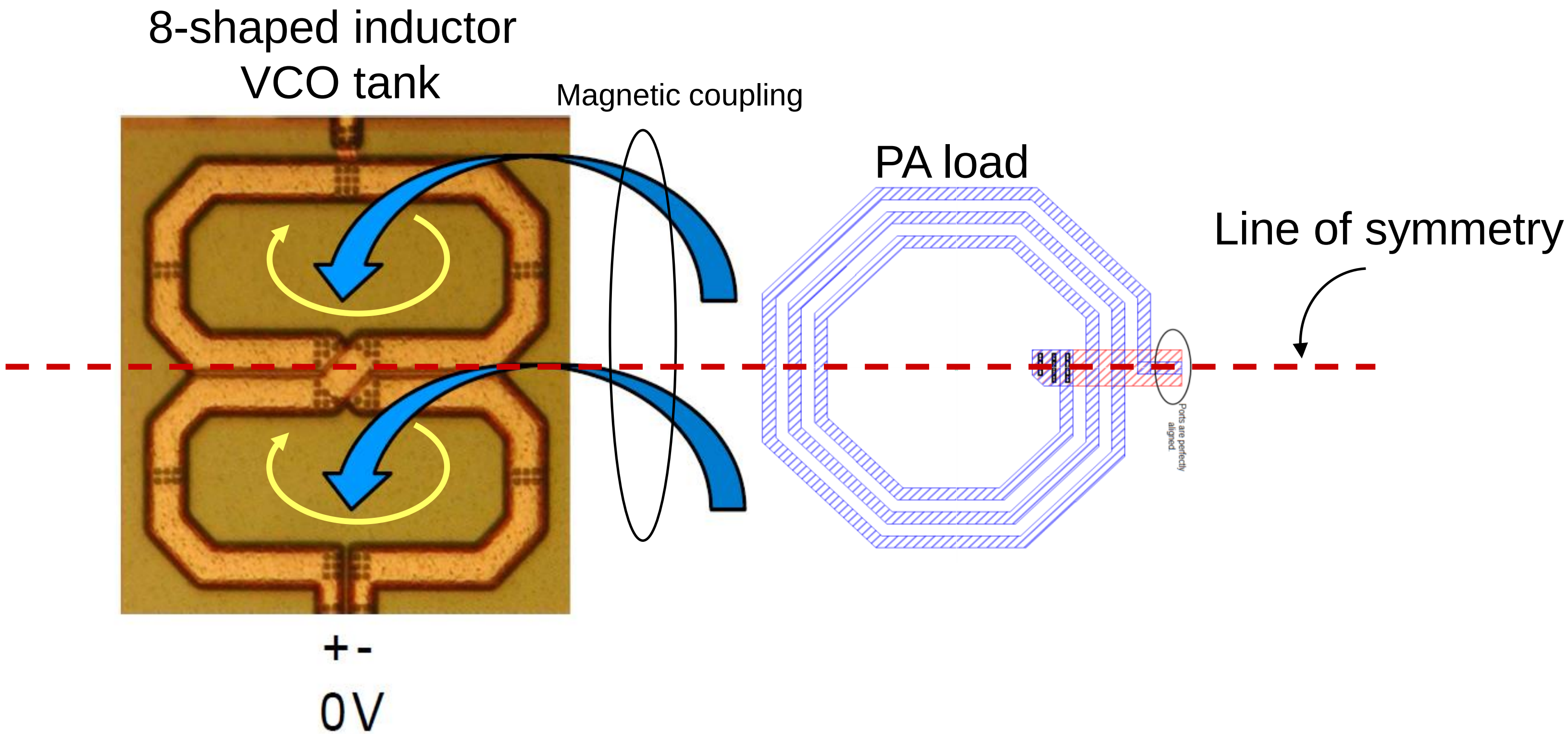
LO generation scheme of VCO running at twice RF frequency:



The VCO tank (centered at $2f_{RF}$) provides ~40dB rejection @ f_{RF}, and therefore provide immunity to VCO pulling due to desired RF receive or transmit signal. This scheme is desirable because of its simplicity, relatively low current consumption, and compact area. The I/Q LO signal can be obtained by dividing the VCO output frequency by 2.

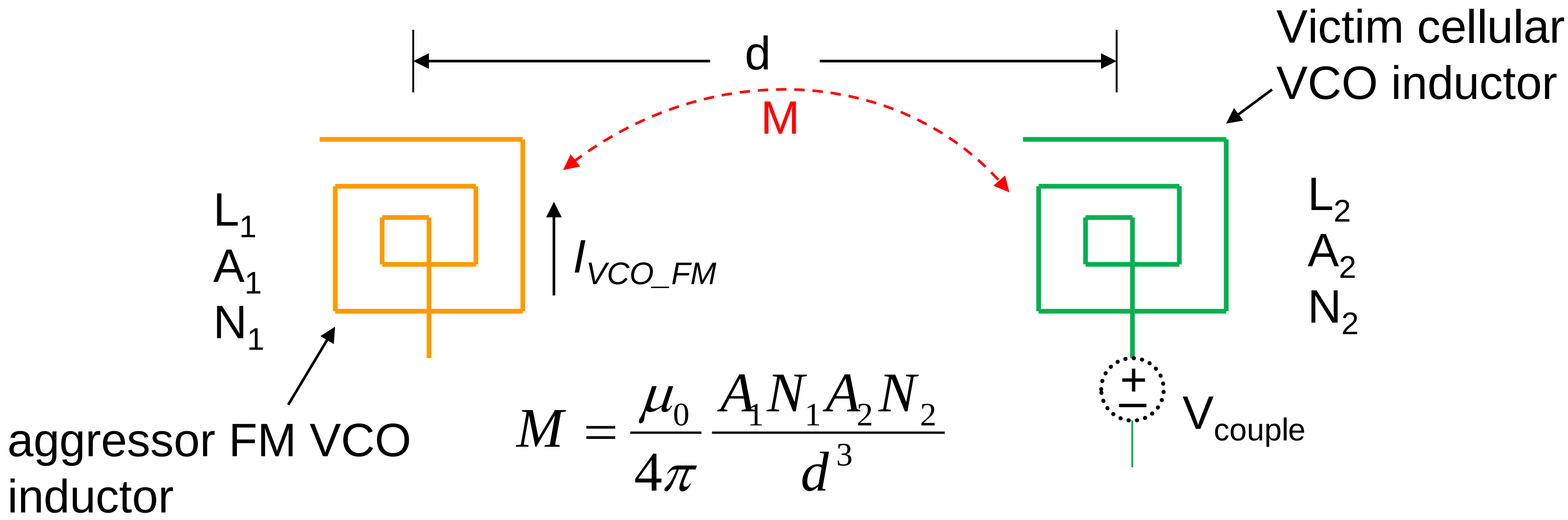
The topology, however, is prone to pulling from the second harmonic of a strong blocker in the receive signal, or the second harmonic of the transmitted signal. This effect can be minimized by:

1. Using differential topology for LNA, PA and VCO. This in turn will transform the second harmonic of both Rx and Tx into a common-mode signal.
2. Use an 8-shape inductor for VCO tank and place it in layout at the line of symmetry to PA balun (or LNA load inductor)



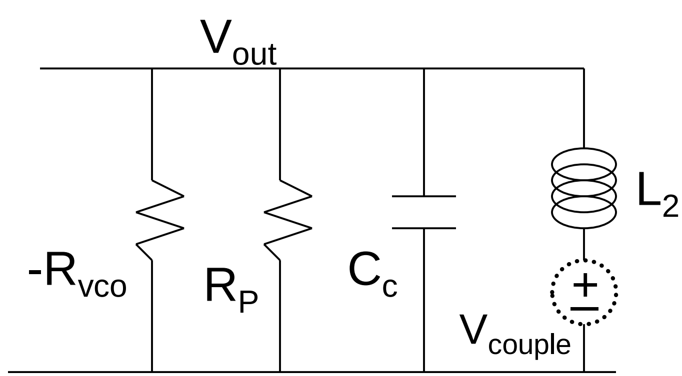
Any current flowing into the PA (or LNA) inductive load results in a magnetic field coupling into the VCO tank. Because of the 8-shaped inductor of VCO, the PA load magnetic coupling results in an equal induced current flowing in opposite direction in each of the two loops comprising the 8-shaped inductor of VCO tank . As a result, these two equal induced currents will cancel each other at the far field. It is important to note that the PA inductor has to be placed at the line of symmetry with respect to the 8-shaped inductor to achieve best rejection (typically >30dB at 2GHz even with 200um distance). However, isolation degrades quite fast if this condition is not met.

So how much aggressor-to-VCO isolation is needed?:



$$M = \frac{\mu_0}{4\pi}\frac{A_1 N_1 A_2 N_2}{d^3}$$

$$V_{couple} = (\omega_{FM} M) I_{VCO\_FM} = \omega_{FM} M \frac{V_{VCO\_FM}}{\omega_{FM} L_1} = \frac{M}{L_1} V_{VCO\_FM}$$



$$Q_{tnk\_loaded} = \frac{R_P}{\omega_{cell} L_2} = R_P \omega_{cell} C_c$$

$$\omega_{cell} = \frac{1}{\sqrt{L_2 C_c}}$$

$|-R_{VCO}| = R_P$ at resonace since VCO negative gm cancels tank loss

Cell VCO tank R-L-C model at resonance

So how much isolation is needed?:

Here we assume that the FM VCO runs at a frequency very close to the cellular VCO. Let us calculate how the V_{couple} signal transfers across the VCO tank of the cell VCO:

$$\frac{V_{out}}{V_{couple}} = \frac{\frac{1}{C_c s}}{sL_2 + \frac{1}{C_c s}} = \frac{1}{C_c L_2 s^2 + 1}; \quad s = j\omega_{FM}$$

$$\Rightarrow \frac{V_{out}}{V_{couple}} = \frac{1}{-C_c L_2 \omega_{FM}^2 + 1}; \quad \text{but } \omega_{cell}^2 = \frac{1}{C_c L_2}, \text{ and let us assume } \omega_{FM} = \omega_{cell} - \Delta\omega_{offset}$$

$$\Rightarrow \frac{V_{out}}{V_{couple}} = \frac{1}{-\frac{(\omega_{cell} - \Delta\omega_{offset})^2}{\omega_{cell}^2} + 1} = \frac{1}{-\frac{(\omega_{cell}^2 - 2\Delta\omega_{cell}\omega_{offset} + \Delta\omega_{offset}^2)}{\omega_{cell}^2} + 1}$$

$$= \frac{1}{-1 + \frac{2\Delta\omega_{offset}}{\omega_{cell}} - \frac{\Delta\omega_{offset}^2}{\omega_{cell}^2} + 1} \approx \frac{1}{\frac{2\Delta f_{offset}}{f_{cell}}}$$

the “gain” by which V_{couple} gets to the VCO tank collapses to

$$\frac{V_{out}}{V_{couple}} \approx \frac{f_{cell}}{2\Delta f_{offset}}$$

Which is very large.

Example on how to calculate spur in VCO due to an aggressor:

A1 = 250μm x 250μm, A2 = 180μm x 180μm
N1 = 2 turns, N2 = 2 turns
d = 4000μm
L1 = 1.5nH, L2 = 1nH
$V_{VCO\_FM} = V_{VCO\_cell}$ = 2V rms
f_{offset} = 400kHz, f_{cell} = 3.5GHz

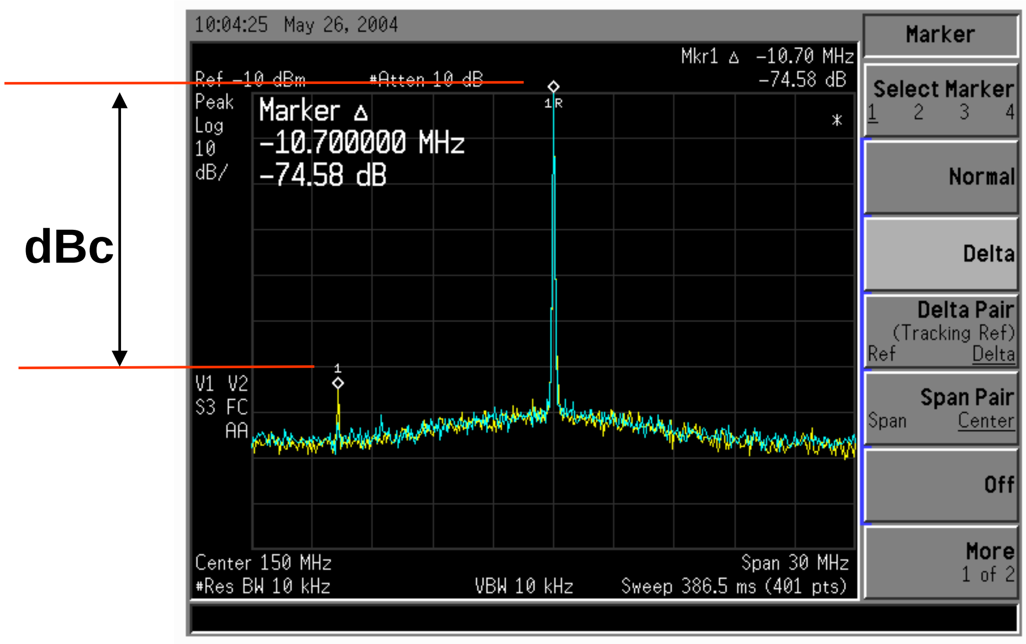
$$\Rightarrow M = \frac{\mu_0}{4\pi}\frac{A_1 N_1 A_2 N_2}{d^3} = 1e^{-7}\frac{4(180)^2(250)^2}{(4000)^3}1e^{-6} = 12.7\,fH$$

$$V_{couple} = \omega_{FM} M I_{VCO\_FM} = \frac{M}{L1}V_{VCO\_FM} = 2\frac{12.7\,fH}{1.5nH} = 16.9\mu V$$

$$V_{out} = V_{spur} = \left|\frac{1}{\frac{2(400k)}{3.5G}}\right| \times 16.9\mu V \approx 74mV$$

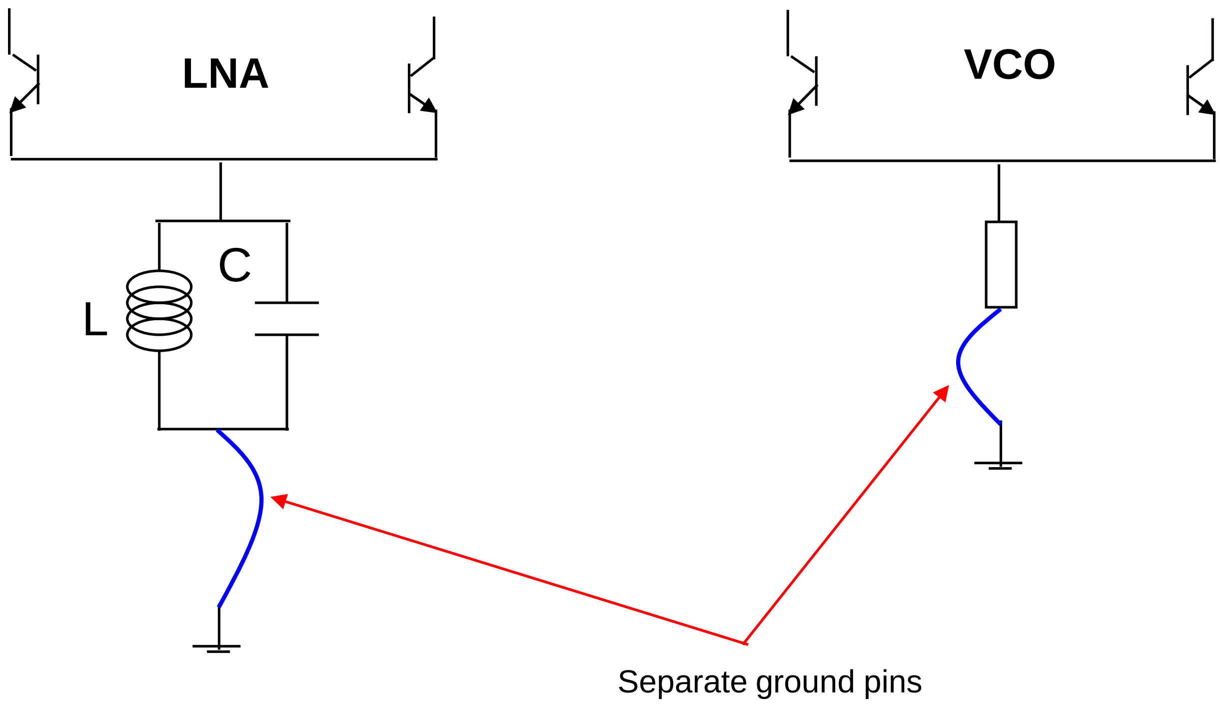
$$\Rightarrow spur_{dBc} = 20\log\left(\frac{74mV}{2V}\right) = -28dB_C$$

How much spur level can be tolerated? (aggressor is a CW tone)

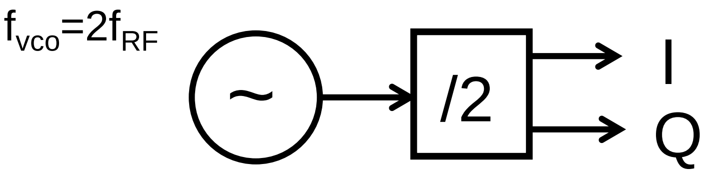


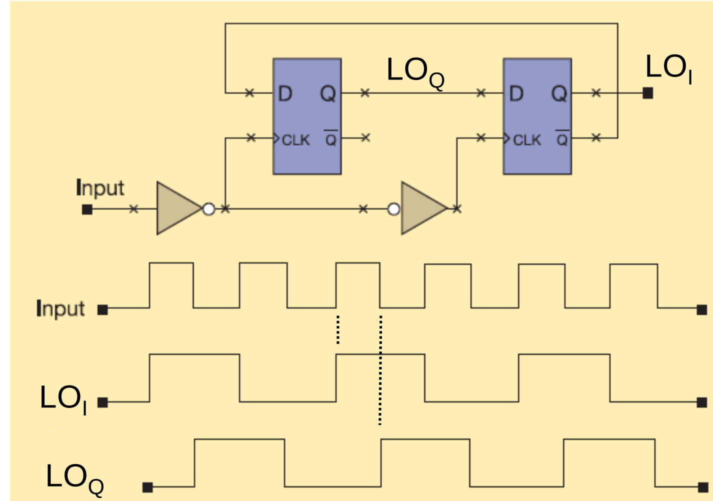
- Tolerable in-band spur level is set by the target EVM of the modulation (Tx) or SNR (Rx). Usually it is best to have the spur level at least 10dB lower than target EVM (or SNR).
- Tolerable out-of-band spur level is set by out of band emission requirement (ACLR, emission mask and FCC regulation).
- A good practice is to use an EM simulator with actual aggressor and VCO Tank layout to find exact coupling and use circuit simulator to find out the spur level and adjust separation in layout accordingly (no guessing).

3. Using a dedicated Vdd and ground pin connection of the VCO core separate from the rest of the chip, especially the Vdd and ground of the PA/LNA/aggressor which are rich of second harmonic components.

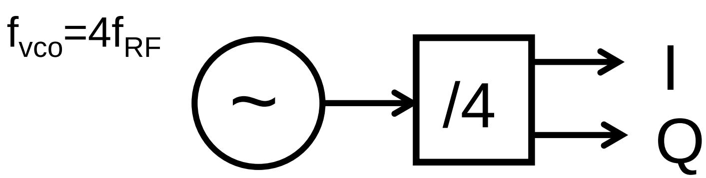


Note that if an exposed paddle package is used, a dedicated ground pin has only 10dB isolation to the paddle at 2GHz. Therefore, a lateral isolation (separation) between ground pins is needed, in addition to having orthogonal orientation to minimize bond-wire coupling.



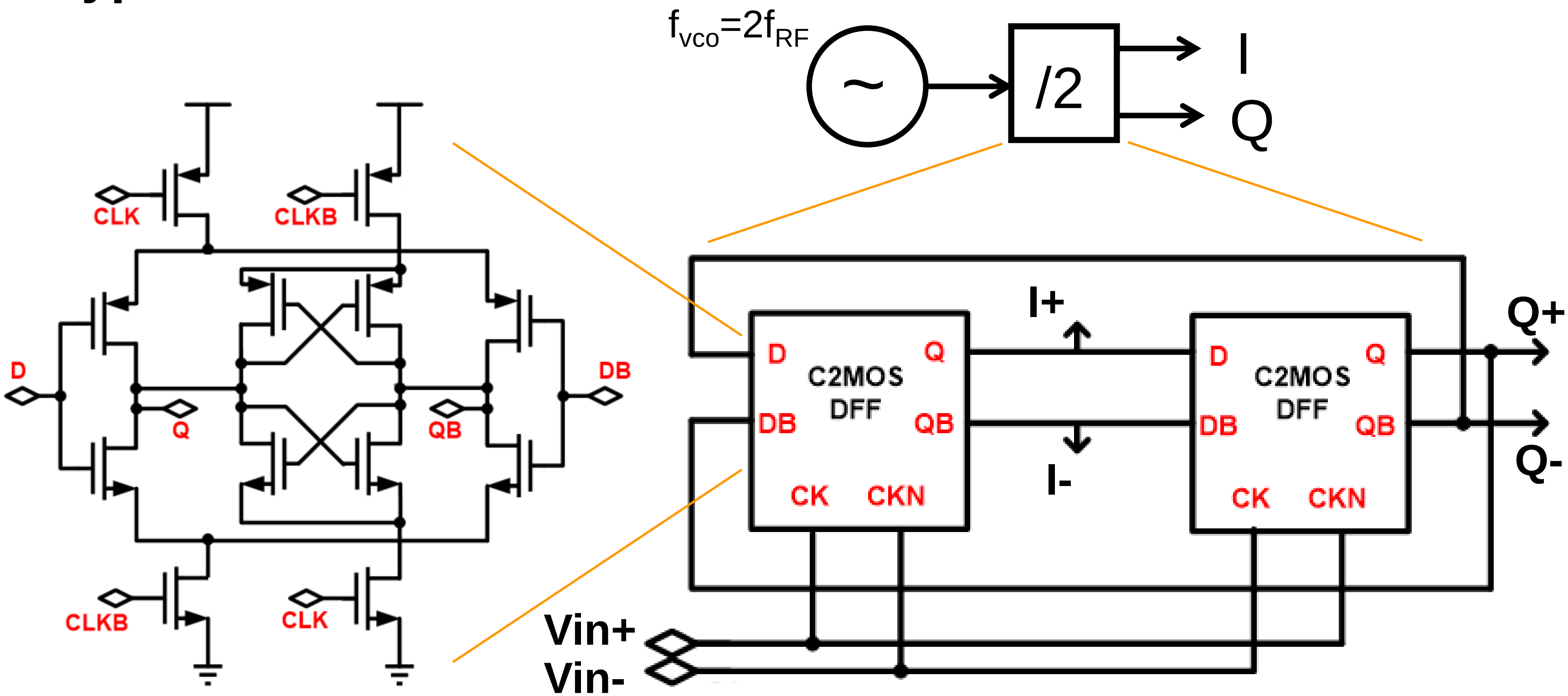


A disadvantage of this scheme is that there is an upper limit on the frequency it can operate at. Up to this moment, reliable low power VCOs with integrated tank run at about 12GHz or lower. This means the maximum RF frequency this scheme can serve is below 6GHz.



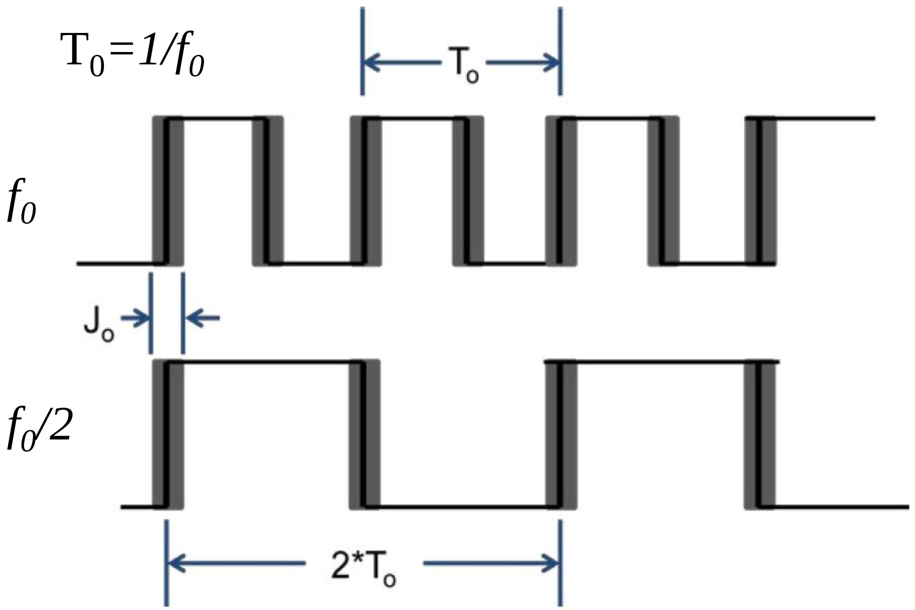
A more aggressive scheme is to use a VCO that runs at 4x the f_{RF}. This way, the VCO is prone to pulling from the 4th harmonic of the PA, which is much smaller than its 2nd harmonic. Such scheme has been successfully used for 2.4GHz Bluetooth transmitters for example with integrated PA delivering 13dBm of power.

A typical /2 circuit:



A master-slave flip-flop architecture is the most common for a divide-by-2 LO generation scheme. The flip flop is implemented using CMOS logic circuit that fits the speed of a given technology (CML, C2MOS, etc.). Device sizing (hence current) is set by divider speed requirement (max input frequency) and I/Q phase matching requirement. Please note that the transition of the divider output is set by the transition (zero-crossing) of the input clock

Input/output phase noise of a divide-by-2:



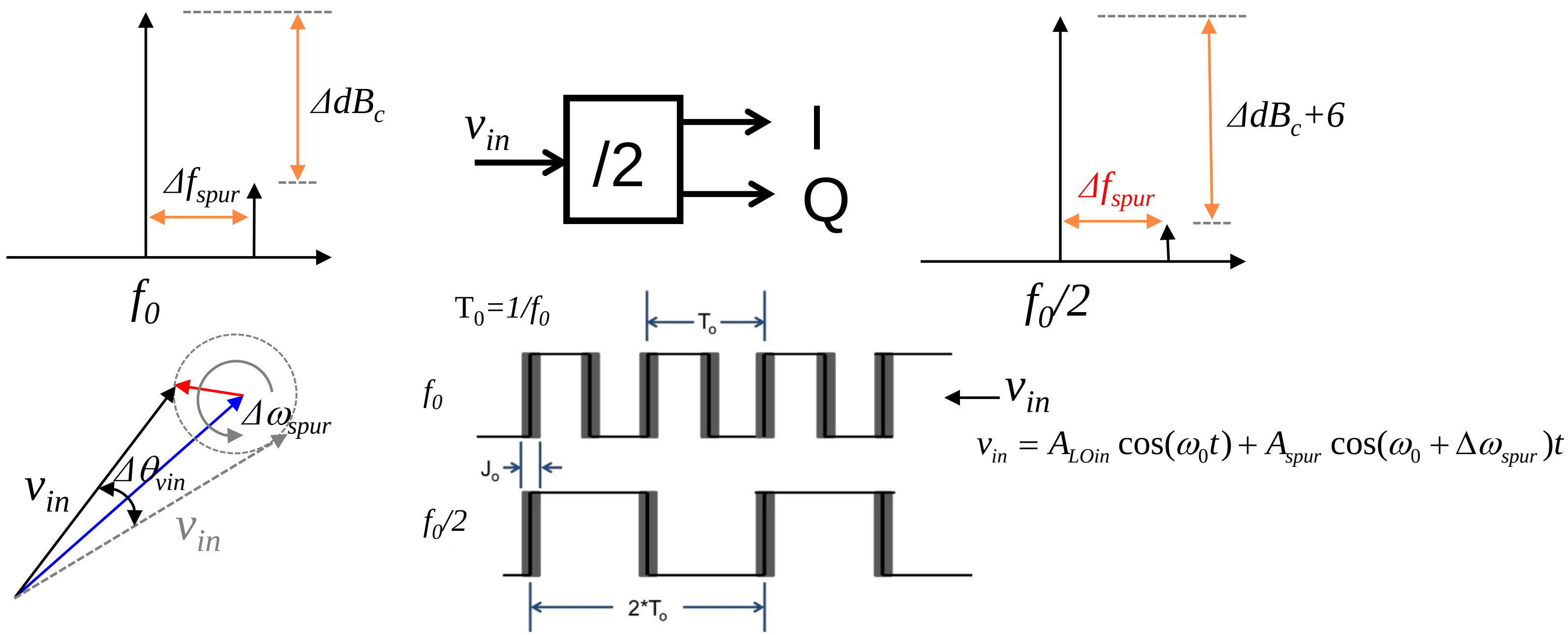
$$T_{jitter} = J_0 = \frac{\phi_{rms}}{2\pi f_0}$$

$$\rightarrow \phi_{rms} = 2\pi f_0 T_{jitter}$$

Φ_{rms} is the DBS integrated phase error in rad

- What a divider does to phase noise is best seen in time domain by looking at the jitter of its input/output signals. Since the divider output logic state changes with rising/falling edge of the input clock, any jitter at that input signal will translate as jitter at the output signal of exact same amount.
- Because the period of the output is twice of that at input, for the same amount of jitter, phase noise improves by 6dB because the phase is jitter times frequency
- For an N-divider, the output phase noise improves by 20logN (ignoring divider's own noise)

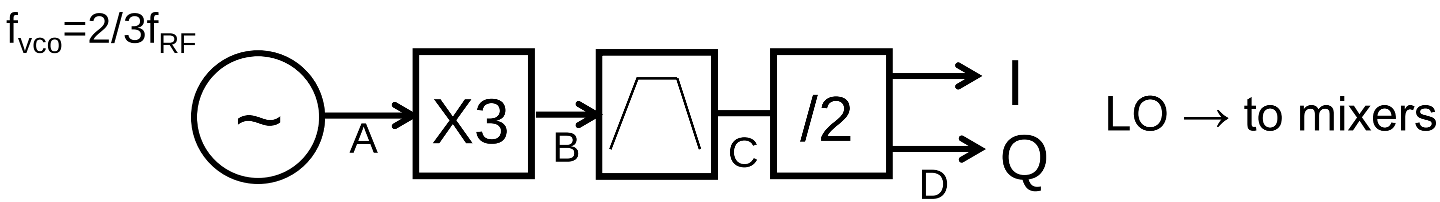
Input/output spur relation of a divide-by-2:



- a close-in spur alters the phase of the input vector sum with a deterministic CW tone rotating at an angular frequency equals to $\Delta\omega_{spur}$. Therefore, its effect is similar to input jitter. Therefore, the divide-by-2 reduces this spur level by 6dB at its output.
- However, the divider does <u>not</u> change the offset frequency of this spur (it remains the same distance from divided LO as it is from input LO. This is because the rate of change of the input zero crossing, which also decides the transition of divider output LO rising/falling edges is the same, $\Delta\omega_{spur}$
- For an N division, the spur level drops by 20log(N)

LO generation scheme of VCO running at non-integer multiple of RF frequency:

Running the VCO at a non-integer multiple of the RF frequency makes it very robust against pulling from the RF signal or its harmonics.



$$A: \cos(\frac{2}{3}\omega_0 t)$$

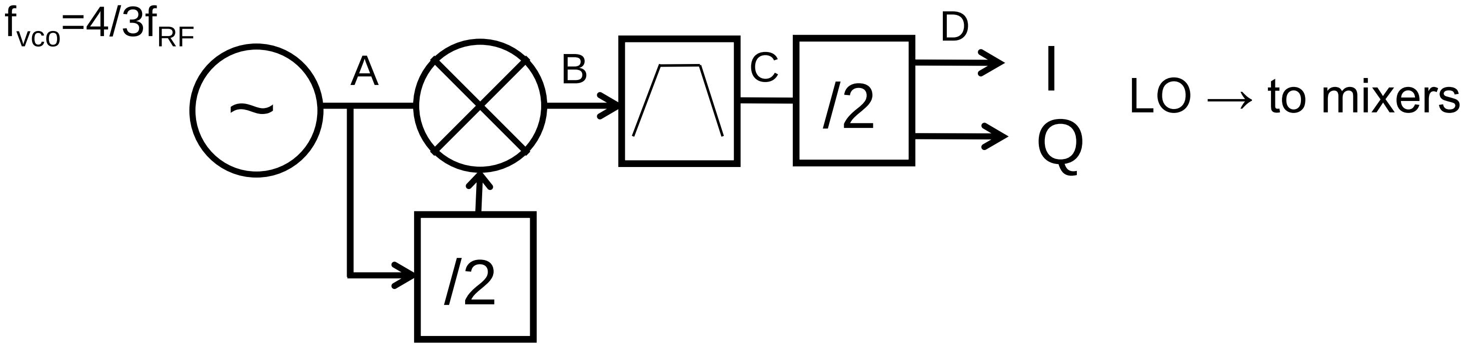
$$B: A^2 A = \cos(2\omega_0 t) + \cos(\frac{2}{3}\omega_0 t)$$

$$C: \cos(2\omega_0 t)$$

$$D: I = \cos(\omega_0 t) \quad , \; Q = \sin(\omega_0 t)$$

Note that the term at (2/3)ω_0 need to be rejected by the tank so as not to desense other radios

Another non-integer scheme:



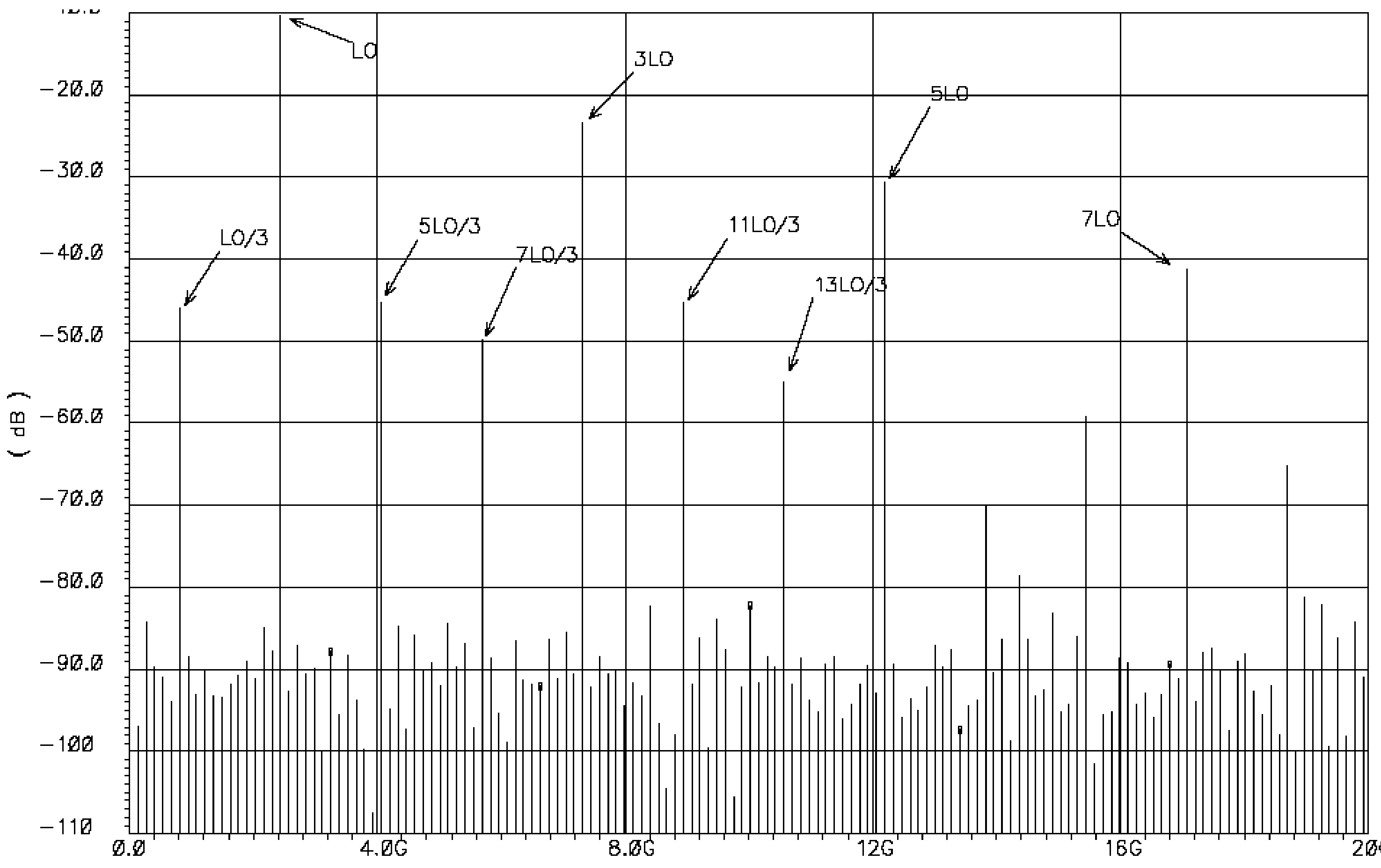
$$A: \cos(\frac{4}{3}\omega_0 t)$$

$$B: \cos(\frac{4}{3}\omega_0 t)\cos(\frac{2}{3}\omega_0 t) = \cos(2\omega_0 t) + \cos(\frac{2}{3}\omega_0 t)$$

$$C: \cos(2\omega_0 t)$$

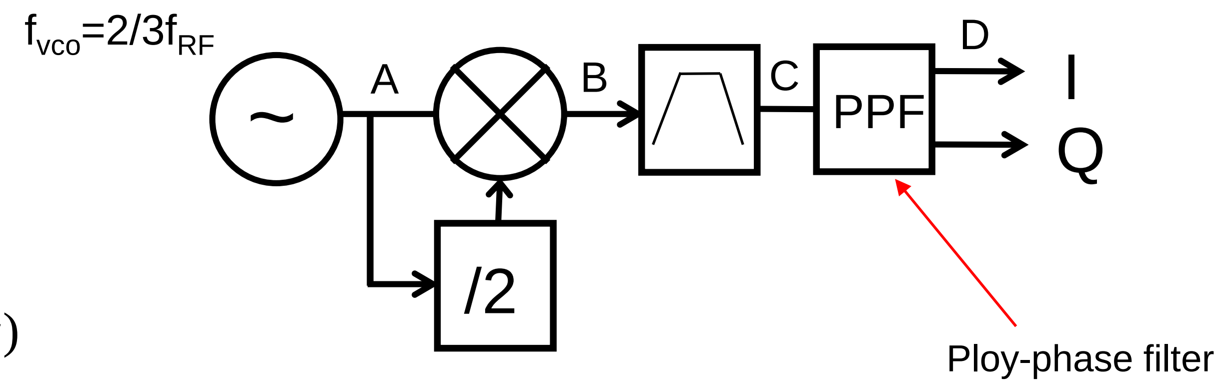
$$D: I = \cos(\omega_0 t) \quad , \ Q = \sin(\omega_0 t)$$

A disadvantage of a non-integer LO generation scheme is the spur generation due to the highly non-linear operation. The most important spur is the image ,$(2/3)\omega_0$, before the bandpass filter and the divide by 2. In fact, a typical spectrum of such a scheme looks like below:



The proposed non-integer schemes still suffers from an upper limit of frequency they can operate at. For example, to generate a 5GHz LO for WLAN 802.11a, a 6.67GHz VCO needs to be built. In addition, a 10GHz divide by 2 needs to be designed, which is not trivial and can be power hungry.

A way to operate at higher LO frequencies, the LO generation scheme is modified to eliminate the divide by 2. To get the I/Q LO, a ploy phase filter is used as follows:



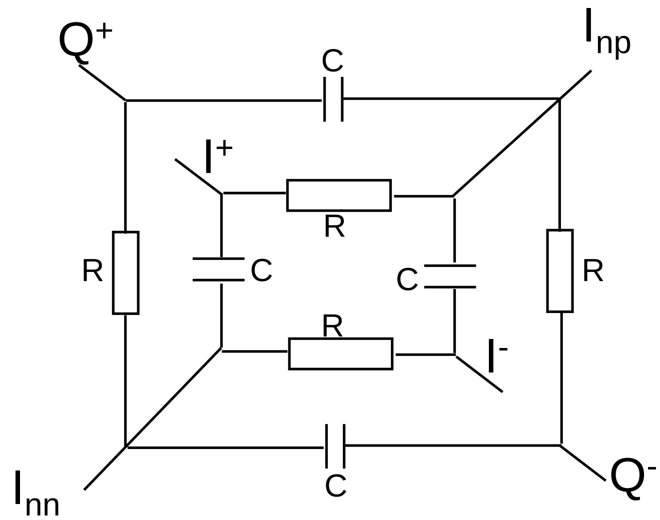
$$A: \cos(\frac{2}{3}\omega_0 t)$$

$$B: \cos(\frac{2}{3}\omega_0 t)\cos(\frac{1}{3}\omega_0 t) = \cos(\omega_0 t) + \cos(\frac{1}{3}\omega_0 t)$$

$$C: \cos(\omega_0 t)$$

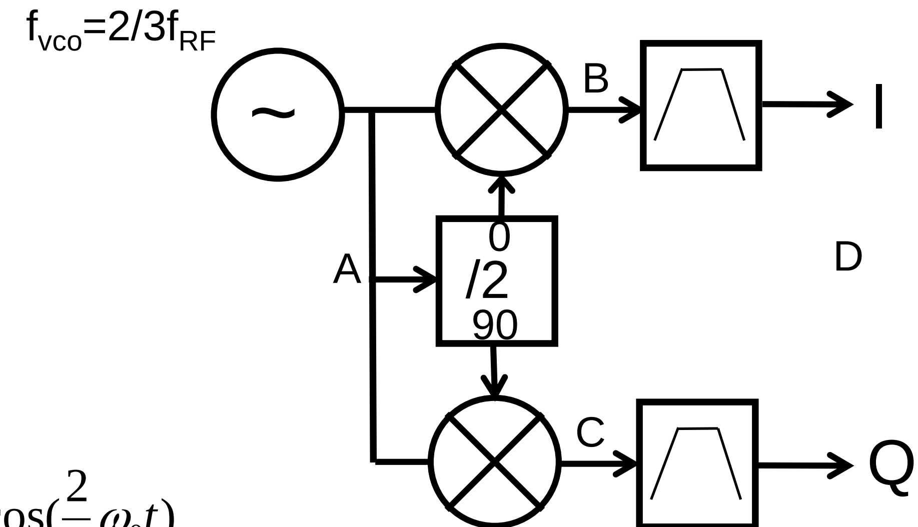
$$D: I = \cos(\omega_0 t) \quad , \; Q = \sin(\omega_0 t)$$

A poly phase filter (sometimes called a Hilbert filter), is passive in nature. One implementation looks like the following:



The problem with polyphase filters is that they are lossy. Therefore, a significant amount of current is needed to drive them. Furthermore, the phase matching expected worsen with higher frequency. A typical I/Q phase matching is 3° for 3-sigma.

A better way to build high frequency non-integer LO generation scheme is by modifying the scheme so as no polyphase filter is needed as follows:



$$A: \cos(\frac{2}{3}\omega_0 t)$$

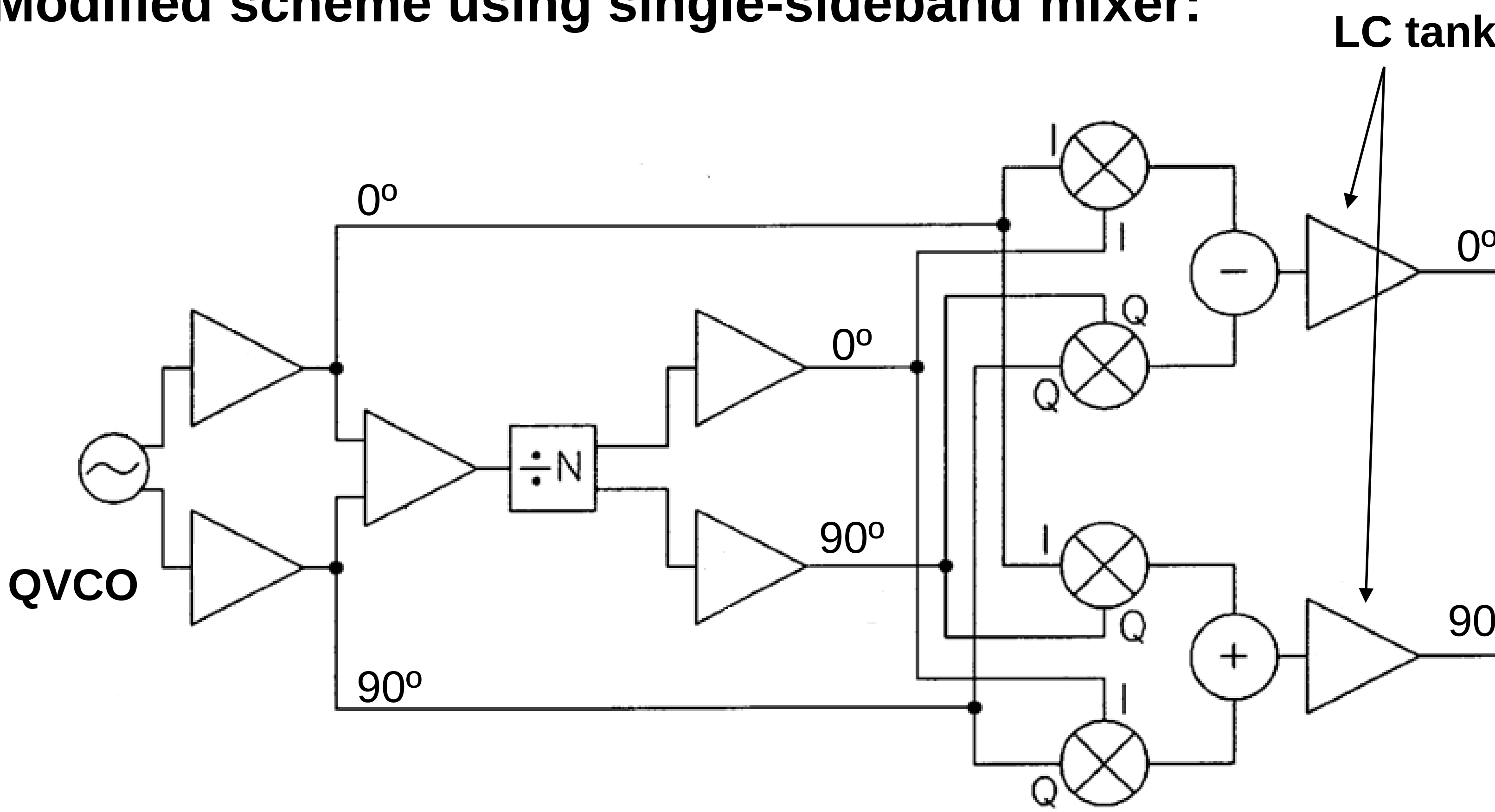
$$B: \cos(\frac{2}{3}\omega_0 t)\cos(\frac{1}{3}\omega_0 t) = \cos(\omega_0 t) + \cos(\frac{1}{3}\omega_0 t)$$

$$C: \cos(\frac{2}{3}\omega_0 t)\sin(\frac{1}{3}\omega_0 t) = \sin(\omega_0 t) - \sin(\frac{1}{3}\omega_0 t)$$

$$D: I = \cos(\omega_0 t) \quad , \ Q = \sin(\omega_0 t)$$

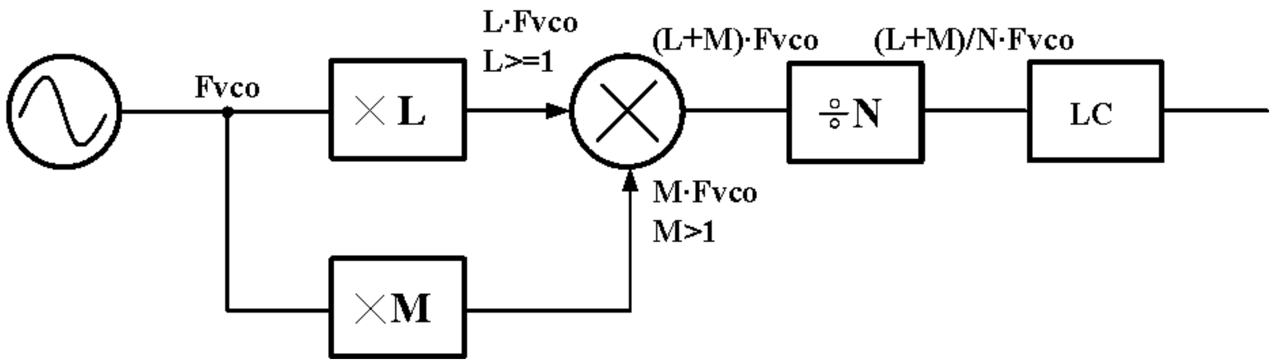
Note that only the tank rejection attenuates the unwanted mixer side band at $(1/3)\omega_0$, which is around ~20dB typical (not sufficient)

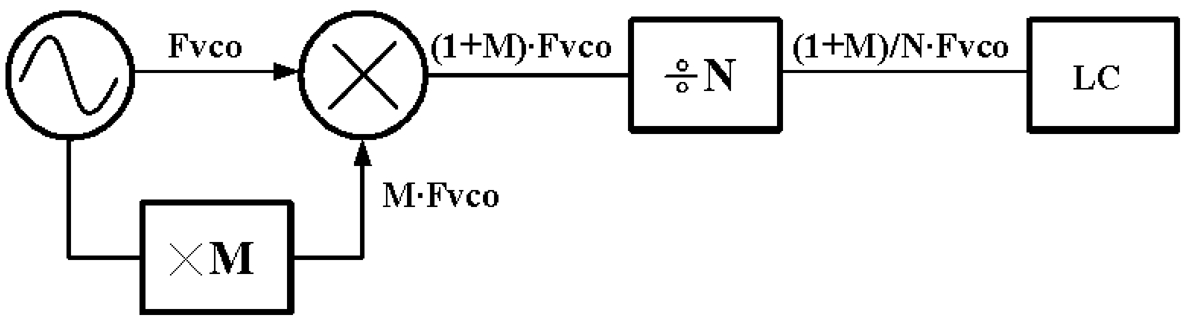
Modified scheme using single-sideband mixer:

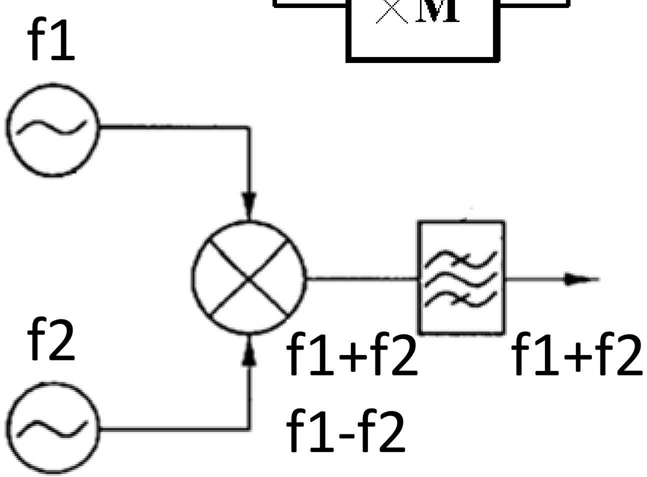


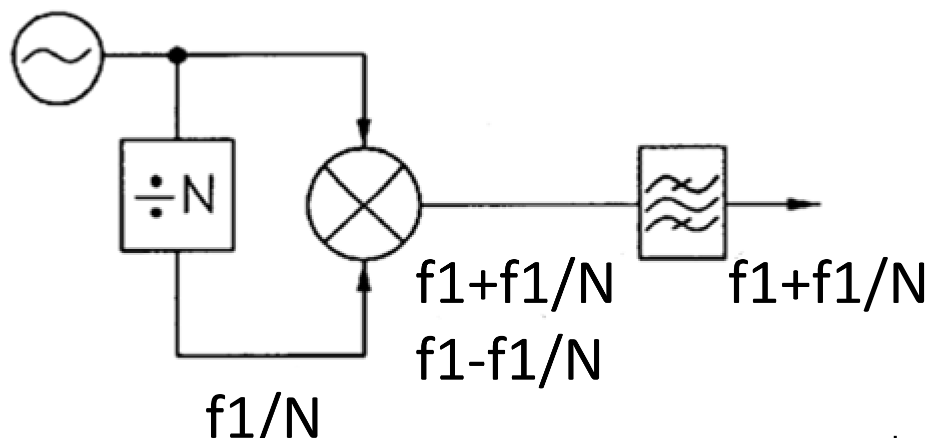
The single-sideband mixer provides additional 20~30dB rejection to the 1/3LO component in addition to the 20dB rejection provided by the LC tank at the output. The disadvantage of this scheme is you need either a quadrature VCO or generate a quadrature VCO signal from VCO output via poly-phase network

Derivatives of such none-integer LO generation schemes:

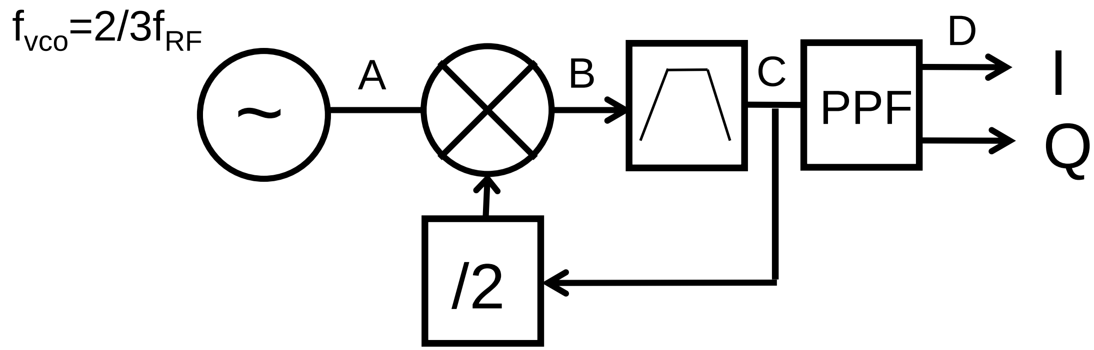








Super-regenerative LO generation scheme



$$B = \frac{C}{2} \times A \Rightarrow f_B = \frac{f_C}{2} + f_A \text{ OR } f_B = f_A - \frac{f_C}{2}$$

$$\text{if} \quad f_B = f_C \Rightarrow f_B = 0.5 f_A \text{ OR } f_B = 1.5 f_A$$

$$A : \cos(\frac{2}{3}\omega_0 t)$$

$$B : \cos(\frac{1}{3}\omega_0 t) \text{ OR } \cos(\omega_0 t)$$

$$C : \cos(\omega_0 t)$$

$$D : I = \cos(\omega_0 t) \quad , \ Q = \sin(\omega_0 t)$$

The super-regenerative scheme is used in ADI's *Othello* direct-conversion receiver series. The main disadvantages of such scheme are:

● start-up. The divider can start oscillation due to its positive feedback nature way before the VCO starts oscillation causing an undesired latchup output

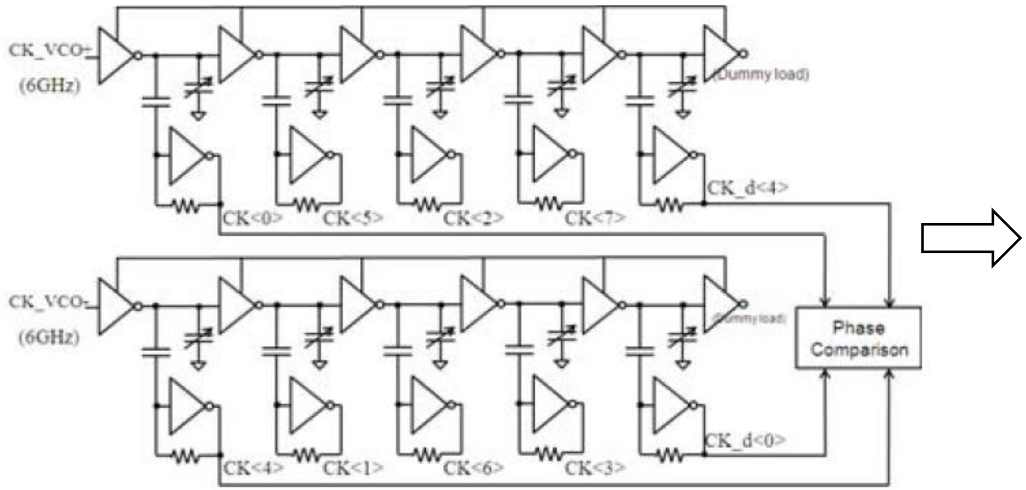
● an image reject mixer is needed to ensure proper operation

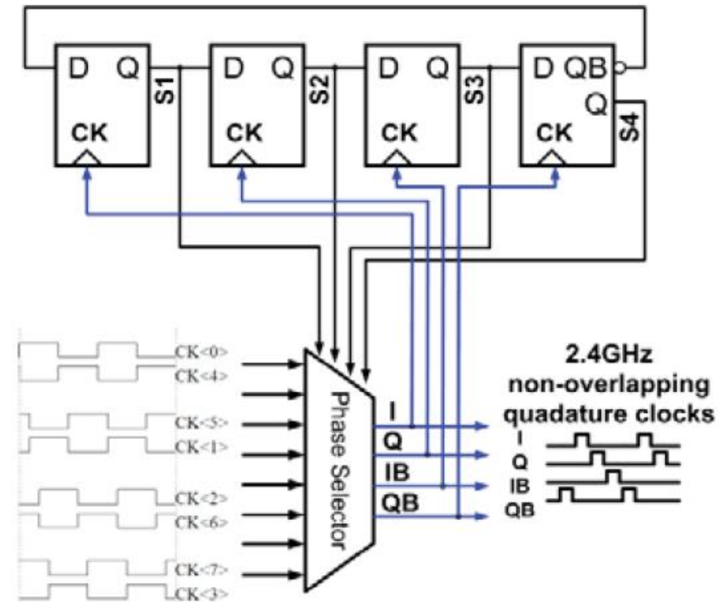
● output spurs are no better than 30dBc.

● poloyphase filter is needed for I/Q output

Voltage Controlled Delay-Line (VCDL) as LOgen:

Yi-Shing Shih, ASSCC 2013





In this example $f_{VCO} = 5/2\ f_{LO}$

- VCO frequency is passed through a delay line to generate 8-equally spaced phases (CK<0:7>). Delay line is calibrated over frequency and temp
- the 8-phases are grouped into a group of 4 complementary phases and fed to a mux (phase selector). Four DFF are connected at the output of the phase selector and generate the S-signals that controls which signal of the phase selector is selected.
- The idea is to select different phases at different times to generate none-overlapping I/Q LO that is 2/5 in frequency compared to VCO's
- >30dB image rejection is achieved

Which LO generation scheme to use? Welcome to the world of frequency planning

- Frequency planning is perhaps one of the most important tasks in an SoC architecture, especially with multiple radios involved in a phone.
- A phone for example has several cellular radios (GSM 2G, 3G, 4G LTE and 5G), a GPS radio, an FM radio, a Bluetooth and WiFi radios. All these need to coexist.
- Each one of these radios can be an aggressor and also it can be a victim.
- As an aggressor, the radio emits Tx power, Tx harmonics, VCO tone, and LO and LO generation spurs.
- For the victim radio, you need to make sure all other radios (as aggressors) do not have anything that falls in band of the victim radio or cause severe desense.
- Usually a matrix listing all frequencies of all these radios are listed and potential problems identified and so different LO scheme is used or an isolation requirement is put in place

References:

[1] A. Behzad et. al., “A direct conversion CMOS transceiver utilizing automatic frequency control for the IEEE 802.11a wireless LAN standard, “ IEEE JSSC, Vol. 38, No. 12, December 2002, pp. 2209-2220.

[2] P. Stroet, Rishi Mohindra, S. Hahn, A. Schurr, E. Riou, "A Zero-IF Single-Chip Tranceiver for up to 22Mb/s QPSK 802.11b Wireless LAN," IEEE ISSCC 2001, pp. 204-205

[3] H. Darrabi et. al., "A 2.4GHz CMOS Transceiver for Bluetooth," IEEE ISSCC 2001, 200-201

[4] A. Ajjikuttira et. al. "A Fully Integrated CMOS RFIC for Bluetooth Applications," IEEE ISSCC 2001, 198-199

[5] F. Eynde et. al., "A fully integrated single-chip SoC for Bluetooth," IEEE ISSCC 2001, 196-197

[6] D. Guermandi, P. Tortori, E. Franchi, A. Gnudi, “A 0.75 to 2.2GHz Continuously-Tunable Quadrature VCO,“ *in Proc. IEEE ISSCC*, 2005, pp. 536-537.

[7] Yi-Shing Shih, *et. al*., “A 55nm 0.6mm^2 Bluetooth SoC Integrated in Cellular Baseband Chip with Enhanced Co-existence,“ *in Proc. IEEE ASSCC*, 2013, pp. 193-196.