

Radio-Frequency IC Design

Lecture 13: Receivers – II

ELEC 404

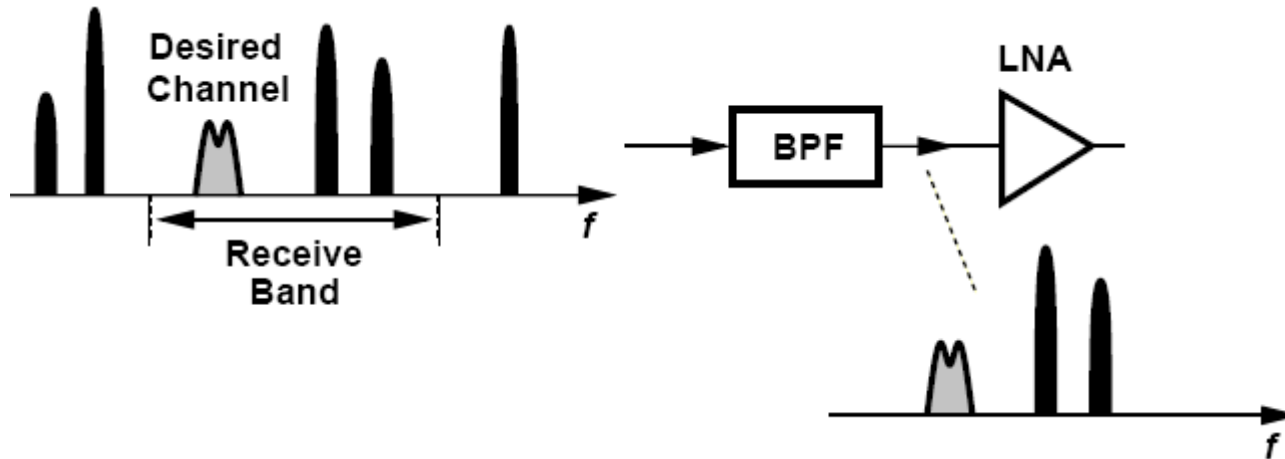
Acknowledgement: B. Razavi's RF Microelectronics



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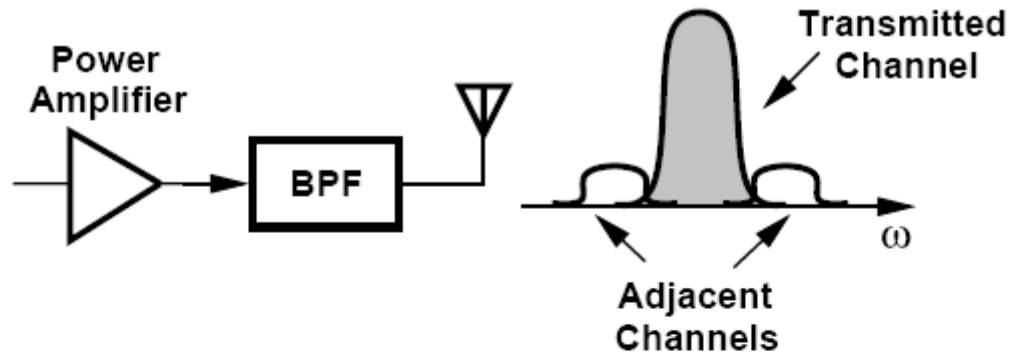
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Band Selection and Channel Selection

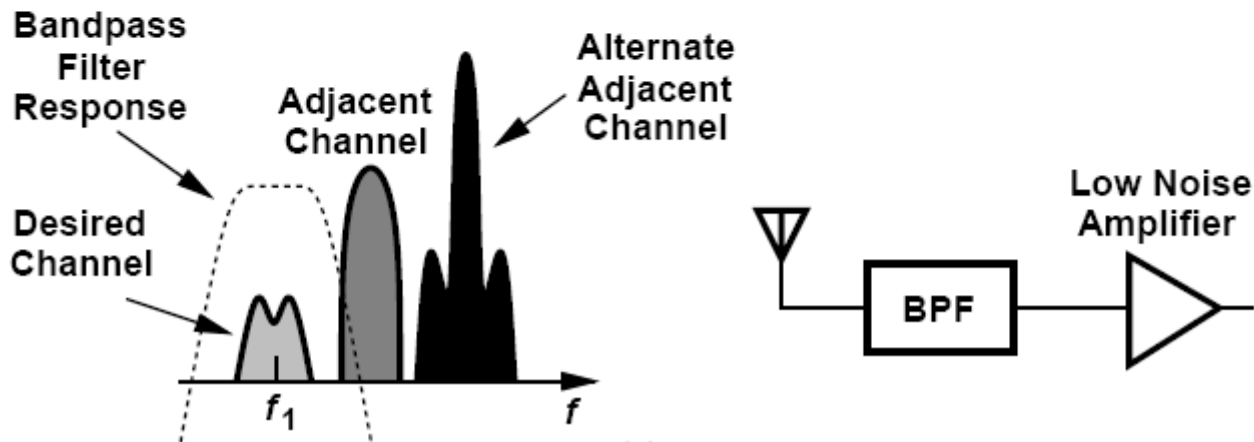


- Most RXFEs incorporate a “band-select” filter (BPF) to reject out-of-band interferers
- Channels have narrow bandwidth, set by the standard

Narrow Channel Bandwidth

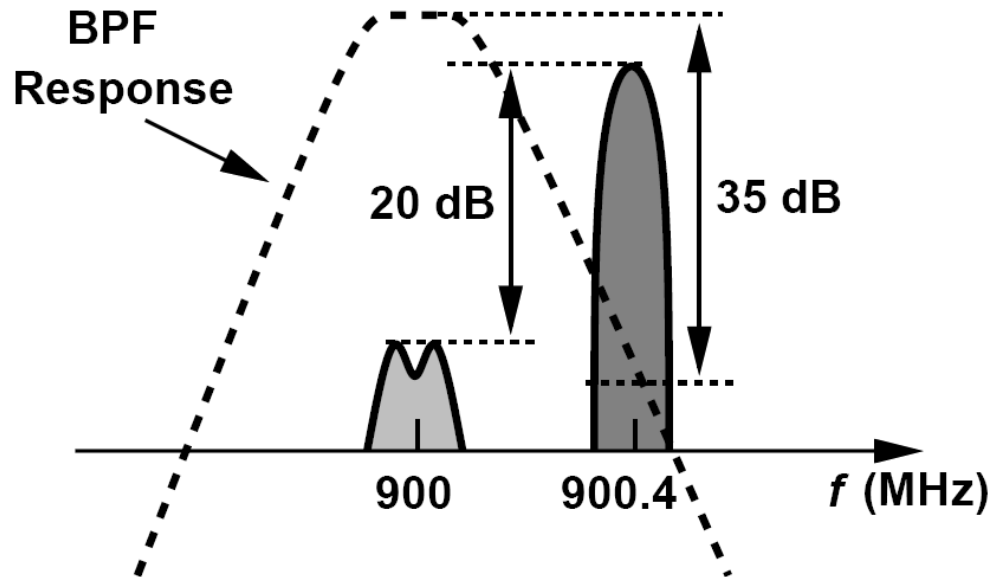


(a)



- TX must employ narrowband modulation and amplification to avoid leakage to adjacent channels
- RX must process desired channel while sufficient rejecting in-band and out-of-band interferers

Channel Selection & Band-Selection?



- Channel selection \rightarrow reject in-band interferers
- Filter needs a variable, yet precise center frequency
- Filter must provide a very high Q
- Channel selection deferred to later stages where center frequency is lower \rightarrow required Q is more reasonable
- All of the stages in the RX chain that precede channel-selection filtering must be sufficiently linear

Filtering In-Band Interferer?

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

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

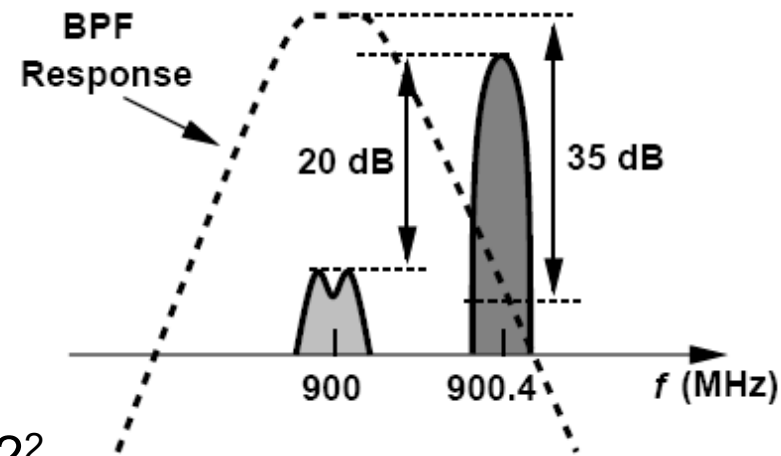
$$35\text{dB} = 56.2, \omega = 900.4\text{MHz}$$

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

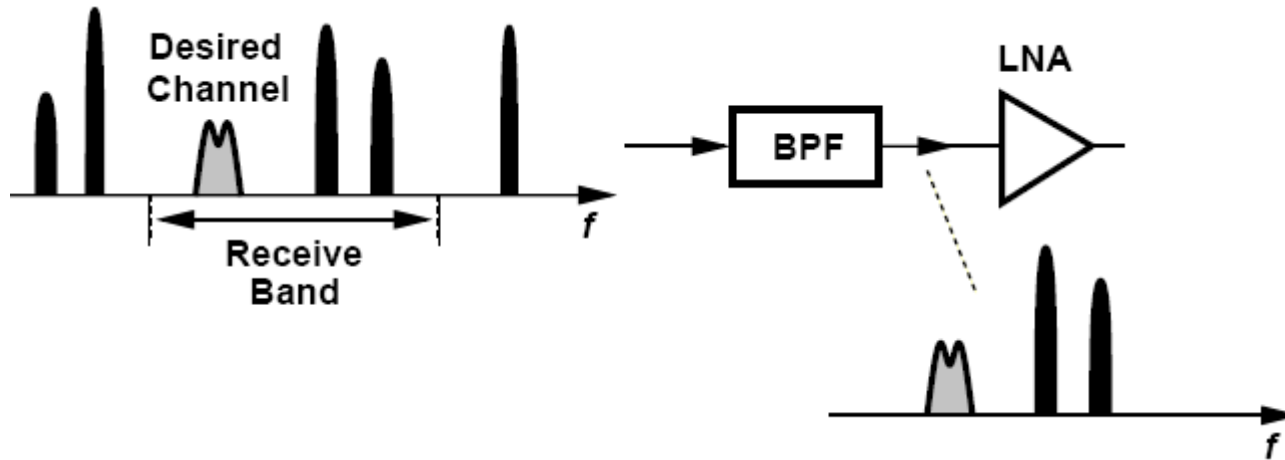
$$|Z_T(j\omega)|^2 = \frac{L^2\omega^2}{(1 - LC\omega^2)^2 + L^2\omega^2/R^2} = R^2/56.2^2$$

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

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

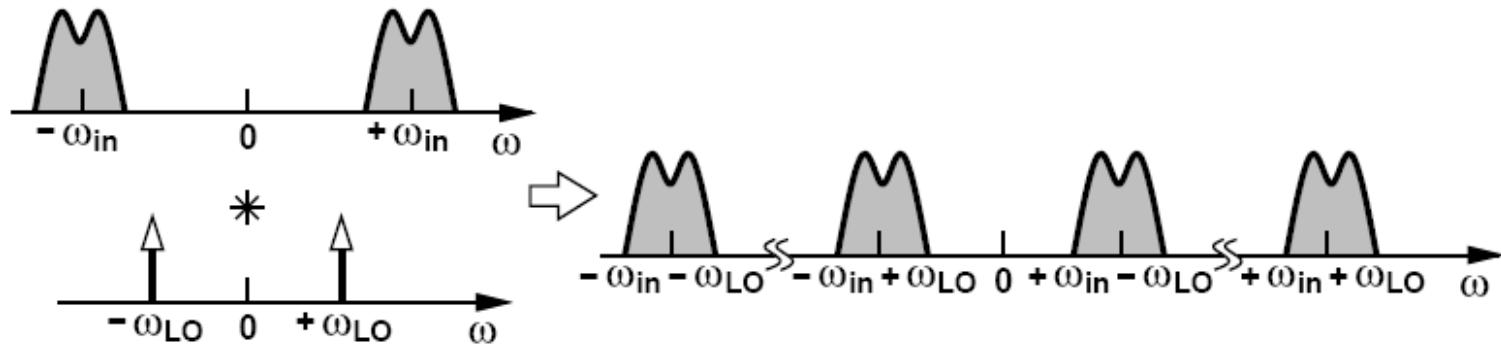
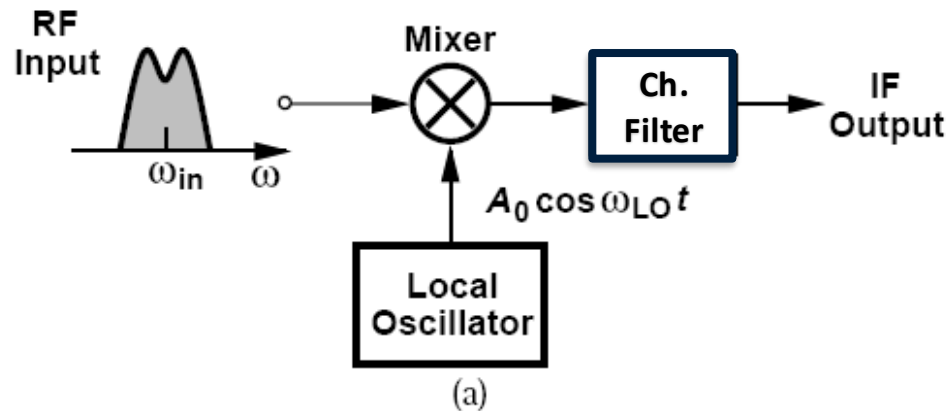


Selectivity vs. In-band Loss Tradeoff



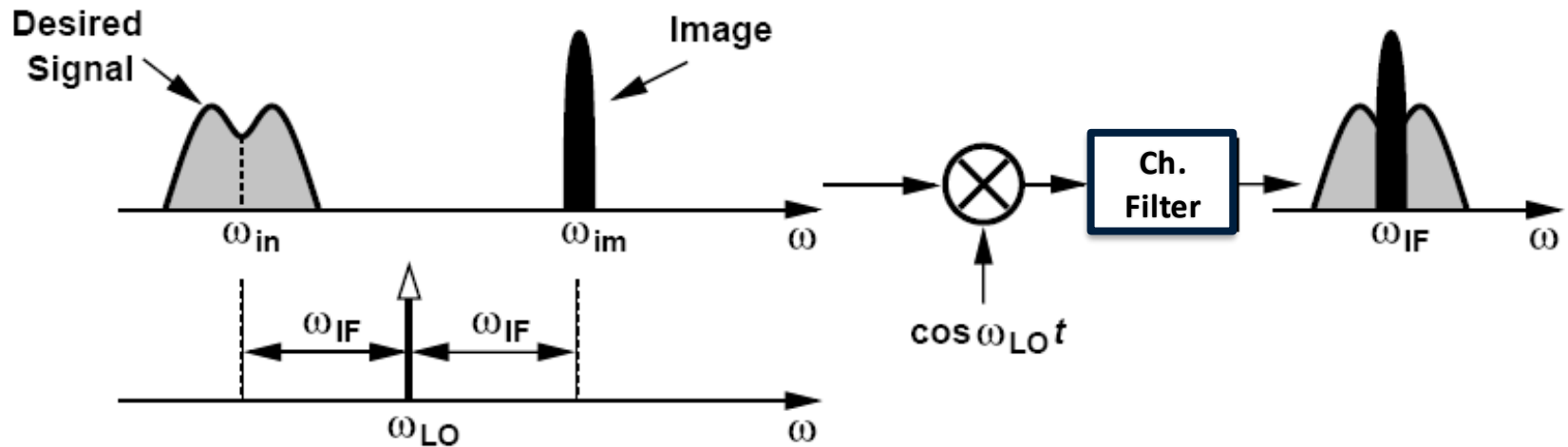
- The edges of the band-pass frequency response can be sharpened only by increasing the order of the filter i.e., number of cascaded sections within the filter.
- Front-end loss directly raises the NF of the entire receiver
- Typical loss = 0.5-1dB

Basic Heterodyne Receivers



- Nonzero IF (Intermediate Frequency) $\rightarrow \omega_{LO} \neq \omega_{in}$
- Due to its high noise, the downconversion mixer is preceded by an LNA

Basic Heterodyne RX: Image Aliasing



$$\begin{aligned}
 A \cos \omega_{IF} t &= A \cos(\omega_{in} - \omega_{LO}) t \\
 &= A \cos(\omega_{LO} - \omega_{in}) t
 \end{aligned}$$

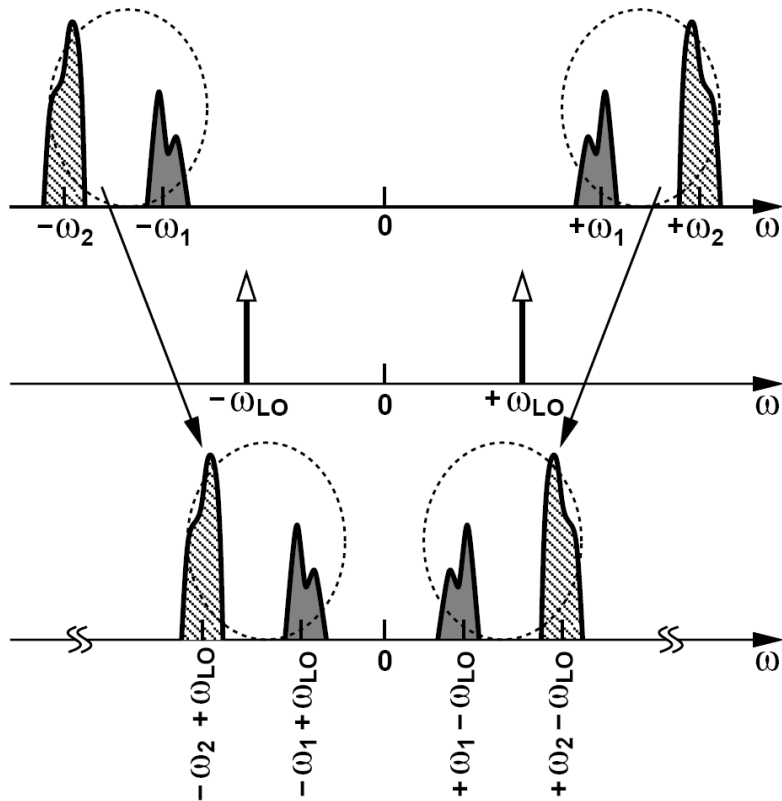
$$\omega_{im} = \omega_{in} + 2\omega_{IF} = 2\omega_{LO} - \omega_{in}$$

- Two spectra located symmetrically around ω_{LO} downconverted to IF
- Image cannot be removed through filtering at the IF output
- Image rejection ratio: $\text{dB}(P_{IF \text{ desired}}/P_{IF \text{ image}})$

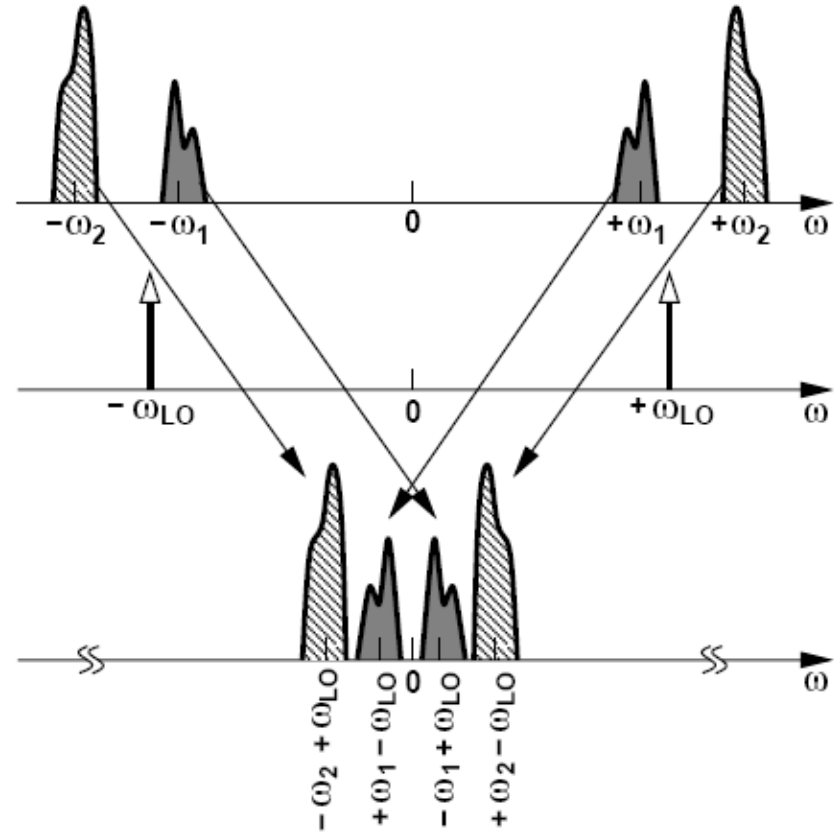
Downconverted Spectrum with LO – I

- Two received channels at ω_1 and ω_2 with $\omega_1 < \omega_2$.
- LO varies from below ω_1 to above ω_2 .

$\omega_{LO} < \omega_1$ (low-side injection)

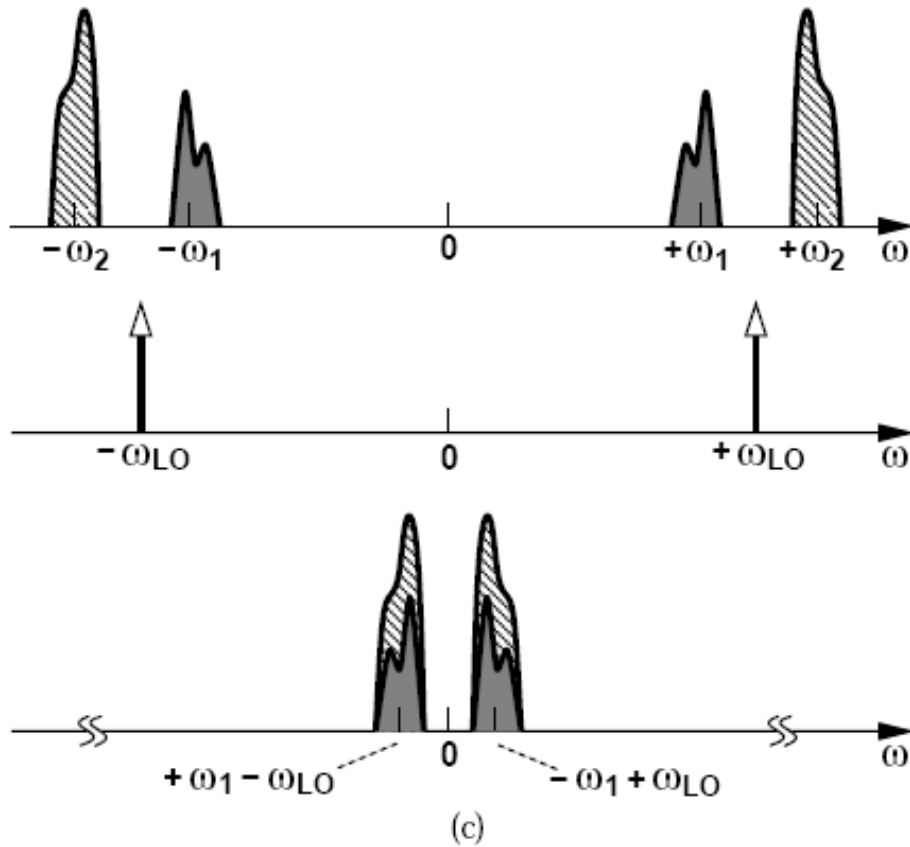


ω_{LO} slightly $> \omega_1$

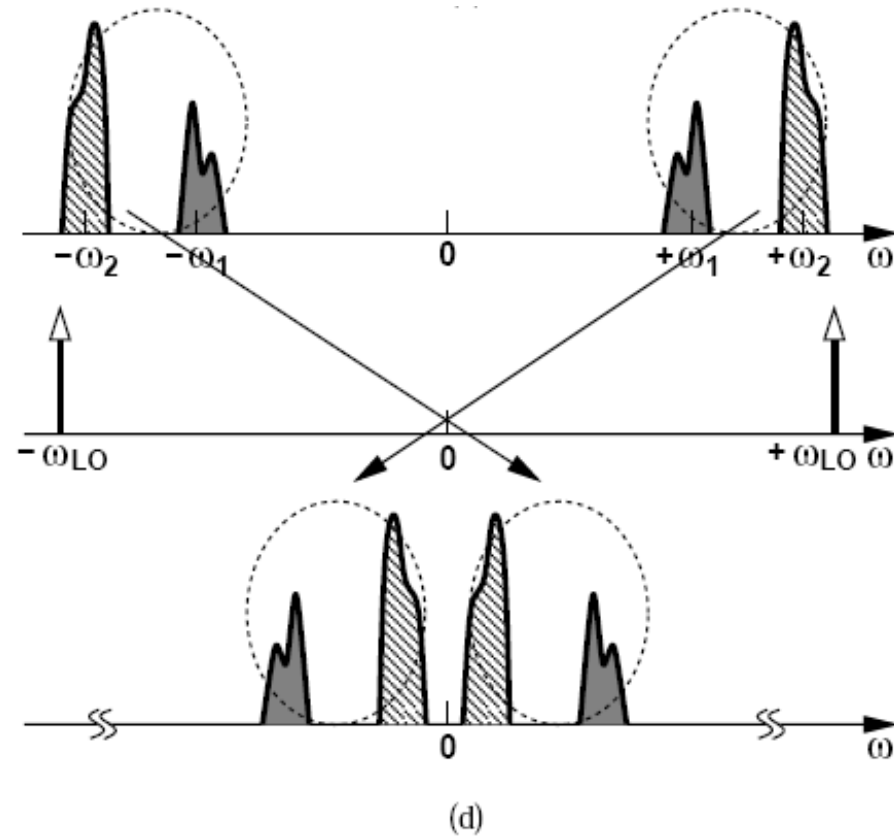


Downconverted Spectrum with LO – II

ω_{LO} midway between ω_1 and ω_2



$\omega_{LO} > \omega_2$ (high-side injection)



High-Side/Low-Side Injection: Image

Can a dual-mode RX designed for both 802.11g and 802.11a operate with a single LO?

- LO = halfway between the two bands
- Covers the 11g band by high-side injection and the 11a band by low-side injection.
- Greatly simplifies system design
- **Makes each band the image of the other.**
- Eg.: If RX is in the 11a mode while an 11g TX operates in close proximity, the reception may be heavily corrupted.
- IF in this case is quite high \rightarrow poor channel selection

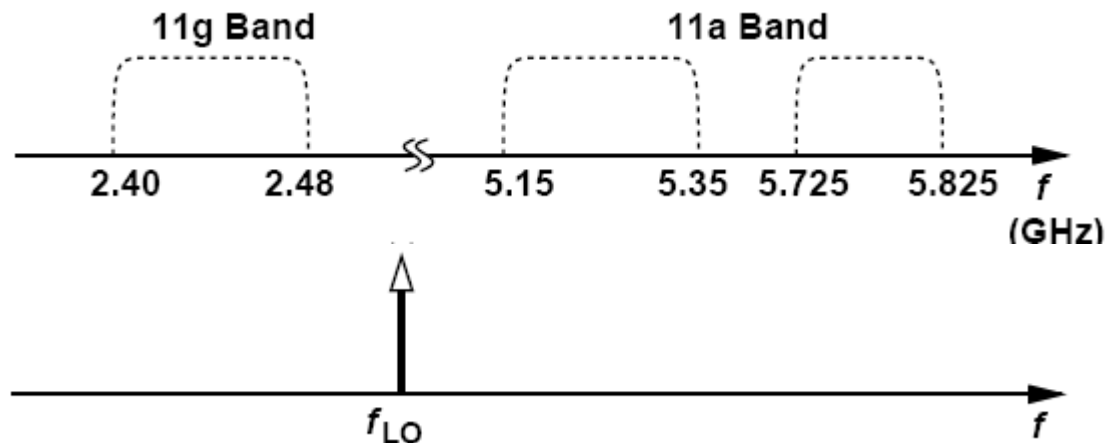


Image Rejection

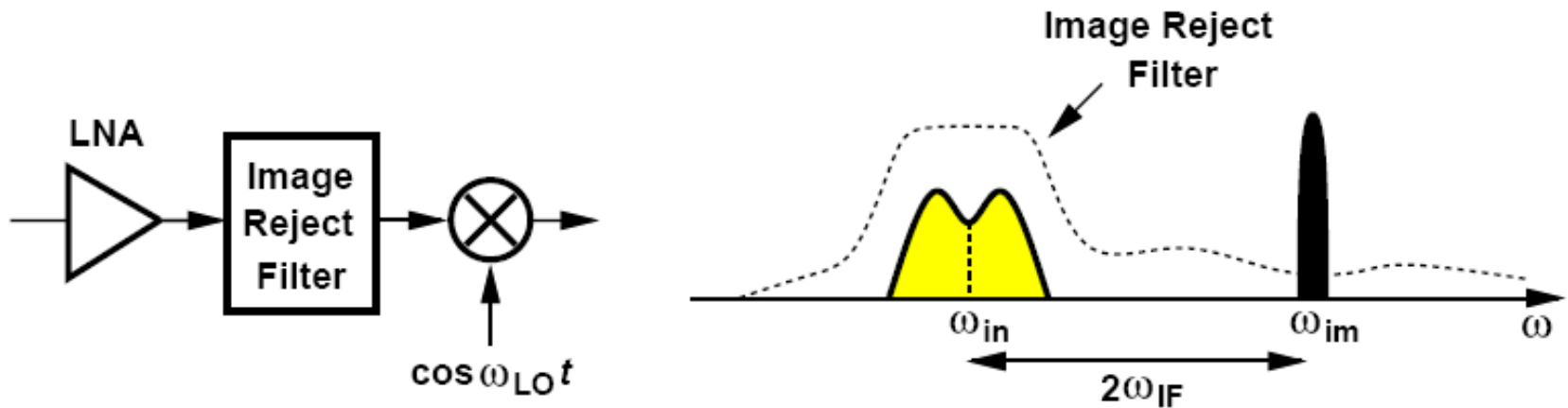
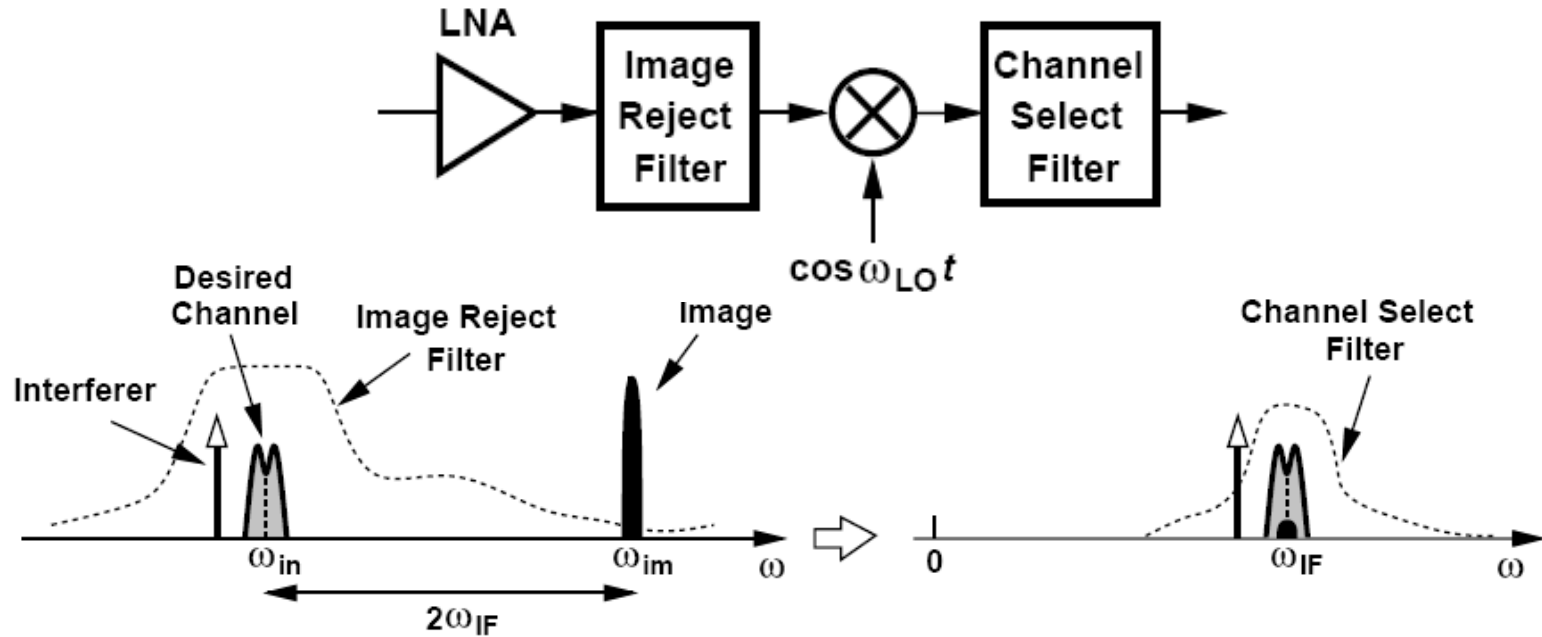


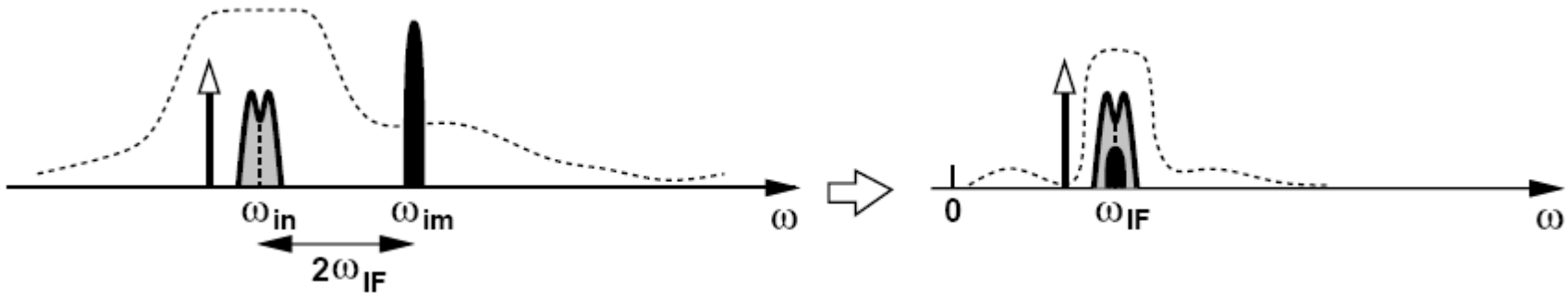
Image-reject filter for high image rejection

- Typically between LNA–Mixer so that LNA gain lowers the filter's contribution to the RX NF
- Linearity and selectivity requirements have dictated passive, off-chip implementations

Image Rejection vs. Channel Selection

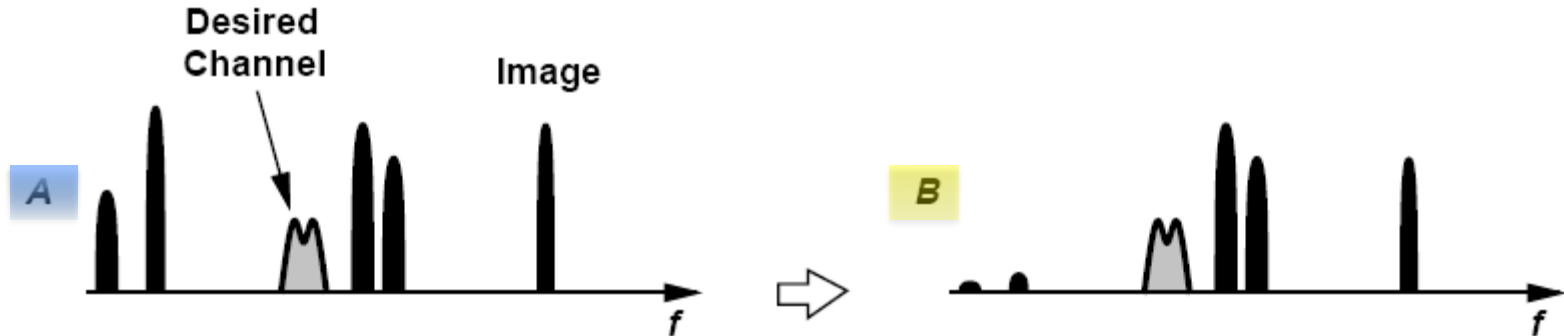
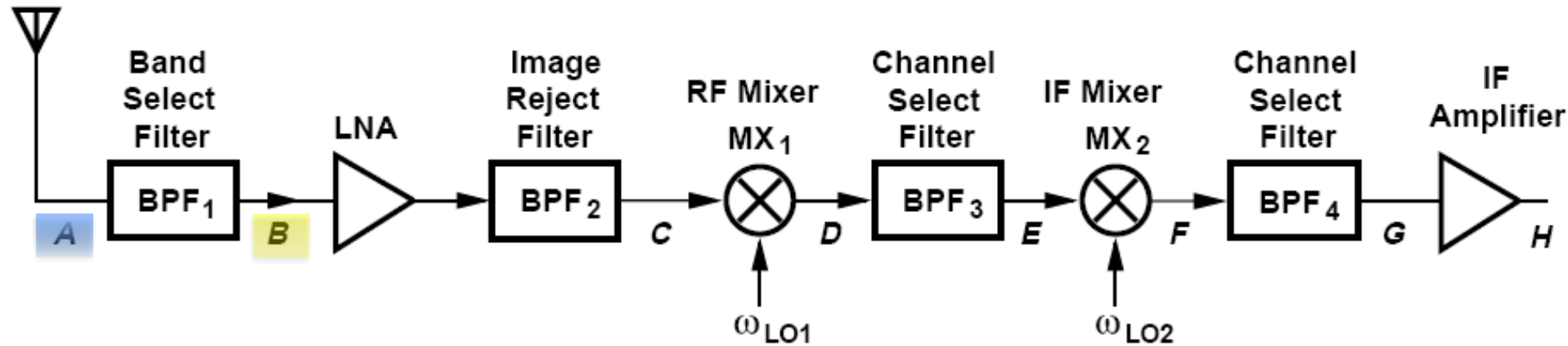


- A high IF allows substantial image-rejection but poor channel selection for given Q_{CSF} .



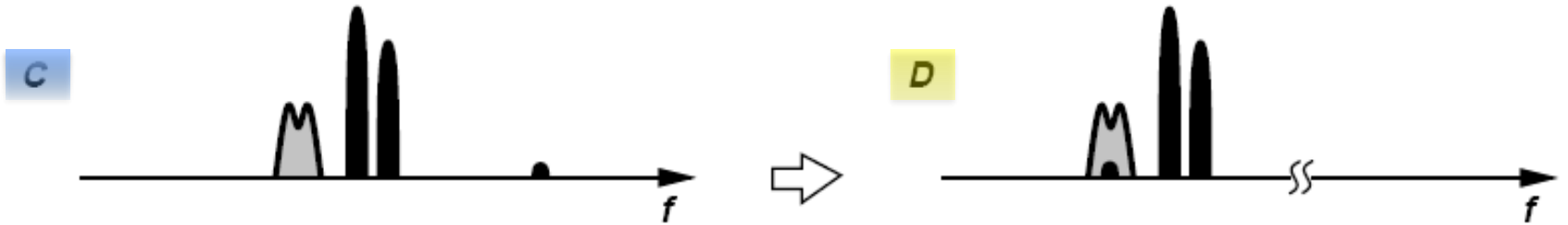
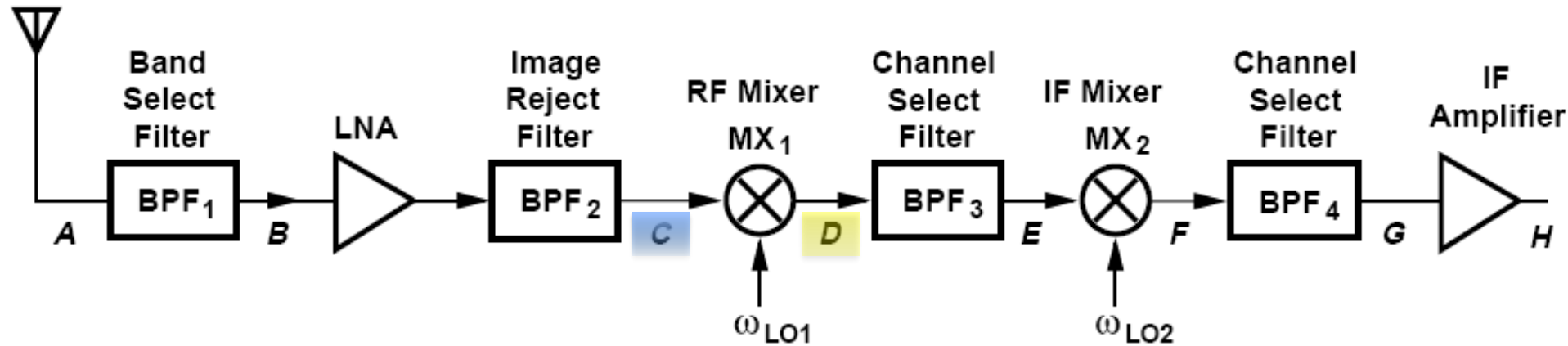
- A low IF helps with the channel selection and suppression of in-band interferers.

Dual Downconversion RX



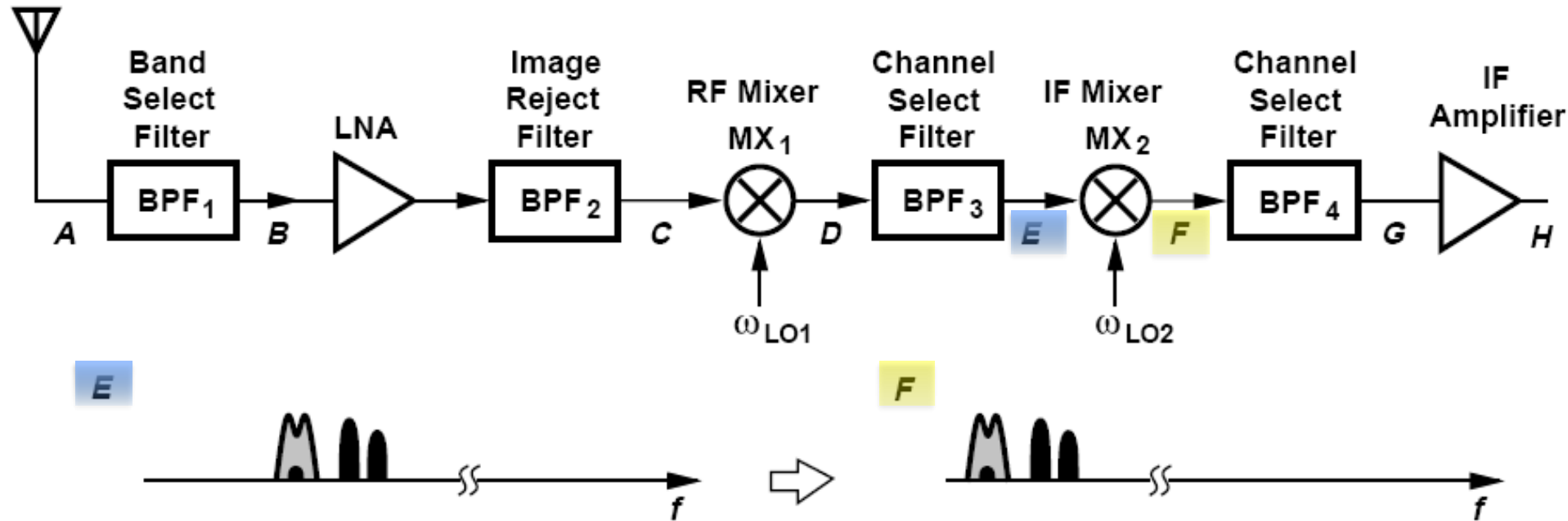
- BPF₁ selects the band while providing some image rejection as well (Point B)

Dual Downconversion RX



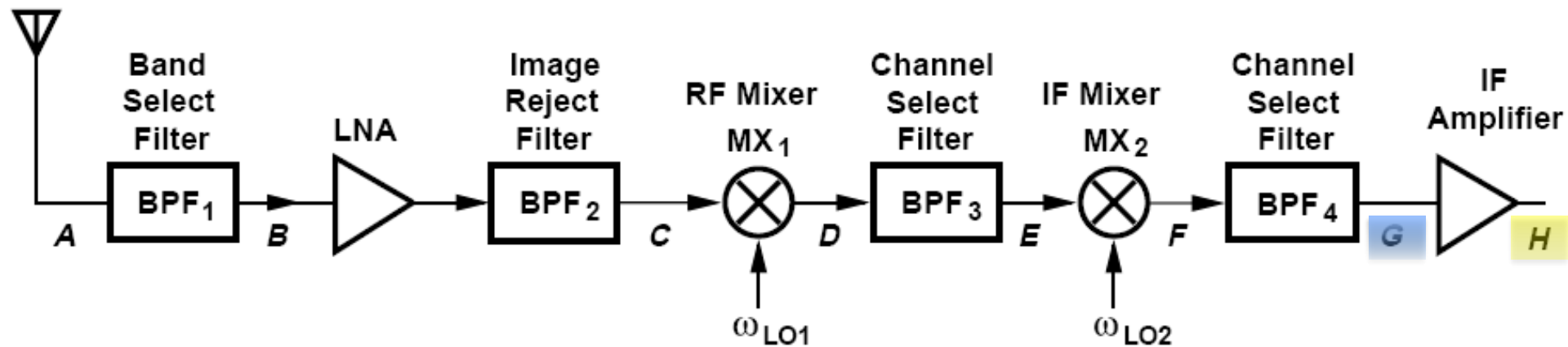
- After amplification and image-reject filtering, spectrum of C obtained
- Sufficiently linear mixer translates desired channel and adjacent interferers to first IF (Point D)

Dual Downconversion RX



- Partial channel selection BPF₃ permits the use of a second mixer with reasonable linearity. (Point E)
- Spectrum is translated to second IF. (Point F)

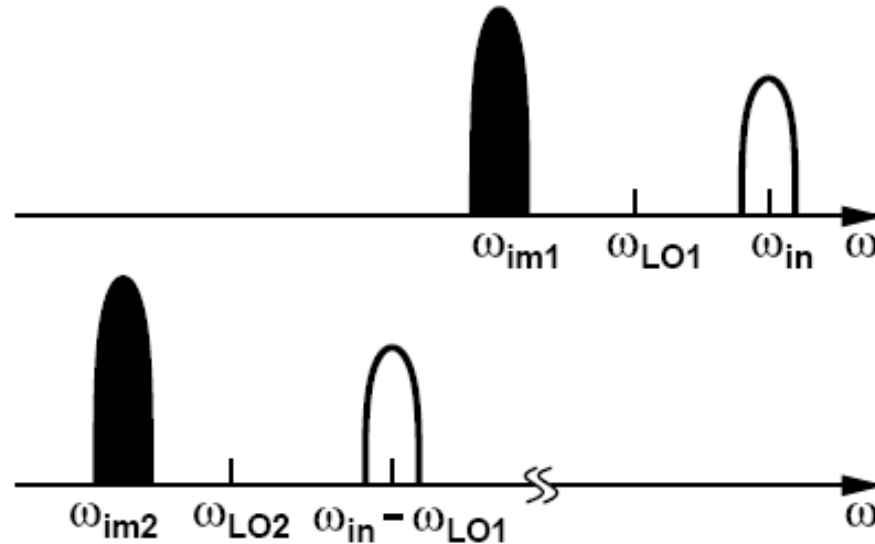
Dual Downconversion RX



- BPF₄ 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.

Dual Image Frequencies

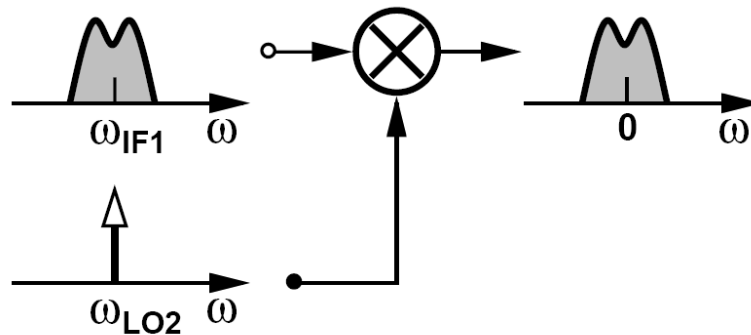
Assuming low-side injection for both downconversion mixers, two image frequencies exist:



1st image $\rightarrow @ 2\omega_{LO1} - \omega_{in}$

2nd image $\rightarrow @ 2\omega_{LO2} - (\omega_{in} - \omega_{LO1})$

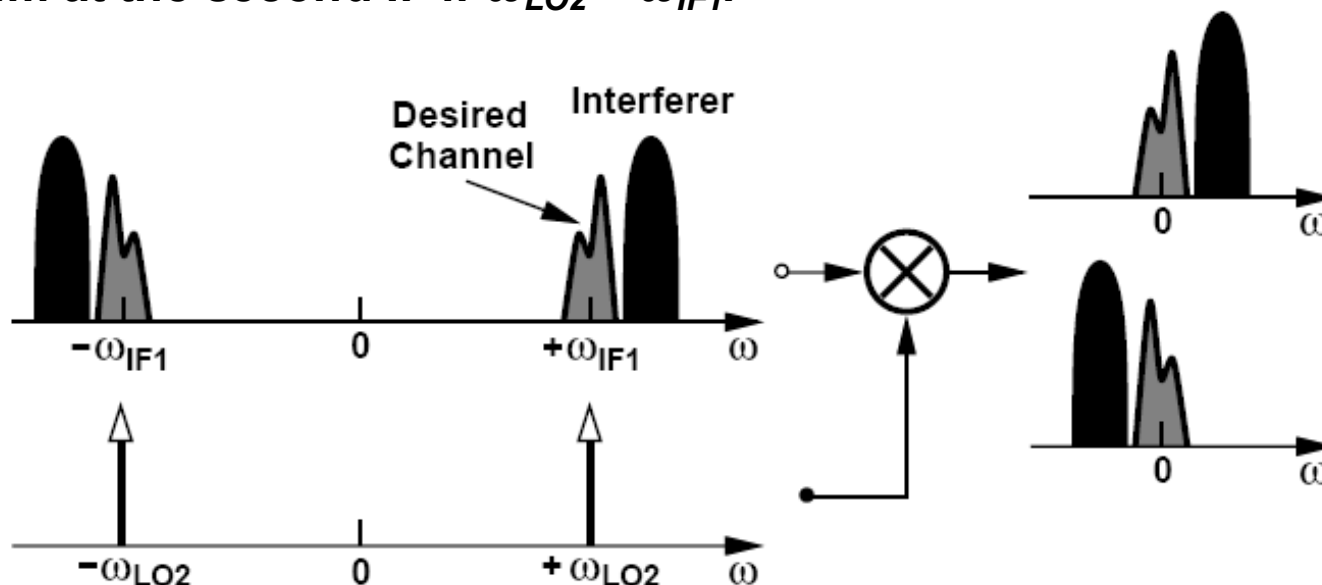
Zero Second IF



- To avoid secondary image, most modern heterodyne RXs 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 $\omega_{LO2} = \omega_{IF1}$

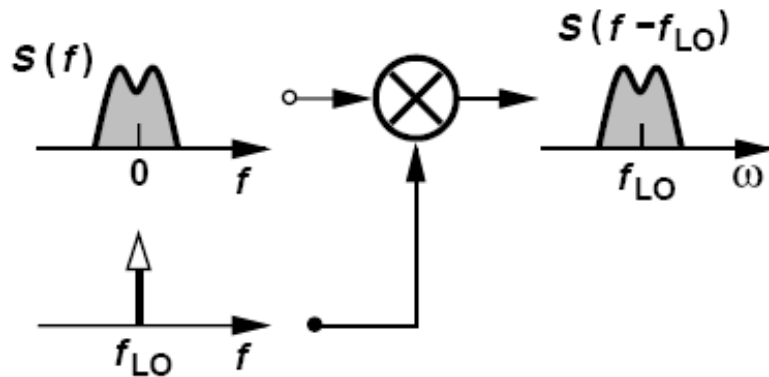
Example of Zero Second IF

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

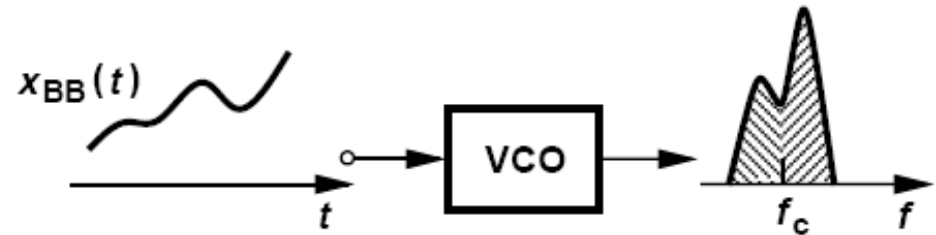


- Desired channel appears at $\pm \omega_{IF1}$ and is accompanied by the interferer.
- Spectrum at -ve freq. is convolved with $+\omega_{LO2}$ and shifted down to zero.
- Spectrum at +ve freq. is convolved with $-\omega_{LO2}$ and shifted down to zero.
- The output consists of two copies of the desired channel surrounded by the interferer spectrum at both +ve and -ve frequencies.

Symmetric/Asymmetric Modulation



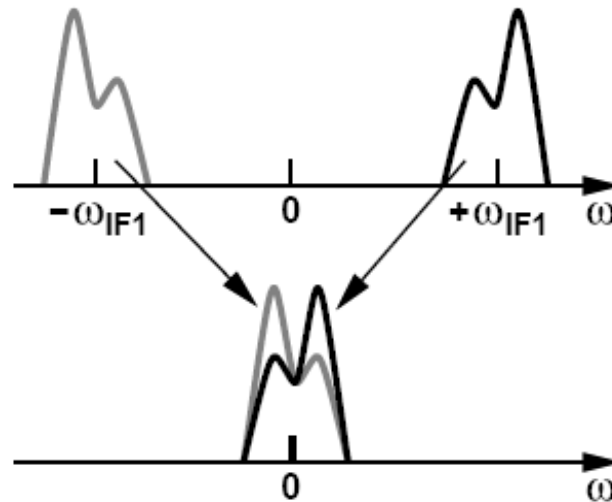
AM signal generation



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 in amplitude and/or phase.

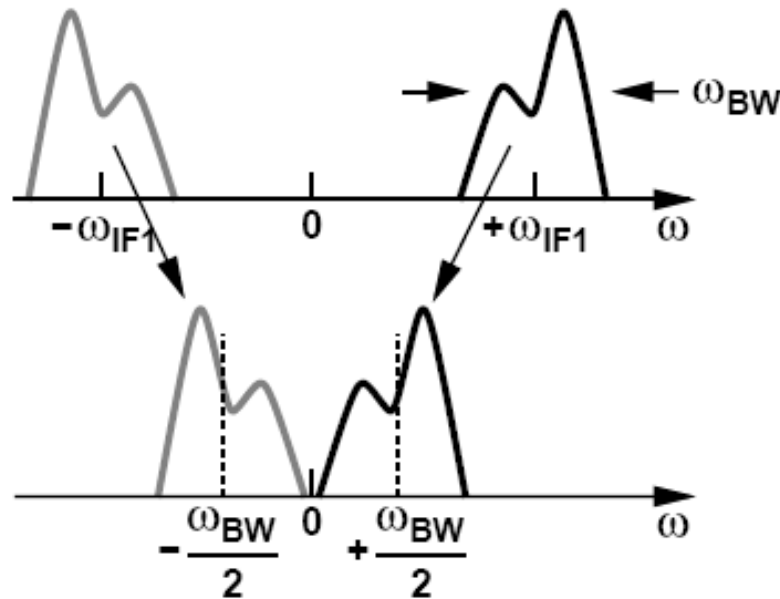
Corruption of Asymmetric Spectrum



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

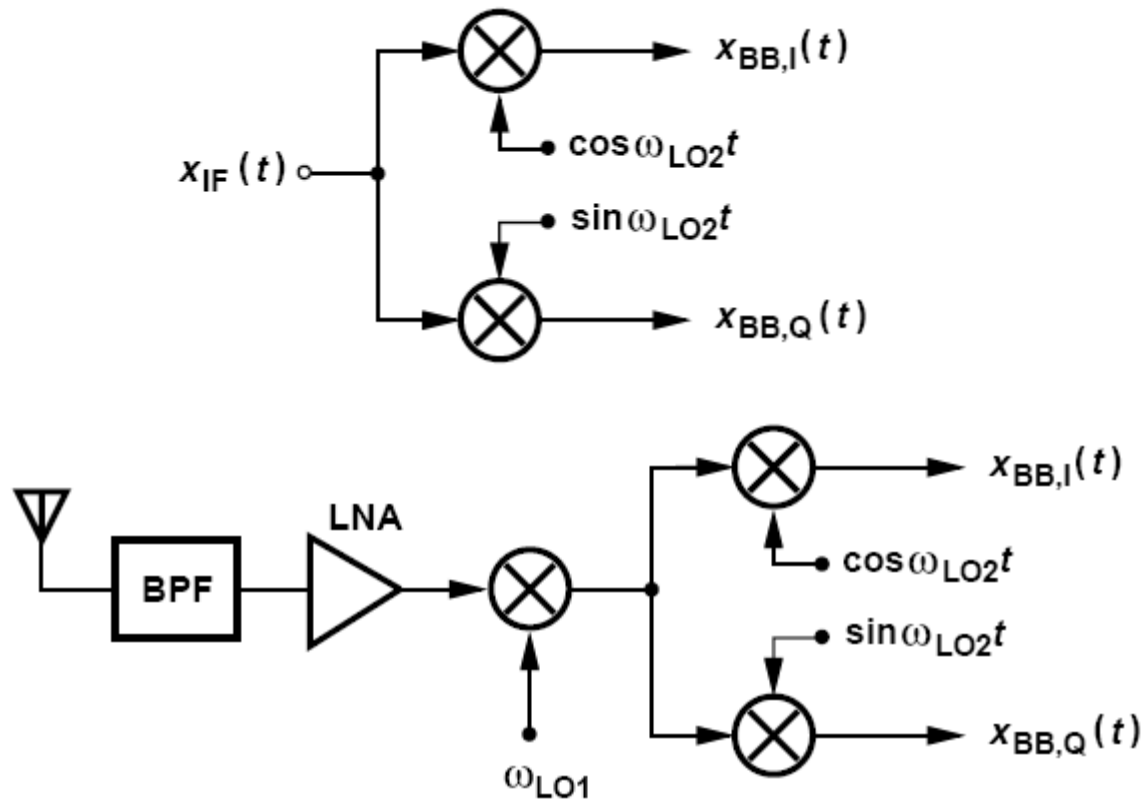
Preventing Self-corruption

What minimum IF avoids self-corruption of asymmetric signals upon downconversion?



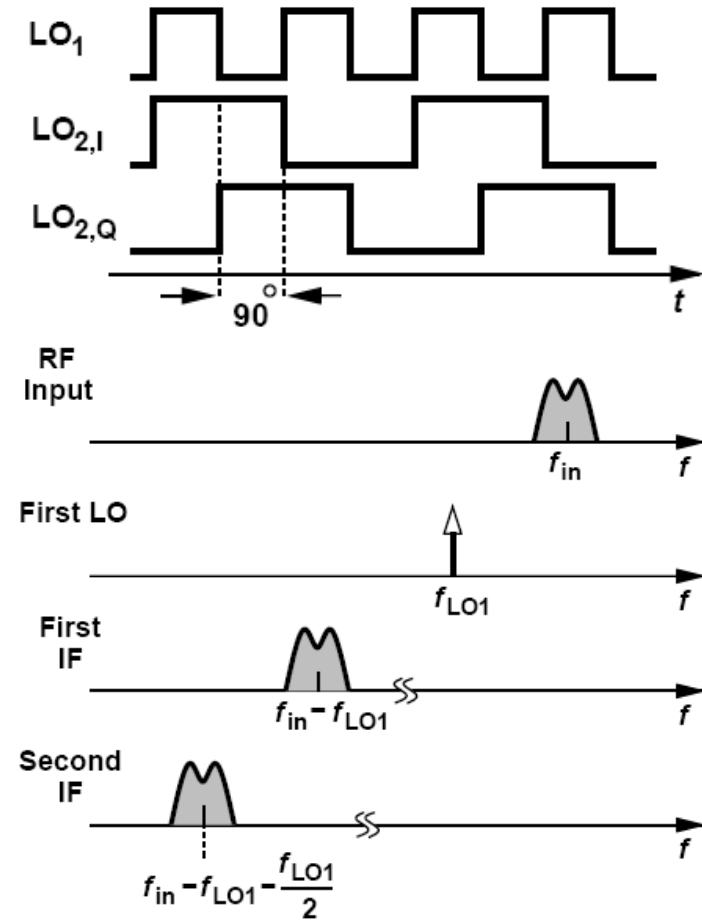
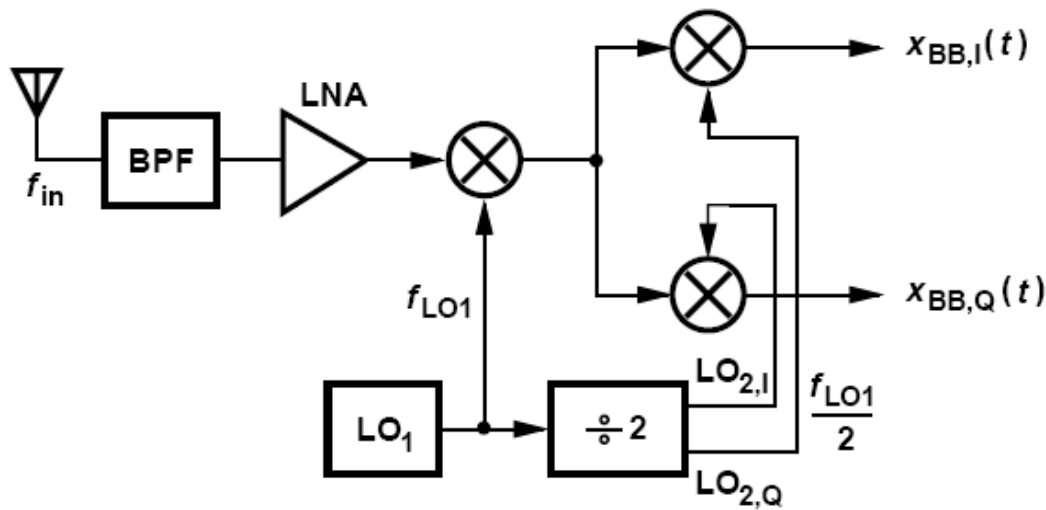
- To avoid overlap, the signal can be downconverted to an $IF = \frac{1}{2}$ signal BW .
- Of course, an interferer may now become the image.

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

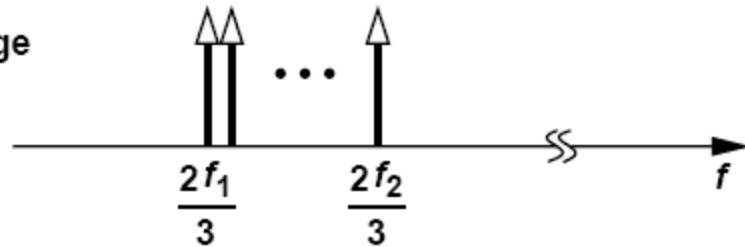
Sliding-IF RX

RF Range



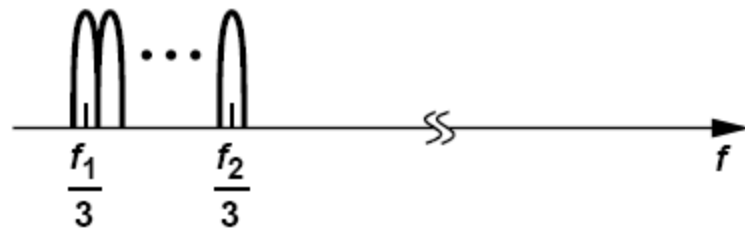
$$f_{LO1} + \frac{1}{2}f_{LO1} = f_{in}$$

LO Range



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

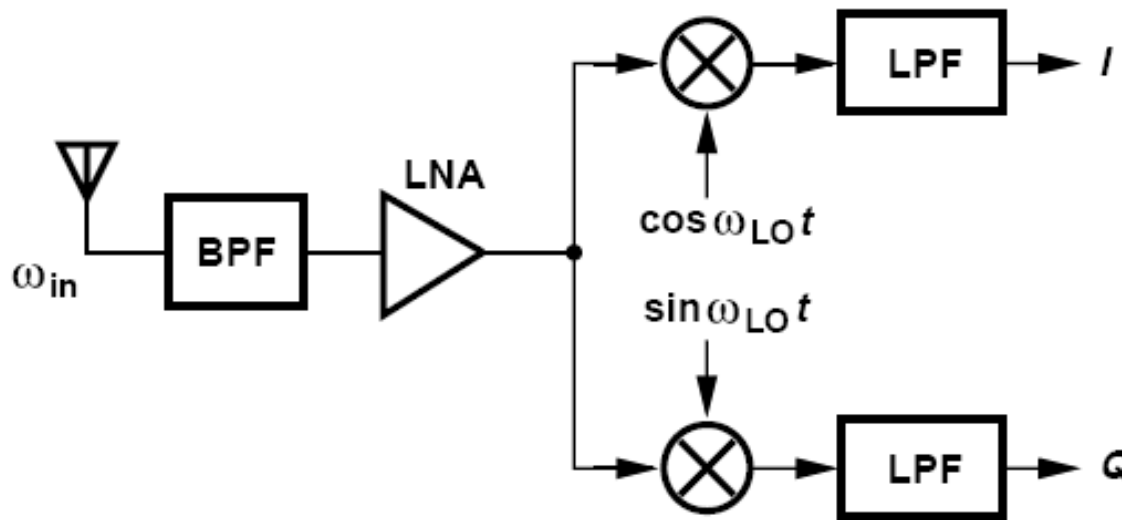
First IF Range



$$\begin{aligned} f_{IF1} &= f_{in} - f_{LO} \\ &= \frac{1}{3}f_{in}. \end{aligned}$$

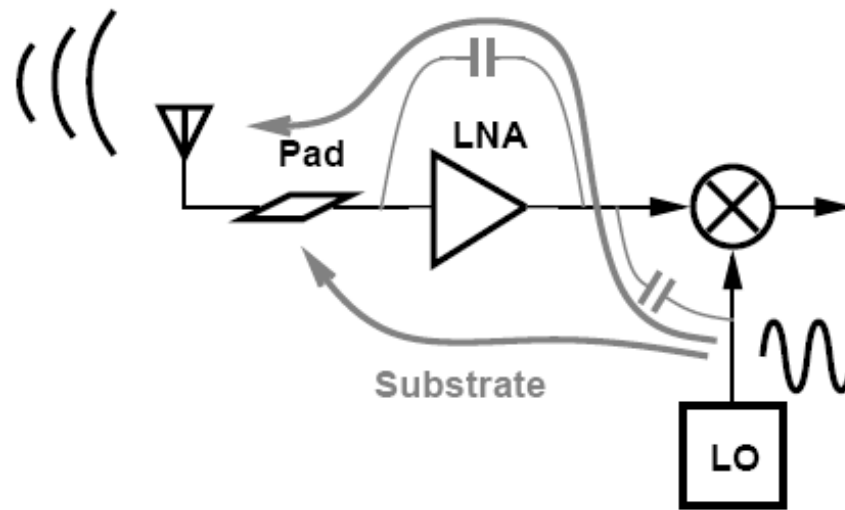
- As f_{in} varies from f_1 to f_2 , f_{IF1} goes from $f_1/3$ to $f_2/3 \rightarrow$ Sliding IF.

Direct-Conversion Receivers



- Absence of an image greatly simplifies the design process
- Channel selection is performed by on-chip LPF
- Mixing spurs are considerably reduced in number
- Quadrature phases to prevent self-corruption of signal (asymmetrically-modulated) upon downconversion

LO Leakage



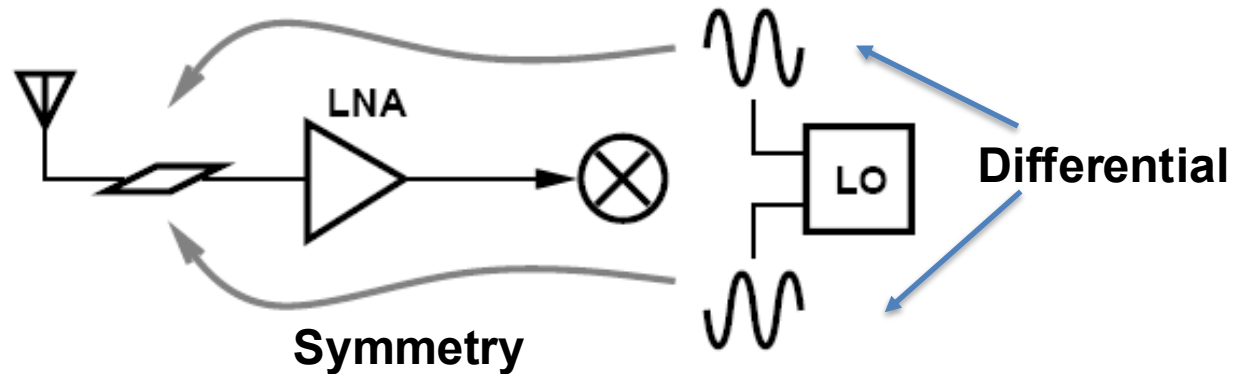
LO couples to the antenna through:

- device cap between LO and RF ports of mixer and device cap or resistances between the output and input of the LNA
- the substrate to the input pad, esp. because the LO employs large on-chip spiral inductors

LO leakage out of antenna can violate emission standards

- Must ensure adequate isolation
- Band-select filter helps in heterodyne Rx but not in DCR

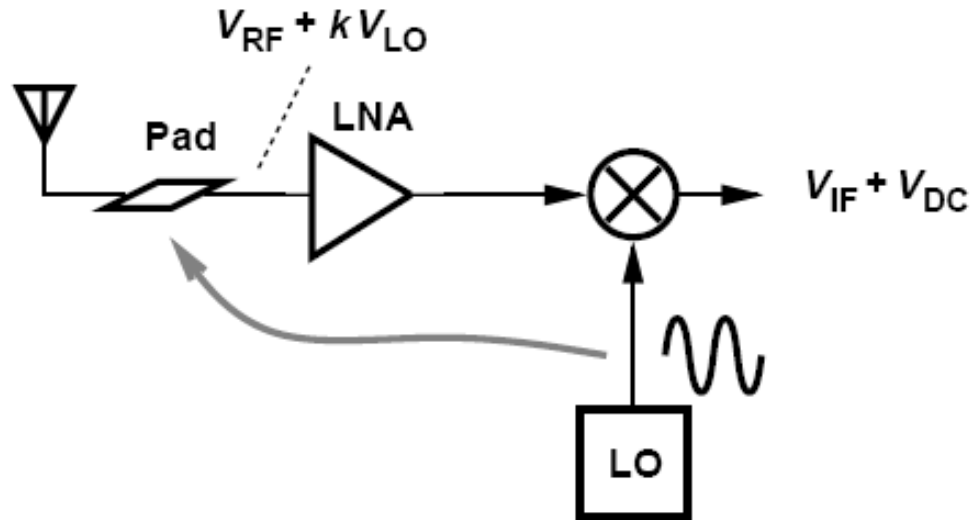
Cancellation of LO Leakage



LO leakage

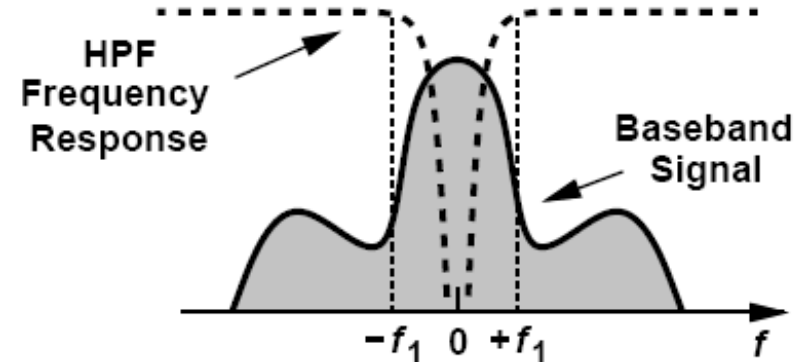
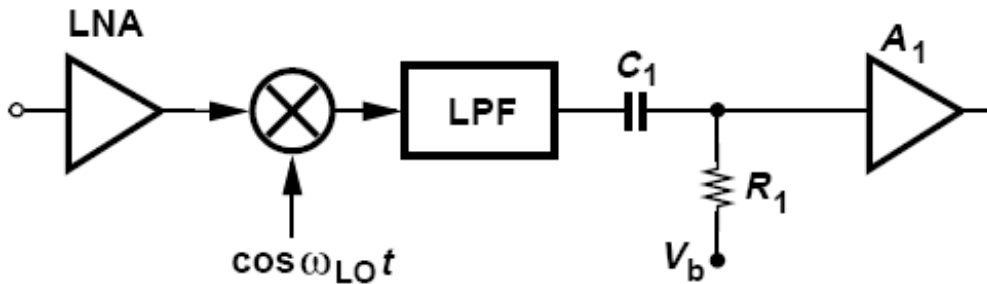
- arises primarily from random or deterministic asymmetries in the circuits and the LO waveform
- minimized through symmetric layout of the oscillator and the RF signal path

DC Offsets due to LO self-mixing

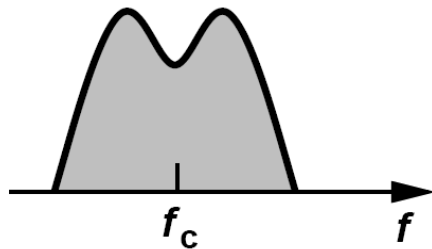


- A finite amount of in-band LO leakage appears at the LNA input. Along with the desired signal, this component is amplified and mixed with LO.
- May saturate baseband circuits, simply prohibiting signal detection.

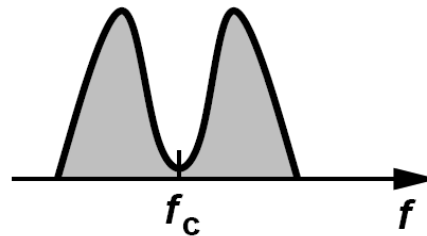
Cancellation of DC Offsets



- Offset cancellation: high-pass filter
- Such network also removes a fraction of the signal's spectrum near zero frequency, introducing intersymbol interference
- Large on-chip capacitor \rightarrow area \rightarrow ac coupling not popular



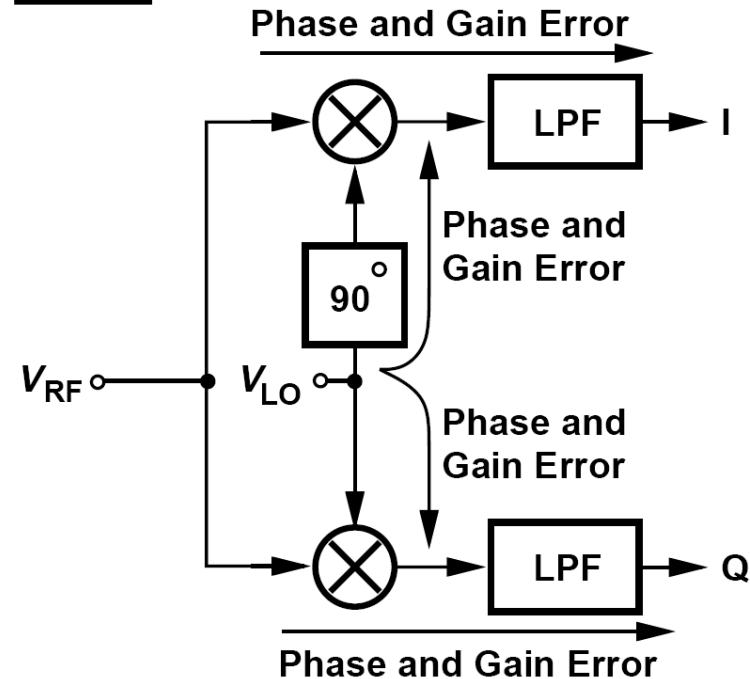
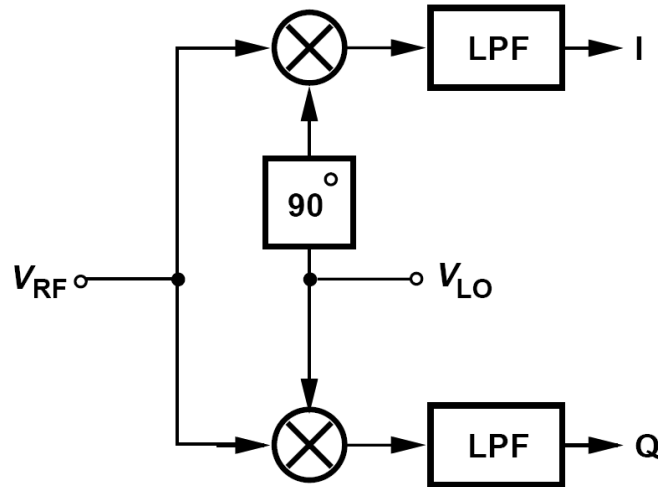
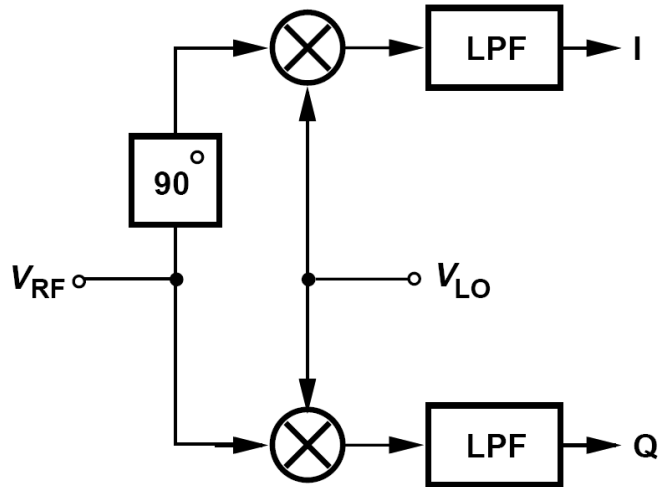
Small modulation index



Large modulation index

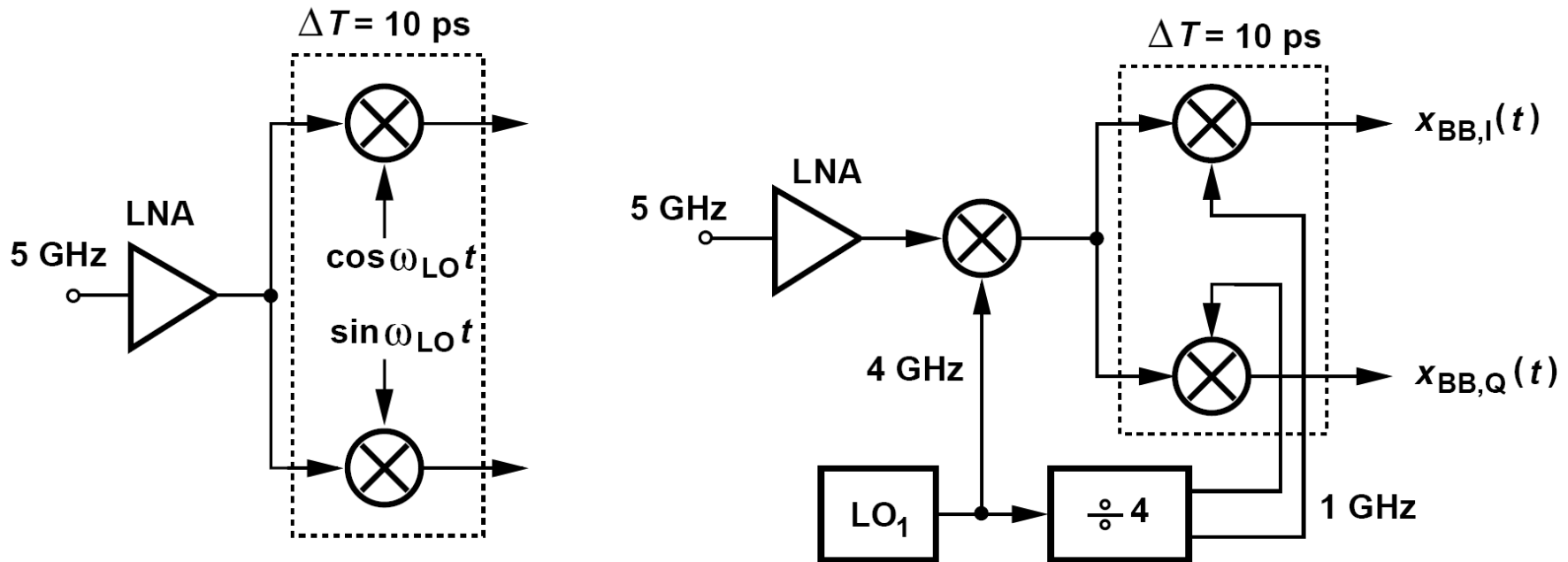
- Modulation schemes that contain little energy around the carrier better lend themselves to ac coupling in the baseband.
- A drawback of ac coupling stems from its slow response to transient input.

I/Q Mismatch: Sources



- Shifting either the RF signal or the LO waveform by 90°
- Errors in the 90° phase shift circuit and mismatches between the quadrature mixers \rightarrow imbalances in the amplitudes and phases of the baseband I and Q outputs.

I/Q Mismatch: DCR vs. Heterodyne RX



- Quadrature mismatches tend to be larger in DCR than in heterodyne topologies.
 - Propagation of a higher frequency (f_{in}) through quadrature mixers experiences greater mismatches
 - Quadrature phases of the LO itself suffer from greater mismatches at higher frequencies

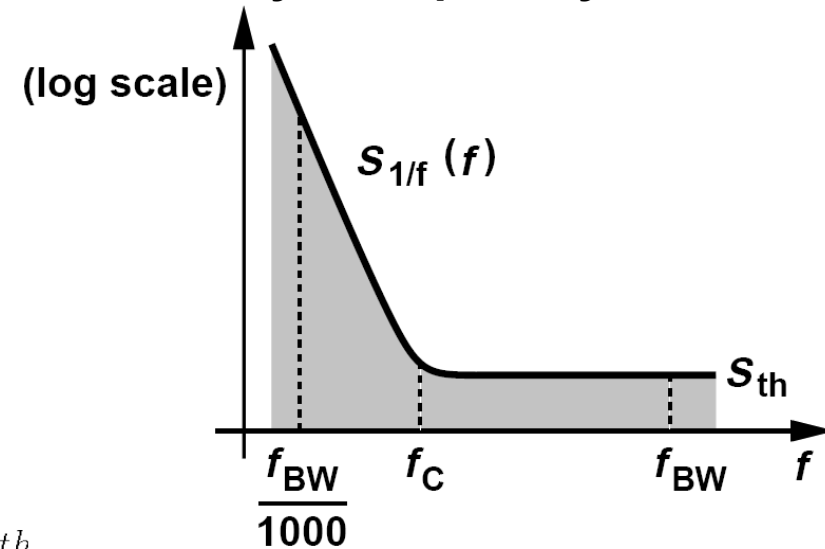
Flicker Noise Penalty

Signal centered around zero frequency can be substantially corrupted by 1/f noise

If $S_{1/f} = \alpha/f$, then at f_c , $\frac{\alpha}{f_c} = S_{th}$

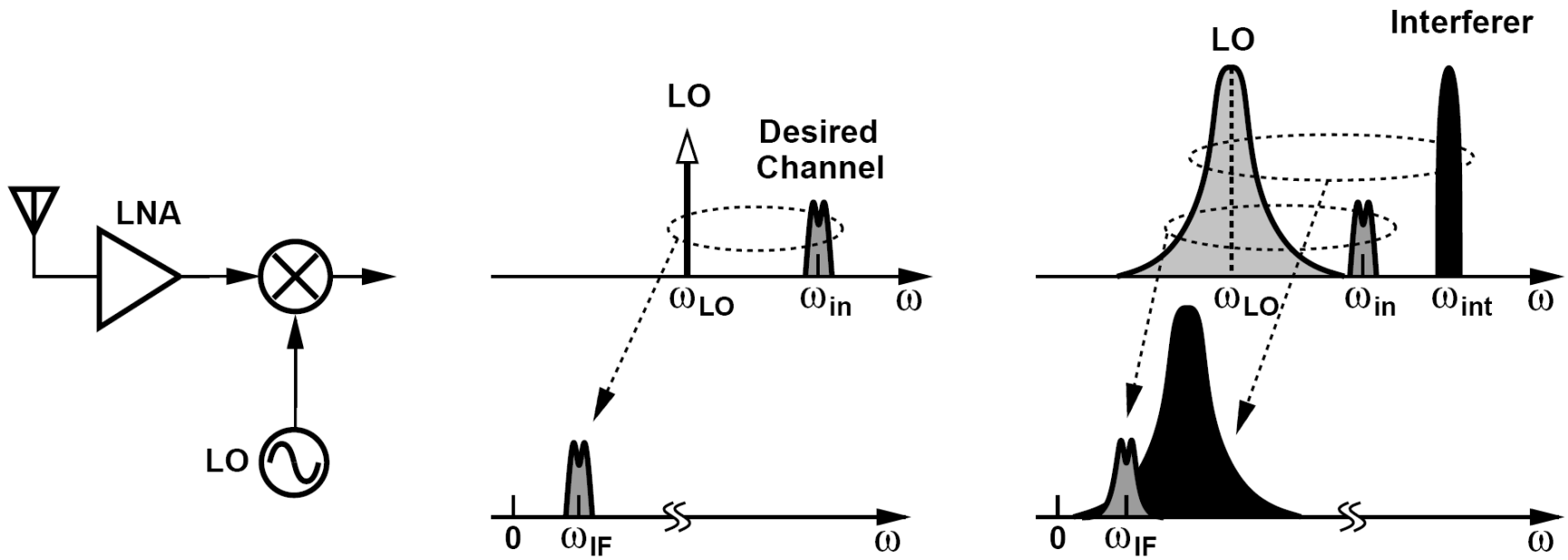
$$\begin{aligned}
 P_{n1} &= \int_{f_{BW}/1000}^{f_c} \frac{\alpha}{f} df + (f_{BW} - f_c) S_{th} \\
 &= \alpha \ln \frac{1000 f_c}{f_{BW}} + (f_{BW} - f_c) S_{th} \\
 &= \left(6.9 + \ln \frac{f_c}{f_{BW}} \right) f_c S_{th} + (f_{BW} - f_c) S_{th} \\
 &= \left(5.9 + \ln \frac{f_c}{f_{BW}} \right) f_c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

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



- Flicker Noise penalty is high for standards with narrow channel BW (GSM) → low-IF is better suited than DCR.

Phase Noise → Reciprocal Mixing



- Ideal case: $\omega_{IF} = \omega_{in} - \omega_{LO}$
- LO suffers from phase noise and the desired signal is accompanied by a large interferer.
- → a *broadened* downconverted interferer whose noise skirt corrupts the desired IF signal
- This phenomenon is called “reciprocal mixing”.

Example of Reciprocal Mixing

A GSM RX must withstand an interferer located three channels away from the desired channel and 45dB higher.

Estimate the maximum tolerable LO phase noise if the corruption due to reciprocal mixing must remain 15dB below the desired signal.

Total noise power introduced by the interferer in the desired channel

$$P_{n,tot} = \int_{f_L}^{f_H} S_n(f) df$$

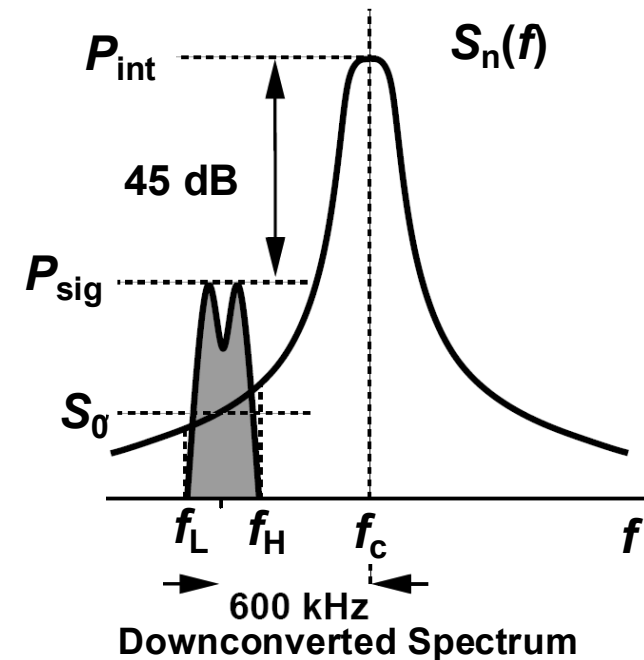
Assume $S_n(f) \sim$ relatively flat in this BW and $= S_0$,

$$\text{SNR} = \frac{P_{sig}}{S_0(f_H - f_L)} = 15\text{dB}$$

$$10 \log \frac{S_0}{P_{sig}} = -15 \text{ dB} - 10 \log(f_H - f_L)$$

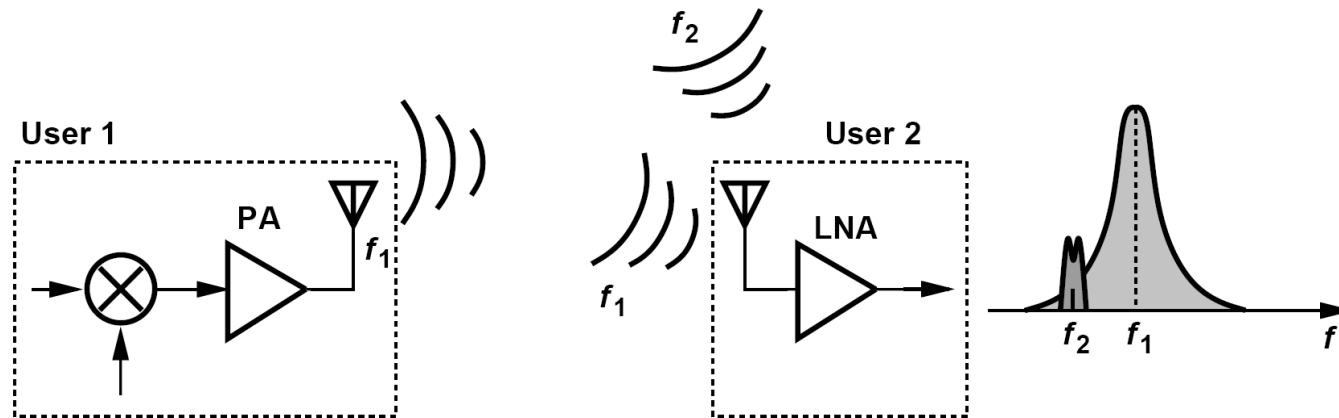
$$10 \log \frac{S_0}{P_{int}} = -15 \text{ dB} - 10 \log(f_H - f_L) - 45 \text{ dB}$$

If $f_H - f_L = 200 \text{ kHz}$, $10 \log \frac{S_0}{P_{int}} = -113 \text{ dBc/Hz}$ at 600-kHz offset

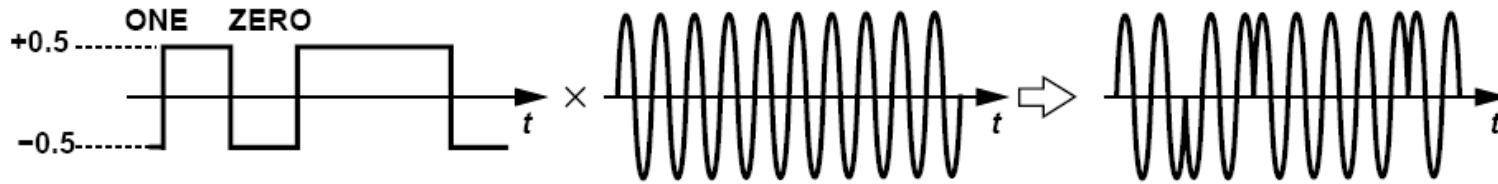


TX Phase Noise → Received Noise @RX

- Two users are located in close proximity. User #1 transmitting a high-power signal at f_1 and user #2 receiving this signal and a weak signal at f_2 .
- If f_1 and f_2 are only a few channels apart, the phase noise skirt around f_1 masks the signal received by user #2, greatly corrupting it even *before* downconversion.



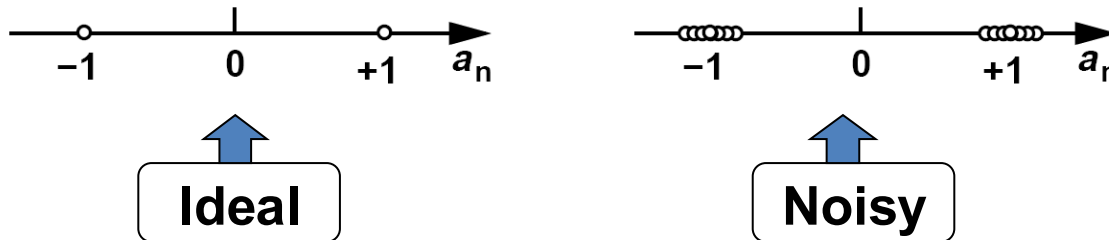
Digital Modulation: Binary PSK



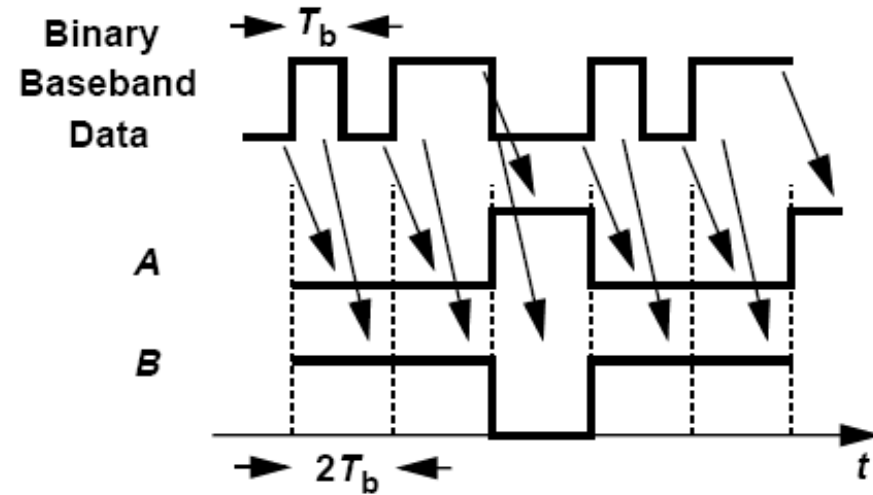
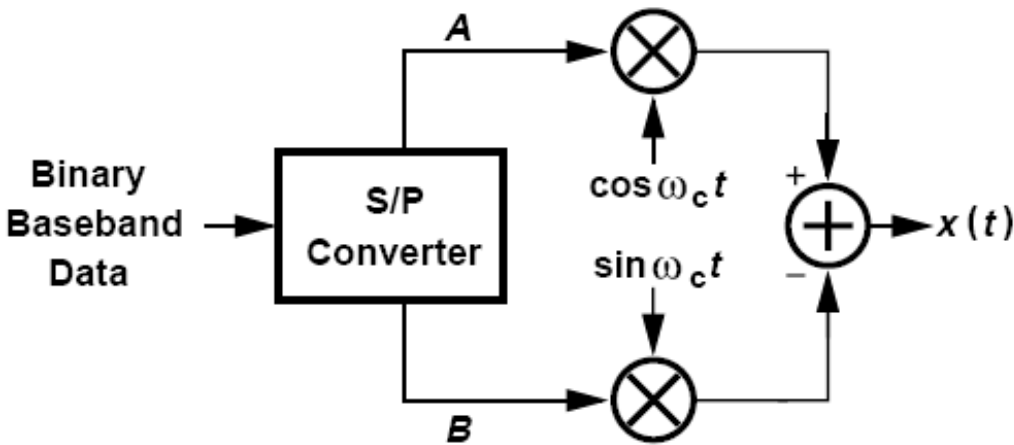
$$x_{PSK}(t) = A_c \cos \omega_c t \quad \text{If data = ZERO}$$
$$= A_c \cos(\omega_c t + 180^\circ) \quad \text{If data = ONE}$$

$$x_{PSK}(t) = a_n \cos \omega_c t \quad a_n = \pm 1$$

Signal Constellation in the presence of amplitude noise:



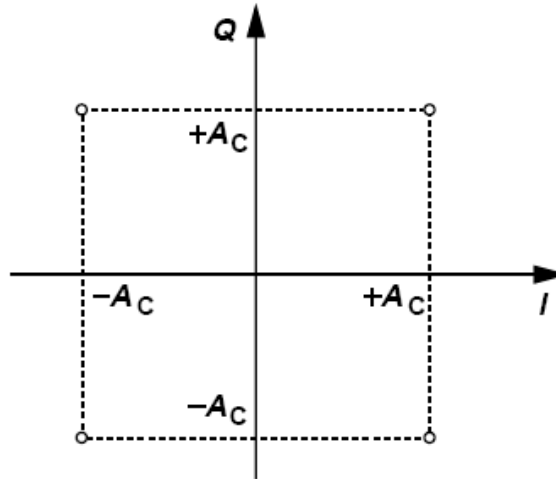
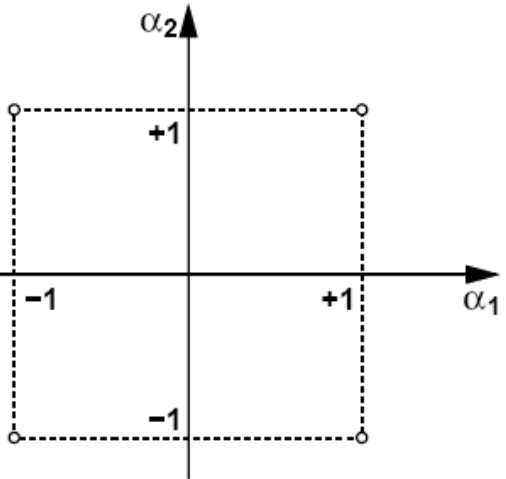
Quadrature Modulation: QPSK



$$x(t) = b_{2m} A_c \cos \omega_c t - b_{2m+1} A_c \sin \omega_c t$$

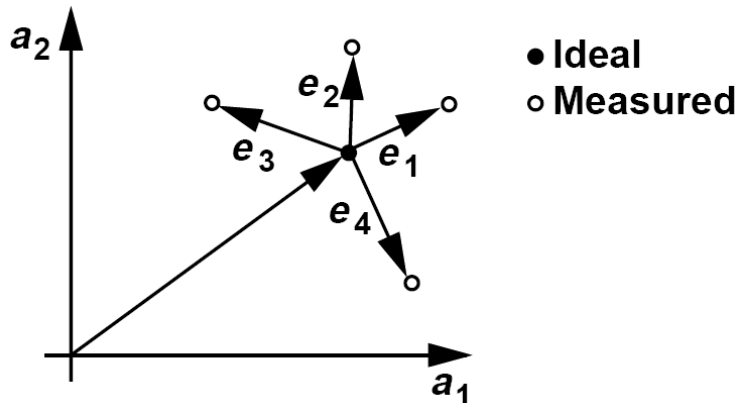
- S/P Converter (DeMUX) stretches each bit duration by 2X
- QPSK halves the occupied bandwidth
- Pulses appear at A and B are called *symbols* rather than *bits*
- For QPSK, Symbol rate = 1/2 Bit-rate

Signal Constellation & EVM



$$x(t) = \alpha_1 A_c \cos(\omega_c t) + \alpha_2 A_c \sin \omega_c t$$

- Error vector magnitude (EVM) represents the deviation of the constellation points from their ideal positions.



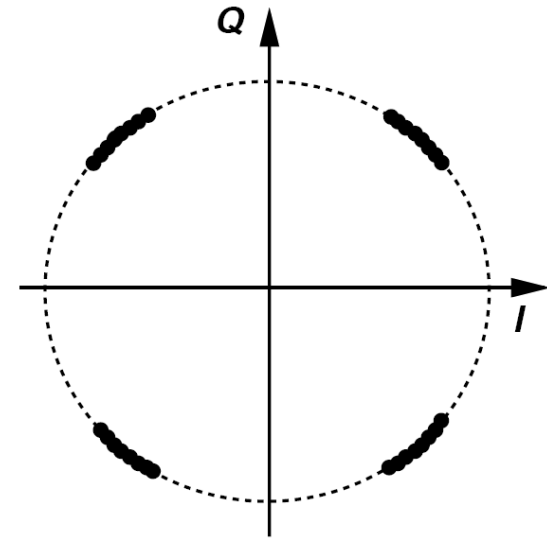
$$EVM_1 = \frac{1}{V_{rms}} \sqrt{\frac{1}{N} \sum_{j=1}^N e_j^2}$$

$$EVM_2 = \frac{1}{P_{avg}} \cdot \frac{1}{N} \sum_{j=1}^N e_j^2$$

Phase Noise Corrupts Sig-Constellation

- Since the phase noise is indistinguishable from PM (or FM), the mixing of the signal with a noisy LO in the TX or RX path corrupts the information carried by the signal.

$$x_{QPSK}(t) = A \cos \left[\omega_c t + (2k + 1) \frac{\pi}{4} + \phi_n(t) \right] \quad k = 0, \dots, 3$$

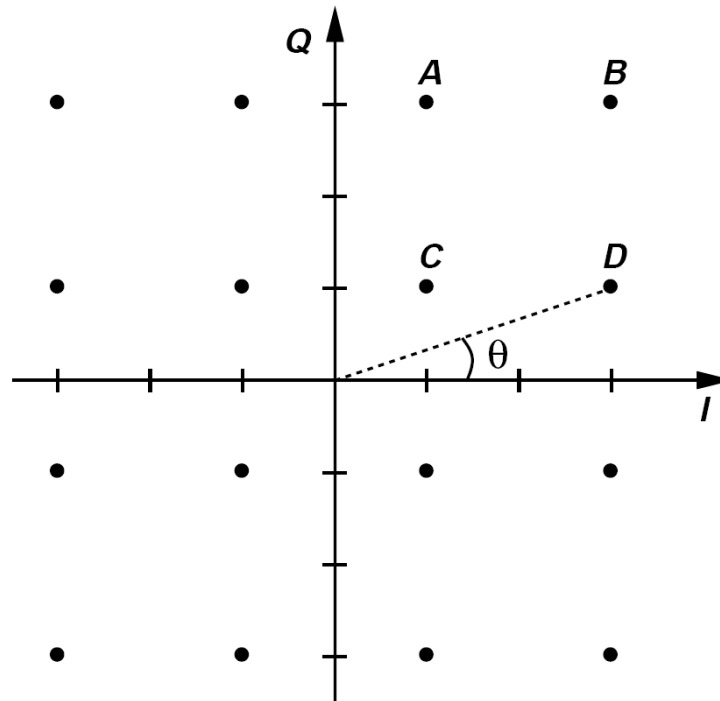


- Constellation points experience only random rotation around the origin.
- If large enough, phase noise and other nonidealities move a constellation point to another quadrant, creating an error.

16-QAM Constellation

Which points in a 16-QAM constellation are most sensitive to phase noise?

- Points *B* and *C* can tolerate a rotation of 45° before they move to adjacent quadrants.
- Points *A* and *D*, on the other hand, can rotate by only $\theta = \tan^{-1}(1/3) = 18.4^\circ$
- Thus, the eight outer points near the *I* and *Q* axes are most sensitive to phase noise.



RX Sensitivity

- Minimum signal level that RX can detect with “acceptable quality.”

$$\begin{aligned} NF &= \frac{SNR_{in}}{SNR_{out}} \\ &= \frac{P_{sig}/P_{RS}}{SNR_{out}} \end{aligned}$$

$$P_{sig} = P_{RS} \cdot NF \cdot SNR_{out}$$

$$P_{sig,tot} = P_{RS} \cdot NF \cdot SNR_{out} \cdot B$$

$$P_{sen}|_{dBm} = P_{RS}|_{dBm/Hz} + NF|_{dB} + SNR_{min}|_{dB} + 10 \log B$$

$$P_{sen} = \underbrace{-174 \text{ dBm/Hz} + NF + 10 \log B}_{\text{Noise Floor}} + SNR_{min}$$

Noise Floor

Example of Sensitivity

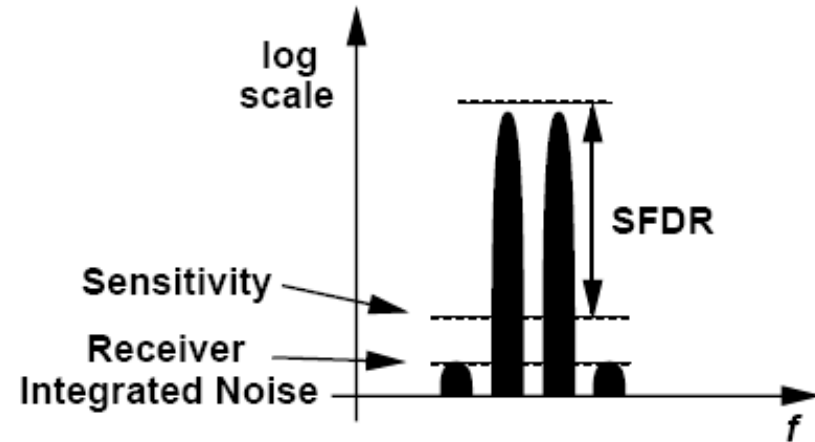
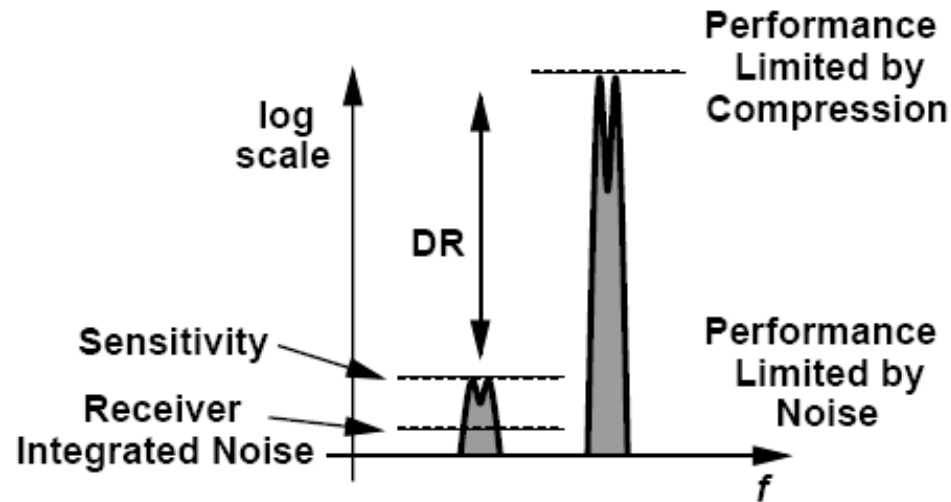
A GSM RX requires a minimum SNR of 12dB and has a channel $BW = 200\text{kHz}$.
A WLAN RX requires a minimum SNR of 23dB and has a channel $BW = 20\text{MHz}$.
Compare the sensitivities if both have an NF of 7dB.

For the GSM RX, $P_{sen} = -102\text{dBm}$,
For the WLAN RX, $P_{sen} = -71\text{dBm}$.

Does this mean that the latter is inferior? No, the latter employs a much wider BW and a more efficient modulation to accommodate a data rate of 54Mb/s. The GSM system handles a data rate of only 270 kb/s. In other words, specifying the sensitivity of a RX without the data rate is meaningless.



Dynamic Range & Spurious-Free DR



- **Dynamic Range:**
Maximum tolerable desired sig power
Minimum tolerable desired sig power (Sensitivity)

- **SFDR:**
Lower end equal to sensitivity.
Higher end defined as maximum input level in a *two-tone* test for which the IM3 products do not exceed the integrated noise of the RX

- **SFDR is a more stringent characteristic of system than DR**

SFDR Calculation

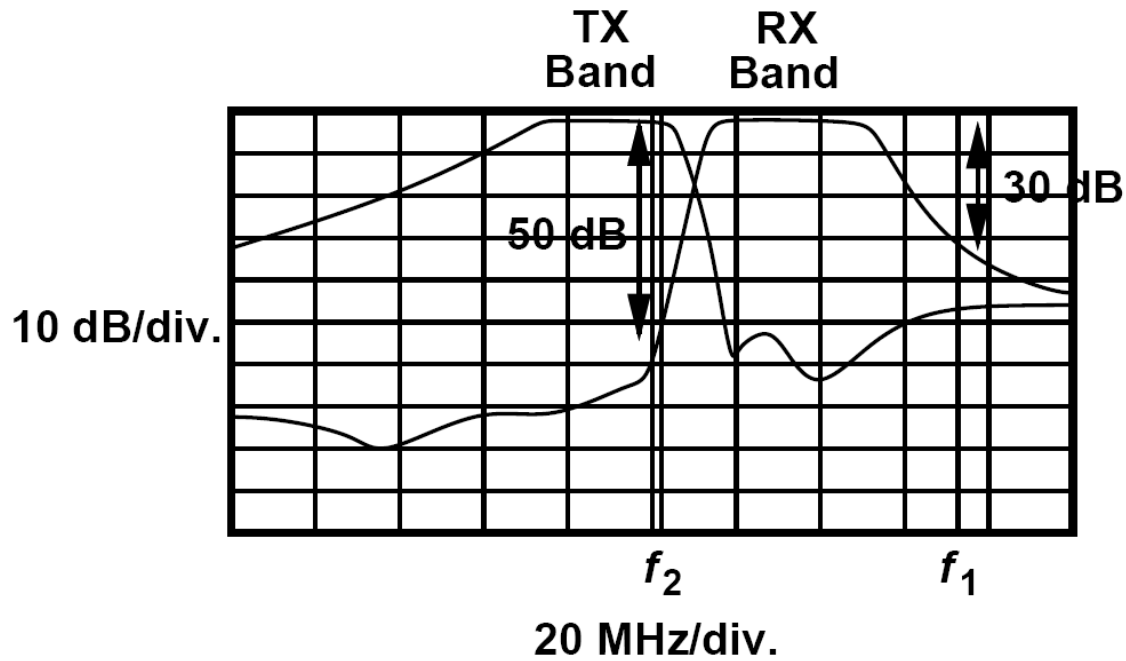
The upper end of the SFDR is that value of P_{in} which makes $P_{IM,in}$ equal to the integrated noise of the RX.

$$P_{in,max} = \frac{2P_{IIP3} + (-174 \text{ dBm} + NF + 10 \log B)}{3}.$$

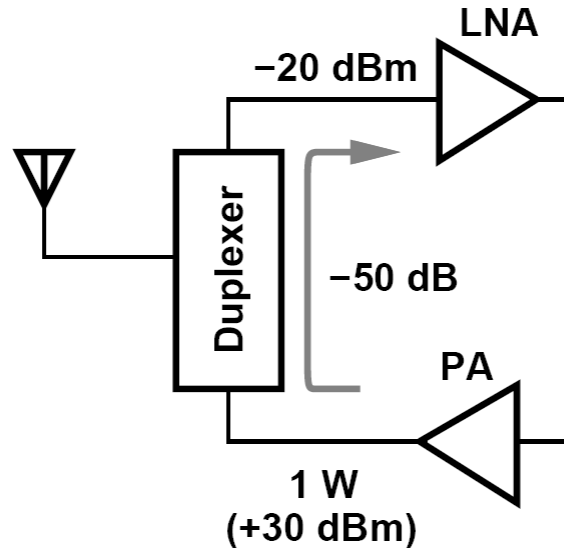
SFDR is the difference in dB between $P_{in,max}$ and the sensitivity:

$$\begin{aligned} SFDR &= P_{in,max} - (-174 \text{ dBm} + NF + 10 \log B + SNR_{min}) \\ &= \frac{2(P_{IIP3} + 174 \text{ dBm} - NF - 10 \log B)}{3} - SNR_{min}. \end{aligned}$$

Duplexer Characteristics



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

A BPF following the LNA can alleviate the TX-RX leakage in a CDMA system.

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.

