Transceiver Architectures

Outline

Heterodyne Receivers

- ✓ Problem of Image
- ✓ Mixing Spurs
- ✓ Sliding-IF RX

Direct-Conversion Receivers

- ✓ LO Leakage and Offsets
- ✓ Even-Order Nonlinearity
- ✓ I/Q Mismatch

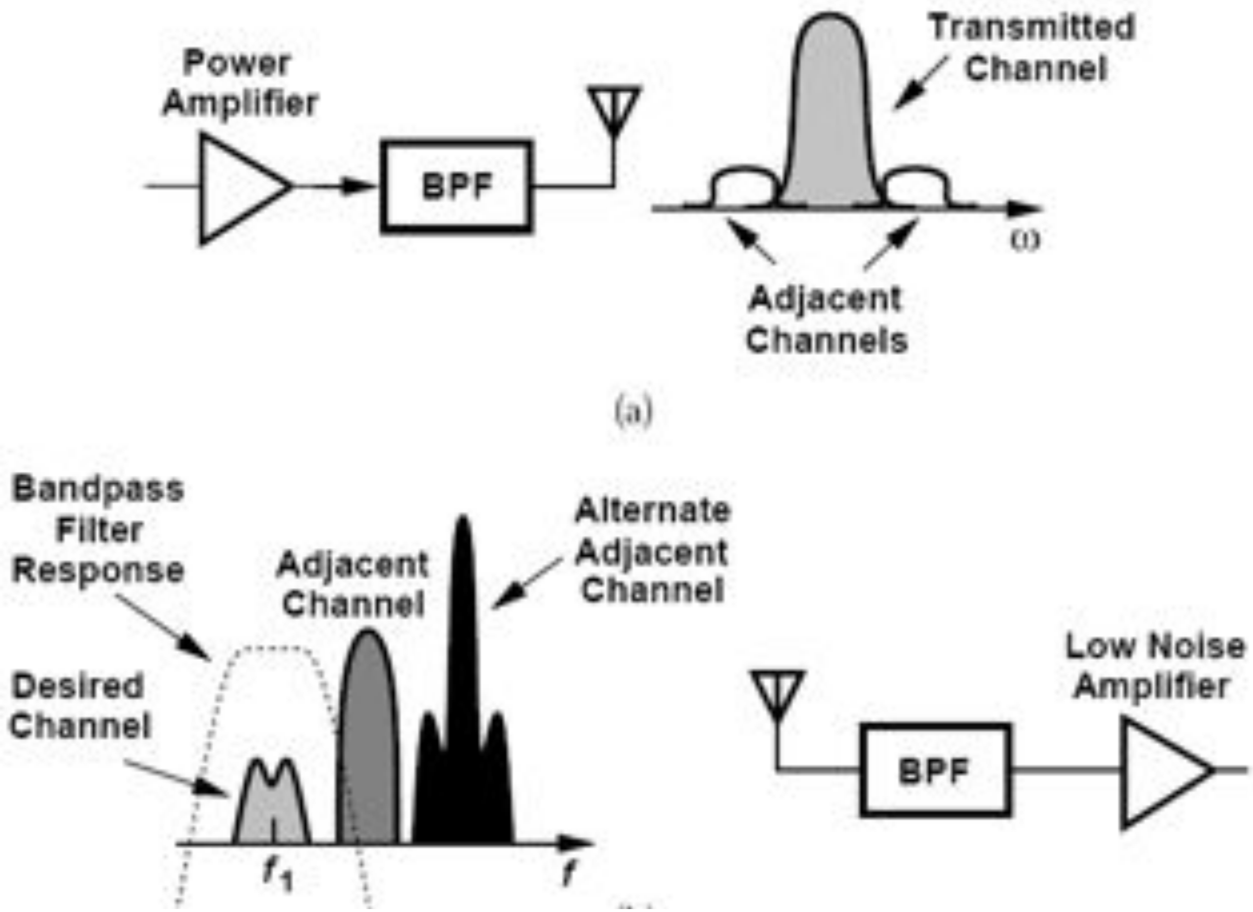
Image-Reject and Low-IF Receivers

- ✓ Hartley and Weaver Receivers
- ✓ Low-IF Receivers
- ✓ Polyphase Filters

Transmitter Architecture

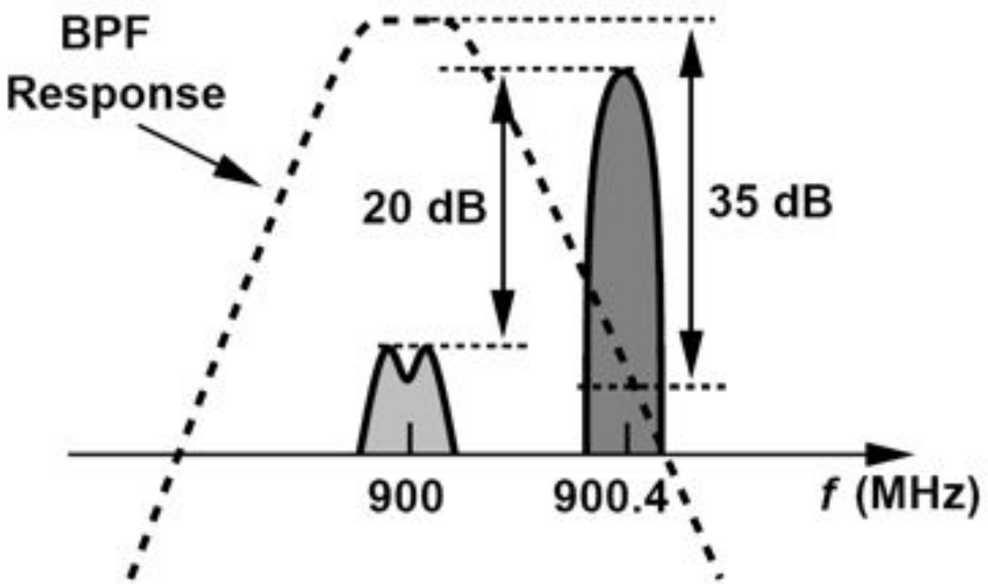
- ✓ TX Baseband Processing
- ✓ Direct-Conversion TX
- ✓ Heterodyne and Sliding-IF TX

General Considerations: Narrow Channel Bandwidth



- Narrow channel bandwidth impacts the RF design of the transceiver.

Can We Simply Filter the Interferers to Relax the Receiver Linearity Requirement?



- **First, the filter must provide a very high *Q***
- **Second, the filter would need a variable, yet precise center frequency**

Example of Interferer-Suppressing Filter

A 900-MHz GSM receiver with 200-kHz channel spacing must tolerate an alternate adjacent channel blocker 20 dB higher than the desired signal. Calculate the *Q* of a second-order LC filter required to suppress this interferer by 35 dB.

Solution:

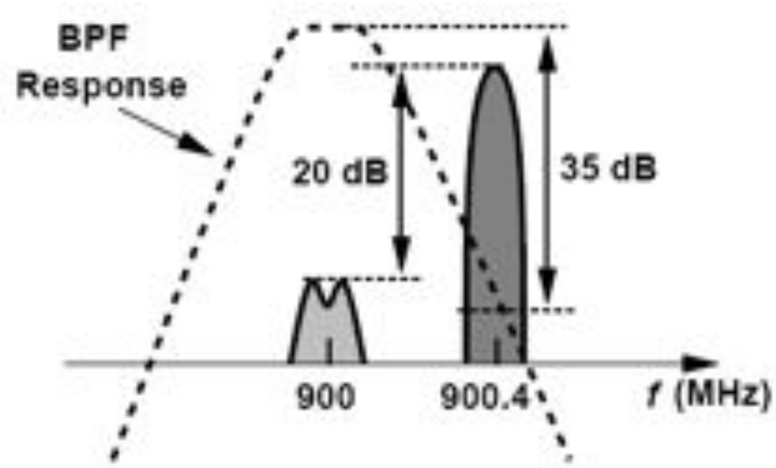
As shown in below, the filter frequency response must provide an attenuation of 35 dB at 400 kHz away from center frequency of 900MHz. For a second-order RLC tank, we write the impedance as

$$Z_T(s) = \frac{RLs}{RLCs^2 + Ls + R}$$

$$\omega_0 = 1/\sqrt{LC} = 2\pi(900\ \text{MHz})$$

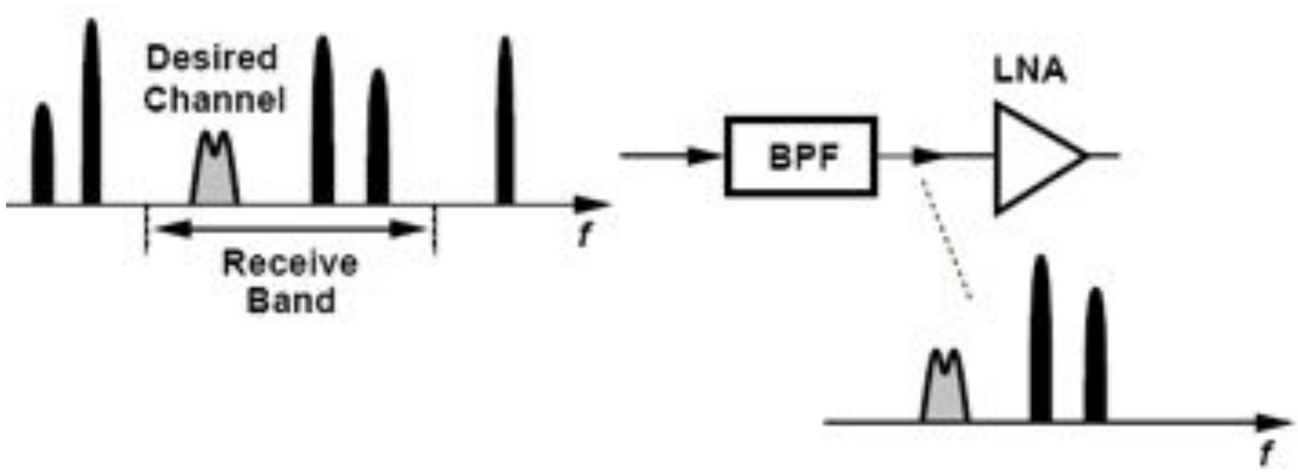
$$|Z_T(j\omega)|^2 = \frac{L^2\omega^2}{(1 - LC\omega^2)^2 + L^2\omega^2/R^2}$$

$$\frac{L^2\omega^2}{R^2} = 2.504 \times 10^{-10}$$



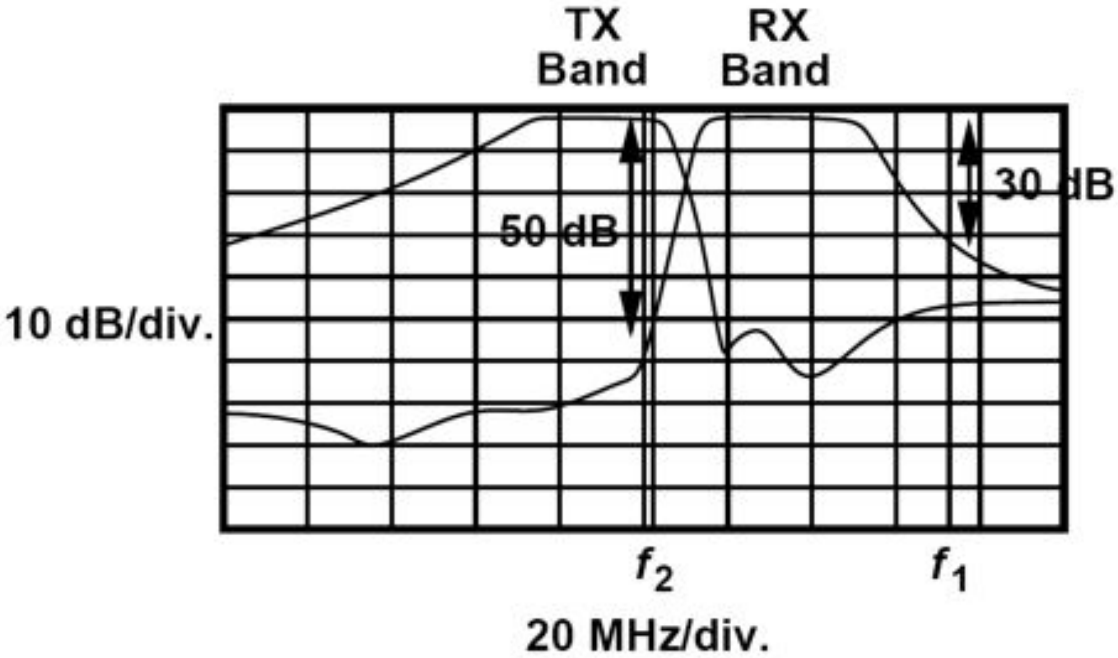
$$Q = R/(L\omega) = 63,200$$

Channel Selection and Band Selection



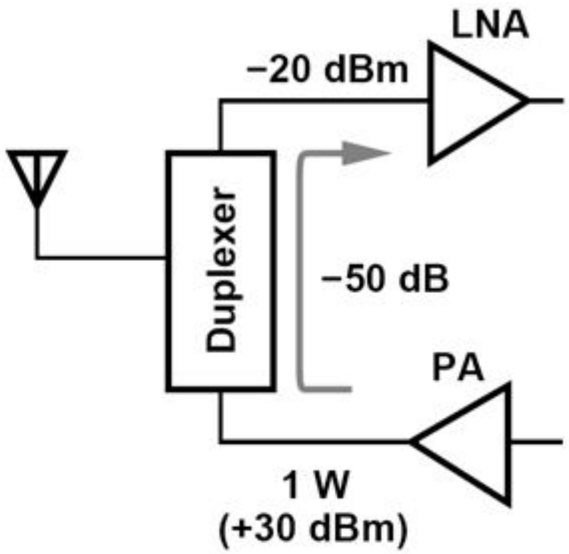
- All of the stages in the receiver chain that precede channel-selection filtering must be sufficiently linear
- Channel selection must be deferred to some other point where center frequency is lower and hence required *Q* is more reasonable
- Most receiver front ends do incorporate a “band-select” filter

Duplexer Characteristics



- The front-end band-select filter suffers from a trade-off between its selectivity and its in-band loss because the edges of the band-pass frequency response can be sharpened only by increasing the order of the filter.
- Front-end loss directly raises the *NF* of the entire receiver

TX-RX Feedthrough



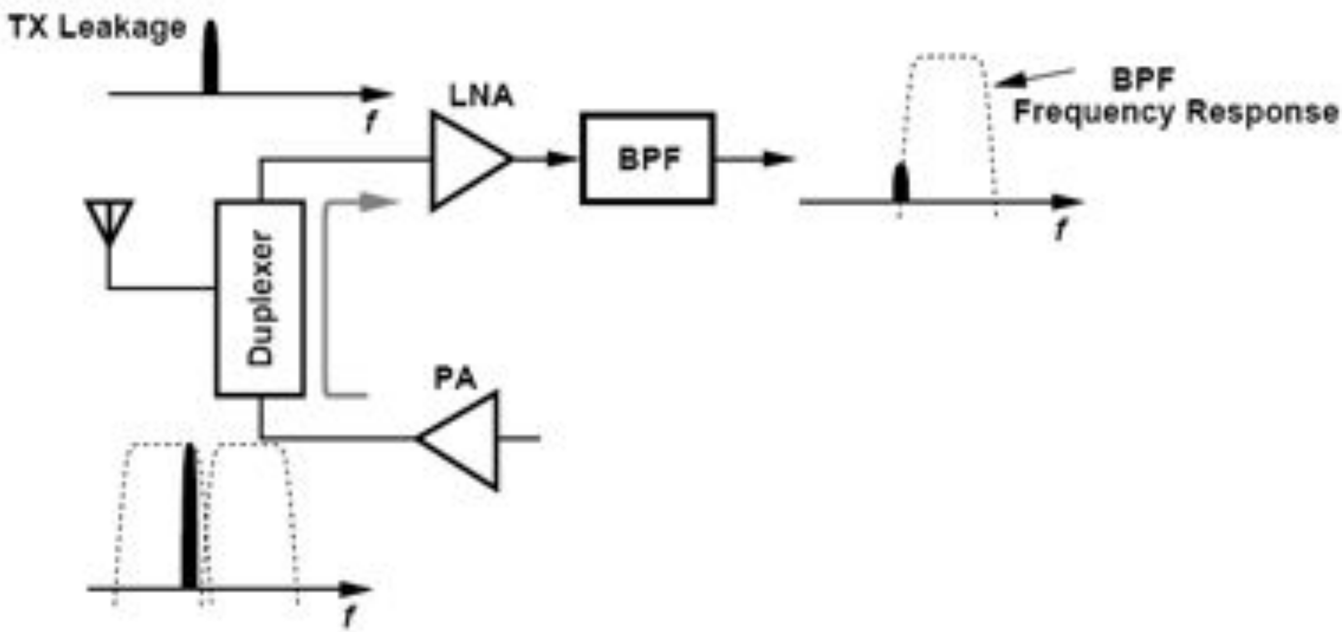
- In full-duplex standards, the TX and the RX operate concurrently.
- With a 1-W TX power, the leakage sensed by LNA can reach -20dBm, dictating a substantially higher RX compression point.

An Example of TX-RX Leakage

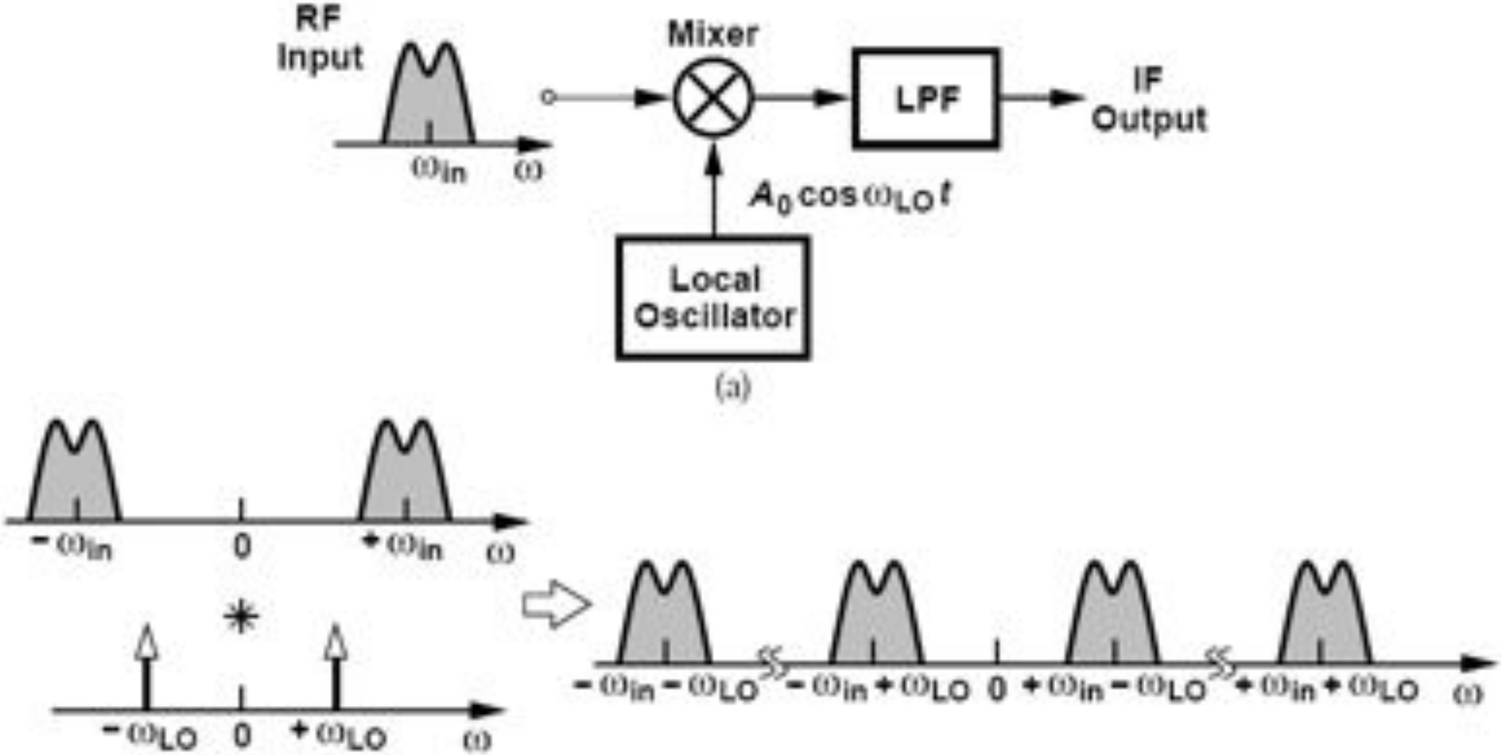
Explain how a band-pass filter following the LNA can alleviate the TX-RX leakage in a CDMA system.

Solution:

As depicted in below, if the BPF provides additional rejection in the TX band, the linearity required of the rest of the RX chain is proportionally relaxed. The LNA compression point, however, must still be sufficiently high.

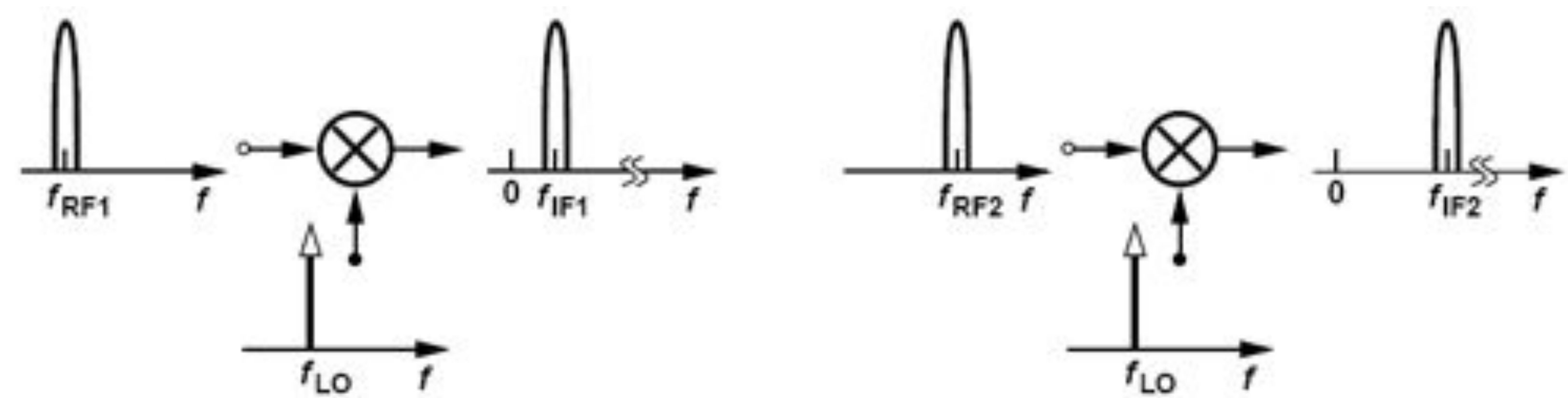


Basic Heterodyne Receivers

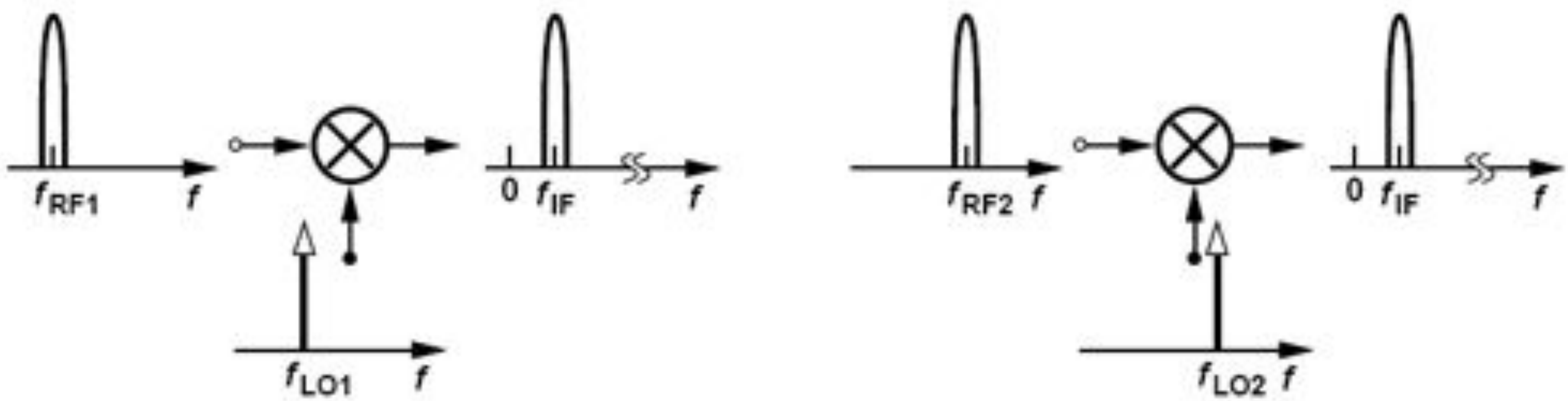


- "Heterodyne" receivers employ an LO frequency unequal to ω_{in} and hence a nonzero IF
- A Mixer performing downconversion.
- Due to its high noise, the downconversion mixer is preceded by a low-noise amplifier

How Does a Heterodyne Receiver Cover a Given Frequency Band?

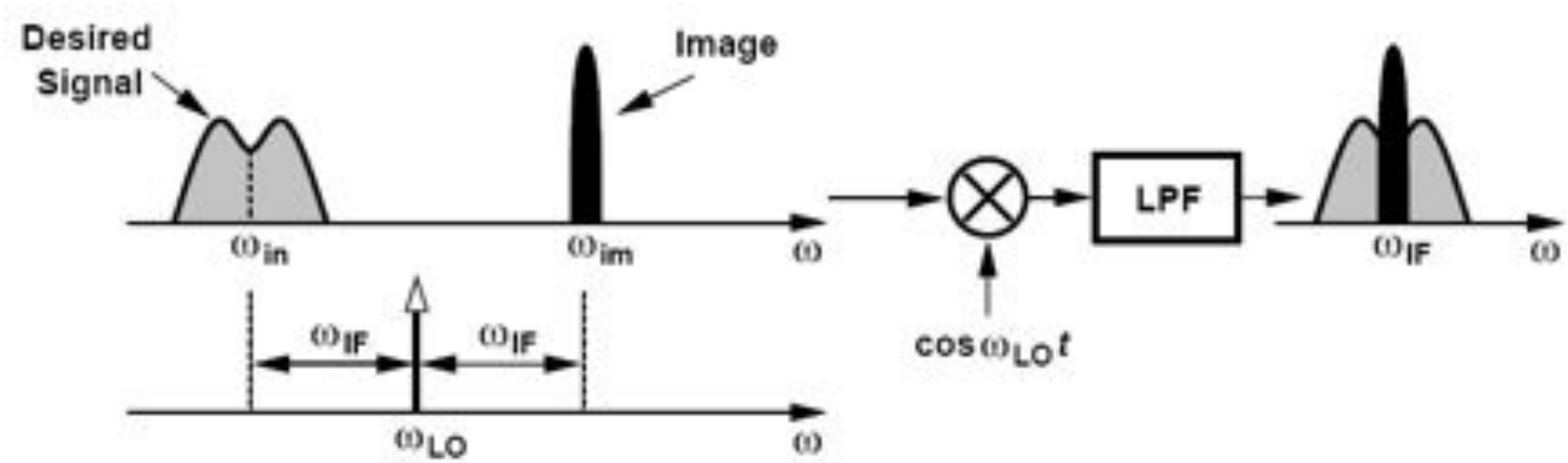


- **Constant LO: each RF channel is downconverted to a different IF channel**



- **Constant IF: LO frequency is variable, all RF channels within the band of interest translated to a single value of IF.**

Basic Heterodyne Receivers: Problem of Image



$$A\cos\omega_{IF}t = A\cos(\omega_{in}-\omega_{LO})t$$
$$= A\cos(\omega_{LO}-\omega_{in})t$$

$$\omega_{im} = \omega_{in} + 2\omega_{IF} = 2\omega_{LO} - \omega_{in}$$

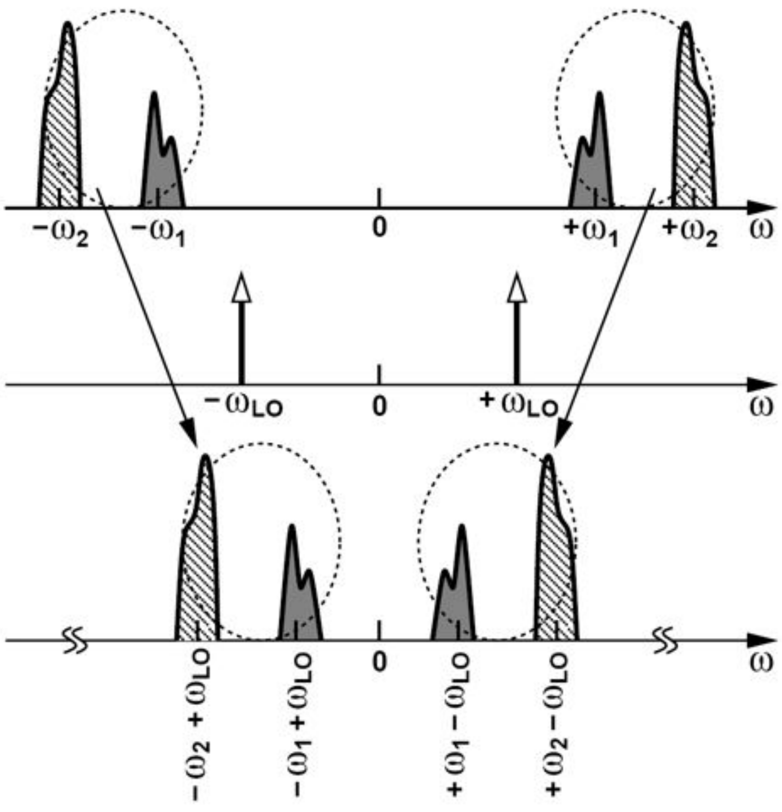
- **Two spectra located symmetrically around ω_{LO} are downconverted to the IF**

An Example of Image (Ⅰ)

Suppose two channels at ω_1 and ω_2 have been received and $\omega_1 < \omega_2$. Study the downconverted spectrum as the LO frequency varies from below ω_1 to above ω_2.

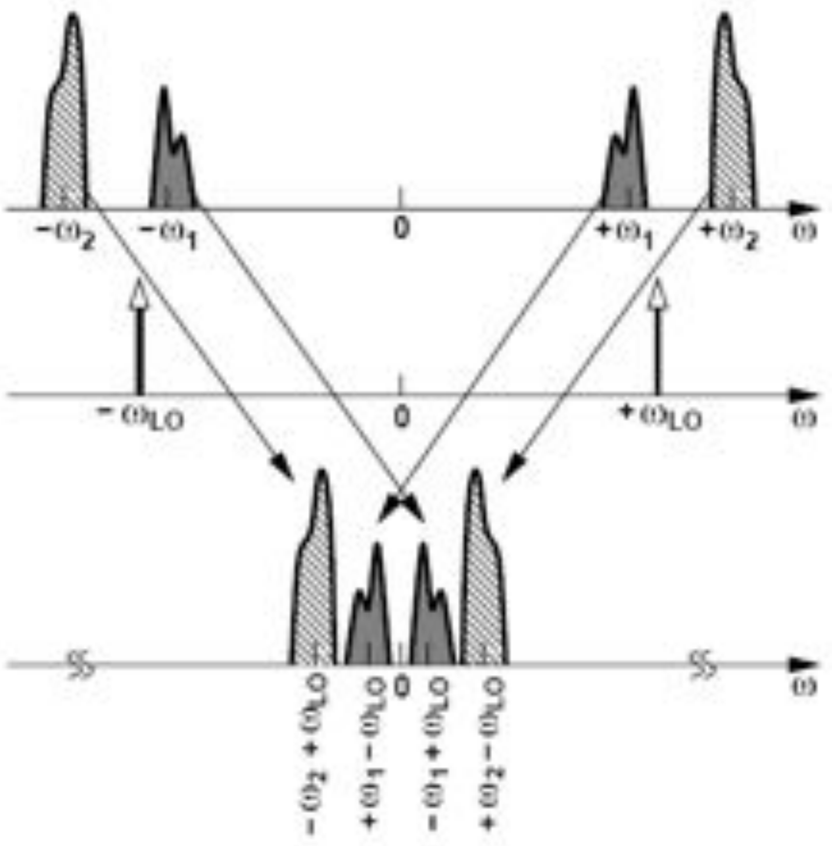
Solution:

$\omega_{LO} < \omega_1$



(a)

ω_{LO} slightly above ω_1



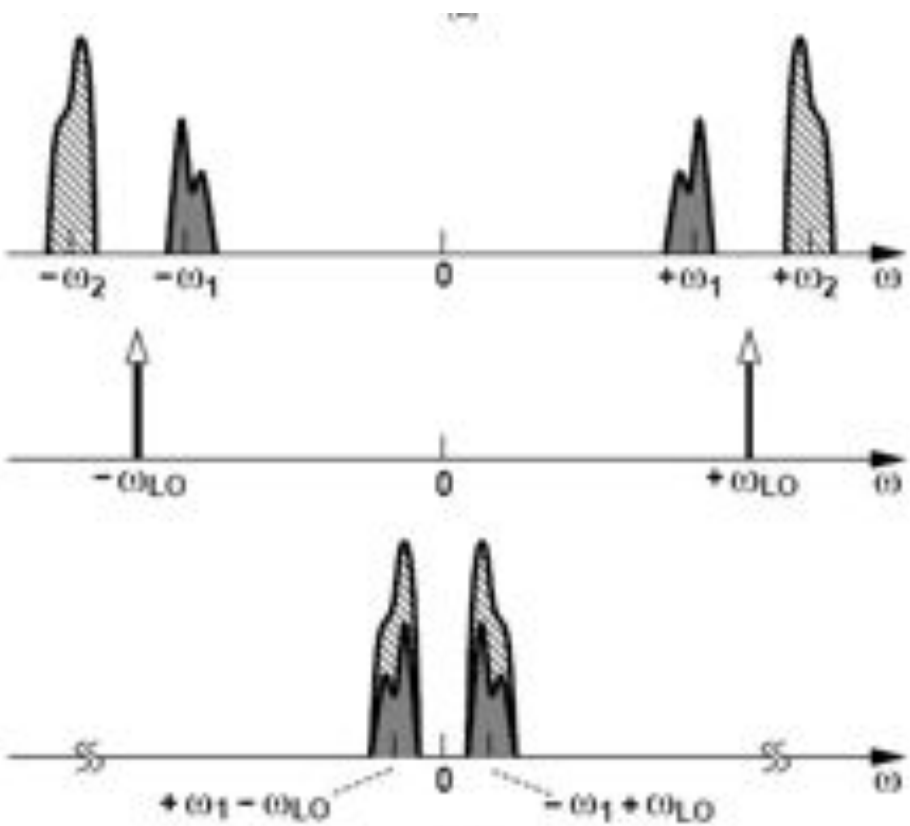
(b)

An Example of Image (Ⅱ)

Suppose two channels at ω_1 and ω_2 have been received and $\omega_1 < \omega_2$. Study the downconverted spectrum as the LO frequency varies from below ω_1 to above ω_2.

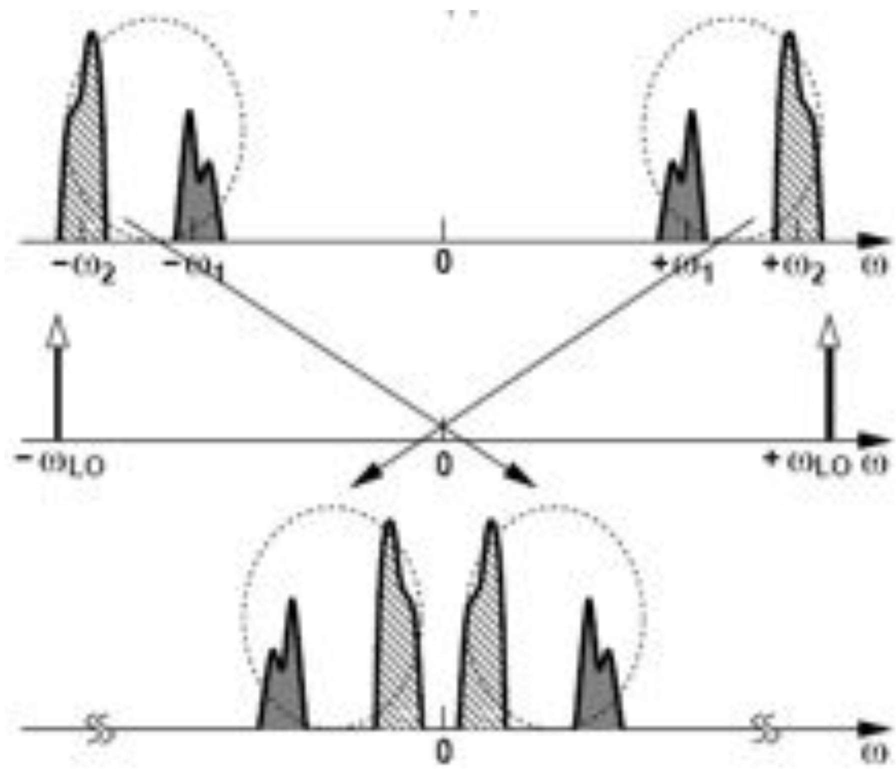
Solution:

ω_{LO} midway between ω_1 and ω_2



(c)

$\omega_{LO} > \omega_2$



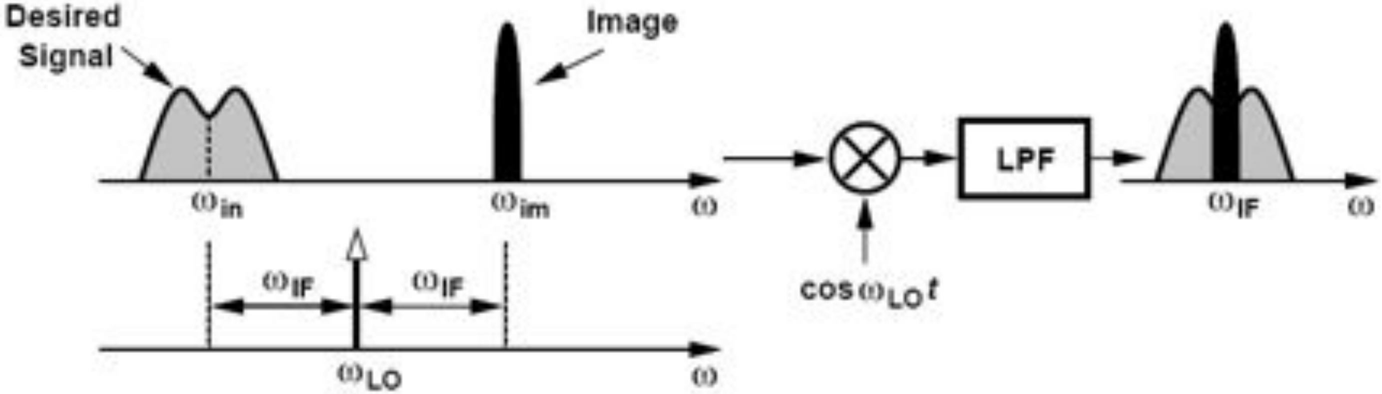
(d)

Another Example of Image

Formulate the downconversion above using expressions for the desired signal and the image.

Solution: $A_{in}(t)\cos[\omega_{in}t + \phi_{in}(t)]$ **and** $A_{im}(t)\cos[\omega_{im}t + \phi_{im}(t)]$

$$\begin{aligned} x_{IF}(t) &= \frac{1}{2}A_{in}(t)A_{LO}\cos[(\omega_{in}+\omega_{LO})t + \phi_{in}(t)] - \frac{1}{2}A_{in}(t)A_{LO}[\cos(\omega_{in}-\omega_{LO})t + \phi_{in}t] \\ &+ \frac{1}{2}A_{im}(t)A_{LO}\cos[(\omega_{im}+\omega_{LO})t + \phi_{im}(t)] \\ &- \frac{1}{2}A_{im}(t)A_{LO}[\cos(\omega_{im}-\omega_{LO})t + \phi_{im}t]. \end{aligned}$$



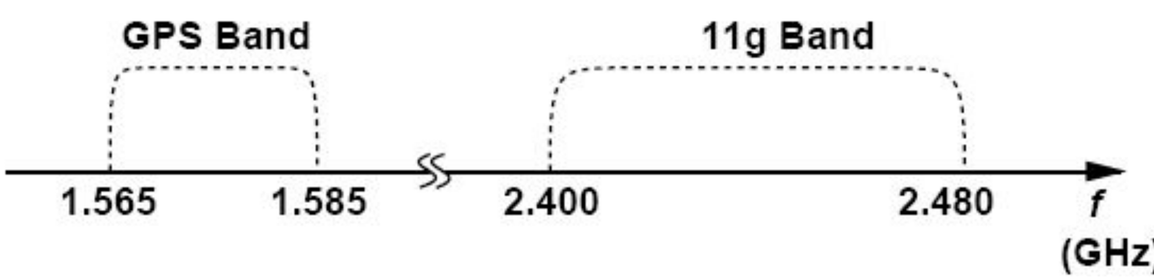
We observe that the components at $\omega_{in} + \omega_{LO}$ and $\omega_{im} + \omega_{LO}$ are removed by low-pass filtering, and those at $\omega_{in} - \omega_{LO} = -\omega_{IF}$ and $\omega_{im} - \omega_{LO} = +\omega_{IF}$ coincide.

An Example of High-Side and Low-Side Injection

The designer of an IEEE802.11g receiver attempts to place the image frequency in the GPS band, which contains only low-level satellite transmissions and hence no strong interferers. Is this possible?

Solution:

The two bands are shown below. The LO frequency must cover a range of 80MHz but, unfortunately, the GPS band spans only 20 MHz. For example, if the lowest LO frequency is chosen so as to make 1.565 GHz the image of 2.4 GHz, then 802.11g channels above 2.42 GHz have images beyond the GPS band.



Another Example of High-Side and Low-Side Injection

A dual-mode receiver is designed for both 802.11g and 802.11a. Can this receiver operate with a single LO?

The following figure (top) depicts the two bands. We choose the LO frequency halfway between the two so that a single LO covers the 11g band by high-side injection and the 11a band by low-side injection. This greatly simplifies the design of the system, but makes each band the image of the other. For example, if the receiver is in the 11a mode while an 11g transmitter operates in close proximity, the reception may be heavily corrupted. Note that also the IF in this case is quite high, an issue revisited later.

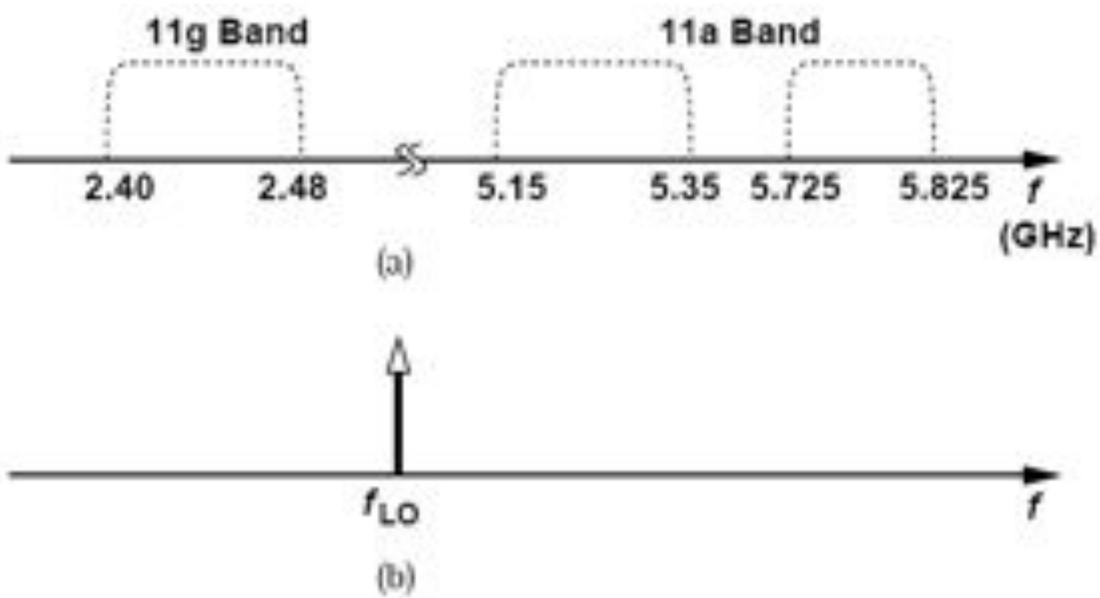
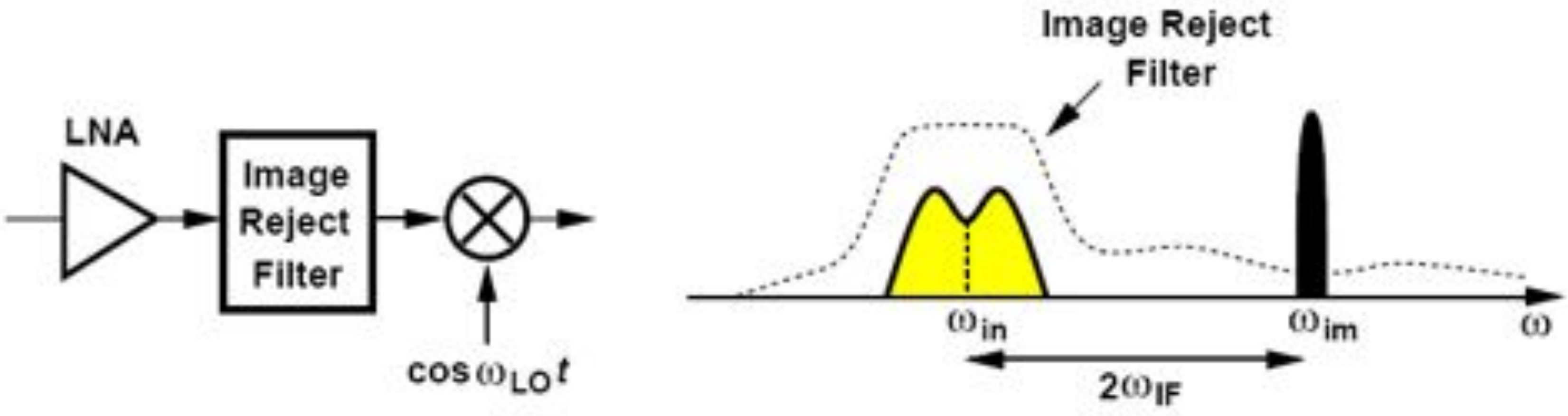
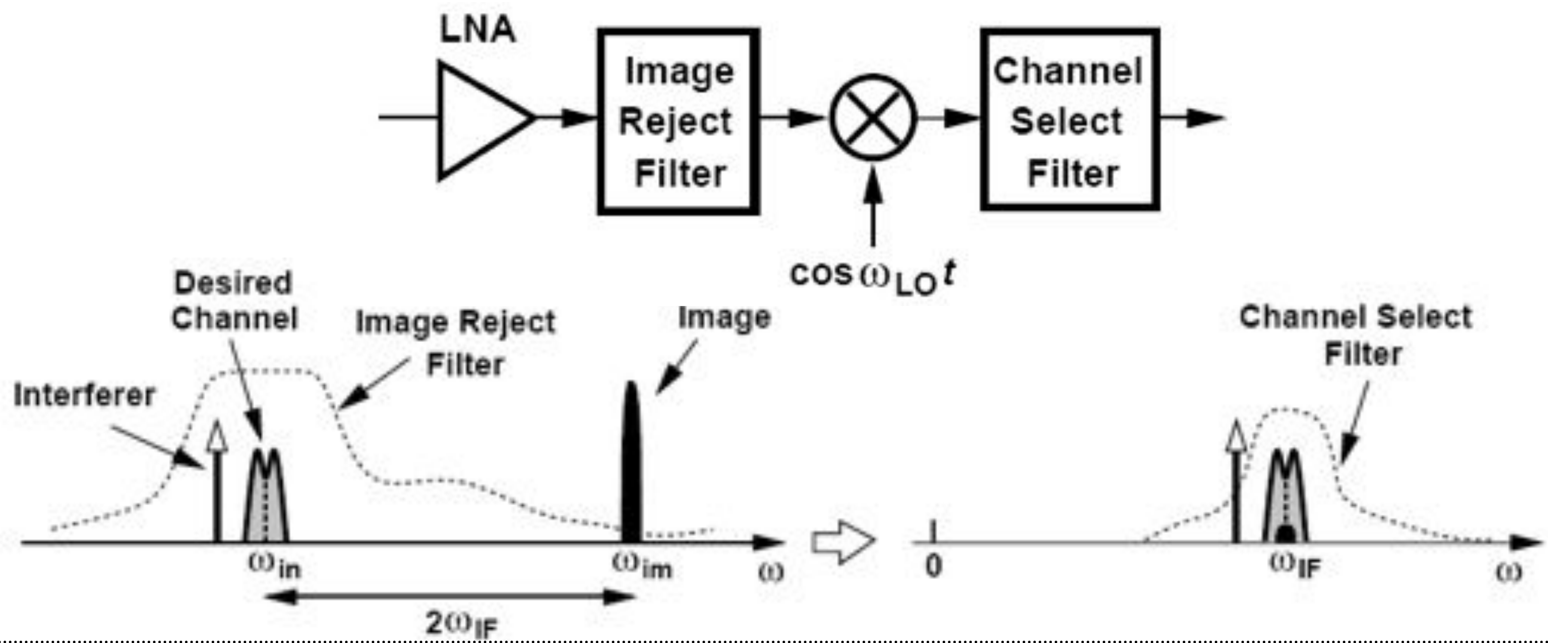


Image Rejection

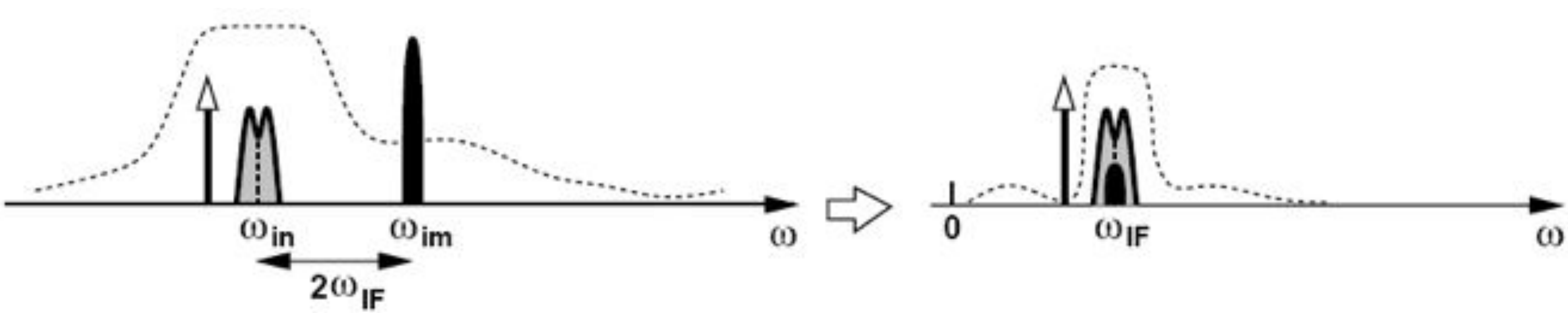


- The most common approach is to precede the mixer with an “image-reject filter”
- A filter with high image rejection typically appears between the LNA and the mixer so that the gain of the LNA lowers the filter’s contribution to the receiver noise figure
- The linearity and selectivity required of the image-reject filter have dictated passive, off-chip implementations.

Image Rejection versus Channel Selection



➢ **A high IF allows substantial rejection of the image.**

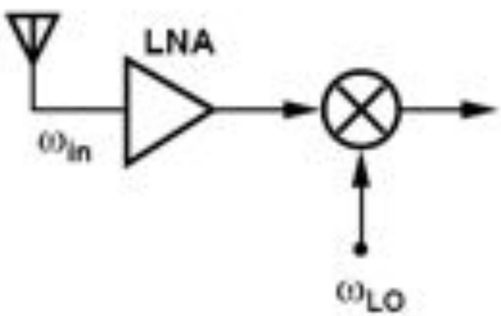


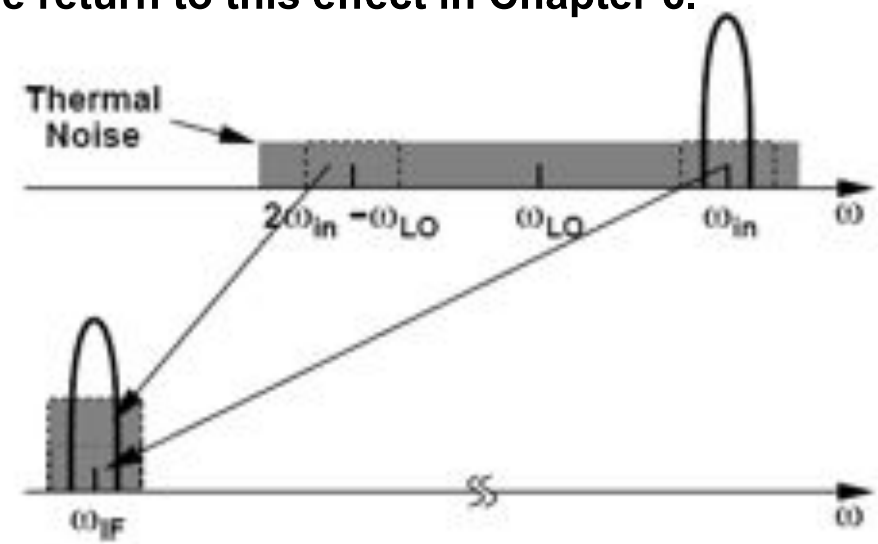
➢ **A low IF helps with the suppression of in-band interferers.**

An Example of Noise Figure in Receiver

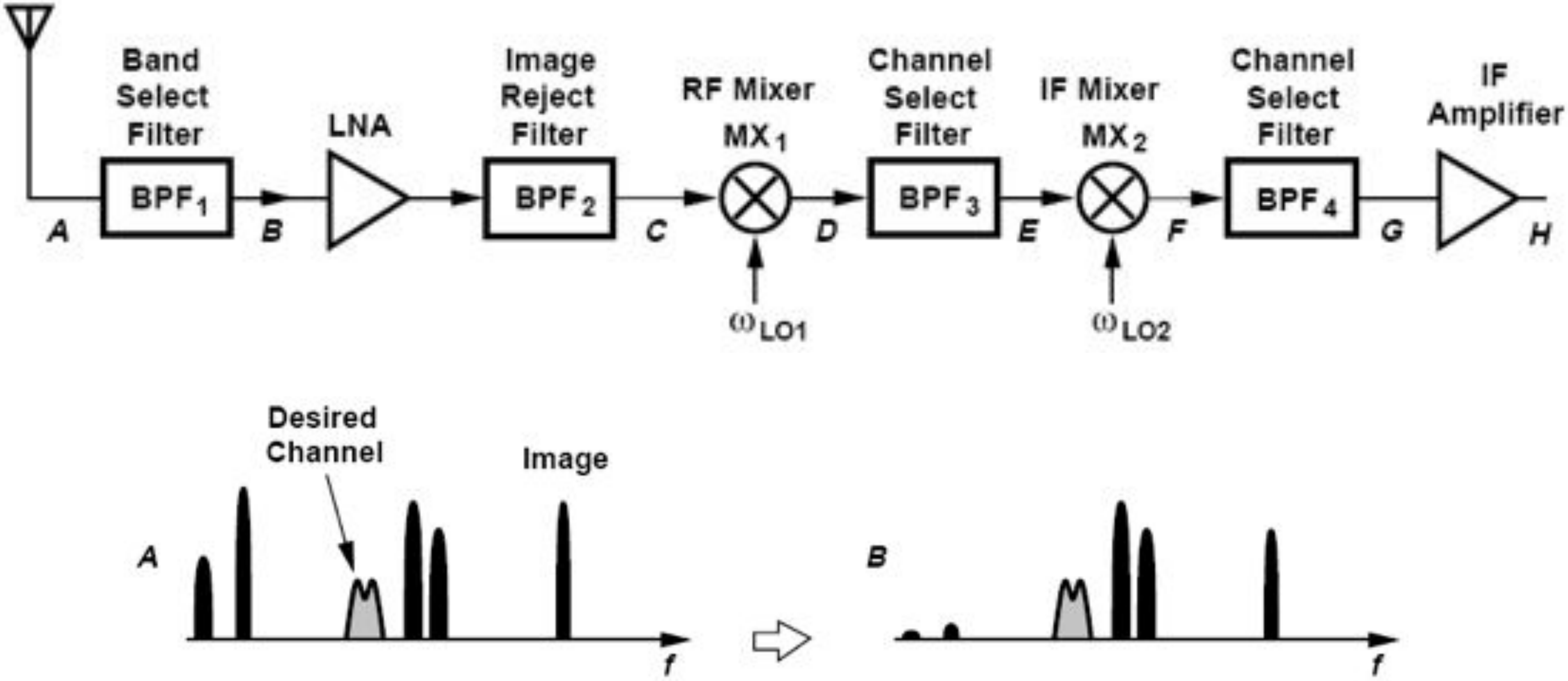
An engineer is to design a receiver for space applications with no concern for interferers. The engineer constructs the heterodyne front end shown in figure below (left), avoiding band-select and image-select filters. Explain why this design suffers from a relatively high noise figure.

Even in the absence of interferers, the thermal noise produced by the antenna and the LNA in the image band arrives at the input of the mixer. Thus, the desired signal, the thermal noise in the desired channel, and the thermal noise in the image band are downconverted to IF, leading to a higher noise figure for the receiver (unless the LNA has such a limited bandwidth that it suppresses the noise in the image band). An image-reject filter would remove the noise in the image band. We return to this effect in Chapter 6.



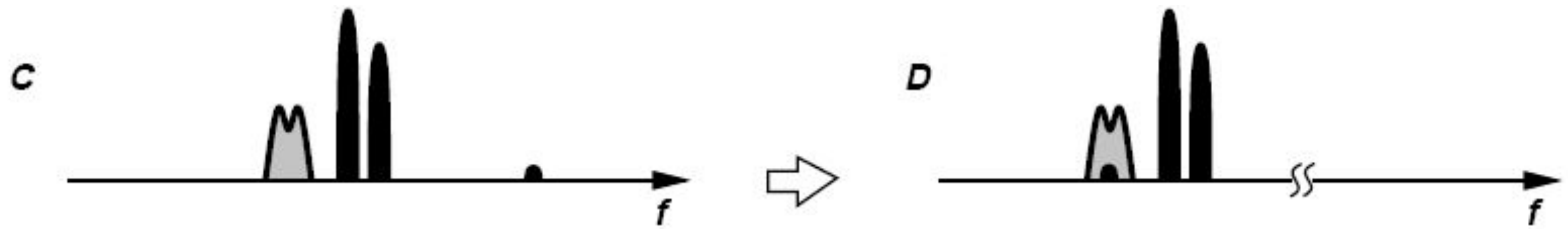


Dual Downconversion (Ⅰ)



- **The front-end filter selects the band while providing some image rejection as well (Point *B*)**

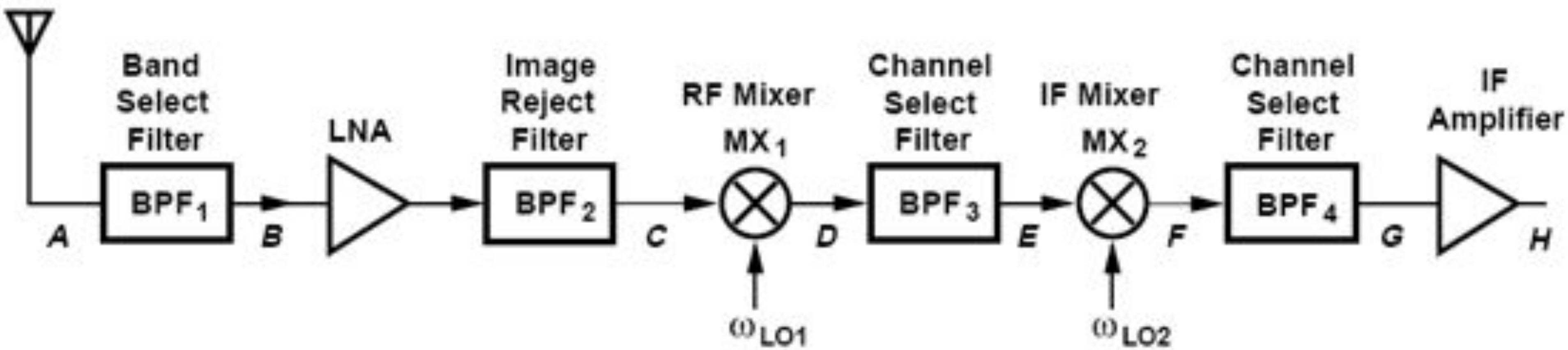
Dual Downconversion (Ⅱ)



- After amplification and image-reject filtering, spectrum of *C* obtained
- Sufficiently linear mixer translates desired channel and adjacent interferers to first IF (Point *D*)

- Partial channel selection BPF_3 permits the use of a second mixer with reasonable linearity. (Point *E*)
- Spectrum is translated to second IF. (Point *F*)

- BPF_4 suppresses the interferers to acceptably low levels (Point *G*)
- An optimum design scales both the noise figure and the IP3 of each stage according to the total gain preceding that stage.

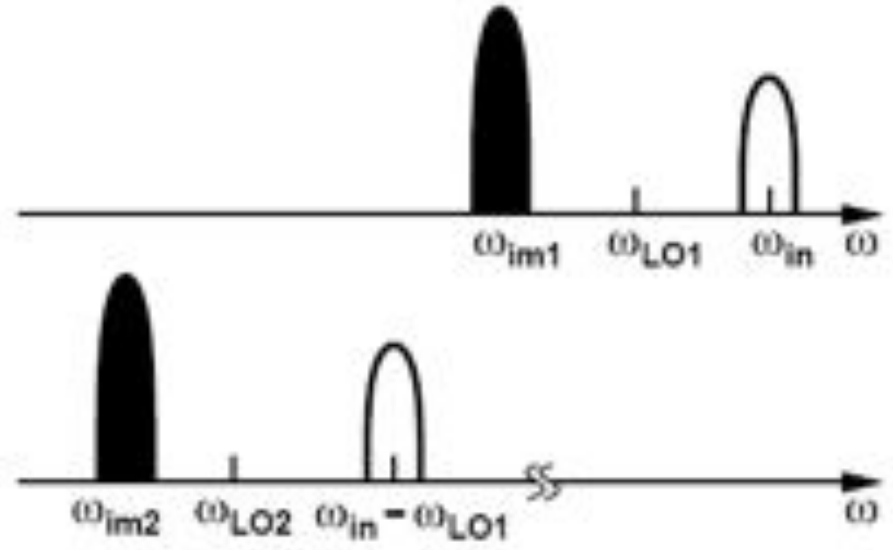


Another Example of Image

Assuming low-side injection for both downconversion mixers in figure above, determine the image frequencies.

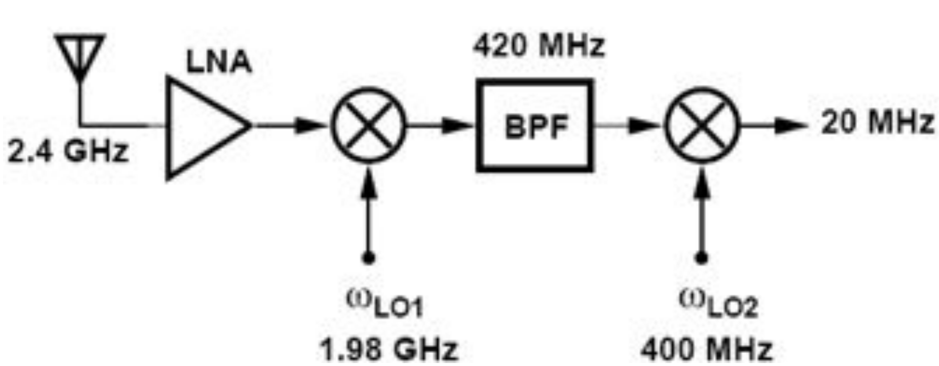
Solution:

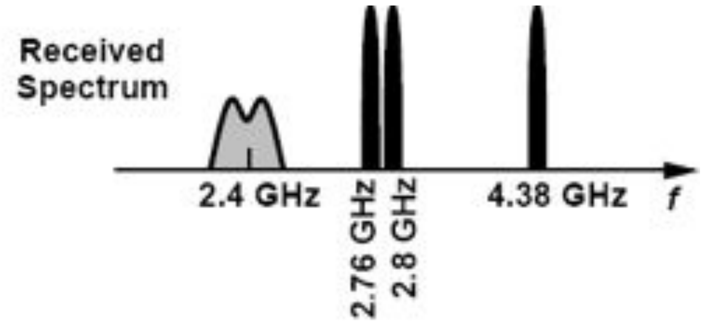
As shown below, the first image lies at $2\omega_{LO1} - \omega_{in}$. The second image is located at $2\omega_{LO2} - (\omega_{in} - \omega_{LO1})$.



Mixing Spurs and an Example

Figure below (left) shows a 2.4-GHz dual downconversion receiver, where the first LO frequency is chosen so as to place the (primary) image in the GPS band for some of the channels. Determine a few mixing spurs.





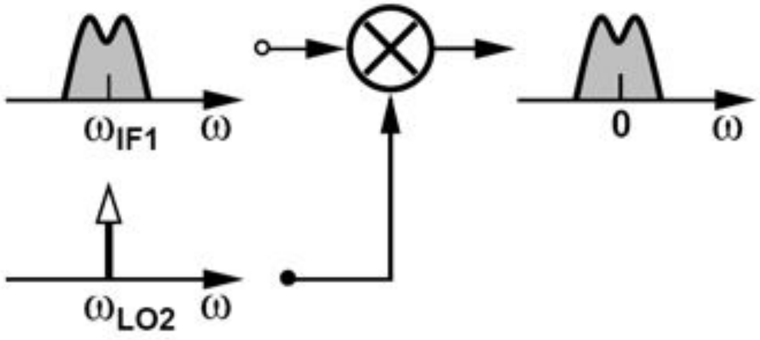
Let us consider the second harmonic of LO_2, 800MHz. If an interferer appears at the first IF at 820MHz or 780MHz, then it coincides with the desired signal at the second IF. In the RF band, the former corresponds to 820MHz+1980MHz = 2.8 GHz and the latter arises from 780MHz +1980MHz = 2.76 GHz. We can also identify the image corresponding to the second harmonic of LO_1 by writing $f_{in} - 2f_{LO1} - f_{LO2}$ = 20 MHz and hence f_{in} = 4.38 GHz. Figure above (right) summarizes these results. We observe that numerous spurs can be identified by considering other combinations of LO harmonics.

Mixing Spurs

$$\omega_{int} \pm m\omega_{LO1} \pm n\omega_{LO2} = \omega_{in} - \omega_{LO1} - \omega_{LO2}$$

- **RF input multiplied by a square-wave LO. Producing harmonics.**
- **If an interferer is downconverted to the same IF, it corrupts the signal**

Zero Second IF

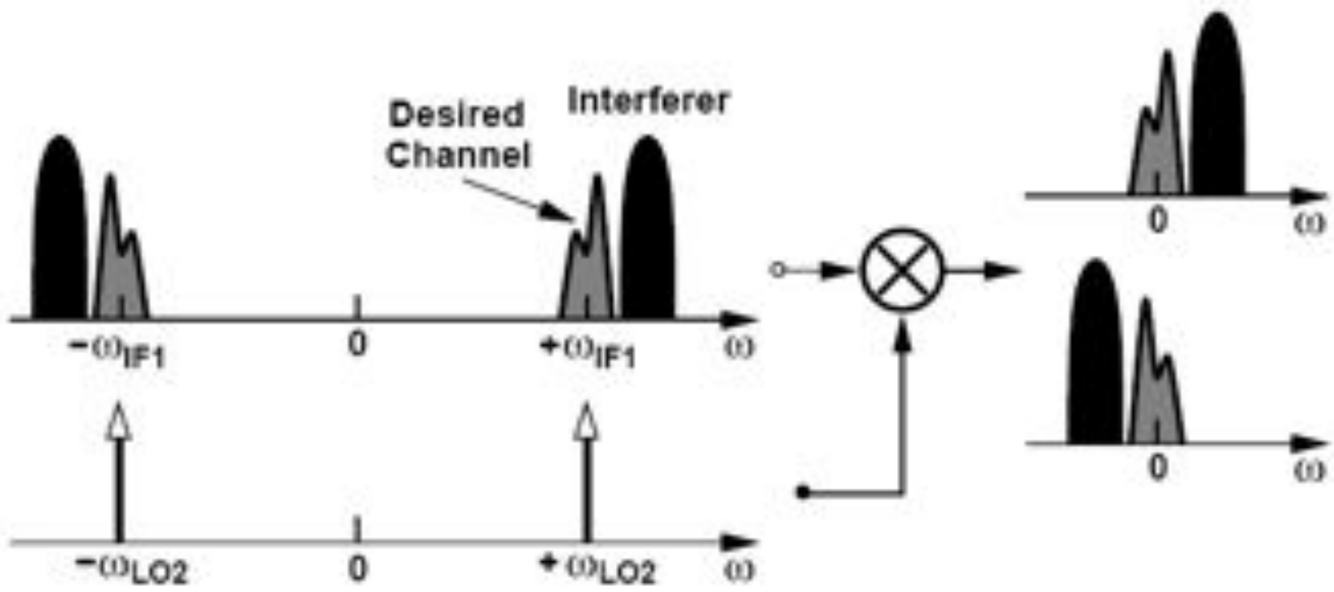


- To avoid secondary image, most modern heterodyne receivers employ a zero second IF.
- In this case, the image is the signal itself. No interferer at other frequencies can be downconverted as an image to a zero center frequency if $\omega_{LO2}=\omega_{IF1}$

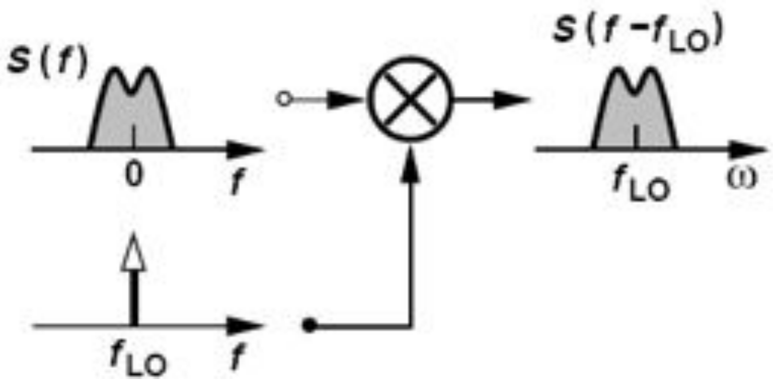
Example of Zero Second IF

Suppose the desired signal in figure above is accompanied by an interferer in the adjacent channel. Plot the spectrum at the second IF if $\omega_{LO2} = \omega_{IF1}$.

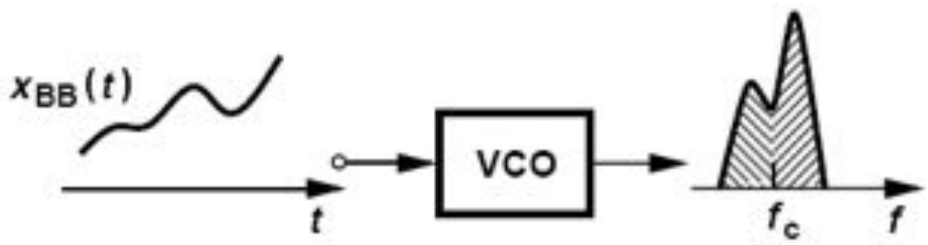
As shown below, the desired channel appears at $\pm\,\omega_{IF1}$ and is accompanied by the interferer. Upon mixing in the time domain, the spectrum at negative frequencies is convolved with the LO impulse at $+\omega_{LO2}$, sliding to a zero center frequency for the desired channel. Similarly, the spectrum at positive frequencies is convolved with the impulse at $-\omega_{LO2}$ and shifted down to zero. The output thus consists of two copies of the desired channel surrounded by the interferer spectrum at both positive and negative frequencies.



Symmetrically-modulated versus Asymmetrically-modulated Signal



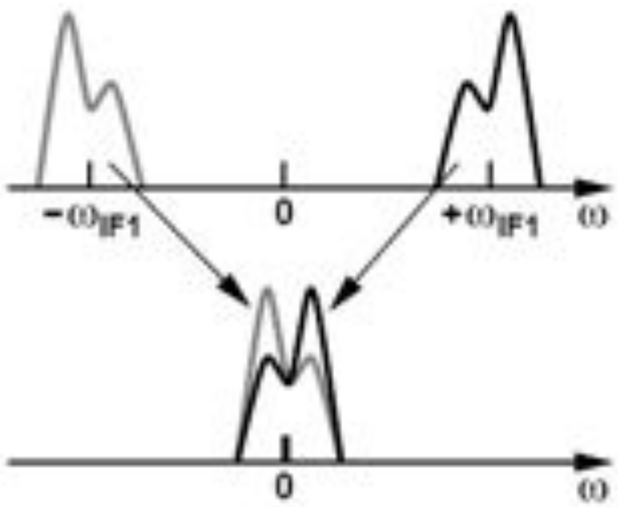
AM signal generation



FM signal generation

- **AM signals are symmetric, FM signals are asymmetric.**
- **Most of today's modulation schemes, e.g., FSK, QPSK, GMSK, and QAM, exhibit asymmetric spectra around carrier frequency.**

Corruption of the Asymmetric Signal Spectrum

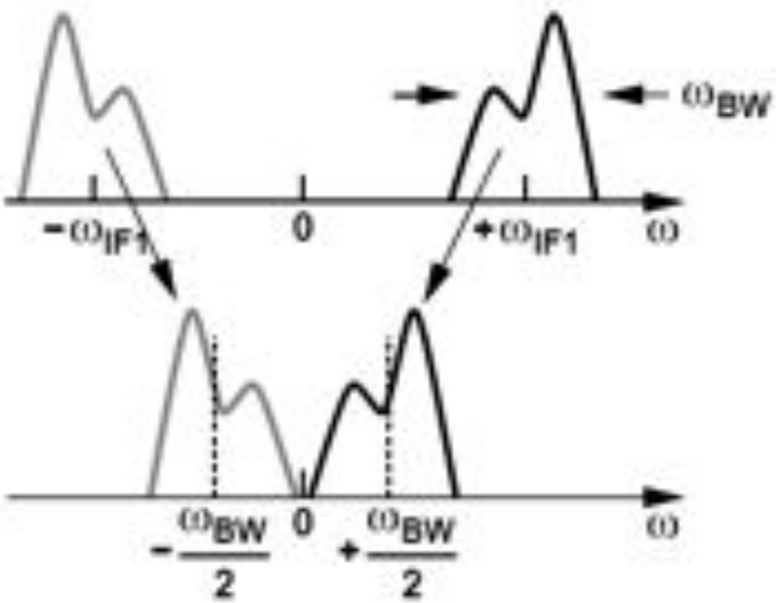


- Downconversion to a zero IF superimposes two copies of the signal
- If the signal spectrum is asymmetric, the original signal spectrum will be corrupted

An Example of Self-corruption

Downconversion to what minimum intermediate frequency avoids self-corruption of asymmetric signals?

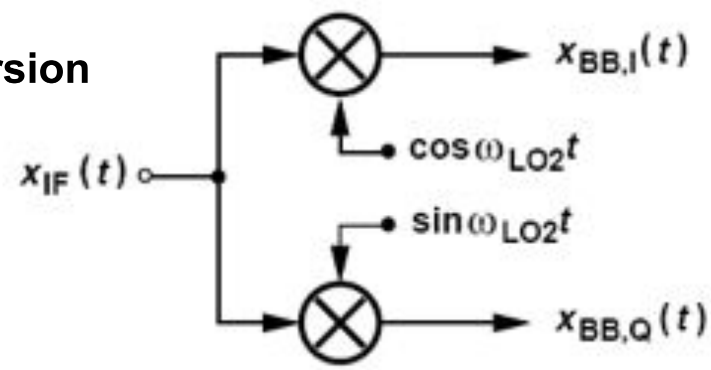
Solution:

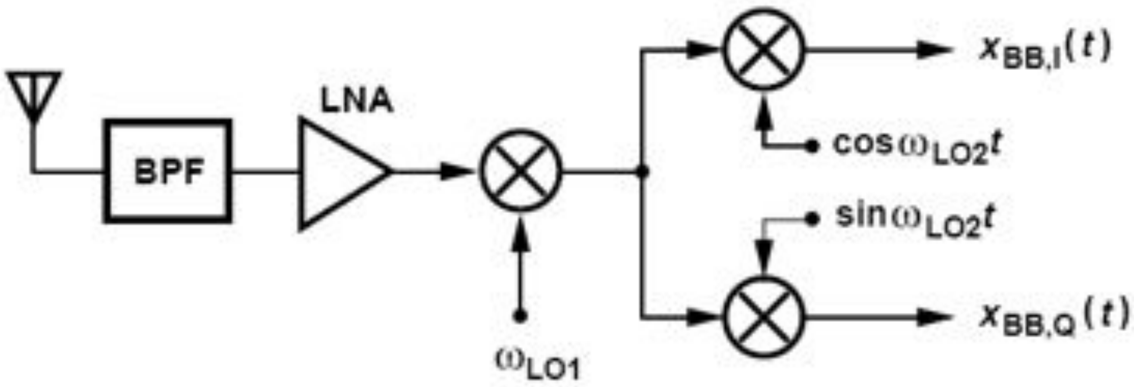


To avoid self-corruption, the downconverted spectra must not overlap each other. Thus, as shown in figure below, the signal can be downconverted to an IF equal to half of the signal bandwidth. Of course, an interferer may now become the image.

How can downconversion to a zero IF avoid self-corruption?

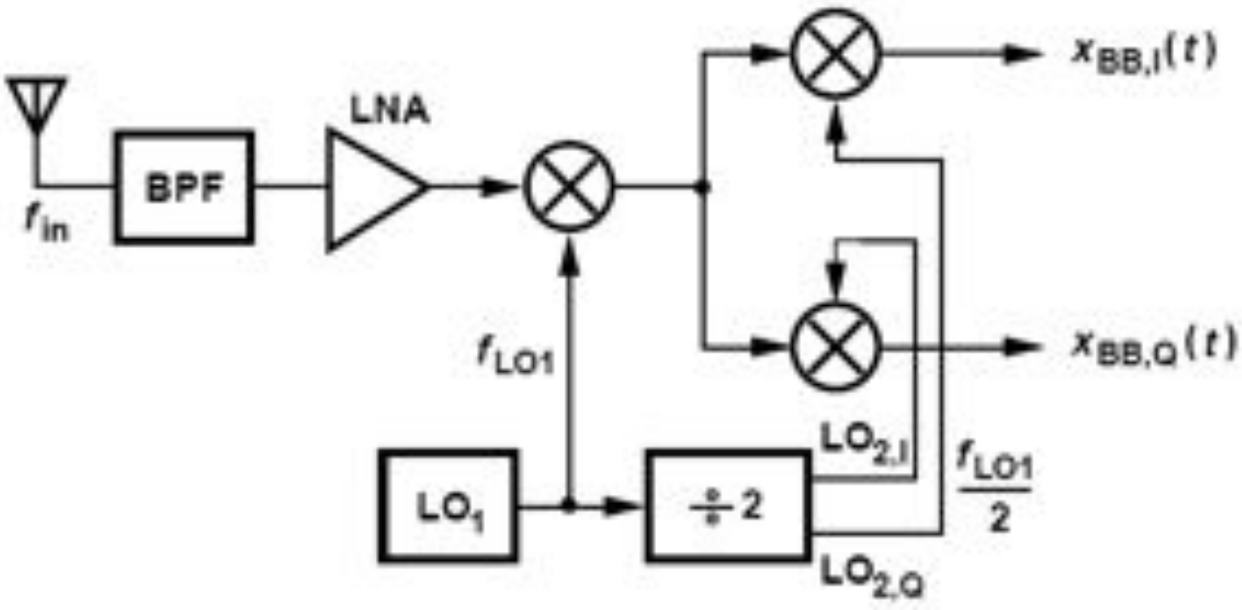
Quadrature downconversion





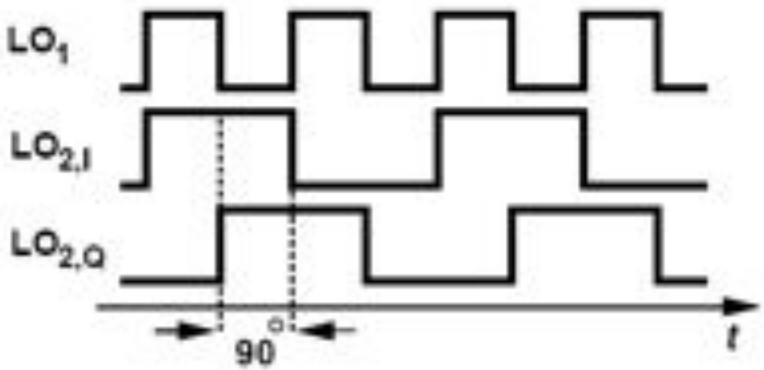
- **By creating two versions of the downconverted signal that have a phase difference of 90** ∘

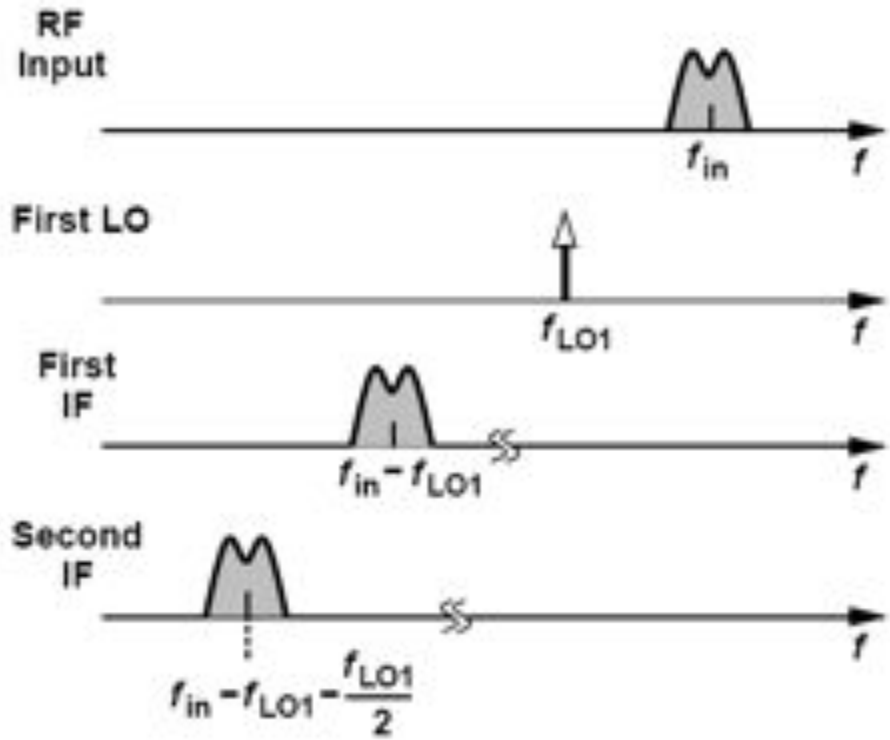
Sliding-IF Receivers



- Modern heterodyne receivers employ only one oscillator
- The second LO frequency is therefore derived from the first by “frequency division”

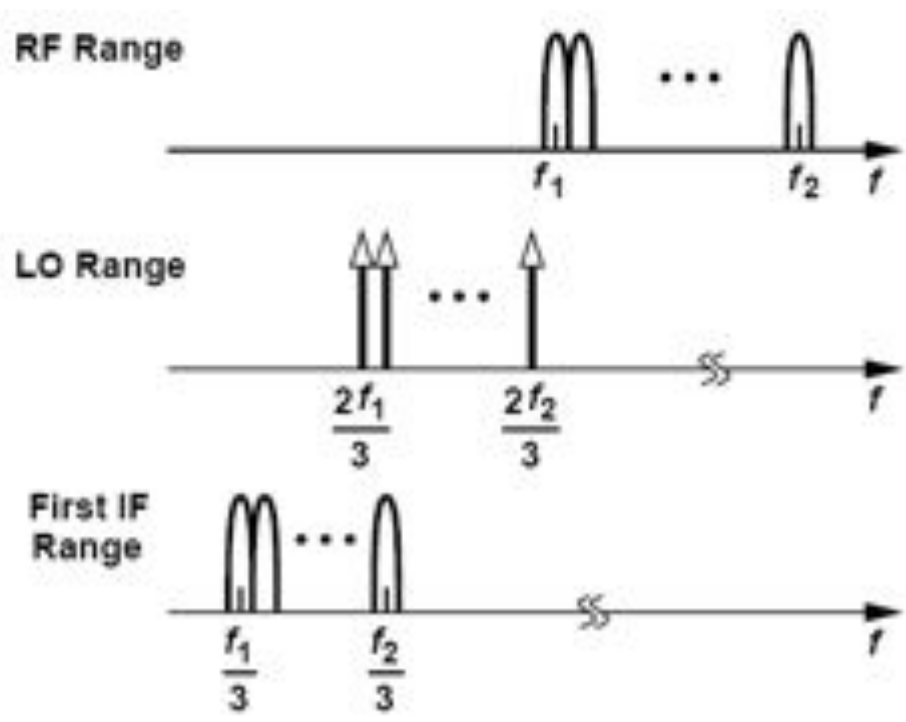
Sliding-IF Receivers: Divide-by-2 Circuit





- Such divide-by-2 topology can produce quadrature output
- The second LO waveforms at a frequency of $f_{LO1}/2$

Interesting Properties of Sliding-IF Receivers



$$f_{LO1} + \frac{1}{2} f_{LO1} = f_{in}$$

$$f_{LO1} = \frac{2}{3} f_{in}$$

$$f_{IF1} = f_{in} - f_{LO} = \frac{1}{3} f_{in}.$$

Fractional bandwidth:

IF

$$\mathrm{BW}_{IF,frac} = \frac{\frac{1}{3} f_2 - \frac{1}{3} f_1}{(\frac{1}{3} f_2 + \frac{1}{3} f_1)/2}$$

RF input

$$\mathrm{BW}_{RF,frac} = \frac{f_2 - f_1}{(f_2 + f_1)/2}$$

An Example of Sliding-IF Receivers

Suppose the input band is partitioned into N channels, each having a bandwidth of $(f_2 - f_1)/N = \Delta f$. How does the LO frequency vary as the receiver translates each channel to a zero second IF?

Solution:

The first channel is located between f_1 and $f_1 + \Delta f$. Thus the first LO frequency is chosen equal to two-third of the center of the channel: $f_{LO} = (2/3)(f_1 + \Delta f/2)$. Similarly, for the second channel, located between $f_1 + \Delta f$ and $f_1 + 2\Delta f$, the LO frequency must be equal to $(2/3)(f_1 + 3\Delta f/2)$. In other words, the LO increments in steps of $(2/3)\Delta f$.

Another Example of Sliding-IF Receivers

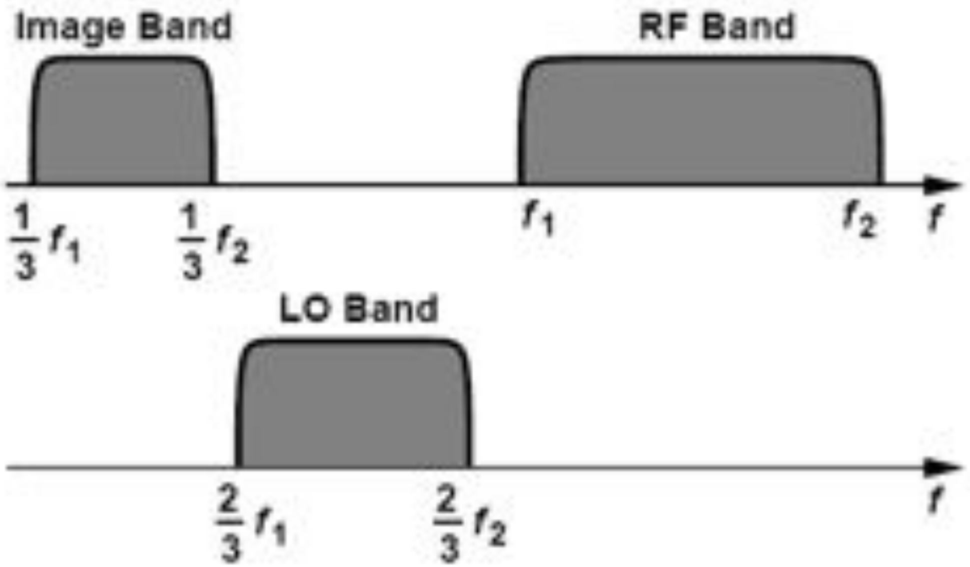
With the aid of the frequency bands shown in figure above, determine the image band for the architecture of the sliding-IF heterodyne RX.

Solution:

For an LO frequency of (2/3)f_1, the image lies at $2f_{LO} - f_{in} = f_1/3$. Similarly, if $f_{LO1} = (2/3)f_2$, then the image is located at $f_2/3$. Thus, the image band spans the range [$f_1/3$ $f_2/3$]. Interestingly, the image band is narrower than the input band.

- **Narrower image band than input band?**

No, recall from the above example that the LO increments by (2/3)Δf as we go from one channel to the next. Thus, consecutive image channels have an overlap of $\Delta f/3$.



Sliding-IF Receivers with Divide Ratio of 4

- May incorporate greater divide ratios, e.g., 4

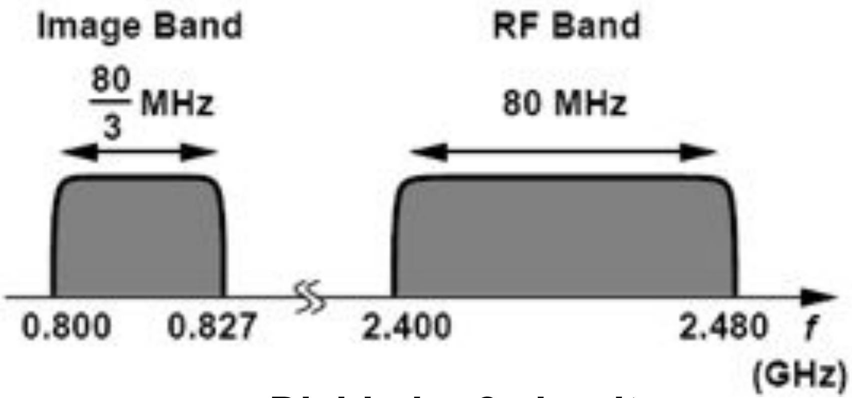
$$f_{LO1} + \frac{1}{4} f_{LO1} = f_{in}$$

$$f_{LO1} = \frac{4}{5} f_{in}.$$

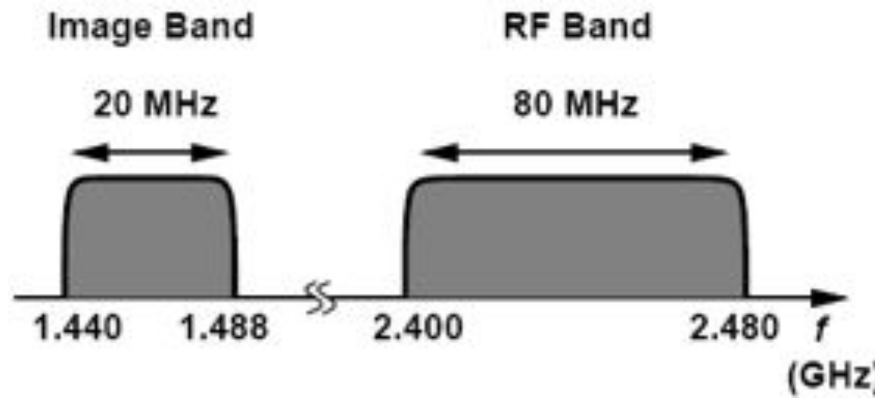
- Second LO $f_{in}/5$, slightly lower, desirable because generation of LO quadrature phases at lower frequencies incurs smaller mismatches
- Reduces the frequency difference between the image and the signal, difficult to reject image.

An Example to Compare the Divide Ratio of 2 and 4

We wish to select a sliding-IF architecture for an 802.11g receiver. Determine the pros and cons of a ÷ 2 or a ÷ 4 circuit in the LO path.



Divide-by-2 circuit

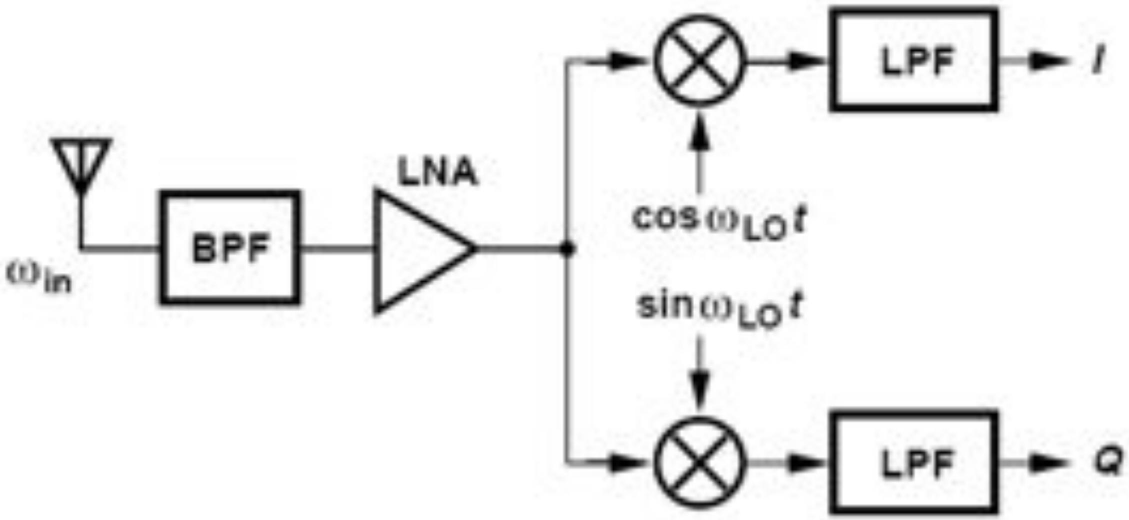


Divide-by-4 circuit

With a ÷2 circuit, the 11g band (2.40-2.48 GHz) requires an LO range of 1.600-1.653 GHz and hence an image range of 800-827 MHz. Unfortunately, since the CDMA transmit band begins at 824 MHz, such a sliding-IF receiver may experience a large image in the range of 824-827 MHz.

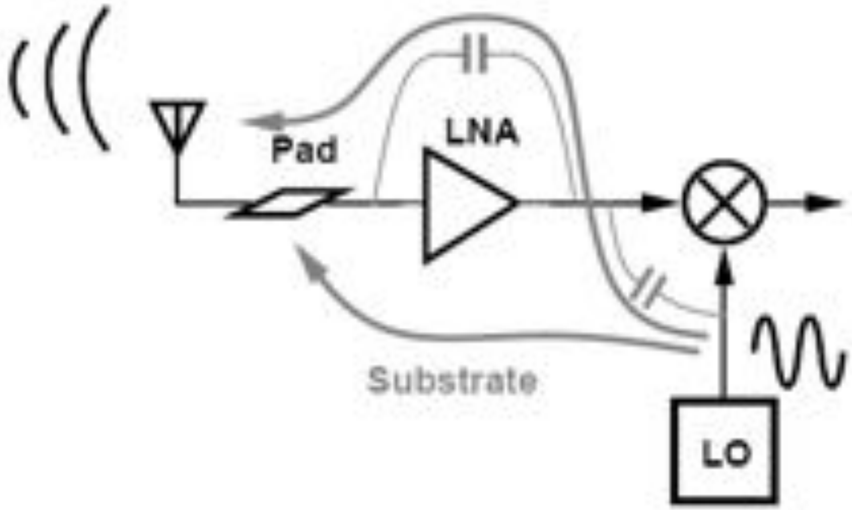
With a ÷4 circuit, the LO range is 1.920-1.984 GHz and the image range, 1.440-1.488 GHz. This image band is relatively quiet. (Only Japan has allocated a band around 1.4 GHz to WCDMA.) Thus, the choice of the ÷4 ratio proves advantageous here if the LNA selectivity can suppress the thermal noise in the image band. The first IF is lower in the second case and may be beneficial in some implementations.

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:

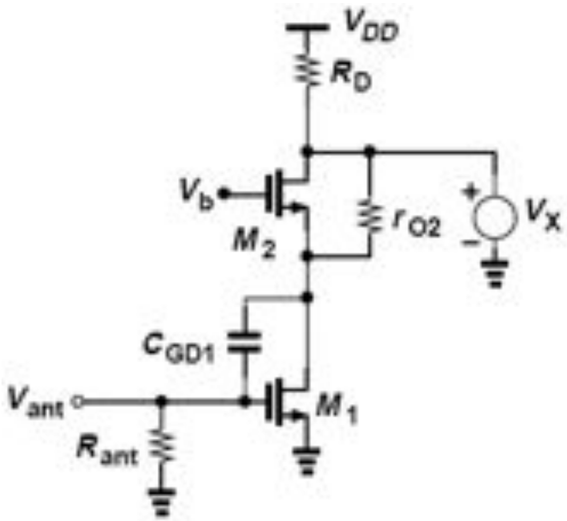
  (a)device capacitances between LO and RF ports of mixer and device capacitances or resistances between the output and input of the LNA

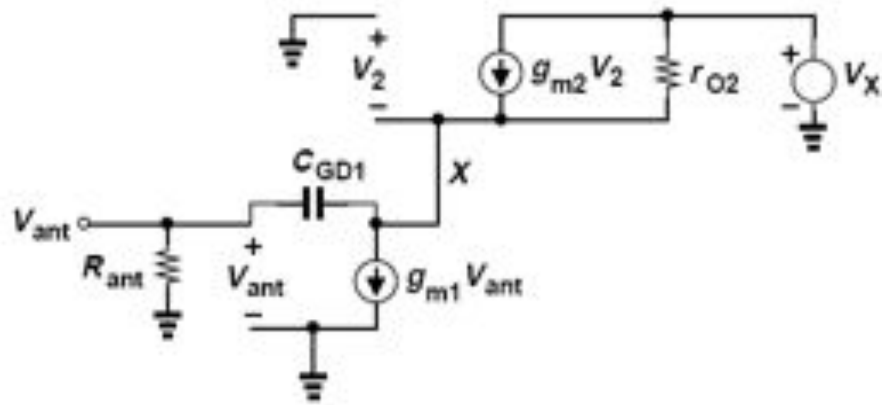
  (b)the substrate to the input pad, especially because the LO employs large on-chip spiral inductors

An Example of LO Leakage

Determine the LO leakage from the output to the input of a cascode LNA.

Solution:





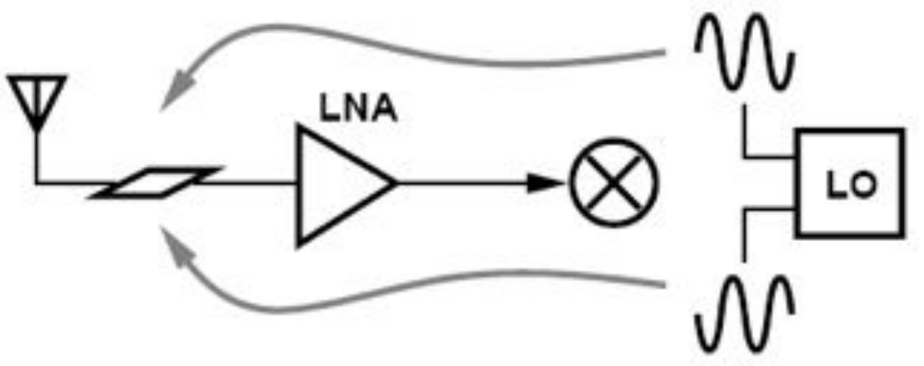
$$V_2 = -[V_{ant} + V_{ant}/(R_{ant}C_{GD1}s)].$$

$$\left(V_{ant} + \frac{V_{ant}}{R_{ant}C_{GD1}s}\right)g_{m2} + \frac{V_{ant}}{R_{ant}} + g_{m1}V_{ant} = \frac{1}{r_{O2}}\left[V_X - \left(V_{ant} + \frac{V_{ant}}{R_{ant}C_{GD1}s}\right)\right]$$

If $g_{m2} >> 1/r_{O2}$

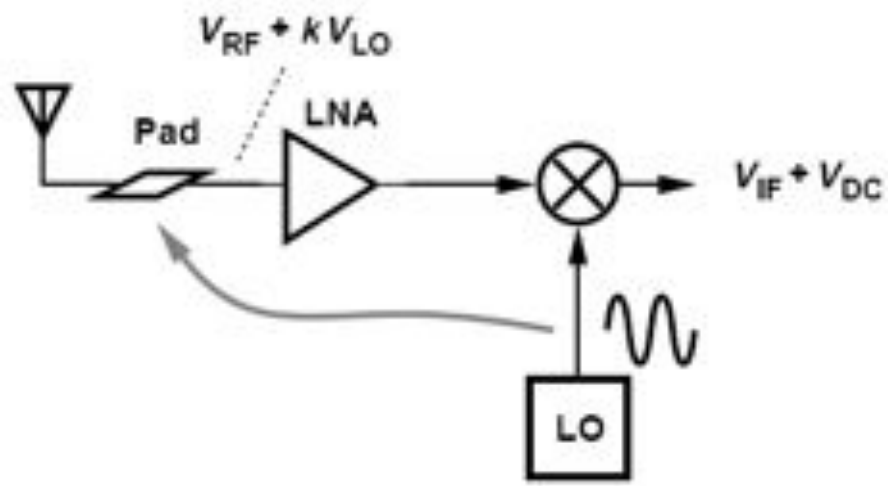
$$\frac{V_{ant}}{V_X} \approx \frac{C_{GD1}s}{(g_{m1}R_{ant} + g_{m2}R_{ant} + 1)C_{GD1}s + g_{m2}} \cdot \frac{R_{ant}}{r_{O2}}.$$

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.

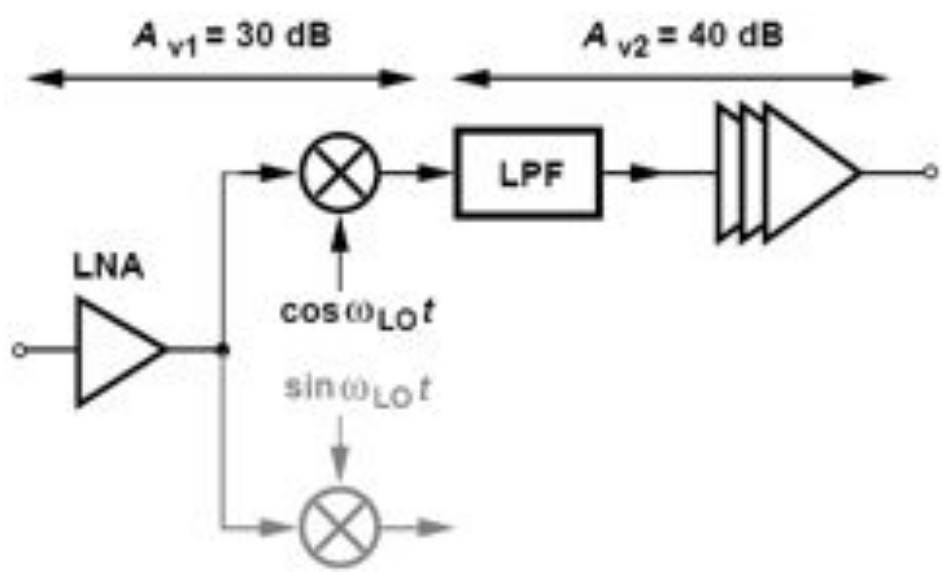
An Example of DC Offsets

A direct-conversion receiver incorporates a voltage gain of 30 dB from the LNA input to each mixer output and another gain of 40 dB in the baseband stages following the mixer. If the LO leakage at the LNA input is equal to -60 dBm, determine the offset voltage at the output of the mixer and at the output of the baseband chain.

Solution:

What does A_{V1} = 30 dB mean? If a sinusoid $V_0 \cos \omega_{in} t$ is applied to the LNA input, then the baseband signal at the mixer output, $V_{bb} \cos(\omega_{in} - \omega_{LO})t$, has an amplitude given by

$$V_{bb} = A_{V1} \cdot V_0.$$



Thus, for an input $V_{leak} \cos \omega_{LO} t$,
the dc value at the mixer output is equal to

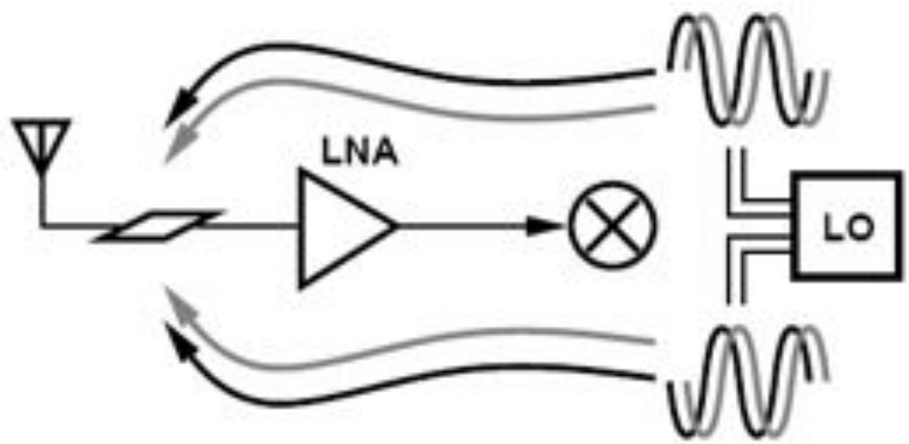
Since A_{V1} = 31.6 and V_{leak} = (632/2) μV,
we have V_{dc} = 10 mV.
Amplified by another 40 dB,
this offset reaches 1 V at the baseband output!

Another Example of DC Offsets

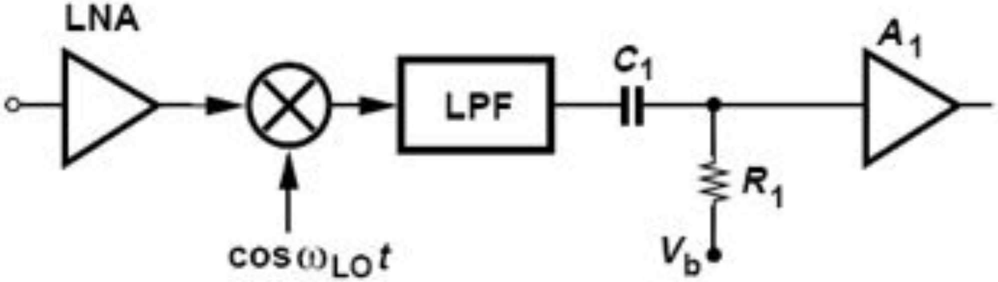
The dc offsets measured in the baseband *I* and *Q* outputs are often unequal. Explain why

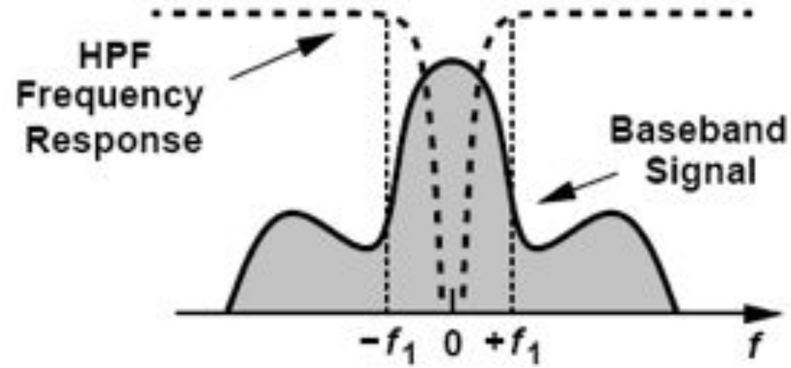
Solution:

Suppose, in the presence of the quadrature phases of the LO, the net LO leakage at the input of the LNA is expressed as $V_{leak} \cos(\omega_{LO}t+\Phi_{leak})$, where Φ_{leak} arises from the phase shift through the path(s) from each LO phase to the LNA input and also the summation of the leakages $V_{LO} \cos \omega_{LO}t$ and $V_{LO} \sin\omega_{LO}t$. The LO leakage travels through the LNA and each mixer, experiencing an additional phase shift, Φ_{ckt}, and is multiplied by $V_{LO} \cos \omega_{LO}t$ and $V_{LO} \sin \omega_{LO}t$. The dc components are therefore given by Thus, the two dc offsets are generally unequal.

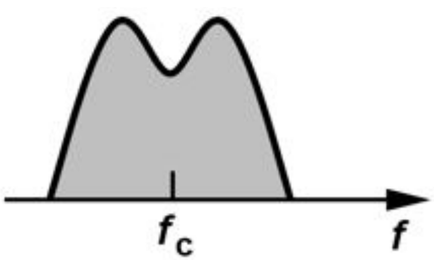


Cancellation of DC Offsets

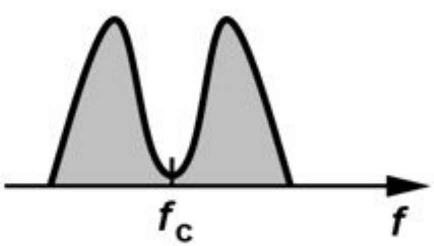




- Offset cancellation: high-pass filter
- Such network also removes a fraction of the signal's spectrum near zero frequency, introducing intersymbol interference



Small modulation index



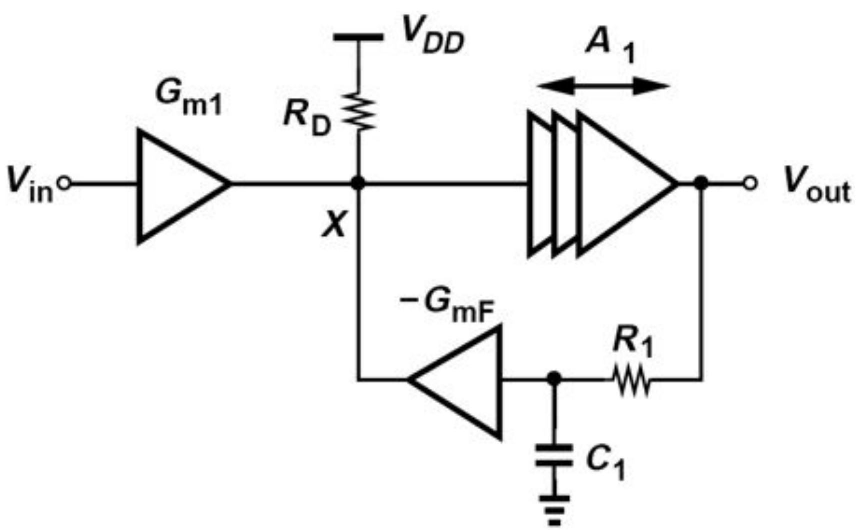
Large modulation index

- Modulation schemes that contain little energy around the carrier better lend themselves to ac coupling in the baseband.
- A drawback of ac coupling stems from its slow response to transient input.

Another Method of Suppressing DC Offsets (Ⅰ)

Figure below shows another method of suppressing dc offsets in the baseband. Here, the main signal path consists of G_{m1} (a transconductance amplifier), R_D, and A_1, providing a total voltage gain of $G_{m1}R_DA_1$. The negative feedback branch comprising R_1, C_1 and $-G_{mF}$ returns a low-frequency current to node *X* so as to drive the dc content of V_{out} toward zero. Note that this topology suppresses the dc offsets of all of the stages in the baseband. Calculate the corner frequency of the circuit.

Solution:



Recognizing that the current returned by $-G_{mF}$ to node *X* is equal to $-G_{mF}V_{out}=(R_1C_1s + 1)$ and the current produced by G_{m1} is given by $G_{m1}V_{in}$, we sum the two at node *X*, multiply the sum by R_D and A_1, and equate the result to V_{out}

$$\left(\frac{-G_{mF}V_{out}}{R_1C_1s+1}+G_{m1}V_{in}\right)R_DA_1=V_{out}$$

It follows:

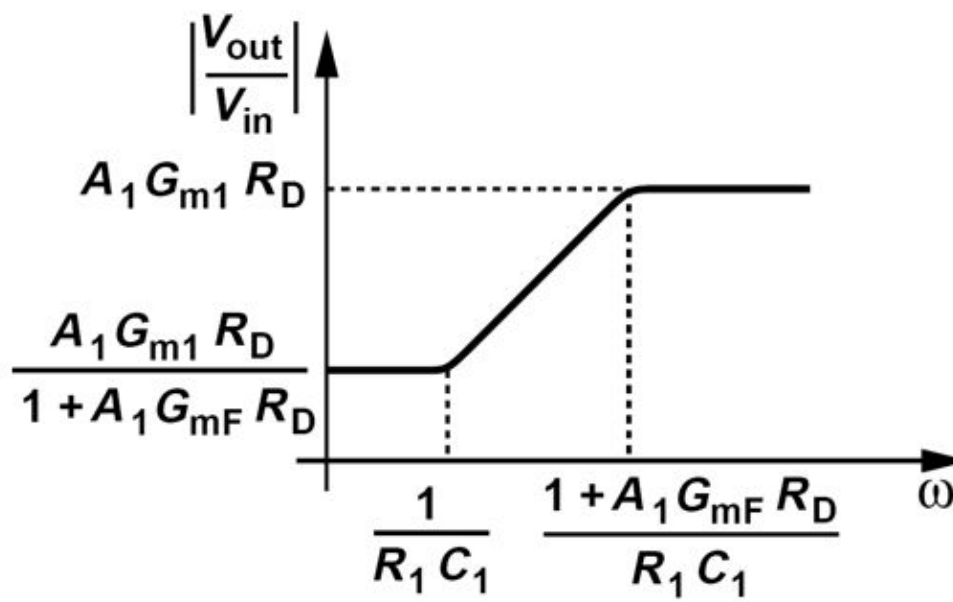
$$\frac{V_{out}}{V_{in}}=\frac{G_{m1}R_DA_1(R_1C_1s+1)}{R_1C_1s+G_{mF}R_DA_1+1}$$

Another Method of Suppressing DC Offsets (Ⅱ)

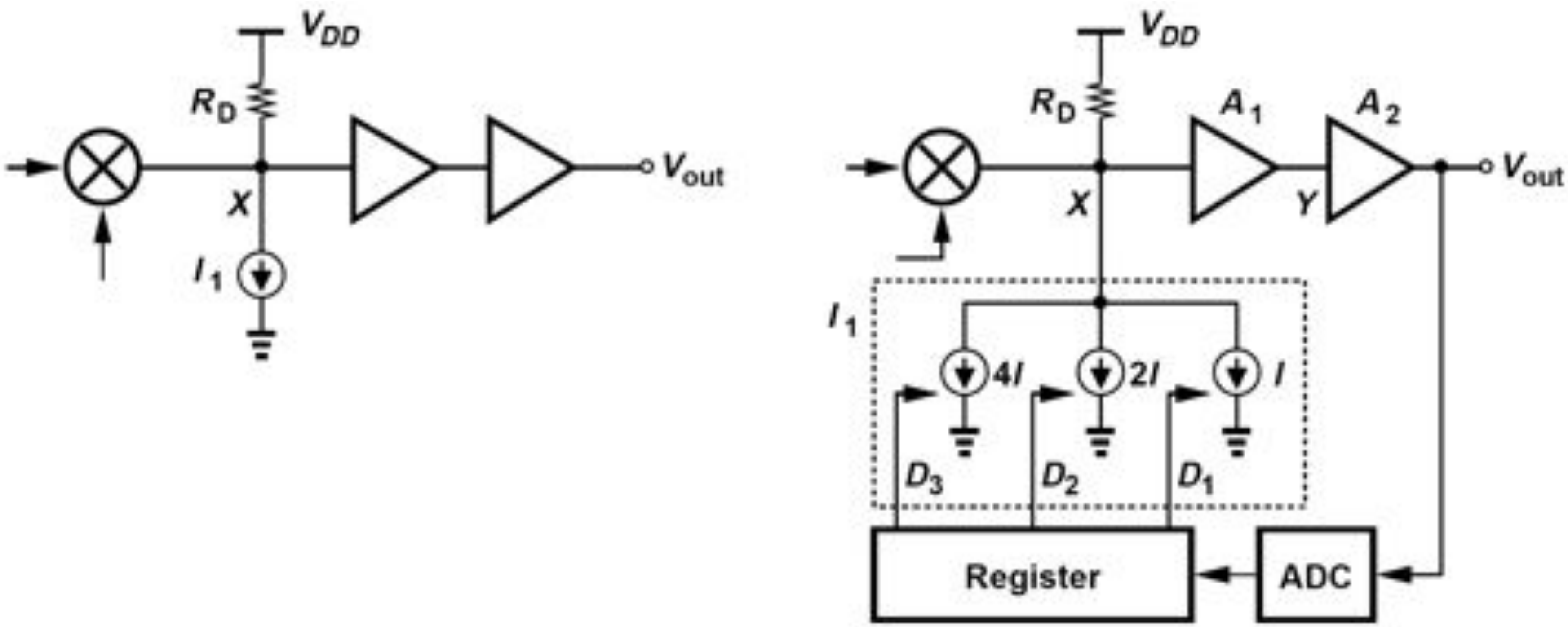
The circuit thus exhibits a pole at $-(1+G_{mF}R_DA_1)/(R_1C_1)$ and a zero at $-1/(R_1C_1)$. The input offset is amplified by a factor of $G_{m1}R_DA_1/(1+G_{mF}R_DA_1) \approx G_{m1}/G_{mF}$ if $G_{mF}R_DA_1 >> 1$. This gain must remain below unity, i.e., G_{mF} is typically chosen larger than G_{m1}. Unfortunately, the high-pass corner frequency is given by

$$f_1 \approx \frac{G_{mF}R_DA_1}{2\pi(R_1C_1)}$$

a factor of $G_{mF}R_DA_1$ higher than that of the passive circuit mentioned before. This "active feedback" circuit therefore requires greater values for R_1 and C_1 to provide a low f_1. The advantage is that C_1 can be realized by a MOSFET.



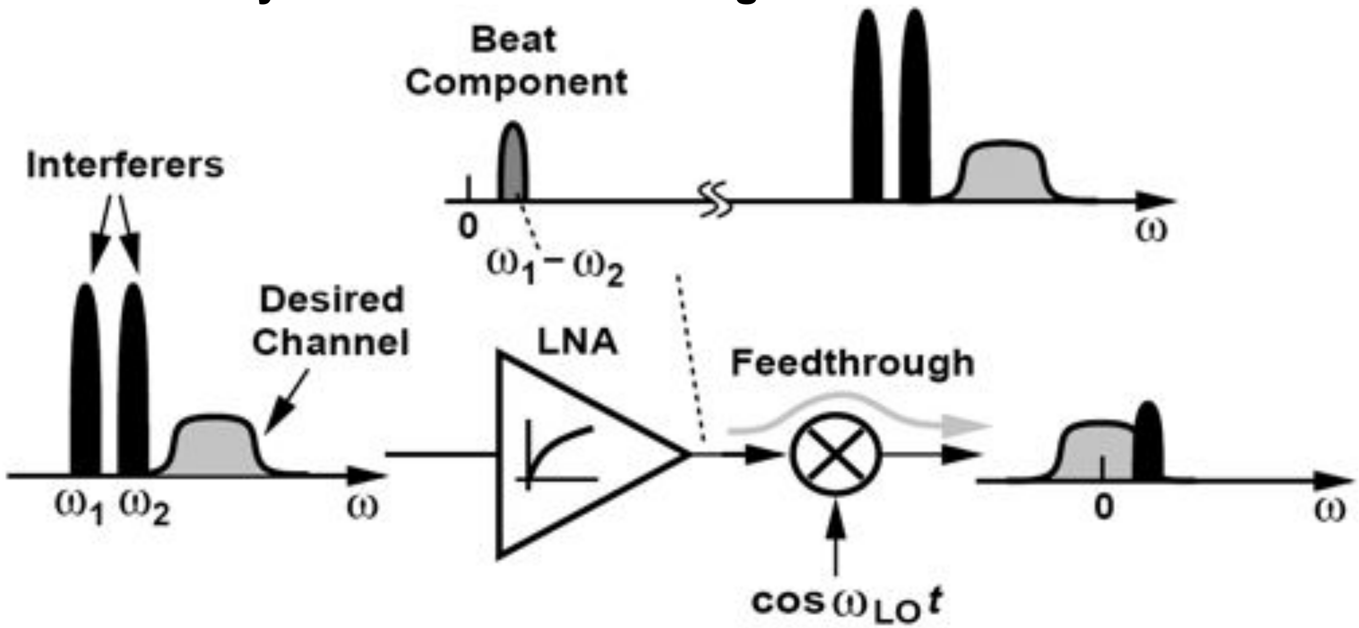
Offset Cancellation Employing DACs to Draw Corrective Current



- The entire negative-feedback loop converges such that V_{out} is minimized. The resulting values are then stored in the register and remain frozen during the actual operation of the receiver.
- The principal drawback of digital storage originates from the finite resolution with which the offset is cancelled. A higher resolution or multiple DACs can be tied to different nodes to alleviate this issue.

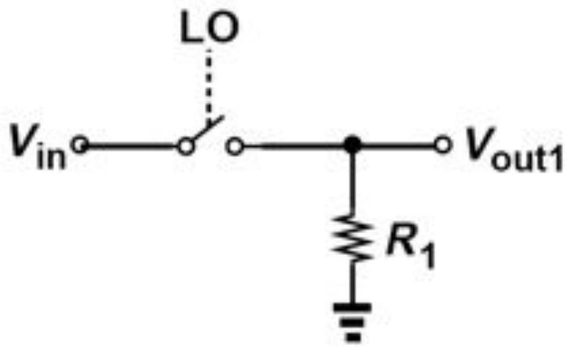
Even-Order Distortion: an Overview

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.

How Asymmetries Give Rise to Direct "Feedthrough"



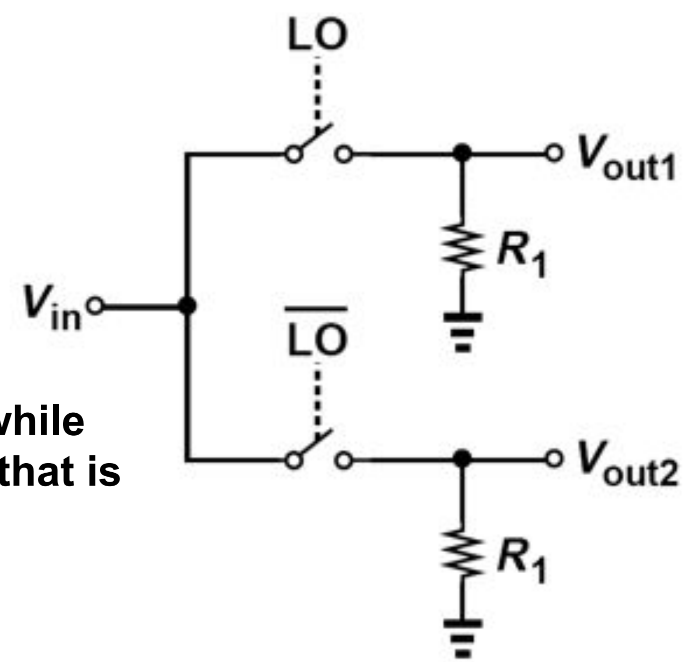
First consider the circuit shown left. The output can be written as the product of V_{in} and an ideal LO, *S(t)*.

$$\begin{aligned} V_{out}(t) &= V_{in}(t) \cdot S(t) \\ &= V_{in}(t)\left[S(t) - \frac{1}{2}\right] + V_{in}(t) \cdot \frac{1}{2}. \end{aligned}$$

Thus, $V_{in}(t) \cdot [S(t) - 1/2]$ contains the product of Vin and the odd harmonics of the LO. The second term, $V_{in}(t) \times 1/2$, denotes the RF feedthrough to the output

Next, consider the circuit shown right. We have

$$\begin{aligned} V_{out1}(t) &= V_{in}(t)S(t) \\ V_{out2}(t) &= V_{in}(t)[1 - S(t)] \end{aligned}$$

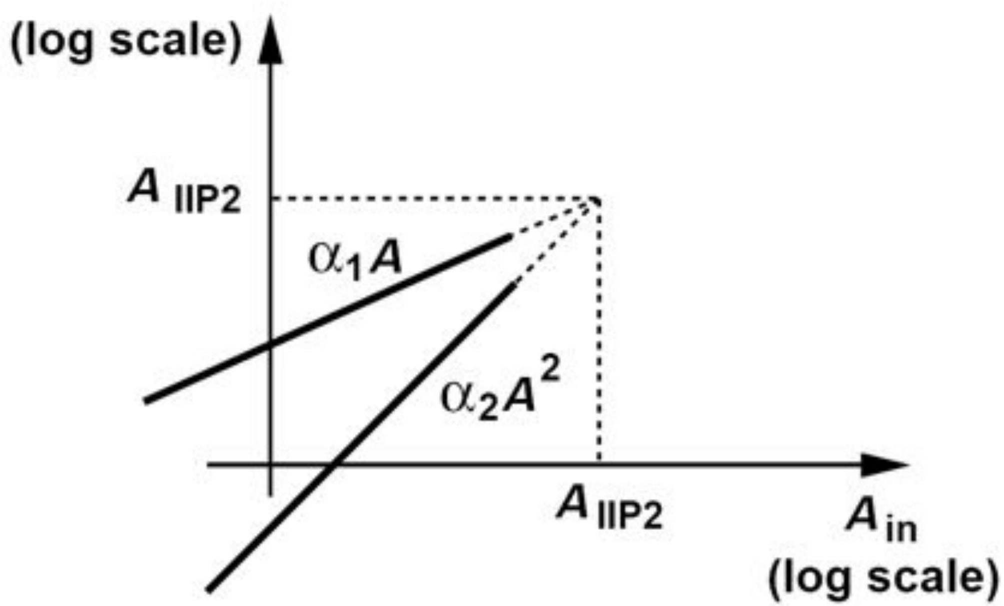


If the output is sensed differentially, the RF feedthroughs in $V_{out1}(t)$ and $V_{out2}(t)$ are cancelled while the signal components add. It is this cancellation that is sensitive to asymmetries.

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

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

Even-Order Distortion in the Absence of Interferers

We express the signal as $x_{in}(t)$ = [A_0+$a(t)$] cos[$\omega_c t$+$\Phi(t)$], where $a(t)$ denotes the envelope and typically varies slowly, i.e., it is a low-pass signal.

$$\alpha_2 x_{in}^2(t) = \alpha_2[A_0^2 + 2A_0 a(t) + a^2(t)]\frac{1+\cos[2\omega_c t + 2\phi(t)]}{2}$$

Both of the terms $\alpha_2 A_0 a(t)$ and $\alpha_2 a^2(t)/2$ are low-pass signals and, like the beat component, pass through the mixer with finite attenuation, corrupting the downconverted signal.

Quantify the self-corruption expressed by equation above in terms of the IP_2.

Assume, that the low-pass components, $\alpha_2 A_0 a(t)$+$\alpha_2 a^2(t)$=2, experience an attenuation factor of k and the desired signal, $\alpha_1 A_0$, sees a gain of unity. Also, typically $a(t)$ is several times smaller than A_0 and hence the baseband corruption can be approximated as $k\alpha_2 A_0 a(t)$. Thus, the signal-to-noise ratio arising from self-corruption is given by

$$\mathrm{SNR} = \frac{\alpha_1 A_0/\sqrt{2}}{k\alpha_2 A_0 a_{rms}} = \frac{A_{IIP2}}{\sqrt{2}a_{rms}},$$

Even-Order Distortion with a Large AM Interferer

Even-order distortion demodulates the AM component of the interferer, and mixer feedthrough allows it to appear in the baseband.

$$\begin{aligned}\text{SNR} &= \frac{\alpha_1 A_{sig}/\sqrt{2}}{k\alpha_2 A_{int} a_{rms}} \\ &= \frac{A_{IIP2} A_{sig}/\sqrt{2}}{A_{int} a_{rms}}.\end{aligned}$$

A desired signal at -100 dBm is received along with an interferer [A_{int} + $a(t)$] cos[$\omega_c t$ + $\Phi(t)$], where A_{int} = 5 mV and a_{rms} = 1 mV. What IP_2 is required to ensure SNR ≥ 20 dB?

$$\begin{aligned}A_{IIP2} &= \text{SNR}\frac{A_{int} a_{rms}}{A_{sig}/\sqrt{2}} \\ &= 22.4\ \text{V} \\ &= +37\ \text{dBm}.\end{aligned}$$

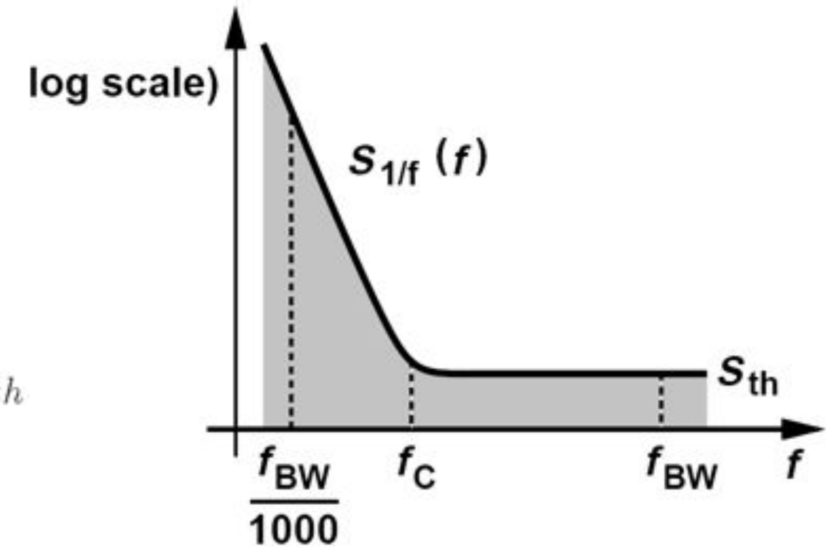
Flicker Noise

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

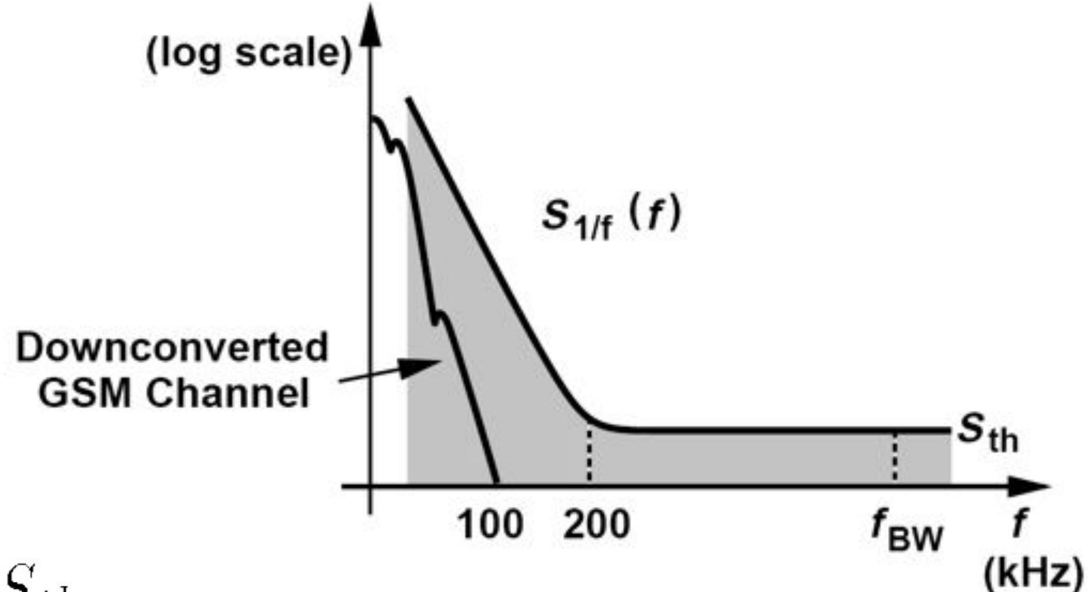
$$\frac{P_{n1}}{P_{n2}} = 1.04$$

Example of GSM Receiver Flicker Noise Penalty

A GSM receiver exhibits a baseband flicker noise corner frequency of 200 kHz. Determine the flicker noise penalty.

Figure below plots the baseband spectra, implying that the noise must be integrated up to 100 kHz. Assuming a lower end equal to about 1/1000 of the bit rate, we write the total noise as

$$P_{n1} = \int_{27\text{ Hz}}^{100\text{ kHz}} \frac{\alpha}{f} df = f_c \cdot S_{th} \ln \frac{100\text{ kHz}}{27\text{ Hz}} = 8.2 f_c S_{th}.$$



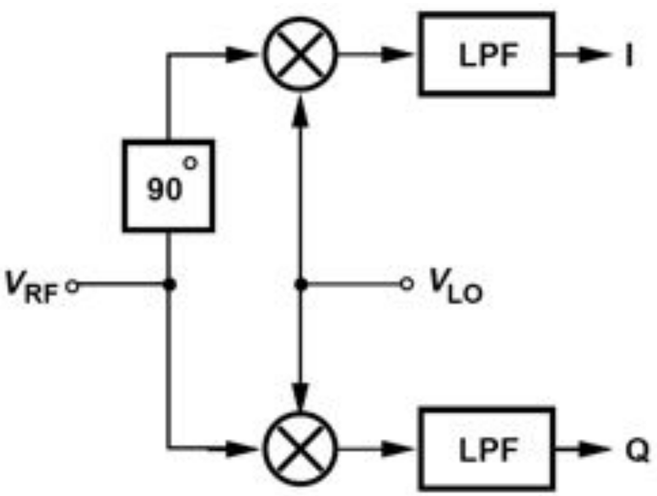
Without flicker noise,

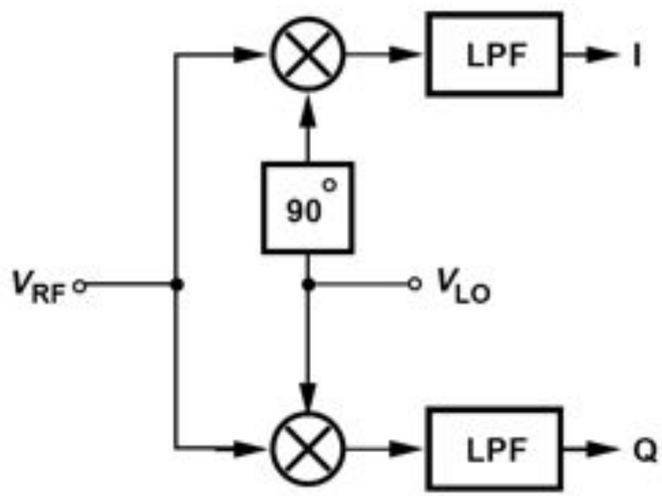
$$P_{n2} \approx (100\text{ kHz}) S_{th}$$

That is, the penalty reaches

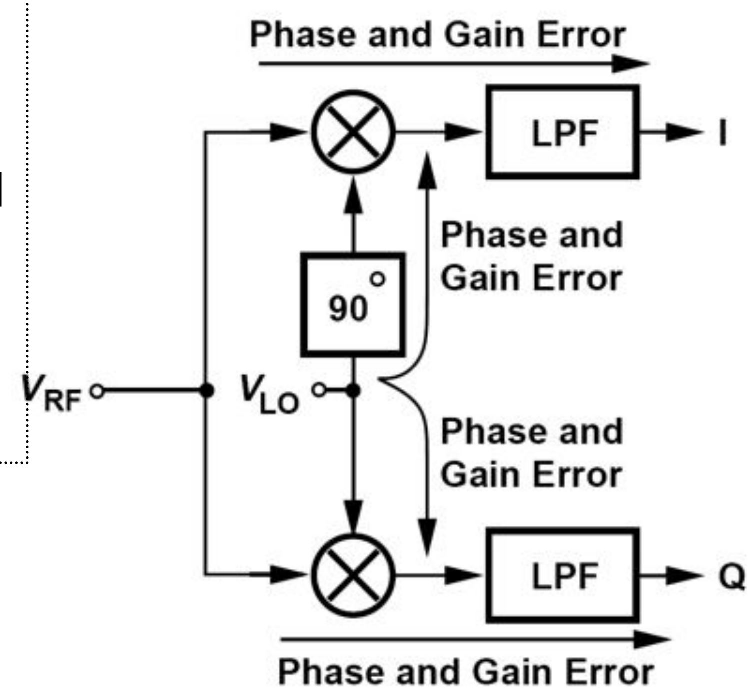
$$\frac{P_{n1}}{P_{n2}} = \frac{8.2 f_c}{100\text{ kHz}} = 16.4.$$

I/Q Mismatch: Sources

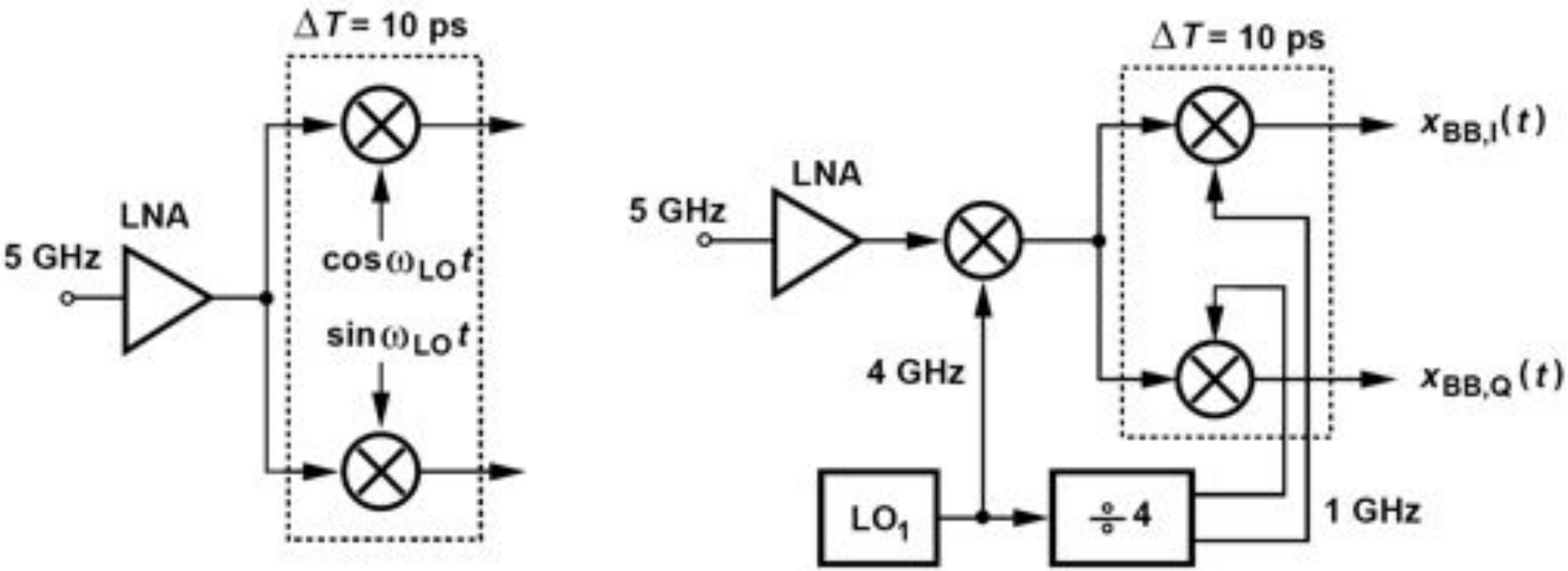




- Separation into quadrature phases can be accomplished by shifting either the RF signal or the LO waveform by 90°.
- Errors in the 90° phase shift circuit and mismatches between the quadrature mixers result in imbalances in the amplitudes and phases of the baseband I and Q outputs.



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



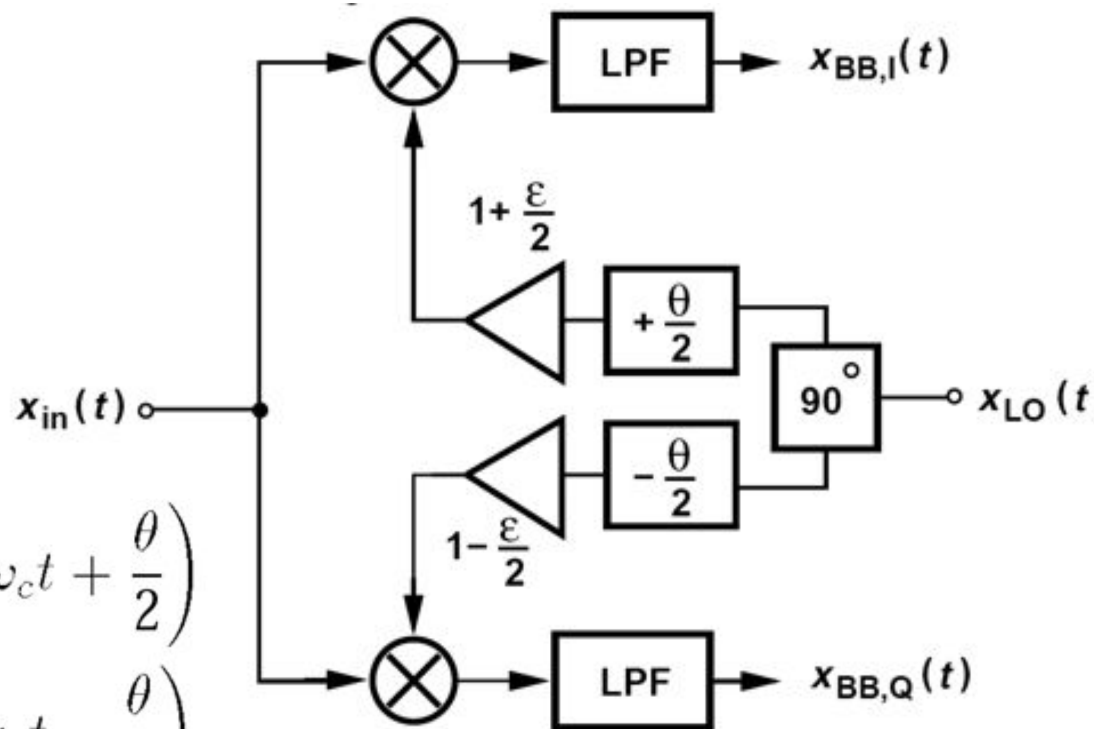
- Quadrature mismatches tend to be larger in direct-conversion receivers than in heterodyne topologies.
- This occurs because

  (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:



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

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

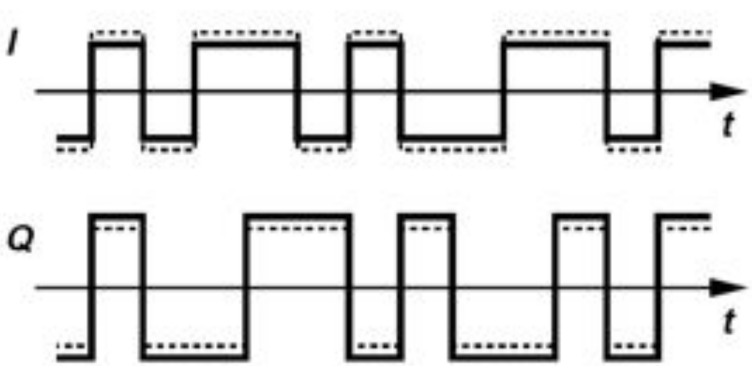
$$x_{BB,I}(t) = a\left(1+\frac{\epsilon}{2}\right)\cos\frac{\theta}{2} - b\left(1+\frac{\epsilon}{2}\right)\sin\frac{\theta}{2}$$

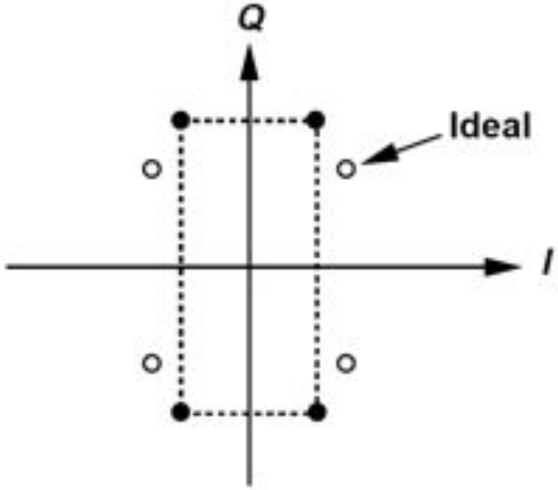
$$x_{BB,Q}(t) = -a\left(1-\frac{\epsilon}{2}\right)\sin\frac{\theta}{2} + b\left(1-\frac{\epsilon}{2}\right)\cos\frac{\theta}{2}$$

Effect of I/Q Mismatch (Ⅱ)

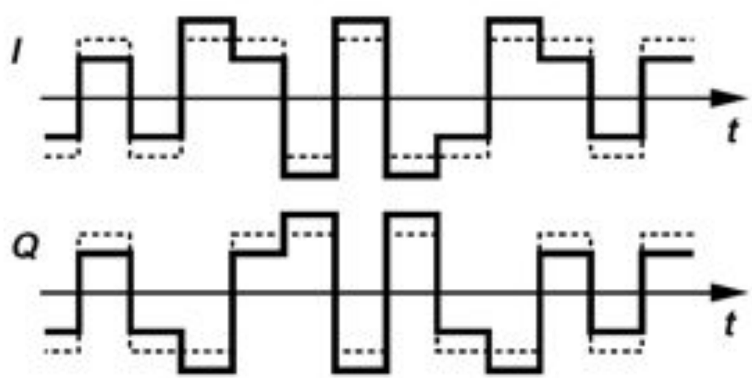
We now examine the results for two special cases:

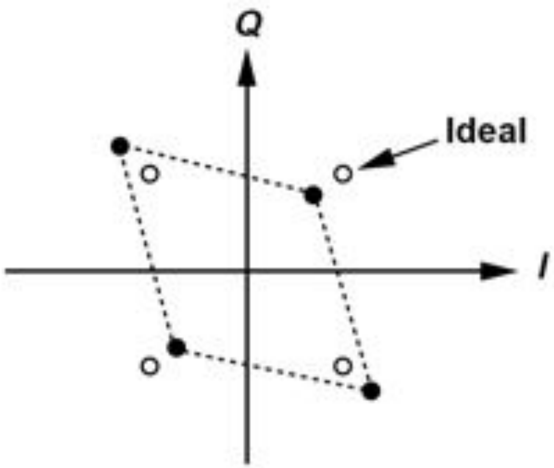
(1) ε ≠ 0, θ = 0 : the quadrature baseband symbols are scaled differently in amplitude,





(2) ε = 0, θ ≠ 0 : each baseband output is corrupted by a fraction of the data symbols in the other output





Example of I/Q Mismatch (Ⅰ)

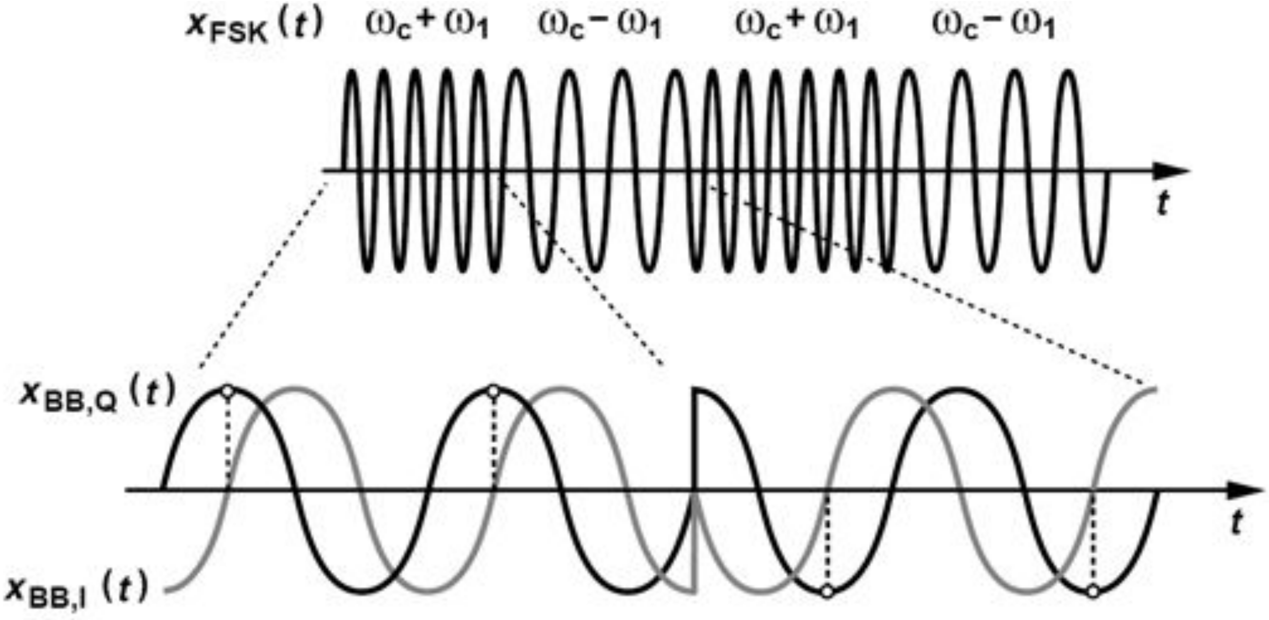
An FSK signal is applied to a direct-conversion receiver. Plot the baseband waveforms and determine the effect of I/Q mismatch.

We express the FSK signal as $x_{FSK}(t) = A_0 \cos[(\omega_c + a\omega_1)t]$, where $a = \pm 1$ represents the binary information; i.e., the frequency of the carrier swings by $+\omega_1$ or $-\omega_1$. Upon multiplication by the quadrature phases of the LO, the signal produces the following baseband components:

$$x_{BB,I}(t) = -A_1 \cos a\omega_1 t$$

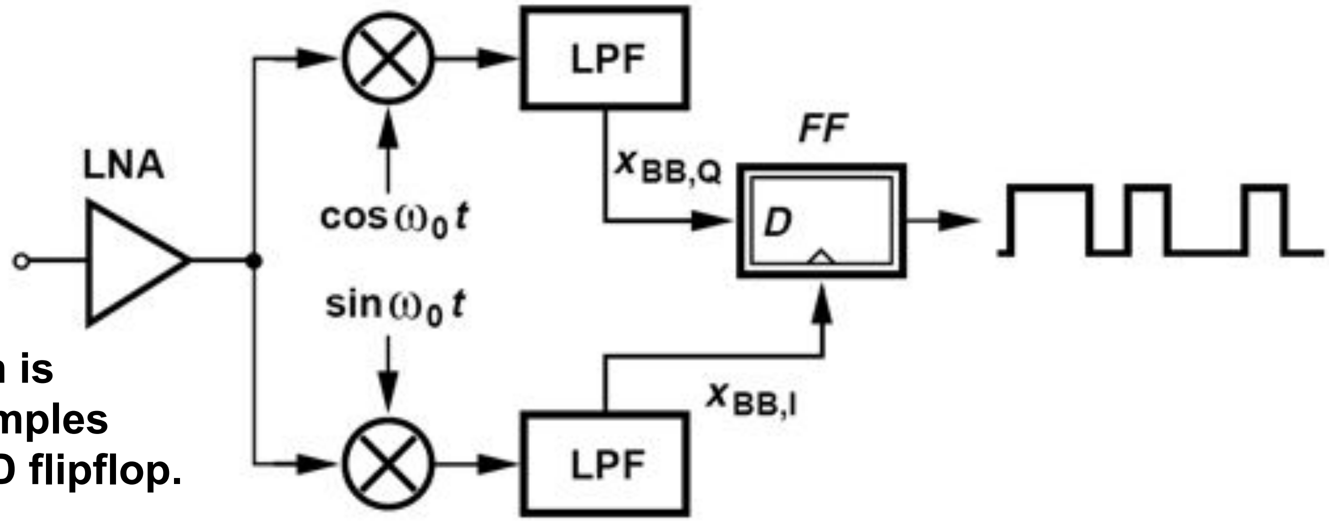
$$x_{BB,Q}(t) = +A_1 \sin a\omega_1 t.$$

Figure on the right illustrates the results: if the carrier frequency is equal to $\omega_c + \omega_1$ (i.e., $a = +1$), then the rising edges of $x_{BB,I}(t)$ coincide with the positive peaks of $x_{BB,Q}(t)$. Conversely, if the carrier frequency is equal to $\omega_c - \omega_1$, then the rising edges of $x_{BB,I}(t)$ coincide with the negative peaks of $x_{BB,Q}(t)$.



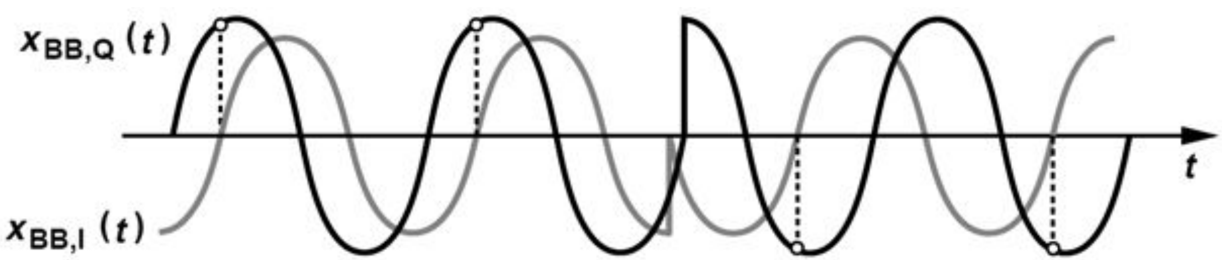
Example of I/Q Mismatch (Ⅱ)

An FSK signal is applied to a direct-conversion receiver. Plot the baseband waveforms and determine the effect of I/Q mismatch.



Thus, the binary information is detected if $x_{BB,I}(t)$ simply samples $x_{BB,Q}(t)$, e.g., by means of a D flipflop.

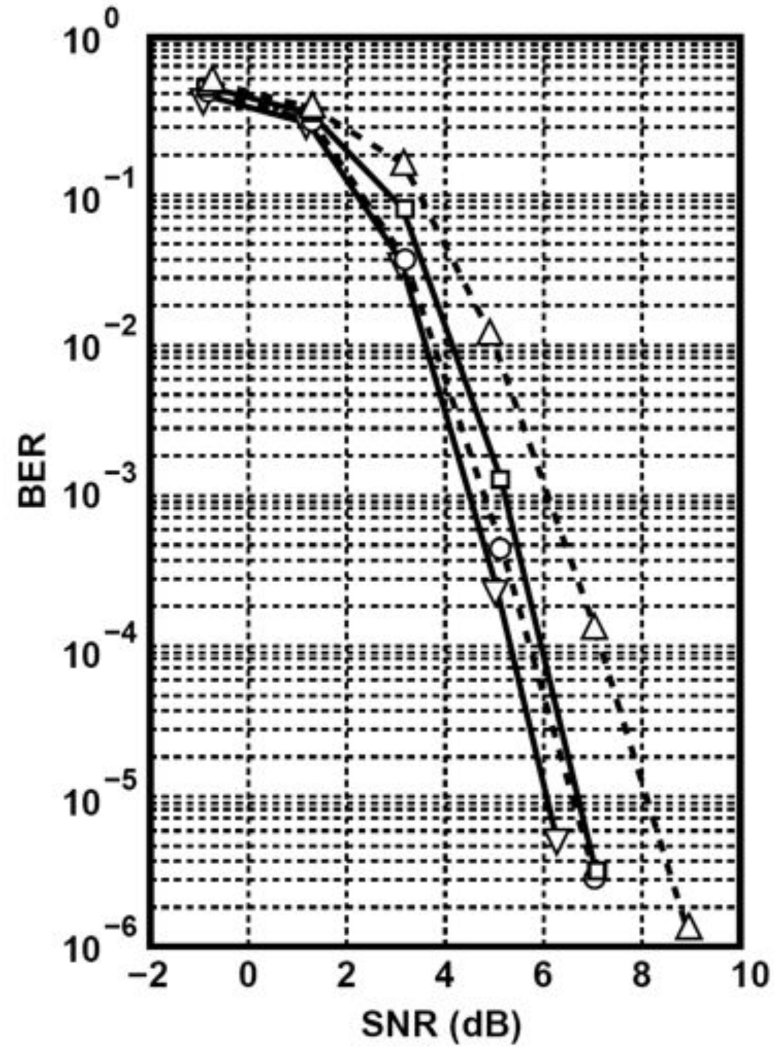
The waveforms above suggest that FSK can tolerate large I/Q mismatches: amplitude mismatch proves benign so long as the smaller output does not suffer from degraded SNR, and phase mismatch is tolerable so long as $x_{BB,I}(t)$ samples the correct polarity of $x_{BB,Q}(t)$. Of course, as the phase mismatch approaches 90_, the additive noise in the receive chain introduces errors.



Requirement of I/Q Mismatch

For complex signal waveforms such as OFDM with QAM, the maximum tolerable I/Q mismatch can be obtained by simulations

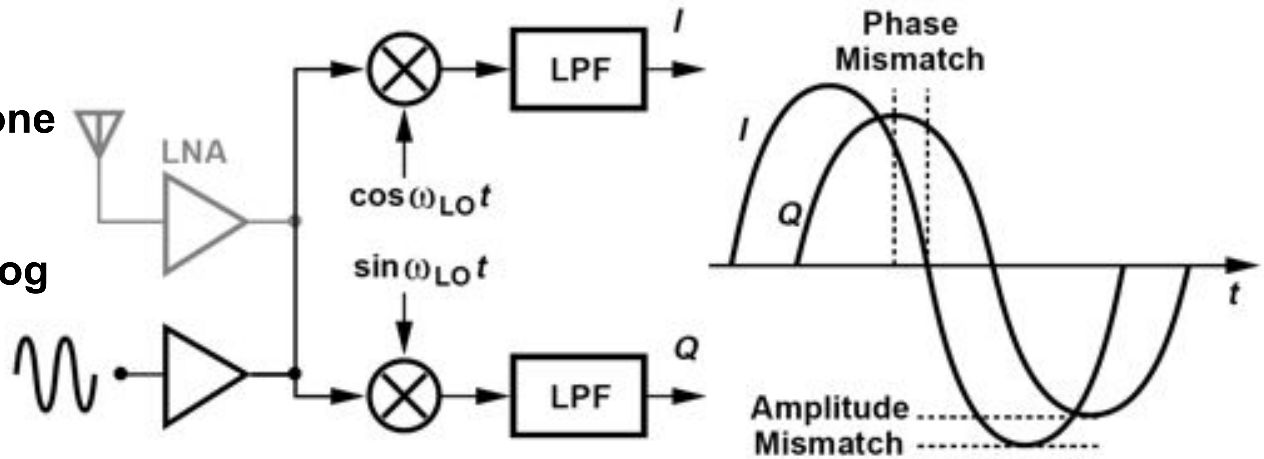
- The bit error rate is plotted for different combinations of gain and phase mismatches, providing the maximum mismatch values that affect the performance negligibly.
- For example, in a system employing OFDM with 128 subchannels and QPSK modulation in each subchannel shown on right, we observe that gain/ phase mismatches below -0.6 dB/6° have negligible effect.



Computation and Correction I/Q Mismatch

In many high performance systems, the quadrature phase and gain must be calibrated—either at power-up or continuously.

Calibration at power-up can be performed by applying an RF tone at the input of the quadrature mixers and observing the baseband sinusoids in the analog or digital domain



With the mismatches known, the received signal constellation is corrected before detection.

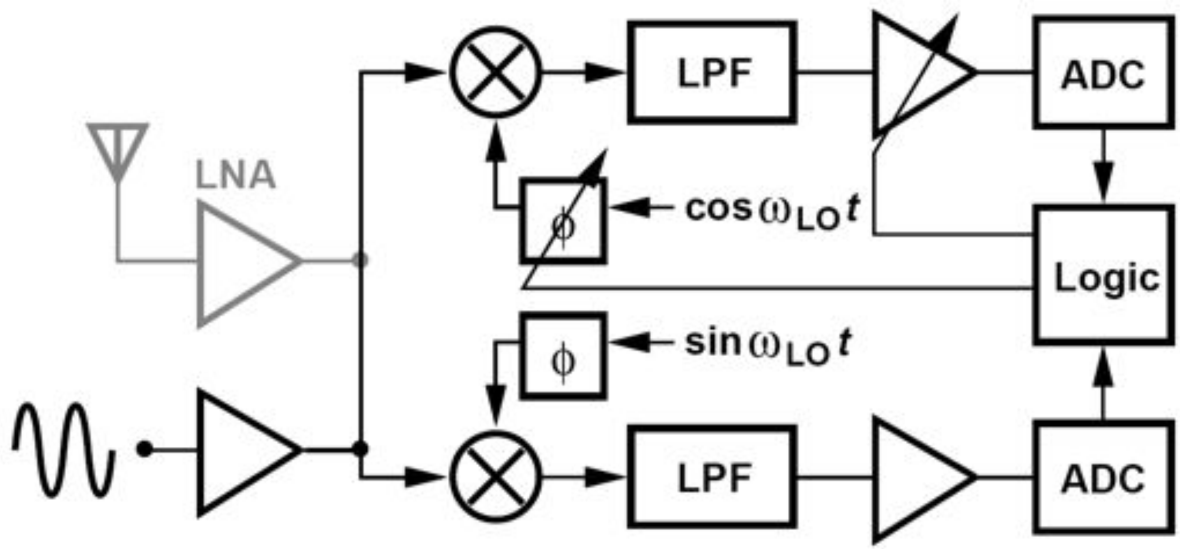
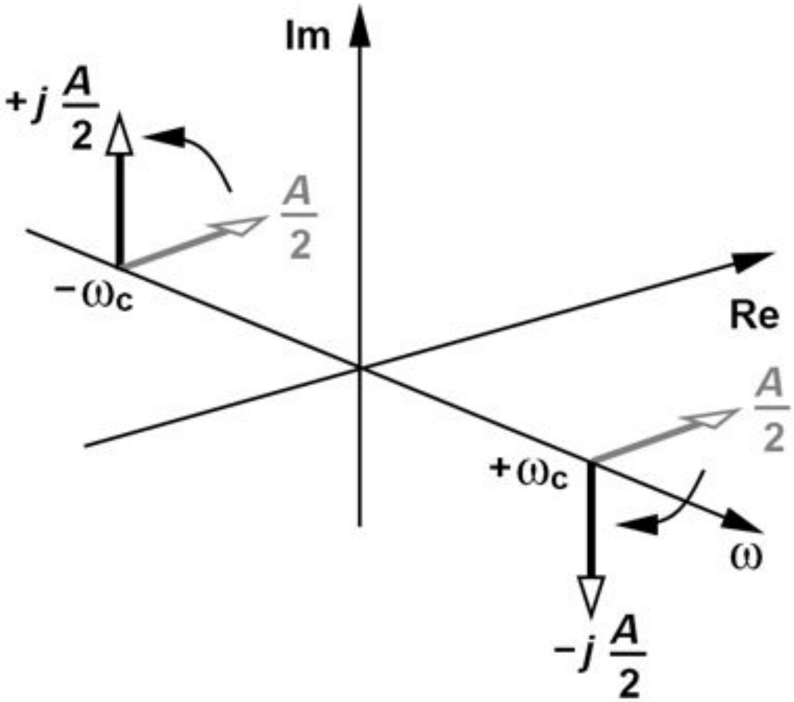


Image-Reject Receivers: 90 °Phase Shift—Cosine Signal

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

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



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

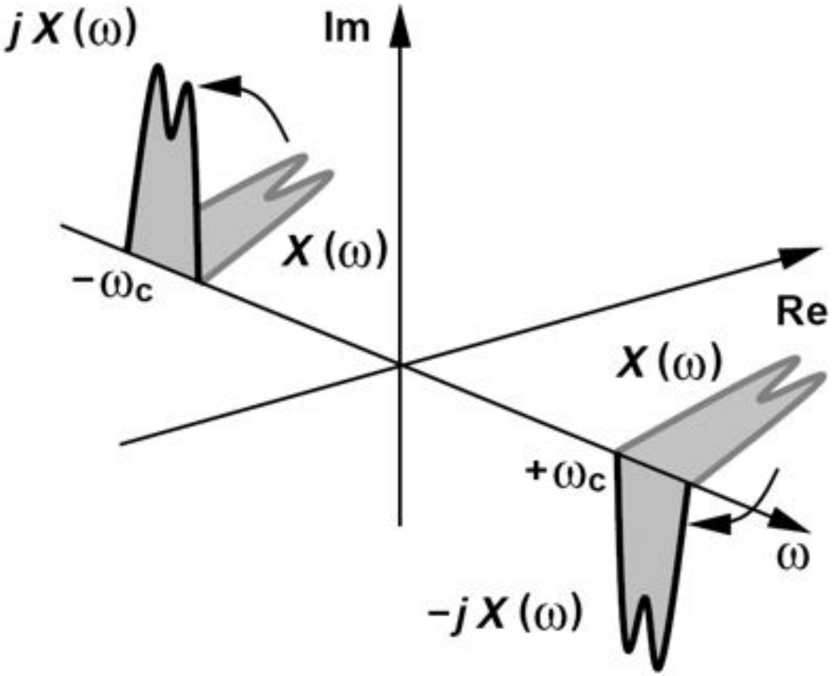
Image-Reject Receivers: 90 °Phase Shift—Modulated Signal

Similarly, for a narrowband modulated signal:

$$A(t)\cos[\omega_c t + \phi(t) - 90°] = A(t)\frac{e^{+j[\omega_c t+\phi(t)-90°]} + e^{-j[\omega_c t+\phi(t)-90°]}}{2}$$

$$= A(t)\frac{-je^{+j[\omega_c t+\phi(t)]} + je^{-j[\omega_c t+\phi(t)]}}{2}$$

$$= A(t)\sin[\omega_c t + \phi(t)].$$



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

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

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

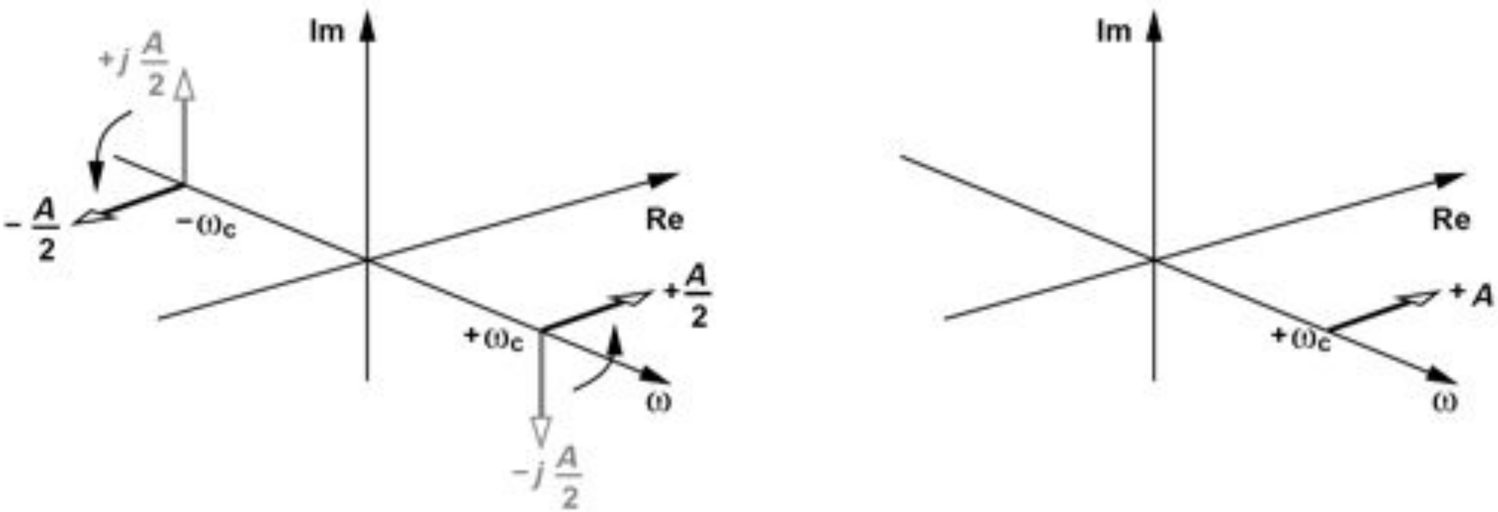
Examples of Hilbert Transform

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 $A\cos\omega_c t + jA\sin\omega_c t$.

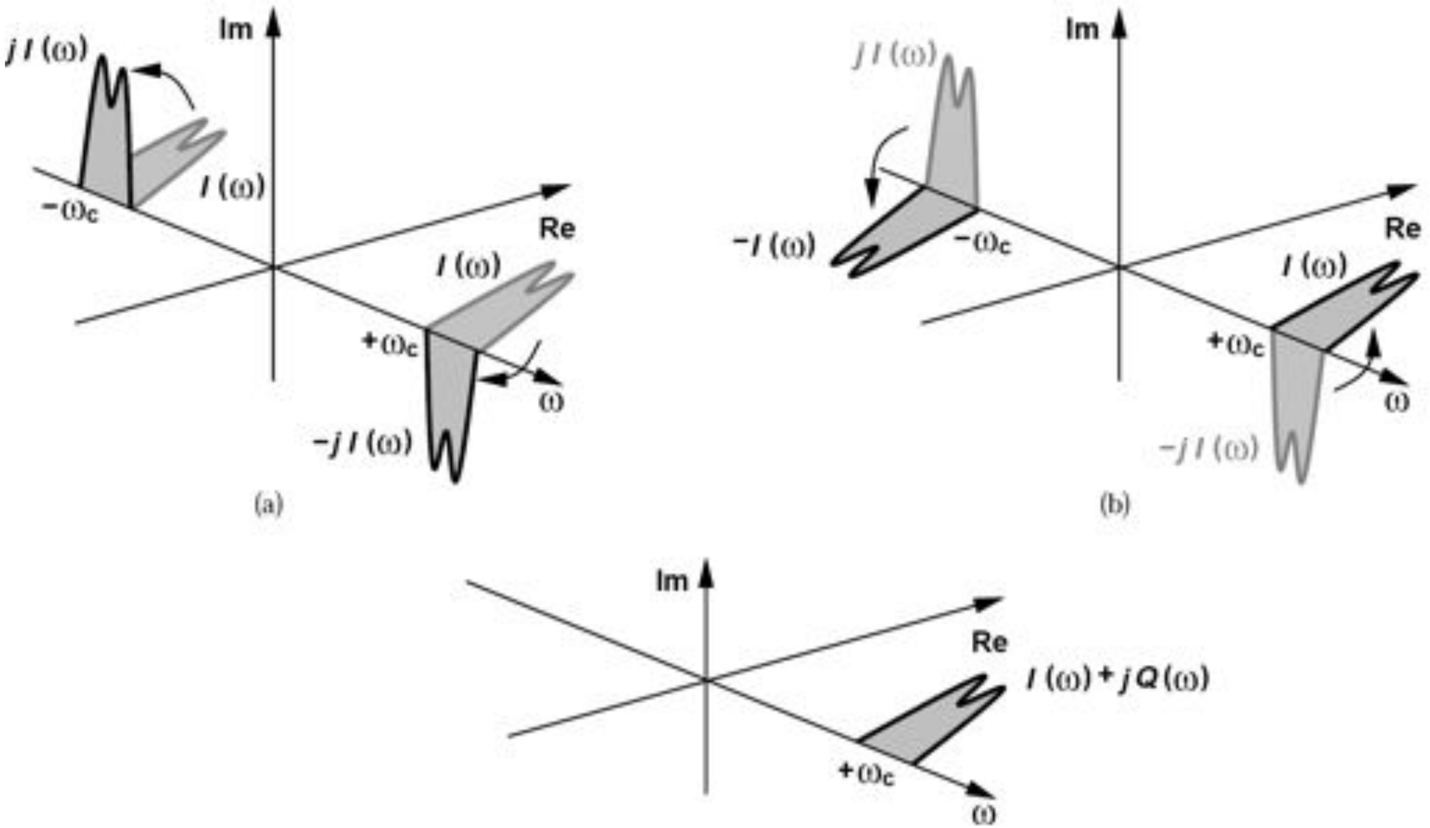
Multiplication of the spectrum of $A\sin\omega_c t$ by j rotates both impulses by 90 ° counterclockwise. Upon adding this spectrum to that $A\cos\omega_c t$, we obtain the one-sided spectrum shown below (right). This is, of course, to be expected because $A\cos\omega_c t + jA\sin\omega_c t = A\exp(-j\omega_c t)$, whose Fourier transform is a single impulse located at $\omega = +\omega_c$.



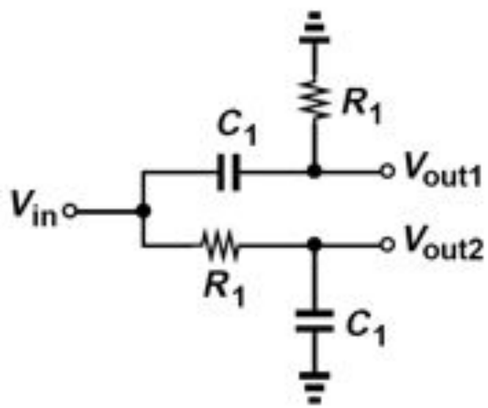
Example of *I* + *jQ*

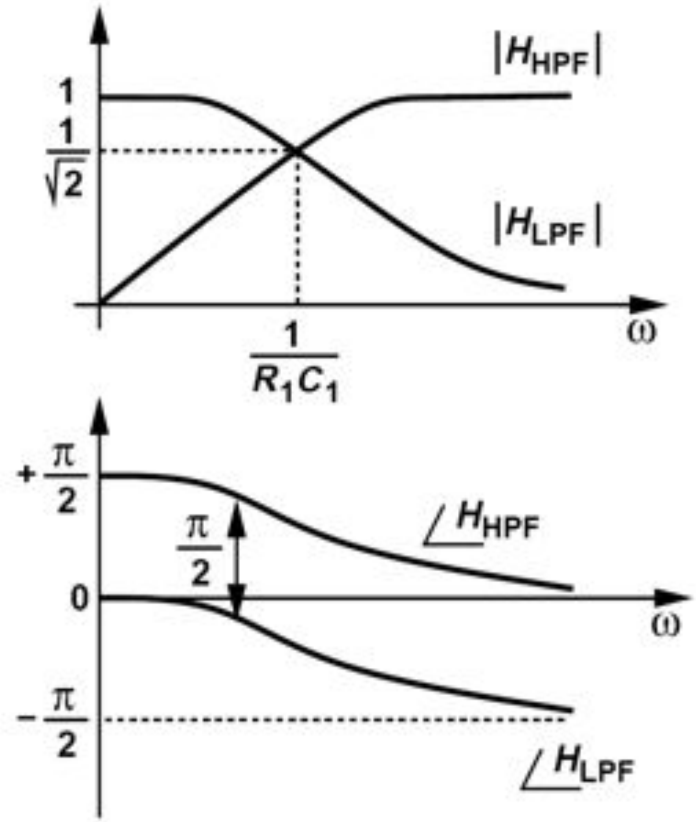
A narrowband signal *I(t)* with a real spectrum is shifted by 90 ° to produce *Q(t)*. Plot the spectrum of *I(t)* + *jQ(t)*.

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.



Implementation of the 90° Phase Shift





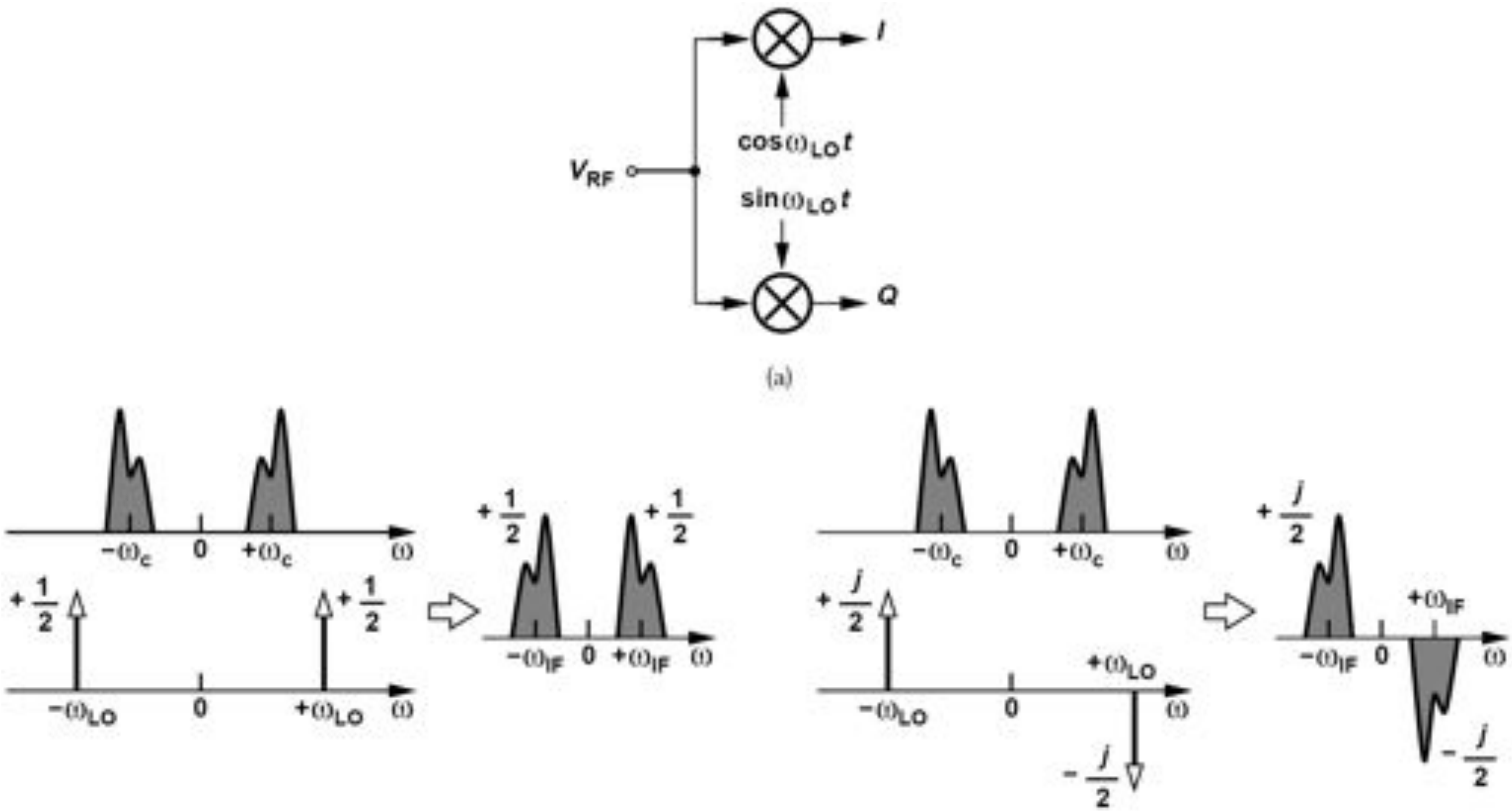
The high-pass and low-pass transfer functions are respectively given by:

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

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

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

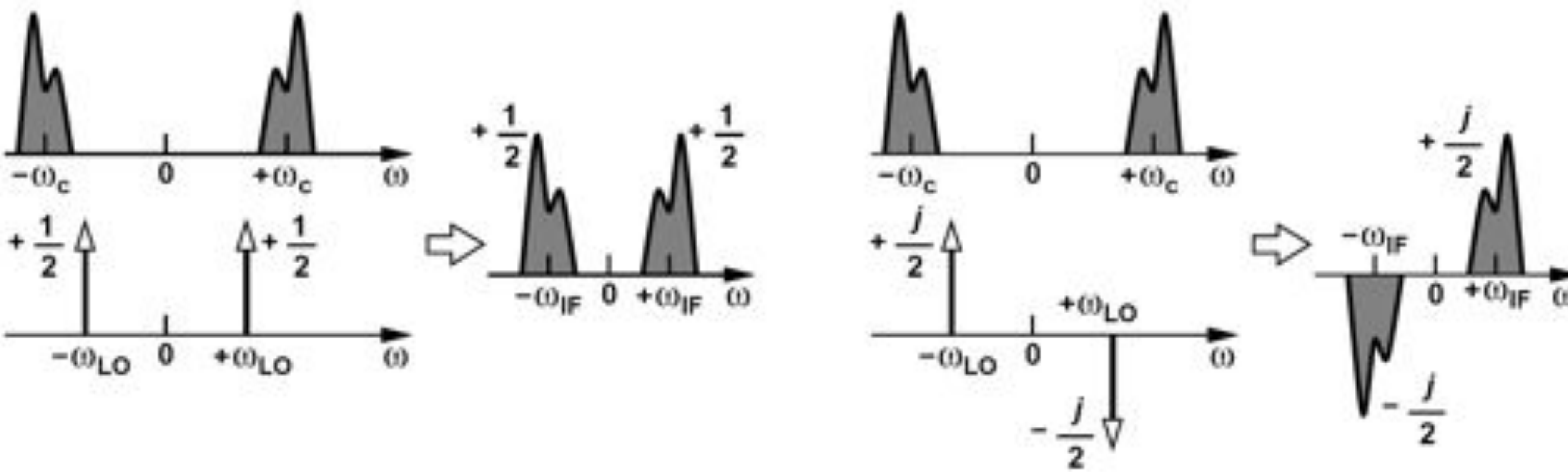


- **The RF input is mixed with the quadrature phases of the LO so as to translate the spectrum to a non-zero IF.**
- **The IF spectrum emerging from the lower arm is the Hilbert transform of that from the upper arm.**

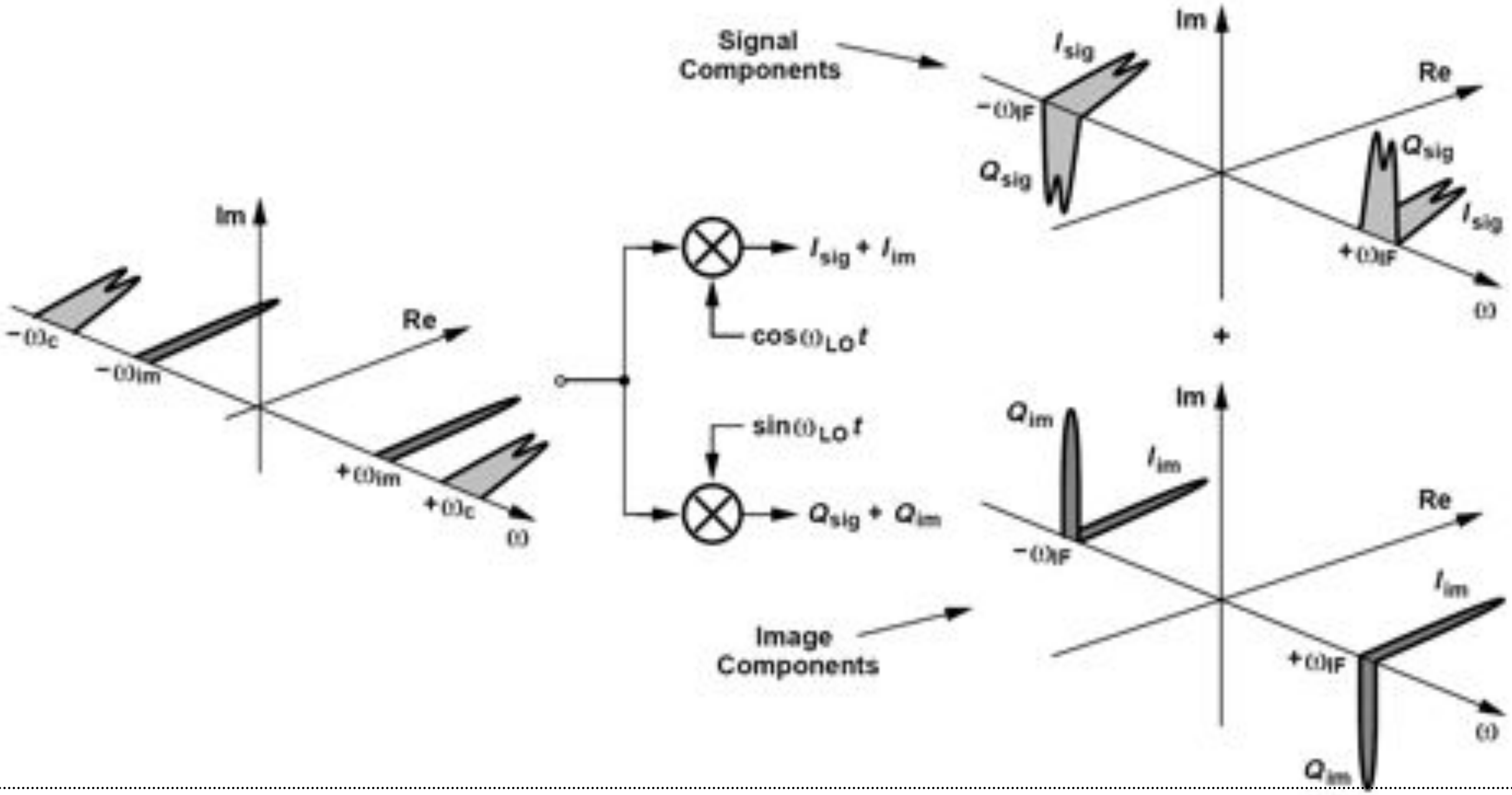
Low-Side Injection of the Above Implementation

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.



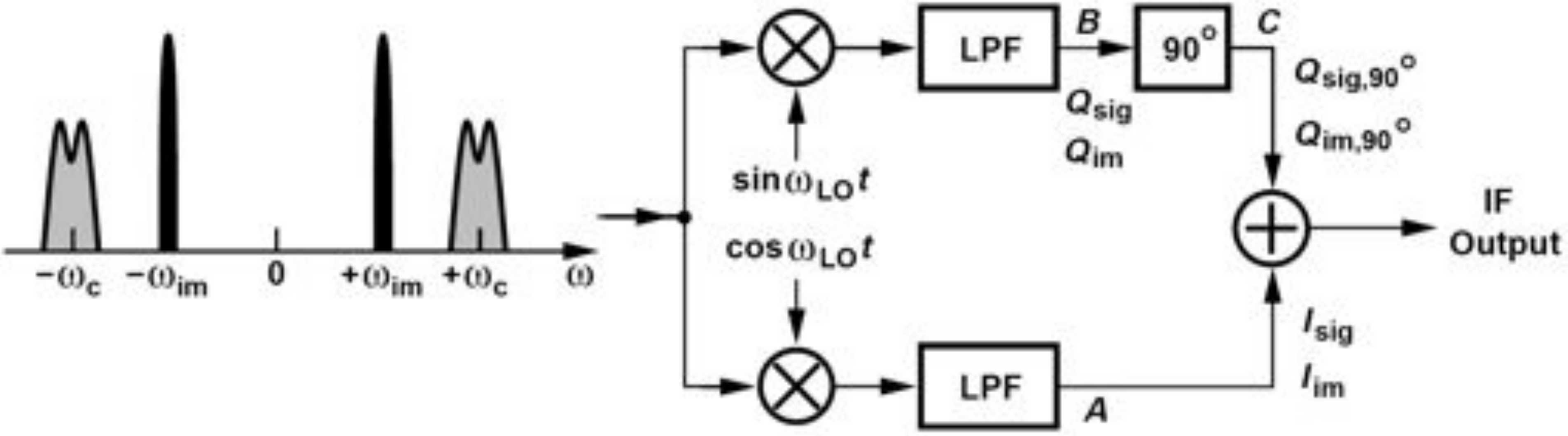
How Can the Image Components Cancel Each Other?



- 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

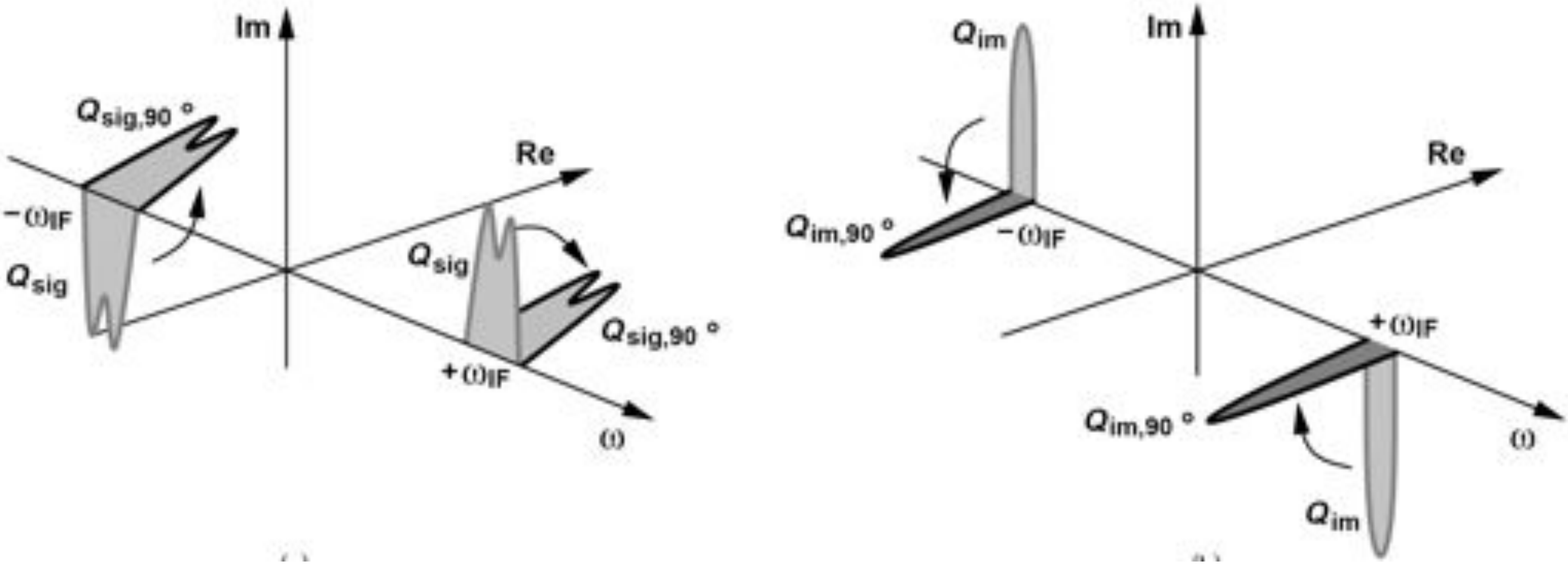
If we shift *I(t)* or *Q(t)* by another 90 ° before adding them, the image may be removed.



- 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:**

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

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

It follows that:

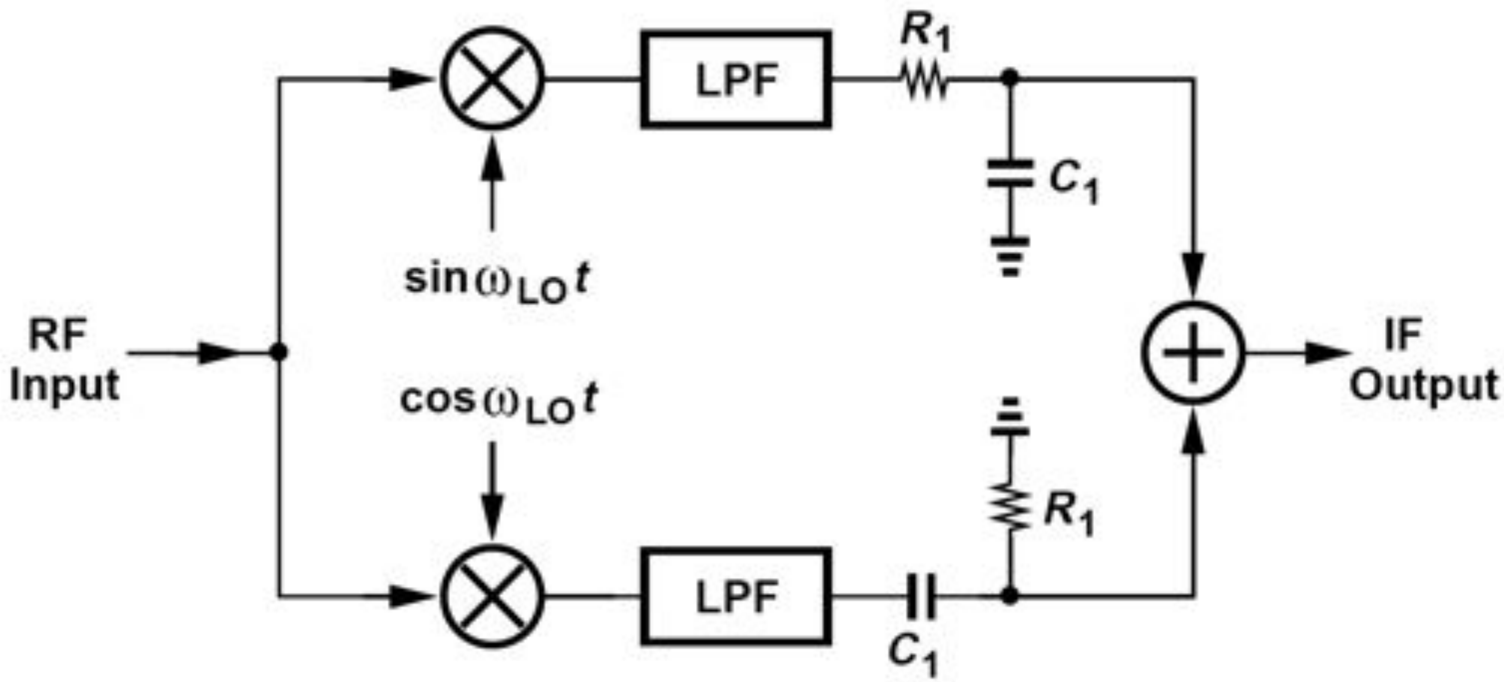
$$x_C(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_{LO})t + \phi_{sig}] - \frac{A_{im}}{2}\cos[(\omega_{im} - \omega_{LO})t + \phi_{im}].$$

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.**

Drawbacks of Hartley Architecture (I)

- **The principal drawback of the Hartley architecture stems from its sensitivity to mismatches**

We lump the mismatches of the receiver as a single amplitude error, *ε*, and phase error, *Δθ*,

$$x_A(t) = \frac{A_{sig}}{2}(1+\epsilon)\cos[(\omega_c-\omega_{LO})t+\phi_{sig}+\Delta\theta]+\frac{A_{im}}{2}(1+\epsilon)\cos[(\omega_{im}-\omega_{LO})t+\phi_{im}+\Delta\theta]$$

$$\begin{aligned} x_{sig}(t) &= \frac{A_{sig}}{2}(1+\epsilon)\cos[(\omega_c-\omega_{LO})t+\phi_{sig}+\Delta\theta] \\ &+ \frac{A_{sig}}{2}\cos[(\omega_c-\omega_{LO})t+\phi_{sig}] \\ x_{im}(t) &= \frac{A_{im}}{2}(1+\epsilon)\cos[(\omega_{im}-\omega_{LO})t+\phi_{im}+\Delta\theta] \\ &- \frac{A_{im}}{2}\cos[(\omega_{im}-\omega_{LO})t+\phi_{im}]. \end{aligned}$$

We divide the image-to-signal ratio at the input by the same ratio at the output, the result is called the "image rejection ratio" (IRR).

$$\frac{P_{im}}{P_{sig}}\Big|_{out} = \frac{A_{im}^2}{A_{sig}^2}\frac{(1+\epsilon)^2-2(1+\epsilon)\cos\Delta\theta+1}{(1+\epsilon)^2+2(1+\epsilon)\cos\Delta\theta+1}$$

$$\Rightarrow \mathrm{IRR} = \frac{(1+\epsilon)^2+2(1+\epsilon)\cos\Delta\theta+1}{(1+\epsilon)^2-2(1+\epsilon)\cos\Delta\theta+1}$$

Simplifying the Expression of IRR

If ε << 1 rad, simplify the expression for IRR.

Solution:

Since cos $\Delta\theta \approx 1 - \Delta\theta^2/2$ for $\Delta\theta$ <<1 rad, we can reduce equation above to

$$\mathrm{IRR} \approx \frac{4 + 4\epsilon + \epsilon^2 - (1+\epsilon)\Delta\theta^2}{\epsilon^2 + (1+\epsilon)\Delta\theta^2}$$

In the numerator, the first term dominates and in the denominator ε << 1, yielding

$$\mathrm{IRR} \approx \frac{4}{\epsilon^2 + \Delta\theta^2}$$

For example, ε = 10% (≈ 0.83 dB) limits the IRR to 26 dB. Similarly, $\Delta\theta$ = 10 ° yields an IRR of 21 dB. While such mismatch values may be tolerable in direct-conversion receivers, they prove inadequate here.

Drawbacks of Hartley Architecture (Ⅱ)

- **Another critical drawback originates from the variation of the absolute values of R_1 and C_1.**

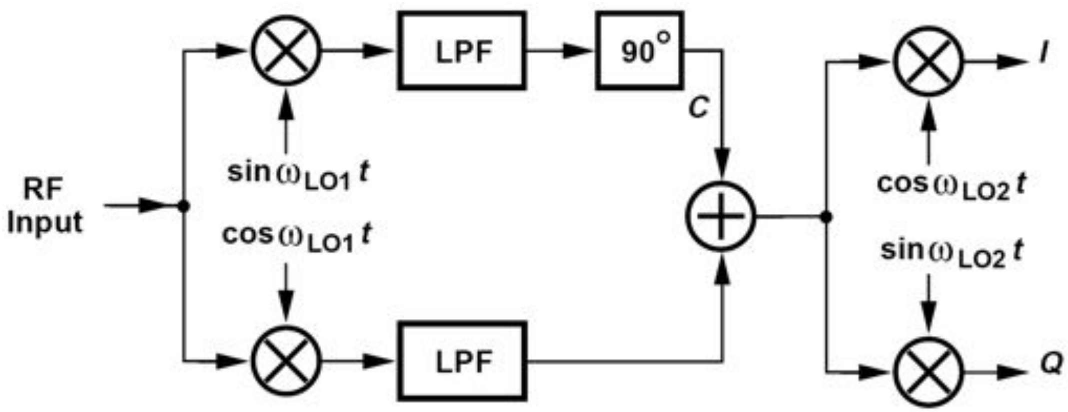
Specifically, if R_1 and C_1 are nominally chosen for a certain IF, $(R_1C_1)^{-1} = \omega_{IF}$, but respectively experience a small change of ΔR and ΔC with process or temperature, then:

$$\left|\frac{H_{HPF}}{H_{LPF}}\right| = (R_1 + \Delta R)(C_1 + \Delta C)\omega_{IF} \approx 1 + \frac{\Delta R}{R_1} + \frac{\Delta C}{C_1}.$$

Thus, the gain mismatch is equal to: $$\epsilon = \frac{\Delta R}{R_1} + \frac{\Delta C}{C_1}$$

- **Another drawback resulting from the *RC-CR* sections manifests itself if the signal translated to the IF has a wide bandwidth.**

The image rejection may degrade substantially near the edges of the channel



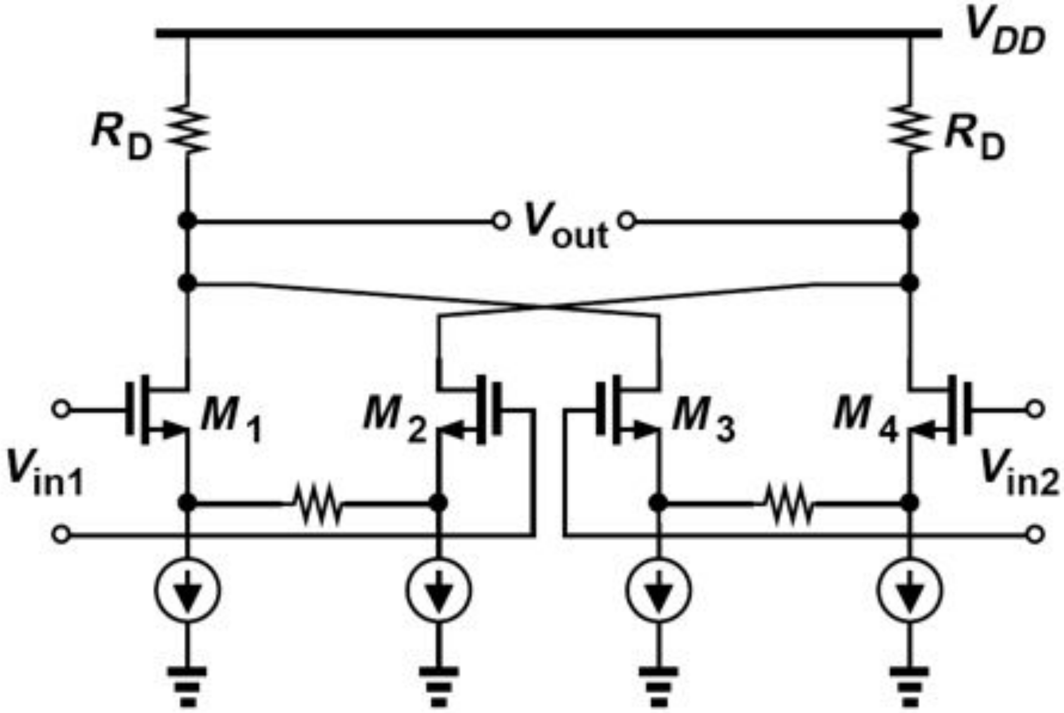
$$\mathrm{IRR} = \left(\frac{\omega_{IF}}{\Delta\omega}\right)^2$$

The limitation expressed above implies that ω_{IF} cannot be zero, dictating a heterodyne approach, one example shown left.

Drawbacks of Hartley Architecture (III)

- The *RC-CR* sections used above also introduce attenuation and noise.

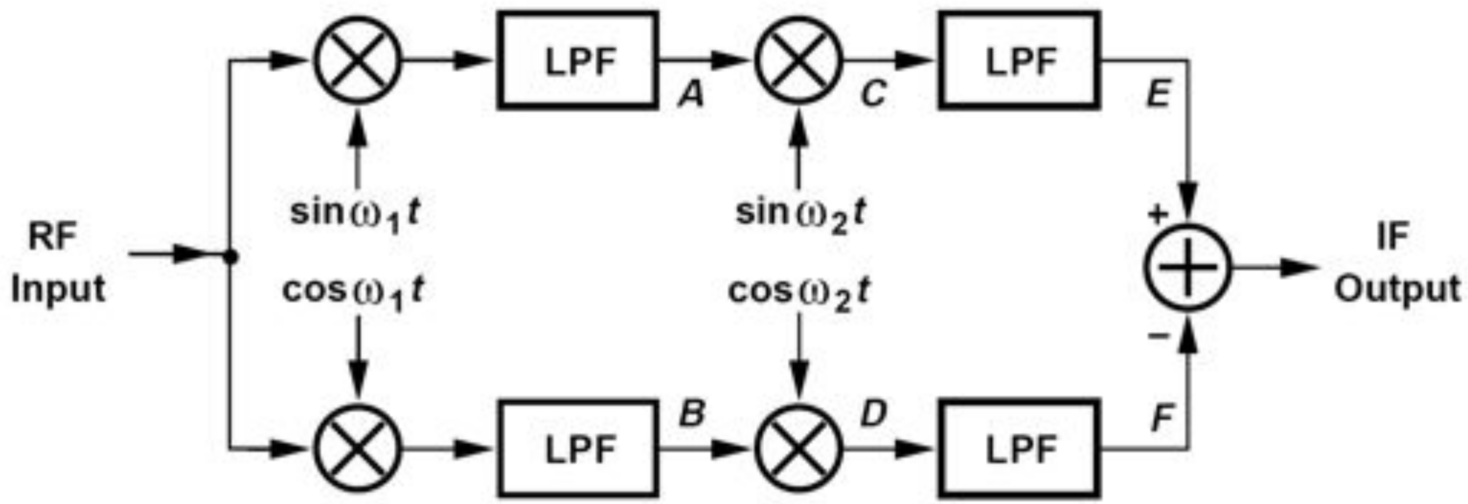
- The voltage adder at the output of the Hartley architecture also poses difficulties as its noise and nonlinearity appear in the signal path.



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.

Weaver Architecture: Overview

The Weaver receiver, derived from its transmitter counterpart, avoids those issues in Hartley architecture.



- Mixing a signal with quadrature phases of an LO takes the Hilbert transform. Depicted above, the Weaver architecture replaces the 90° phase shift network with quadrature mixing.

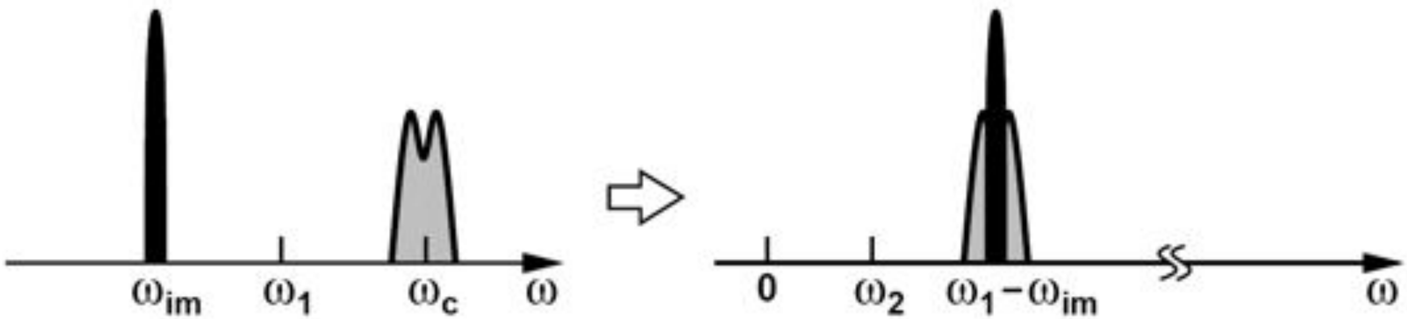
Weaver Architecture: Formulating the Circuit's Behavior

We perform the second quadrature mixing operation, arriving at

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

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

Let us assume low-side injection for both mixing stages.

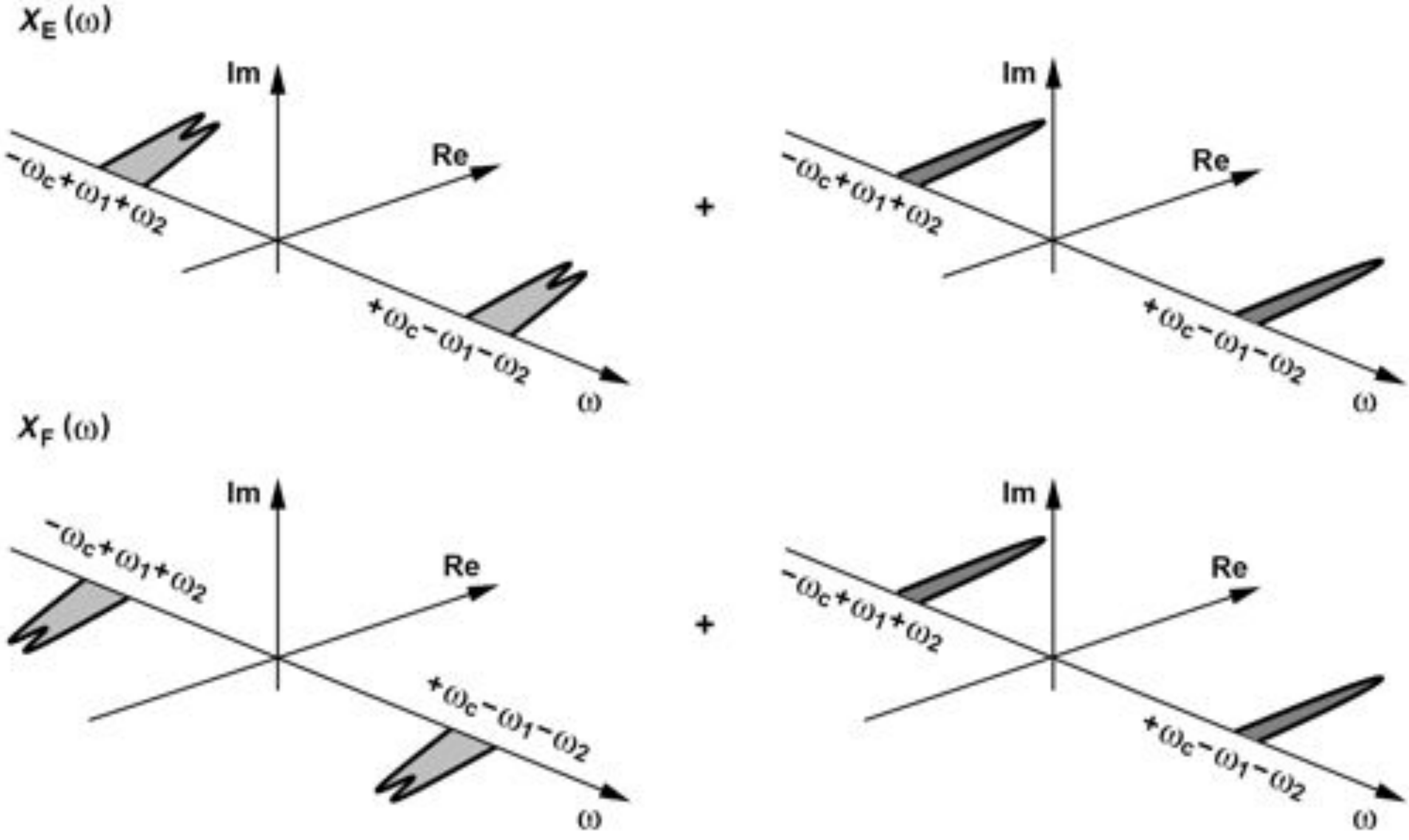


$$x_E(t) - x_F(t) = \frac{A_{sig}}{2}\cos[(\omega_c - \omega_1 - \omega_2)t + \phi_{sig}]$$

Graphically Analysis of Weaver Architecture

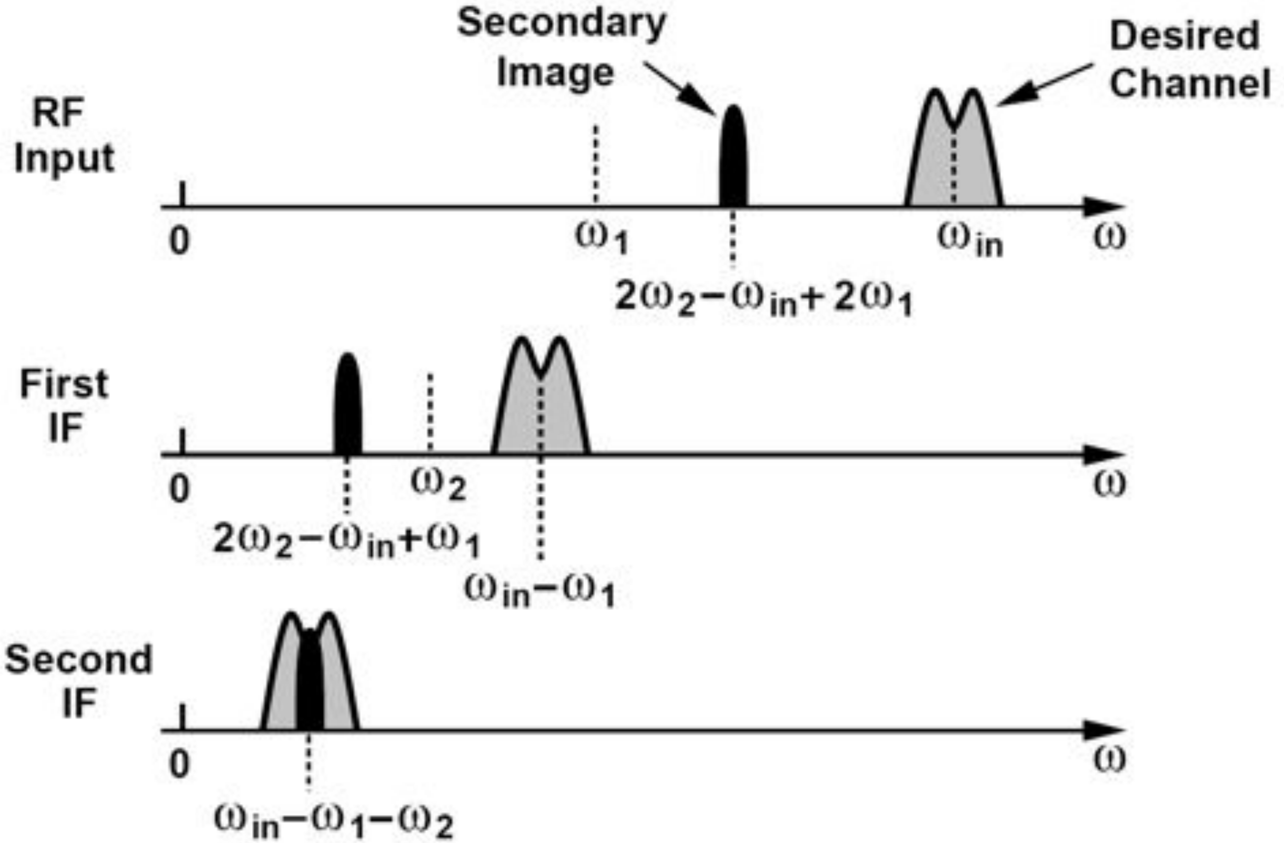
Perform the above analysis graphically. Assume low-side injection for both mixing stages.

Recall that low-side injection mixing with a sine multiplies the spectrum by +(*j*/2)sgn(*ω*). Subtraction of $X_F(f)$ from $X_E(f)$ thus yields the signal and removes the image.



Secondary Image in Weaver Architecture

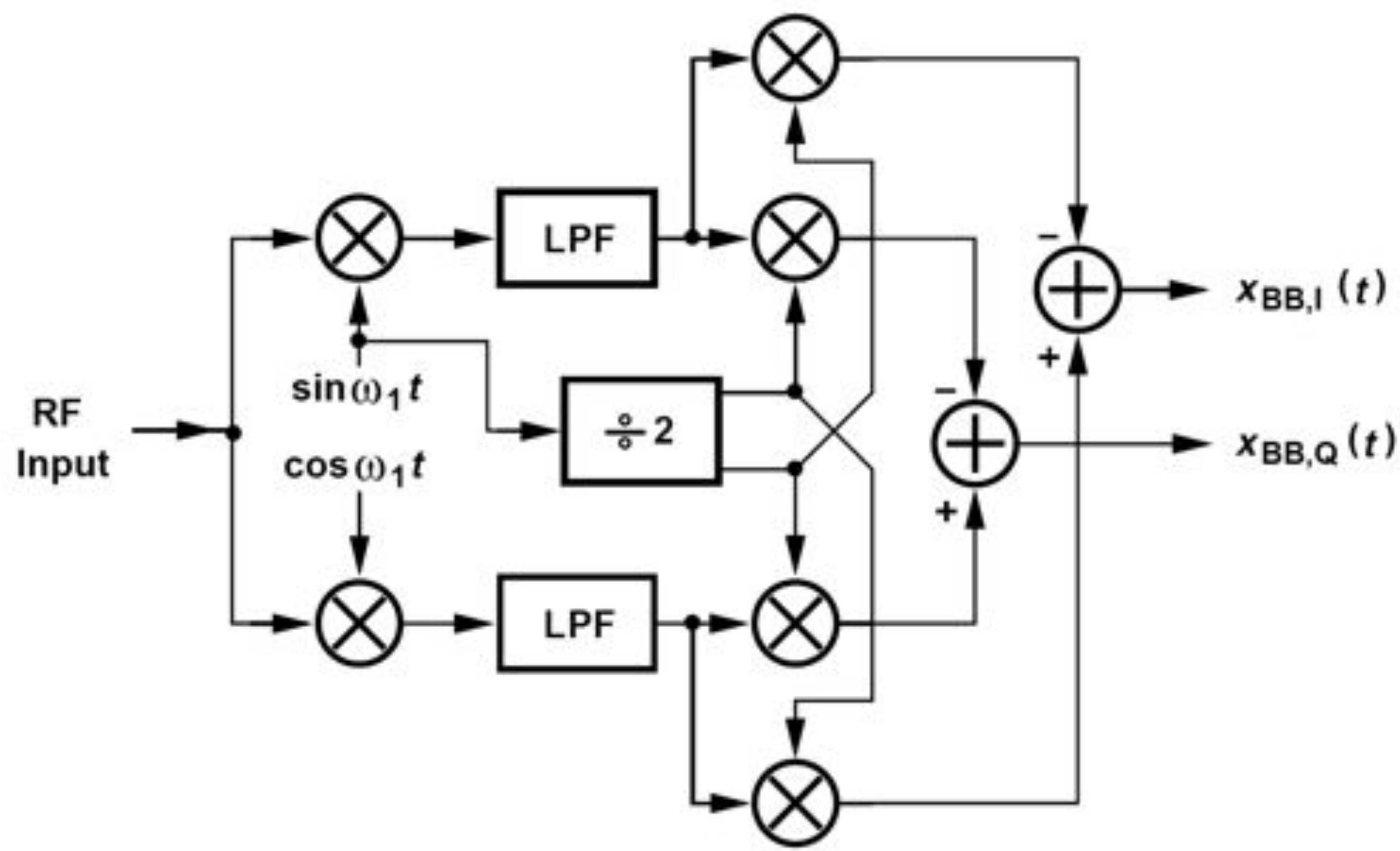
The Weaver architecture must deal with a secondary image if the second IF is not zero.



- **This effect arises if a component at $2\omega_2 - \omega_{in} + 2\omega_1$ accompanies the RF signal. Downconversion to the first IF translates this to image of the signal with respect to ω_2, and mixing with ω_2 brings it to the same IF at which the signal appears.**

Double Quadrature Downconversion Weaver Architecture

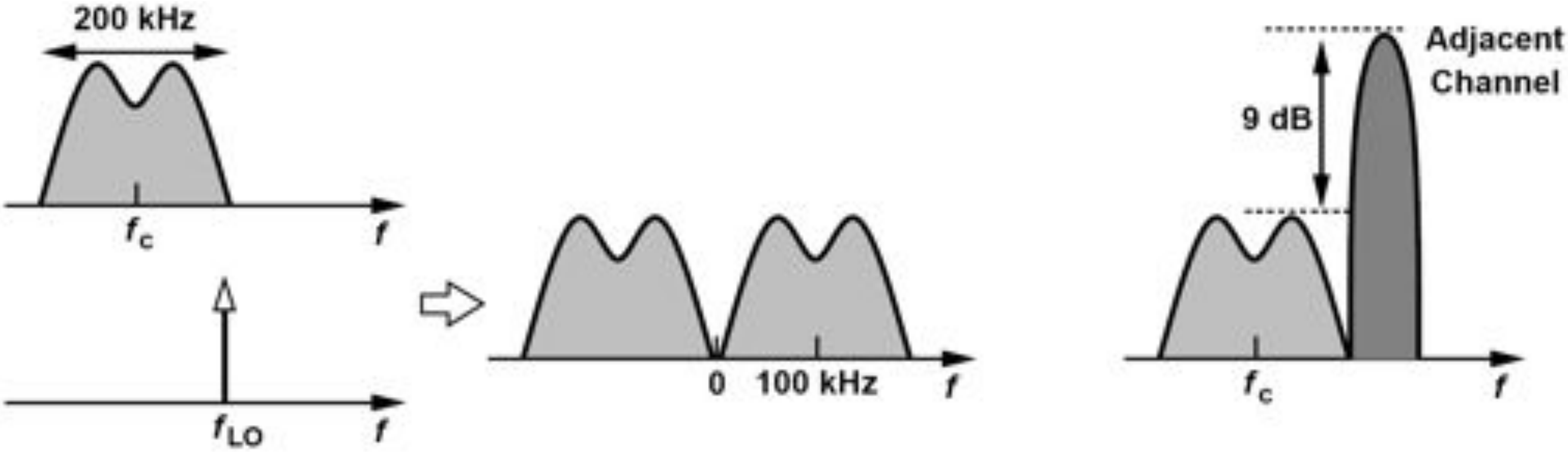
For above reason, the second downconversion preferably produces a zero IF, in which case it must perform quadrature separation as well.



Low- IF Receivers: Overview

We noted that it is undesirable to place the image within the signal band because the image thermal noise of the antenna, the LNA, and the input stage of the RF mixer would raise the overall noise figure by approximately 3 dB.

Now suppose the LO frequency is placed at the edge of the desired (200-kHz) channel:



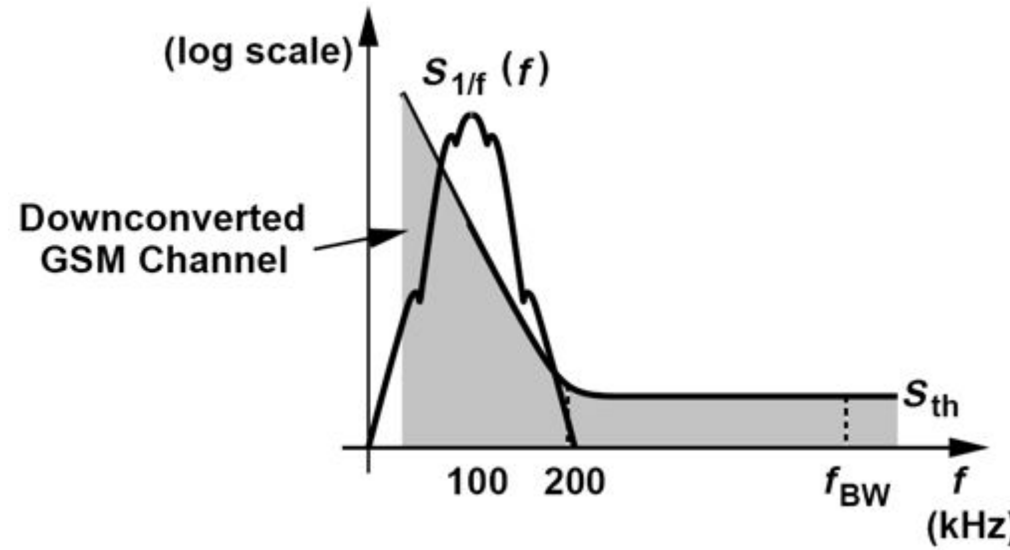
- The 1/*f* noise penalty is much less severe. Also, on-chip high-pass filtering of the signal becomes feasible.

Flicker Noise Penalty of Low-IF Receiver

Repeat the flicker noise penalty analysis for a low-IF receiver.

Assuming that high-pass filtering of dc offsets also removes the flicker noise up to roughly 20 kHz, we integrate the noise from 20 kHz to 200 kHz

$$
\begin{aligned}
P_{n1} &= \int_{20\text{ kHz}}^{200\text{ kHz}} \frac{\alpha}{f} df \\
&= f_c \cdot S_{th} \ln 10 \\
&= 2.3 f_c S_{th}.
\end{aligned}
$$



Without flicker noise,

$$P_{n2} \approx (200\text{ kHz}) S_{th}$$

It follows that

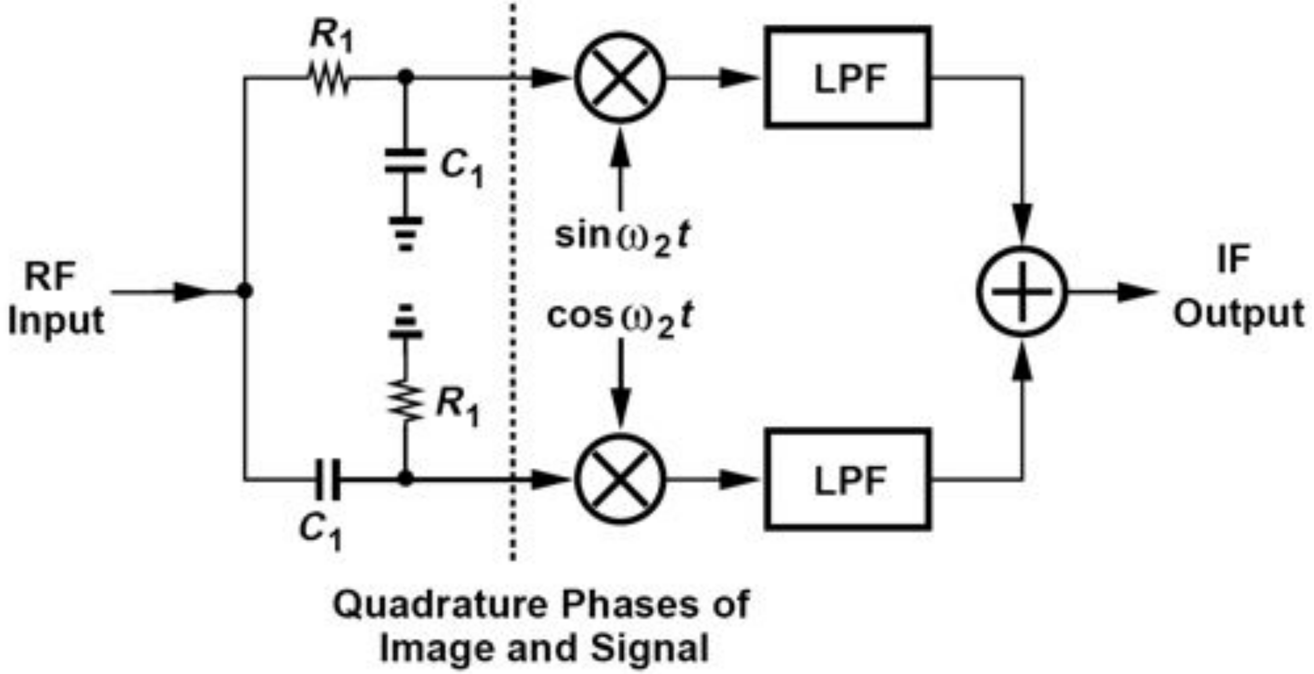
$$\frac{P_{n1}}{P_{n2}} = 2.3 (= 3.62\text{ dB})$$

The flicker noise penalty is therefore much lower in this case.

Image Rejection in Low- IF Receivers (Ⅰ)

The IF spectrum in a low-IF RX may extend to zero frequency, making the Hartley architecture impossible to maintain a high IRR across the signal bandwidth.

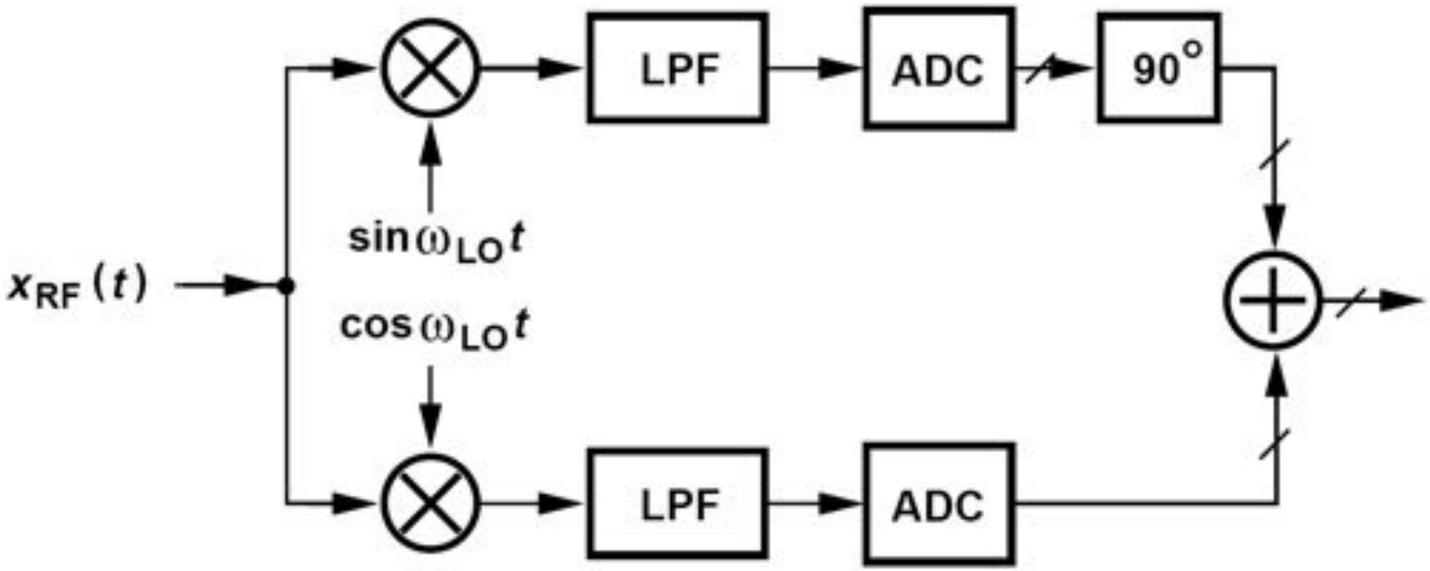
- **One possible remedy is to move the 90 ° phase shift in the Hartley architecture from the IF path to the RF path.**



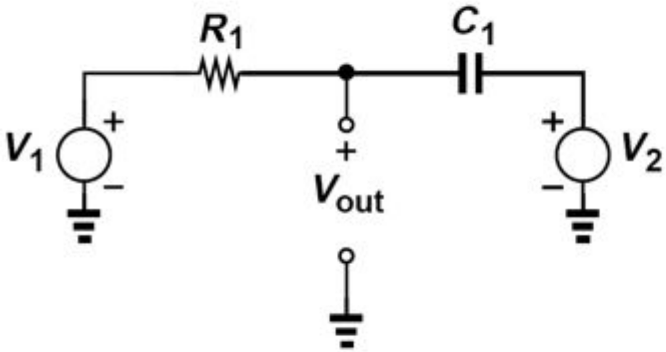
The *RC-CR* network is centered at a high frequency and can maintain a reasonable IRR across the band.

Image Rejection in Low- IF Receivers (Ⅱ)

- Another variant of the low-IF architecture is shown below, the downconverted signals are applied to channel-select filters and amplifiers as in a direct-conversion receiver. The results are then digitized and subjected to a Hilbert transform in the digital domain before summation.



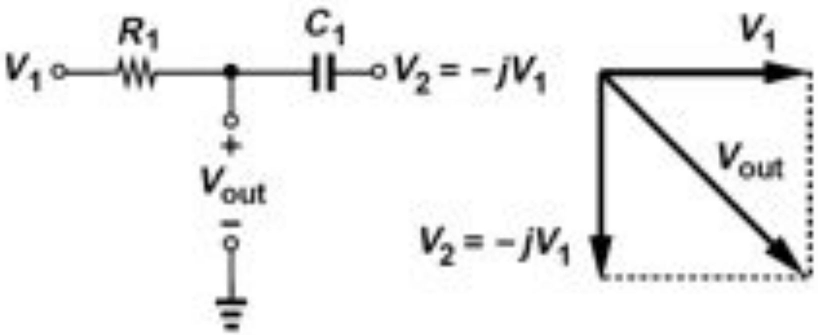
Polyphase Filters-How it Works



V_{out} **can be viewed as a weighted sum of** V_1 **and** V_2

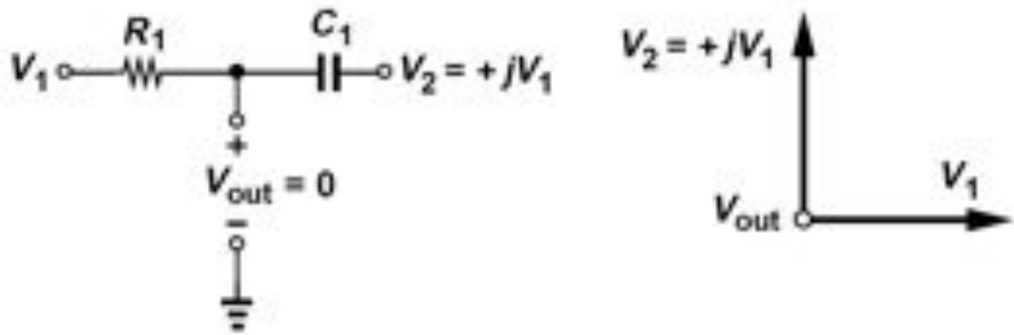
$$V_{out} = \frac{V_1 + R_1C_1sV_2}{R_1C_1s + 1}$$

We consider two special cases:



$$V_{out} = \frac{V_1 + R_1C_1sV_2}{R_1C_1s + 1}$$

In this case, the circuit simply computes the vector summation of V_1 **and** $V_2 = -jV_1$**.**



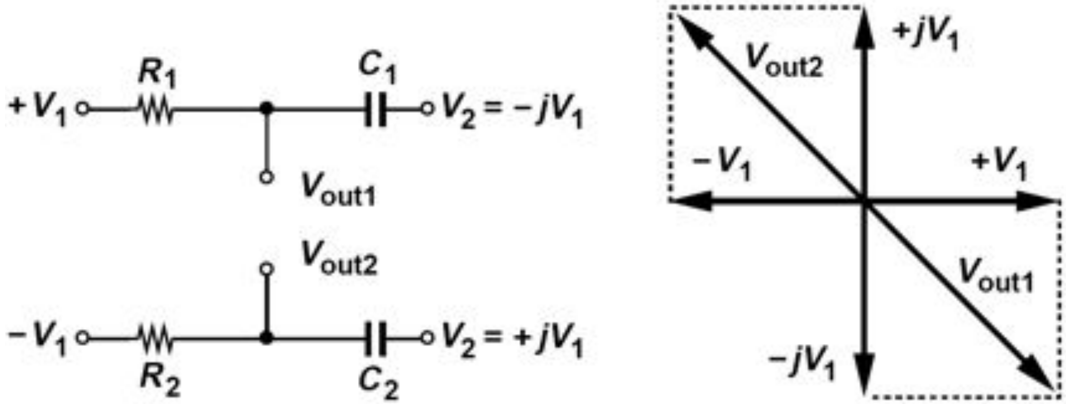
$$V_{out} = V_1 \frac{R_1C_1\omega + 1}{jR_1C_1\omega + 1}$$

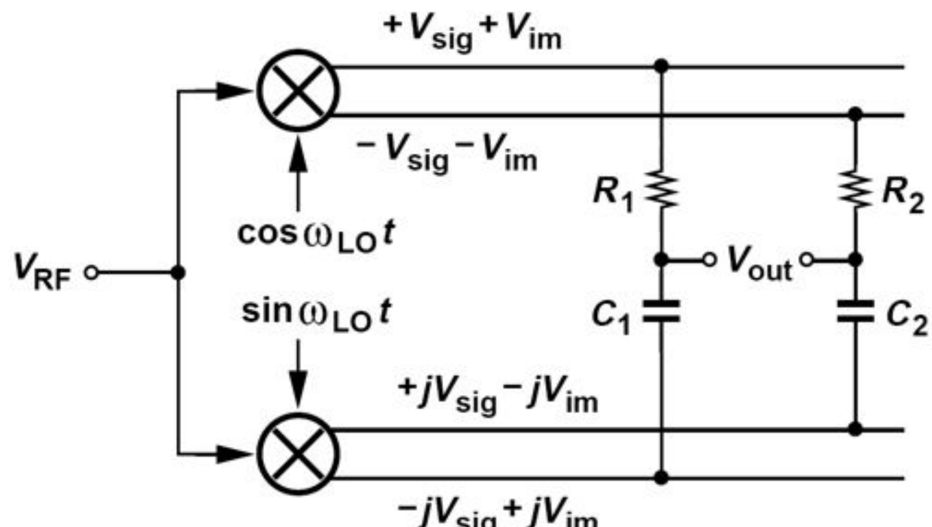
C_1 **rotates** V_2 **by another 90 ° so that the result cancels the effect of** V_1 **at the output node**

Differential Form of Polyphase Filters

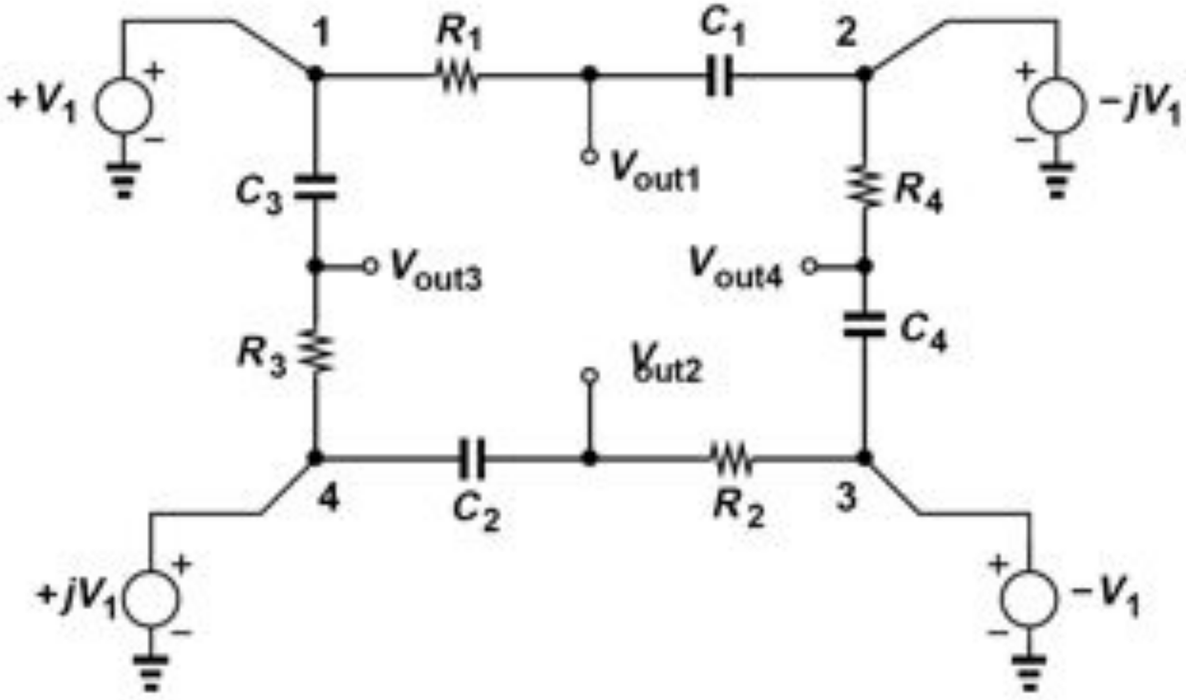
Extend the topology above if V_1 and $-jV_1$ are available in differential form and construct an image-reject receiver.

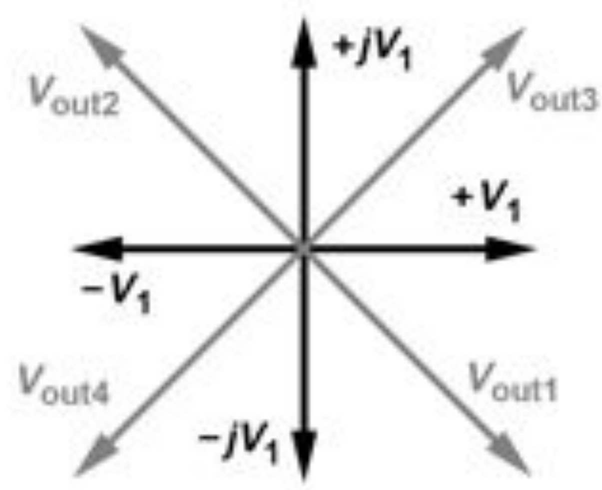
Figure below (top) shows the arrangement and the resulting phasors if $R_1 = R_2 = R$ and $C_1 = C_2 = C$. The connections to quadrature downconversion mixers are depicted in figure below (bottom).





RC Network Sensing Differential Quadrature Phases

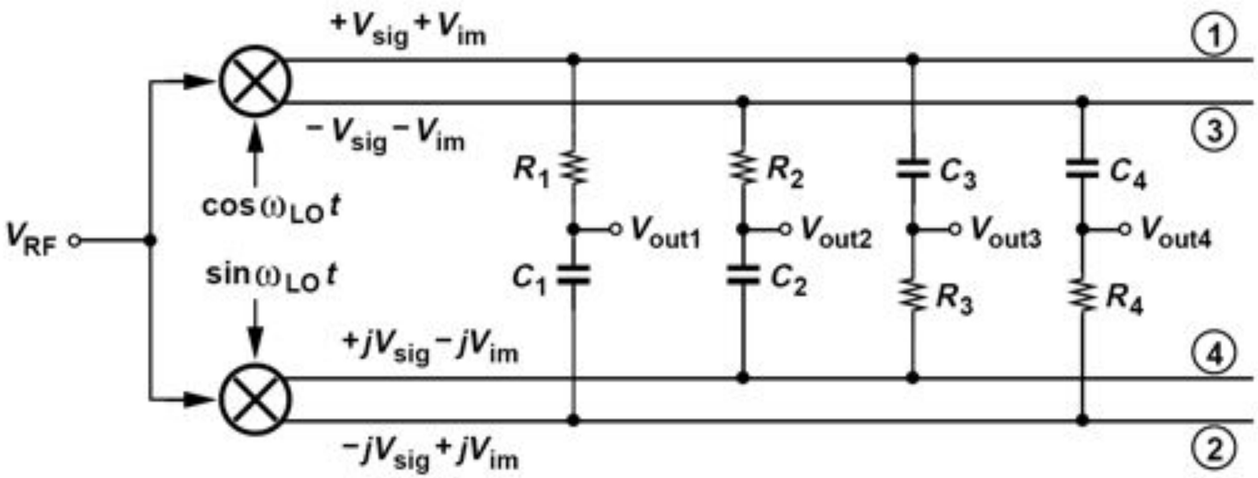


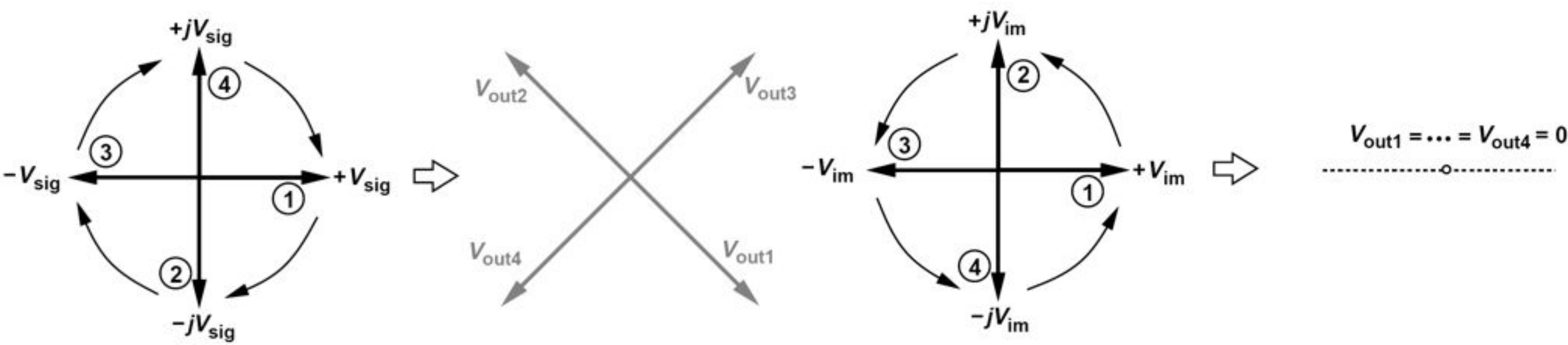


- The circuit produces quadrature outputs that are 45 ° out of phase with respect to the quadrature inputs.

Example of Quadrature Downconverter Driving RC Sections

The outputs of a quadrature downconverter contain the signal, V_{sig}, and the image, V_{im}, and drive the circuit above as shown in figure below. Determine the outputs, assuming all capacitors are equal to C and all resistors equal to R.





Effect of Polyphase Filter at a Frequency Offset

We Substituting $\omega = (R_1C_1)^{-1} + \Delta\omega$ in Eq. (4.93), we have:

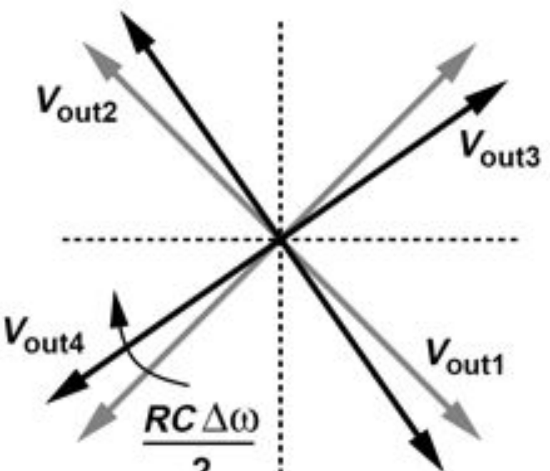
$$V_{out1} = V_{sig}\frac{2 + RC\Delta\omega}{1 + j(1 + RC\Delta\omega)}$$

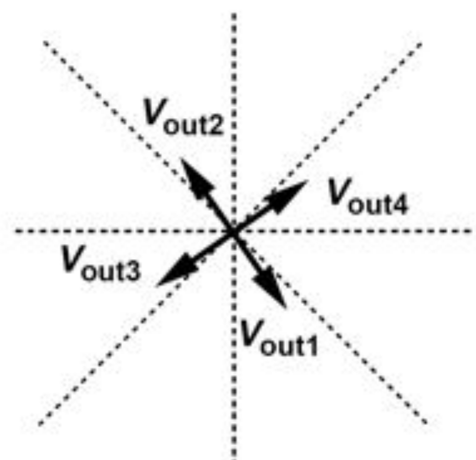
Hence,

$$\begin{aligned} |V_{out1}|^2 &= |V_{sig}|^2 \frac{4 + 4RC\Delta\omega + R^2C^2\Delta\omega^2}{2 + 2RC\Delta\omega + R^2C^2\Delta\omega^2} \\ &\approx 2|V_{sig}|^2\left(1 + RC\Delta\omega + \frac{R^2C^2\Delta\omega^2}{4}\right)\left(1 - RC\Delta\omega - \frac{R^2C^2\Delta\omega^2}{2}\right) \\ &\approx 2|V_{sig}|^2\left(1 - \frac{5}{4}R^2C^2\Delta\omega^2\right). \end{aligned}$$

$$|V_{out1}| \approx \sqrt{2}|V_{sig}|\left(1 - \frac{5}{8}R^2C^2\Delta\omega^2\right)$$

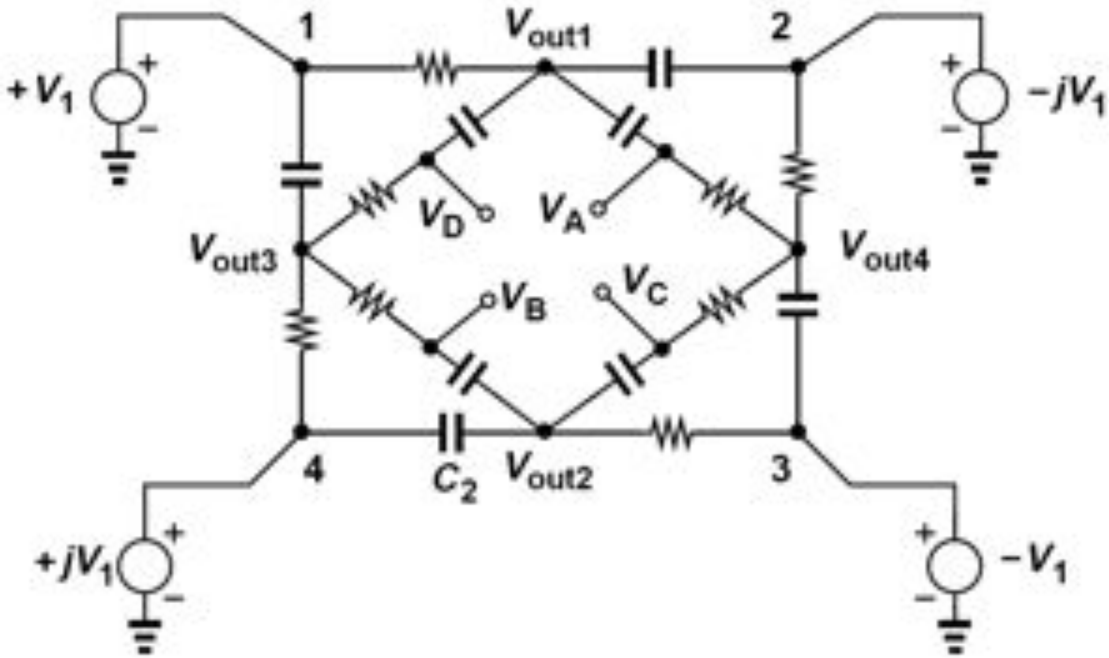
$$\angle V_{out1} = \angle V_{sig} - \tan^{-1}(1 + RC\Delta\omega) \qquad \angle V_{out1} = \angle V_{sig} - \left(\frac{\pi}{4} + \frac{RC\Delta\omega}{2}\right)$$

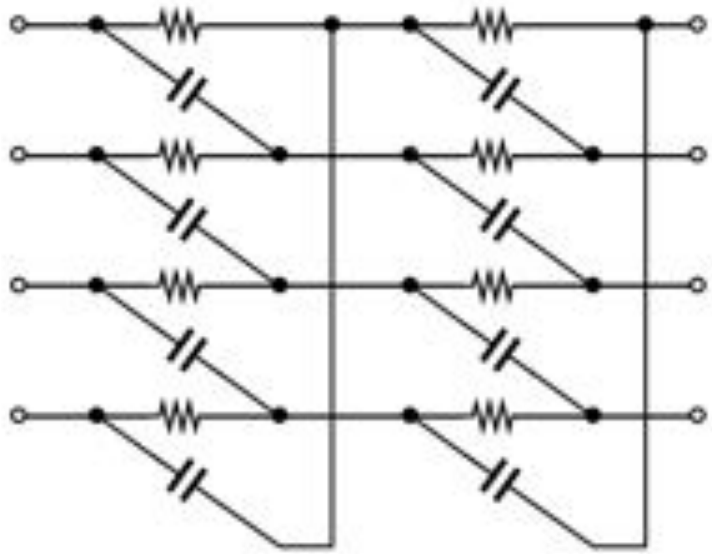




Cascaded Polyphase Sections- Overview

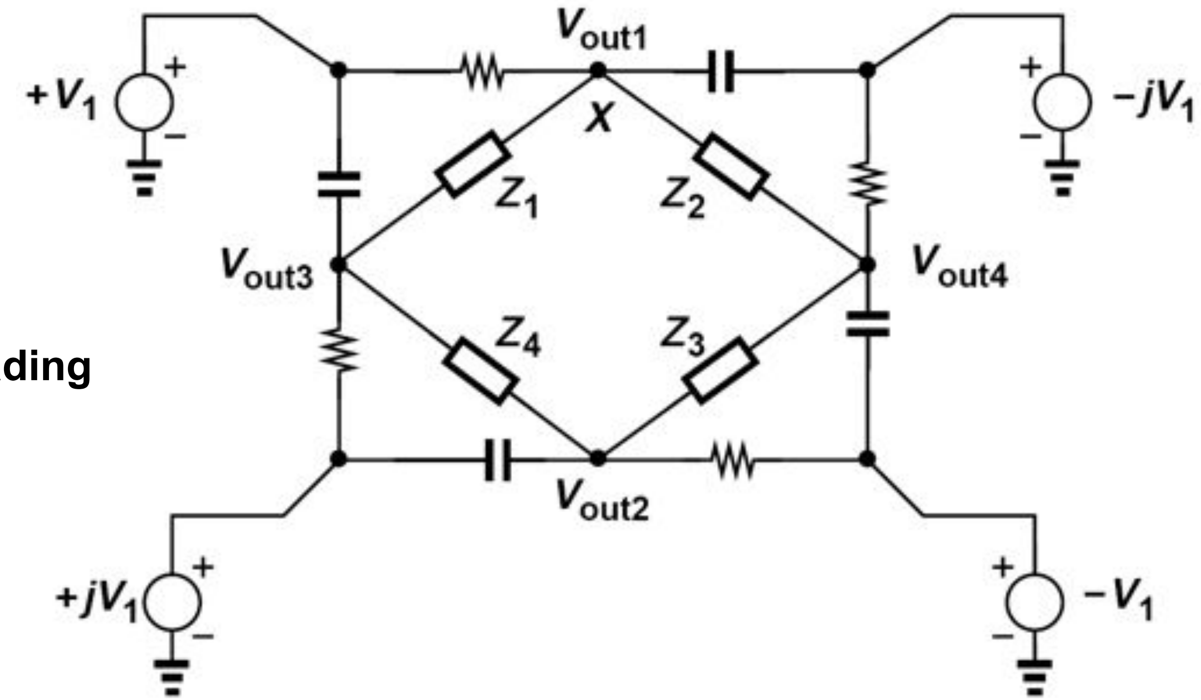
The output signal and image components exhibit opposite sequences. We therefore expect that if this polyphase filter is followed by another, then the image can be further suppressed.





- **We must answer two questions:**
  (1) how should we account for the loading of the second stage on the first?
  (2) how are the *RC* values chosen in the two stages?

Effect of Loading of Second Polyphase Section



If $Z_1 = \cdots = Z_4 = Z$, then, V_{out1}-V_{out4} experience no rotation, but the loading may reduce their magnitudes.

$$\frac{\alpha(1-j)V_1 - V_1}{R} + [\alpha(1-j)V_1 + jV_1]Cj\omega + \frac{\alpha(1-j)V_1 - \alpha(1+j)V_1}{Z}$$
$$+ \quad \frac{\alpha(1-j)V_1 + \alpha(1+j)V_1}{Z} = 0.$$

If $RC\omega = 1$, the expression reduces to

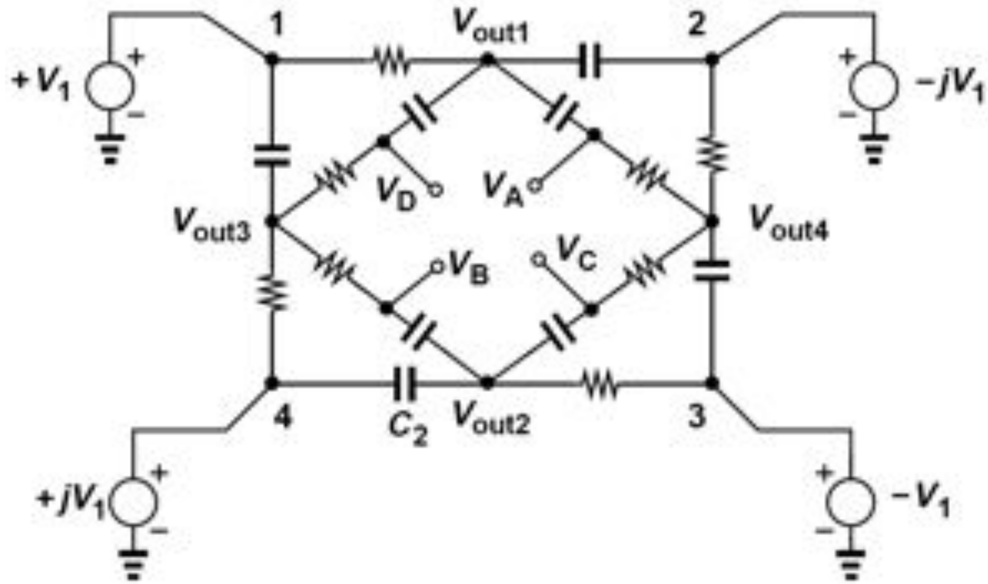
$$2\alpha - 2 + \frac{2\alpha(1-j)R}{Z} = 0.$$

$$\Rightarrow \quad \alpha = \frac{Z}{Z + (1-j)R}$$

Example of Loading of the Second Polyphase Section

If $Z = R + (jC\omega)^{-1}$ and $RC\omega = 1$, determine V_A in figure below.

Solution:



We have $V_{out1} = (1/2)(1 - j)V_1$ and $V_{out4} = (1/2)(-1 - j)V_1$, observing that V_{out1} and V_{out2} have the same phase relationship $-j$. Thus, V_A is simply the vector sum of V_{out1} and V_{out4}:

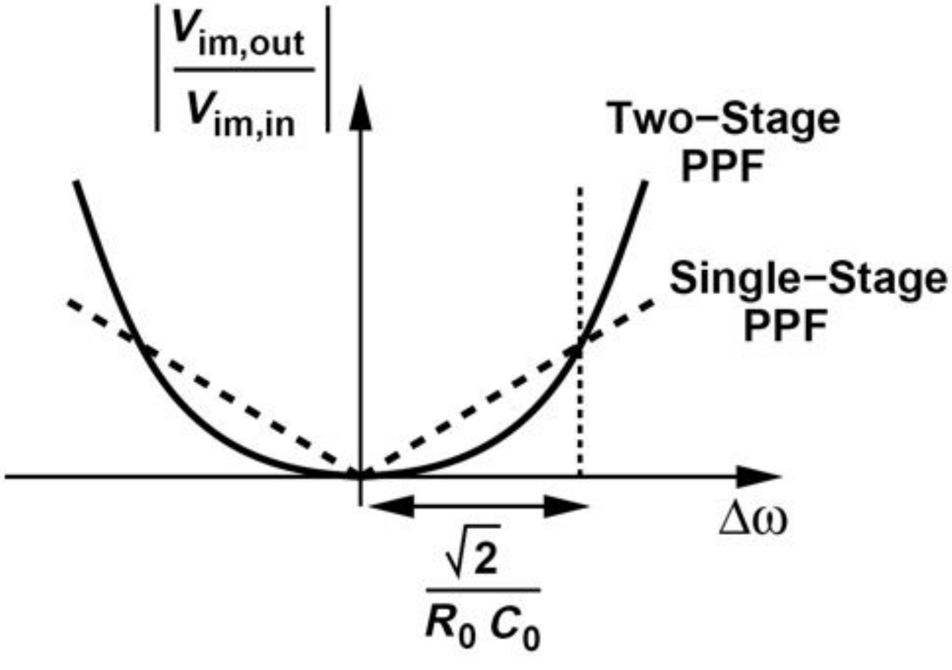
$$V_A = -jV_1$$

In comparison with single-section polyphase filter, we note that a two-section polyphase filter produces an output whose magnitude is $\sqrt{2}$ times smaller than that of a single-section counterpart. We say each section attenuates the signal by a factor of $\sqrt{2}$.

Choice of *RC* Values

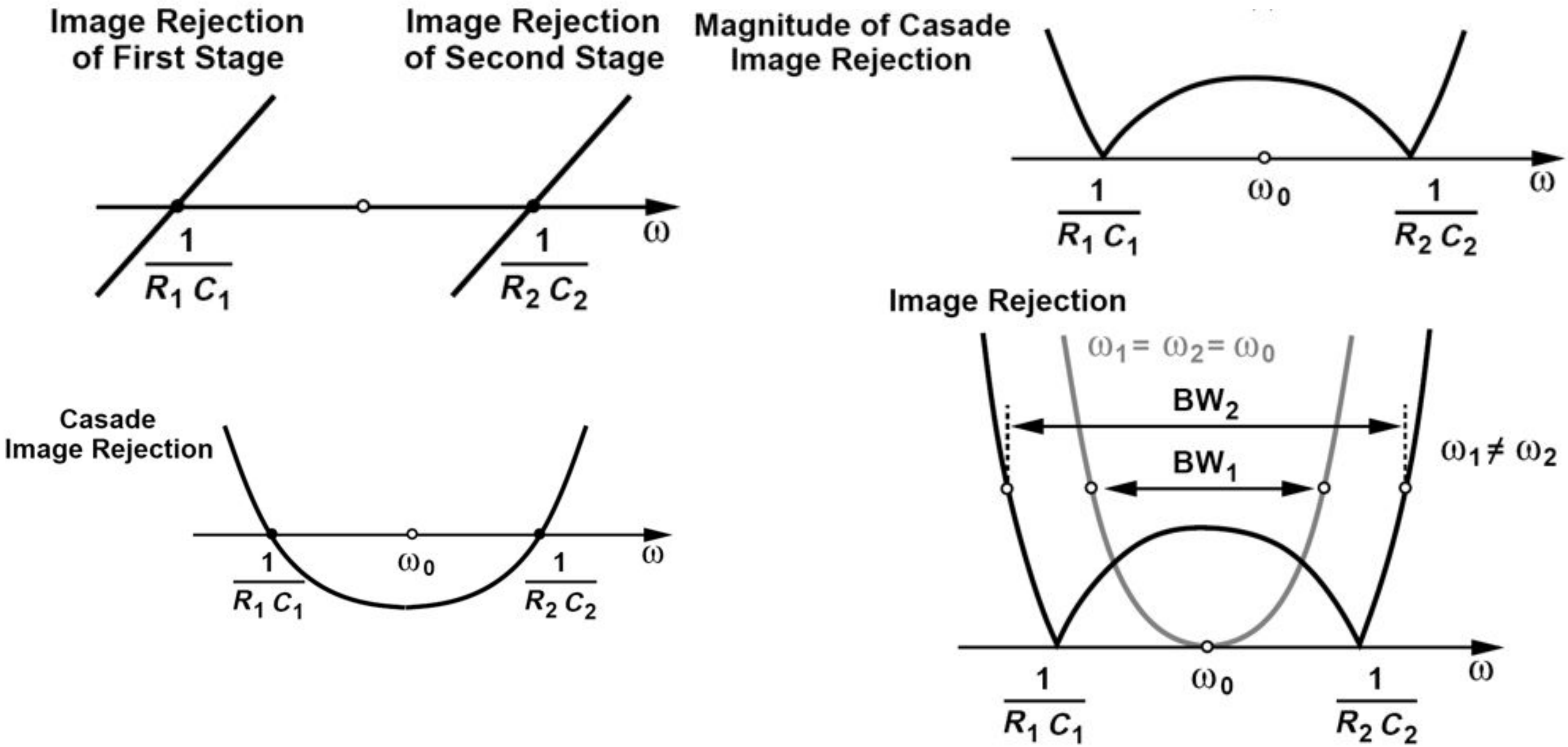
The cascade of two stages yields an image attenuation at a frequency of $(R_0C_0)^{-1} + \Delta\omega$ equal to:

$$\left|\frac{V_{im,out}}{V_{im,in}}\right| \approx \frac{(R_0C_0\Delta\omega)^2}{2 + 2R_0C_0\Delta\omega}$$



Two Unidentical Polyphase Sections

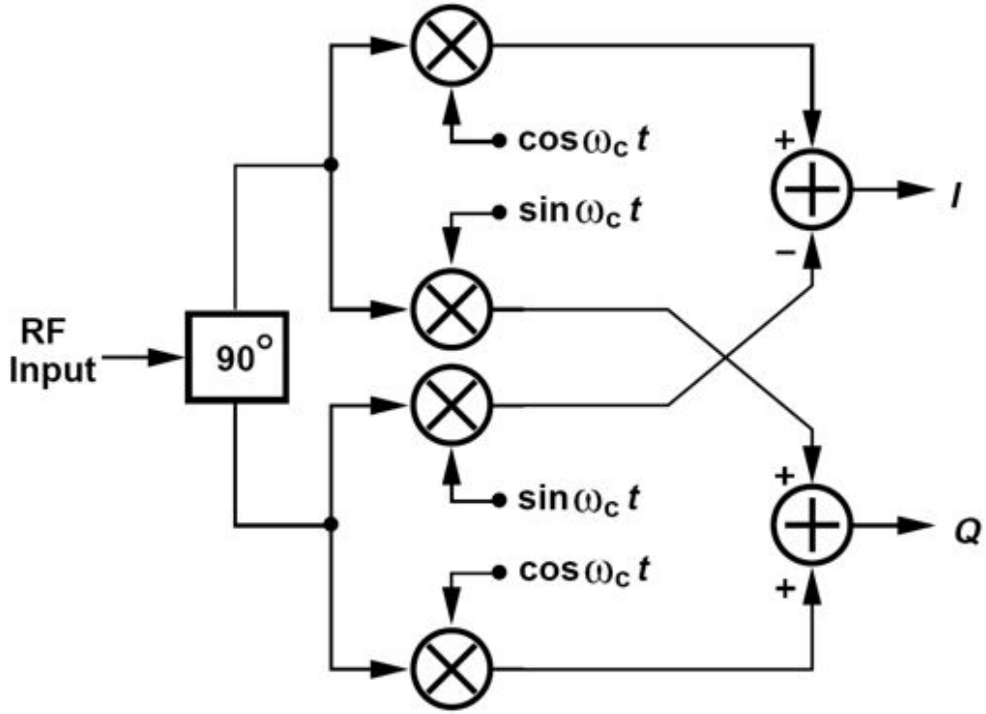
What happens if the two stages use different time constants?



- **The advantage of splitting the cut-off frequencies of the two stages is the wider achievable bandwidth.**

Double-Quadrature Downconversion

A method of reducing the effect of mismatches incorporates “double-quadrature” downconversion



The overall gain and phase mismatches of this topology are given by:

$$\frac{\Delta A}{A} = \frac{\Delta A_{RF}}{A_{RF}} \cdot \frac{\Delta A_{LO}}{A_{LO}} + \frac{\Delta G_{mix}}{G_{mix}}$$

$$\tan(\Delta\phi) = \tan(\Delta\phi_{RF}) \cdot \tan(\Delta\phi_{LO}) + \frac{\tan(\Delta\phi_{mix})}{2}$$

Transmitter Architecture: General Considerations

An RF transmitter performs modulation, upconversion, and power amplification.

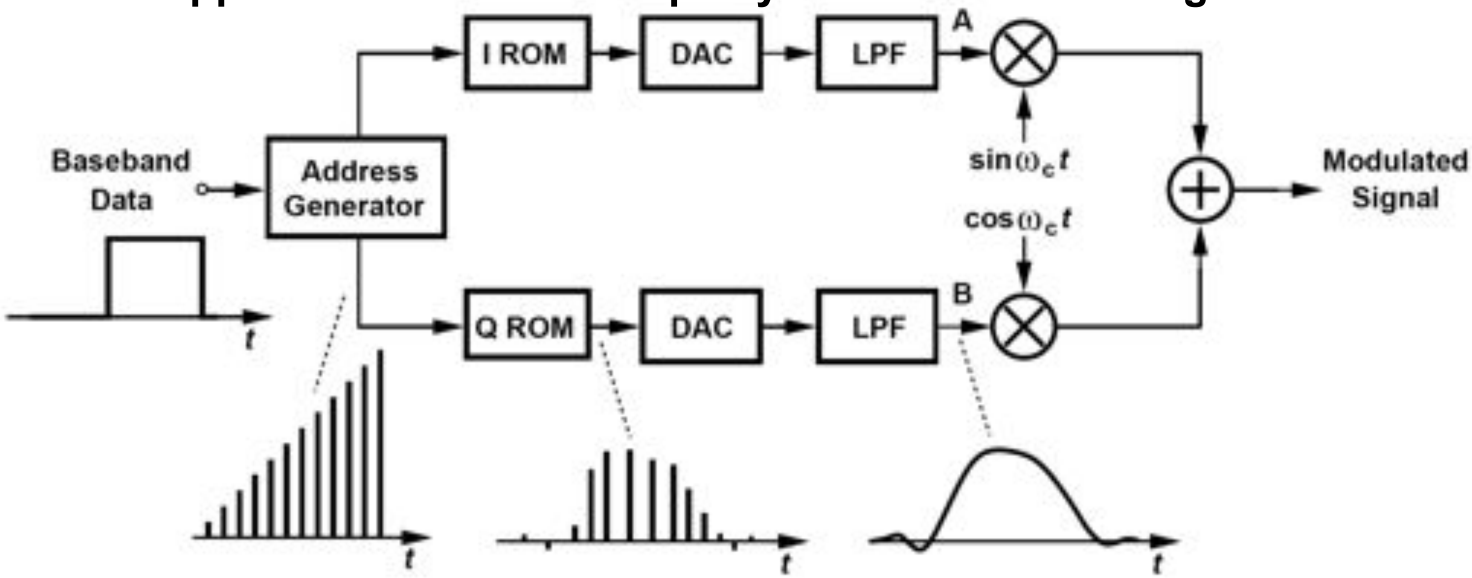
The GMSK waveform in GSM can be expanded as

$$\begin{aligned} x_{GMSK}(t) &= A\cos[\omega_c t + m\int x_{BB}(t) * h(t)dt] \\ &= A\cos\omega_c t\cos\phi - A\sin\omega_c t\sin\phi \end{aligned}$$

where $\phi = m\int x_{BB} * h(t)dt.$

Thus, ***cosΦ*** and ***sinΦ*** are produced from $x_{BB}(t)$ by the digital baseband processor, converted to analog form by D/A converters, and applied to the transmitter.

Each incoming pulse is mapped to the desired shape by a combination of digital and analog techniques:



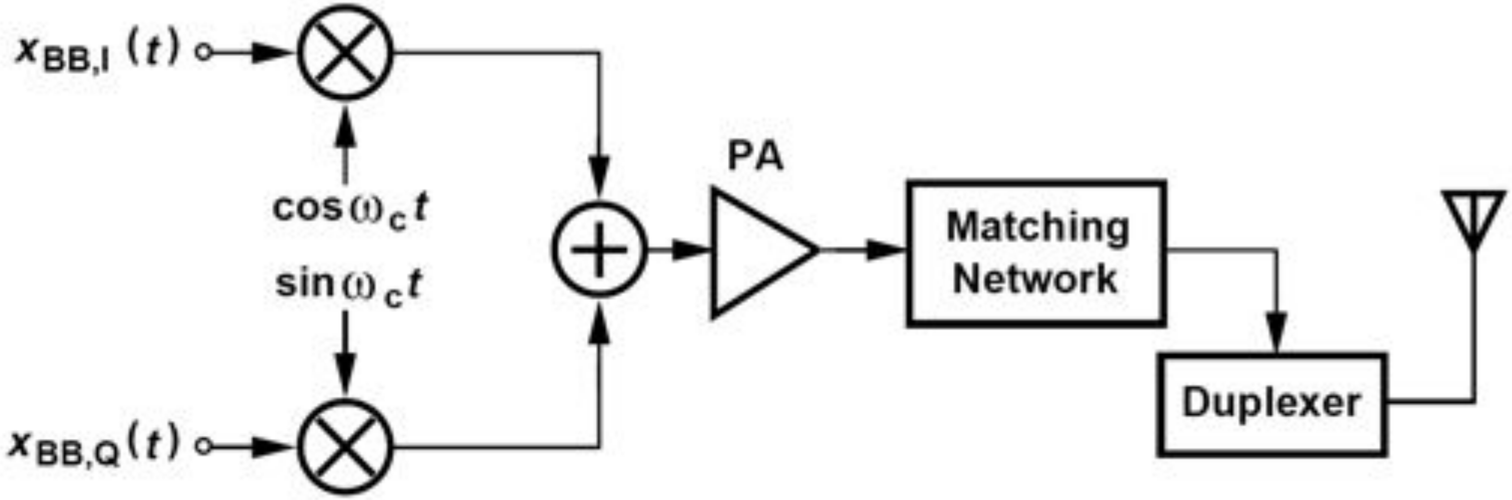
Direct-Conversion Transmitters: Overview

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

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

We therefore define the quadrature baseband signals as

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



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

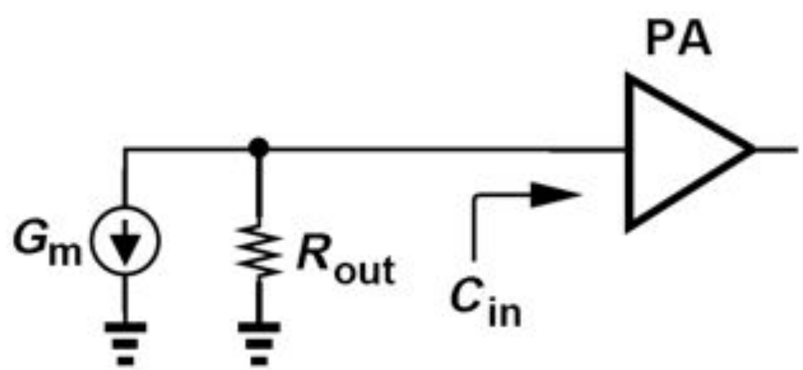
Example of Scaling of Quadrature Upconverter

A student decides to omit the predriver and simply "scale up" the upconverter so that it can drive the PA directly. Explain the drawback of this approach.

Solution:

In order to scale up the upconverter, the width and bias current of each transistor are scaled up, and the resistor and inductor values are proportionally scaled down. For example, if the upconverter is modeled as a transconductance G_m and an output resistance R_{out}, then R_{out} can be reduced to yield adequate bandwidth with the input capacitance of the PA, and G_m can be enlarged to maintain a constant $G_m R_{out}$ (i.e., constant voltage swings). In practice, the upconverter employs a resonant LC load, but the same principles still apply.

The scaling of the transistors raises the capacitances seen at the baseband and LO ports of the mixers in figure above. The principal issue here is that the LO now sees a large load capacitance, requiring its own buffers. Also, the two mixers consume a higher power.



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:

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

I/Q Mismatch: Another Approach of Quantification

Another approach to quantifying the I/Q mismatch in a transmitter involves applying two tones $V_0 \cos\omega_{in} t$ and $V_0 \sin\omega_{in} t$ to the *I* and *Q* inputs and examining the output spectrum.

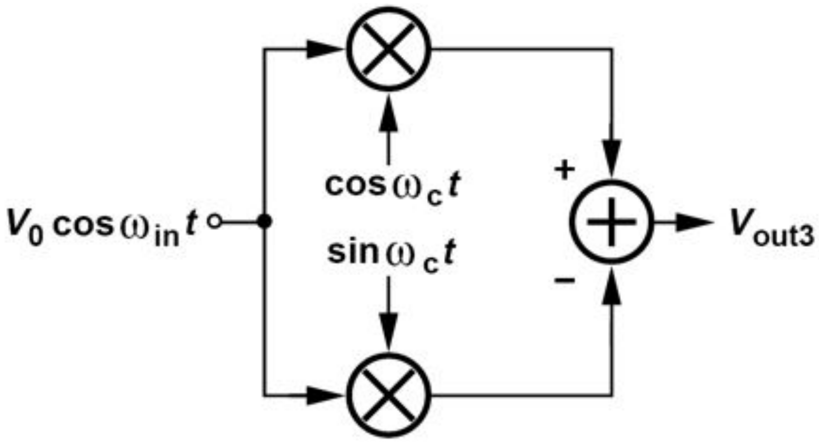
$$\begin{aligned} V_{out}(t) &= V_0(1+\varepsilon)\cos\omega_{in}t\cos(\omega_c t+\Delta\theta) - V_0\sin\omega_{in}t\sin\omega_c t \\ &= \frac{V_0}{2}[(1+\varepsilon)\cos\Delta\theta + 1]\cos(\omega_c t+\omega_{in})t \\ &- \frac{V_0}{2}(1+\varepsilon)\sin\Delta\theta\sin(\omega_c+\omega_{in})t \\ &+ \frac{V_0}{2}[(1+\varepsilon)\cos\Delta\theta - 1]\cos(\omega_c-\omega_{in})t \\ &- \frac{V_0}{2}(1+\varepsilon)\sin\Delta\theta\sin(\omega_c-\omega_{in})t. \end{aligned}$$

The power of the unwanted sideband at $\omega_c - \omega_{in}$ divided by that of the wanted sideband at $\omega_c + \omega_{in}$ is given by

$$\begin{aligned} \frac{P_-}{P_+} &= \frac{[(1+\varepsilon)\cos\Delta\theta - 1]^2 + (1+\varepsilon)^2\sin^2\Delta\theta}{[(1+\varepsilon)\cos\Delta\theta + 1]^2 + (1+\varepsilon)^2\sin^2\Delta\theta} \\ &= \frac{(1+\varepsilon)^2 - 2(1+\varepsilon)\cos\Delta\theta + 1}{(1+\varepsilon)^2 + 2(1+\varepsilon)\cos\Delta\theta + 1}, \end{aligned}$$

I/Q Mismatch Calibration: Phase Mismatch

Let us now apply a single sinusoid to both inputs of the upconverter.



$$
\begin{aligned}
V_{out3}(t) &= V_0(1+\varepsilon)\cos\omega_{in}t\cos(\omega_c t+\Delta\theta) - V_0\cos\omega_{in}t\sin\omega_c t \\
&= V_0\cos\omega_{in}t(1+\varepsilon)\cos\Delta\theta\cos\omega_c t \\
&- V_0\cos\omega_{in}t[(1+\varepsilon)\sin\Delta\theta+1]\sin\omega_c t.
\end{aligned}
$$

It can be shown that the output contains two sidebands of equal amplitudes and carries an average power equal to:

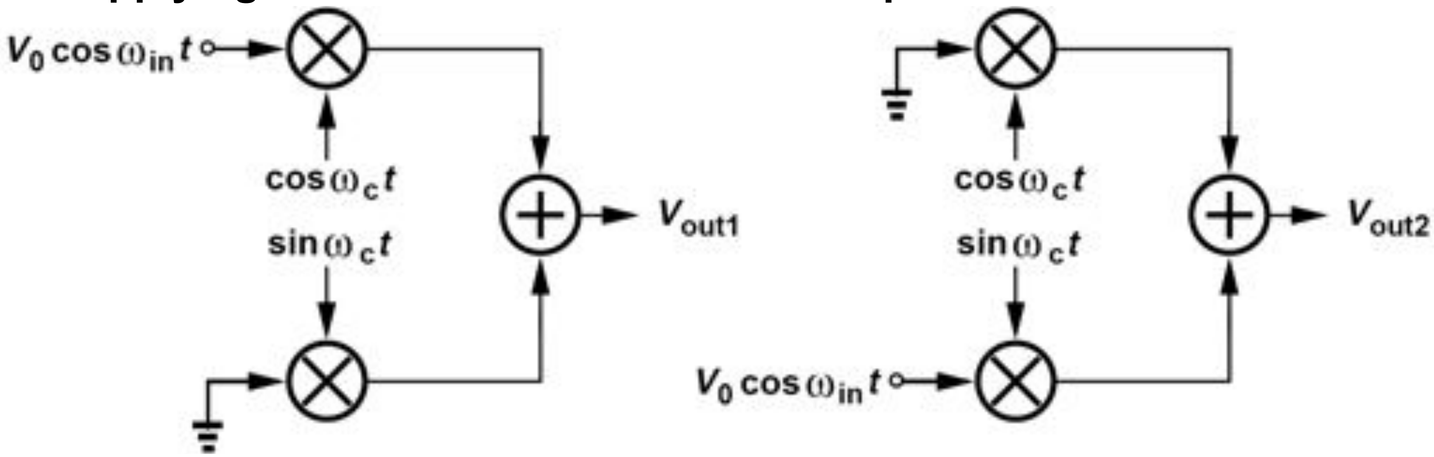
$$\overline{V_{out3}^2(t)} = V_0^2[1+(1+\varepsilon)\sin\Delta\theta].$$

ε is forced to zero as described above, then

$$\overline{V_{out3}^2} - 2\overline{V_{out1}^2} = \sin\Delta\theta$$

I/Q Mismatch Calibration: Gain Mismatch

The tests entail applying a sinusoid to one baseband input while the other is set to zero.



$$V_{out1}(t) = V_0(1+\varepsilon)\cos\omega_{in}t\cos(\omega_c t + \Delta\theta)$$

yielding an average power of $\overline{V_{out1}^2(t)} = \dfrac{V_0^2}{2} + V_0^2\varepsilon$

In figure above (right): $V_{out2}(t) = V_0\cos\omega_{in}t\sin\omega_c t,$

$$\overline{V_{out2}^2(t)} = \frac{V_0^2}{2} \quad \Rightarrow \quad \overline{V_{out1}^2(t)} - \overline{V_{out2}^2(t)} = V_0^2\varepsilon$$

suggesting that the gain mismatch can be adjusted so as to drive this difference to zero.

Carrier Leakage: Definition

The analog baseband circuitry producing the quadrature signals in the transmitter exhibits dc offsets, and so does the baseband port of each upconversion mixer.

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

The upconverter output therefore contains a fraction of the *unmodulated* carrier:

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

Called "carrier leakage," and quantified as:

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

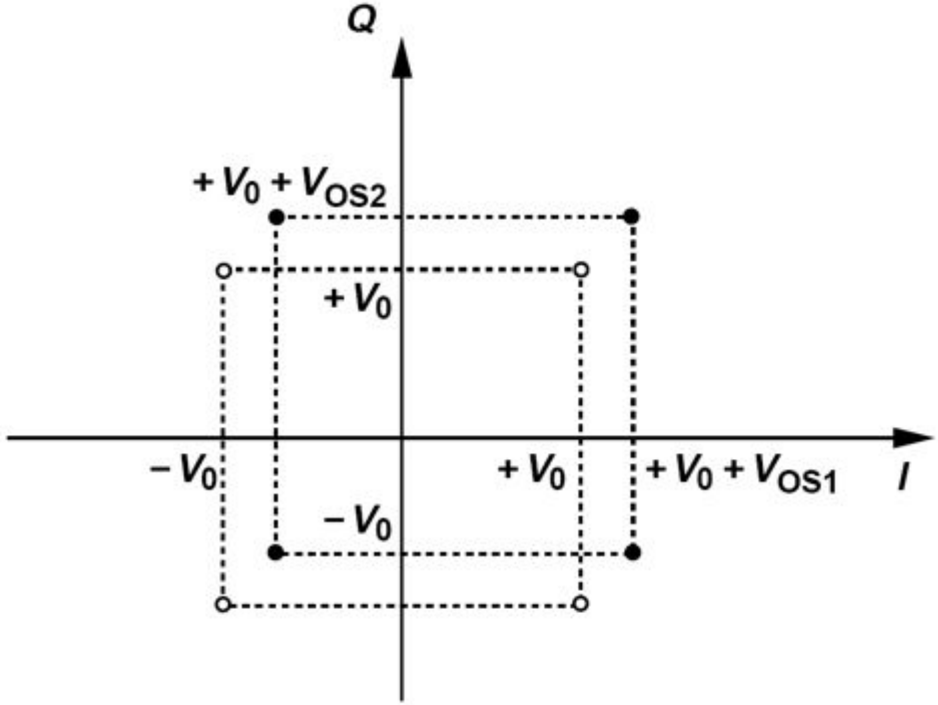
- Carrier Leakage will lead to tow adverse effects: distorting the signal constellation and making it difficult for power control.

Effect of Carrier Leakage(Ⅰ)

- **First, it distorts the signal constellation, raising the error vector magnitude at the TX output.**

For a QPSK signal:

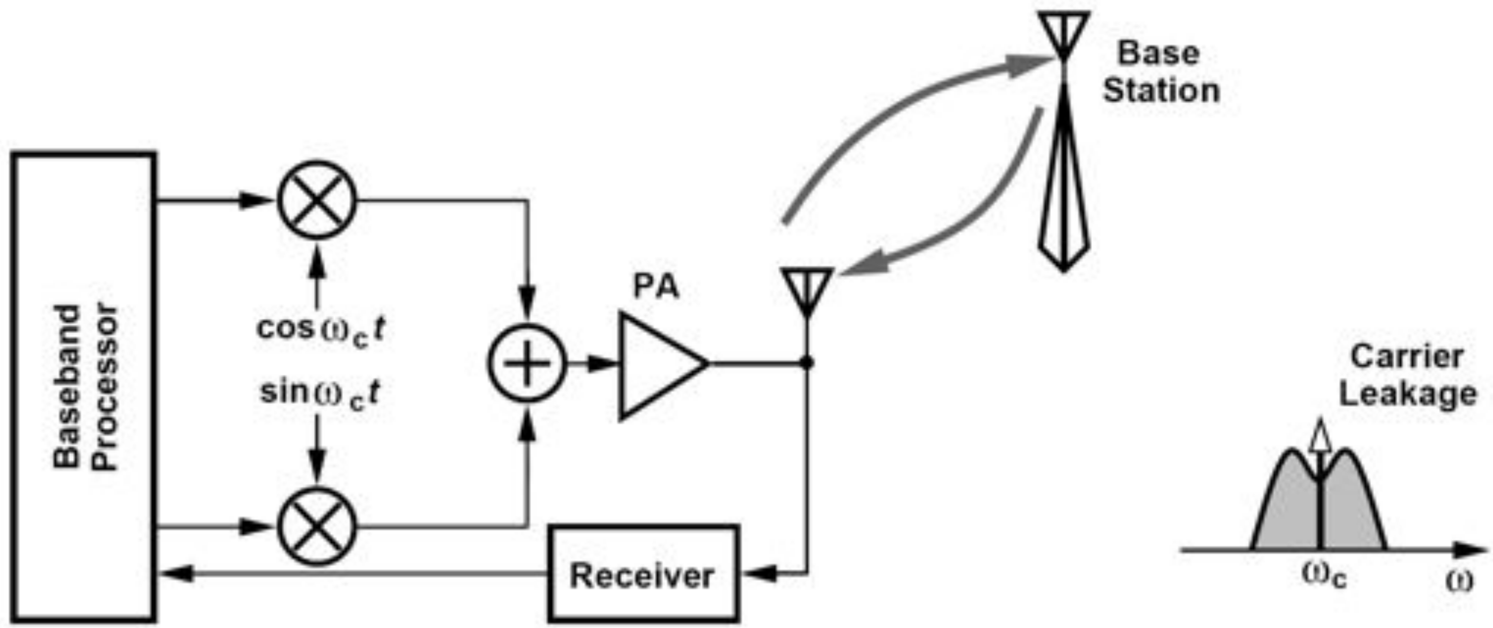
$$V_{out}(t) = \alpha_1(V_0 + V_{OS1})\cos\omega_c t - \alpha_2(V_0 + V_{OS2})\sin\omega_c t$$



The baseband quadrature outputs suffer from dc offsets, i.e., horizontal and vertical shifts in the constellation.

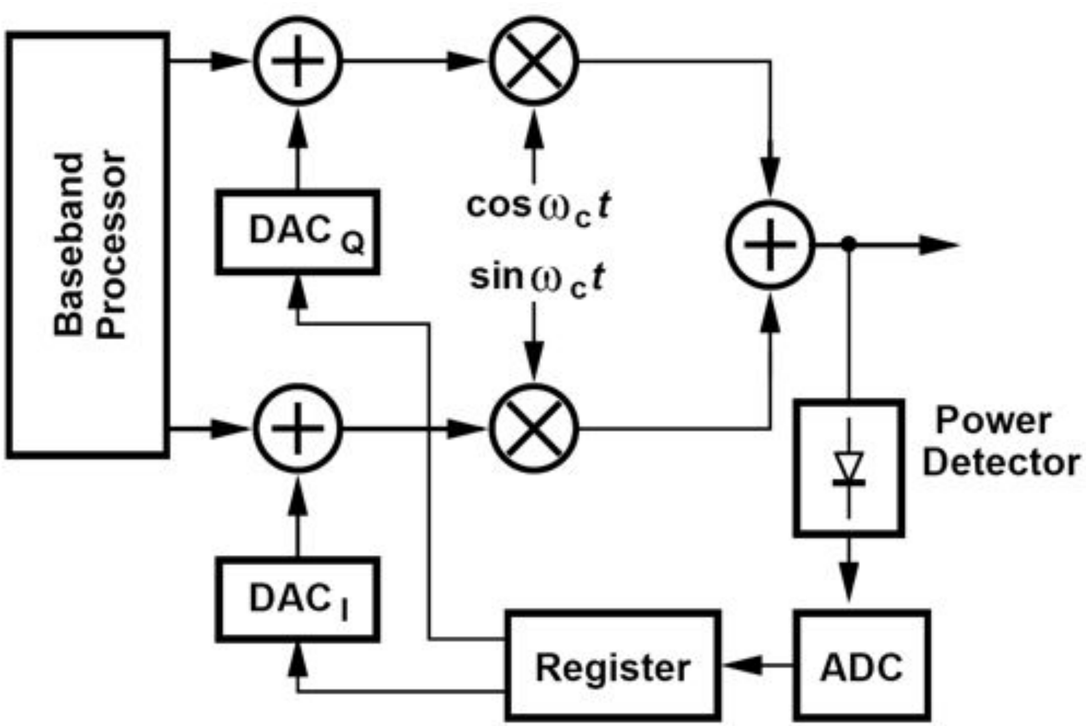
Effect of Carrier Leakage(Ⅱ)

- The second effect manifests itself if the output power of the transmitter must be varied across a wide range by varying the amplitude of the baseband signals.



With a short distance between the base station and the mobile, the carrier power dominates, making it difficult to measure the actual signal power.

Reduction of Carrier Leakage



- The loop consisting of the TX, the detector, and the DACs drives the leakage toward zero, with the final settings of the DACs stored in the register.

Is it possible to cancel the carrier leakage by means of a single DAC?

No, it is not. Previous equation implies that no choice of V_{OS1} or V_{OS2} can force $V_{OS1} \cos \omega_c t - V_{OS2} \sin\omega_c t$ to zero if the other remains finite.

Mixer Linearity

- Excessive nonlinearity in the baseband port of upconversion mixers can corrupt the signal or raise the adjacent channel power

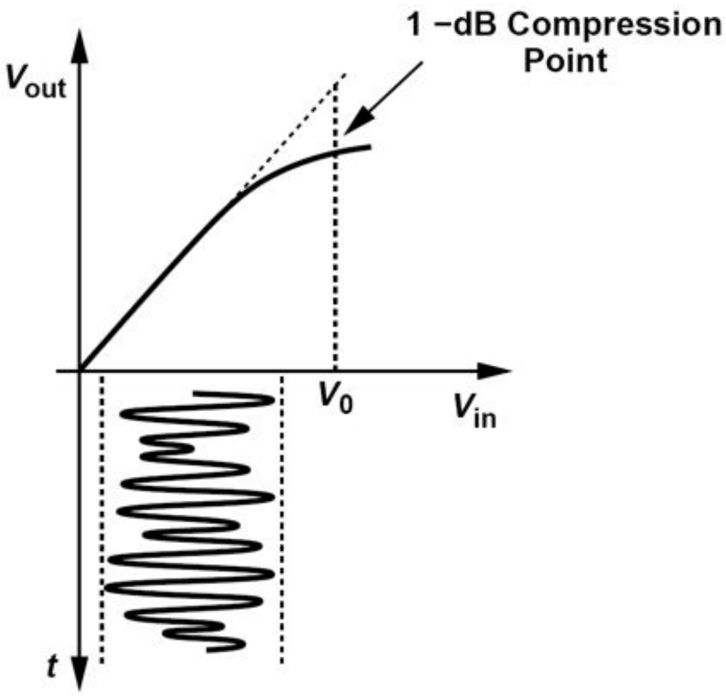
Consider the GMSK signal and suppose the baseband I/Q inputs experience a nonlinearity given by $\alpha_1 x + \alpha_3 x^3$. The upconverted signal assumes the form:

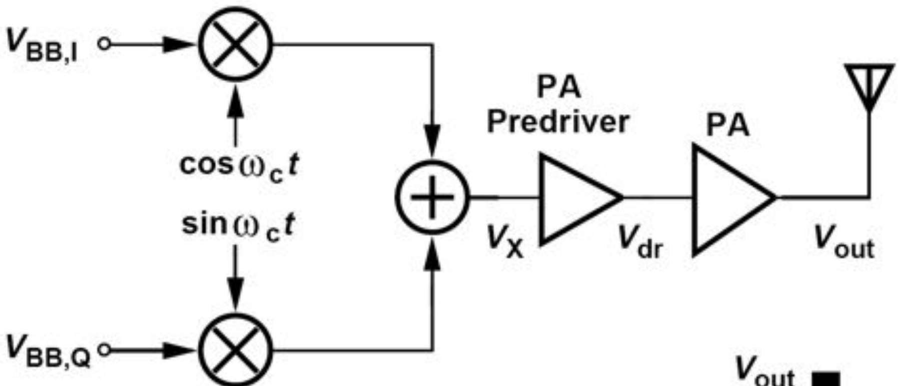
$$\begin{aligned} V_{out}(t) &= (\alpha_1 A \cos\phi + \alpha_3 A^3 \cos^3\phi)\cos\omega_c t - (\alpha_1 A \sin\phi + \alpha_3 A^3 \sin^3\phi)\sin\omega_c t \\ &= \left(\alpha_1 A + \frac{3}{4}\alpha_3 A^3\right)\cos(\omega_c t + \phi) + \frac{\alpha_3 A^3}{4}\cos(\omega_c t - 3\phi). \end{aligned}$$

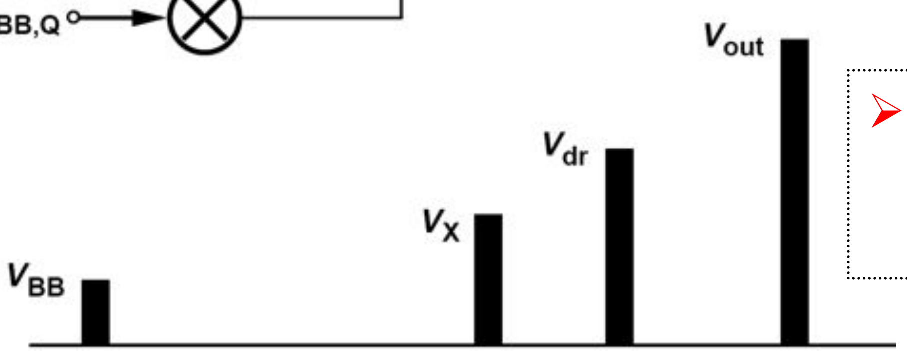
- The second term also represents a GMSK signal but with a threefold modulation index, thereby occupying a larger bandwidth.
- For variable-envelope signals, $A^3(t)$ appears in both terms of equation above, exacerbating the effect.

TX Linearity

- The distortion of a variable-envelope signal is typically characterized by the compression that it experiences.







- We must *maximize* the gain of the PA and *minimize* the output swing of the predriver and the stages preceding it.

Example of TX Linearity

If the predriver and the PA exhibit third-order characteristics, compute the 1-dB compression point of the cascade of the two.

Solution:

Assuming a nonlinearity of $\alpha_1 x+\alpha_3 x^3$ for the predriver and $\alpha_1 x+\alpha_3 x^3$ for the PA, we write the PA output as

$$y(t) = \beta_1(\alpha_1 x + \alpha_3 x^3) + \beta_3(\alpha_1 x + \alpha_3 x^3)^3$$
$$= \beta_1\alpha_1 x + (\beta_1\alpha_3 + \beta_3\alpha_1^3)x^3 + \cdots$$

then the input 1-dB compression point is given by:

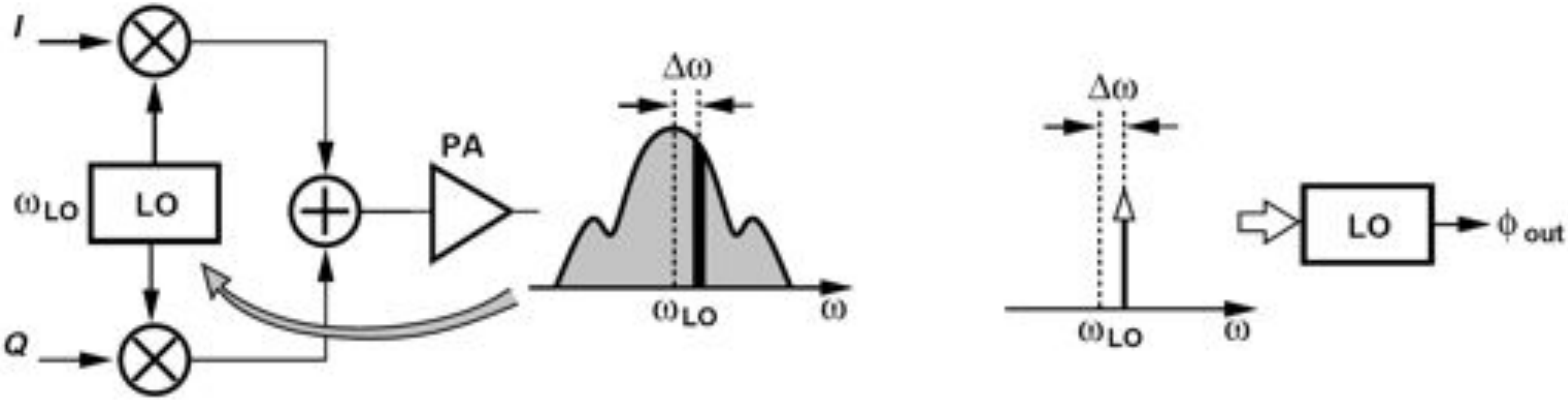
$$A_{1dB,in} = \sqrt{0.145\left|\frac{\beta_1\alpha_1}{\beta_1\alpha_3 + \beta_3\alpha_1^3}\right|}.$$

In transmitters, the output power is of interest, suggesting that the compression behavior must also be quantified at the output. :

$$A_{1dB,out} = A_{1dB,in} \times |\alpha_1\beta_1| \times \frac{1}{1.12}$$

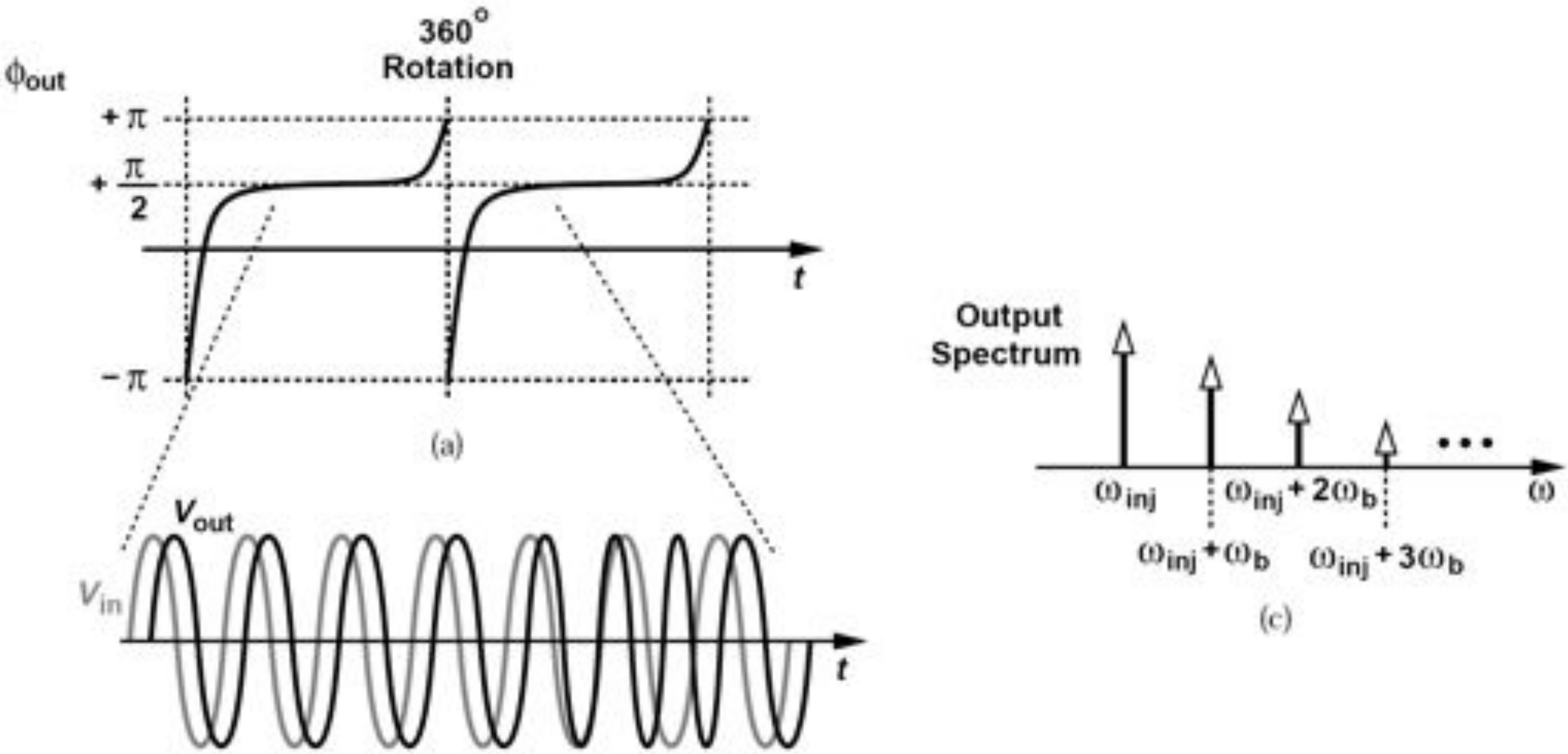
$$\Rightarrow A_{1dB,out} = \frac{0.34|\alpha_1\beta_1|\sqrt{|\beta_1\alpha_1|}}{\sqrt{\beta_1\alpha_3 + \beta_3\alpha_1^3}}$$

Oscillator Pulling



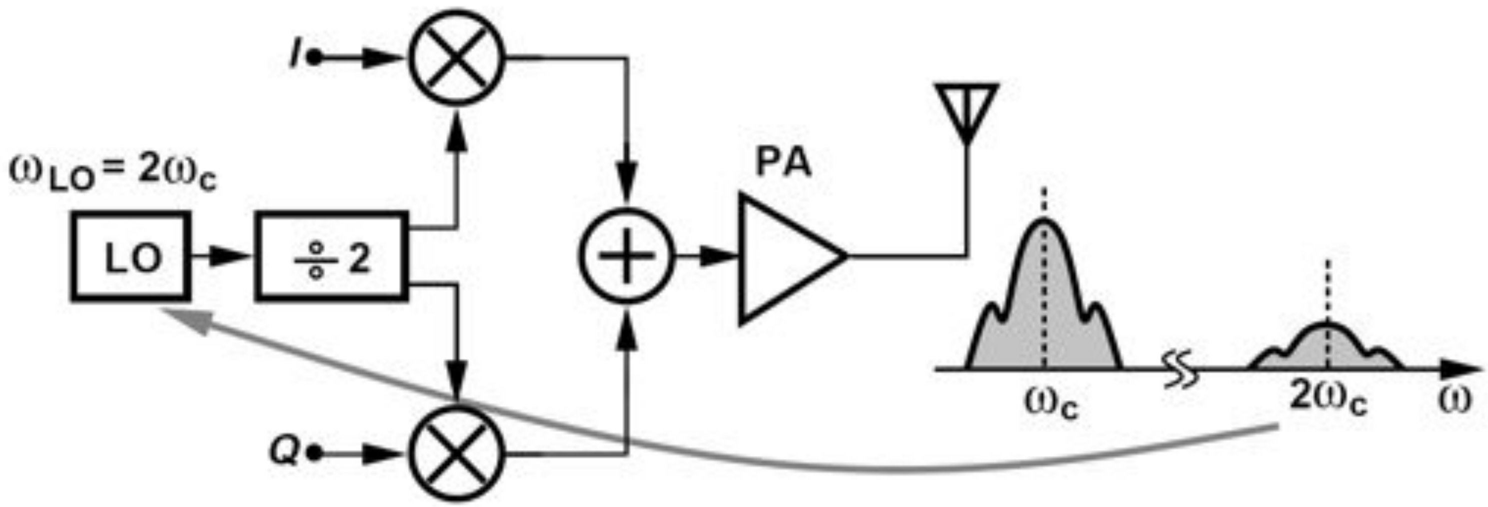
- The PA output exhibits very large swings, which couple to various parts of the system through the silicon substrate, package parasitics, and traces on the printed-circuit board. Thus, it is likely that an appreciable fraction of the PA output couples to the local oscillator.

Effect of Oscillator Pulling



- The output phase of the oscillator, Φ_{out}, is modulated periodically.
- In order to avoid injection pulling, the PA output frequency and the oscillator frequency must be made sufficiently different

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

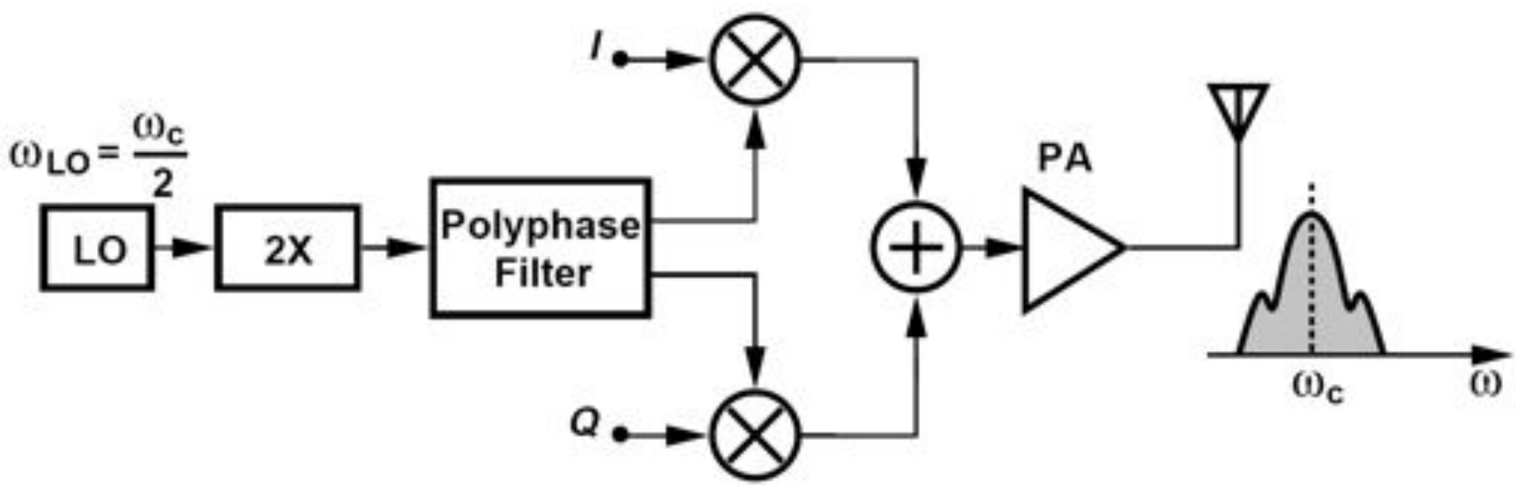
  (2) the divider readily provides quadrature phases of the carrier

Is it Possible to Use a Frequency Doubler?

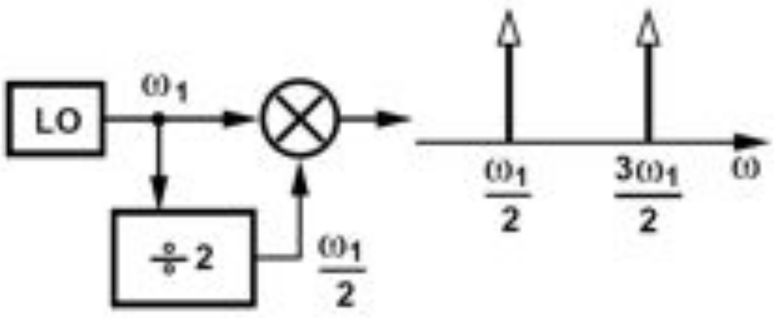
Is it possible to choose $\omega_{LO} = \omega_c/2$ and use a frequency doubler to generate ω_c?

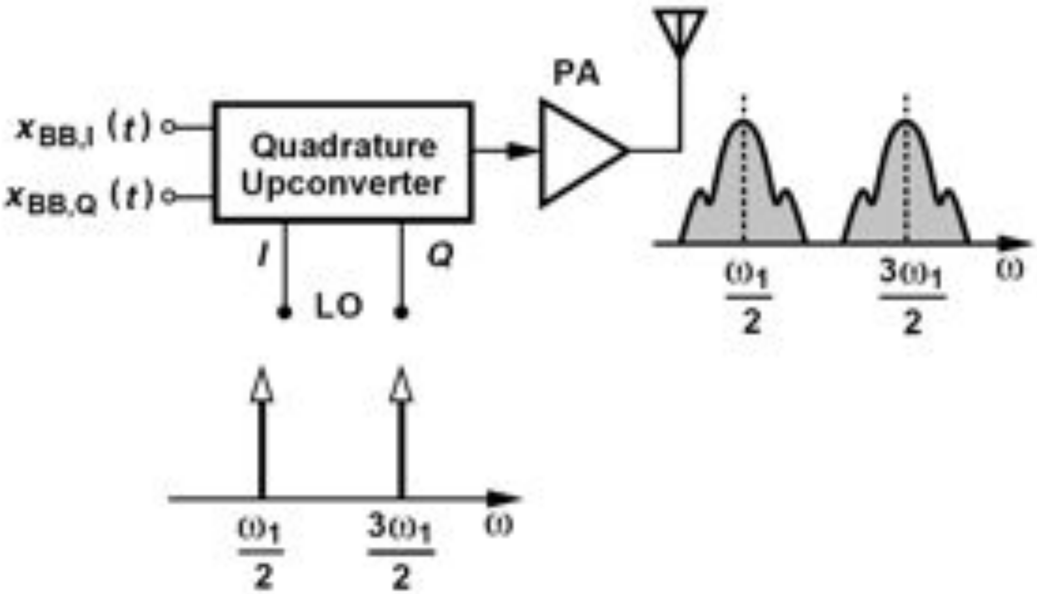
Solution:

It is possible, but the doubler typically does not provide quadrature phases, necessitating additional quadrature generation stages. Figure below shows an example where the doubler output is applied to a polyphase filter. The advantage of this architecture is that no harmonic of the PA output can pull the LO. The serious disadvantage is that the doubler and the polyphase filter suffer from a high loss, requiring the use of power-hungry buffers.



Use Mixing to Derive Frequencies





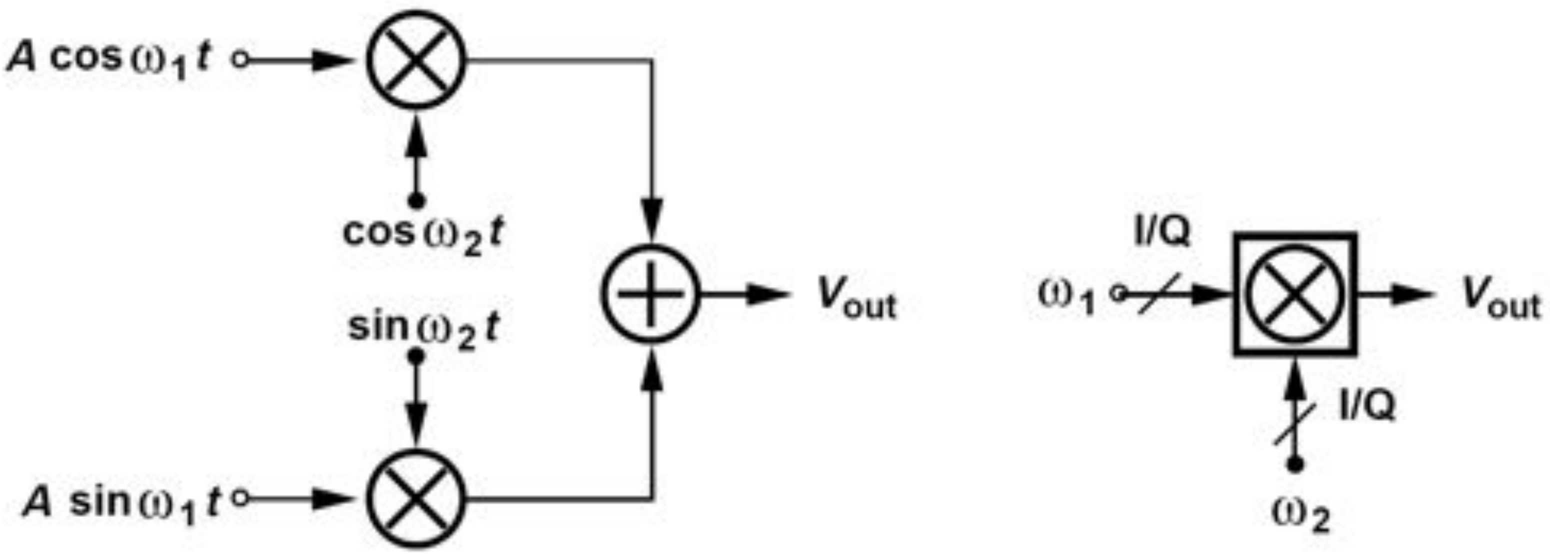
- The oscillator frequency is divided by 2 and the two outputs are mixed. The result contains components at ω1 ± ω1/2 with equal magnitudes.
- Can both components be retained?

  (1) half of the power delivered to the antenna is wasted.

  (2) the power transmitted at the unwanted carrier frequency corrupts communication in other channels or bands.

Single-Sideband Mixing

An alternative method of suppressing the unwanted sideband incorporates "single-sideband" (SSB)mixing.



➢ **Gain and phase mismatches lead to an unwanted sideband**

Corruption from Harmonics of the Input (I)

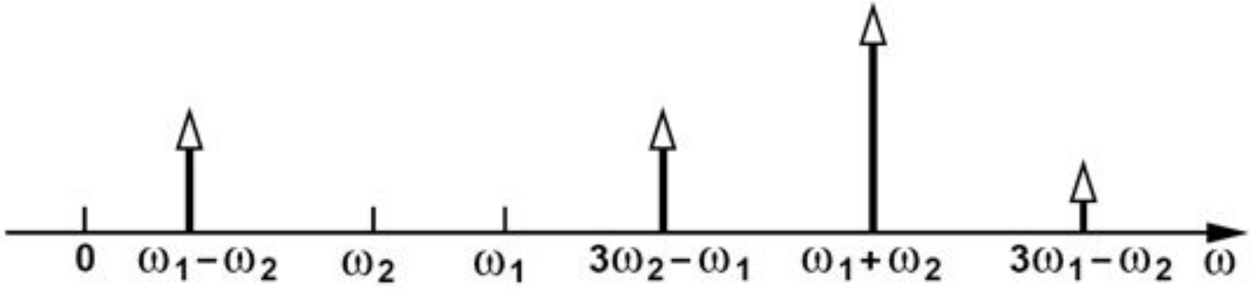
Suppose each mixer exhibits third-order nonlinearity If the nonlinearity is of the form $\alpha_1 x + \alpha_3 x^3$,

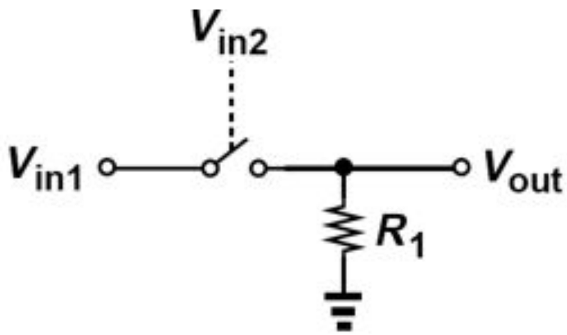
$$
\begin{aligned}
V_{out}(t) &= (\alpha_1 A \cos\omega_1 t + \alpha_3 A^3 \cos^3\omega_1 t)\cos\omega_2 t \\
&- (\alpha_1 A \sin\omega_1 t + \alpha_3 A^3 \sin^3\omega_1 t)\sin\omega_2 t \\
&= \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\cos\omega_1 t\cos\omega_2 t - \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\sin\omega_1 t\sin\omega_2 t \\
&+ \frac{\alpha_3 A^3}{4}\cos 3\omega_1 t\cos\omega_2 t + \frac{\alpha_3 A^3}{4}\sin 3\omega_1 t\sin\omega_2 t \\
&= \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\cos(\omega_1 + \omega_2)t + \frac{\alpha_3 A^3}{4}\cos(3\omega_1 - \omega_2)t.
\end{aligned}
$$

- **The output spectrum contains a spur at $3\omega_1 - \omega_2$. Similarly, with third-order nonlinearity in the mixer ports sensing sin $\omega_2 t$ and cos $\omega_2 t$, a component at $3\omega_2 - \omega_1$ arises at the output.**

Corruption from Harmonics of the Input (Ⅱ)

The overall output spectrum:

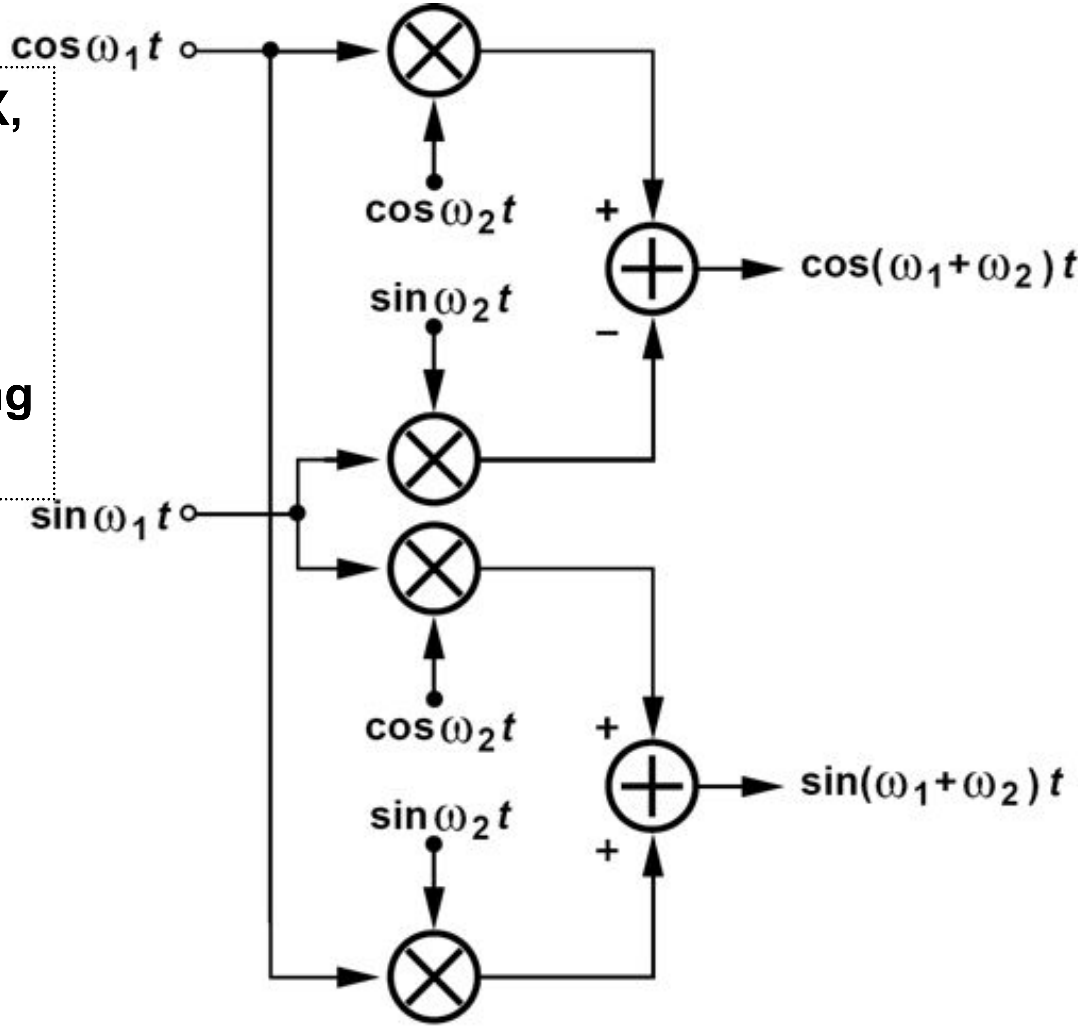




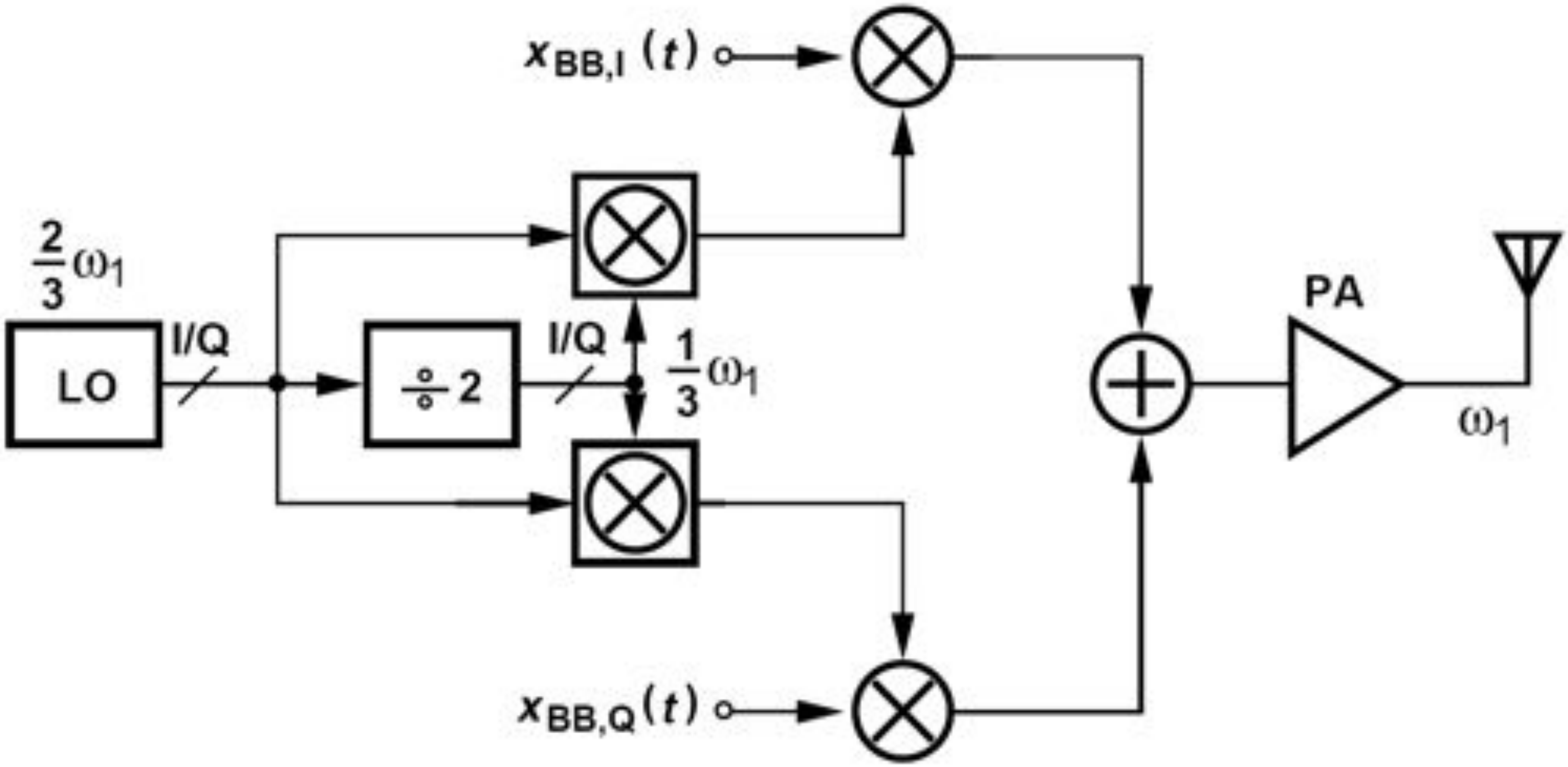
- **Between the two spurs at $3\omega_1 - \omega_2$ and $3\omega_2 - \omega_1$, only one can be reduced to acceptably low levels while the other remains only 10 dB (a factor of one-third) below the desired component.**

SSB Mixer Providing Quadrature Outputs

- For use in a direct-conversion TX, the SSB mixer must provide the quadrature phases of the carrier. This is accomplished by noting that $\sin \omega_1 t \cos \omega_2 t + \cos \omega_1 t \sin \omega_2 t = \sin(\omega_1 + \omega_2)t$ and duplicating the SSB mixer as shown right.



Direct-Conversion TX Using SSB Mixing in LO Path



- Since the carrier and LO frequencies are sufficiently different, this architecture remains free from injection pulling.

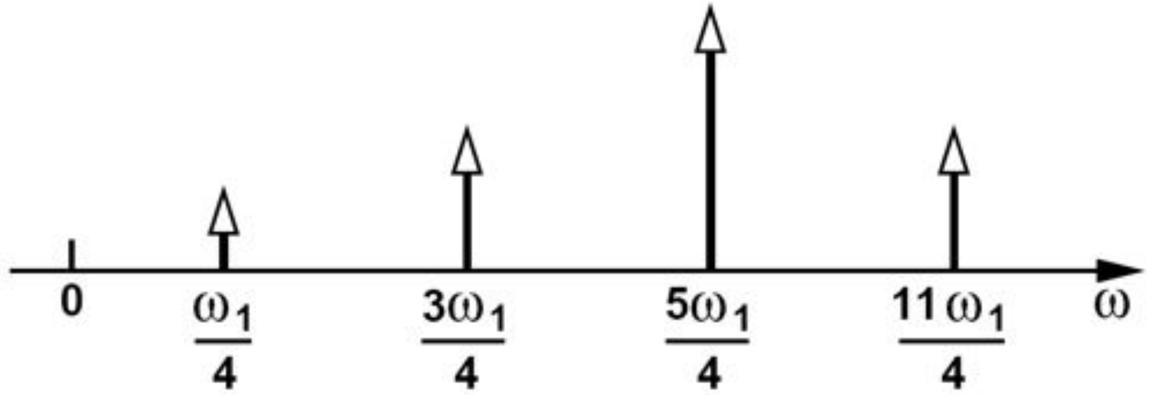
Example Using ÷4 Circuit instead of ÷2 Circuit

A student replaces the ÷ 2 circuit in figure above with a ÷ 4 topology. Analyze the unwanted components in the carrier.

Solution:

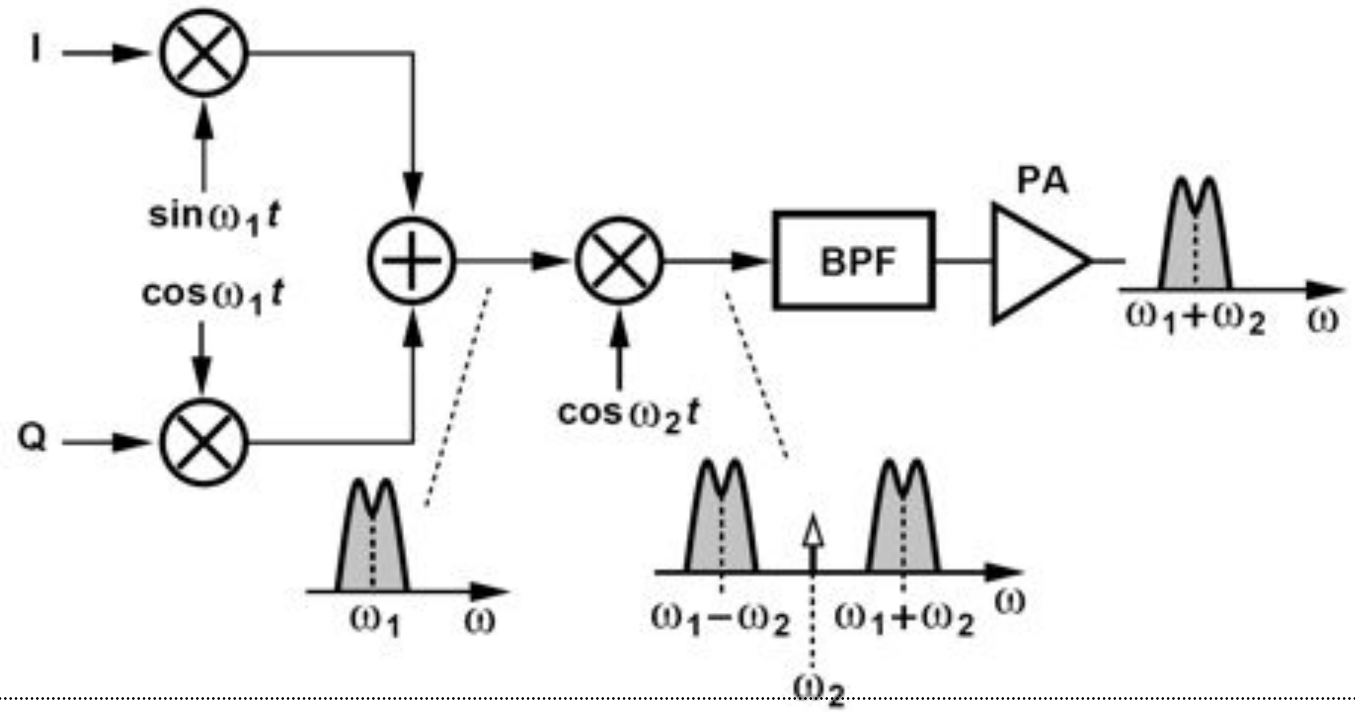
Upon mixing ω_1 and $\omega_1/4$, the SSB mixer generates $5\omega_1/4$ and, due to mismatches, $3\omega_1/4$. In the previous case, these values were given by $3\omega_1/2$ and $\omega_1/2$, respectively. Thus, filtering the unwanted sideband is more difficult in this case because it is closer to the wanted sideband.

As for the effect of harmonics, the output contains spurs at $3\omega_1-\omega_2$ and $3\omega_2-\omega_1$, which are respectively equal to $11\omega_1/4$ and $\omega_1/4$ if $\omega_2 = \omega_1/4$. The spur at $11\omega_1/4$ remains slightly higher than its counterpart in the previous case ($5\omega_1/2$), while that at $\omega_1/4$ is substantially lower and can be filtered more easily. Figure below summarizes the output components.



Heterodyne Transmitters

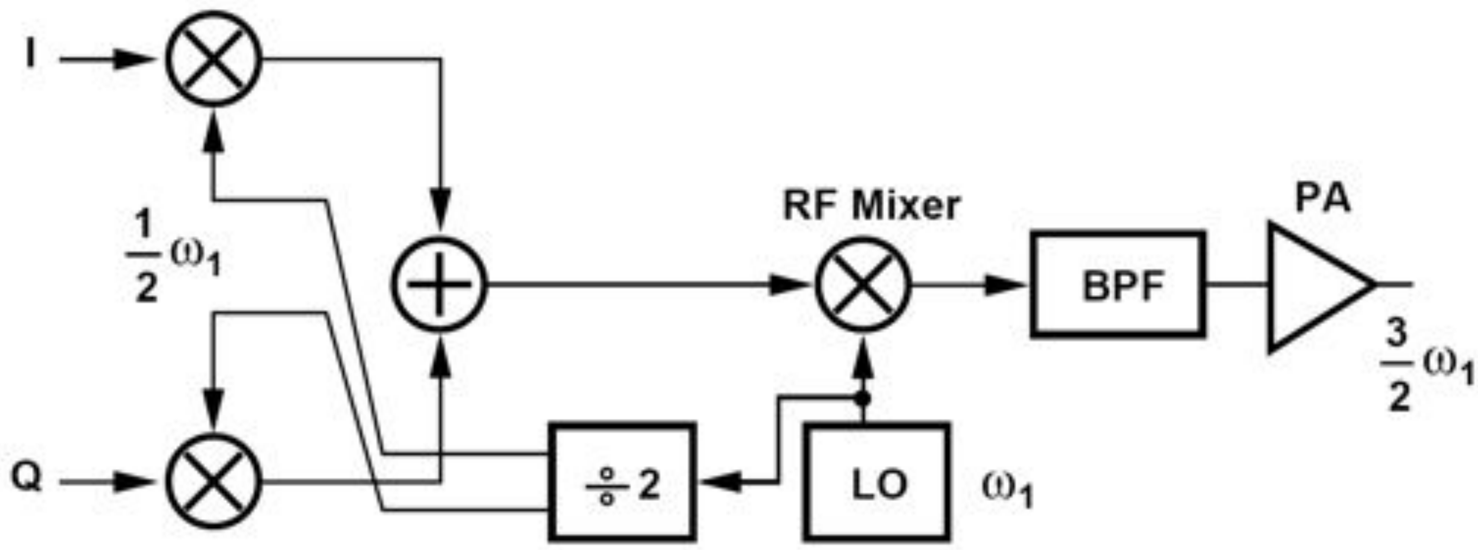
Perform the signal upconversion in two steps so that the LO frequency remains far from the PA output spectrum



- **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.**

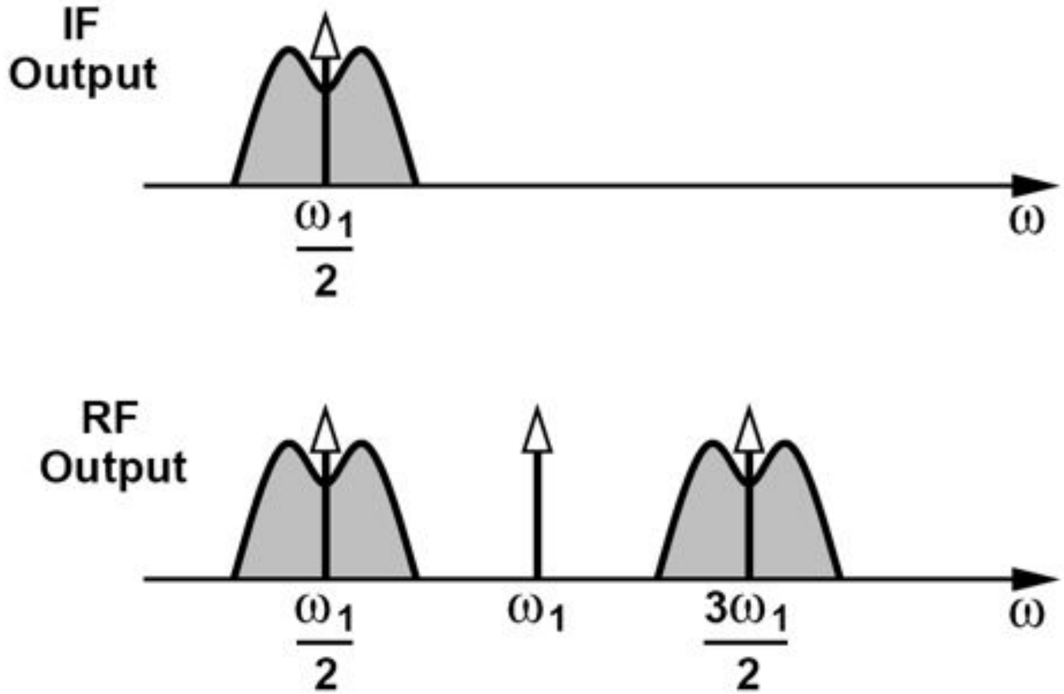
Sliding-IF TX

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.**

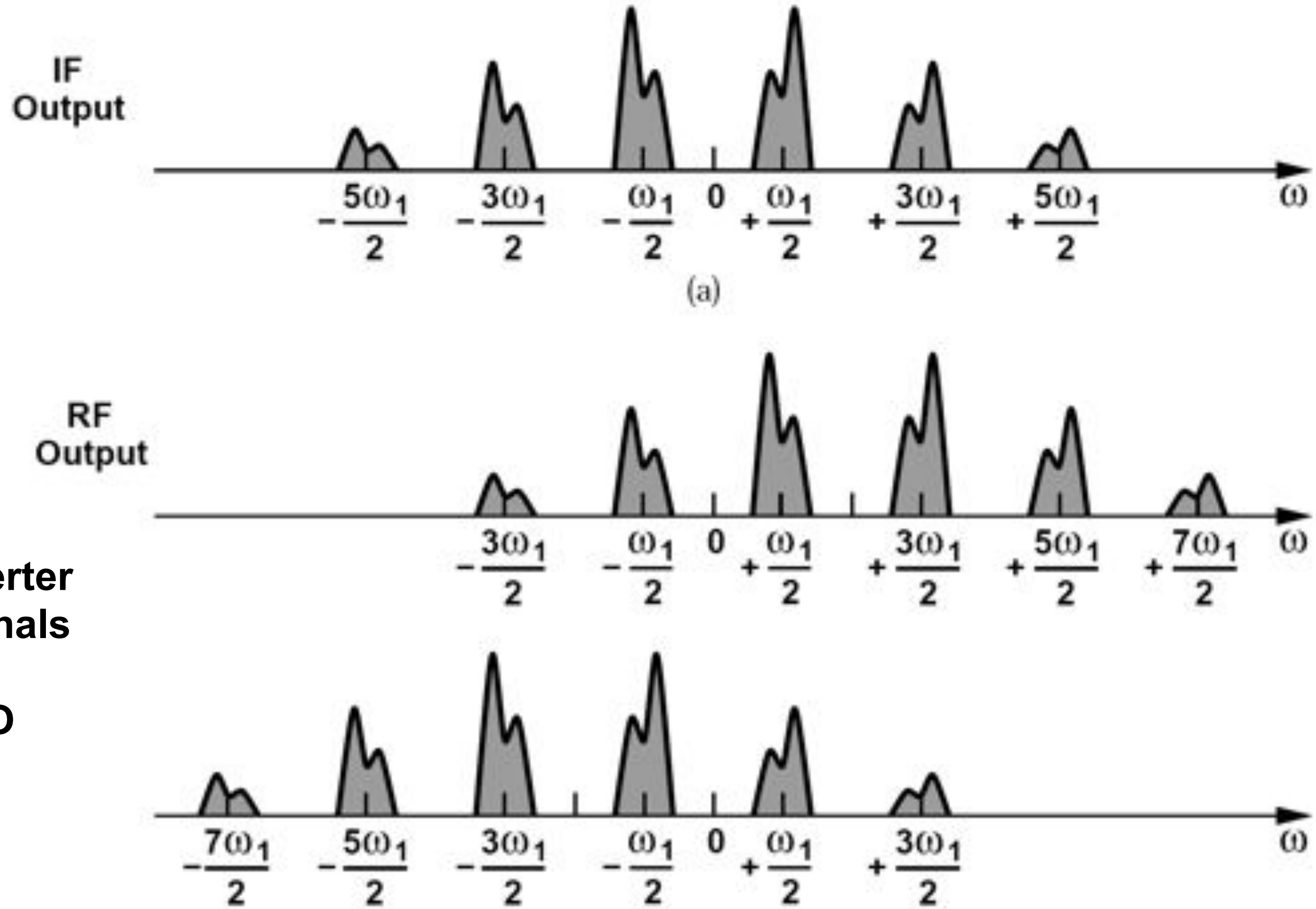
Carrier Leakage



- The dc offsets in the baseband yield a component at $\omega_1/2$ at the output of the quadrature upconverter, and the dc offset at the input of the RF mixer produces another component at ω_1
- The former can be minimized as described before. The latter, and the lower sideband at $\omega_1/2$, must be removed by filtering

Mixing Spurs: the Harmonics of the First LO

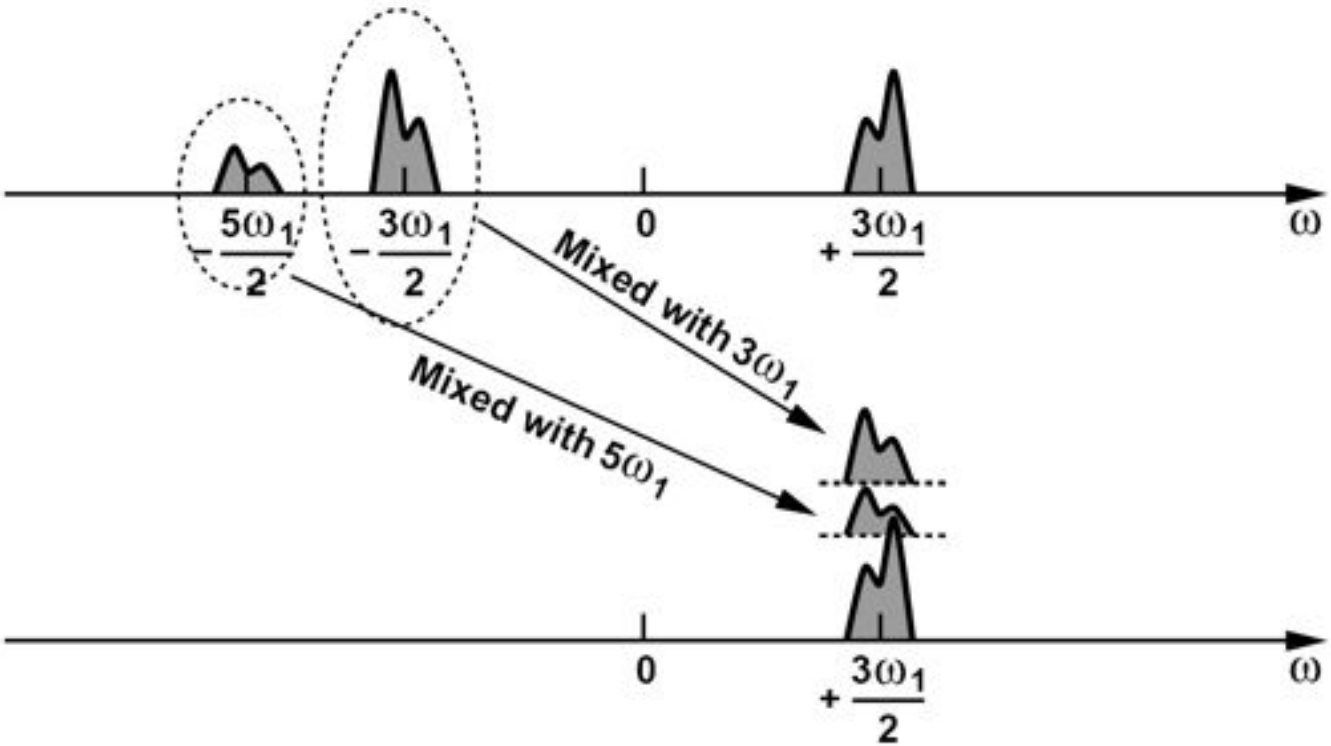
- The spurs arise from two mechanisms: the harmonics of the first LO and the harmonics of the second LO.



The quadrature upconverter mixes the baseband signals with the third and fifth harmonics of the first LO

Mixing Spurs: the Harmonics of the Second LO

- The second mechanism relates to the harmonics of the second LO. That is, the spectrum shown in figure above is mixed with not only ω_1 but $3\omega_1$, $5\omega_1$, etc.

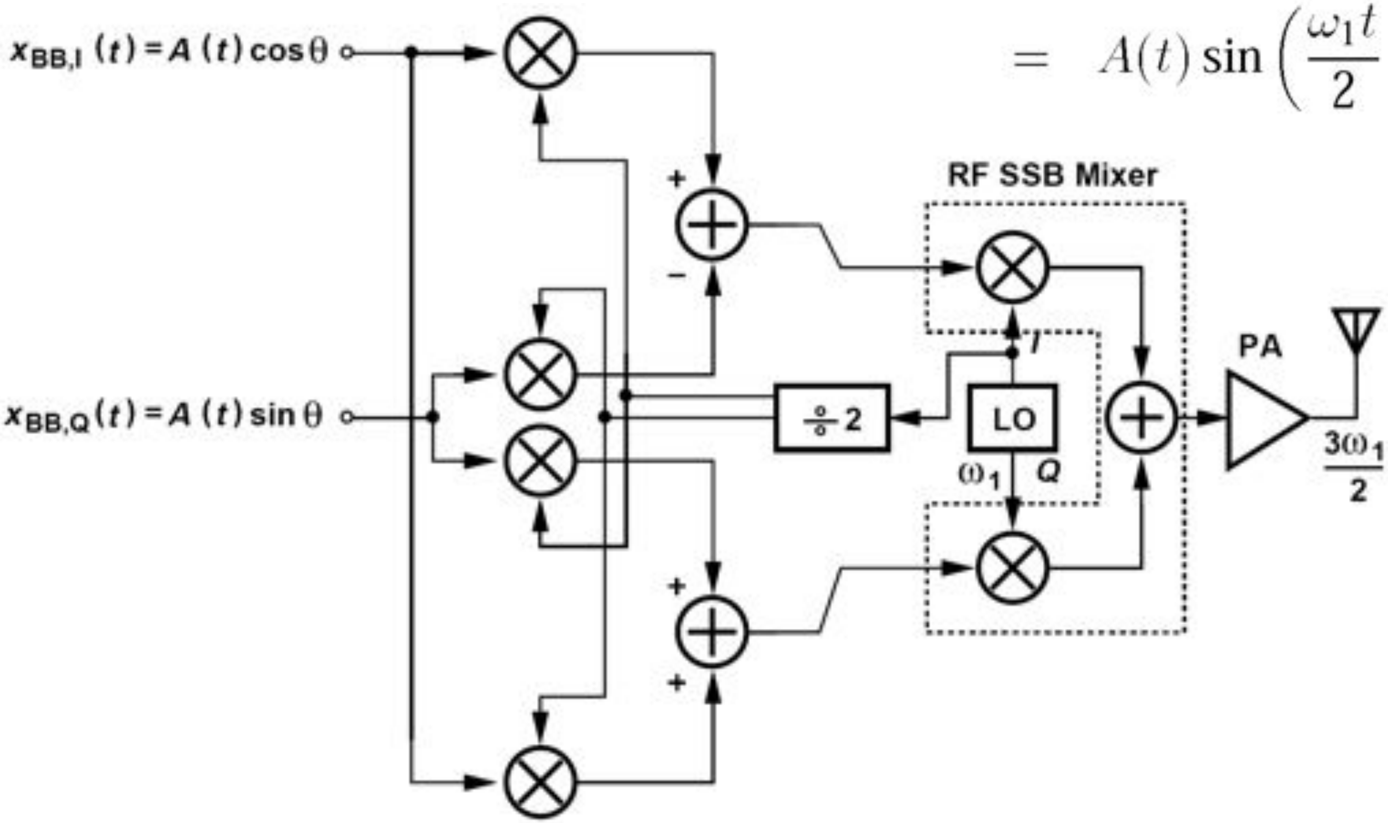


- Upon mixing with $+3\omega_1$, the IF sideband at $-3\omega_1/2$ is translated to $+3\omega_1/2$, thereby corrupting the wanted sideband

Use of SSB Mixing to Suppress the Unwanted Sideband

Two quadrature upconverters provide the quadrature components of the IF signal:

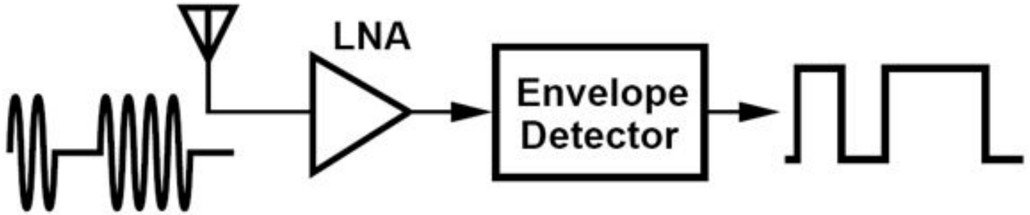
$$\begin{aligned} x_{IF,I}(t) &= A(t)\cos\theta\cos\frac{\omega_1 t}{2} - A(t)\sin\theta\sin\frac{\omega_1 t}{2} \\ &= A(t)\cos\left(\frac{\omega_1 t}{2}+\theta\right) \\ x_{IF,Q}(t) &= A(t)\cos\theta\sin\frac{\omega_1 t}{2} + A(t)\sin\theta\cos\frac{\omega_1 t}{2} \\ &= A(t)\sin\left(\frac{\omega_1 t}{2}+\theta\right). \end{aligned}$$



Other TX Architectures: OOK Transceivers

- "On-off keying" (OOK) modulation is a special case of ASK where the carrier amplitude is switched between zero and maximum.

- When LO is directly turned on and off by the binary baseband data (figure above left), If the LO swings are large enough, the PA also experiences relatively complete switching and delivers an OOK waveform to the antenna. figure above (right) can avoid the issue that LO cannot be easily controlled by a PLL.



- An LNA followed by an envelope detector can recover the binary data.