Transceiver Architectures (2/2)

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Trigonometric Functions

$A\cos(\omega_C t) = (A/2)e^{+j\omega_C t} + (A/2)e^{-j\omega_C t}$

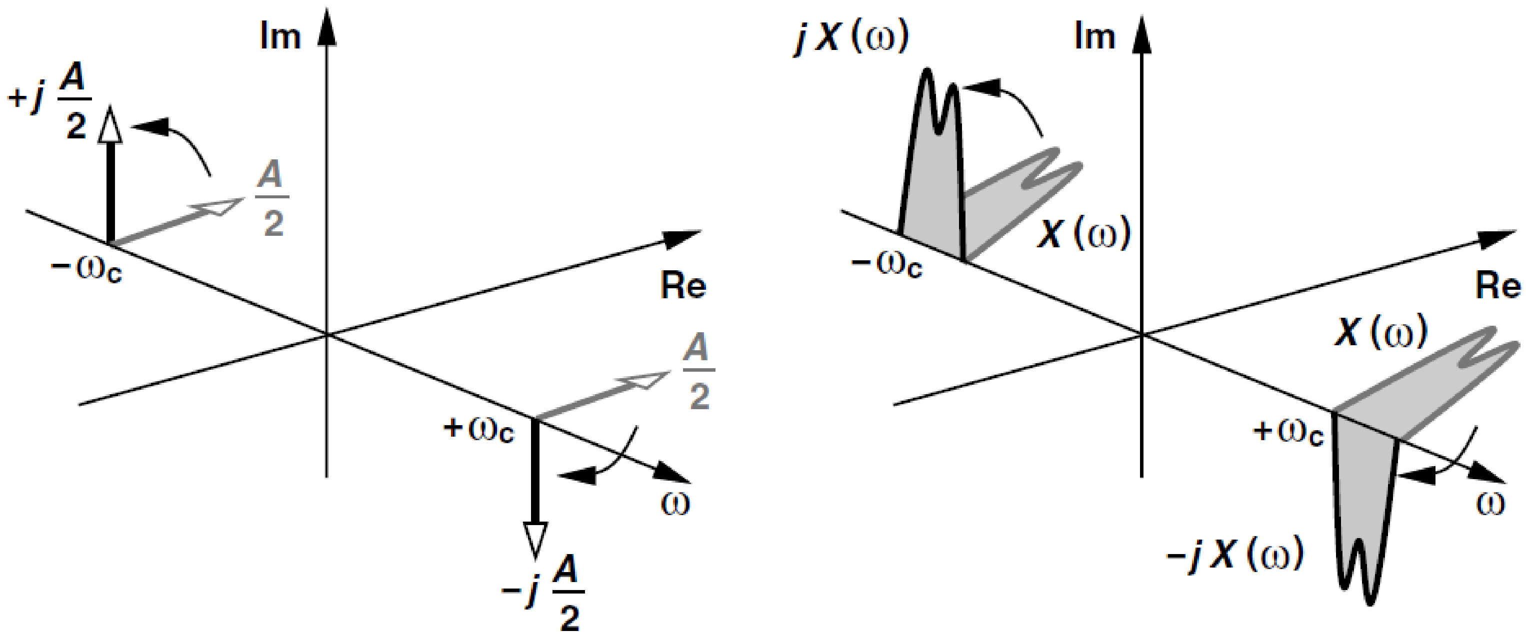
- **The impulses at both $+\omega_c$ and $-\omega_c$ are multiplied by 1/2**
- **The phases of impulses at both $+\omega_c$ and $-\omega_c$ are unchanged**

$$A\cos(\omega_c t - 90^\circ) = A\frac{e^{+j(\omega_c t - 90^\circ)} + e^{-j(\omega_c t - 90^\circ)}}{2}$$
$$= -\frac{A}{2}je^{+j\omega_c t} + \frac{A}{2}je^{-j\omega_c t}$$
$$= A\sin\omega_c t$$

- **The impulse at $+\omega_c$ is multiplied by –j/2 and that at $-\omega_c$ by +j/2**
- **The impulse at $+\omega_c$ is rotated clockwise and that at $-\omega_c$ counterclockwise.**

[B. Razavi, RF Microelectronics]

Image-Reject Receivers

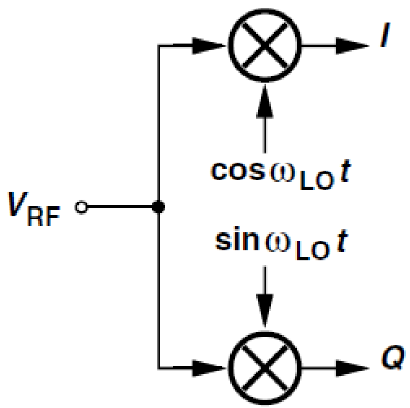


❑**Image-reject architectures are another class of receivers that suppress the image without filtering, which can be realized with a 90-degree phase shift.**

[B. Razavi, RF Microelectronics]

Quadrature Downconvert

- If $\omega_c < \omega_{LO}$, then:



$$\cos(\omega_C t)\cos(\omega_{LO} t)$$

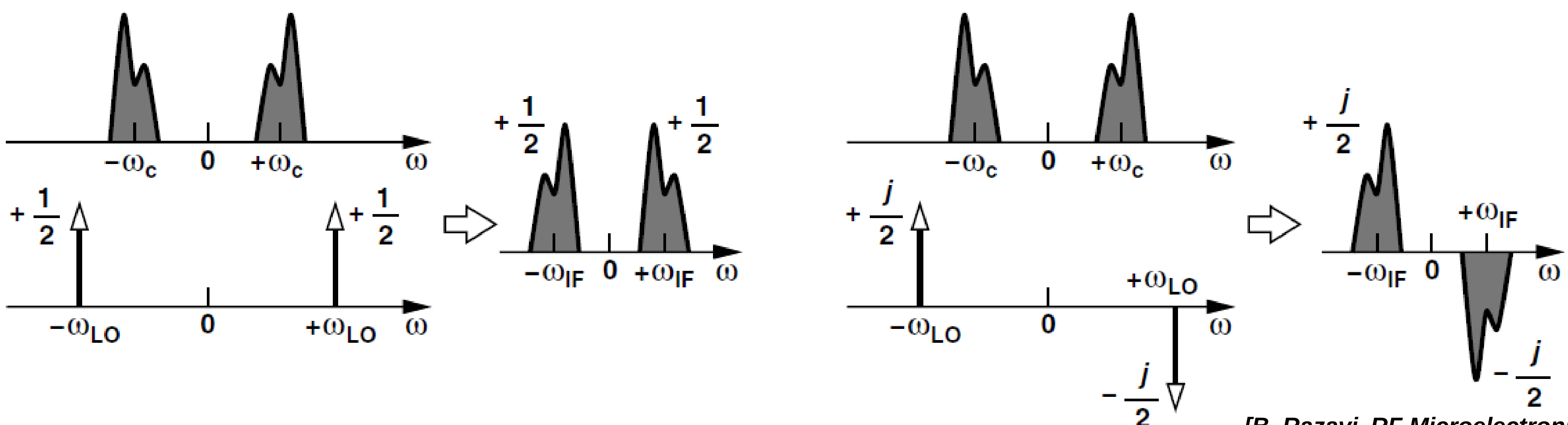
$$= [\cos(\omega_C+\omega_{LO})t + \cos(\omega_C-\omega_{LO})t]/2$$

$$\rightarrow\ (1/2)e^{+j(\omega_C-\omega_{LO})t} + (1/2)e^{-j(\omega_C-\omega_{LO})t}$$

$$= (1/2)e^{-j\omega_{IF}t} + (1/2)e^{j\omega_{IF}t} \quad (\omega_{IF}=\omega_{LO}-\omega_C)$$

$$\cos(\omega_C t)\sin(\omega_{LO} t)$$

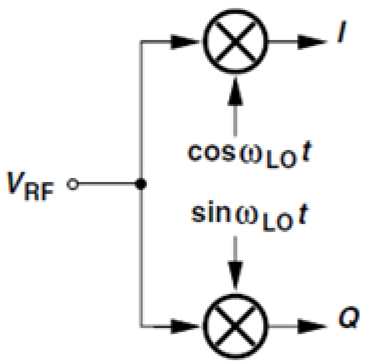
$$= (1/2)je^{-j\omega_{IF}t} + (-1/2)je^{j\omega_{IF}t} \quad (\omega_{IF}=\omega_{LO}-\omega_C)$$



[B. Razavi, RF Microelectronics]

Quadrature Downconvert

- If $\omega_c > \omega_{LO}$, then:



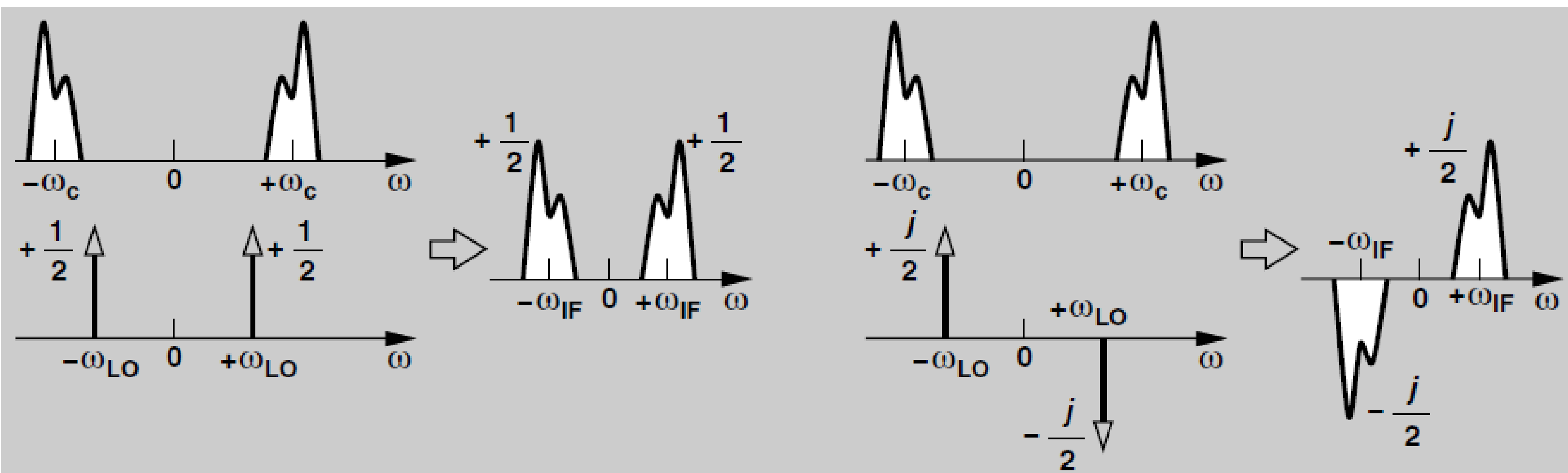
$\cos(\omega_C t)\cos(\omega_{LO} t)$

$\rightarrow (1/2)e^{+j(\omega_C-\omega_{LO})t} + (1/2)e^{-j(\omega_C-\omega_{LO})t}$

$= (1/2)e^{j\omega_{IF}t} + (1/2)e^{-j\omega_{IF}t}\quad (\omega_{IF}=\omega_C-\omega_{LO})$

$\cos(\omega_C t)\sin(\omega_{LO} t) \rightarrow (1/2)\sin(\omega_{LO}-\omega_C)t$

$= (1/2)je^{j\omega_{IF}t} + (-1/2)je^{-j\omega_{IF}t}\quad (\omega_{IF}=\omega_C-\omega_{LO})$



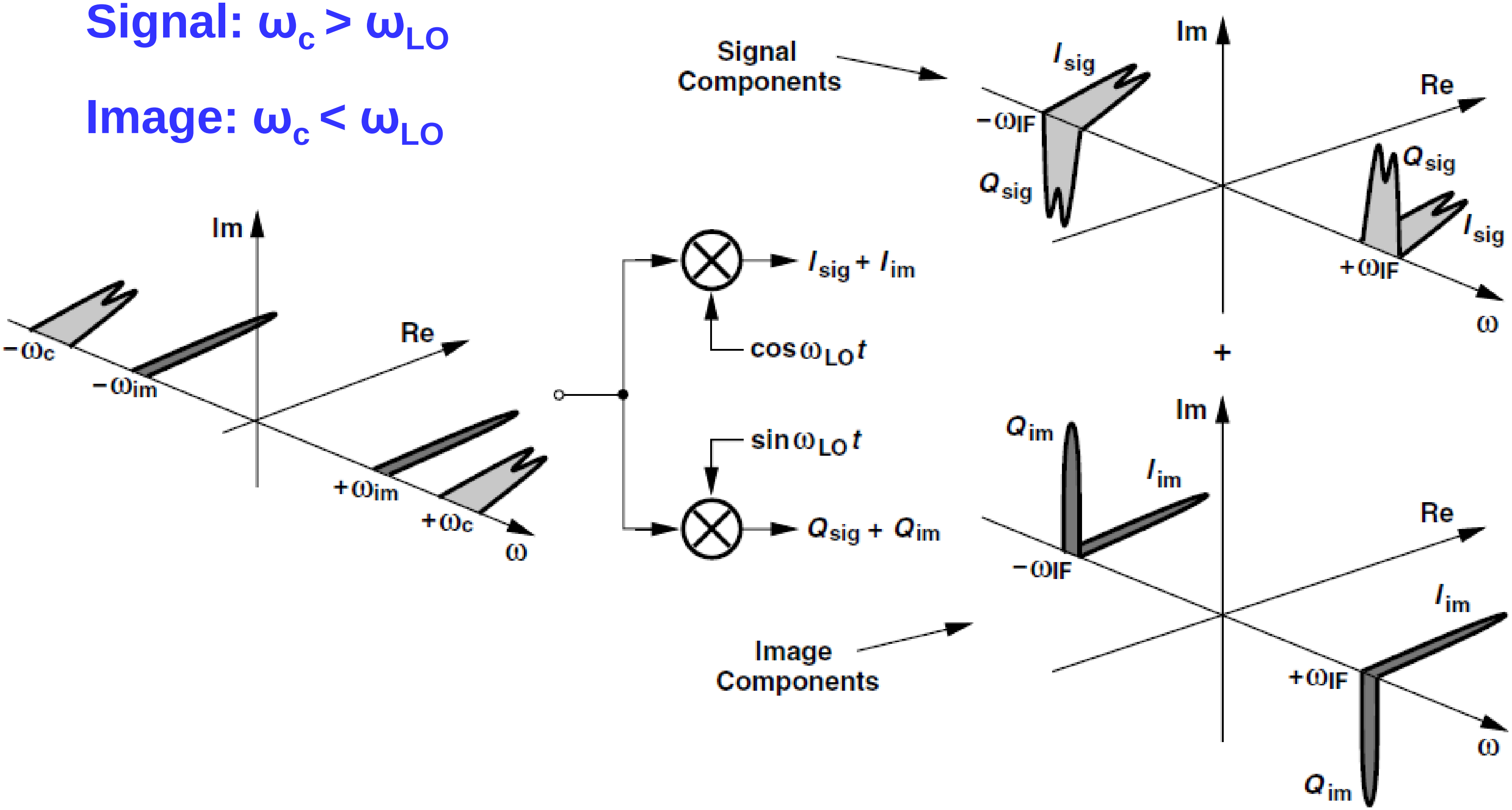
[B. Razavi, RF Microelectronics]

Quadrature Downconvert

❑As the signal and image locate at separated sides of ω_{LO} :

Signal: $\omega_c > \omega_{LO}$

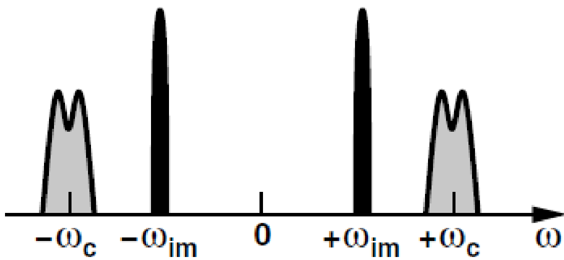
Image: $\omega_c < \omega_{LO}$

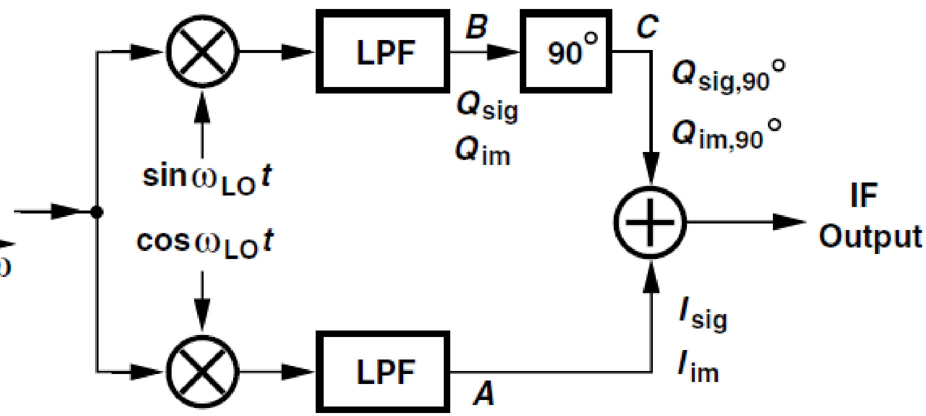


[B. Razavi, RF Microelectronics]

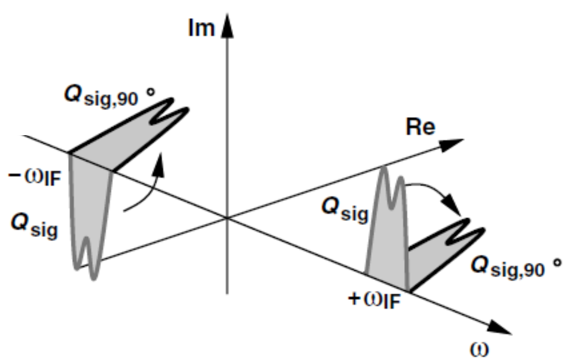
Hartley Architecture

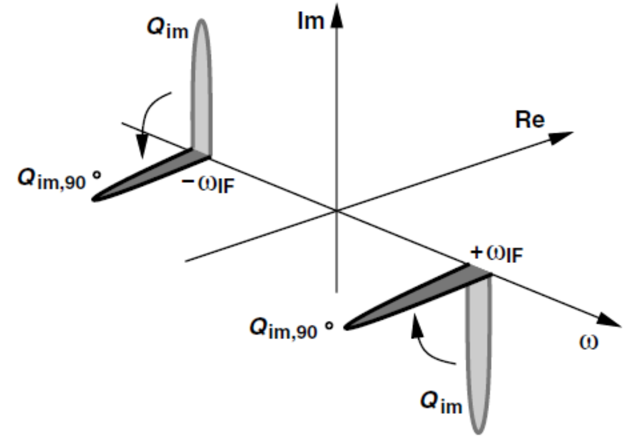
❑If we shift one path by another 90 degree before adding them, the image may be removed:





$$A\cos(\omega_c t - 90^\circ) = A\frac{e^{+j(\omega_c t - 90^\circ)} + e^{-j(\omega_c t - 90^\circ)}}{2}$$
$$= -\frac{A}{2}je^{+j\omega_c t} + \frac{A}{2}je^{-j\omega_c t}$$
$$= A\sin\omega_c t$$

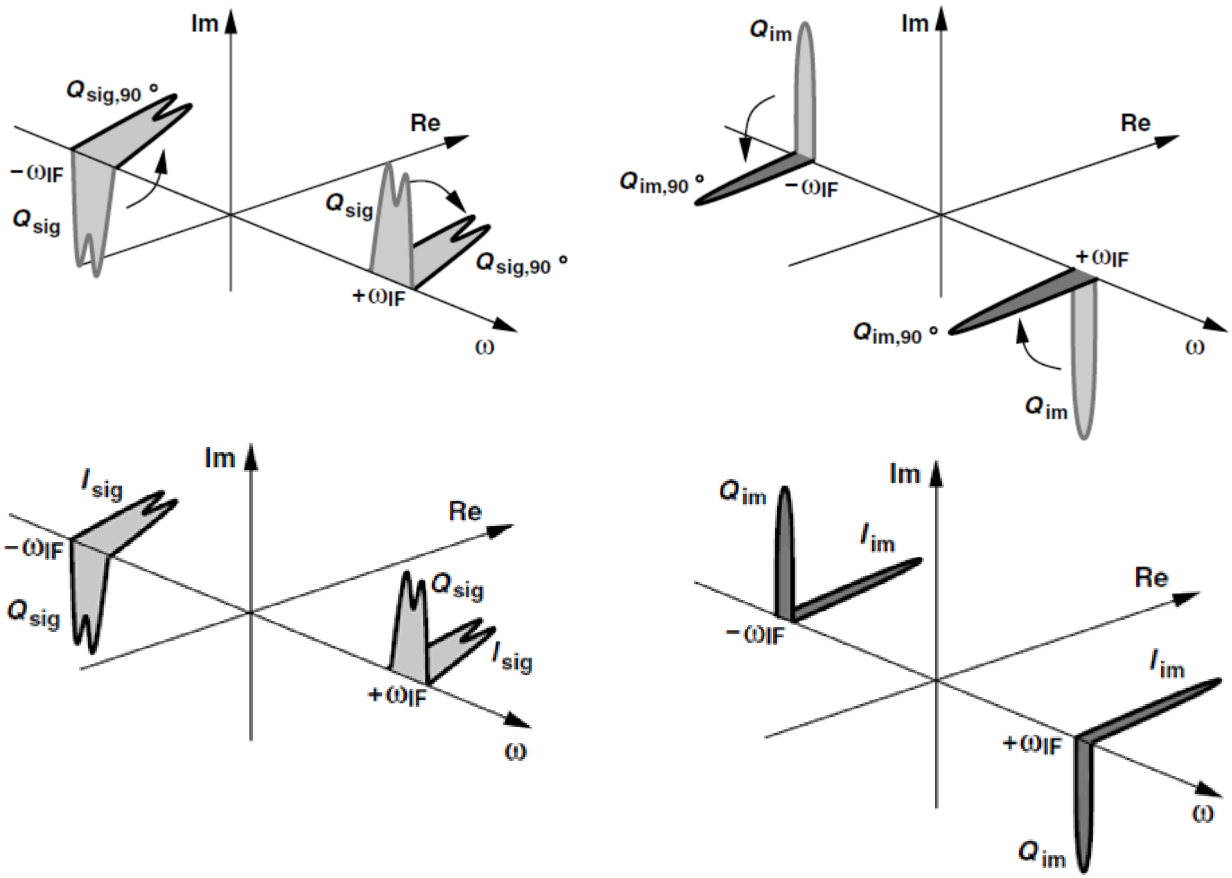




[B. Razavi, RF Microelectronics]

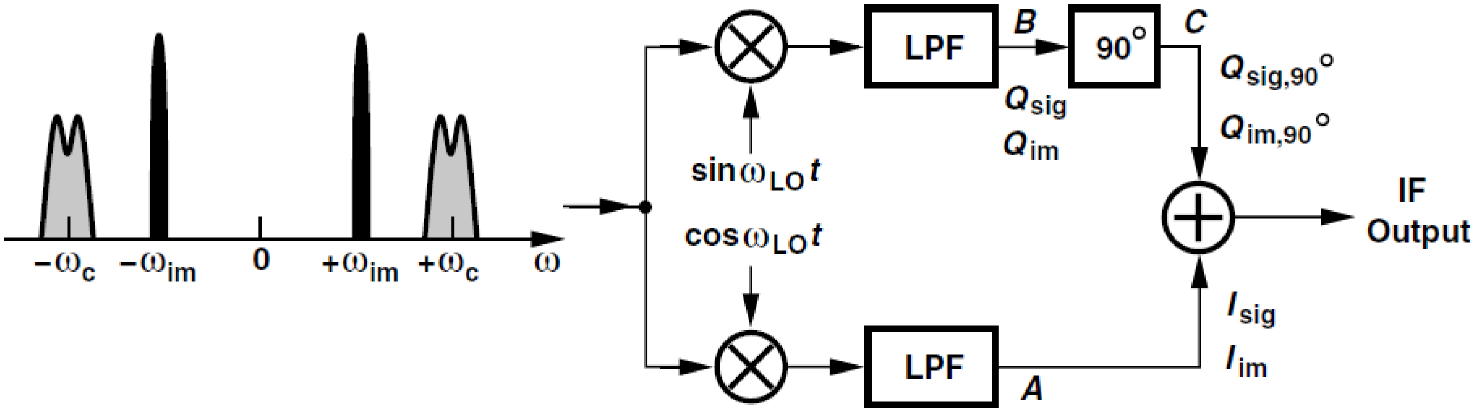
Hartley Architecture

❑The sum of I and Q results in cancellation of image:



I_{sig} and Q_{sig} Doubled **I_{im} and Q_{im} Cancelled** *[B. Razavi, RF Microelectronics]*

Hartley Architecture



❑**The input signal and image:** $x(t) = A_{sig}\cos(\omega_c t + \phi_{sig}) + A_{im}\cos(\omega_{im} t + \phi_{im})$

$$x_A(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_{LO})t + \phi_{sig}] + \frac{A_{im}}{2}\cos[(\omega_{im} - \omega_{LO})t + \phi_{im}]$$

$$x_B(t) = -\frac{A_{sig}}{2}\sin[(\omega_c - \omega_{LO})t + \phi_{sig}] - \frac{A_{im}}{2}\sin[(\omega_{im} - \omega_{LO})t + \phi_{im}]$$

$$x_C(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_{LO})t + \phi_{sig}] - \frac{A_{im}}{2}\cos[(\omega_{im} - \omega_{LO})t + \phi_{im}]$$

❑**The image-rejection ratio:** $\mathrm{IRR} = \dfrac{(1+\epsilon)^2 + 2(1+\epsilon)\cos\Delta\theta + 1}{(1+\epsilon)^2 - 2(1+\epsilon)\cos\Delta\theta + 1}$

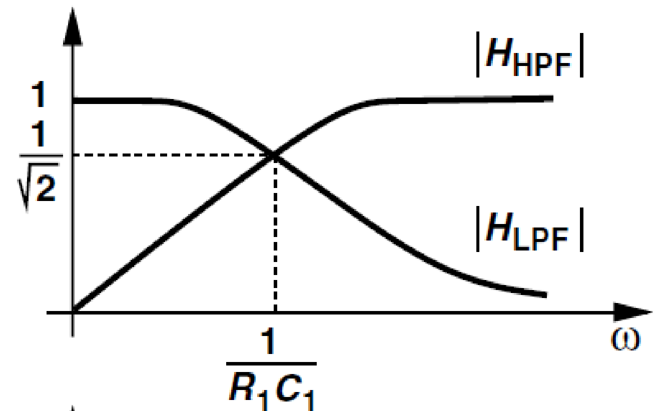
[B. Razavi, RF Microelectronics]

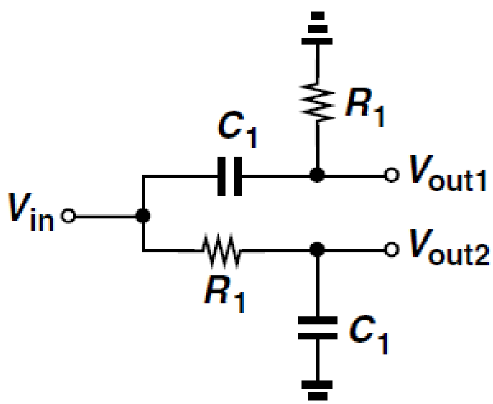
Realization of 90^o Phase Shift

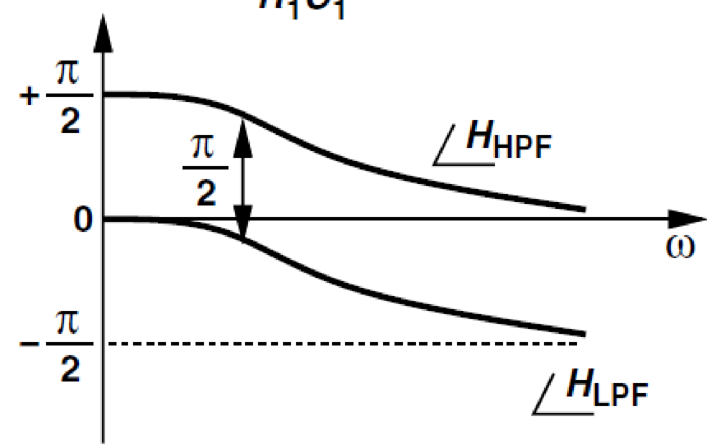
❑With a RC-CR network, the high-pass and low-pass transfer functions are respectively given by

$$H_{HPF}(s) = \frac{V_{out1}}{V_{in}} = \frac{R_1C_1s}{R_1C_1s+1}$$

$$H_{LPF}(s) = \frac{V_{out2}}{V_{in}} = \frac{1}{R_1C_1s+1}$$



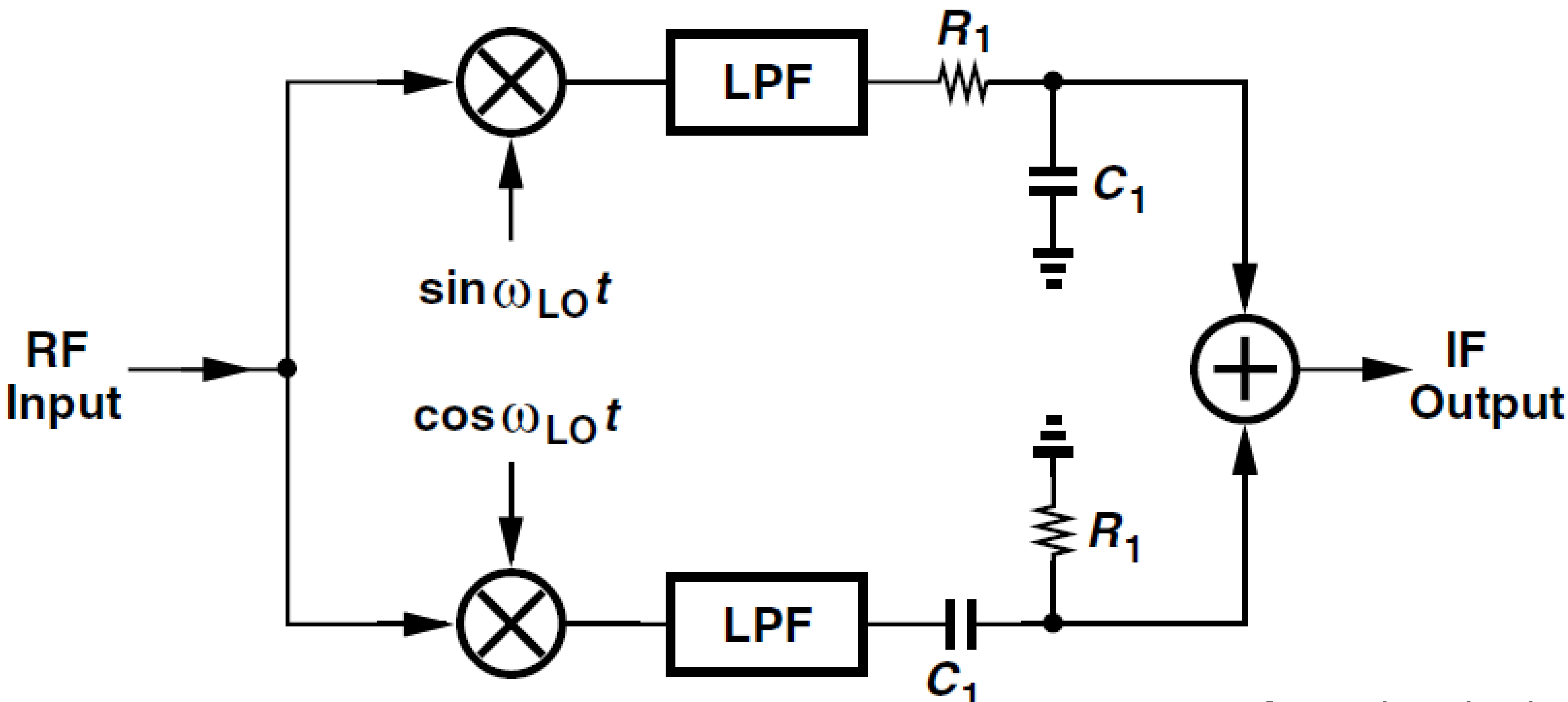




[B. Razavi, RF Microelectronics]

Hartley Architecture

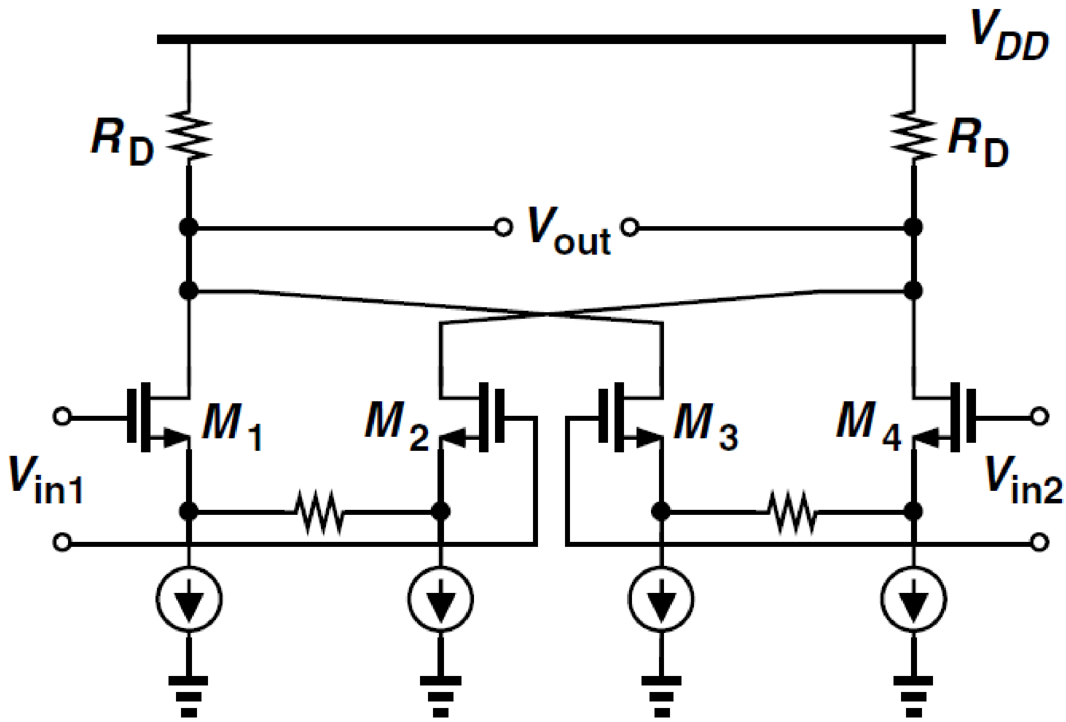
❑The 90° phase shift is typically realized as a 45° shift in one path and 45° shift in the other. This is because it is difficult to shift a single signal by 90° .



[B. Razavi, RF Microelectronics]

Voltage Adder

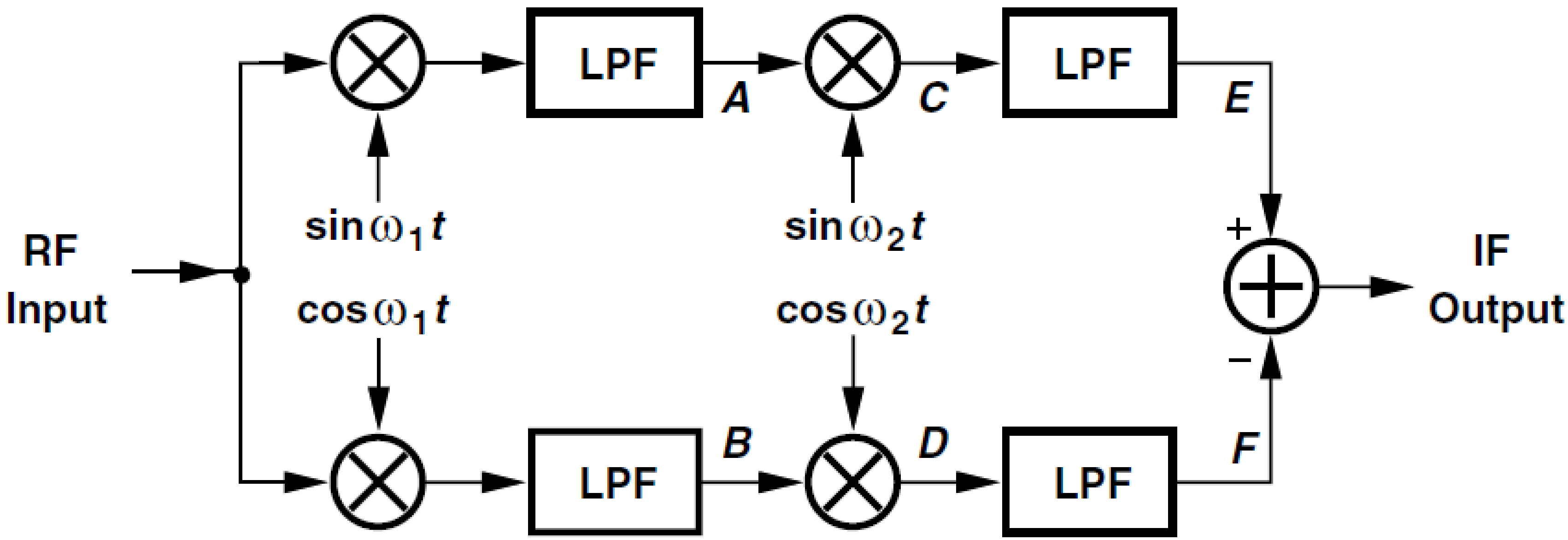
❑The summation is typically realized by differential pairs, which convert the signal voltages to currents, sum the currents, and convert the result to a voltage.



[B. Razavi, RF Microelectronics]

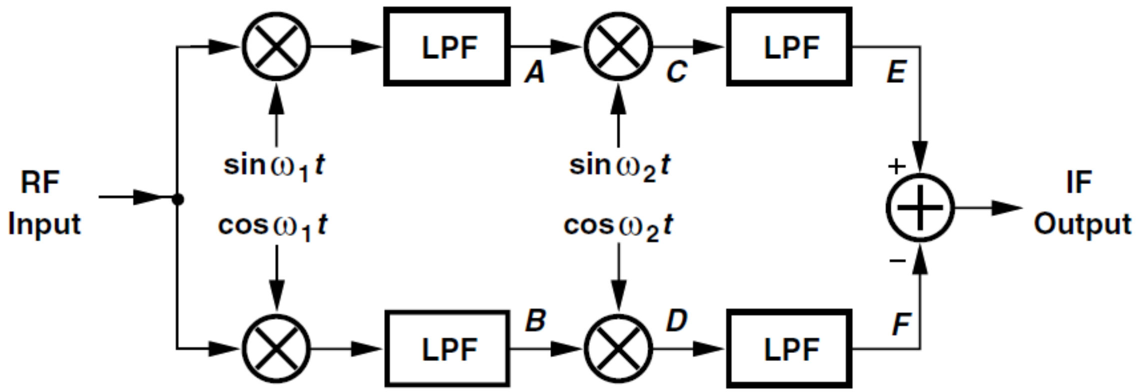
Weaver Architecture

❑The Hartley architecture has revealed mismatch issues induced by the RC phase shift networks. The Weaver receiver avoids these issues:



[B. Razavi, RF Microelectronics]

Weaver Architecture



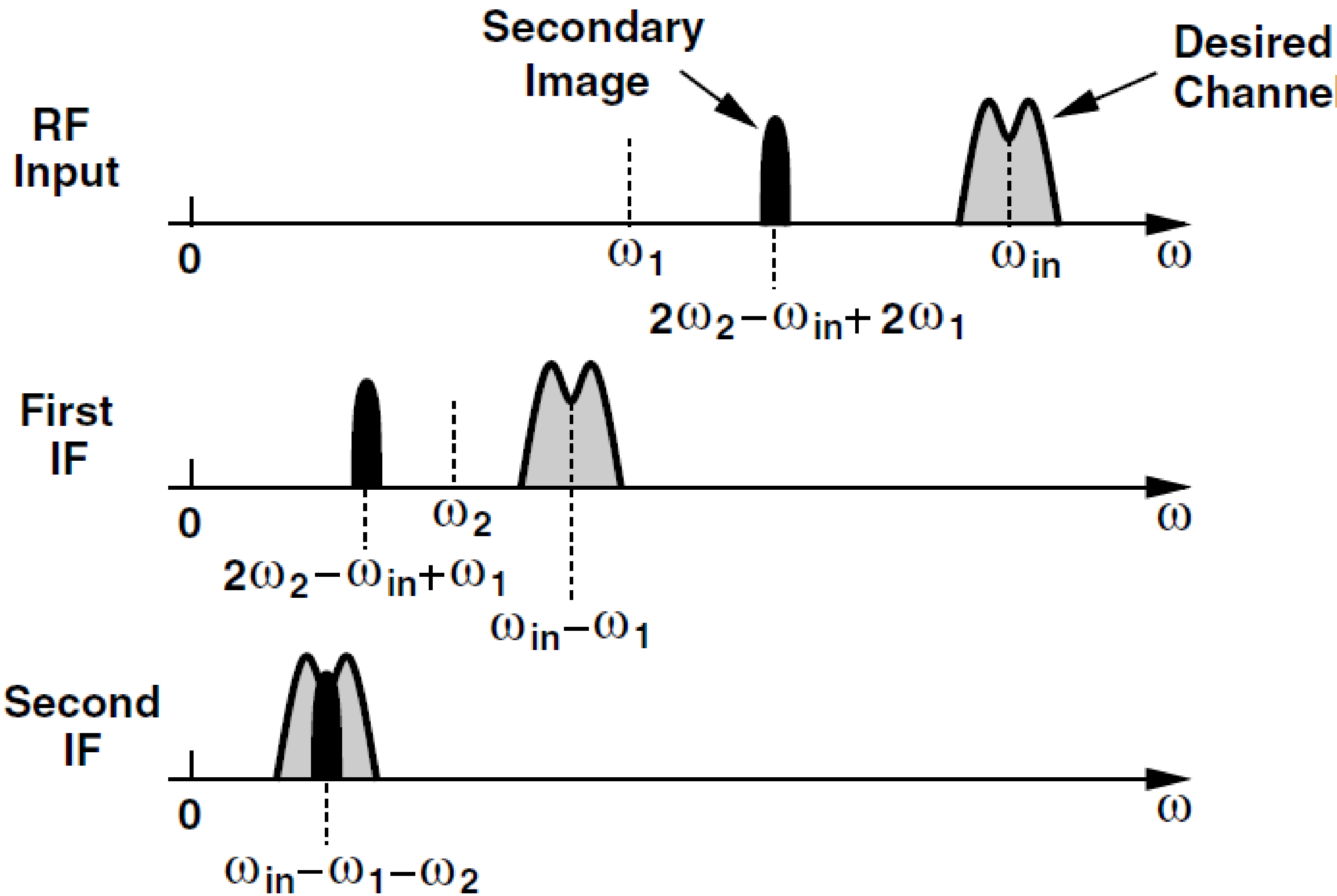
$$x_C(t) = \frac{A_{sig}}{4}\cos[(\omega_c - \omega_1 - \omega_2)t + \phi_{sig}] + \frac{A_{im}}{4}\cos[(\omega_{im} - \omega_1 - \omega_2)t + \phi_{im}]$$
$$+ \frac{A_{sig}}{4}\cos[(\omega_c - \omega_1 + \omega_2)t + \phi_{sig}] + \frac{A_{im}}{4}\cos[(\omega_{im} - \omega_1 + \omega_2)t + \phi_{im}]$$

$$x_D(t) = -\frac{A_{sig}}{4}\cos[(\omega_c - \omega_1 - \omega_2)t + \phi_{sig}] - \frac{A_{im}}{4}\cos[(\omega_{im} - \omega_1 - \omega_2)t + \phi_{im}]$$
$$+ \frac{A_{sig}}{4}\cos[(\omega_c - \omega_1 + \omega_2)t + \phi_{sig}] + \frac{A_{im}}{4}\cos[(\omega_{im} - \omega_1 + \omega_2)t + \phi_{im}]$$

$$x_E(t) - x_F(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_1 - \omega_2)t + \phi_{sig}]$$

[B. Razavi, RF Microelectronics]

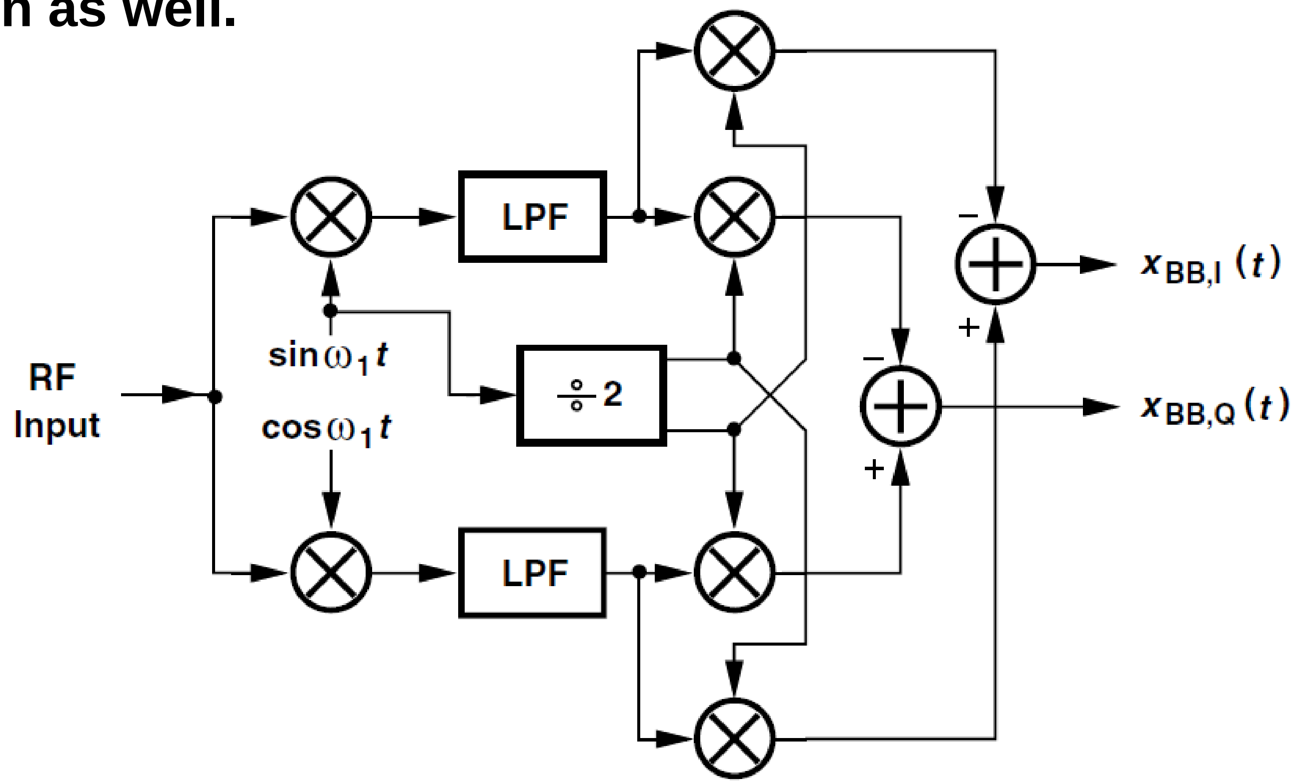
Secondary Image



[B. Razavi, RF Microelectronics]

Secondary Image

❑For this reason, the second downconversion preferably produces a zero IF, in which case it must perform quadrature separation as well.

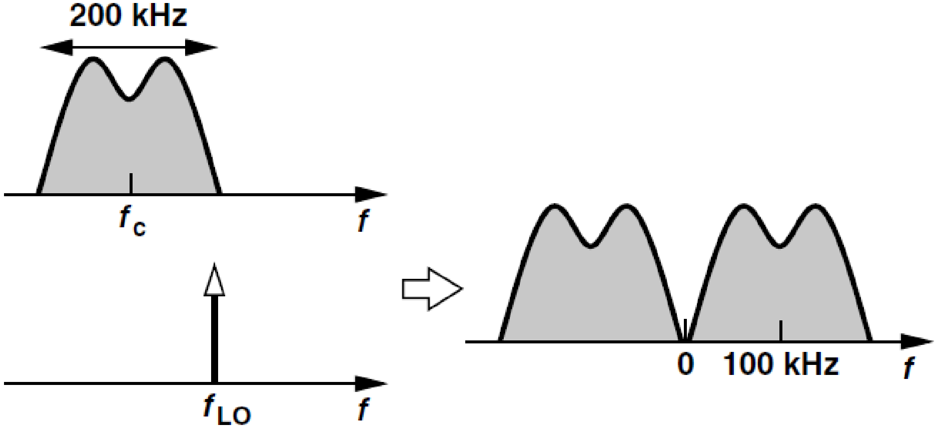


[B. Razavi, RF Microelectronics]

Low-IF RX

❑In the direct-conversion RX, converting the desired RF channel to a zero IF may significantly corrupt the signal by 1/f noise. Furthermore, the removal of the DC offset proves difficult.

❑Translating the RF signal to an low IF, the 1/f noise penalty is much less severe. Also, on-chip high-pass filtering of the signal becomes feasible.



[B. Razavi, RF Microelectronics]

Image Rejection in Low-IF RX

- The Weaver architecture must deal with the secondary image if the second IF is not zero

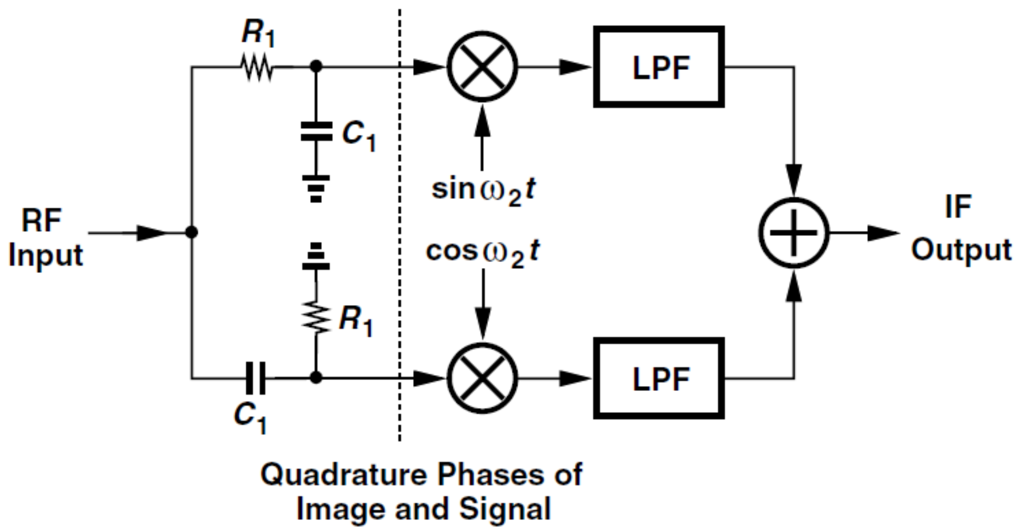
- There are still flicker noise issue in low-IF RX.

- The Hartley architecture employing the RC-CR network appears to be a candidate, but the IF spectrum in a low-IF RX may extend to zero frequency, making it impossible to maintain a high IRR across the signal bandwidth. (The high-pass Section exhibits zero gain near 0 frequency!)

[B. Razavi, RF Microelectronics]

Phase Shift in RF

❑By moving the 90-degree phase shift in the Hartley architecture from the IF path to the RF path, the RC-CR network is centered at a high frequency and can maintain a reasonable IRR across the band.



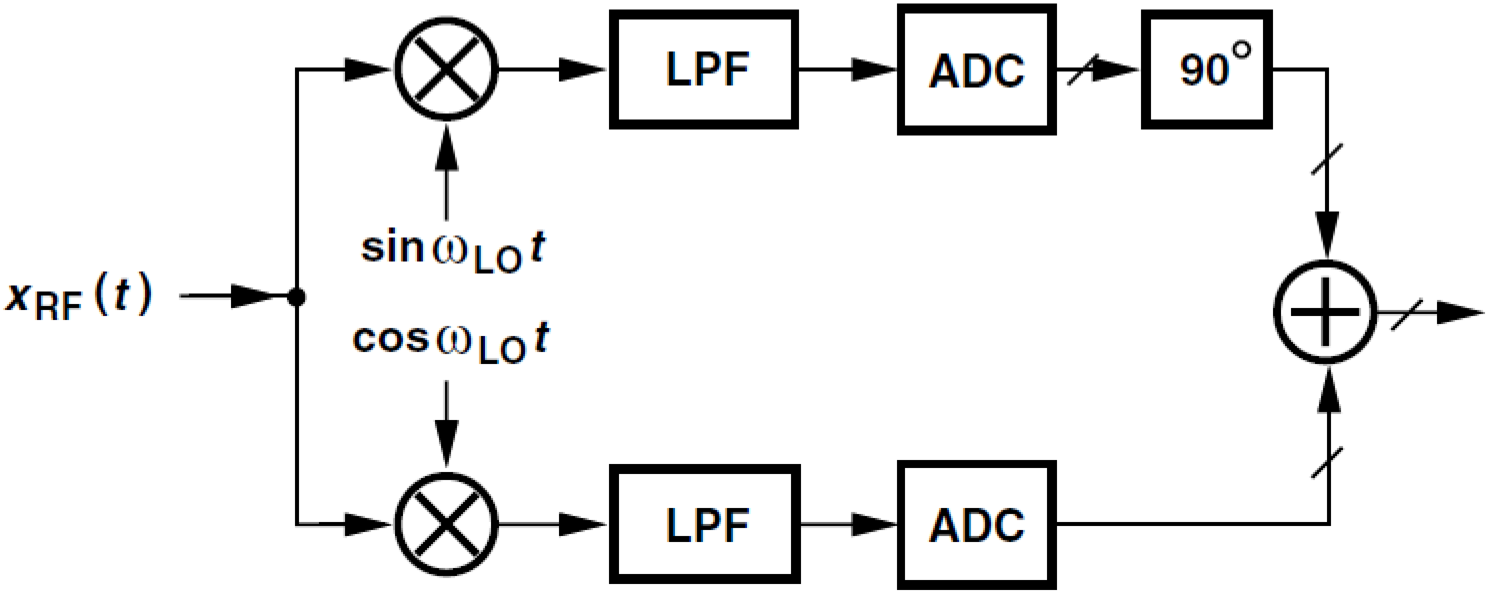
❑However, the variation of R_1 and C_1 limits the IRR to 20 dB

❑How about noise figure?

[B. Razavi, RF Microelectronics]

Phase Shift in Digital

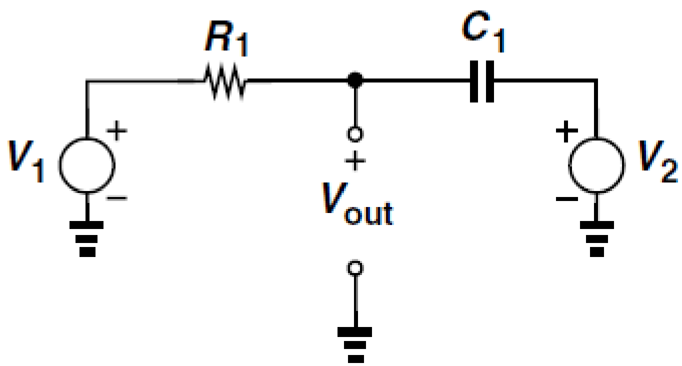
❑The phase shift and summation can be conducted in the digital domain after ADC:



❑The power consumption of ADC is higher than that in direct-conversion RX due to the higher IF.

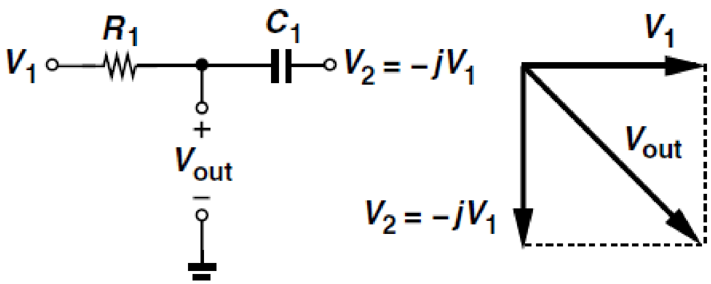
[B. Razavi, RF Microelectronics]

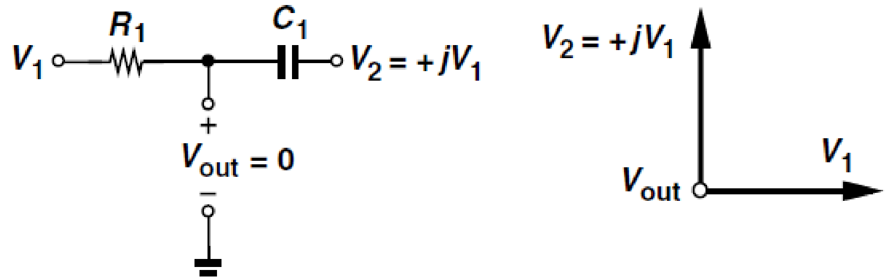
Polyphase (Complex) Filters



$$V_{out} = \frac{V_1 + R_1 C_1 s V_2}{R_1 C_1 s + 1}$$

$$\omega = \frac{1}{R_1 C_1}$$

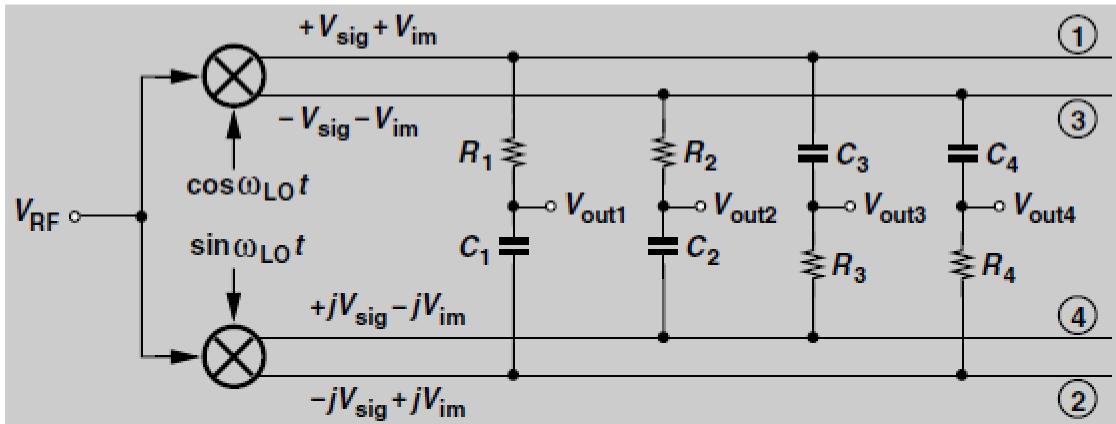


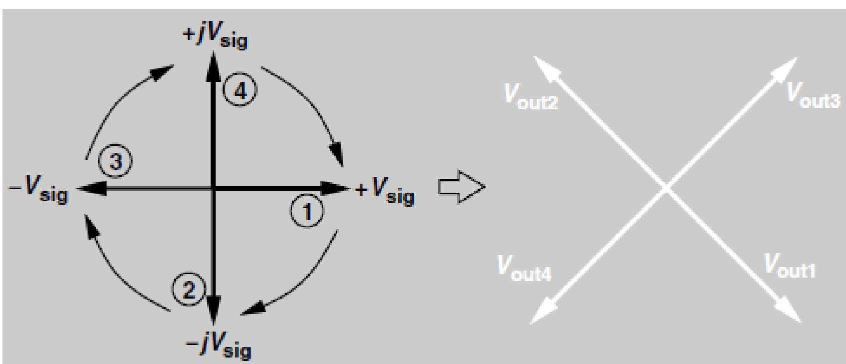


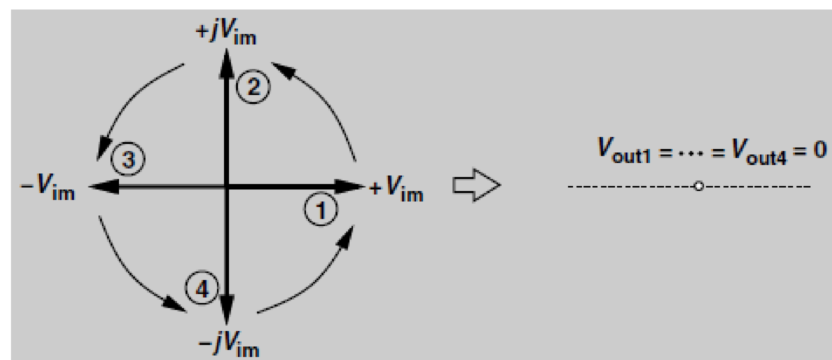
❑If the former is set for the desired signal and the latter is set for image, then the image can be filtered out.

[B. Razavi, RF Microelectronics]

Polyphase Filters



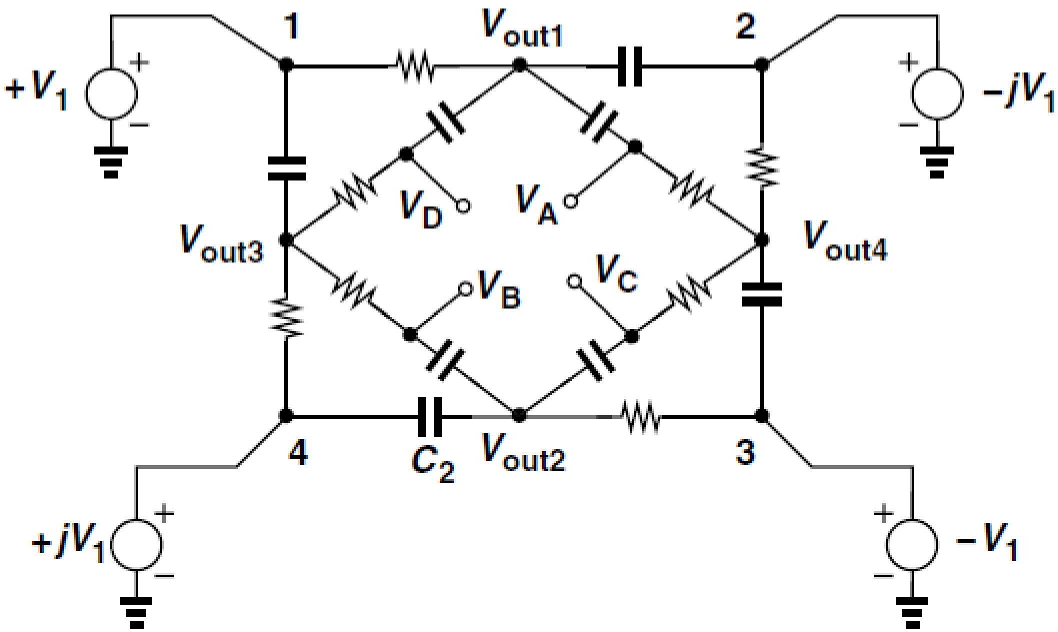


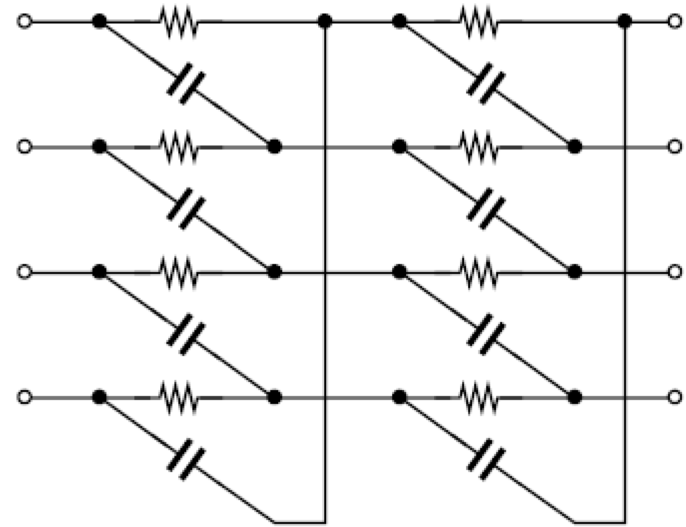


❑**The image components, on the other hand, yield a zero output.**

[B. Razavi, RF Microelectronics]

Cascaded Polyphase Sections





❑The image can be further suppressed.

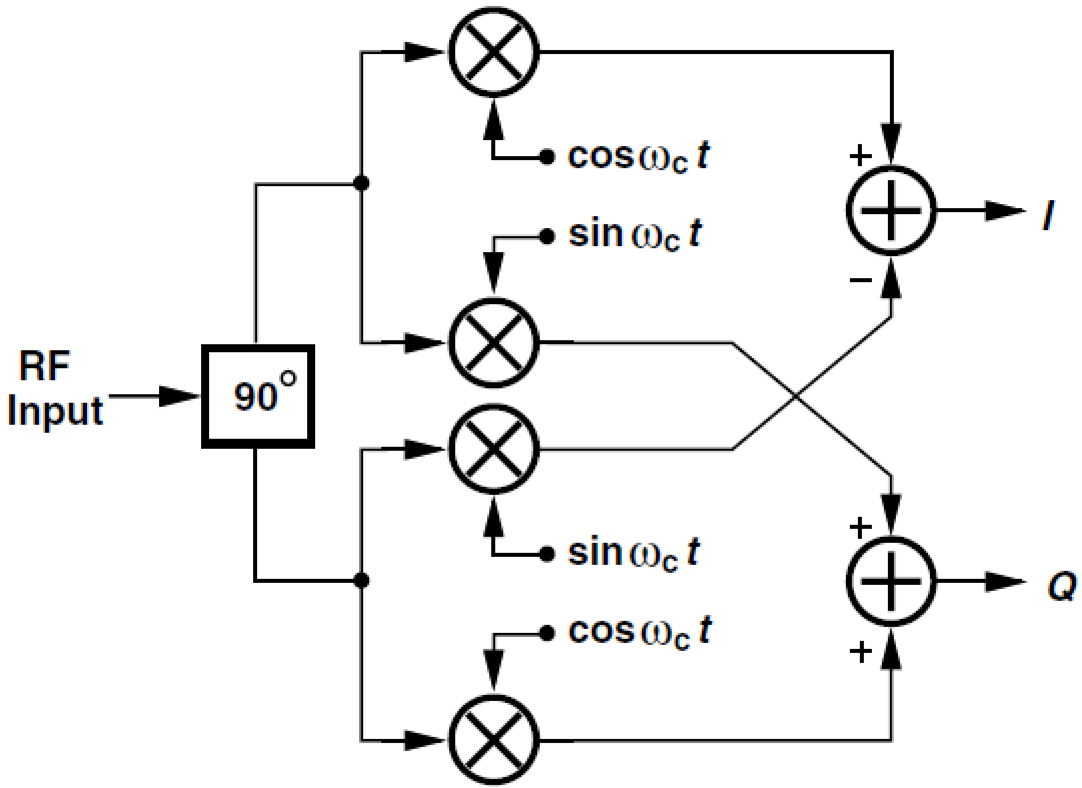
❑The poles of each stage can be set to determine the bandwidth and filter topologies (Butterworths, Chebyshev, etc.)

❑Polyphase filtering can also be done in digital domain!

[B. Razavi, RF Microelectronics]

Double Quadrature Conversion

❑"Double-quadrature" downconversion helps to reduce the effect of mismatches.



❑The circuit decomposes the RF signal into quadrature components, performs quadrature downconversion on each of the RF components, and subtracts and adds the results to produce net quadrature IF outputs.

[B. Razavi, RF Microelectronics]

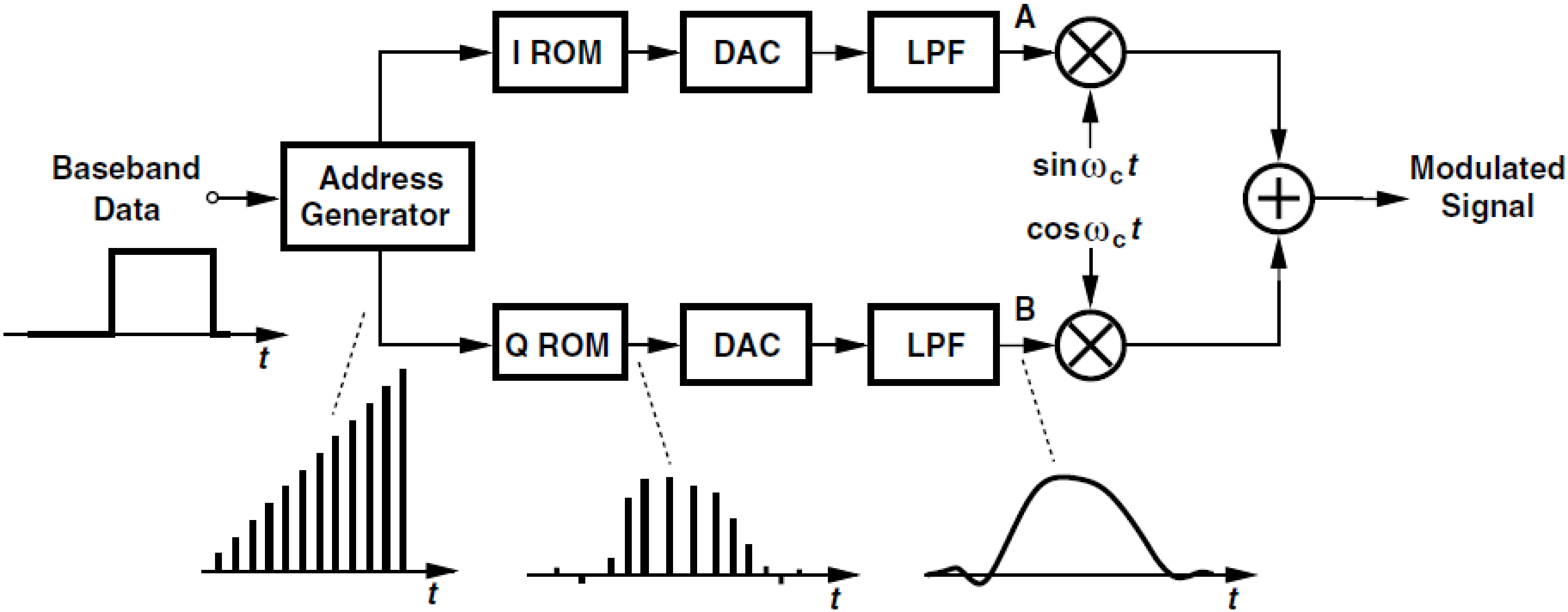
Summary of Low-IF RX

- ❑DC offset reduction (low IF) ☺
- ❑Low flicker noise ☺
- ❑Low even-order distortion ☺
- ❑Low low-frequency interferences ☺
- ❑Lower power than heterodyne RX ☺

- ❑Image issue (Phase shift in RF or digital domains, Polyphase filter, mismatch, double conversion) ☹
- ❑Higher power than direct-conversion RX ☹

[B. Razavi, RF Microelectronics]

Basic Transmitter Architecture



❑The GMSK waveform in GSM can be expanded as

$$x_{GMSK}(t) = A\cos[\omega_c t + m\int x_{BB}(t) * h(t)dt]$$

$$= A\cos\omega_c t\cos\phi - A\sin\omega_c t\sin\phi$$

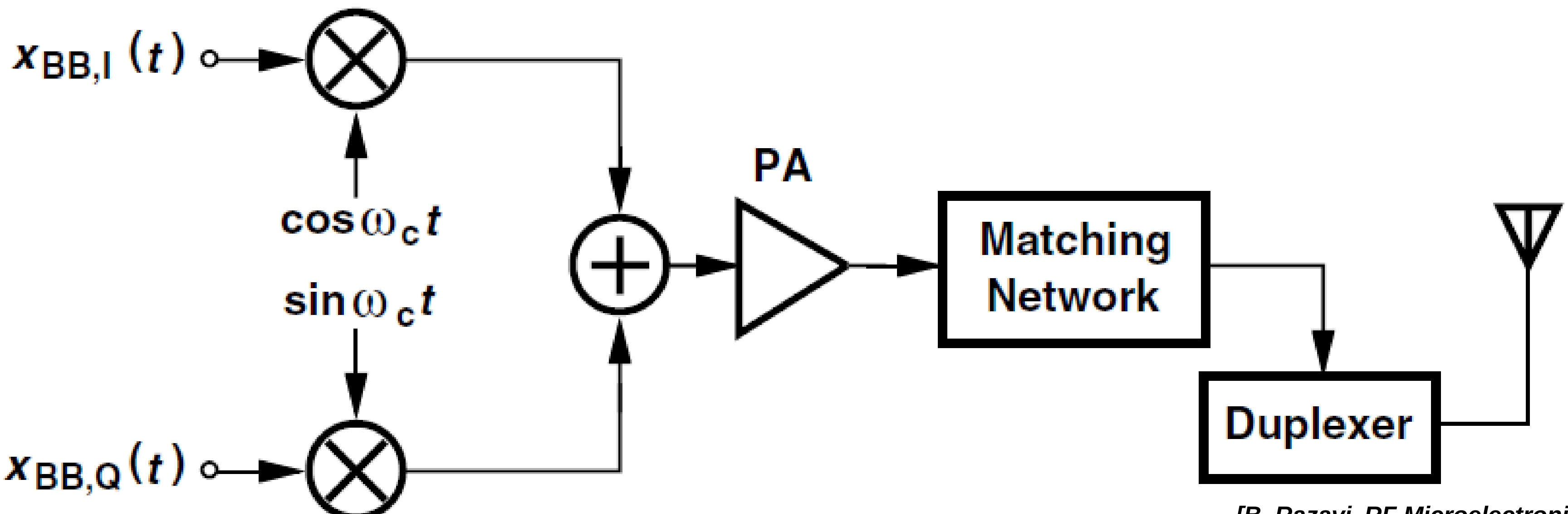
[B. Razavi, RF Microelectronics]

Direct-Conversion Transmitter

$$x(t) = A(t)\cos[\omega_c t + \phi(t)]$$
$$= A(t)\cos\omega_c t\cos[\phi(t)] - A(t)\sin\omega_c t\sin[\phi(t)]$$

$$x_{BB,I}(t) = A(t)\cos[\phi(t)]$$
$$x_{BB,Q}(t) = A(t)\sin[\phi(t)]$$

❑A "direct-conversion" transmitter directly translates the baseband spectrum to the RF carrier by means of a "quadrature upconverter."



[B. Razavi, RF Microelectronics]

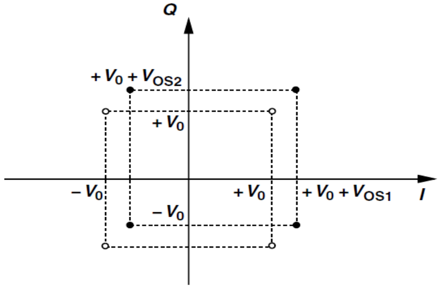
I/Q Mismatch & Carrier Leakage

- ❑The I/Q mismatch in transmitter will enlarge the error vector magnitude (EVM) of the transmitting signals.
- ❑Similar to a receiver, the unwanted sideband the “image” lead to unwanted transmitting signals.
- ❑The analog baseband circuitry producing the quadrature signals in the transmitter exhibits DC offsets

$$V_{out}(t) = [A(t)\cos\phi + V_{OS1}]\cos\omega_c t - [A(t)\sin\phi + V_{OS2}]\sin\omega_c t$$

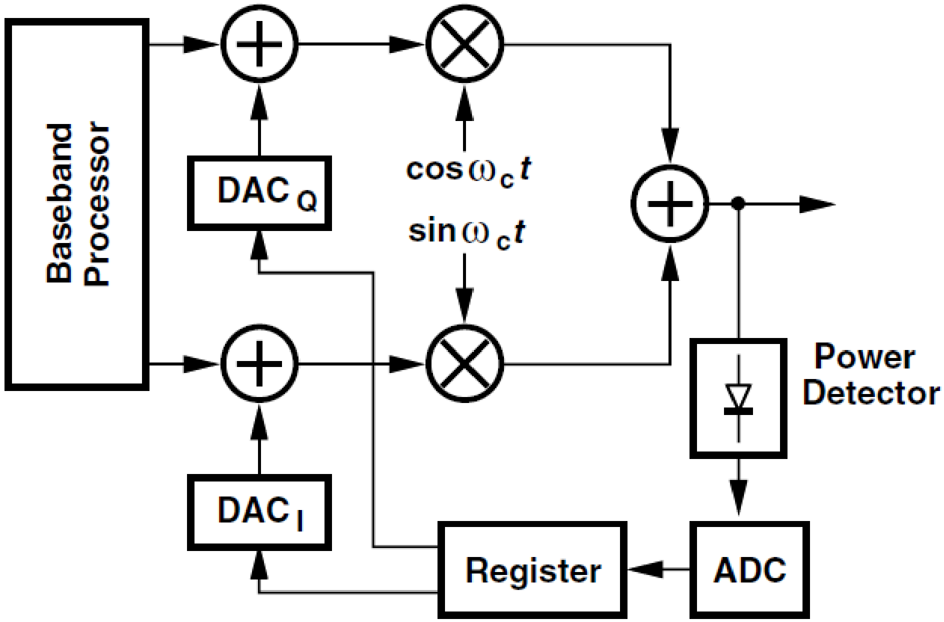
$$V_{out}(t) = A(t)\cos(\omega_c t + \phi) + V_{OS1}\cos\omega_c t - V_{OS2}\sin\omega_c t$$

$$\text{Relative Carrier Leakage} = \frac{\sqrt{V_{OS1}^2 + V_{OS2}^2}}{\sqrt{\overline{A^2(t)}}}$$



[B. Razavi, RF Microelectronics]

Baseband Control



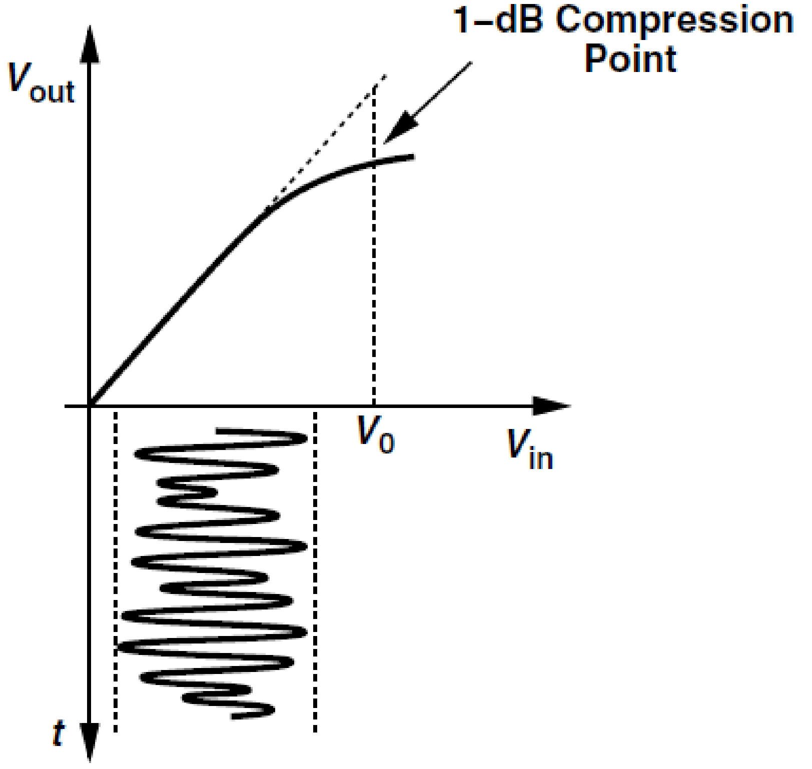
❑Two DACs trims the DC offset and a power detector monitors the output level, and its output is digitized.

❑During carrier leakage cancellation, the baseband produces a zero output so that the detector measures only the leakage.

[B. Razavi, RF Microelectronics]

TX Linearity

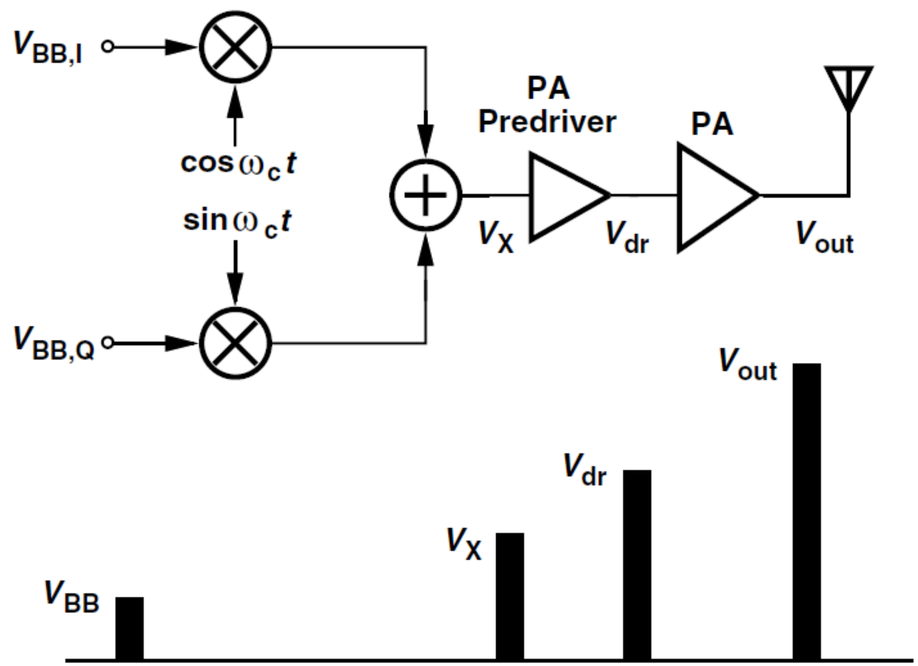
- Unlike downconversion mixers in a receiver, upconversion mixers in a transmitter sense no interferers.
- Excessive nonlinearity in the baseband port of upconversion mixers can corrupt the signal or raise the adjacent channel power.
- The distortion of a variable-envelope signal is typically characterized by the compression that it experiences.



[B. Razavi, RF Microelectronics]

TX Linearity

- In general case, the preceding stages must remain well below compression as the PA output approaches the 1-dB compression point. We must maximize the gain of the PA and minimize the output swing of the predriver and the stages preceding it. This requirement places additional burden on the PA design.



[B. Razavi, RF Microelectronics]

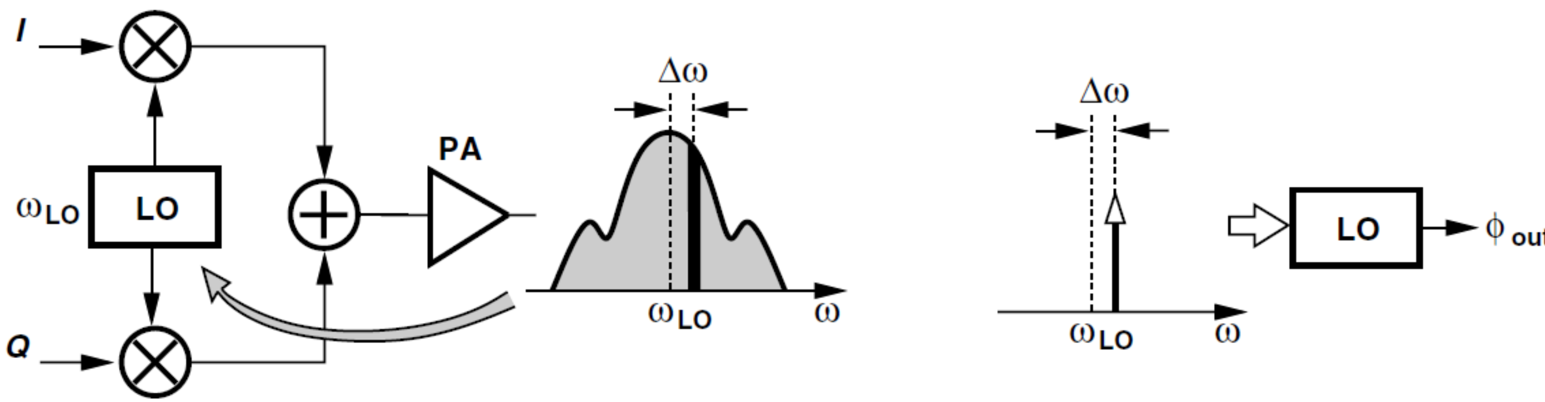
TX Noise

- Some standards (e.g., GSM) specify the maximum noise that a TX can transmit in the RX band.
- In a direct-conversion transmitter, the baseband circuits, the upconverter, and the PA may create significant noise in the RX band.
- To resolve this issue, noise shaping techniques can be adopted.

[B. Razavi, RF Microelectronics]

Oscillator Pulling

- The PA output exhibits very large swings, which couple to the oscillator through the silicon substrate, package parasitics, and traces on the PCB.

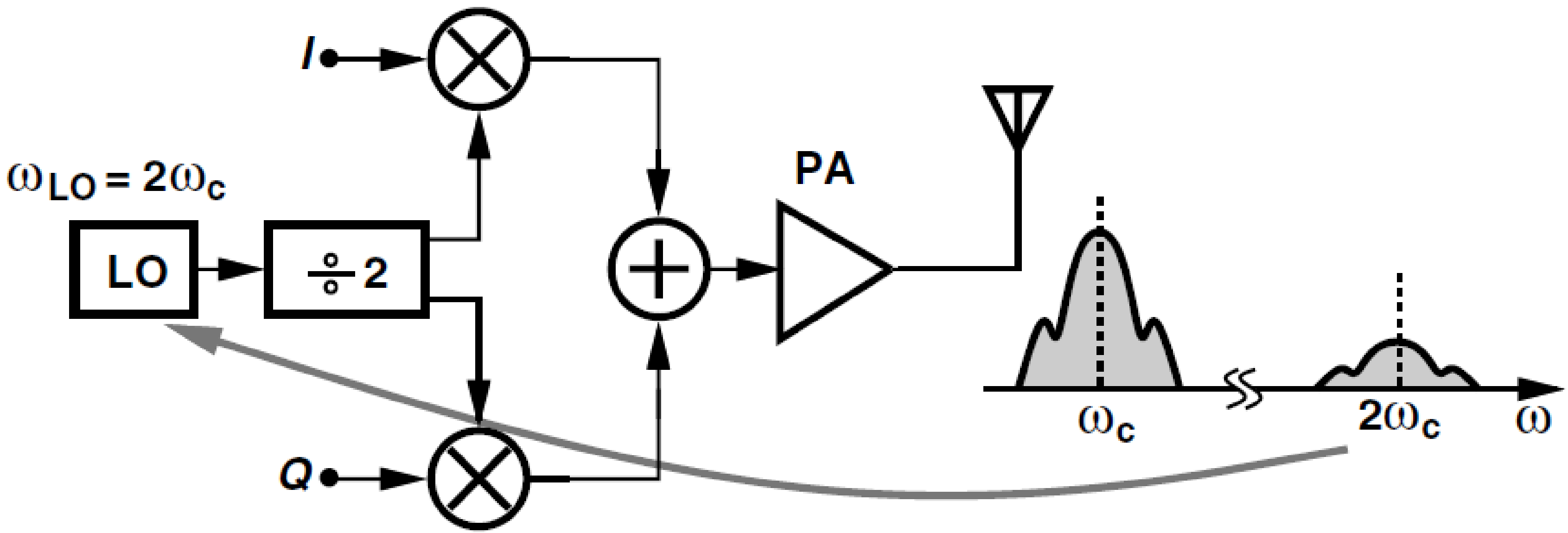


- In order to avoid injection pulling, the PA output frequency and the oscillator frequency must be made sufficiently different

[B. Razavi, RF Microelectronics]

Oscillator Pulling

❑The PA frequency is two times of the LO frequency:



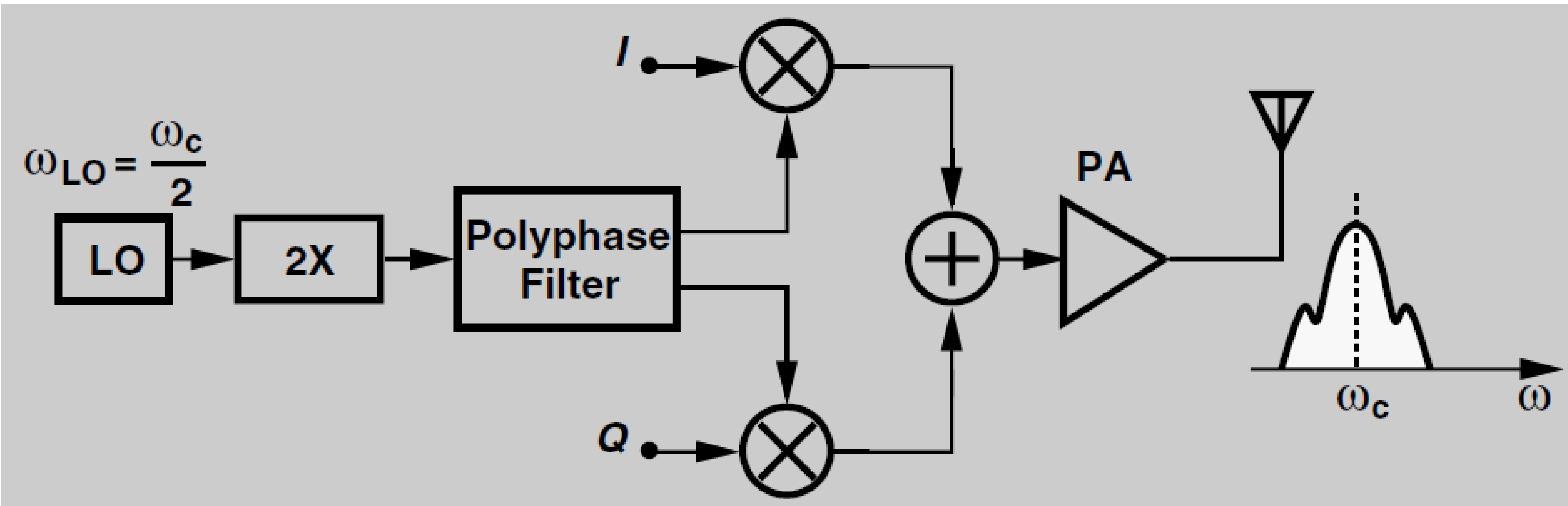
❑Injection pulling is greatly reduced.

❑Since the PA nonlinearity produces a finite amount of power at the second harmonic of the carrier, the LO may still be pulled.

[B. Razavi, RF Microelectronics]

Oscillator Pulling

❑The PA frequency is half of the LO frequency:



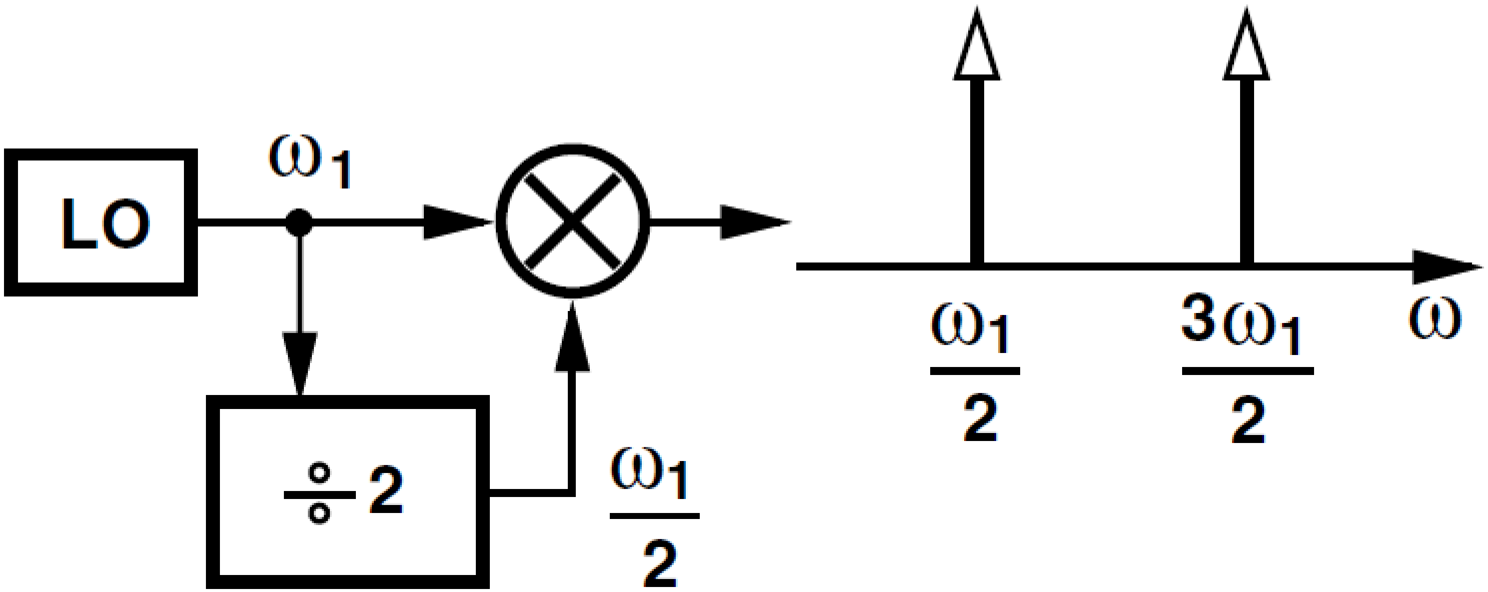
❑No harmonic of the PA output can pull the LO.

❑The doubler typically does not provide quadrature phases, necessitating additional quadrature generation stages, which is power hungry.

[B. Razavi, RF Microelectronics]

Sideband Issue

❑The PA frequency can be generated by mixing:



❑Half of the power delivered to the antenna is wasted.

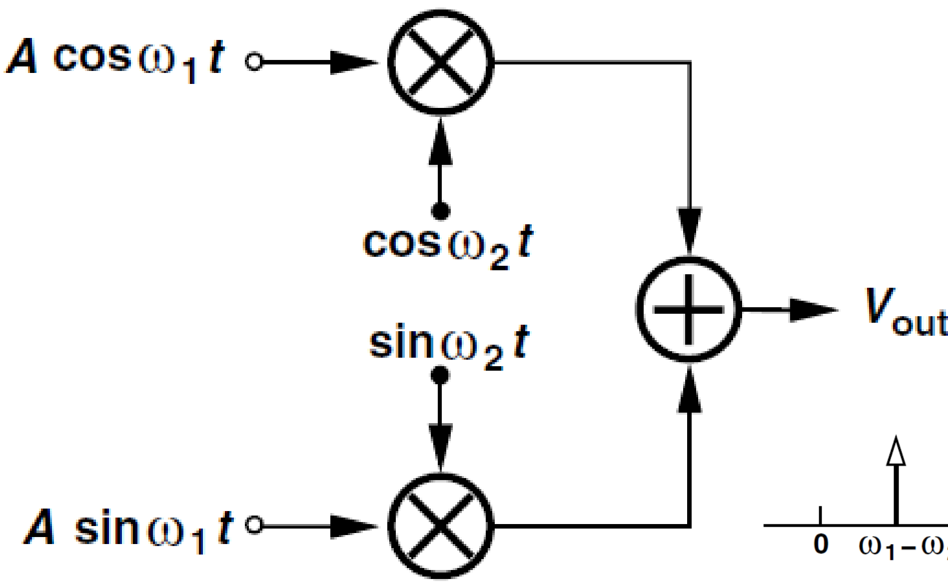
❑The power transmitted at the unwanted carrier frequency corrupts communication in other channels or bands.

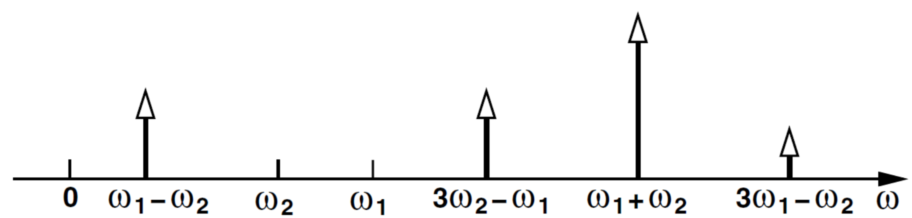
❑One component must therefore be suppressed.

[B. Razavi, RF Microelectronics]

Single-Sideband (SSB) Mixing

❑SSB mixing involves multiplying the quadrature phases of ω_1 and ω_2 and subtracting the results:





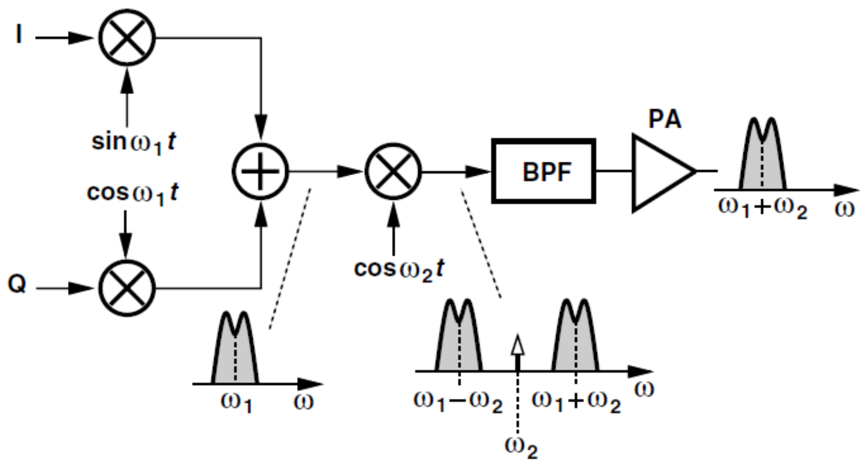
❑Gain and phase mismatches lead to an unwanted sideband

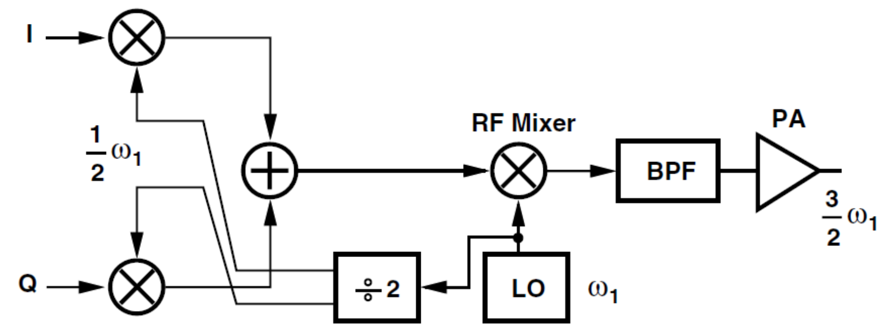
❑Harmonics of the input frequencies also corrupt the output of an SSB mixer

[B. Razavi, RF Microelectronics]

Heterodyne TX

❑An approach to avoiding injection pulling involves performing the signal upconversion in two steps so that the LO frequency remains far from the PA output spectrum:

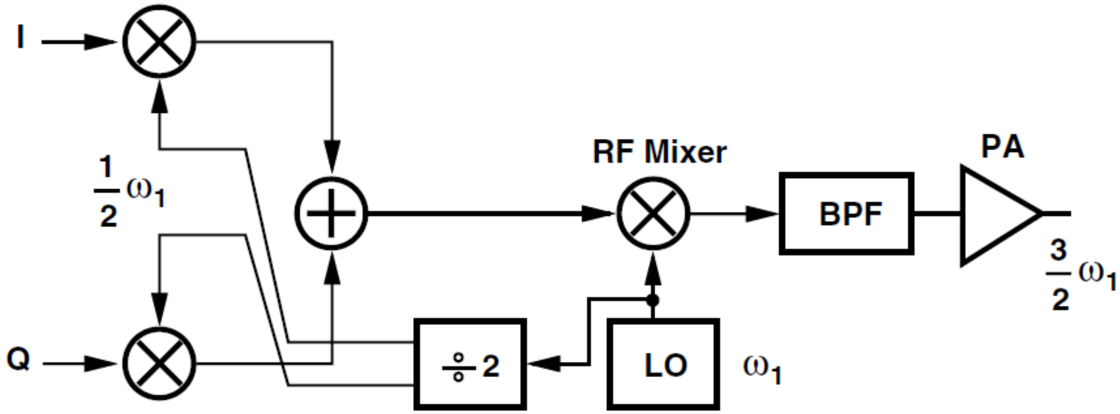




❑The I/Q upconversion occurs at a significantly lower frequency than the carrier, exhibiting smaller gain and phase mismatches.

[B. Razavi, RF Microelectronics]

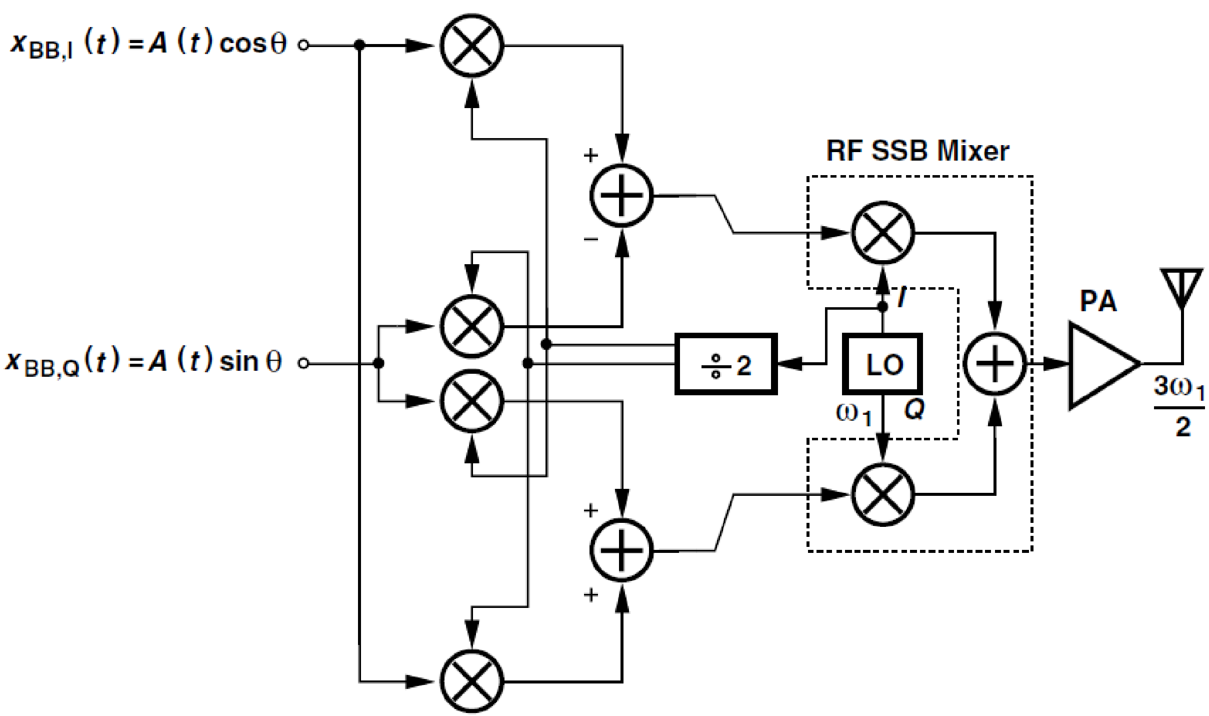
Carrier Leakage and Mixing Spurs



- The DC offsets in the baseband yield a component at $\omega_1/2$ at the output of the quadrature upconverter, which can be minimized by baseband control.
- The DC offset at the input of the RF mixer produces another component at ω_1, which must be removed by filtering.
- The mixing spurs arise the harmonics of the first LO and the second LO.

[B. Razavi, RF Microelectronics]

Cascaded SSB upconversion



❑The unwanted sideband produced by the RF mixer can be greatly suppressed through SSB mixing.

❑Mismatch can be reduced by the double-quadrature upconversion.

[B. Razavi, RF Microelectronics]