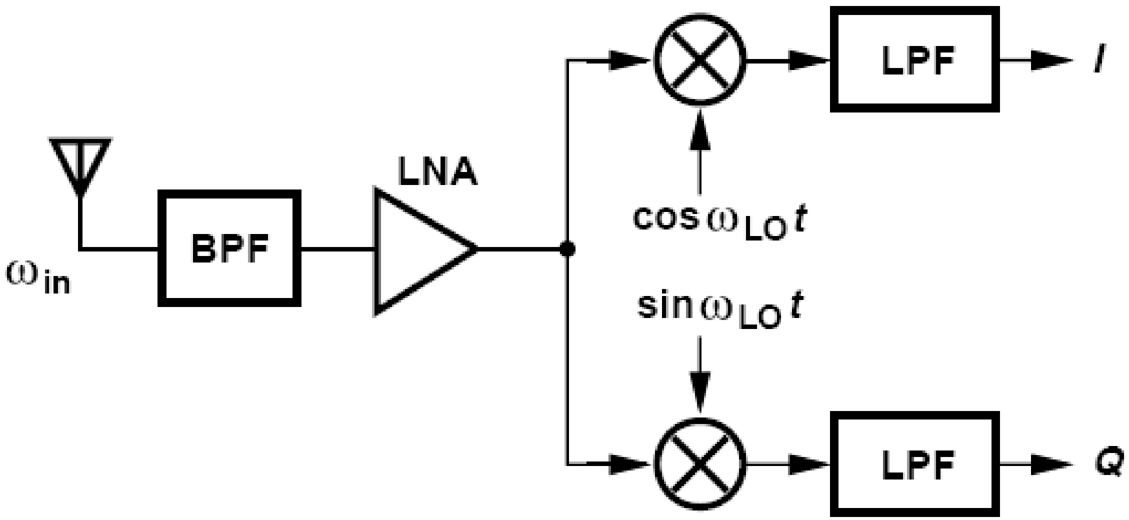
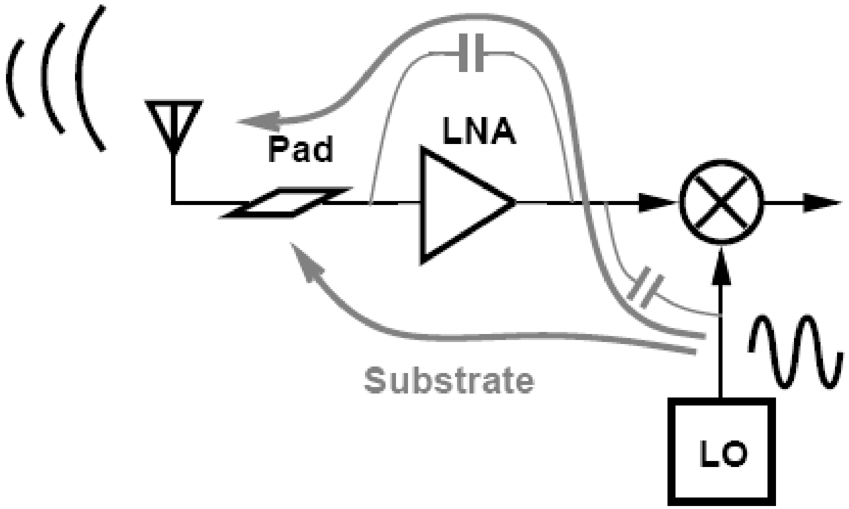
Direct-Conversion Receivers



- Absence of an image greatly simplifies the design process
- Channel selection is performed by on-chip low-pass filter
- Mixing spurs are considerably reduced in number

LO Leakage

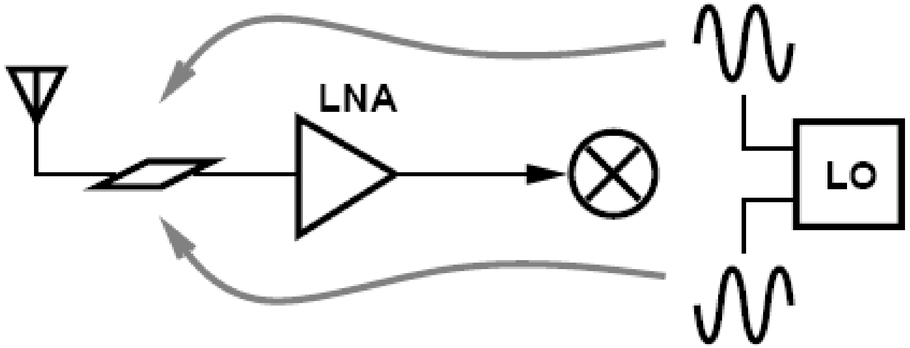


- **LO couples to the antenna through:**

  (a)device capacitances between LO and RF ports of mixer and device capacitances or resistances between the output and input of the LNA

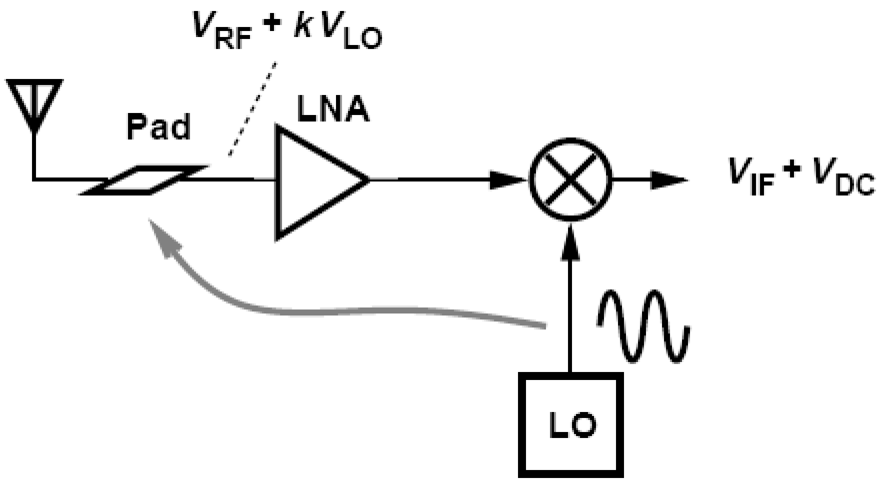
  (b)the substrate to the input pad, especially because the LO employs large on-chip spiral inductors

Cancellation of LO Leakage



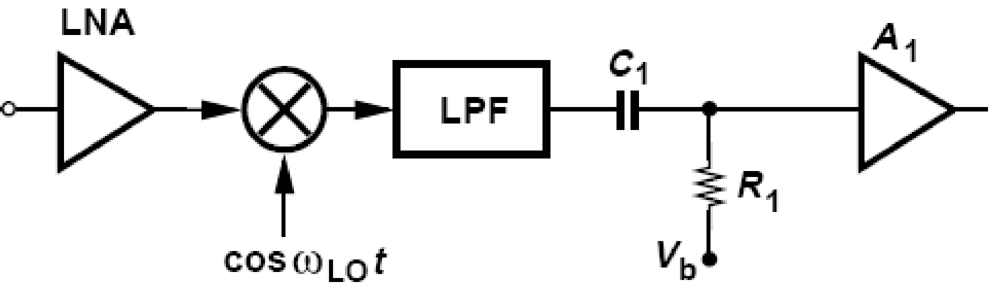
- LO leakage can be minimized through symmetric layout of the oscillator and the RF signal path
- LO leakage arises primarily from random or deterministic asymmetries in the circuits and the LO waveform

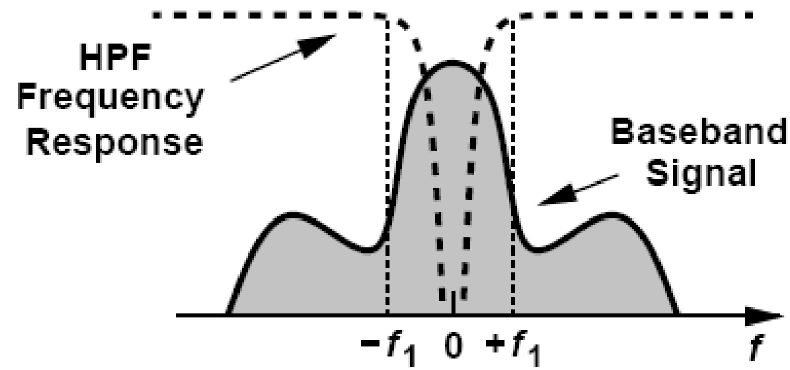
DC Offsets



- A finite amount of in-band LO leakage appears at the LNA input. Along with the desired signal, this component is amplified and mixed with LO.
- May saturates baseband circuits, simply prohibiting signal detection.

Cancellation of DC Offsets

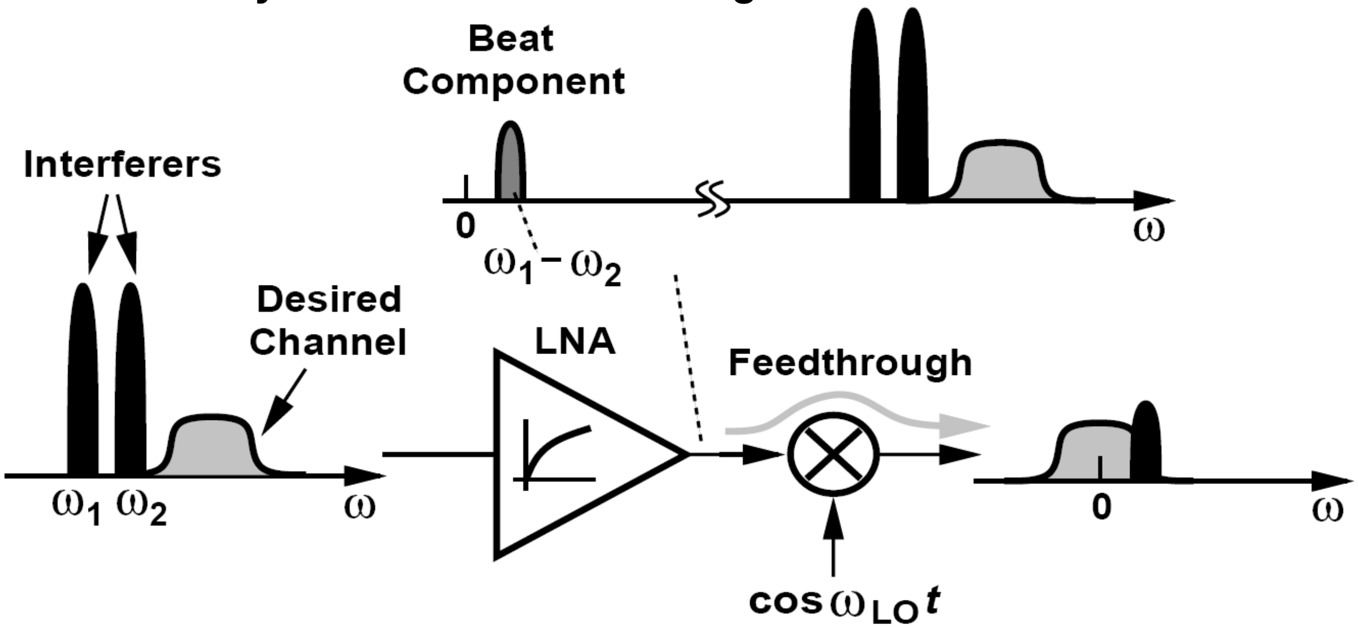




- **Offset cancellation: high-pass filter**
- **A drawback of ac coupling stems from its slow response to transient input.**

Even-Order Distortion: an Overview

Direct-conversion receivers are additionally sensitive to even-order nonlinearity in the RF path, and so are heterodyne architectures having a second zero IF.

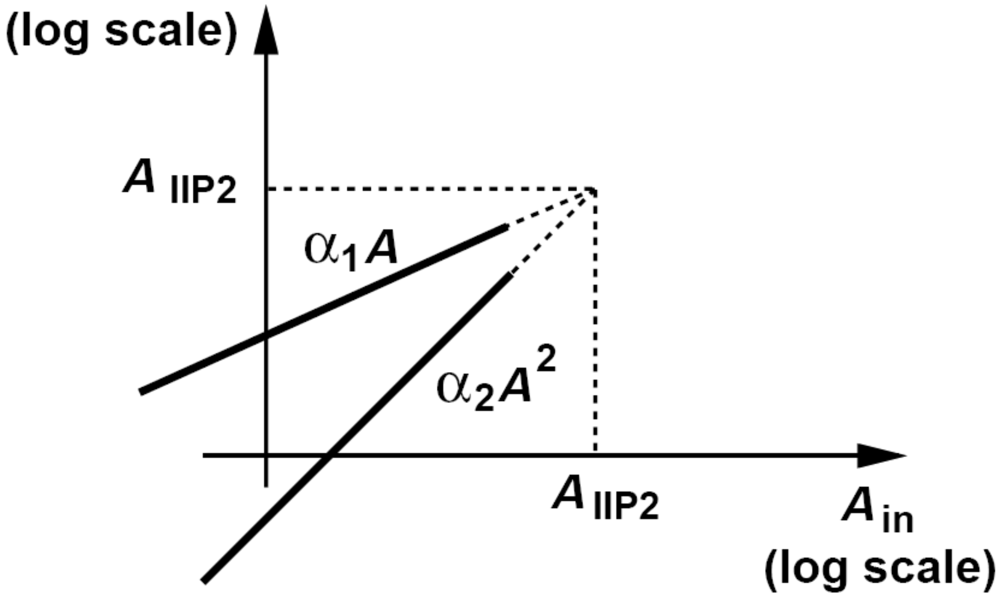


- **Asymmetries in the mixer or in the LO waveform allow a fraction of the RF input of the mixer to appear at the output without frequency translation, corrupting the downconverted signal.**
- **The beat generated by the LNA can be removed by ac coupling, making the input transistor of the mixer the dominant source of even-order distortion.**

Second Intercept Point (IP_2)

If $V_{in}(t)$ = $A\cos\omega_1 t + A\cos\omega_2 t$, then the LNA output is given by

$$
\begin{aligned}
V_{out}(t) &= \alpha_1 V_{in}(t) + \alpha_2 V_{in}^2(t) \\
&= \alpha_1 A(\cos\omega_1 t + \cos\omega_2 t) + \alpha_2 A^2 \cos(\omega_1 + \omega_2)t \\
&+ \alpha_2 A^2 \cos(\omega_1 - \omega_2)t + \cdots,
\end{aligned}
$$



Beat amplitude grows with the *square* of the amplitude of the input tones.

Example of Calculation of IP_2

Since the feedthrough of the beat depends on the mixer and LO asymmetries, the beat amplitude measured in the baseband depends on the device dimensions and the layout and is therefore difficult to formulate.

Suppose the attenuation factor experienced by the beat as it travels through the mixer is equal to *k* whereas the gain seen by each tone as it is downconverted to the baseband is equal to unity. Calculate the IP_2.

Solution:

From equation above, the value of A that makes the output beat amplitude, $k\alpha_2 A^2$, equal to the main tone amplitude, $\alpha_1 A$, is given by

$$k\alpha_2 A_{IIP2}^2 = \alpha_1 A_{IIP2}$$

hence

$$A_{IIP2} = \frac{1}{k} \cdot \frac{\alpha_1}{\alpha_2}$$

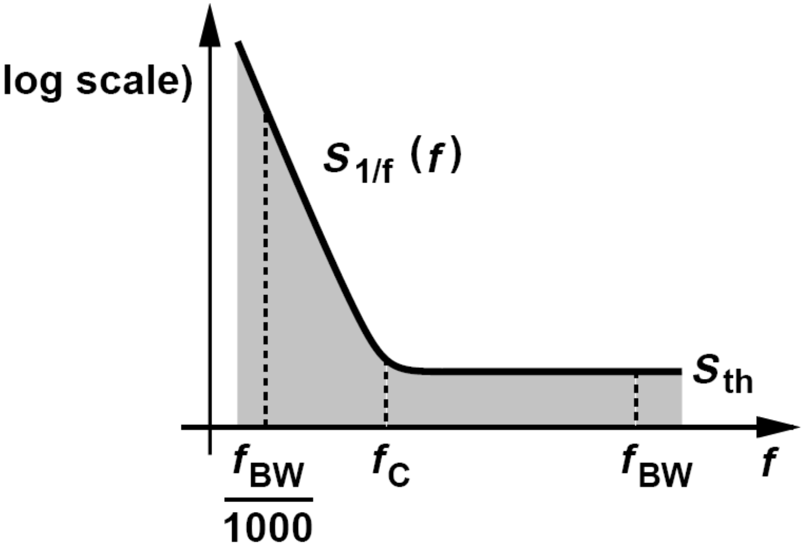
Flicker Noise

Since the signal is centered around zero frequency, it can be substantially corrupted by flicker noise.

We note that if $S_{1/f} = \alpha/f$, then at f_c,

$$\frac{\alpha}{f_c} = S_{th}$$

$$\begin{aligned} P_{n1} &= \int_{f_{BW}/1000}^{f_c} \frac{\alpha}{f} df + (f_{BW} - f_c) S_{th} \\ &= \alpha \ln \frac{1000 f_c}{f_{BW}} + (f_{BW} - f_c) S_{th} \\ &= \left(6.9 + \ln \frac{f_c}{f_{BW}}\right) f_c S_{th} + (f_{BW} - f_c) S_{th} \\ &= \left(5.9 + \ln \frac{f_c}{f_{BW}}\right) f_c S_{th} + f_{BW} S_{th}. \end{aligned}$$



$$P_{n2} \approx f_{BW} S_{th}$$

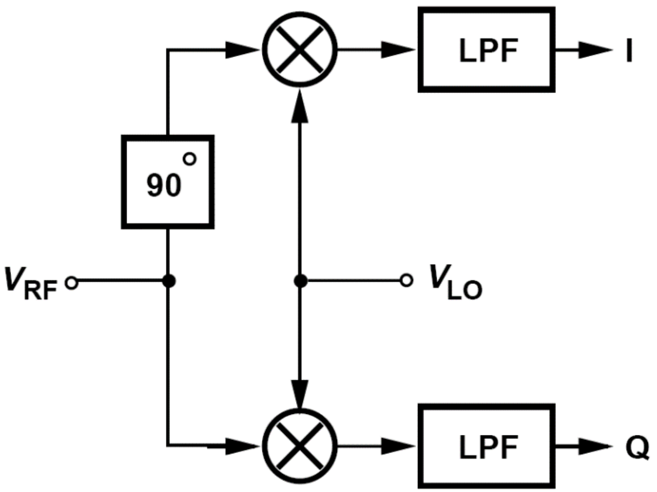
$$\frac{P_{n1}}{P_{n2}} = 1 + \left(5.9 + \ln \frac{f_c}{f_{BW}}\right) \frac{f_c}{f_{BW}}$$

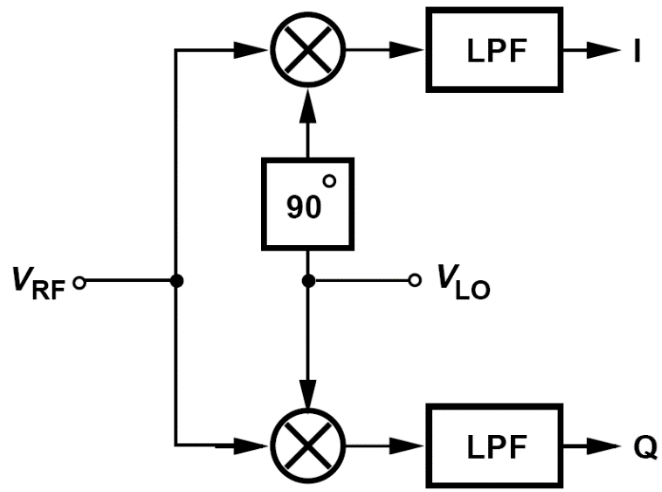
An 802.11g receiver exhibits a baseband flicker noise corner frequency of 200 kHz. Determine the flicker noise penalty

We have f_{BW} = 10 MHz, f_c = 200 kHz, and hence

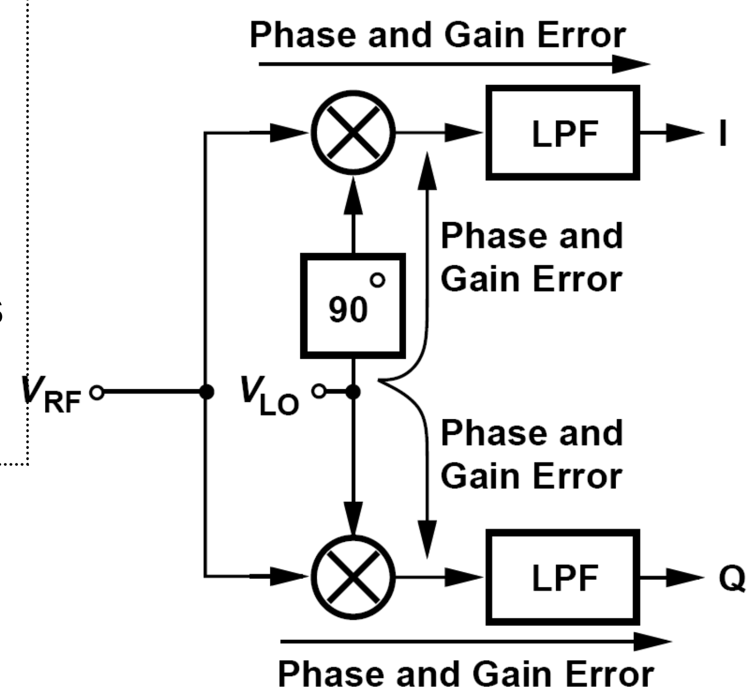
$$\frac{P_{n1}}{P_{n2}} = 1.04$$

I/Q Mismatch: Sources

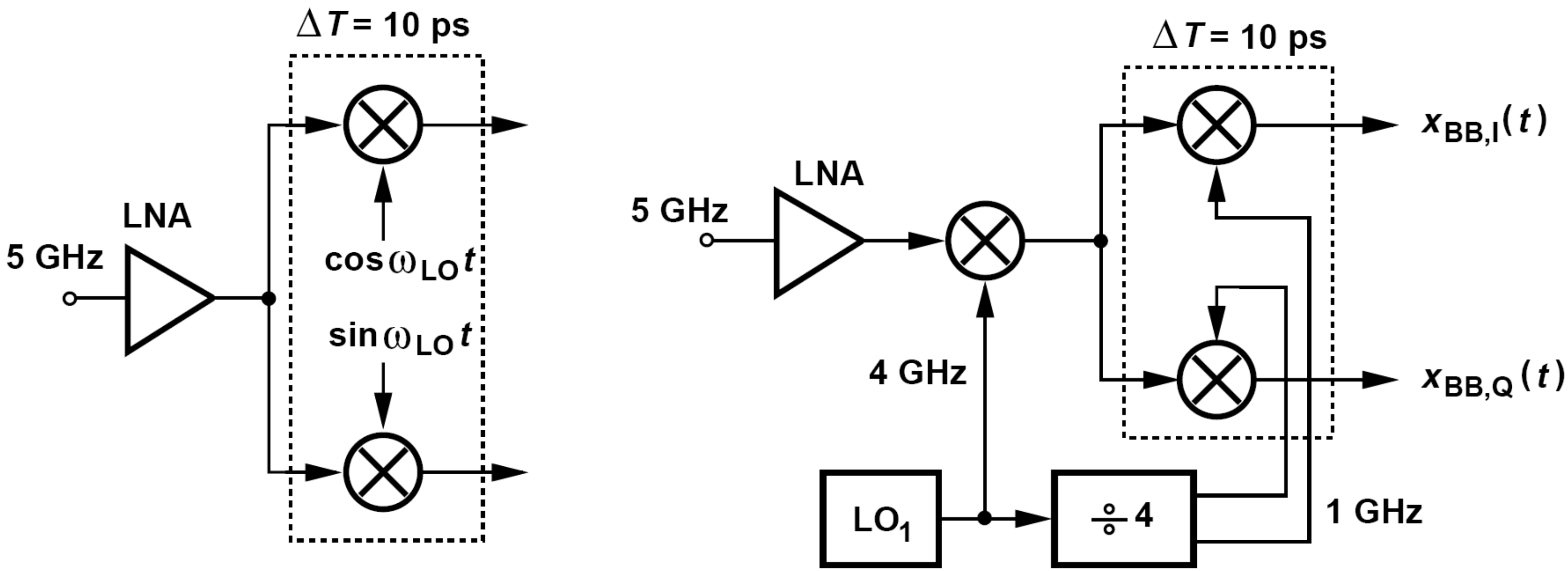




- Separation into quadrature phases can be accomplished by shifting either the RF signal or the LO waveform by 90° .
- Errors in the 90° phase shift circuit and mismatches between the quadrature mixers result in imbalances in the amplitudes and phases of the baseband I and Q outputs.



I/Q Mismatch in Direct-Conversion Receivers And Heterodyne Topologies



- Quadrature mismatches tend to be larger in direct-conversion receivers than in heterodyne topologies.
- This occurs because

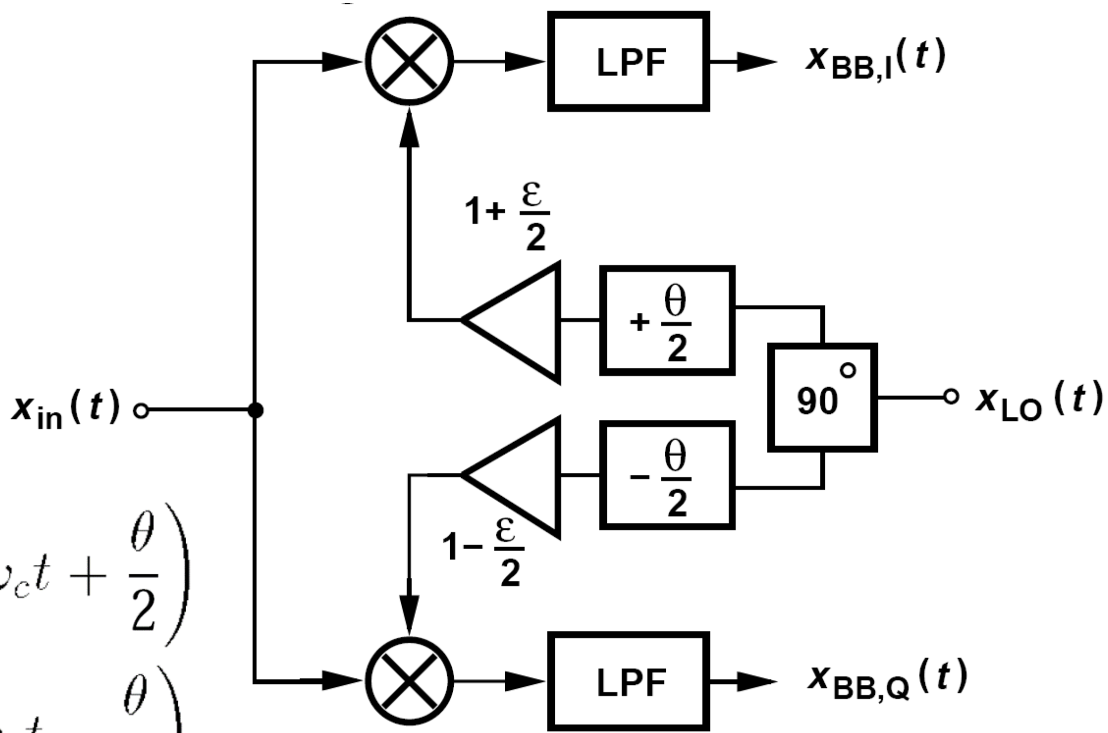
  (1) the propagation of a higher frequency (f_{in}) through quadrature mixers experiences greater mismatches;

  (2) the quadrature phases of the LO itself suffer from greater mismatches at higher frequencies;

Effect of I/Q Mismatch (I)

Let us lump all of the gain and phase mismatches shown below:

$$a\cos\omega_c t + b\sin\omega_c t,$$



$$x_{LO,I}(t) = 2\left(1+\frac{\epsilon}{2}\right)\cos\left(\omega_c t + \frac{\theta}{2}\right)$$

$$x_{LO,Q}(t) = 2\left(1-\frac{\epsilon}{2}\right)\sin\left(\omega_c t - \frac{\theta}{2}\right),$$

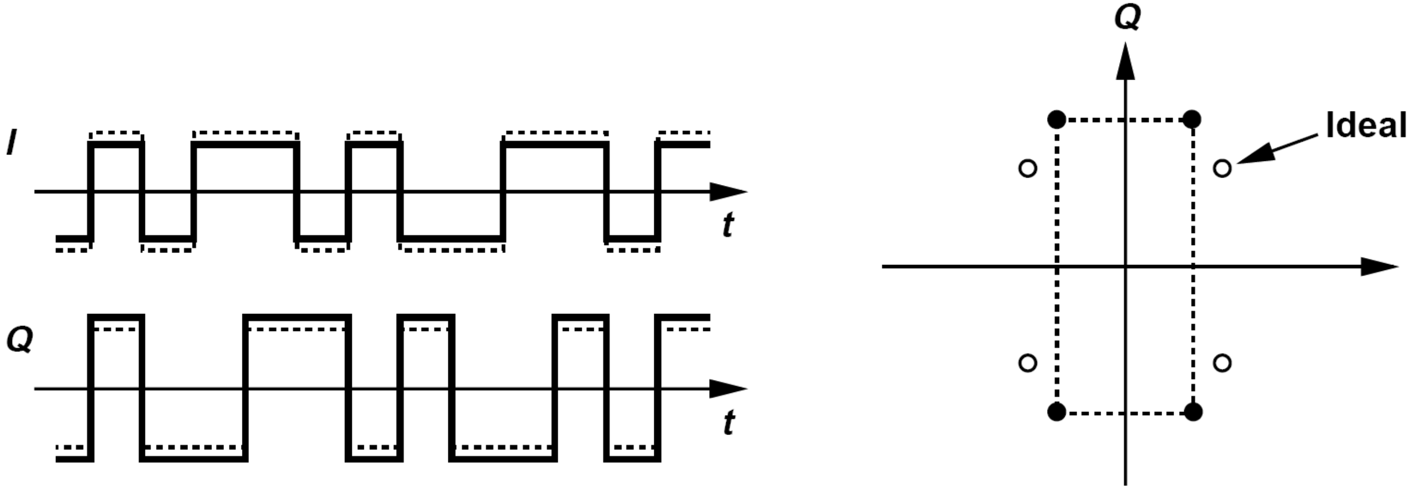
$$x_{BB,I}(t) = a\left(1+\frac{\epsilon}{2}\right)\cos\frac{\theta}{2} - b\left(1+\frac{\epsilon}{2}\right)\sin\frac{\theta}{2}$$

$$x_{BB,Q}(t) = -a\left(1-\frac{\epsilon}{2}\right)\sin\frac{\theta}{2} + b\left(1-\frac{\epsilon}{2}\right)\cos\frac{\theta}{2}$$

Effect of I/Q Mismatch (Ⅱ)

We now examine the results for two special cases:

(1) $\varepsilon \neq 0, \theta = 0$: the quadrature baseband symbols are scaled differently in amplitude,



(2) $\varepsilon = 0, \theta \neq 0$: each baseband output is corrupted by a fraction of the data symbols in the other output

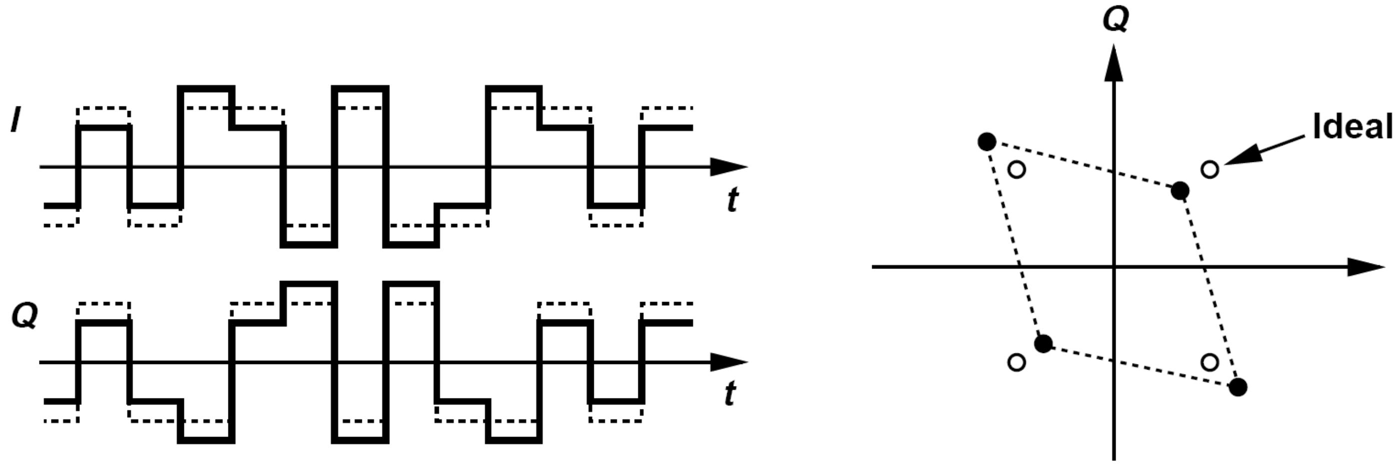
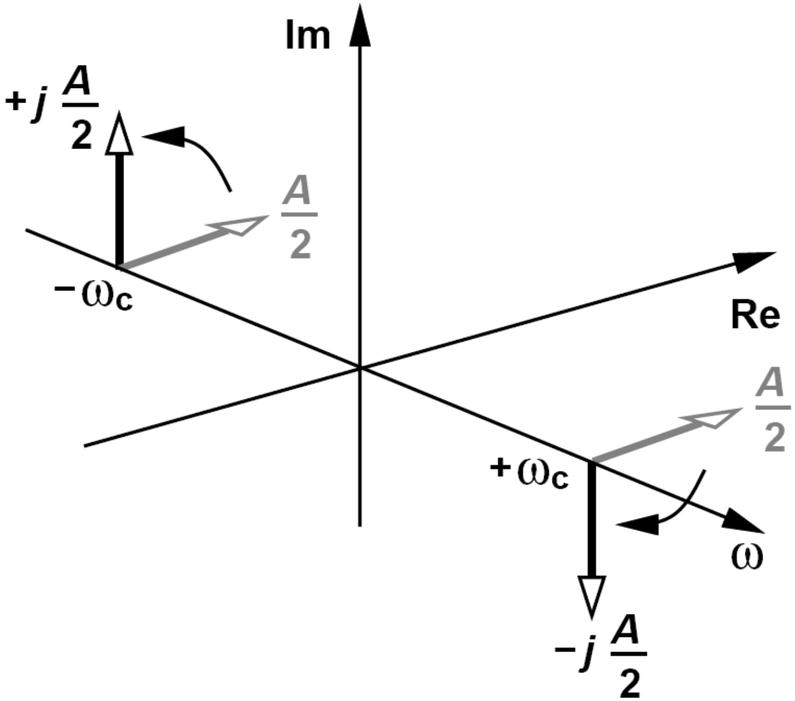


Image-Reject Receivers: 90° Phase Shift—Cosine Signal

Before studying these architectures, we must define a "shift-by-90° " operation.

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

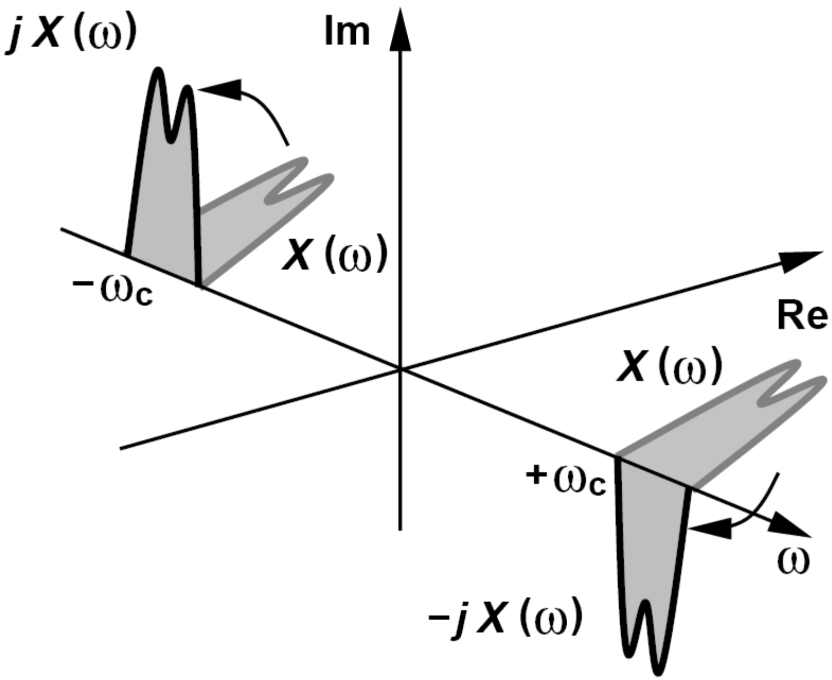


- **The impulse at $+\omega_c$ is rotated clockwise and that at $-\omega_c$ counterclockwise**

Image-Reject Receivers: 90° Phase Shift—Modulated Signal

Similarly, for a narrowband modulated signal:

$$
\begin{aligned}
A(t)\cos[\omega_c t + \phi(t) - 90^\circ] &= A(t)\frac{e^{+j[\omega_c t + \phi(t) - 90^\circ]} + e^{-j[\omega_c t + \phi(t) - 90^\circ]}}{2} \\
&= A(t)\frac{-je^{+j[\omega_c t + \phi(t)]} + je^{-j[\omega_c t + \phi(t)]}}{2} \\
&= A(t)\sin[\omega_c t + \phi(t)].
\end{aligned}
$$



- **We write in the frequency domain:**

$$X_{90^\circ}(\omega) = X(\omega)[-j\,\mathrm{sgn}(\omega)]$$

- **The shift-by-90° operation is also called the "Hilbert transform".**

Examples of Hilbert Transform

In phasor diagrams, we simply multiply a phasor by -*j* to rotate it by 90° clockwise. Is that inconsistent with the Hilbert transform?

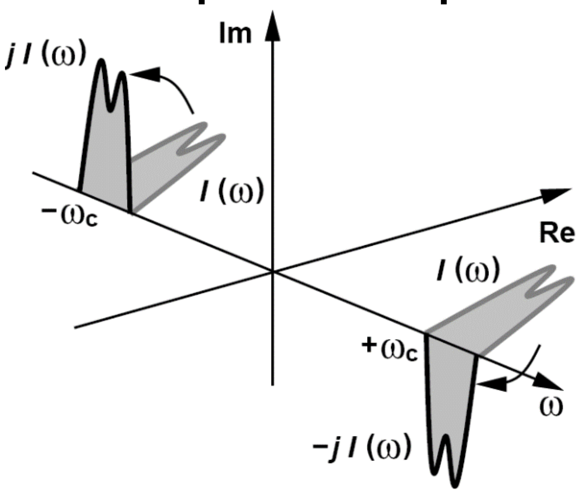
No, it is not. A phasor is a representation of $A\exp(j\omega_c t)$, i.e., only the positive frequency content. That is, we implicitly assume that if $A\exp(j\omega_c t)$ is multiplied by -*j*, then $A\exp(-j\omega_c t)$ is also multiplied by +*j*.

Plot the spectrum of $A\cos\omega_c t + jA\sin\omega_c t$.

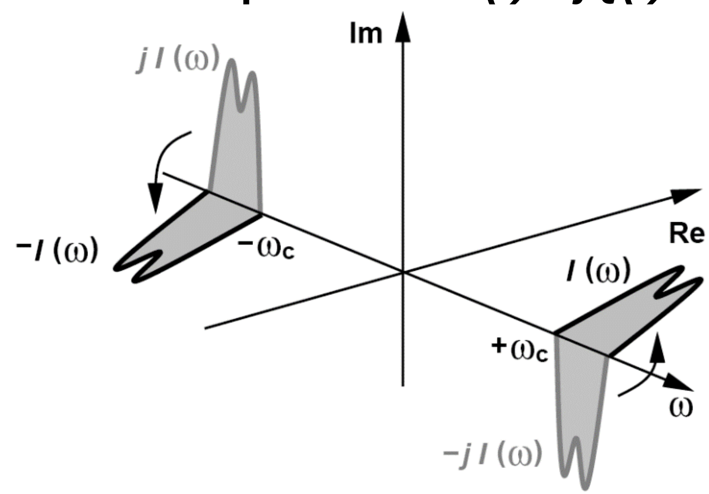
Example of *I* + *jQ*

A narrowband signal *I(t)* with a real spectrum is shifted by 90° to produce *Q(t)*. Plot the spectrum of *I(t)* + *jQ(t).*

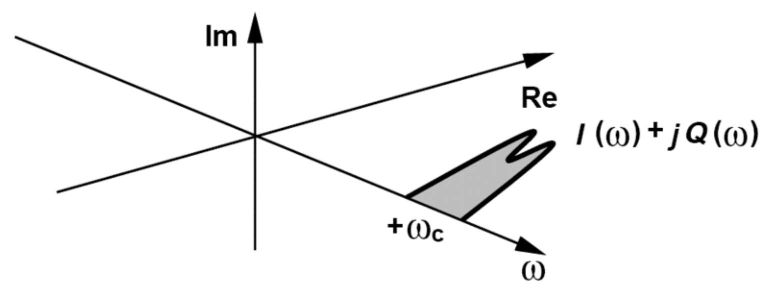
We first multiply *I(ω)* by -*j*sgn*(ω)* and then, in a manner similar to the previous example, multiply the result by *j*. The spectrum of *jQ(t)* therefore cancels that of *I(t)* at negative frequencies and enhances it at positive frequencies. The one-sided spectrum of *I(t)* + *jQ(t)* proves useful in the analysis of transceivers.



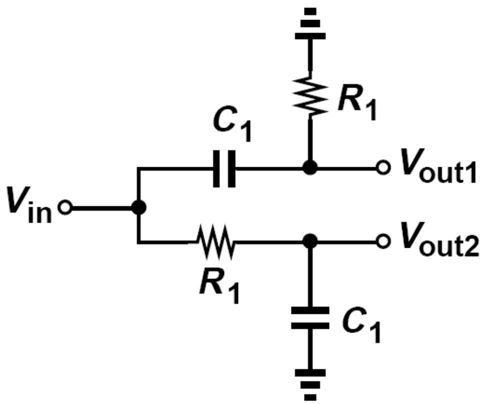
(a)

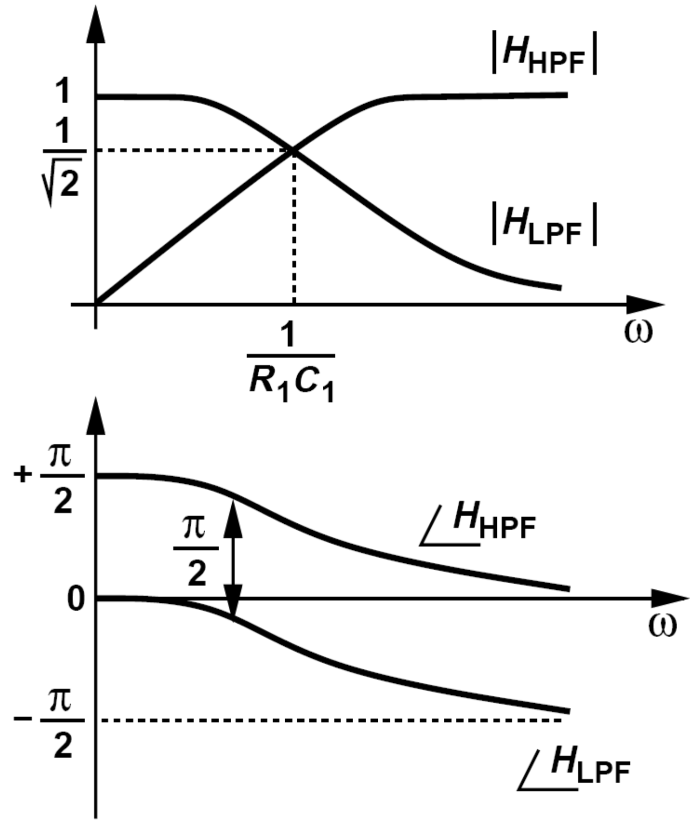


(b)



Implementation of the 90° Phase Shift





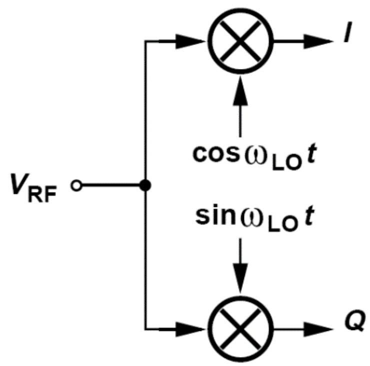
The high-pass and low-pass transfer functions are respectively given by:

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

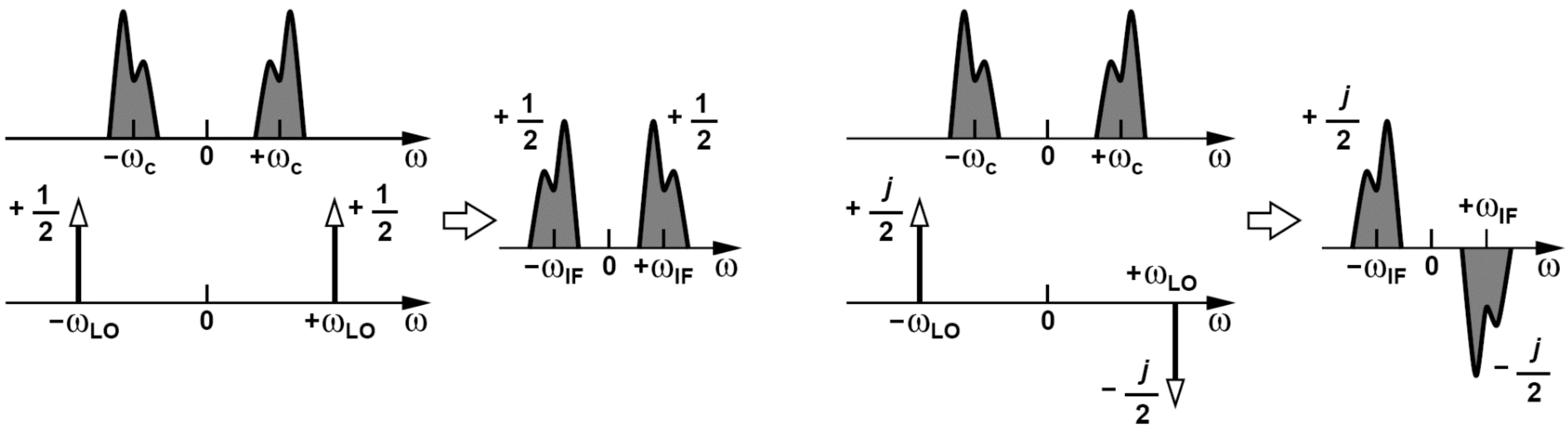
$$H_{LPF}(s) = \frac{V_{out2}}{V_{in}} = \frac{1}{R_1C_1s+1}$$

We can therefore consider V_{out2} as the Hilbert transform of V_{out1} at frequencies close to $(R_1C_1)^{-1}$

Another Approach to Implement the 90° Phase Shift



(a)

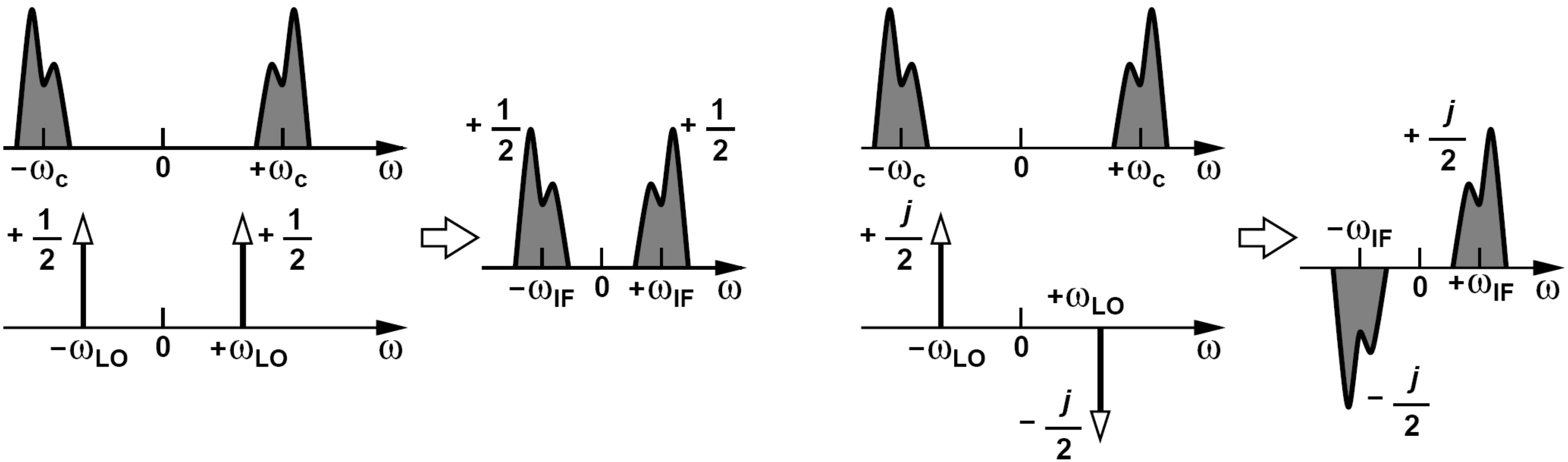


- The RF input is mixed with the quadrature phases of the LO so as to translate the spectrum to a non-zero IF.
- The IF spectrum emerging from the lower arm is the Hilbert transform of that from the upper arm.

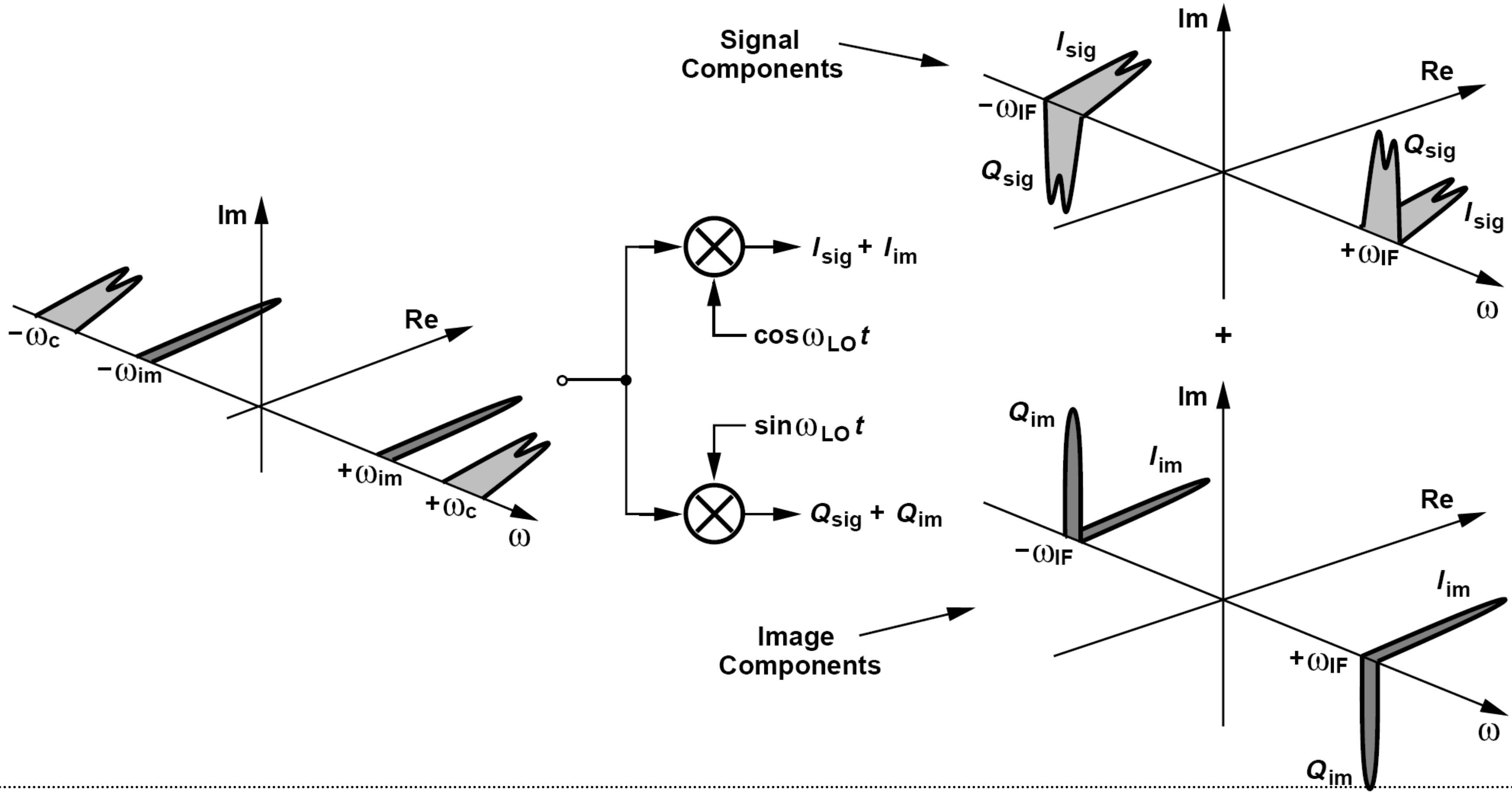
Low-Side Injection of the Above Implementation

The realization above assumes high-side injection for the LO. Repeat the analysis for low-side injection.

Figures below show the spectra for mixing with cos $\omega_{LO}t$ and sin $\omega_{LO}t$, respectively. In this case, the IF component in the lower arm is the negative of the Hilbert transform of that in the upper arm.



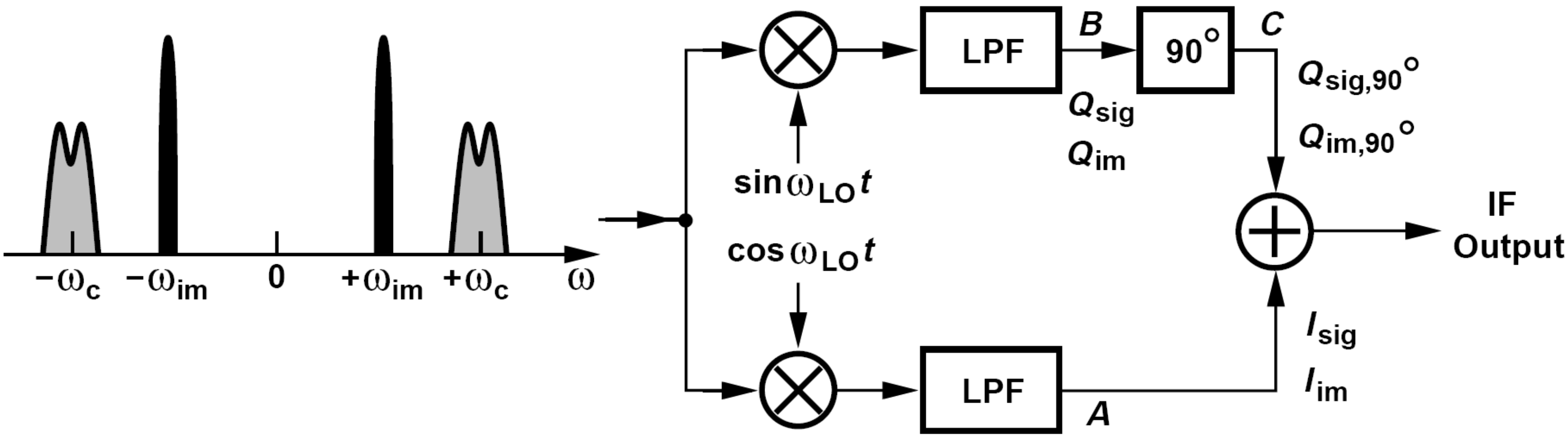
How Can the Image Components Cancel Each Other?



- Is *I(t)* + *Q(t)* free from the image? Since the image components in *Q(t)* are 90° out of phase with respect to those in *I(t)*, this summation still contains the image.

Hartley Architecture

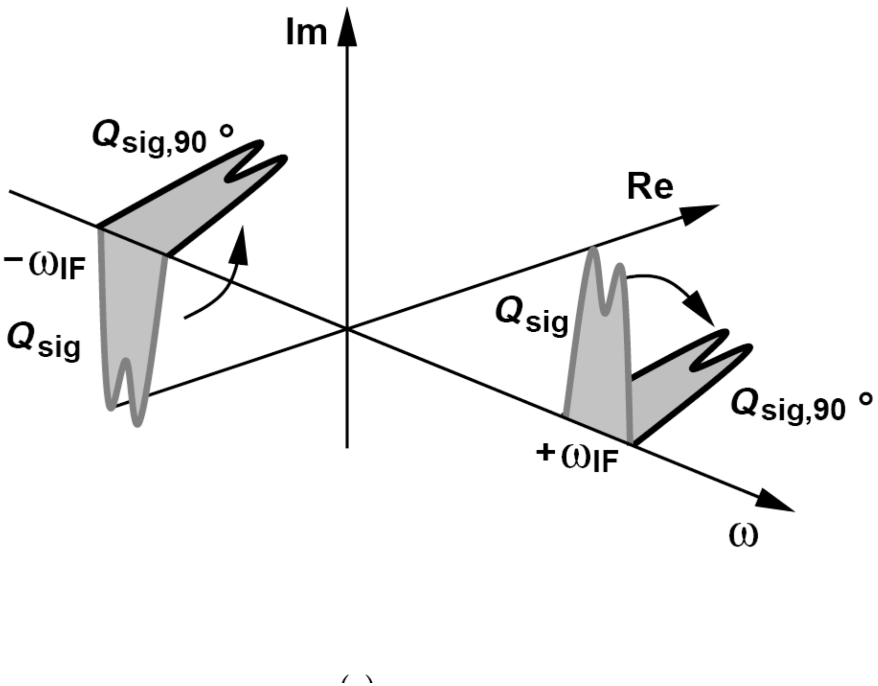
If we shift *I(t)* or *Q(t)* by another 90 ° before adding them, the image may be removed.

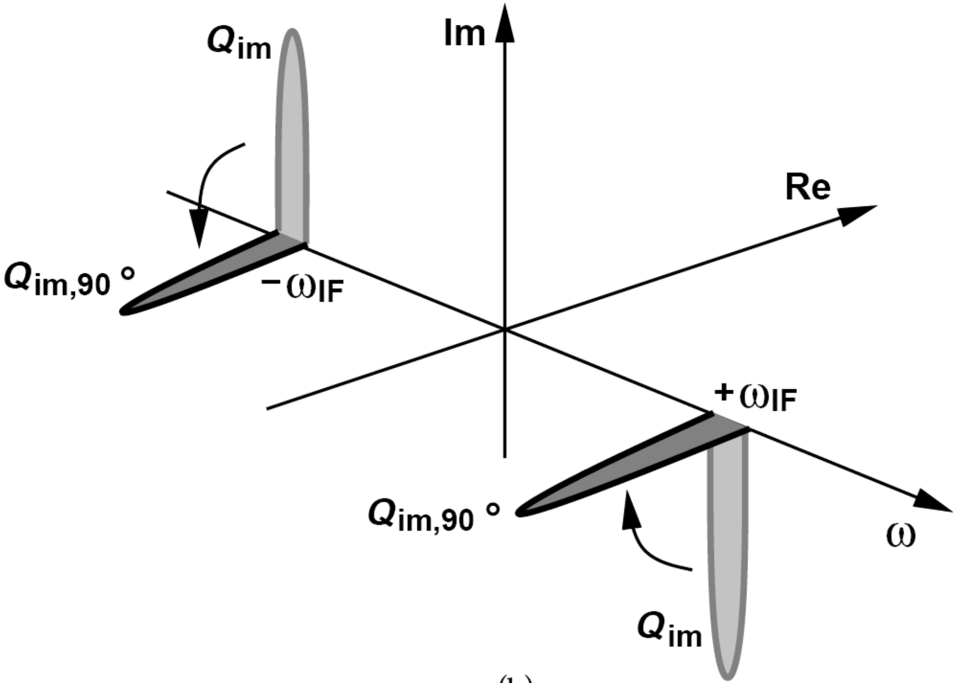


- **The low-pass filters are inserted to remove the unwanted high-frequency components generated by the mixers**

Operation of Hartley's Architecture

We assume low-side injection and apply a 90° phase shift to the Hilbert transforms of the signal and the image (the *Q* arm)





- **Multiplication of Q_{sig} by *-j*sgn(*ω*) rotates and superimposes the spectrum of Q_{sig} on that of I_{sig}, doubling the signal amplitude. On the other hand, multiplication of Q_{im} by –*j*sgn(*ω*) creates the opposite of I_{im}, canceling the image.**

Analytical Expression of Hartley's Architecture

An eager student constructs the Hartley architecture but with high-side injection. Explain what happens.

We note that the quadrature converter takes the Hilbert transform of the signal and the negative Hilbert transform of the image. Thus, with another 90° phase shift, the outputs *C* and *A* in figure above contain the signal with opposite polarities and the image with the same polarity. The circuit therefore operates as a "signal-reject" receiver! Of course, the design is salvaged if the addition is replaced with subtraction.

Represent the received signal and image as x(t) = $A_{sig}\cos(\omega_c t + \Phi_{sig}) + A_{im}\cos(\omega_{im} t + \Phi_{im})$, obtaining the signal at point A and B:

$$\begin{aligned} x_A(t) &= \frac{A_{sig}}{2}\cos[(\omega_c - \omega_{LO})t + \phi_{sig}] + \frac{A_{im}}{2}\cos[(\omega_{im} - \omega_{LO})t + \phi_{im}] \\ x_B(t) &= -\frac{A_{sig}}{2}\sin[(\omega_c - \omega_{LO})t + \phi_{sig}] - \frac{A_{im}}{2}\sin[(\omega_{im} - \omega_{LO})t + \phi_{im}], \end{aligned}$$

It follows that:

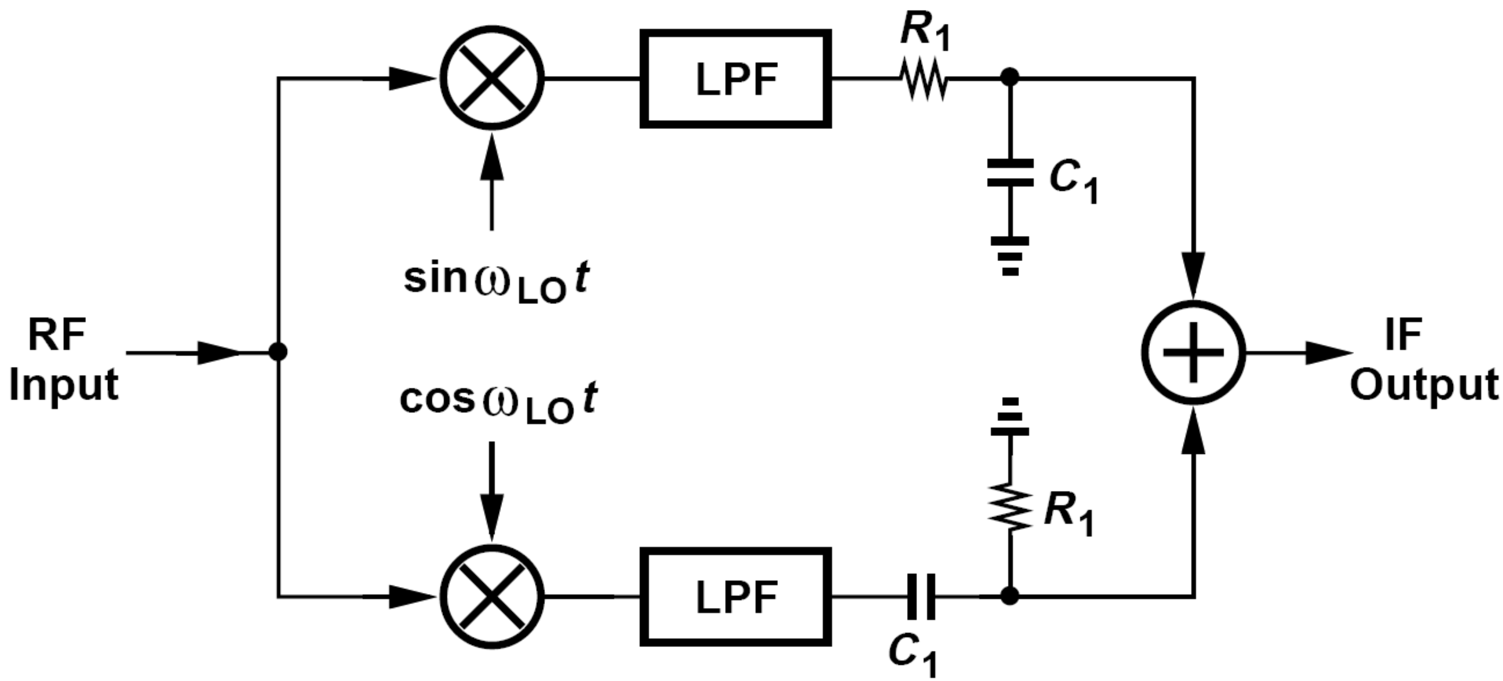
$$x_C(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_{LO})t + \phi_{sig}] - \frac{A_{im}}{2}\cos[(\omega_{im} - \omega_{LO})t + \phi_{im}].$$

Upon addition of $x_A(t)$ and $x_C(t)$, we retain the signal and reject the image.

The

Realization of 90° Phase Shift of Hartley Architecture

The 90° phase shift depicted before 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° while circuit components vary with process and temperature.

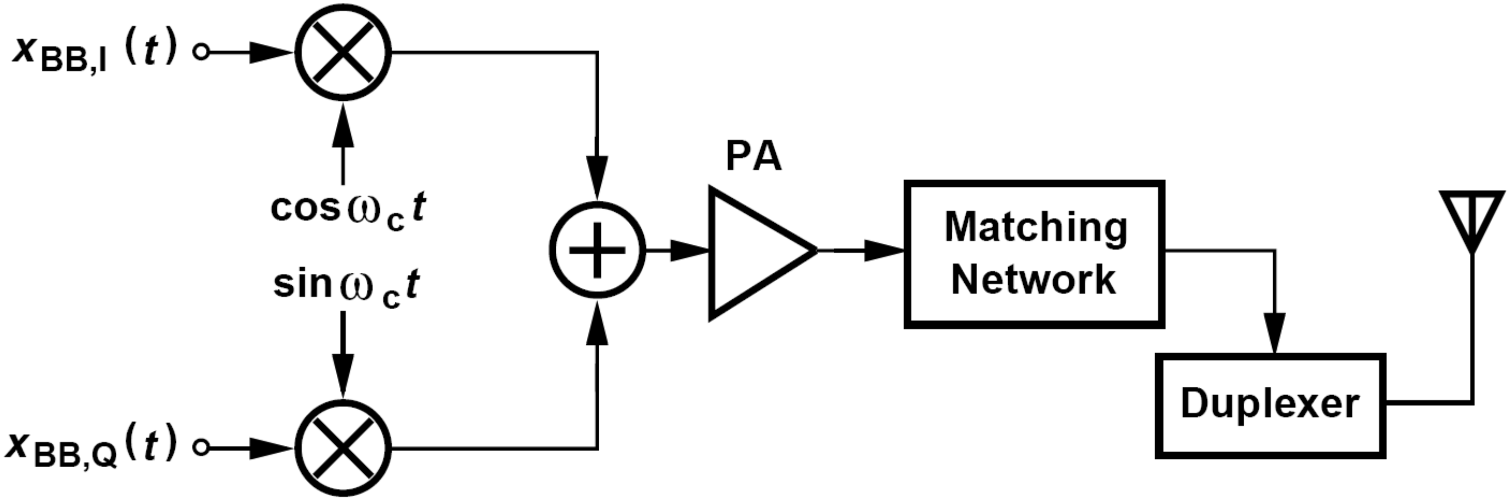
Direct-Conversion Transmitters: Overview

The above expression of a GMSK waveform can be generalized to any narrowband modulated signal:

$$\begin{aligned} x(t) &= A(t)\cos[\omega_c t + \phi(t)] \\ &= A(t)\cos\omega_c t\cos[\phi(t)] - A(t)\sin\omega_c t\sin[\phi(t)] \end{aligned}$$

We therefore define the quadrature baseband signals as

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



This topology directly translates the baseband spectrum to the RF carrier by means of a "quadrature upconverter".

Direct-Conversion Transmitters: I/Q Mismatch

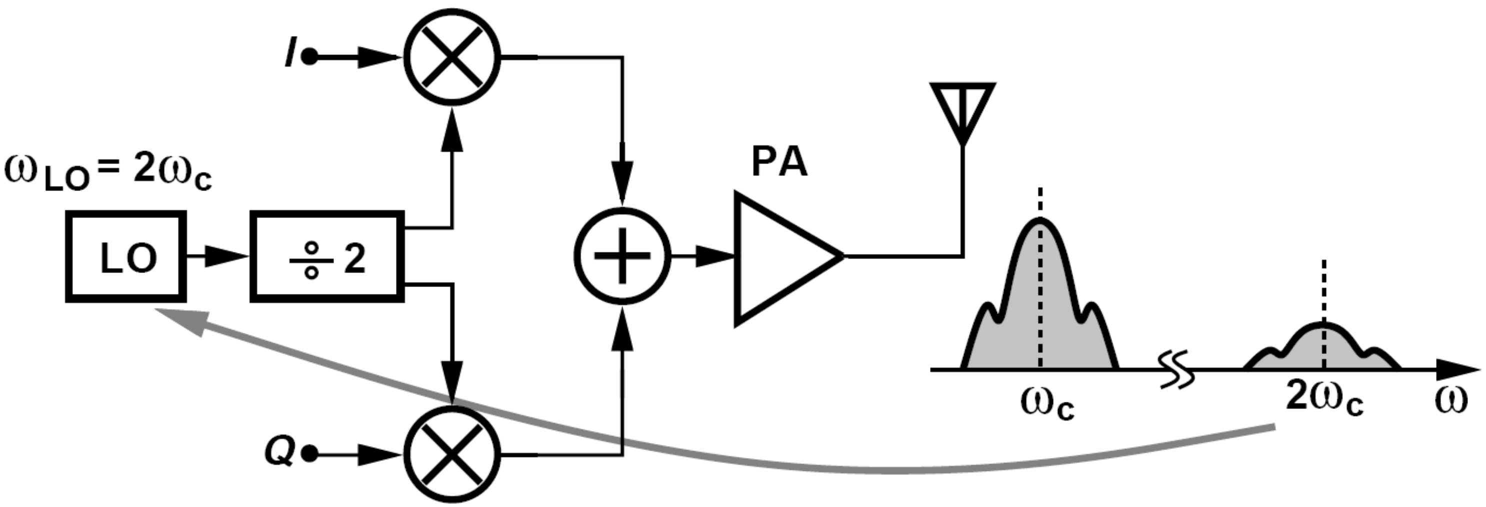
The I/Q mismatch in direct-conversion receivers results in “cross-talk” between the quadrature baseband outputs or, equivalently, distortion in the constellation.

$$
\begin{aligned}
x(t) &= \alpha_1(A_c+\Delta A_c)\cos(\omega_c t+\Delta\theta)+\alpha_2 A_c\sin\omega_c t \\
&= \alpha_1(A_c+\Delta A_c)\cos\Delta\theta\cos\omega_c t+[\alpha_2 A_c-\alpha_1(A_c+\Delta A_c)\sin\Delta\theta]\sin\omega_c t.
\end{aligned}
$$

For the four points in the constellation:

$$
\begin{aligned}
\beta_1 &= +\left(1+\frac{\Delta A_c}{A_c}\right)\cos\Delta\theta,\ \beta_2=1-\left(1+\frac{\Delta A_c}{A_c}\right)\sin\Delta\theta \\
\beta_1 &= +\left(1+\frac{\Delta A_c}{A_c}\right)\cos\Delta\theta,\ \beta_2=-1-\left(1+\frac{\Delta A_c}{A_c}\right)\sin\Delta\theta \\
\beta_1 &= -\left(1+\frac{\Delta A_c}{A_c}\right)\cos\Delta\theta,\ \beta_2=1+\left(1+\frac{\Delta A_c}{A_c}\right)\sin\Delta\theta \\
\beta_1 &= -\left(1+\frac{\Delta A_c}{A_c}\right)\cos\Delta\theta,\ \beta_2=-1+\left(1+\frac{\Delta A_c}{A_c}\right)\sin\Delta\theta.
\end{aligned}
$$

Modern Direct-Conversion Transmitters



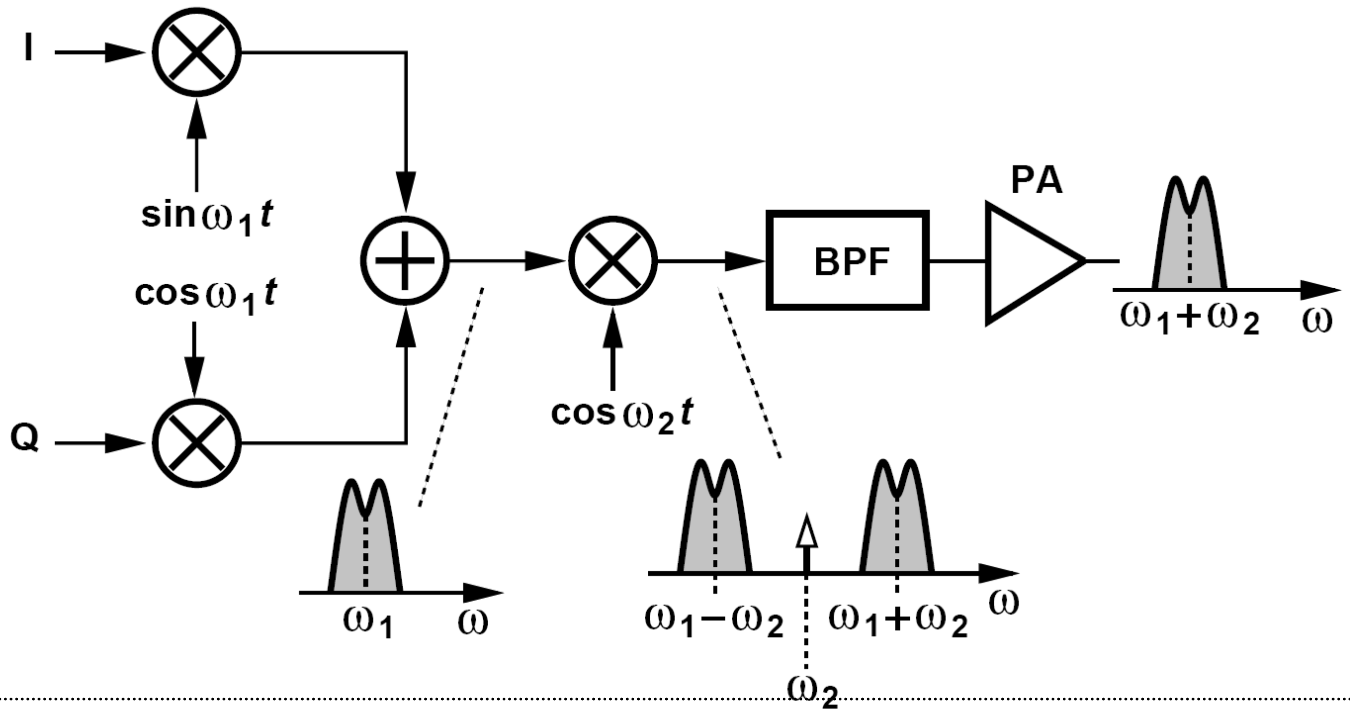
- Most of today's direct-conversion transmitters avoid an oscillator frequency equal to the PA output frequency.
- This architecture is popular for two reasons:

  (1) injection pulling is greatly reduced

  (2) the divider readily provides quadrature phases of the carrier

Heterodyne Transmitters

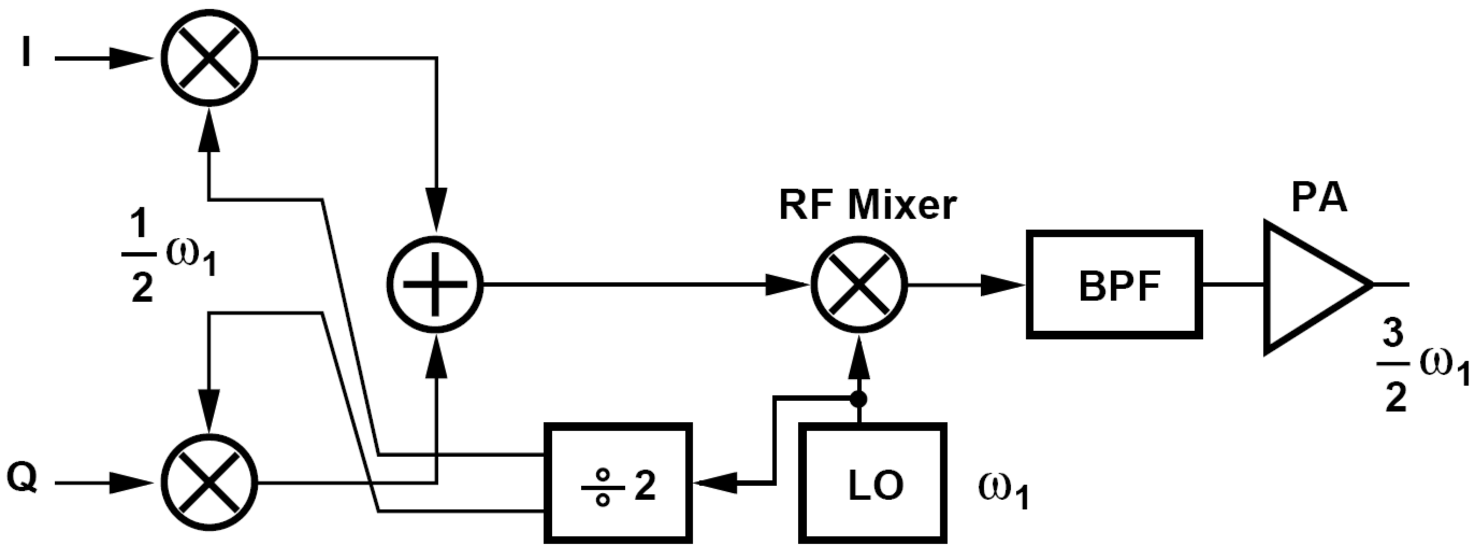
Perform the signal upconversion in two steps so that the LO frequency remains far from the PA output spectrum



- **As with the receiver counterpart, one advantage of this architecture is that the *I/Q* upconversion occurs at a significantly lower frequency than the carrier, exhibiting smaller gain and phase mismatches.**

Sliding-IF TX

In analogy with the sliding-IF receiver architecture, we eliminate the first oscillator in the above TX and derive the required phases from the second oscillator



- **We call the LO waveforms at $\omega_1/2$ and ω_1 the first and second LOs, respectively.**