MICRO-461
Low-power Radio Design for the IoT

7. RF Transceiver Architectures

Christian Enz

Integrated Circuits Lab (ICLAB), Institute of Microengineering (IMT), School of Engineering (STI)

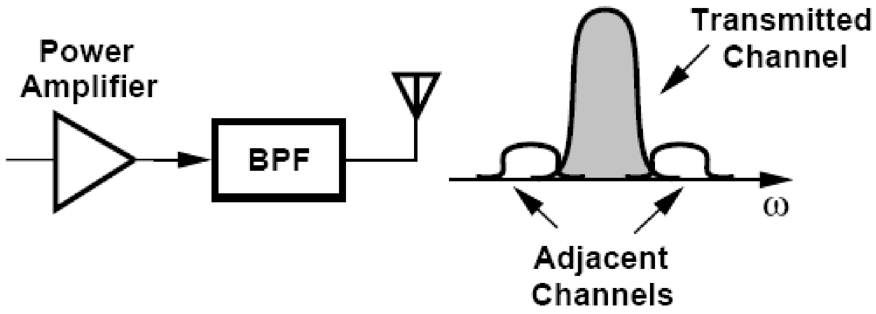
Swiss Federal Institute of Technology, Lausanne (EPFL), Switzerland

EPFL

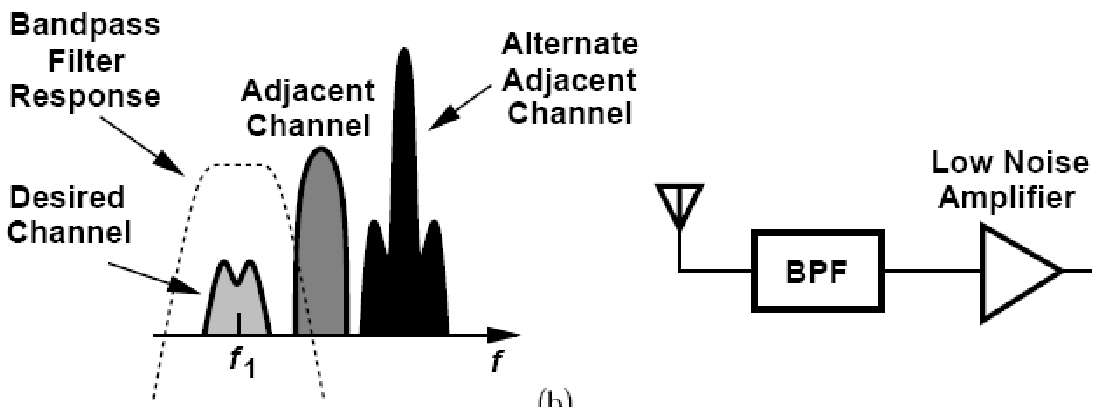
Outline

- **General Considerations**
- Heterodyne Receivers
- Homodyne Receivers
- Image-Reject Receivers
- Low-IF Receivers
- Discrete-time Receivers

Limited Bandwidth



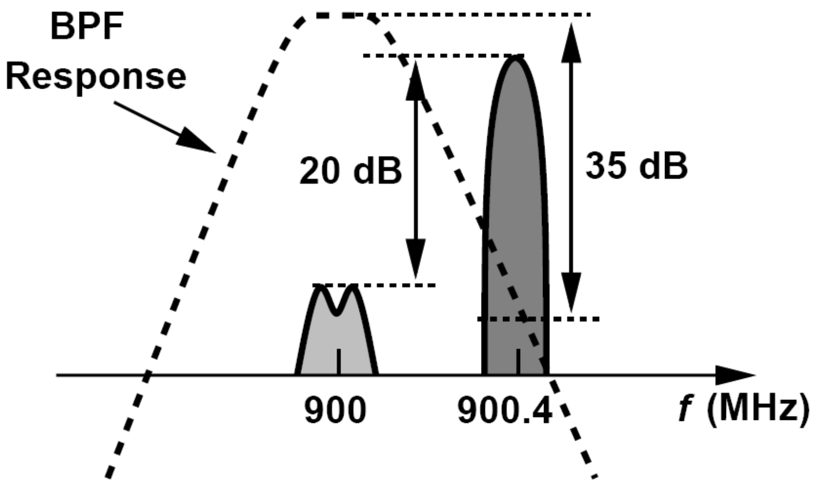
(a)



(b)

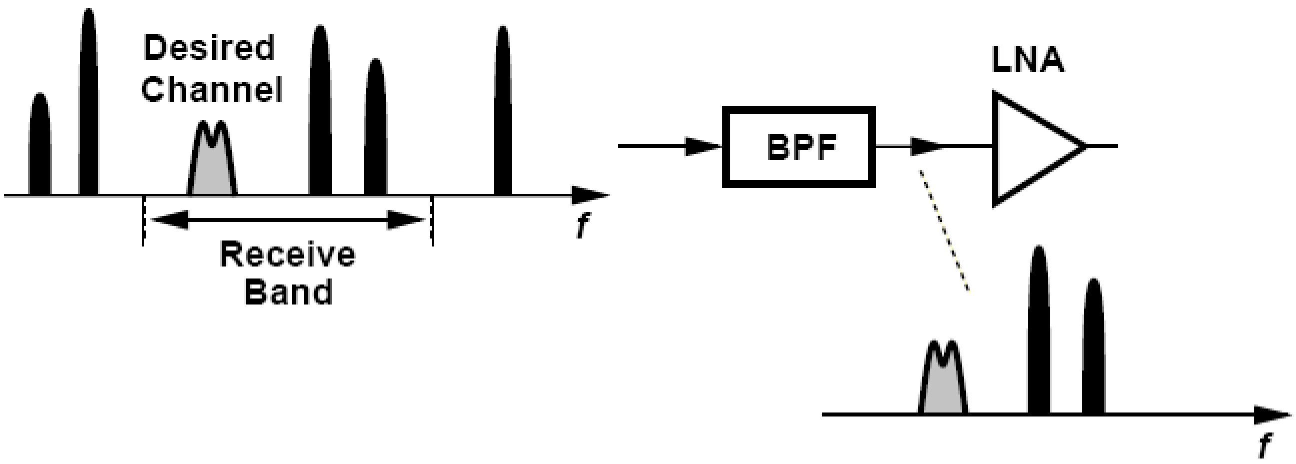
- Transmitter: narrowband modulation, amplification and filtering
- Receiver: process desired channel, while sufficiently rejecting strong neighboring interferers

Front-End Filtering



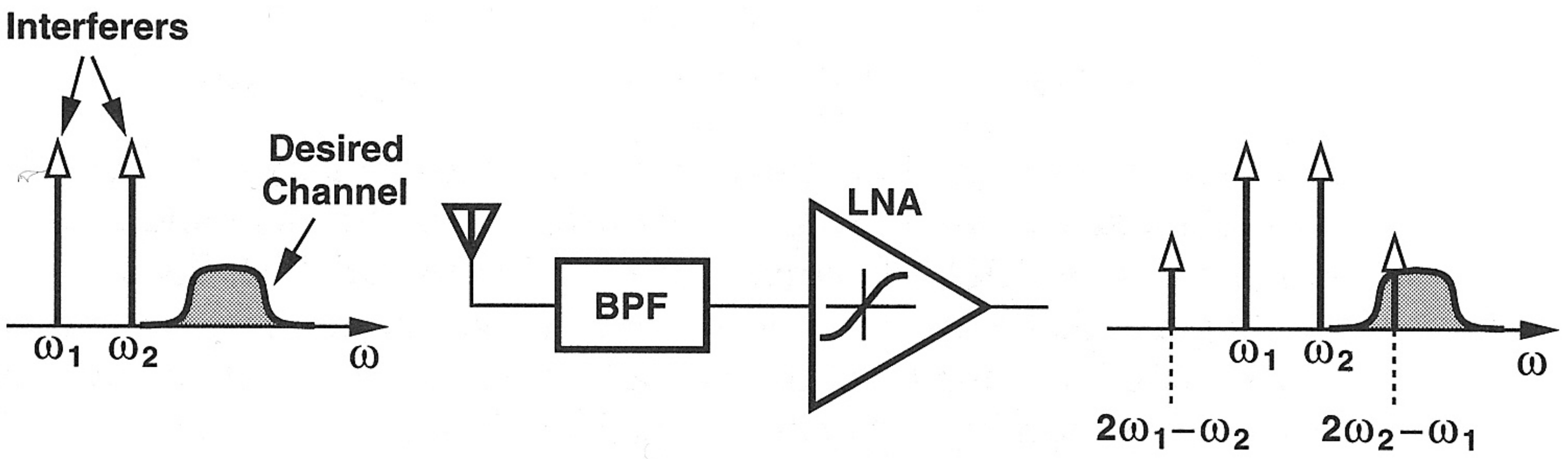
- Assume a 900-MHz GSM receiver with 200-kHz channel spacing must tolerate an alternate adjacent channel blocker 20 dB higher than the desired signal
- The Q of a second-order LC filter required to suppress this interferer by 35 dB is 63'400!
- The required Q is therefore **prohibitively high**
- Can't vary center frequency

Band Selection vs. Channel Selection



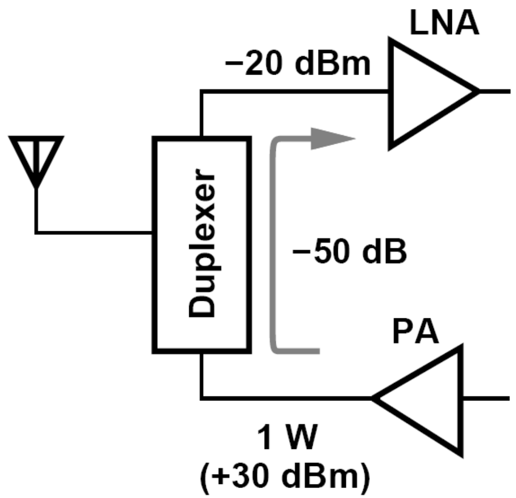
- **Band**: entire spectrum in which users of a particular standard are allowed to communicate (e.g. 935 to 960 MHz for GSM)
- **Channel**: signal bandwidth of only one user in system (e.g. 200 kHz in GSM)
- Channel selection cannot be done at the RF front-end because it would require a prohibitively high-Q, it is therefore postponed to lower center frequency

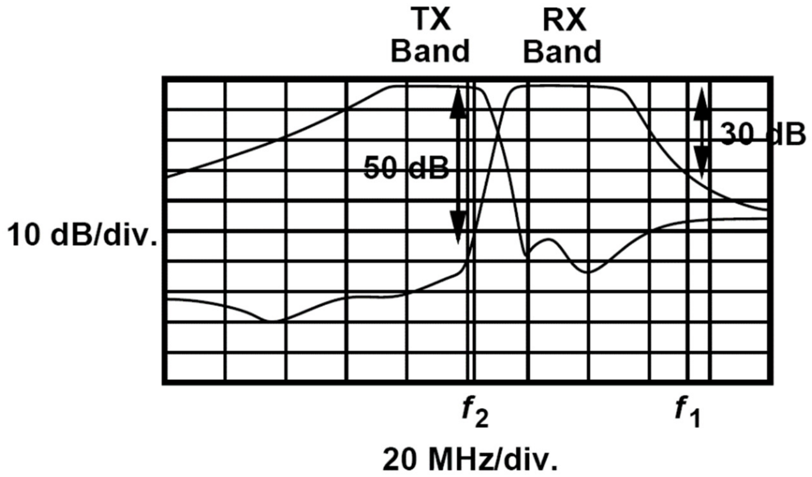
Third-order Nonlinearity



- Odd-order nonlinearities yield **intermodulation** products that **fall in the desired channel**
- Since 3rd-order distortion is usually dominant, the IP3 of each stage must be sufficiently high in order to avoid any corruption of the signal by the intermodulation products

LNA Desensitization by PA



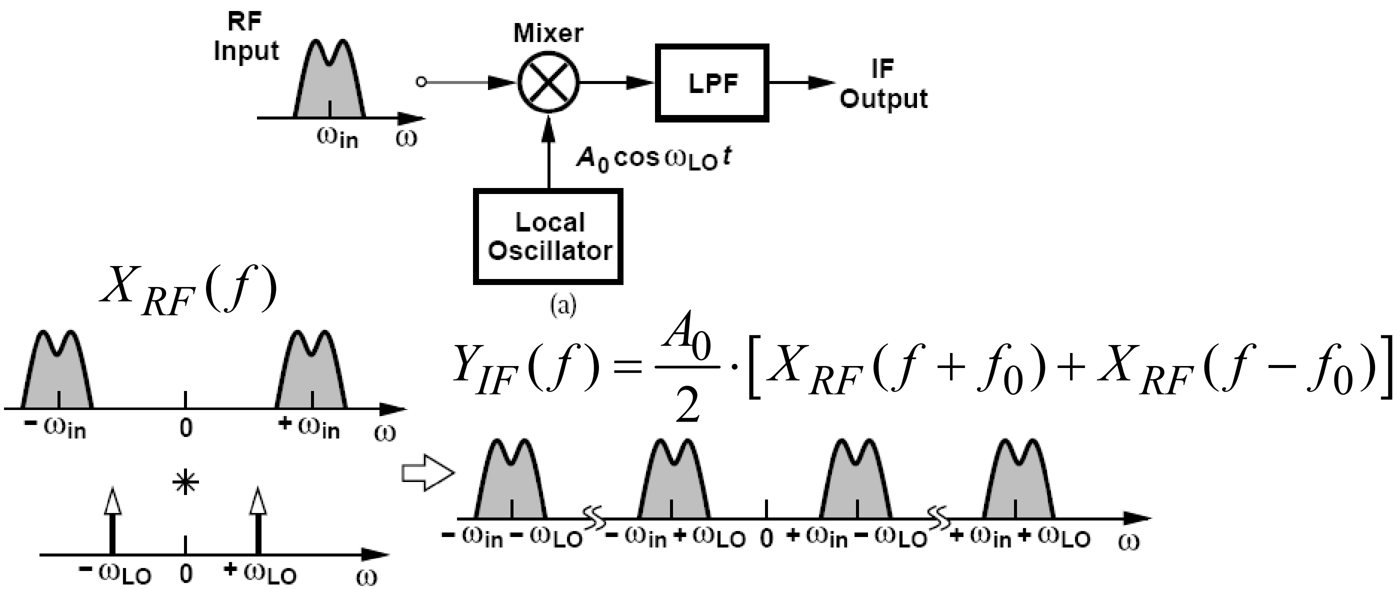


- High signal dynamic range (typically greater than 100 dB)
- If Tx PA delivers 1 W to 50 Ω antenna, the peak-to-peak voltage is equal to 20 V
- The duplexer attenuation is only about 30 dB, so that the leakage to the receive path is on the order of 30 mV_{pp} (≈−26 dBm). Since the LNA 1 dB compression point is in the vicinity of −25 dBm, the leakage signal may already significantly desensitize the LNA
- This problem is avoided if Tx and Rx time slots are offset

Outline

- General Considerations
- **Heterodyne Receivers**
- Homodyne Receivers
- Image-Reject Receivers
- Low-IF Receivers
- Discrete-time Receivers

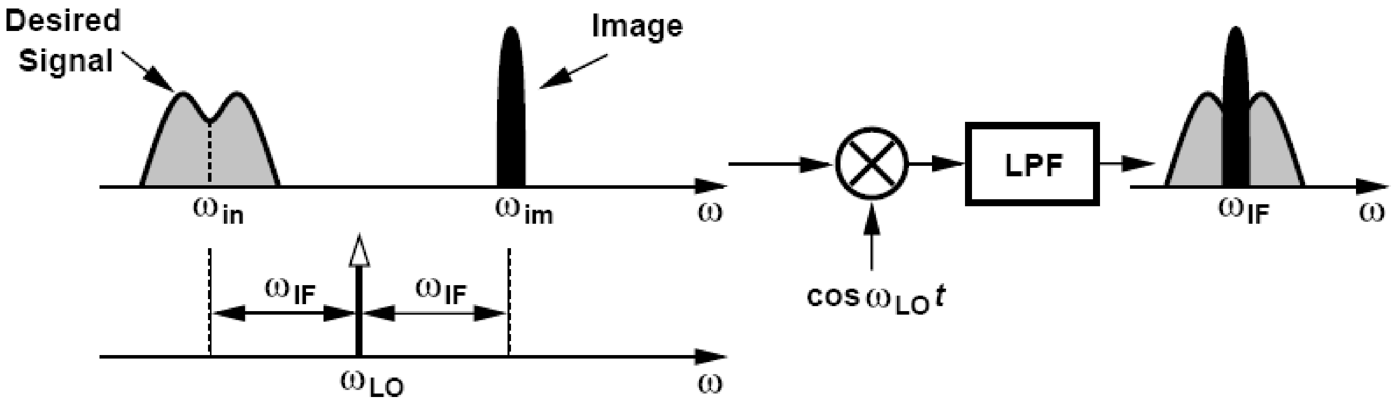
Heterodyne Receivers – Basic Principle



$$Y_{IF}(f) = \frac{A_0}{2} \cdot \left[X_{RF}(f + f_0) + X_{RF}(f - f_0) \right]$$

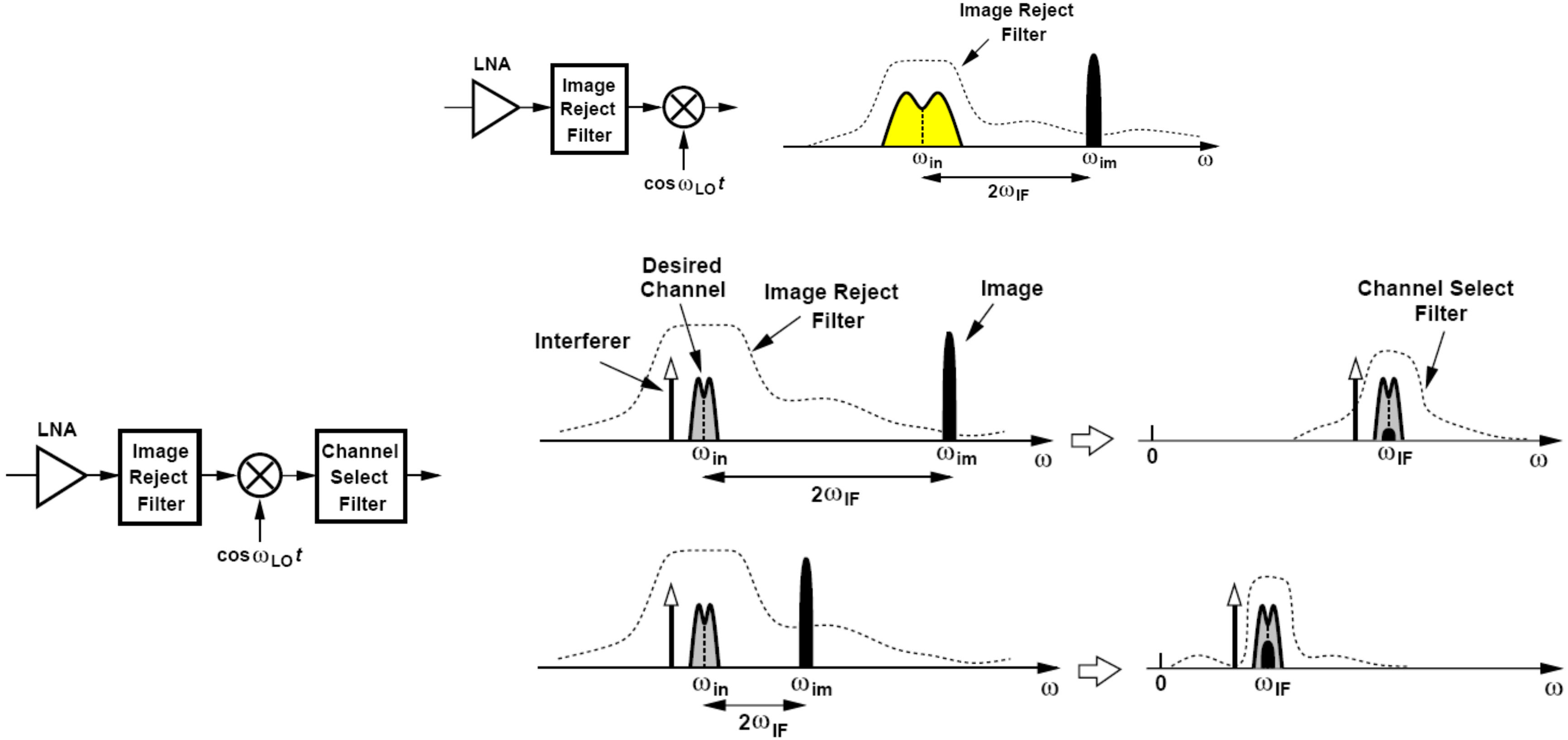
- RF signal transposed to lower frequency (**intermediate frequency IF**) ω_{IF} to avoid high-Q requirement of channel filtering at RF
- Frequency translation obtained by mixing the RF signal centered around ω_{in} with a cosine at ω_{LO}, where $\omega_{LO} = \omega_{in} - \omega_{IF}$ ($\omega_{IF} = \omega_{in} - \omega_{LO}$) yielding a component at $\omega_{IF} = \omega_{in} - \omega_{LO}$ and another one at $\omega_{in} + \omega_{LO} = 2\omega_{in} - \omega_{IF}$
- ω_{LO} is the **local oscillator** (LO) frequency and ω_{IF} is the **intermediate frequency** (IF)

Heterodyne Receivers – Image Problem



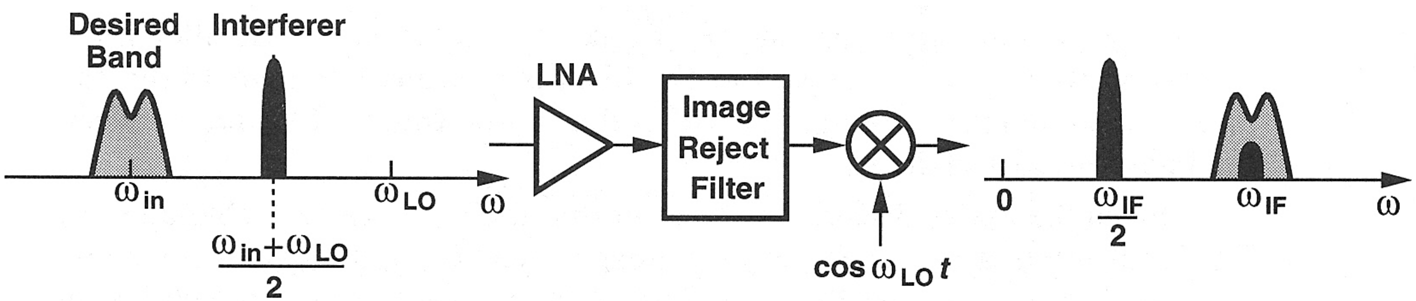
- The bands located at $\omega_{LO} + \omega_{IF}$ and $\omega_{LO} - \omega_{IF}$ are downconverted to the same IF frequency ω_{IF} with no way to distinguish them
- In the case of **high-side injection**, ω_{LO} is at a higher frequency than the band of interest centered around ω_1. In this case $\omega_{IF} = \omega_{LO} - \omega_1$ or $\omega_1 = \omega_{LO} - \omega_{IF}$ and the image around $\omega_{LO} + \omega_{IF} = 2\omega_{LO} - \omega_1$
- In the case of **low-side injection**, ω_{LO} is at a lower frequency than ω_1. In this case $\omega_{IF} = \omega_1 - \omega_{LO}$ or $\omega_1 = \omega_{LO} + \omega_{IF}$ and the image around $\omega_{LO} - \omega_{IF} = 2\omega_{LO} - \omega_1$

Heterodyne Receivers – Image Reject Filter



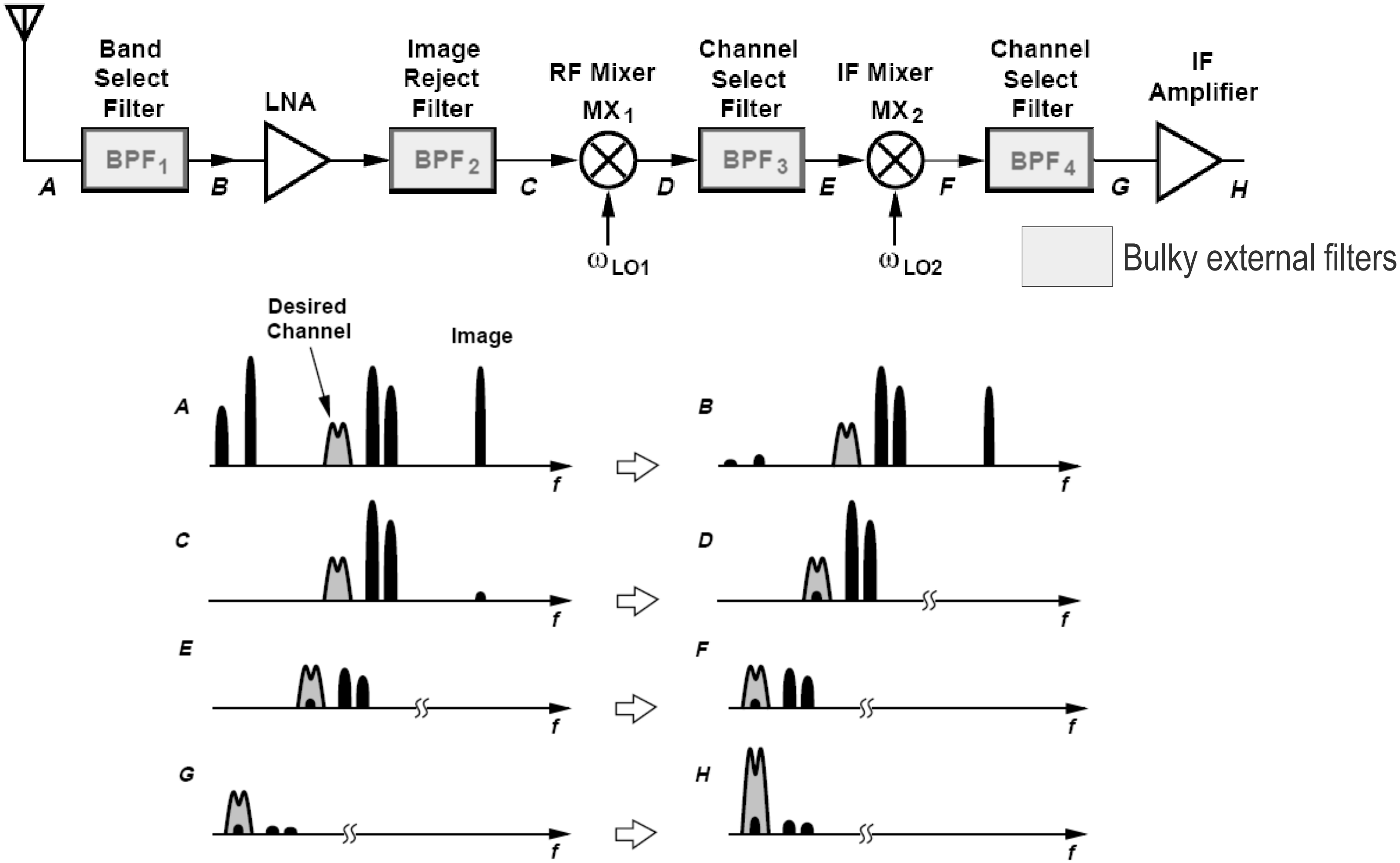
- Choice of the IF is a **trade-off** among the amount of image noise, the spacing between the desired band and the image, and the loss of the image-reject filter
- Basically a trade-off between image rejection (i.e. **sensitivity**) and channel selection (i.e. **selectivity**)

Heterodyne Receivers – Problem of Half Image



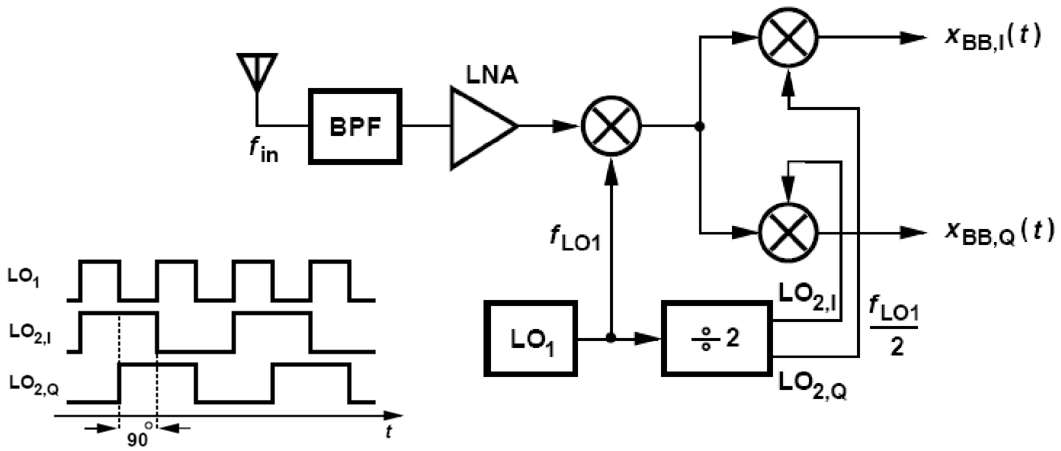
- If in the downconversion path the interferer at $(\omega_{in} + \omega_{LO})/2$ experiences 2nd-order distortion and the LO contains a significant 2nd-order harmonic the IF exhibits a component at $|2\,(\omega_{in} + \omega_{LO})/2 - 2\omega_{LO}| = \omega_{LO} - \omega_{in} = \omega_{IF}$
- Another possibility is that the interferer is translated to $(\omega_{in} - \omega_{LO})/2 = \omega_{IF}/2$ and then undergoes 2nd-order distortion in the baseband causing its 2nd-order harmonic to fall into the downconverted band of interest

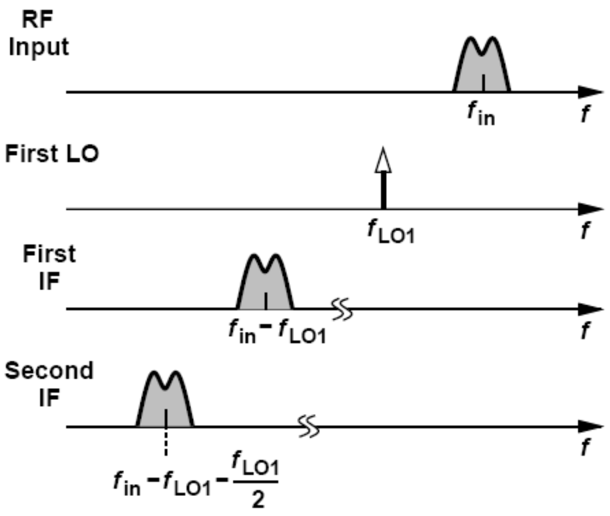
Heterodyne Receivers – Dual-IF Heterodyne Receiver



- Downconversion done in several steps to relax the Q required by each filter

Sliding-IF Receivers





- Modern heterodyne receivers employ only one oscillator
- The second LO frequency is therefore derived from the first by “frequency division”
- Such divide-by-2 topology can produce quadrature output
- The second LO waveforms at a frequency of $f_{LO1}/2$

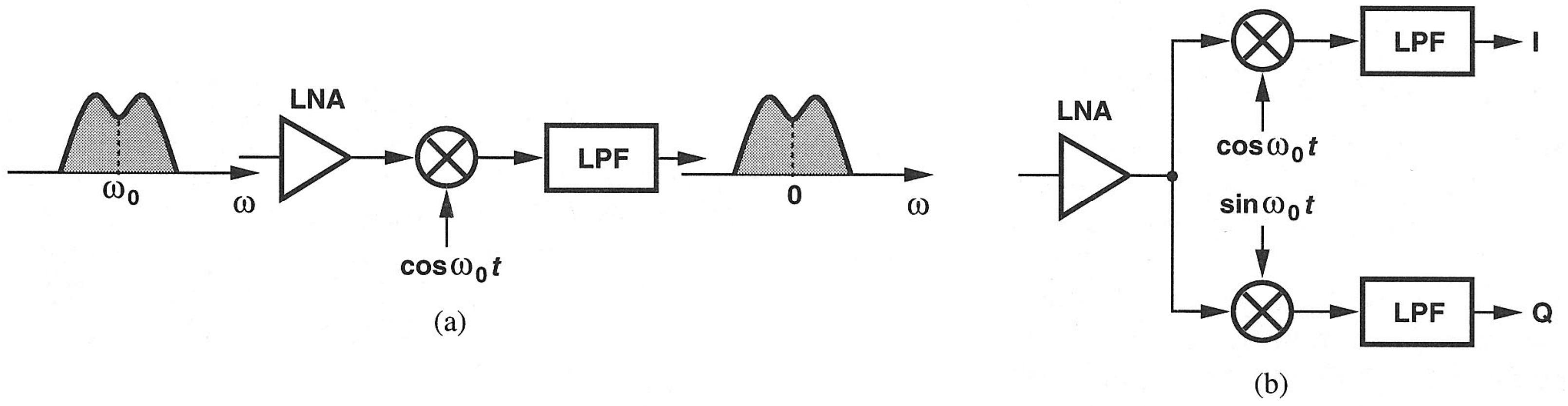
Heterodyne Receivers – Main Issues

- Trade-off between **image rejection** and **channel selection**
- Requires many **external bulky filters** and hence **not well suited for integration**
- LNA and mixers must drive 50 Ω which is **against low power consumption**
- Distribution of gain, noise and nonlinearity along the receive chain is rather complex
- Many spurious components resulting from intermixing of harmonics of interferers and LO signals may corrupt the signal

Outline

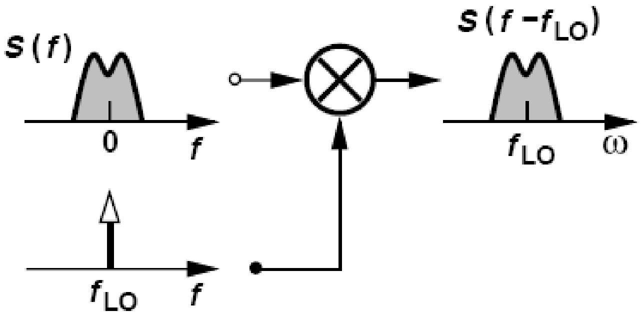
- General Considerations
- Heterodyne Receivers
- **Homodyne Receivers**
- Image-Reject Receivers
- Low-IF Receivers
- Discrete-time Receivers

Homodyne Receivers

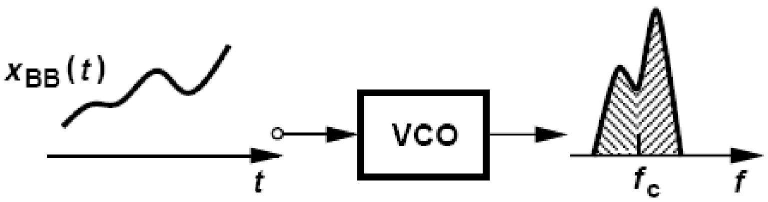


- **Homodyne** receivers are also called **direct conversion** or **zero-IF** receivers because the LO frequency is equal to the carrier frequency converting the RF signal directly to DC
- **No image problem** since $\omega_{IF} = 0$ and therefore no image reject filter required
- In principle, channel selection requires only a low-pass filter and therefore IF filter replaced by low-pass filter making zero-IF receivers **fully integrable on-chip**
- Circuit in Fig. (a) only works with double-sideband AM, circuit in Fig. (b) is required for frequency and phase-modulated signals and requires quadrature for avoiding loss of information carried by the phase

Symmetrically- versus Asymmetrically-modulated Signal



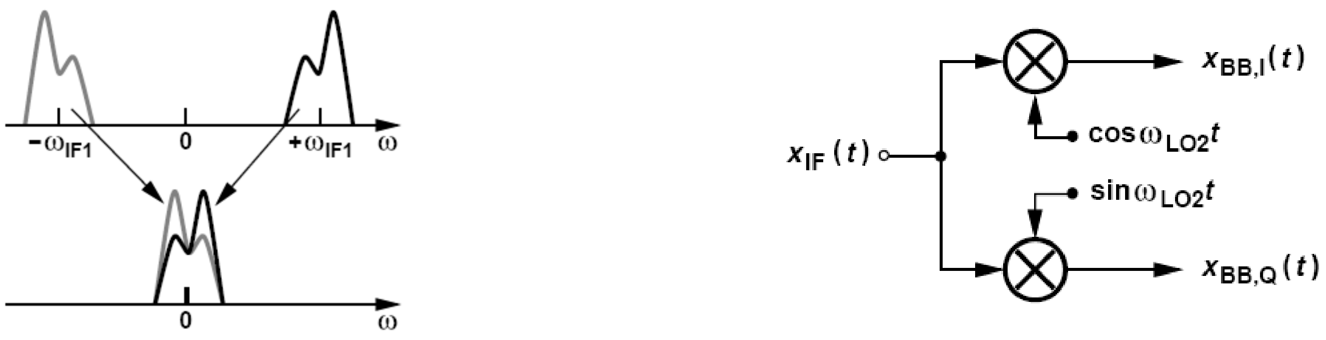
AM signal generation

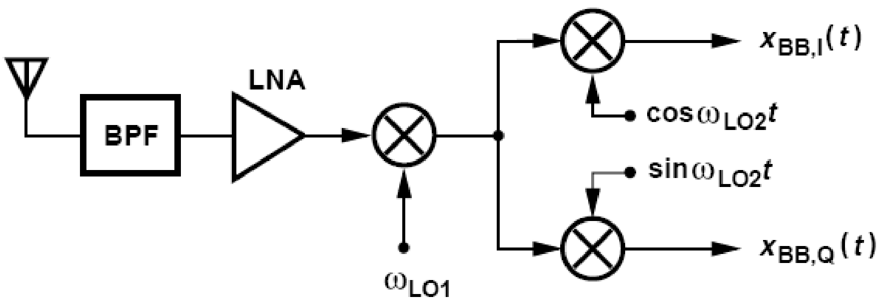


FM signal generation

- 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

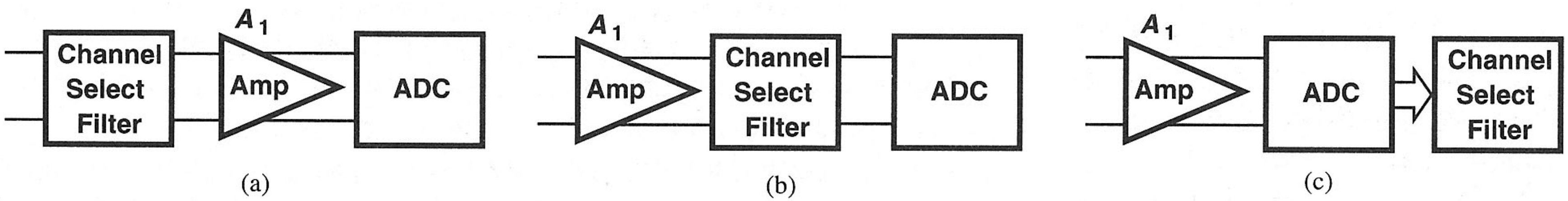
Corruption of the Asymmetric Signal Spectrum





- Downconversion to a zero IF superimposes two copies of the signal
- If the signal spectrum is **asymmetric**, which is the case for most frequency and phase modulations, the original signal spectrum will be **corrupted**
- This can be avoided by creating two versions of the downconverted signal that are in **quadrature**

Homodyne Receivers – Channel Selection

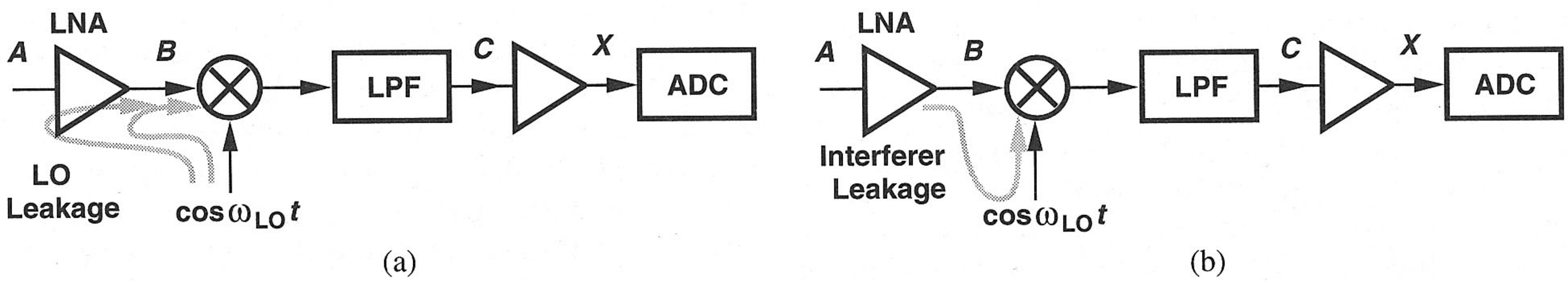


- LPF suppresses out-of-channel interferers
- A_1 can be a nonlinear but high-gain amplifier
- ADC with moderate dynamic range (4 to 10 bits)
- However, LPF imposes tight noise-linearity trade-offs

- Relaxes the LPF noise requirements
- Demanding a higher performance in the amplifier
- Another amplifier can be added between the LPF and the ADC to overcome the noise of the ADC

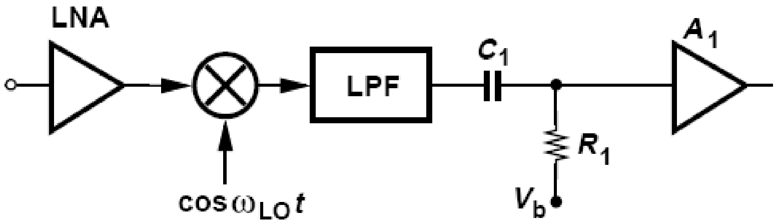
- All the channel filtering is moved to the digital domain
- High dynamic range and highly linear ADC required in order to digitize the signal with minimal inter-modulation of interferers
- ADC thermal and quantization noise floor well below the signal level

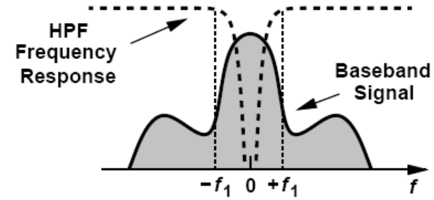
Homodyne Receivers – DC Offset

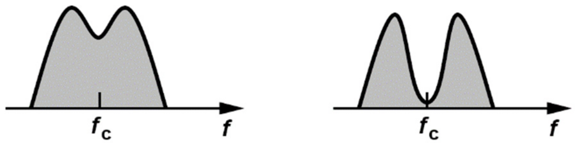


- LO leakage from the LO to point A and B is **self-mixed** with the LO signal producing a **DC component** at point C (Fig. (a))
- Similar effect if large interferer leaks from the LNA or mixer input to the LO port and is multiplied by itself (Fig. (b))
- The total gain from the antenna to point X is very large (80 to 100dB) from which typically 20 to 30 dB is contributed by the LNA-mixer combination
- If the LO signal is 0.63V_{pp} (≈0 dBm) and the attenuation from LO to point A is 60 dB, assuming a gain of the LNA and mixer of 30 dB, the DC offset at the mixer output is about 10 mV while the desired signal can typically be 30 µV_{rms}. If the remaining gain is 50 to 70 dB, the **offset will saturate the following stages**

Homodyne Receivers – DC Offset Cancellation



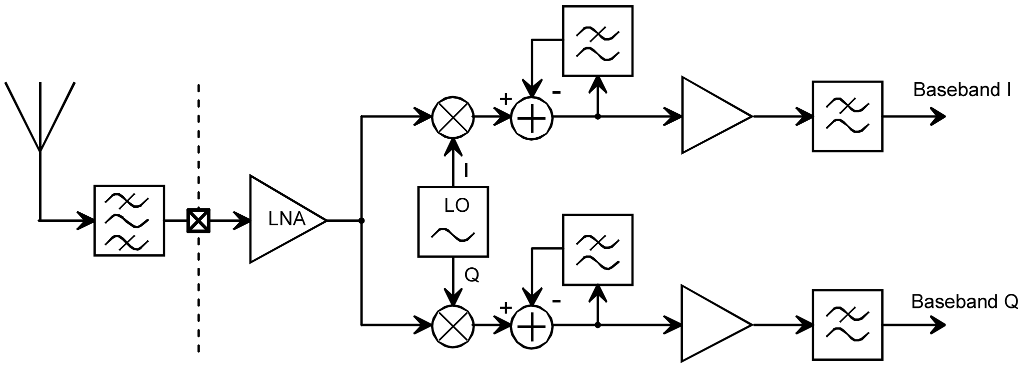


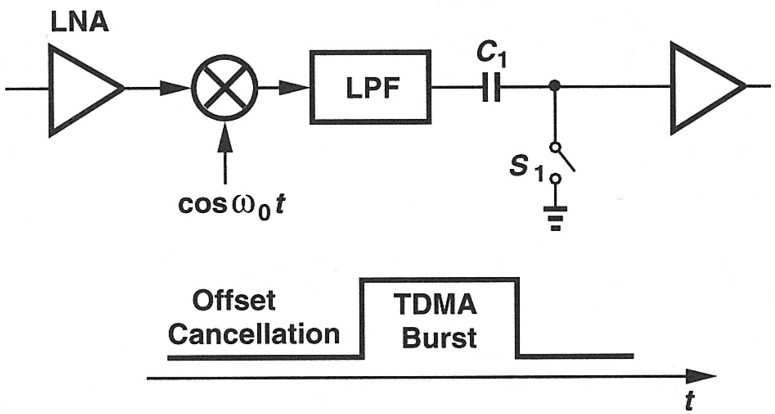


Small modulation index Large modulation index

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

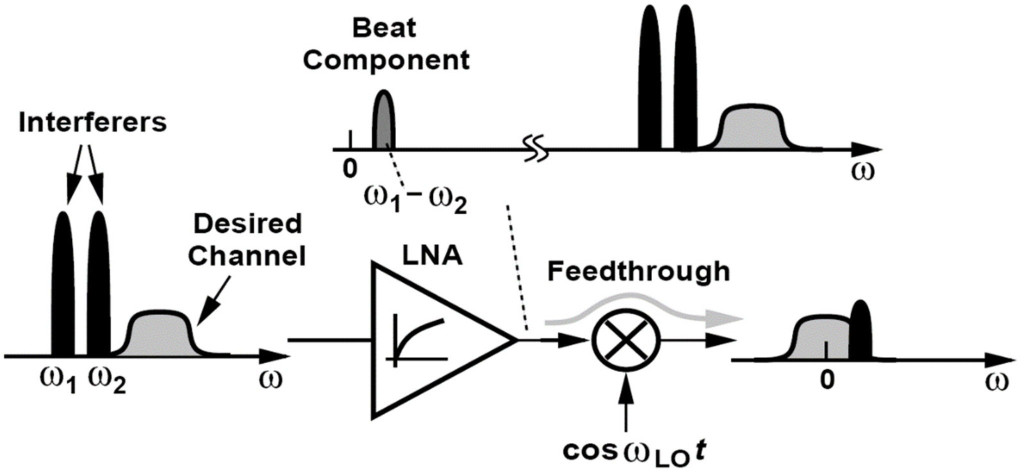
Homodyne Receivers – DC Offset Cancellation





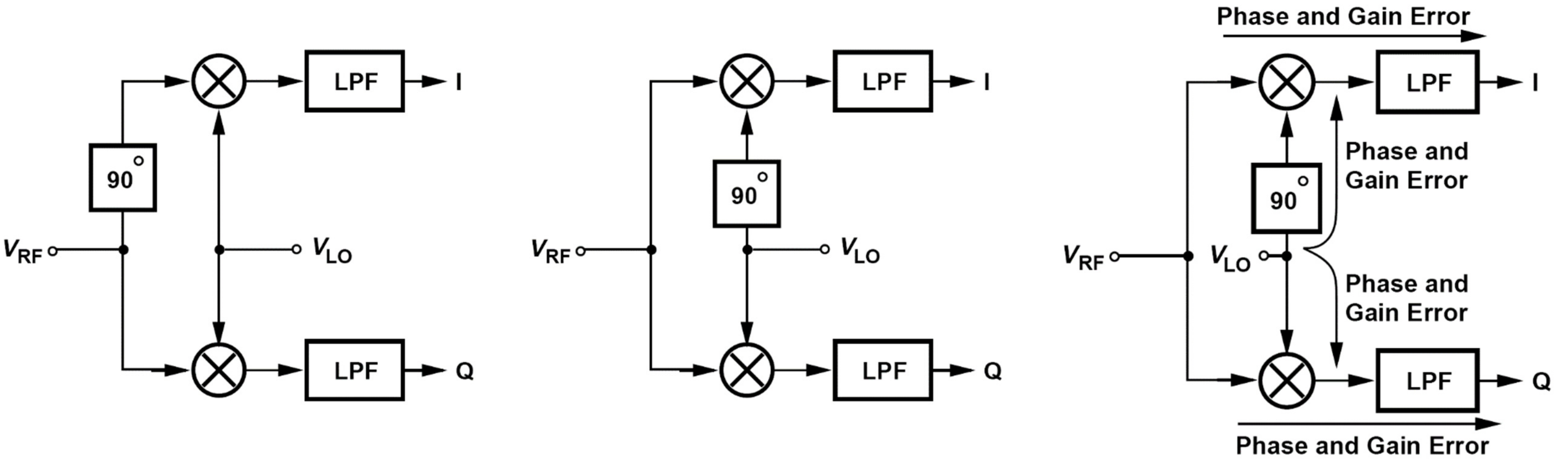
- DC offset can be cancelled by high-pass filtering (capacitive coupling being the simplest implementation)
- Spectra of many modulated signals exhibit a peak at zero frequency imposing a low cut-off frequency (of the high-pass filter) and hence a very large coupling capacitance or a high-order high-pass filter which is not always acceptable for integration
- This method also fails at tracking large variations of the DC offset voltage
- Can be alleviated by
  - Encoding the signal at the transmitter such that the modulated signal after downconversion only contains little energy at DC
  - In TDMA system by calibrating the offset between consecutive bursts (offset cancellation as shown in right Fig.)

Homodyne Receivers – Even-Order Distortion



- Two interferers close to the channel $A_1 \cos(\omega_1 t)$ and $A_2 \cos(\omega_2 t)$ experience a 2nd-order nonlinearity
- The output signal $y(t) = \alpha_1 x(t) + \alpha_2 x^2(t)$ contains a term $\alpha_2 A_1 A_2 \cos\big((\omega_1 - \omega_2)t\big)$
- Upon multiplication by $\cos(\omega_{LO} t)$ in an ideal mixer, this term is translated to high frequencies
- Now, because of feed through, this component might fall in the desired channel

Homodyne Receivers – I/Q Mismatch



$$x_{in}(t) = a\cos(\omega_c t) + b\sin(\omega_c t)$$

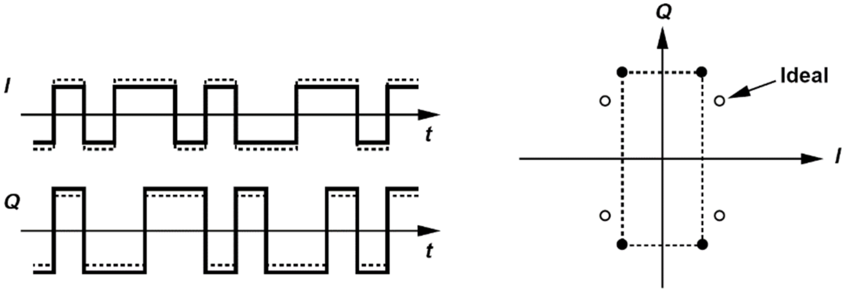
$$x_{LO,I}(t) = 2\left(1+\frac{\varepsilon}{2}\right)\cos\left(\omega_c t + \frac{\theta}{2}\right)$$

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

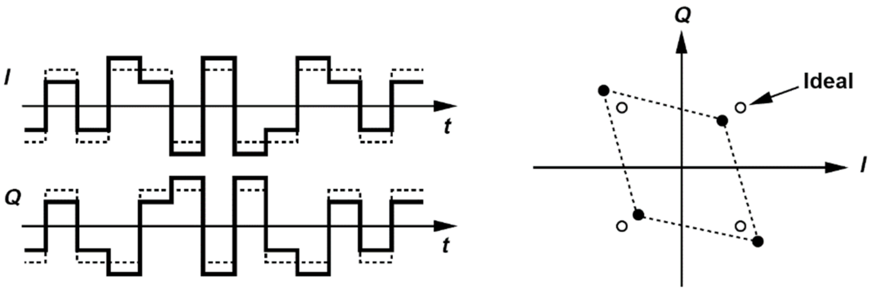
$$x_{BB,I}(t) = a\left(1+\frac{\varepsilon}{2}\right)\cos\frac{\theta}{2} - b\left(1+\frac{\varepsilon}{2}\right)\sin\frac{\theta}{2}$$

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

Amplitude mismatch only ($\varepsilon \neq 0$, $\theta = 0$)



Phase mismatch only ($\varepsilon = 0$, $\theta \neq 0$)



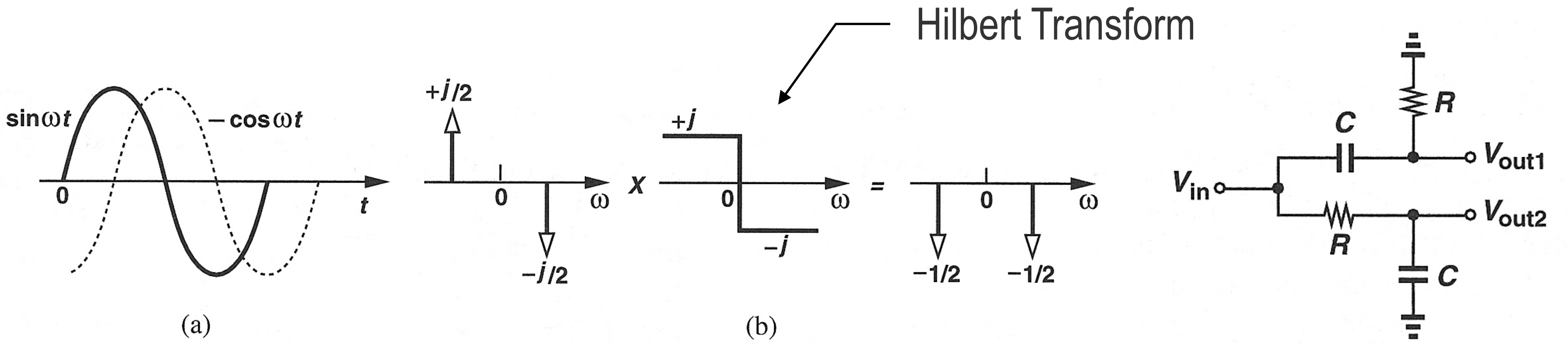
Homodyne Receivers – Flicker Noise and LO Leakage

- The gain of the RF front-end (LNA + mixer) is typically in the order of 30 dB
- The **noise** of the following stages, e.g. filter and amplifiers, is therefore still **critical**
- Since the signal is downconverted around DC, the **flicker noise** may substantially corrupt the signal, particularly in deep-submicron MOS implementation, where flicker noise dominates at low frequency and is a severe problem
- **LO leakage** is another issue, since in homodyne receivers the LO is set to the carrier frequency, leakage of the LO from the oscillator to the antenna and radiation may create interferers in the band of other receivers using the same wireless standard. The design of the wireless standard and regulation impose upper bounds for the in-band LO radiation in the order of −80 dBm to −50 dBm

Outline

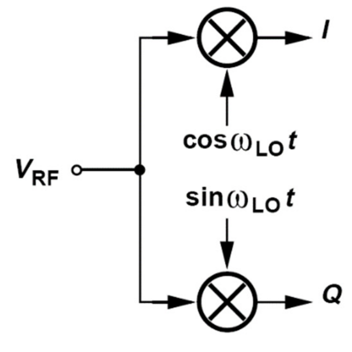
- General Considerations
- Heterodyne Receivers
- Homodyne Receivers
- **Image-Reject Receivers**
- Low-IF Receivers
- Discrete-time Receivers

Image-Reject Receivers – Quadrature Generation



- The basic idea of image-reject receivers is to **process the signal and its image differently** allowing cancellation of the image by its negated replica
- Image reject receivers are based on **quadrature signals**
- Such signals can be generated from a single signal by shifting it by 90° within a band of interest (typically the signal bandwidth) (Hilbert transform)
- This operation can be done for example by the circuit in the right Fig., but it is **only valid at a single frequency** given by 1/(RC)

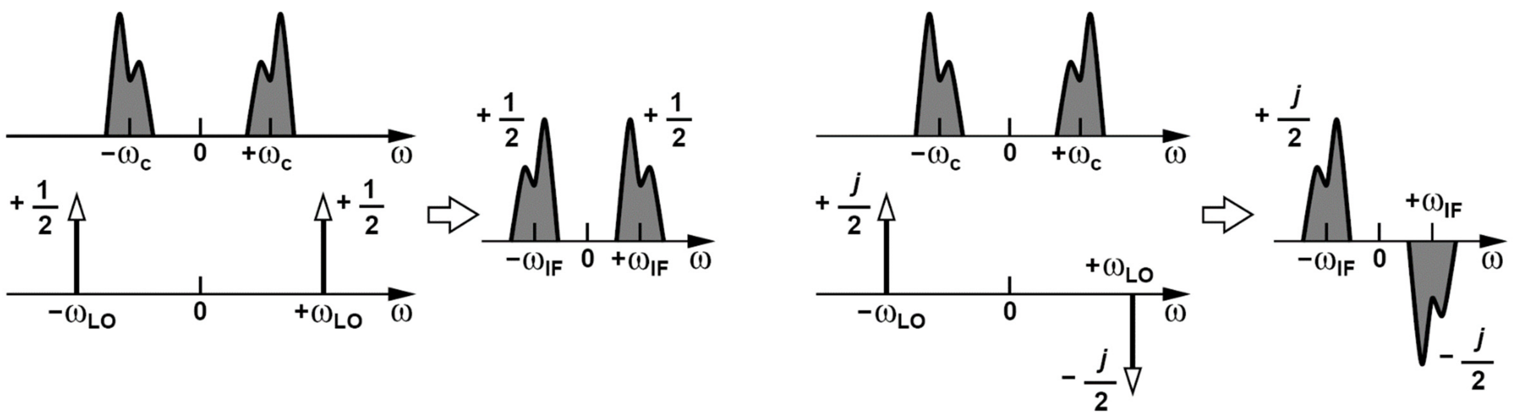
Quadrature Downconversion



Upper arm (I signal path)

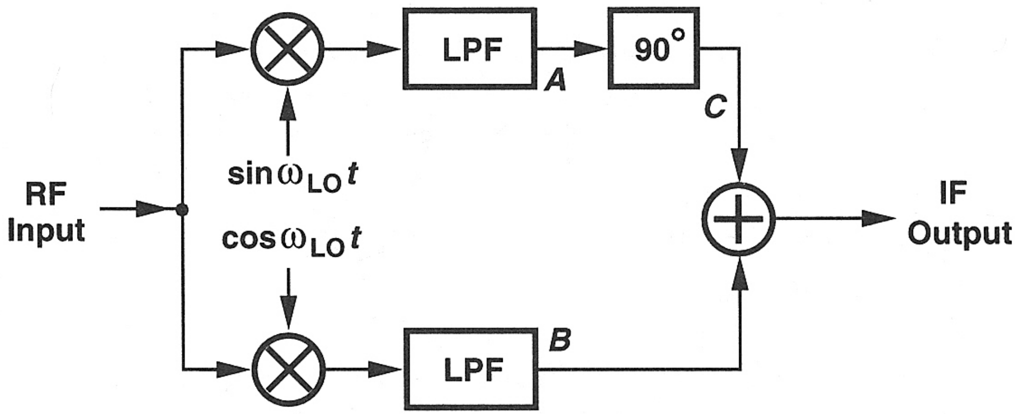
Lower arm (Q signal path)

(a)



- The RF input is mixed with the quadrature phases of the LO so as to translate the spectrum to a non-zero IF (the above figure assumes high side injection $\omega_c < \omega_{LO}$)
- The IF spectrum emerging from the lower arm is the Hilbert transform of that from the upper arm

The Hartley Architecture – Principle



Assuming low-side injection:

$$\omega_{IF} = \omega_{RF} - \omega_{LO} = \omega_{LO} - \omega_{im}$$

$$x(t) = A_{RF}\cos(\omega_{RF}t) + A_{im}\cos(\omega_{im}t)$$

Point A: $x_A(t) = \frac{A_{RF}}{2}\sin\left[(\omega_{LO} - \omega_{RF})t\right] + \frac{A_{im}}{2}\sin\left[(\omega_{LO} - \omega_{im})t\right]$

$= -\frac{A_{RF}}{2}\sin\left[(\omega_{RF} - \omega_{LO})t\right] + \frac{A_{im}}{2}\sin\left[(\omega_{LO} - \omega_{im})t\right]$

Point B: $x_B(t) = \frac{A_{RF}}{2}\cos\left[(\omega_{LO} - \omega_{RF})t\right] + \frac{A_{im}}{2}\cos\left[(\omega_{LO} - \omega_{im})t\right]$

Point C: $x_C(t) = +\frac{A_{RF}}{2}\cos\left[(\omega_{RF} - \omega_{LO})t\right] - \frac{A_{im}}{2}\cos\left[(\omega_{LO} - \omega_{im})t\right]$

Signal at B and C have same polarity whereas image have opposite polarity

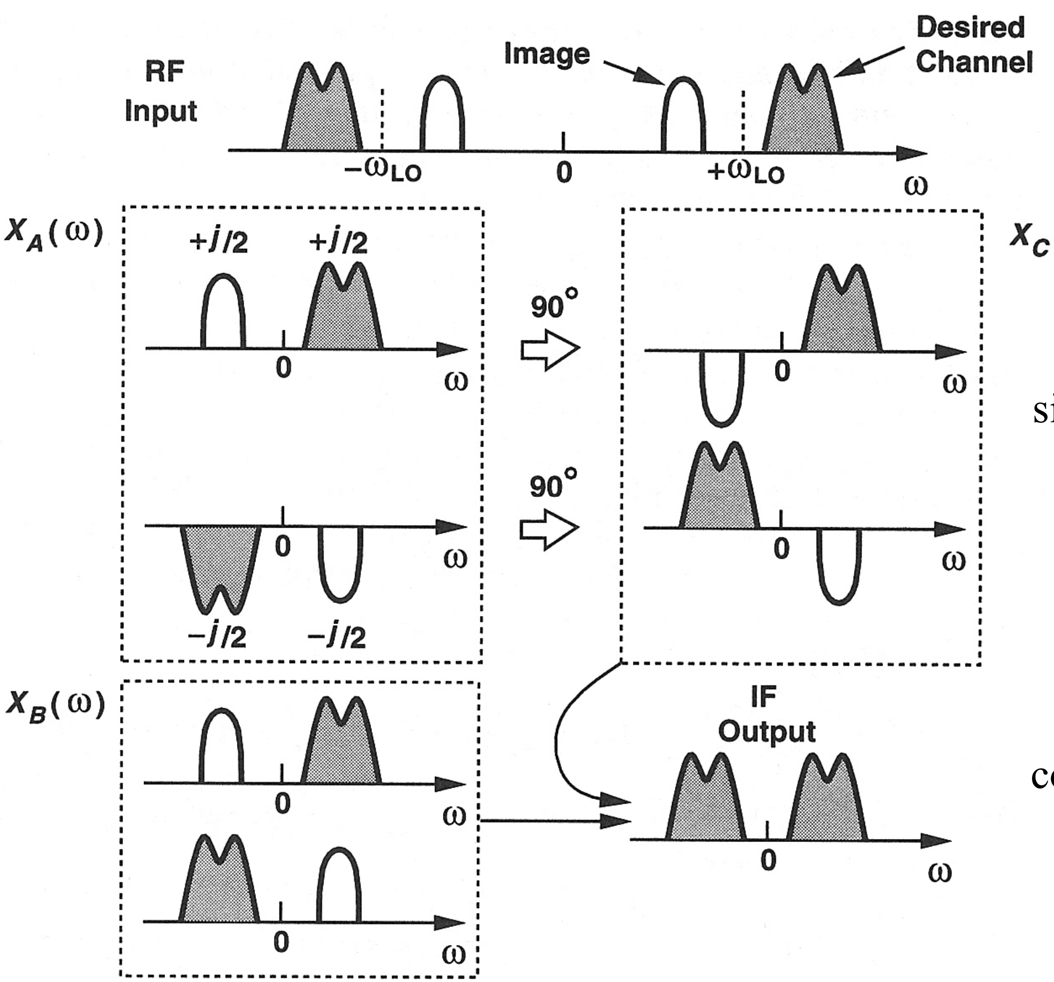
$$\sin\alpha\cdot\cos\beta = \frac{1}{2}\sin(\alpha-\beta) + \frac{1}{2}\sin(\alpha+\beta)$$

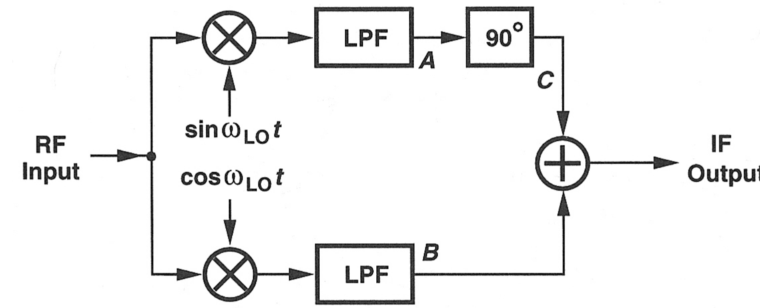
$$\cos\alpha\cdot\cos\beta = \frac{1}{2}\cos(\alpha-\beta) + \frac{1}{2}\cos(\alpha+\beta)$$

Finally we get: $x_B(t) + x_C(t) = A_{RF}\cos\left[(\omega_{LO} - \omega_{RF})t\right]$

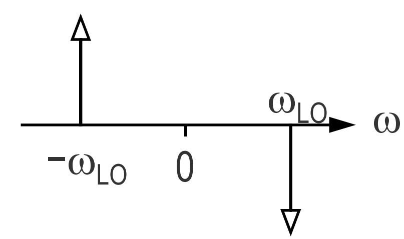
The Hartley Architecture – Graphical Analysis

- Hartley's circuit can also be analyzed graphically

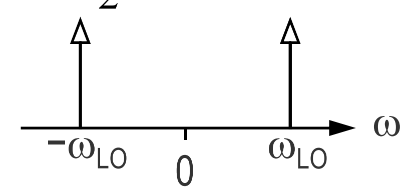




$$\sin(2\pi f_{LO} t) \leftrightarrow \frac{j}{2}\left[\delta(f+f_{LO})-\delta(f-f_{LO})\right]$$



$$\cos(2\pi f_{LO} t) \leftrightarrow \frac{1}{2}\left[\delta(f+f_{LO})+\delta(f-f_{LO})\right]$$



The Hartley Architecture – Limitations

- The main drawback of the Hartley architecture is its **sensitivity to amplitude and phase mismatches**
- Assuming the LO signals are not exactly in quadrature and have slightly different amplitude

$$x_{LO,I}(t) = A_{LO}\sin(\omega_{LO}t) \quad \text{and} \quad x_{LO,Q}(t) = A_{LO}(1+\varepsilon)\sin(\omega_{LO}t+\theta)$$

- The downconverted desired signal and the image are then given by

$$x_{sig}(t) = \frac{A_{LO}(1+\varepsilon)A_{RF}}{2}\cos[(\omega_{LO}-\omega_{RF})t+\theta] + \frac{A_{LO}A_{RF}}{2}\cos[(\omega_{LO}-\omega_{RF})t]$$

$$x_{im}(t) = \frac{A_{LO}(1+\varepsilon)A_{im}}{2}\cos[(\omega_{LO}-\omega_{im})t+\theta] - \frac{A_{LO}A_{im}}{2}\cos[(\omega_{LO}-\omega_{im})t]$$

- The image and signal average powers are then given by

$$P_{im} = \frac{A_{im}^2 A_{LO}^2}{8}\cdot\left[(1+\varepsilon)^2 - 2(1+\varepsilon)\cos\theta + 1\right]$$

$$P_{sig} = \frac{A_{RF}^2 A_{LO}^2}{8}\cdot\left[(1+\varepsilon)^2 + 2(1+\varepsilon)\cos\theta + 1\right]$$

The Hartley Architecture – Image Rejection Ratio

- The image-to-signal power ratio is given by

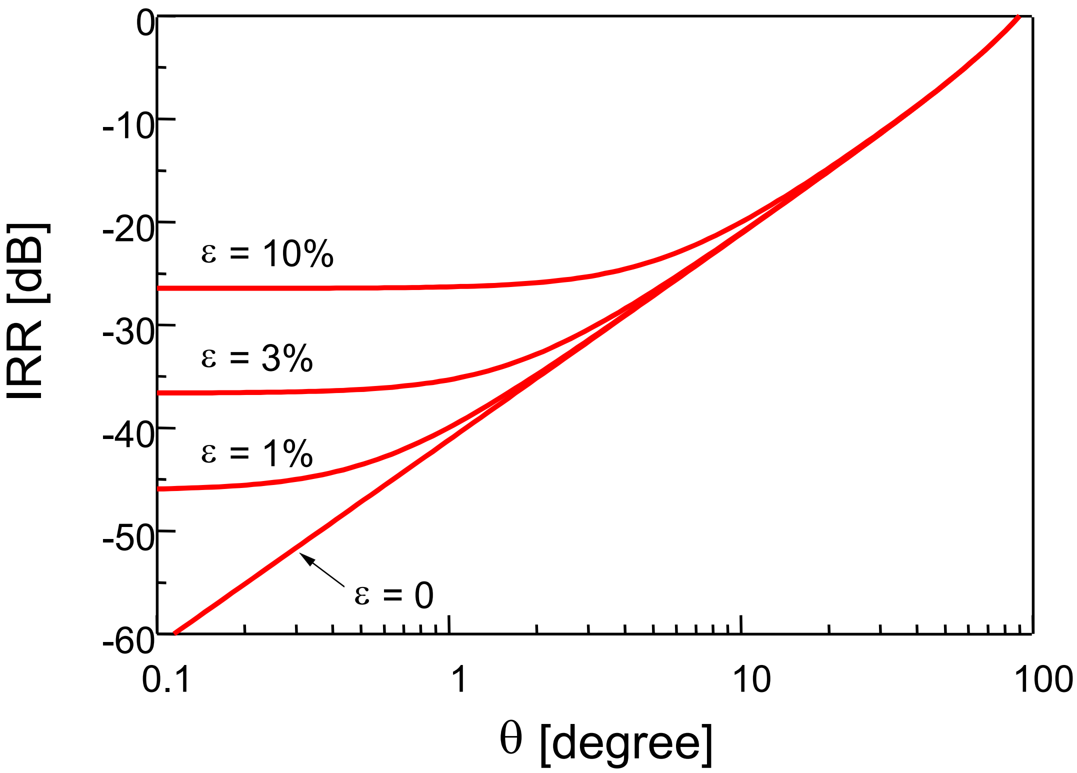
$$\left.\frac{P_{im}}{P_{sig}}\right|_{out} = \frac{A_{im}^2}{A_{RF}^2} \cdot \frac{(1+\varepsilon)^2 - 2(1+\varepsilon)\cos\theta + 1}{(1+\varepsilon)^2 + 2(1+\varepsilon)\cos\theta + 1}$$

- Noting that A_{im}^2/A_{RF}^2 is the image-to-signal power ratio at the input, we define the **image rejection ratio** (IRR) as

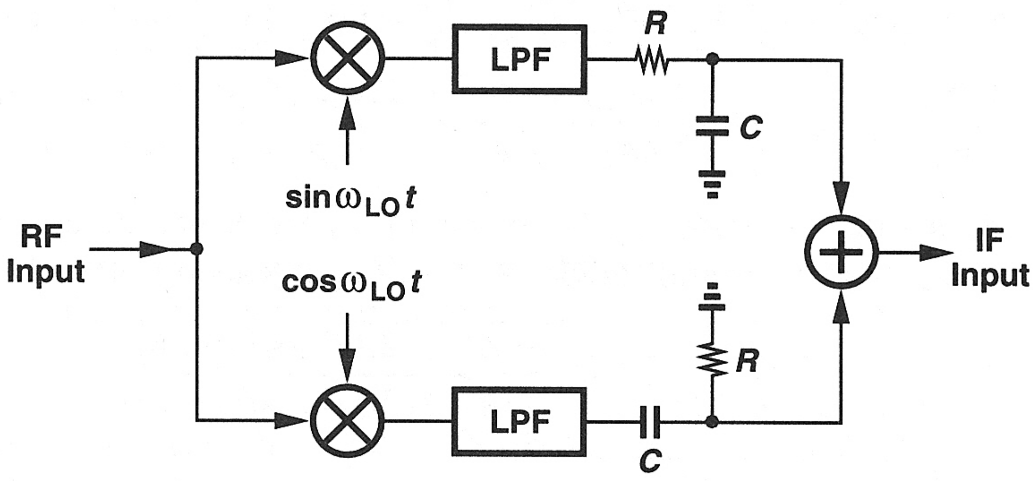
$$IRR \triangleq \frac{\left.\frac{P_{im}}{P_{sig}}\right|_{out}}{\left.\frac{P_{im}}{P_{sig}}\right|_{in}} = \frac{\left.\frac{P_{im}}{P_{sig}}\right|_{out}}{\frac{A_{im}^2}{A_{RF}^2}} = \frac{(1+\varepsilon)^2 - 2(1+\varepsilon)\cos\theta + 1}{(1+\varepsilon)^2 + 2(1+\varepsilon)\cos\theta + 1} \cong \frac{\varepsilon^2 + \theta^2}{4}$$

The Hartley Architecture – Image Rejection Ratio

$$IRR = \frac{(1+\varepsilon)^2 - 2(1+\varepsilon)\cos\theta + 1}{(1+\varepsilon)^2 + 2(1+\varepsilon)\cos\theta + 1} \cong \frac{\varepsilon^2 + \theta^2}{4}$$



The Hartley Architecture – Image Rejection Ratio



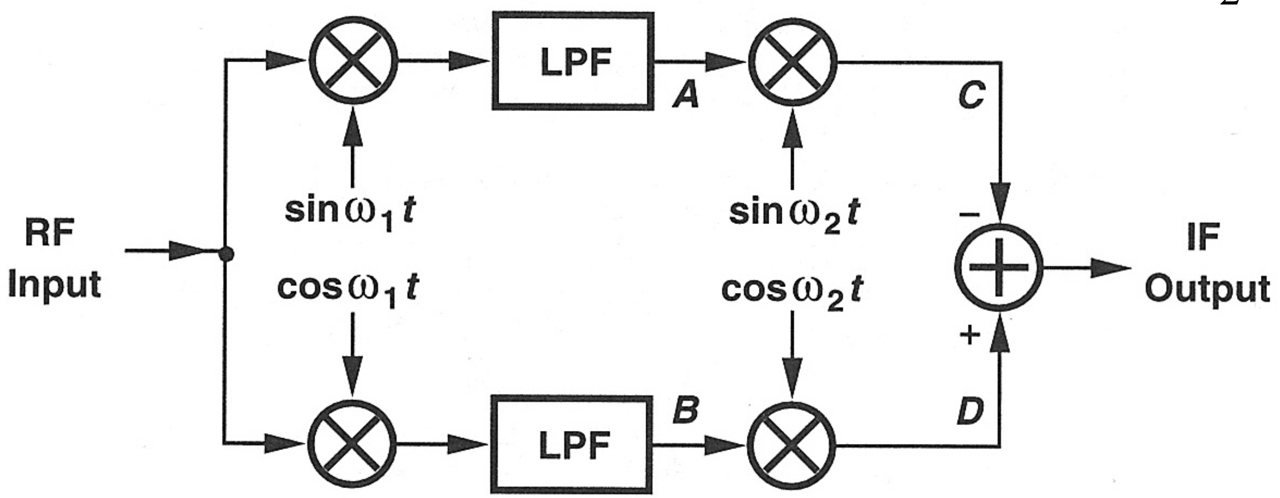
Amplitude mismatch due to R and C mismatch between I and Q path approximately given by

$$\frac{\Delta A}{A} \cong \frac{\Delta R}{R} + \frac{\Delta C}{C}$$

- RC-CR network limits image rejection to 20 dB if RC varies by 20%
- Low-IF requires large fractional bandwidth and hence RC-CR network provides inadequate IRR
- Replace RC-CR network with **multisection polyphase RC filter**, with sections “staggered” to provide **broadband image rejection**
- But each section introduces **3 dB of loss**

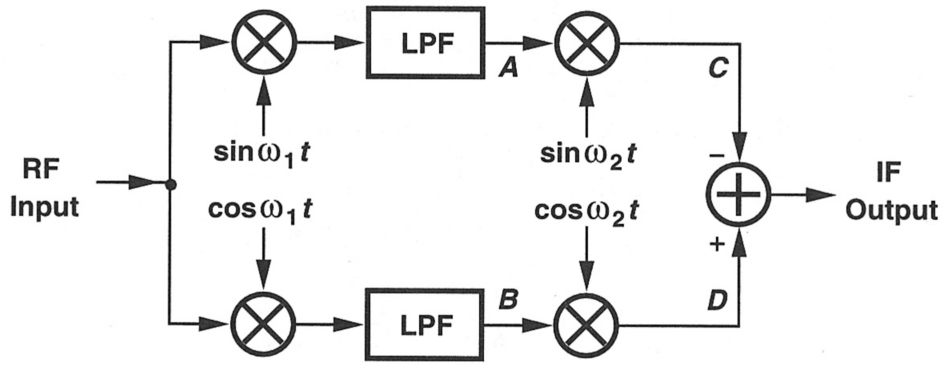
The Weaver Architecture – Principle

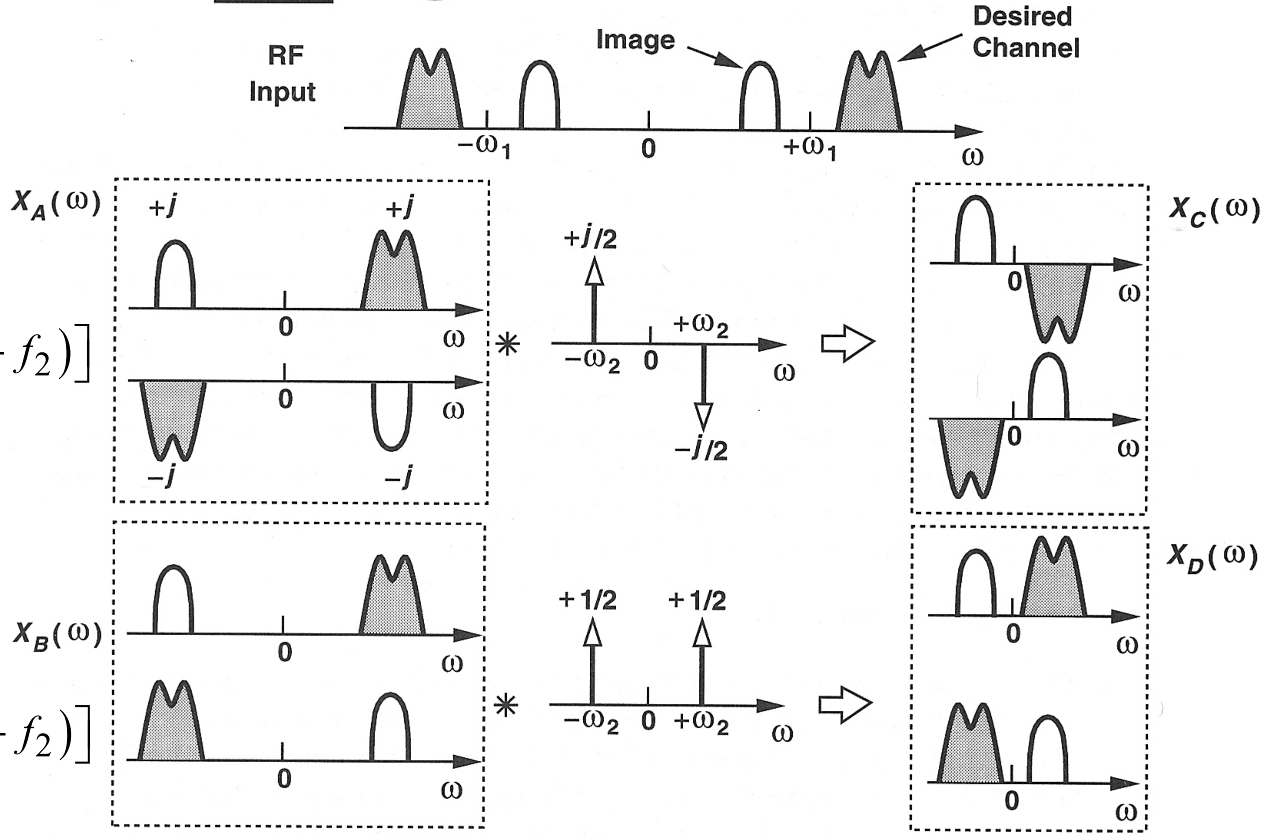
$\omega_2 \ll \omega_1$



- 90º phase shift of the Hartley image reject mixer is replaced by a second **image rejection mixer**

The Weaver Architecture – Graphical Analysis

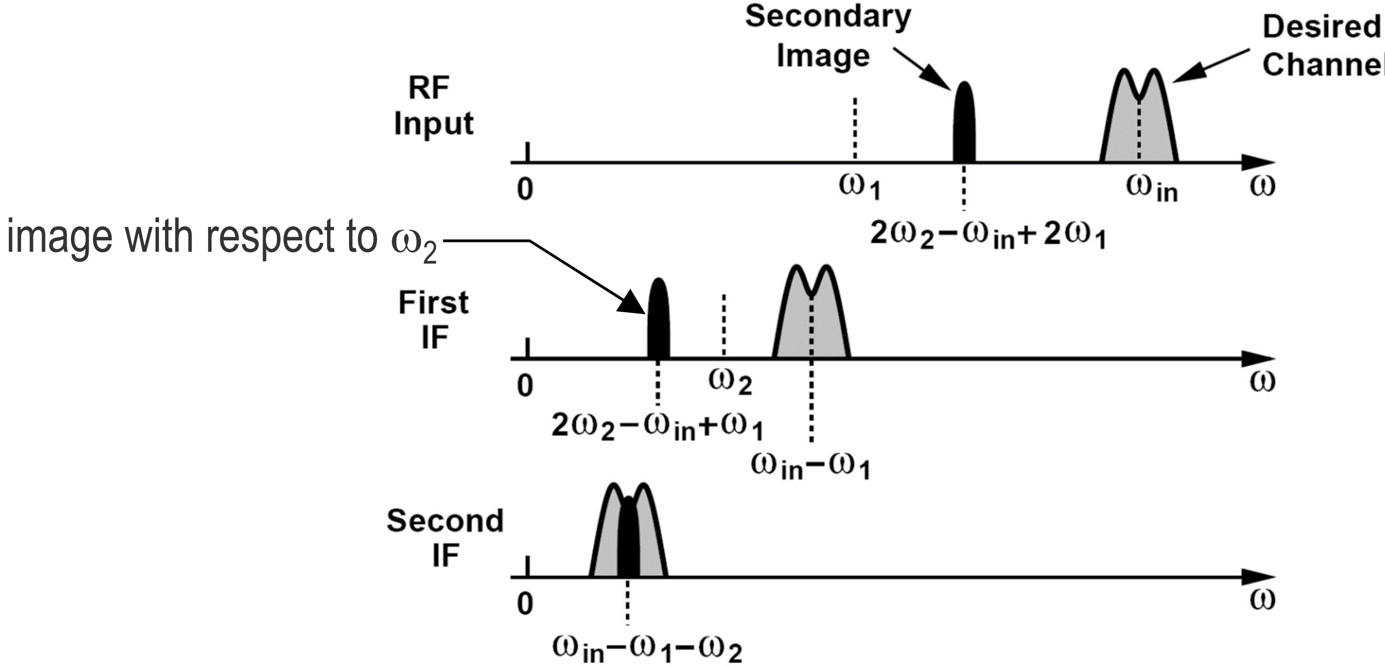




$$\sin(2\pi f_2 t) \leftrightarrow \frac{j}{2}\left[\delta(f+f_2) - \delta(f-f_2)\right]$$

$$\cos(2\pi f_2 t) \leftrightarrow \frac{1}{2}\left[\delta(f+f_2) + \delta(f-f_2)\right]$$

The Weaver Architecture – Secondary Image Problem

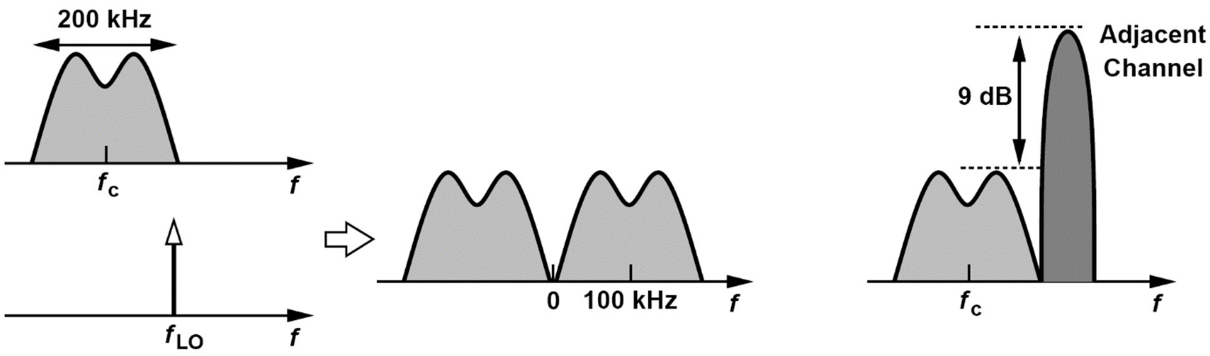


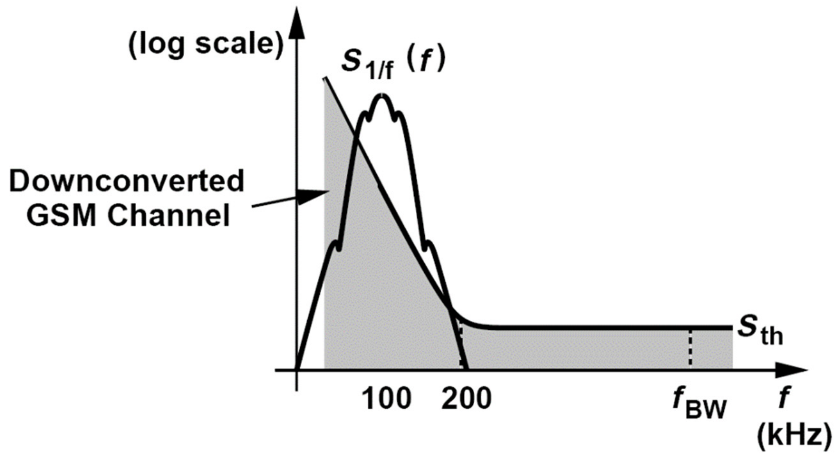
- Second mixing entails the problem of secondary image
- Interferer at $2\omega_2 - \omega_{in} + 2\omega_1$ is first downconverted to $2\omega_2 - \omega_{in} + \omega_1$ and then to $\omega_{in} - \omega_1 - \omega_2$ which is in the band of interest
- The LPF filter at the first IF ω_2 must therefore be replaced by a BPF
- Or the second IF can be set to zero (direct conversion) avoiding any secondary image

Outline

- General Considerations
- Heterodyne Receivers
- Homodyne Receivers
- Image-Reject Receivers
- **Low-IF Receivers**
- Discrete-time Receivers

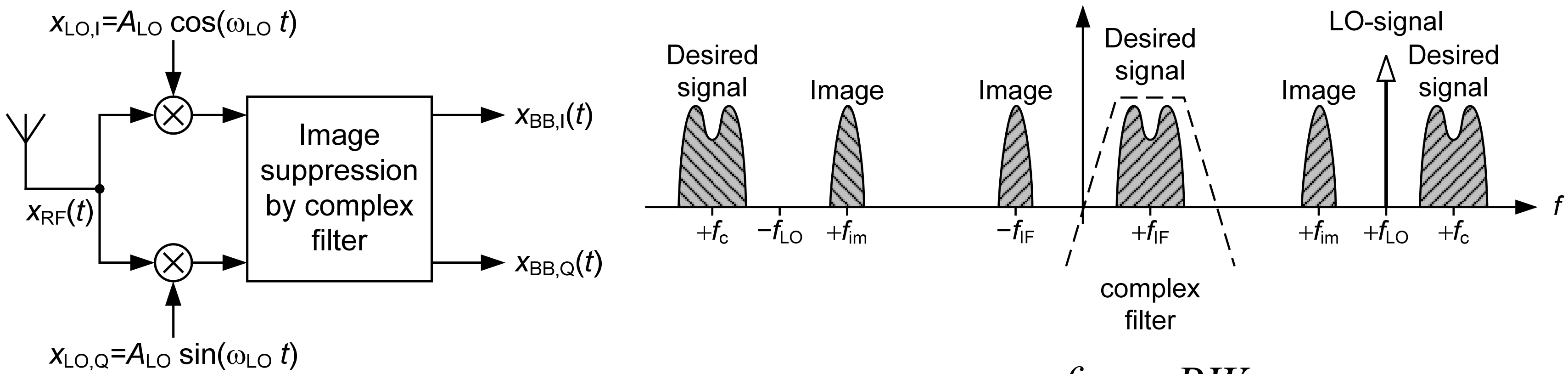
Low-IF Receiver – Principle





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

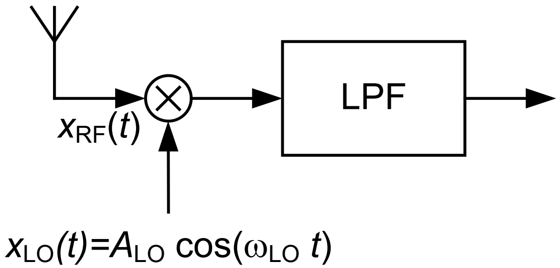
Low-IF Receiver – Polyphase Filters



$$f_{IF} > BW$$

- Image suppression by **bandpass polyphase filter**
- Allows high-pass filtering in baseband to remove DC offsets and part of flicker noise spectrum
- LO leakage, LO pulling by PA and even-order distortion are still problematic
- Requires **more stringent I/Q matching** than zero-IF Rx
- Baseband ADC must run twice as fast as in zero-IF Rx

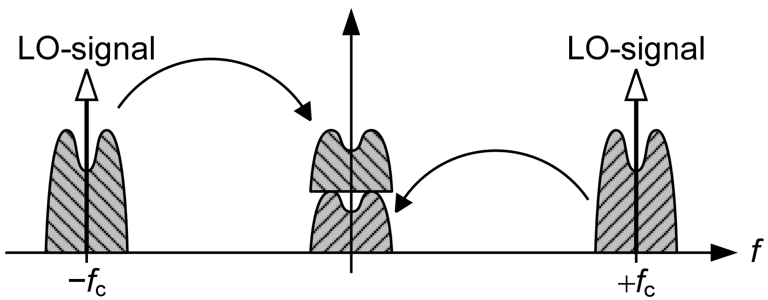
Why Quadrature Receivers? – The Homodyne Receiver



$$x_{BB}(t)=x_{RF}(t)\cdot x_{LO}(t)=x_{RF}(t)\cdot A_{LO}\cos(\omega_{LO}\,t)$$

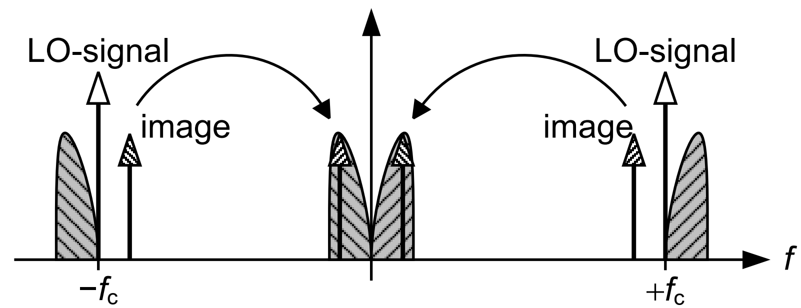
$$X_{BB}(f)=X_{RF}(f)*X_{LO}(f)=X_{RF}(f)*\frac{A_{LO}}{2}\left[\delta(f+f_{LO})+\delta(f-f_{LO})\right]=$$

$$=\frac{A_{LO}}{2}\left[X_{RF}(f+f_{LO})+X_{RF}(f-f_{LO})\right]$$



Double side band modulated signals

$$f_{LO}=f_c$$



Single side band modulated signals

- Homodyne receivers with simple real downconversion **only work with double-side band modulated signals** (upper and lower sidebands are symmetrical)
- Homodyne receivers with simple real downconversion do not work in the case where the upper and lower sidebands are asymmetrical
- For this reason the downconversion has to be made in **quadrature**

Transceiver Description Using Complex Signals

- Transceivers using quadrature signals can advantageously be described using **complex signals**
- The **in-phase** signal I constitutes the **real part**, whereas the **quadrature-phase** Q signal represents the **imaginary part**

$$x(t) = x_I(t) + jx_Q(t) \quad \Leftrightarrow \quad X(f) = X_I(f) + jX_Q(f)$$

- A multi-path (I and Q) quadrature transceiver can then be described with a single path using complex signals
- The quality of multi-path signal processing depends heavily on the matching between the I and Q path
- In digital systems this matching can be perfect, but in analog signal processing the I/Q path matching is fundamentally limited by the component matching

📖 K. Martin, "Complex Signal Processing is not – Complex," ESSCIRC 2003.
📖 K. Martin, "Complex signal processing is not complex," IEEE Trans. Circuits Syst. I, Reg. Papers, vol. 51, no. 9, pp. 1823–1826, Sep. 2004.

Complex Amplifier

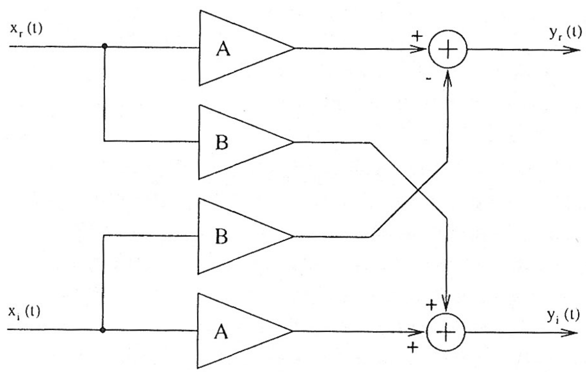


Figure 3.1. Amplification of a complex signal with a complex constant $A + jB$.

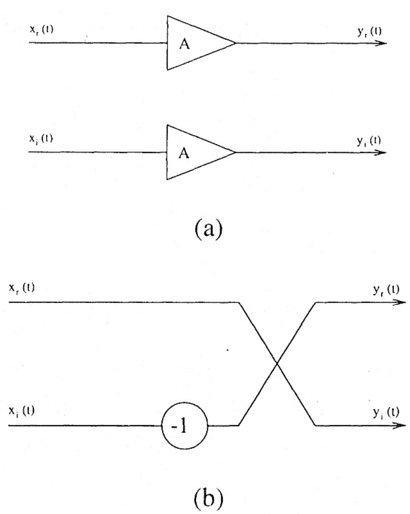


Figure 3.2. Special cases of complex amplification : a) a real amplifier and b) a 90 ^{o} phase shift.

$$y(t) = Z \cdot x(t) = (A + jB) \cdot [x_r(t) + j\,x_i(t)]$$

$$= \underbrace{A\,x_r(t) - B\,x_i(t)}_{=y_r(t)} + j\underbrace{[B\,x_r(t) + A\,x_i(t)]}_{=y_i(t)}$$

$$y_r(t) = A\,x_r(t) - B\,x_i(t)$$

$$y_i(t) = B\,x_r(t) + A\,x_i(t)$$

$$Z = A \quad (B = 0)$$

$$y_r(t) = A\,x_r(t)$$

$$y_i(t) = A\,x_i(t)$$

$$Z = +j \quad (A = 0, B = 1)$$

$$y_r(t) = -x_i(t)$$

$$y_i(t) = x_r(t)$$

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Complex Mixer – Complex Input and Output Signals

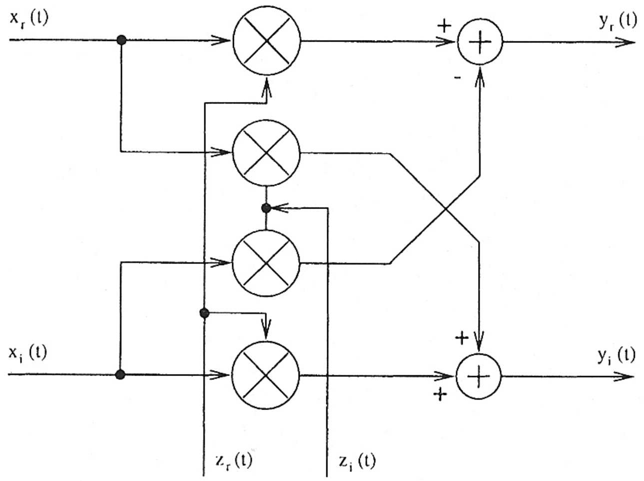
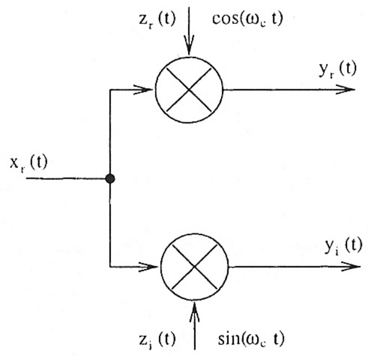


Figure 3.3. Block schematic of a complex mixer.

$$y(t) = z(t) \cdot x(t) = [z_r(t) + j\,z_i(t)] \cdot [x_r(t) + j\,x_i(t)]$$
$$= \underbrace{z_r(t)\,x_r(t) - z_i(t)\,x_i(t)}_{=y_r(t)} + j\underbrace{[z_i(t)\,x_r(t) + z_r(t)\,x_i(t)]}_{=y_i(t)}$$

$$y_r(t) = z_r(t)\,x_r(t) - z_i(t)\,x_i(t)$$
$$y_i(t) = z_i(t)\,x_r(t) + z_r(t)\,x_i(t)$$



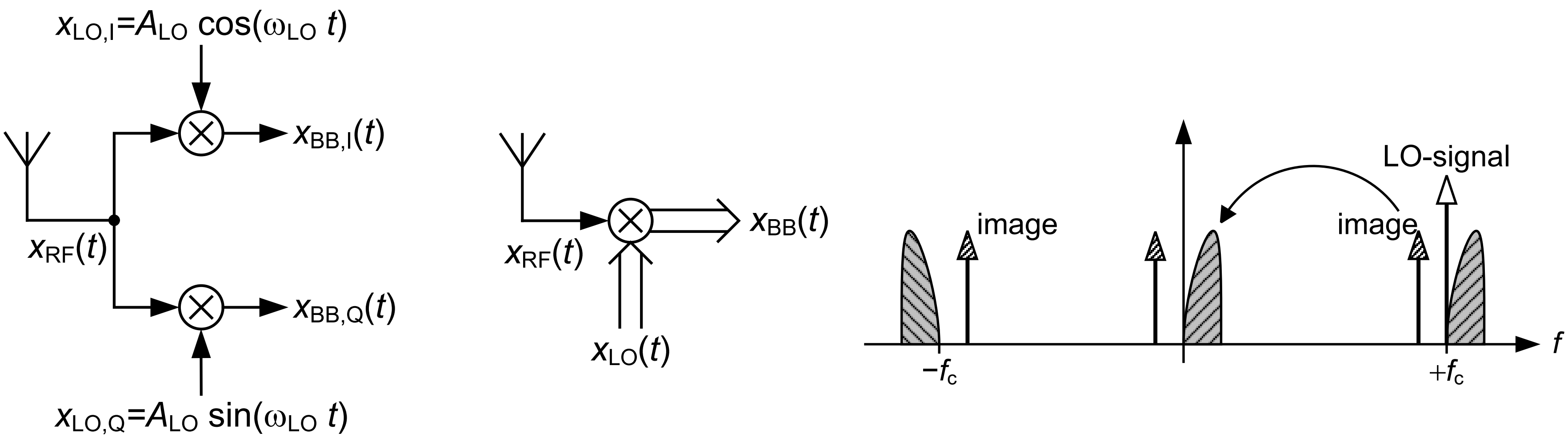
$$x_i(t) = 0$$

$$y_r(t) = z_r(t)\,x_r(t)$$
$$y_i(t) = z_i(t)\,x_r(t)$$

Figure 3.4. The multiplication of a real signal with a complex signal (a positive frequency, $f_c = \omega_c/2\pi$), as used in the zero-IF receiver.

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Complex Mixer – Real Input (RF) and Complex Output



$$x_{LO}(t) \triangleq x_{LO,I}(t) + jx_{LO,Q}(t) = A_{LO}\left[\cos(\omega_{LO} t) + j\sin(\omega_{LO} t)\right] = A_{LO}\, e^{j\omega_{LO} t}$$

$$X_{LO}(f) = X_{LO,I}(f) + jX_{LO,Q}(f) =$$

$$= \frac{A_{LO}}{2}\left[\delta(f + f_{LO}) + \delta(f - f_{LO})\right] - \frac{A_{LO}}{2}\left[\delta(f + f_{LO}) - \delta(f - f_{LO})\right] = A_{LO}\,\delta(f - f_{LO})$$

$$x_{BB}(t) = x_{RF}(t) \cdot x_{LO}(t)$$

$$X_{BB}(f) = X_{RF}(f) * X_{LO}(f) = X_{RF}(f) * A_{LO}\,\delta(f - f_{LO}) = A_{LO}\, X_{RF}(f - f_{LO})$$

- Complex mixer translates the entire spectrum of the real RF signal by f_{LO}
- If $f_{LO} = f_c$ the RF signal is translated directly to DC without mixing with the lower (upper) sideband

Complex Filters

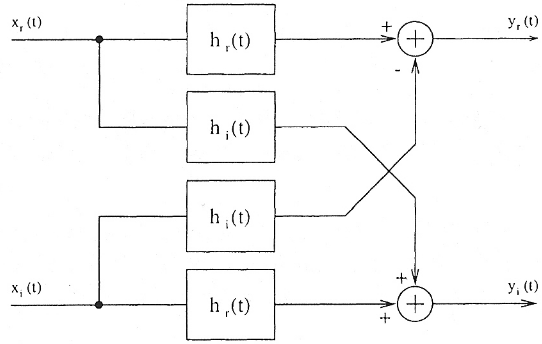


Figure 3.5. General block schematic of a complex filter.

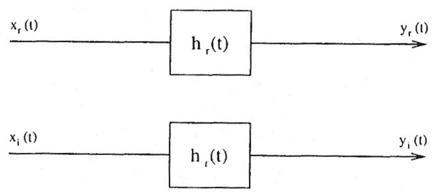


Figure 3.6. A special complex filter : the real filter for complex signals.

- Real and imaginary parts of a complex signal are filtered by a real filter having impulse response $h_r(t)$ and $h_i(t)$

$$h(t) = h_r(t) + jh_i(t) \quad \leftrightarrow \quad H(j\omega) = H_r(j\omega) + jH_i(j\omega)$$

- The output complex signal is then given by

$$Y(j\omega) = H(j\omega) \cdot X(j\omega)$$

$$Y_r(j\omega) = H_r(j\omega) \cdot X_r(j\omega) - H_i(j\omega) \cdot X_i(j\omega)$$
$$Y_i(j\omega) = H_i(j\omega) \cdot X_r(j\omega) + H_r(j\omega) \cdot X_i(j\omega)$$

$$y_r(t) = h_r(t) * x_r(t) - h_i(t) * x_i(t)$$
$$y_i(t) = h_i(t) * x_r(t) + h_r(t) * x_i(t)$$

- Special cases are:
  - Real filter filtering a complex signal
  - Complex filtering of a real input signal giving rise to a complex output signal

📖 J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Complex Filters

- The main application of complex filters is the suppression of positive or negative frequency components (such as image) by a bandpass filter
- The latter can be obtained by simple frequency translation of a low-pass filter

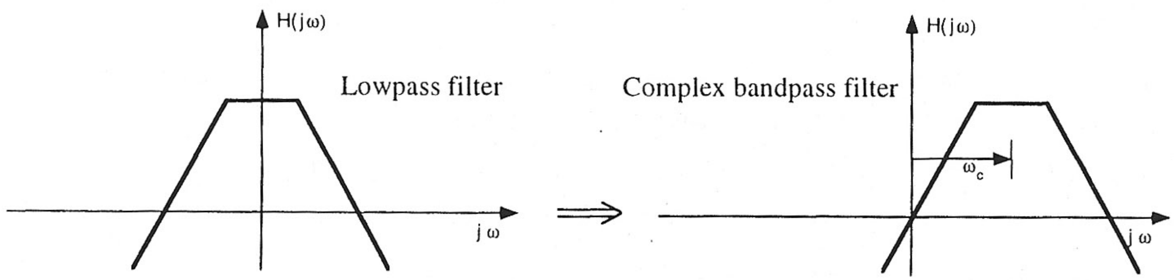


Figure 3.7. A linear lowpass to bandpass transformation, resulting in a complex band-pass filter with a transfer function which passes only positive frequencies.

$$j\omega \;\rightarrow\; j\omega - j\omega_c$$

which introduces complex coefficients of $j\omega$

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Direct Synthesis of Complex Filters

- The most efficient way to realize a 2nd-order complex bandpass filter is by direct synthesis

$$\frac{Y(j\omega)}{X(j\omega)} = H_{BP}(j\omega) = \frac{1}{1 - j2Q + j\omega/\omega_0} \quad \text{with} \quad \omega_0 = \frac{1}{RC} \quad \text{and} \quad Q = \frac{\omega_c}{2\omega_0}$$

- ω_0 is the bandwidth of the original lowpass filter and $2\omega_0$ is the bandwidth of the transformed bandpass filter centered around ω_c
- Can be rewritten as

$$j\omega/\omega_0 \cdot Y(j\omega) = X(j\omega) + \left(j2Q - 1\right) \cdot Y(j\omega)$$

- Which can be synthesized by an integrator and a complex amplifier

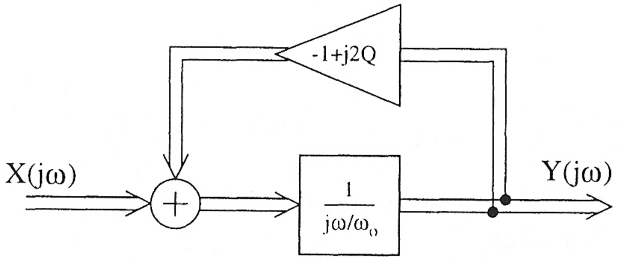


Figure 3.8. Direct synthesis of a second order complex bandpass filter for positive frequencies.

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Complex Filters

$$\frac{Y(j\omega)}{X(j\omega)} = H_{BP}(j\omega) = \frac{1}{1 - j2Q + j\,\omega/\omega_0} \quad \text{with} \quad \omega_0 = \frac{1}{RC} \quad \text{and} \quad Q = \frac{\omega_c}{2\omega_0}$$

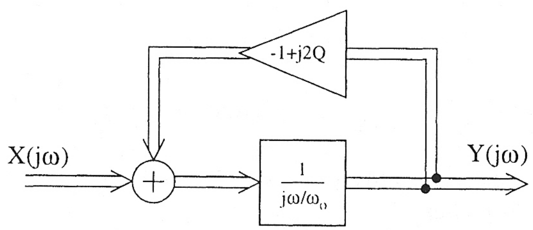


Figure 3.8. Direct synthesis of a second order complex bandpass filter for positive frequencies.

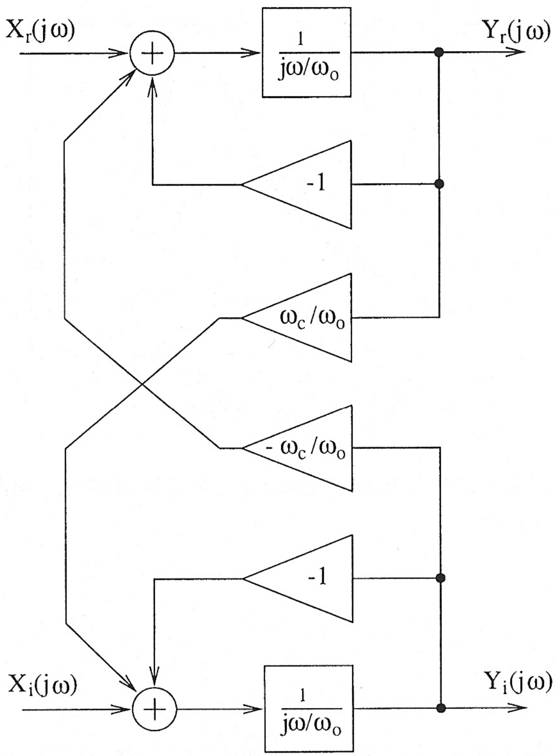


Figure 3.9. Expanded block schematic of a second-order complex bandpass filter.

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Example of a 5th-order Complex Filters

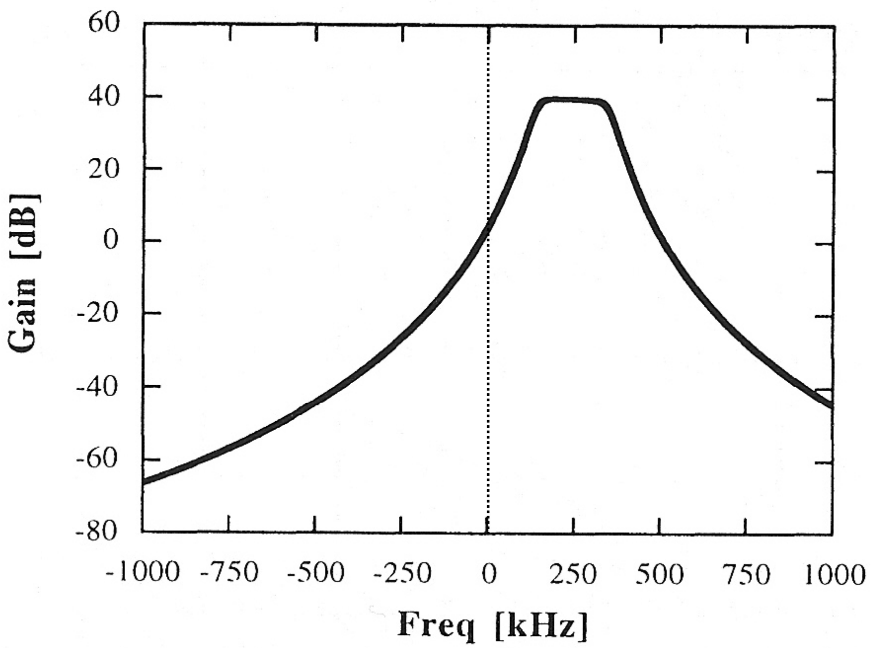


Figure 3.11. The transfer function for positive and negative frequencies of a 5^{th}-order complex bandpass filter.

J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Polyphase Filters

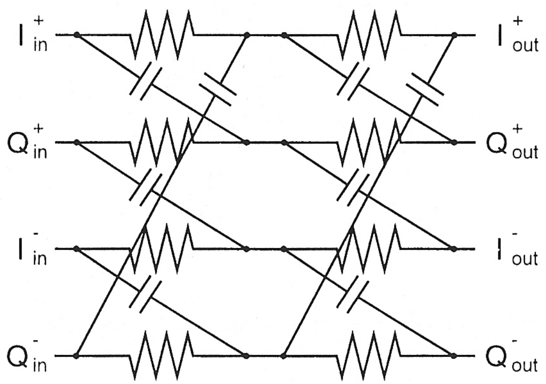


Figure 3.12. A two-stage passive sequence asymmetric polyphase filter.

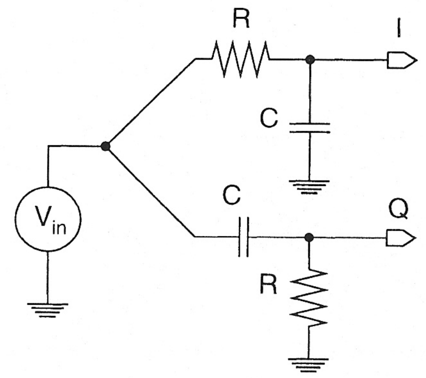


Figure 3.14. An RC-CR filter, classically used for the generation of quadrature oscillator signals.

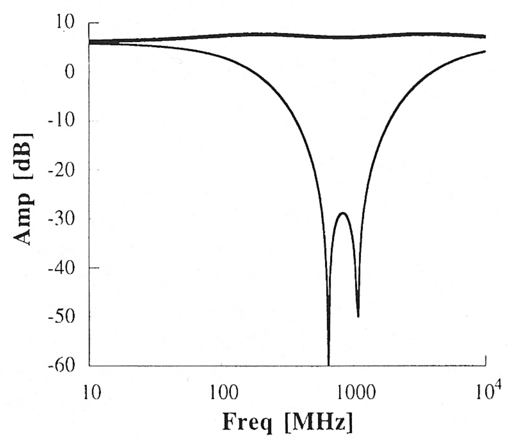


Figure 3.13. The transfer function of a 2-stage polyphase filter for positive frequencies (thick line) and negative frequencies (thin line).

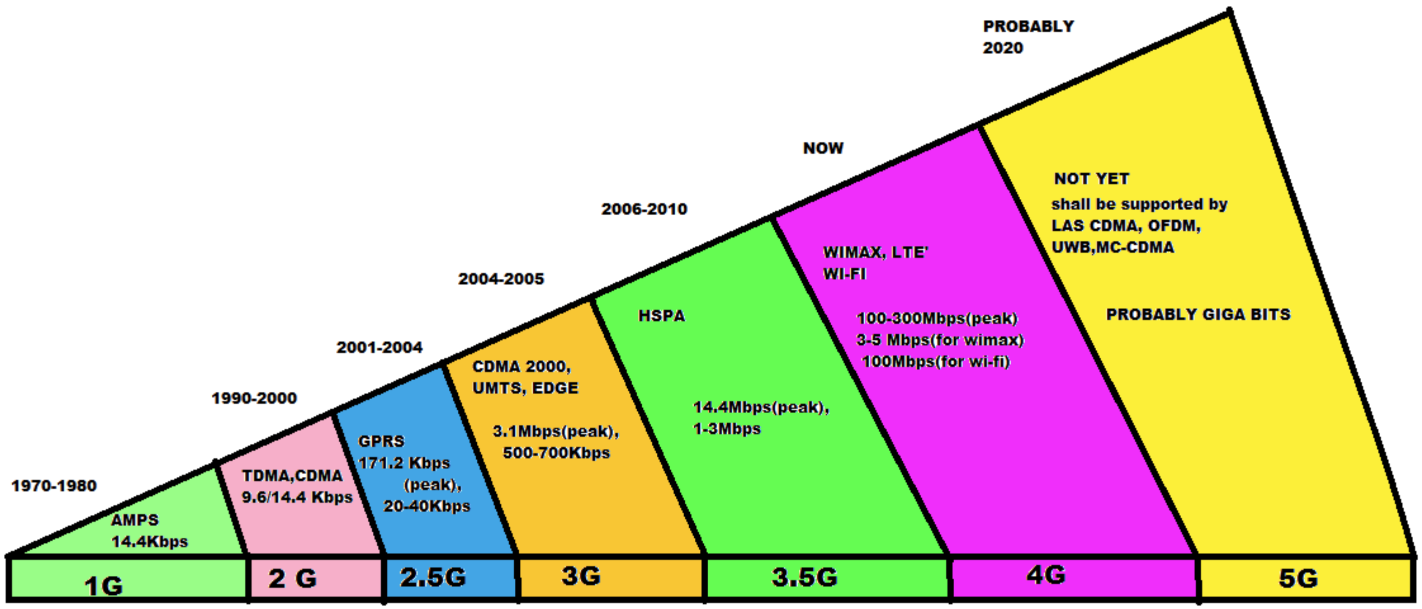
J. Crols and M. Steyaert, CMOS Wireless Transceiver Design, Kluwer Academic Press, 1997.

Outline

- General Considerations
- Heterodyne Receivers
- Homodyne Receivers
- Image-Reject Receivers
- Low-IF Receivers
- **Discrete-time Receivers**

Introduction – Need for Software Defined Radio (SDR)

- Driving force behind the rapid expansion of the Wireless Communication field is the huge demand to consume wide variety of data (video, gaming) on the move
- Emergence of various new standards and protocols owing to this demand
- Need for a receiver able to receive any channel in any band of any modulation
- The hardware of the receiver should be programmable (Software) to receive the wanted signal across the whole spectrum irrespective of the modulation used



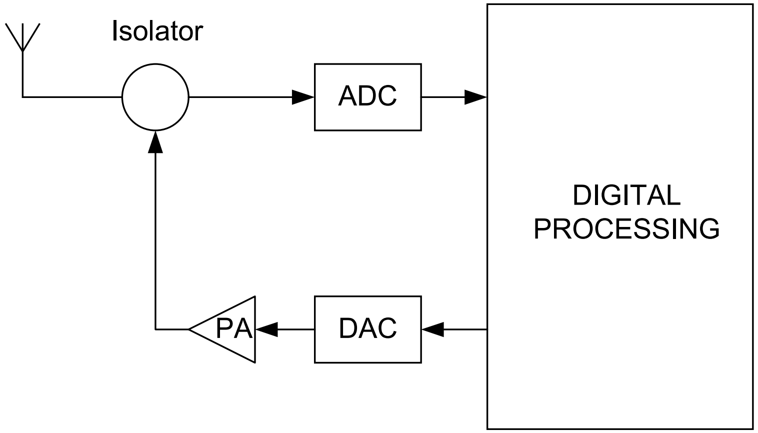
Multi-mode

Multi-band

Multi-network

Introduction – What is SDR?

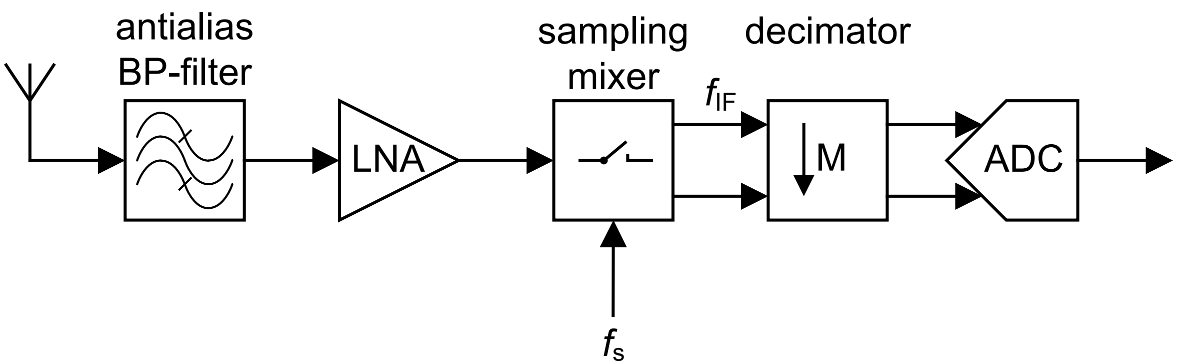
Mitola's receiver front end



- SDR or a Digital Radio performs all the signal processing in the digital domain
- This idea of digitizing the entire spectrum received by the antenna and then demodulating in the digital domain was first proposed by J. Mitola
- This is not a practical solution as the power requirements on the ADC are totally non-realistic (due to the very high sampling rate)
- So, many receivers based on this idea, but with some signal processing done in the analog domain (discrete time) and feeding the processed samples to the ADC have been proposed and designed

J. Mitola, IEEE Communication Mag, vol. 33, no. 5, pp. 26-38, May 1995

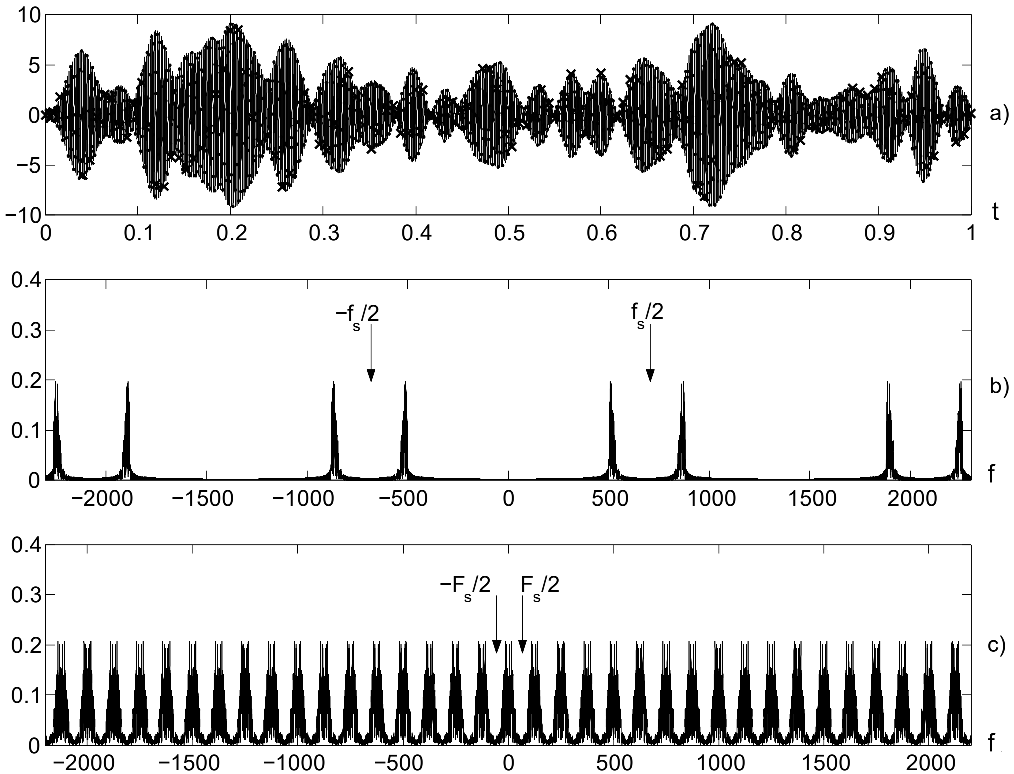
Sub-sampling Receiver Architecture



- The mixing is achieved by sub-sampling directly the RF input signal
- To avoid aliasing of unwanted bands into the wanted region of the spectrum (as a result of sampling), a band-pass pre-filter used as an anti-aliasing filter is a MUST
- The samples at the output of the mixer cannot be directly fed to the ADC (because of the high sampling rate) $\Rightarrow$ decimation and/or down-sampling is required
- The band at the output of the mixer also needs to be brought down to baseband $\Rightarrow$ downconversion
- Both decimation and downconversion can normally be achieved in a single circuitry (for example switched capacitors circuits)

Bandpass Sampling

- **Band-pass sampling** (BPS) is a technique for undersampling a band-pass signal to realize **frequency down-conversion** through intentional aliasing with the sampling rate being slightly larger than twice the signal bandwidth $f_s \geq 2B$ and much smaller than the classical low-pass sampling (LPS) frequency f_s' (i.e. $f_s \ll f_s'$)



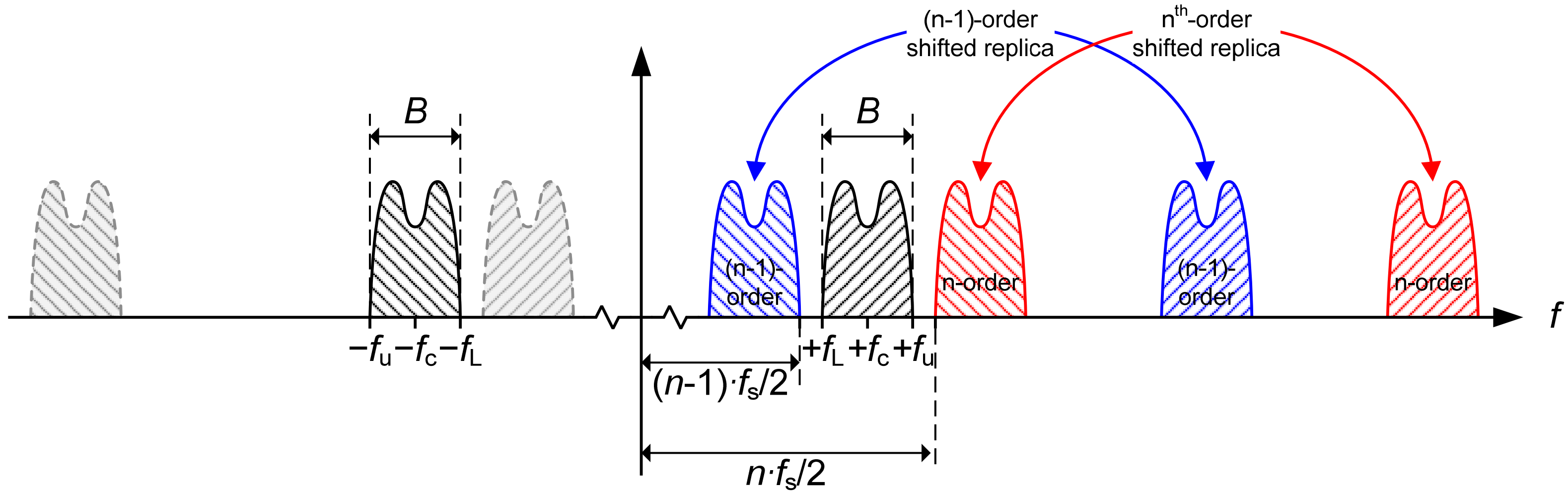
Random generated single-side band suppressed-carrier (SSB-SC) AM band-pass signal with bandwidth $B = 50$ and $f_c = 500$

Low-pass sampling (LPS) of the above signal with $f_s' = 2.5(f_c + B) = 1375$ (slightly larger than the minimum $f_{smin}' = 2(f_c + B) = 1100$

Band-pass sampling (BPS) of the above signal with $f_s = 2.5B = 125$, (slightly larger than the minimum $f_{smin} = 2B = 100$)

Yi-Ran Sun, PhD Thesis, KTH, 2006.

Bandpass Sampling – Avoid Spectra Overlapping



- The condition $f_s \geq 2B$ is not sufficient because of **spectra folding**
- To avoid harmful signal spectral overlapping and have proper interlacing of the shifted spectra resulting from the sampling process, the following constraints should apply to the nth and (n-1) order shifted spectra

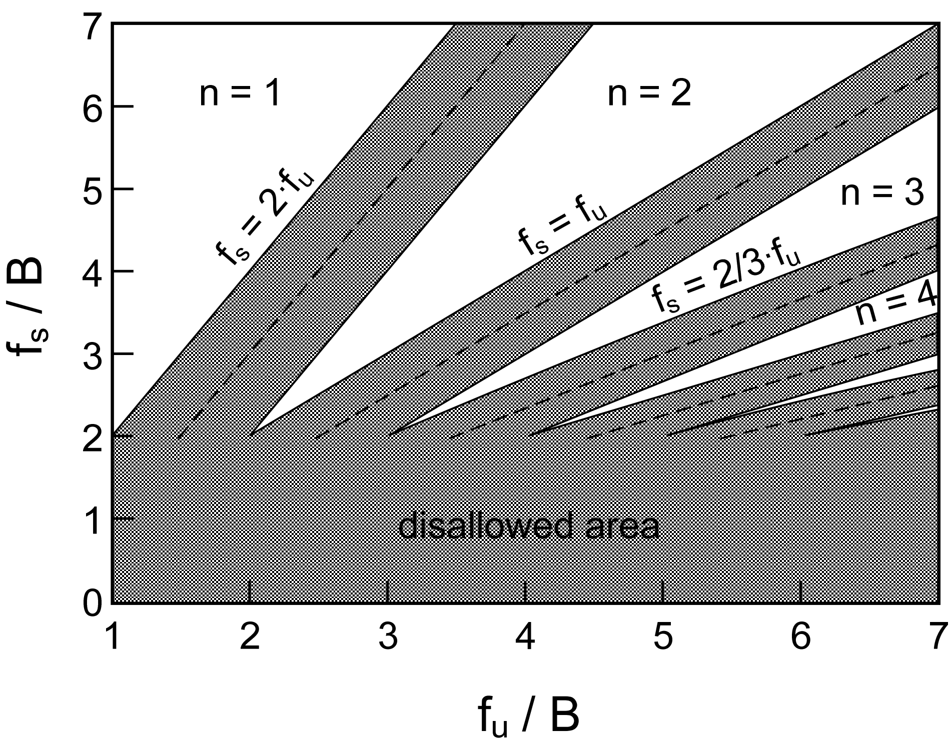
$$\begin{cases} n \cdot \frac{f_s}{2} \geq f_u \\ (n-1) \cdot \frac{f_s}{2} \leq f_L \end{cases} \quad \text{or} \quad \frac{2f_u}{n} \leq f_s \leq \frac{2f_L}{n-1} \quad \text{with} \quad 1 \leq n \leq \left\lfloor \frac{f_u}{B} \right\rfloor$$

where $\lfloor f_u / B \rfloor$ denotes the largest integer less or equal than f_u / B

R. G. Vaughan, N. L. Scott and D. R. White, "The Theory of Bandpass Sampling," Trans. on Signal Proc., Sept. 1991.

Bandpass Sampling – Acceptable Sampling Rates

- The condition for acceptable uniform sampling rate are represented below, where the gray area represent the disallowed area



$$\frac{f_{s\min}}{B} = \frac{2}{n}\cdot\frac{f_u}{B} \qquad \frac{f_{s\max}}{B} = \frac{2}{n-1}\cdot\left(\frac{f_u}{B}-1\right) \qquad \frac{f_{smid}}{B} = \frac{1}{n}\cdot\left(\frac{2 f_u}{B}-1\right)$$

- The dashed lines correspond to a particular acceptable sampling rate for symmetrical double side-band (DSB) and corresponding to direct down-conversion (homodyne)
- The acceptable sampling rate in this case is given by

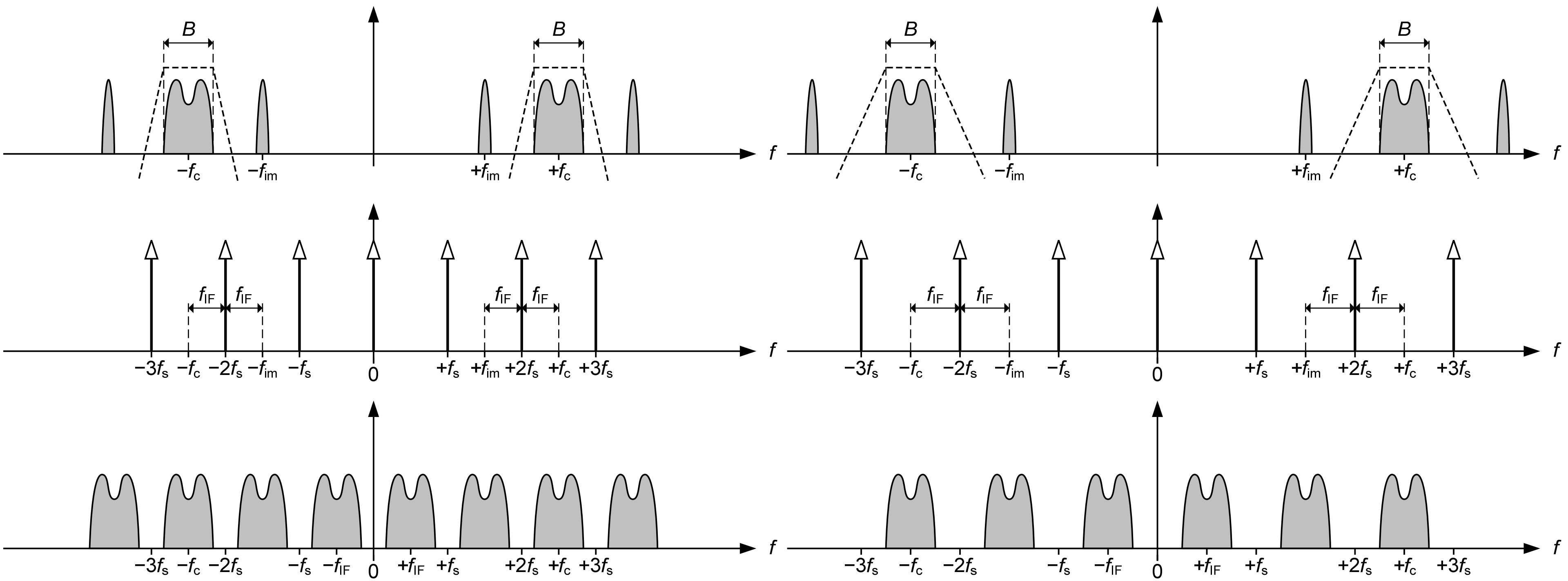
$$f_s = \frac{2\cdot f_c}{n}$$

where f_c is the center frequency defined as

$$f_c = \frac{f_L + f_u}{2} = f_L + \frac{B}{2} = f_u - \frac{B}{2}$$

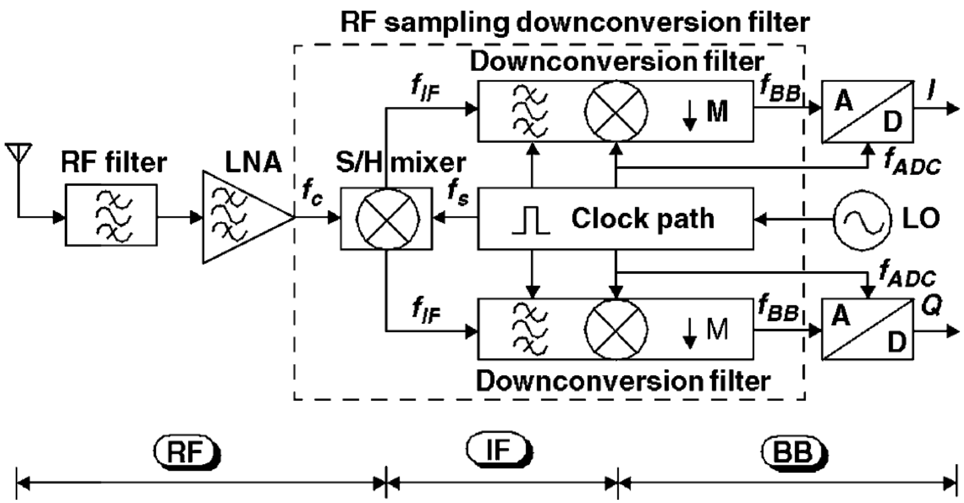
R. G. Vaughan, N. L. Scott and D. R. White, "The Theory of Bandpass Sampling," Trans. on Signal Proc., Sept. 1991.

Sampling Frequency and Anti-aliasing Filter



- Increasing the sampling frequency f_s moves the unwanted alias frequency bands (image bands) farther from the signal resulting in a relaxed anti aliasing bandpass filter and reduces the effect of clock jitter
- Lowering f_s results in lower power consumption and easier circuit design (ADC and frequency synthesizers' specifications are relaxed)

Implementation of the Receiver Front End



- Receiver front end designed for 802.11 b/g
- Channel selection achieved by changing the sampling frequency

Radio frequency bandwidth	2.4-2.484 GHz
Center frequency	2.412 GHz
Signal bandwidth	22 MHz
Sampling rate	1072 MS/s
Closest RF image spacing	±562 MHz
Intermediate frequency	268 MHz
Anti-alias bandwidth	157 MHz max
Decimated sample rate	89.3MS/s

D. Jakonis, JSSC, vol 40, no. 6 June 2005

References

[1] B. Razavi, *RF Microelectronics*, 1st ed. Prentice Hall, 1998.

[2] B. Razavi, *RF Microelectronics*, 2nd ed. Pearson, 2012.

[3] J. Crols and M. Steyaert, *CMOS Wireless Transceiver Design,* Kluwer, 1997.