Homodyne or Zero IF Transceivers

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Homodyne Receivers

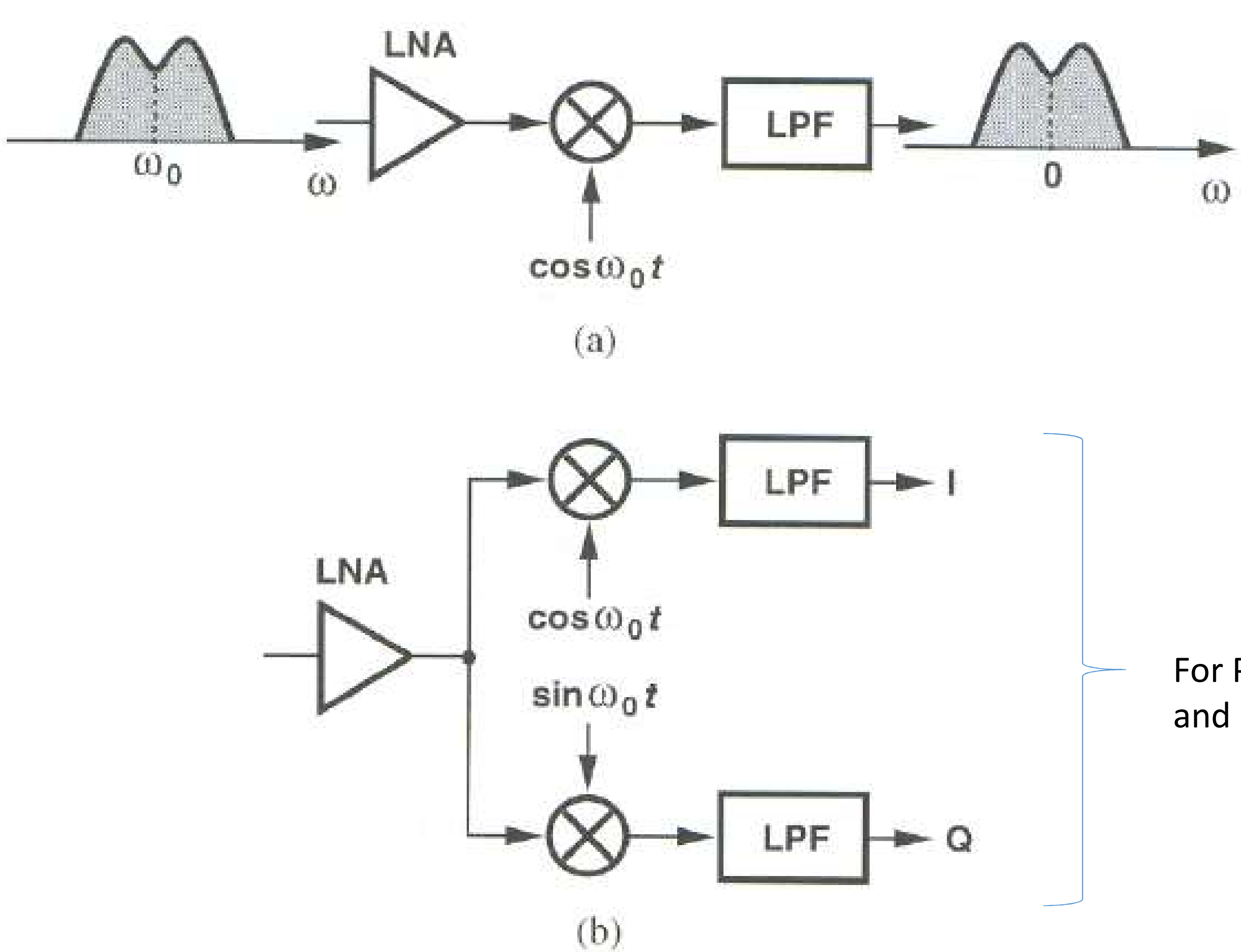
- Homodyne <--> direct-conversion <--> zero-IF architecture.
- For **double-sideband AM signals** the topology on Figure 5.14a can be used.
- For **frequency and phase modulated signals**, the down-conversion must provide quadrature output signal because the 2 sides of FM or QPSK spectra carry different information: see Figure 5.14b.

Advantages:

- No image problem (since $\omega_{IF} = 0$) so no image filter needed.
- IF SAW filters and down-conversion stages replaced by baseband lowpass filter and amplifier: amenable to monolithic integration (i.e on chip integration).

Drawbacks:

- DC offsets
- I/Q mismatch
- 1/f noise
- Even order distortion



For AM

For PM
and FM

Figure 5.14 (a) Simple homodyne receiver, (b) homodyne receiver with quadrature downconversion.

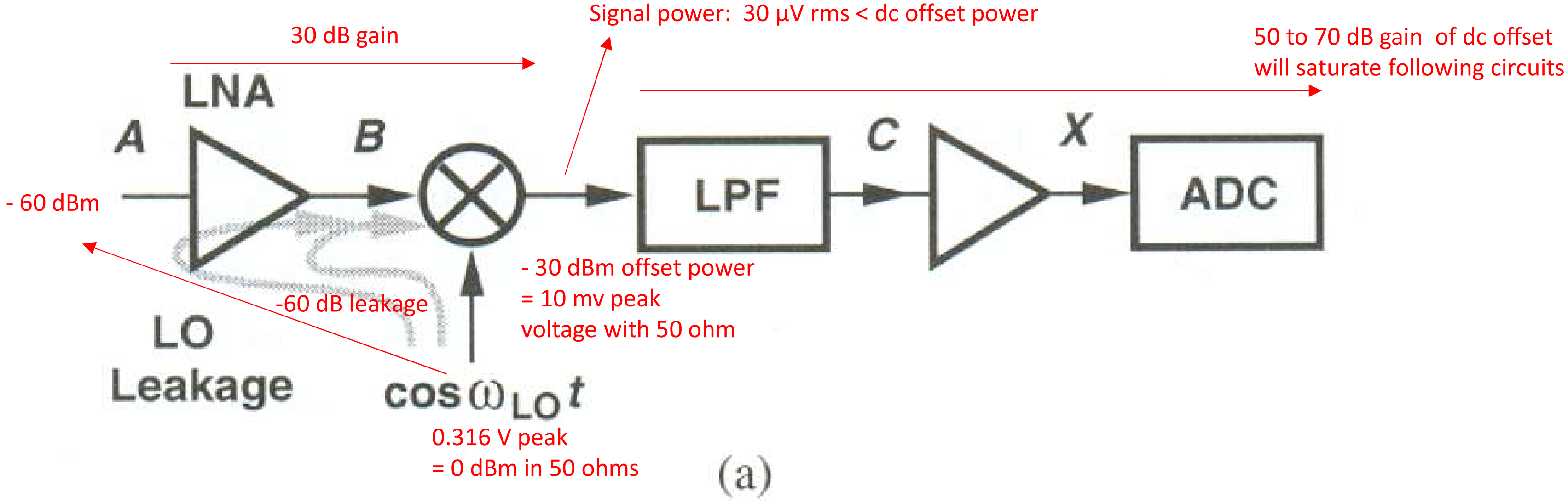
DC Offsets

DC offsets are due to **self-mixing** (signal multiplied by itself) originating from (see Figure 5.16):

- LO leakage (capacitive/substrate/bond-wire coupling) to i/p of lna or mixer.
- Large interferer coupling in the LO port

Example:
LO o/p of 0 dBm with a 60 dB leakage giving -60 dm at the LNA input. Then amplified by 30 dB by LNA/mixer giving -30 dBm. The remaining amplification of 50-70 dB will saturate the rest of the baseband circuits.

LO leakage can also be a problem since leakage of the signal to the antenna and radiation therefrom creates interference in the band of other receivers using the same wireless standard.



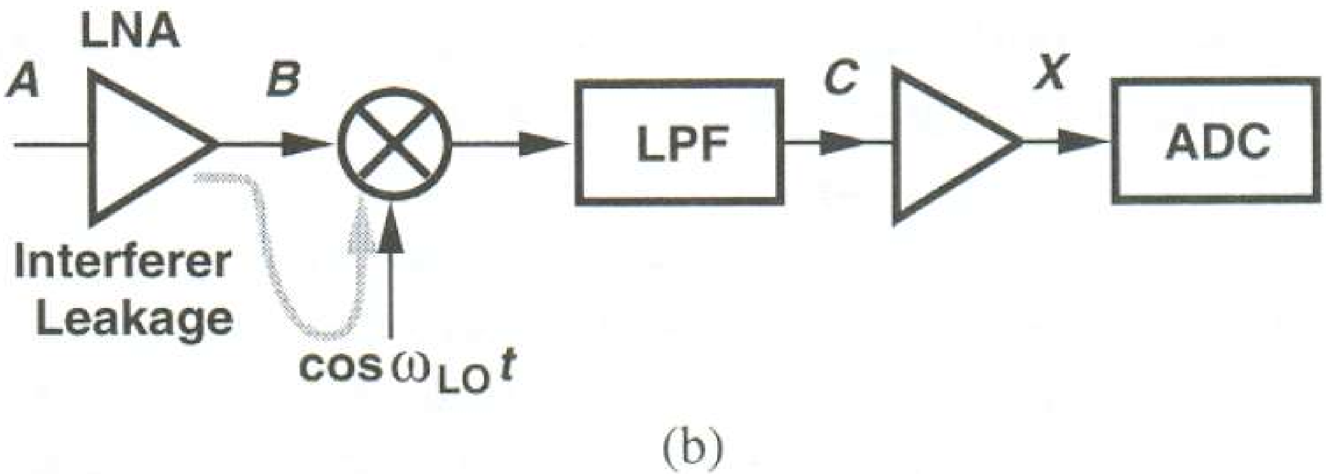


Figure 5.16 Self-mixing of (a) LO signal, (b) a strong interferer [5].

DC Offset (..Contd)

- The problem is exacerbated when the leakage signal varies rapidly -> difficult to distinguish original signal from leakage
  e.g, car moving at high speed -> leakage signal can be reflected multiple times.

Heterodyne receivers face very less this problem because

- LO frequency is different from RF frequency.
- Self mixing only at interferer frequency – offsets can be removed since IF is far from dc.

Offset Reduction and Cancellation:

Leakage control.

· Design LNA with high isolation (S_{12}) (cascode configuration) so that LO leakage is less.

· For external LO: use $2 \times \omega_{LO}$ followed by divide by 2 (inside chip) to mitigate bondwire coupling (with chip metal layers) so that bondwire coupling is with $2\omega_{LO}$ instead of ω_{LO}.

· Highpass filtering:

- Removing the band 0-20 Hz increases the BER to 10^{-3} so corner frequency has to be very low
- Base band needs to be encoded such that it contains less power near DC.
- Calls for prohibitively large capacitors.
- coarse cancellation i.e with high C values time constant is very high $\rightarrow$ fast changing dc offsets cannot be filtered.

· Store offset in idle time interval (e.g., in between TDMA bursts): see Figure 5.17

- Need large capacitor to reduce thermal noise kT/C.
- Say P_{in} (before LNA) = -120 dBm and LNA gain = 30dB $\Rightarrow$ P_{out} of LNA = -90 dBm.

  Now say we want to keep capacitor noise power at least 16 dB below P_{out} $\Rightarrow$ C = 200 pF or higher since kT/C for C = 200 pF = -106.87 dBm.
- Problem: time varying interferers are stored as well.

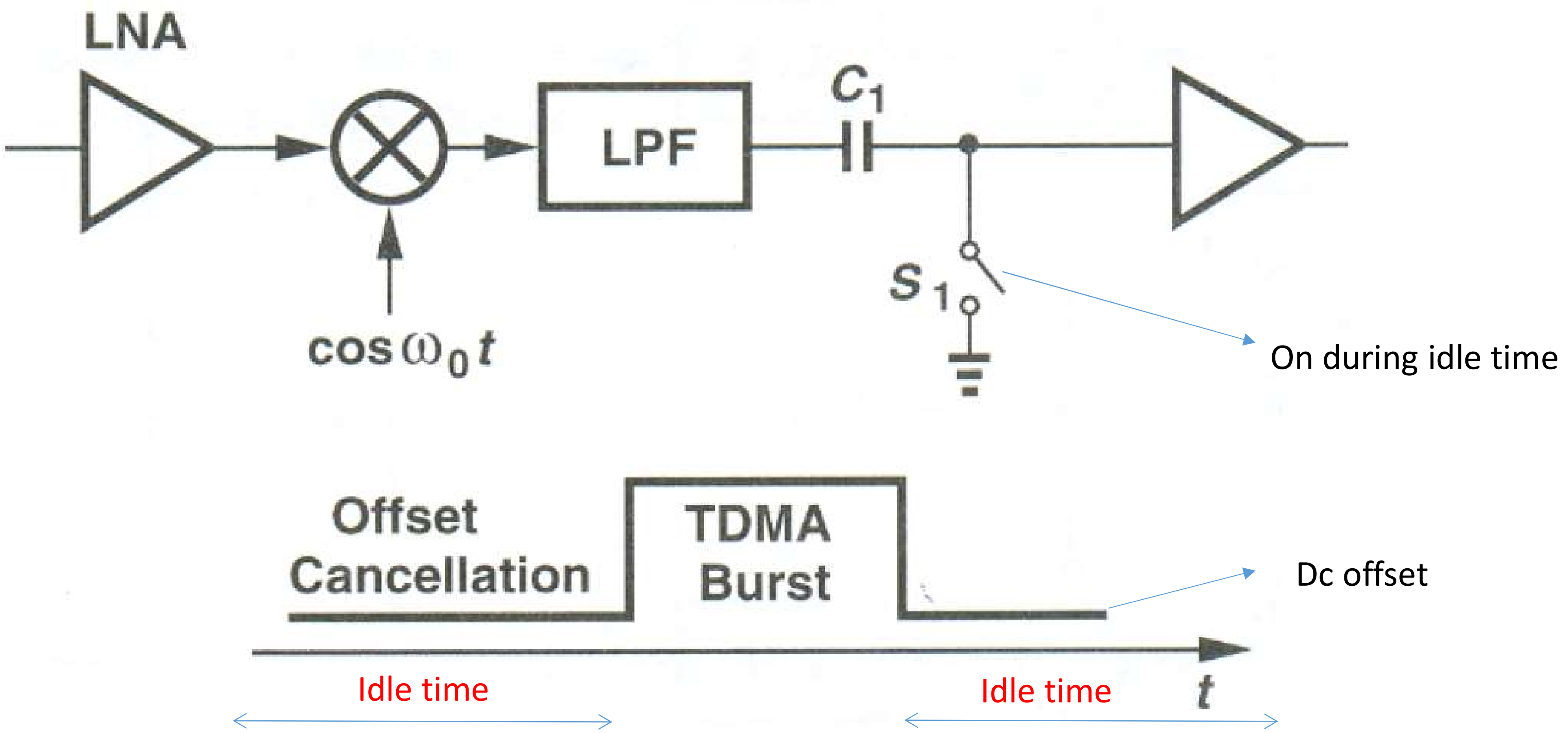


Figure 5.17 Simple offset cancellation in a TDMA system.

- Using feed back

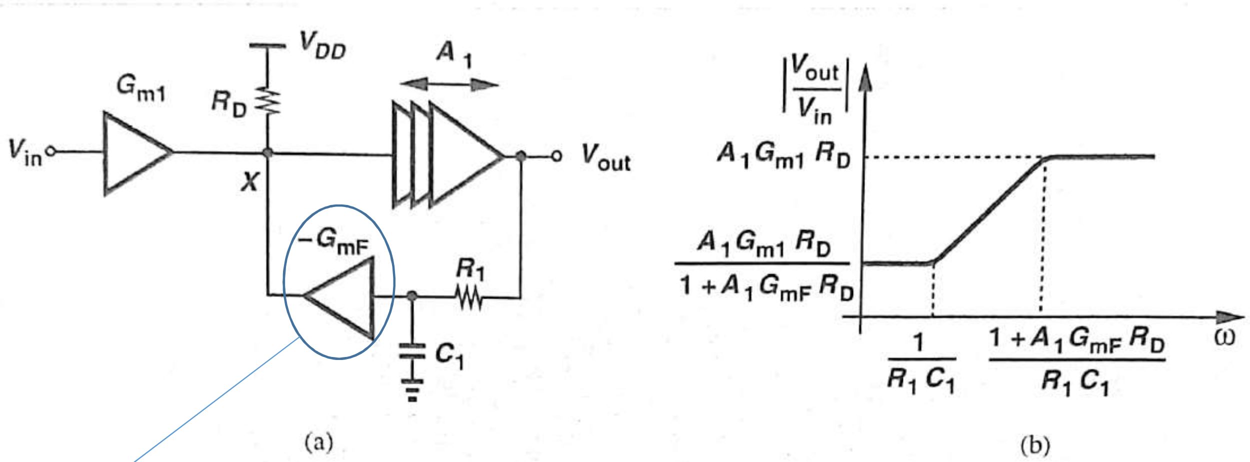


Figure 4.39 *(a) Offset cancellation by feedback, (b) resulting frequency response.*

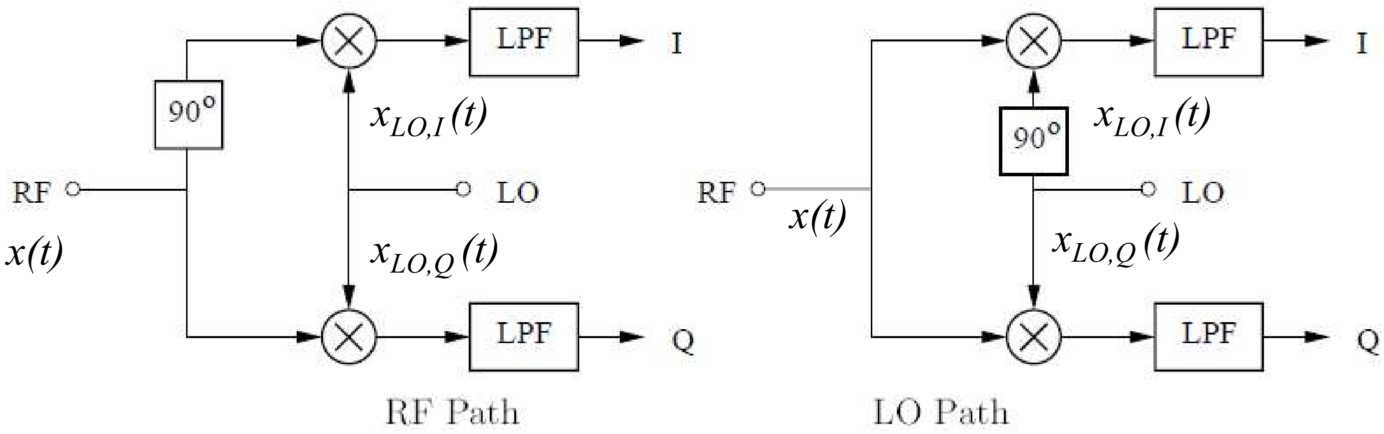
- Adjust $-G_{mF}$ to control HPF corner frequency
- Acts like a dc current sink to clear dc offset voltage

$$\frac{V_{out}}{V_{in}} = \frac{G_{m1} R_D A_1 (R_1 C_1 s + 1)}{R_1 C_1 s + G_{mF} R_D A_1 + 1}$$

- Can be implemented digitally (see Fig 4.40 in new edition of book by Razavi).

Quadrature Mixing

For phase and frequency modulation, quadrature mixing is required. Two options are possible.



The LO path is used to avoid adding loss to the RF path.

I/Q Mismatch

- The quadrature mixer can contribute a gain and phase error.

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

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

- The resulting I and Q baseband components after quadrature mixing are then:

$$x(t) = b_{2n}\cos(\omega_c t) + b_{2n+1}\sin(\omega_c t) \text{ Ideally } b_{2n} \text{ and } b_{2n+1} \text{ can take values of } \pm 1$$

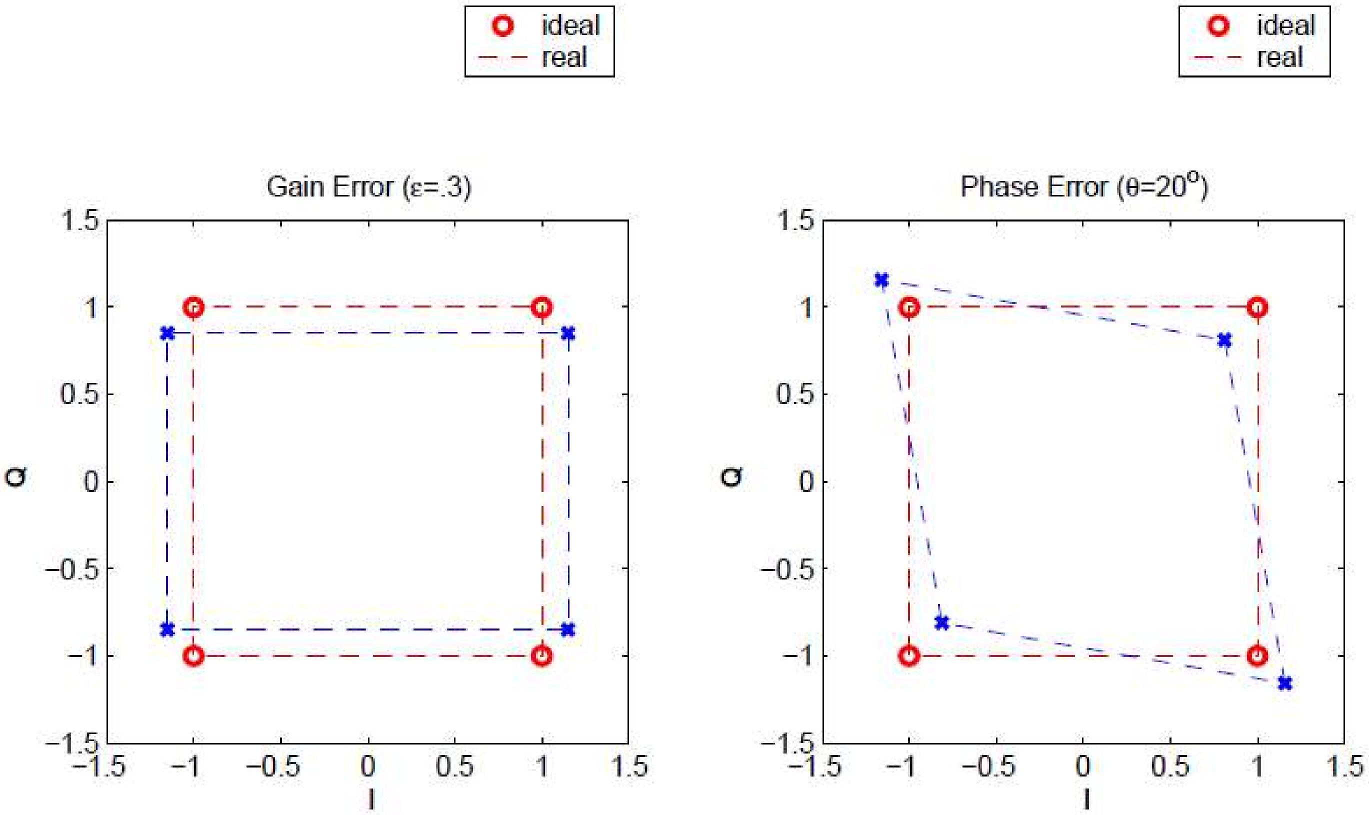
After mixing and LP filtering

$$x(t)\times x_{LO,I} + \text{LPF} \rightarrow x_{BB,I} = b_{2n}\left(1+\frac{\varepsilon}{2}\right)\cos\left(\frac{\theta}{2}\right) - b_{2n+1}\left(1+\frac{\varepsilon}{2}\right)\sin\left(\frac{\theta}{2}\right) \textit{ instead of } b_{2n}$$

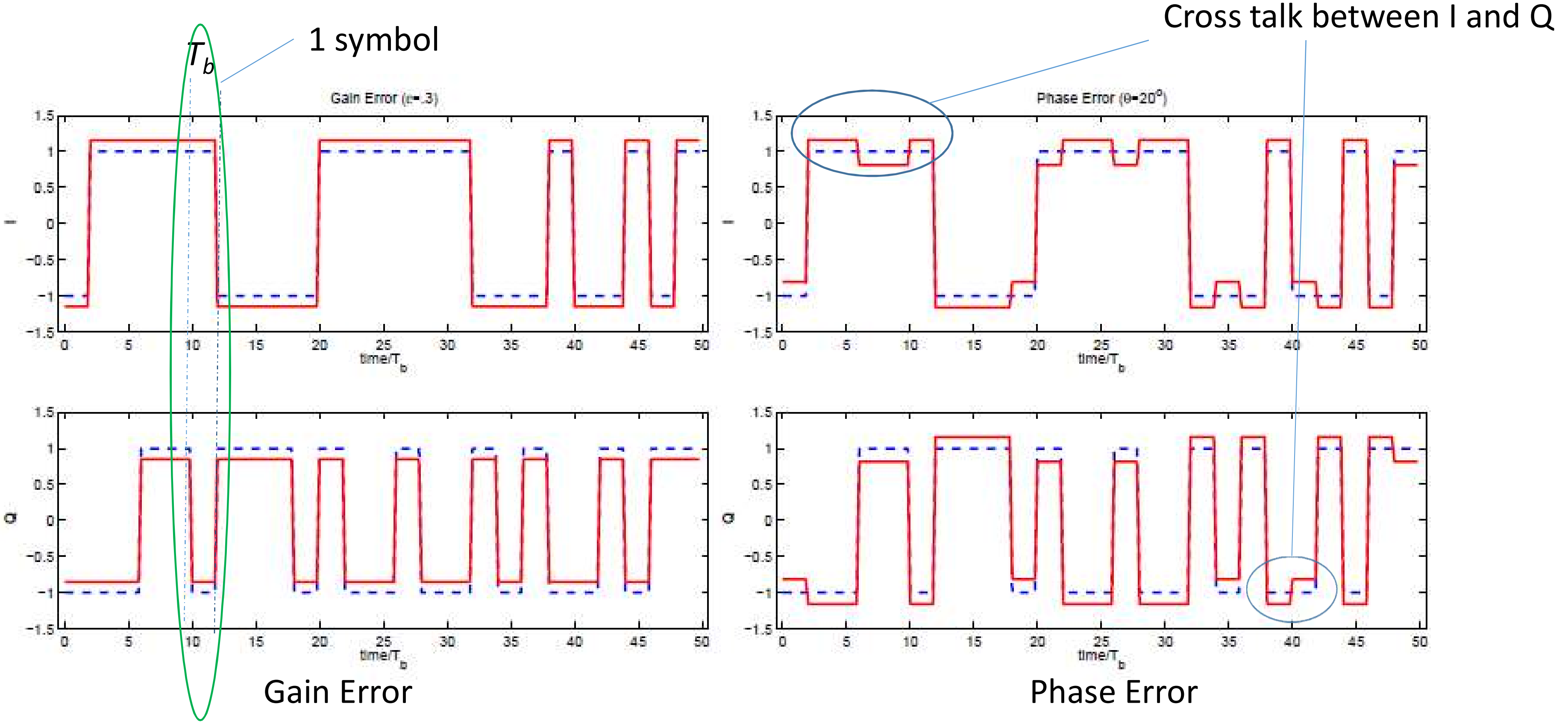
$$x(t)\times x_{LO,Q} + \text{LPF} \rightarrow x_{BB,Q} = -b_{2n}\left(1-\frac{\varepsilon}{2}\right)\sin\left(\frac{\theta}{2}\right) + b_{2n+1}\left(1-\frac{\varepsilon}{2}\right)\cos\left(\frac{\theta}{2}\right) \textit{ instead of } b_{2n+1}$$

Channels are corrupted by cross talk.

Effect of I/Q Mismatch on QPSK Signal Constellation



Effect of I/Q Mismatch on Time Domain Output of QPSK



Typical maximum target is 1 dB gain error and 5° phase error.

I/Q mismatch in heterodyne receivers (..Contd)

- Problem is less severe in heterodyne as the I/Q separation is done at IF for which LO frequency is much lesser than for homodyne.
- I/Q separation is done down the receiver after undergoing approx 50 to 60 dB of gain. In homodyne the gain is provided after I/Q separation.
- In heterodyne signal can be digitized at IF and I/Q separation can be done in digital domain.

Even Order Distortion in LNA or Mixer

Consider 2 strong interferers close to the channel :

$$x(t) = A_1 \cos \omega_1 t + A_2 \ \cos \omega_2 t$$

The second order non - linearity in the LNA : $y(t) = \alpha_1 x(t) + \alpha_2 x^2(t)$

generates a component :

$$\alpha_2 A_1 A_2 \cos(\underbrace{\omega_1 - \omega_2}_{\text{near dc}})t$$

which can then directly feedthrough from the RF input to the IF output through the mixer. see Figure 5.22

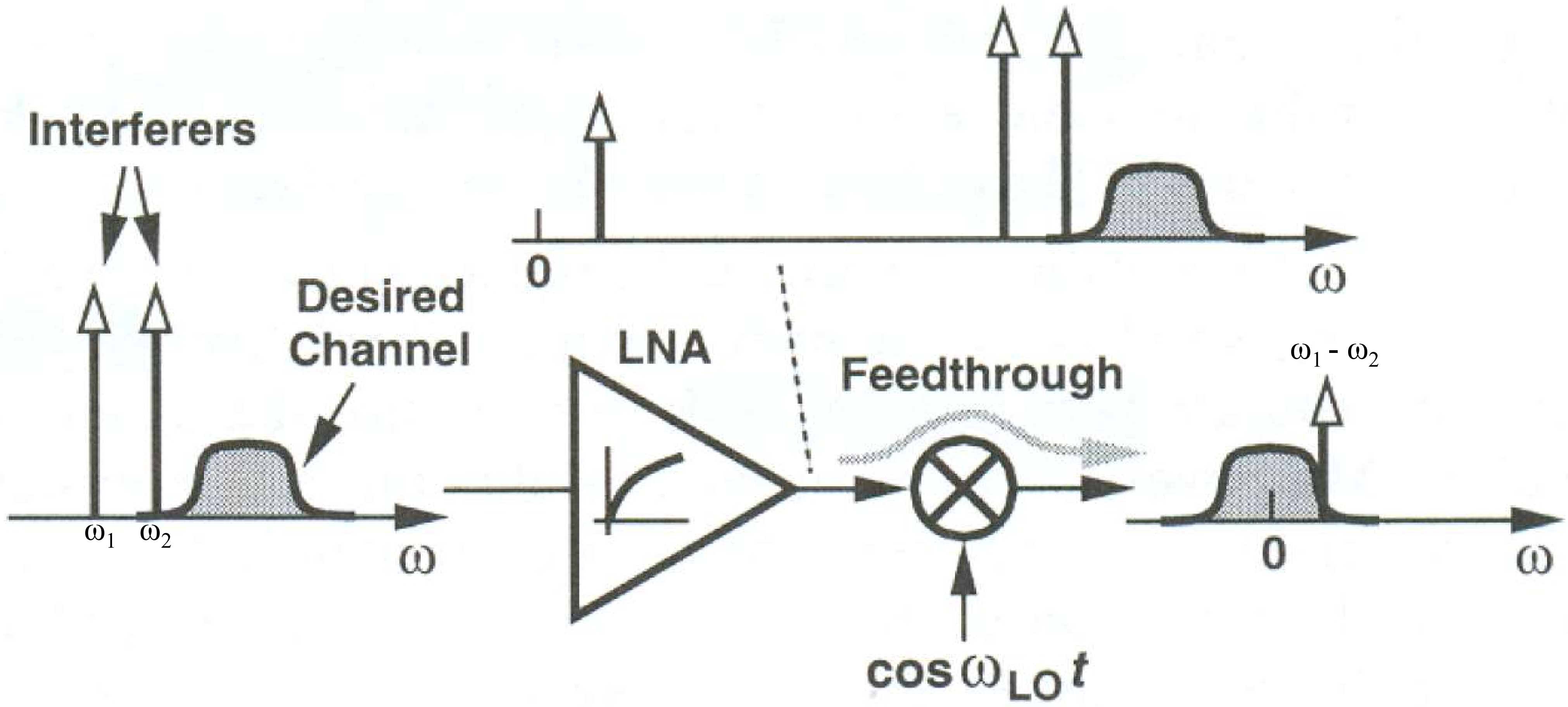


Figure 5.22 Effect of even-order distortion on interferers.

Even Order Distortion (.. Contd)

Say if x(t) is:

$$x(t) = \left(A + \varepsilon \cos \omega_m t\right)\left(a \cos \omega_{RF} t + b \sin \omega_{RF} t\right)$$

$$= \left(A + \varepsilon \cos \omega_m t\right)\sqrt{a^2 + b^2} \sin\left(\omega_{RF} t + \varphi\right)$$

The second order linearity in the receiver: $y(t) = \alpha_1 x(t) + \alpha_2 x^2(t)$

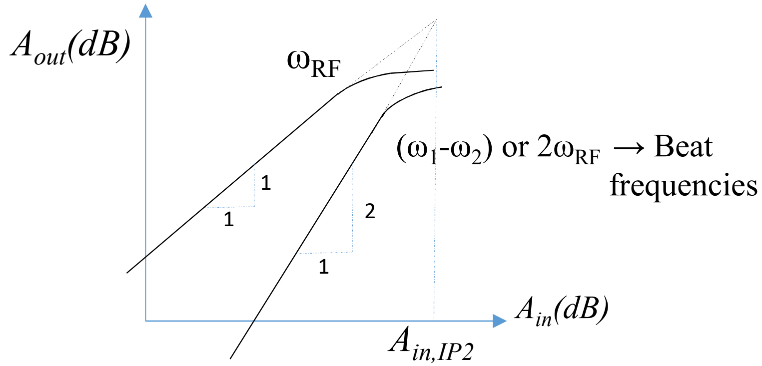
generates a component:

$$\alpha_2 A\varepsilon \left(a^2 + b^2\right) \cos(\underbrace{\omega_m}_{\substack{\text{within}\\ \text{baseband}}})t \rightarrow \text{ demodulation due to even order distortion}$$

which can then directly feedthrough from the RF input to the IF output with finite attenuation.

$\rightarrow$ corrupts downconverted signal.

Even-Order Distortion



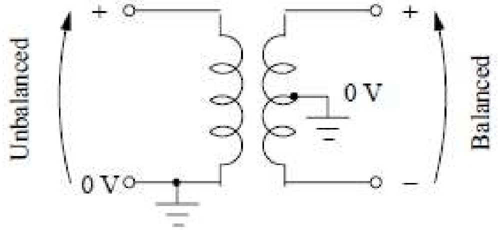
Remediation

- Need low second order non linearity for the LNA and mixer.
- Use BALUN (Balanced to unbalanced transformer) for converting antenna signals (single ended or unbalanced) to double ended (or balanced).
- Use differential LNA and mixers to suppress even order distortion.

Drawbacks

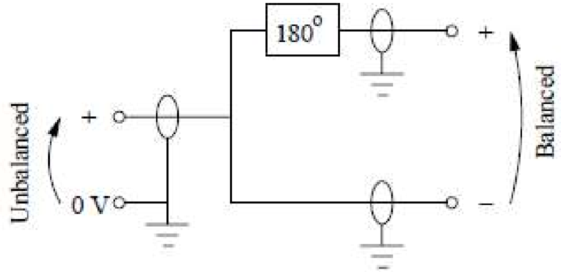
- A passive or active BALUN (balanced to unbalanced transformer is required)
- BALUN introduces a few decibels of loss degrading the SNR and consequently increases NF.
- Differential LNA and mixers require more power than single-ended devices.

Transformer based BALUN (external component)



(a)

Phase shifter based BALUN (can be on chip)



(b)

1/f Noise

1/f Noise Issue

The baseband signal after LNA and mixer can be on the order of 10 μV. Flicker noise in the baseband chain can then be a problem in particular for MOS implementation.

Remediation

- Use large devices in the baseband (less 1/f noise).
- Use active mixers rather than passive mixers to increase signal level.
- Use high-pass filtering if data is DC free.

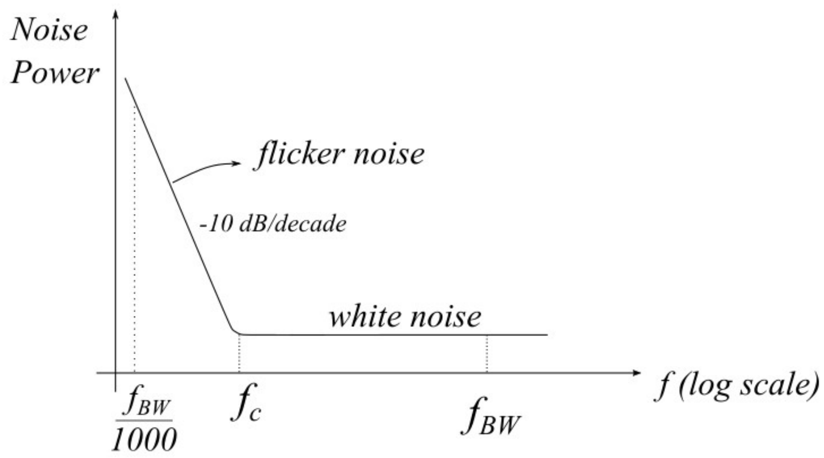
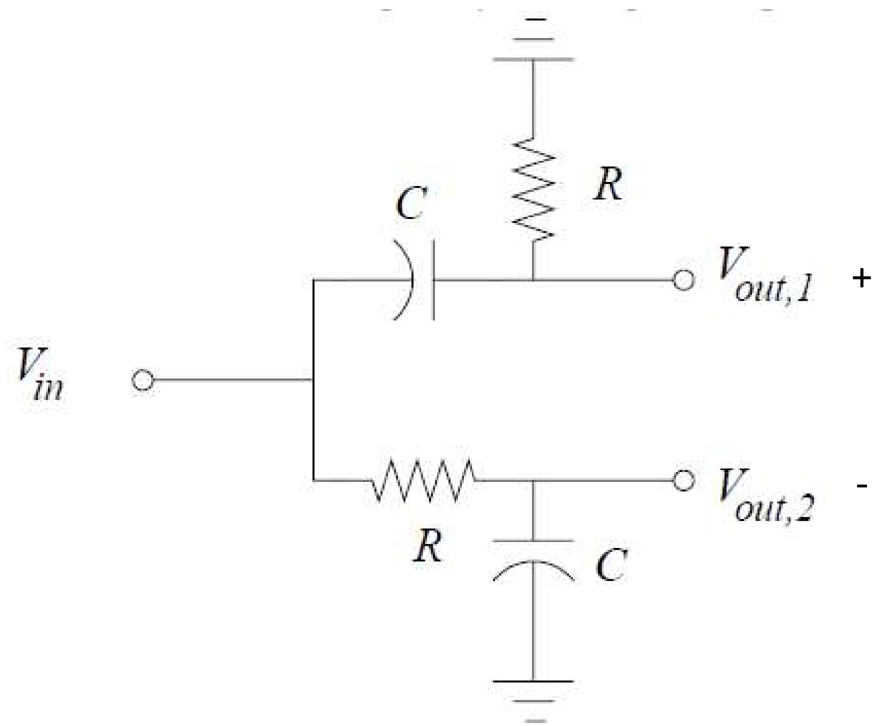


Image Reject Mixers

90° Phase Shifter

Often implemented with a bridge: $\frac{V_{out1}}{V_{out2}} = \mathrm{j}\omega\,\mathrm{RC}$

A 90° phase shift is obtained for all frequency and equal amplitude for $\omega\ RC = 1$



Transfer Function of Ideal 90° Phase Shifter

The transfer function of an ideal 90° phase shifter is:

$$G(\omega) = -j\ \mathrm{sgn}(\omega)$$

$$\sin(\omega_0 t) \rightarrow \sin\left(\frac{\pi}{2} + \omega_0 t\right) = -\cos(\omega_0 t)$$

$$\sin(\omega_0 t) = \frac{e^{j\omega_0 t} - e^{-j\omega_0 t}}{2j} \rightarrow \frac{-j}{2}\left[\delta(\omega - \omega_0) - \delta(\omega + \omega_0)\right] \quad \text{(frequency response)}$$

$$\frac{1}{2j}\left[\delta(\omega - \omega_0) - \delta(\omega + \omega_0)\right] \times -j\ \mathrm{sgn}(\omega)$$

$$= \frac{-j \times \delta(\omega - \omega_0) - j \times \delta(\omega + \omega_0)}{2j}$$

$$= \frac{-\delta(\omega - \omega_0) - \delta(\omega + \omega_0)}{2} \rightarrow -\frac{e^{j\omega_0 t} + e^{-j\omega_0 t}}{2} = -\cos\omega_0 t$$

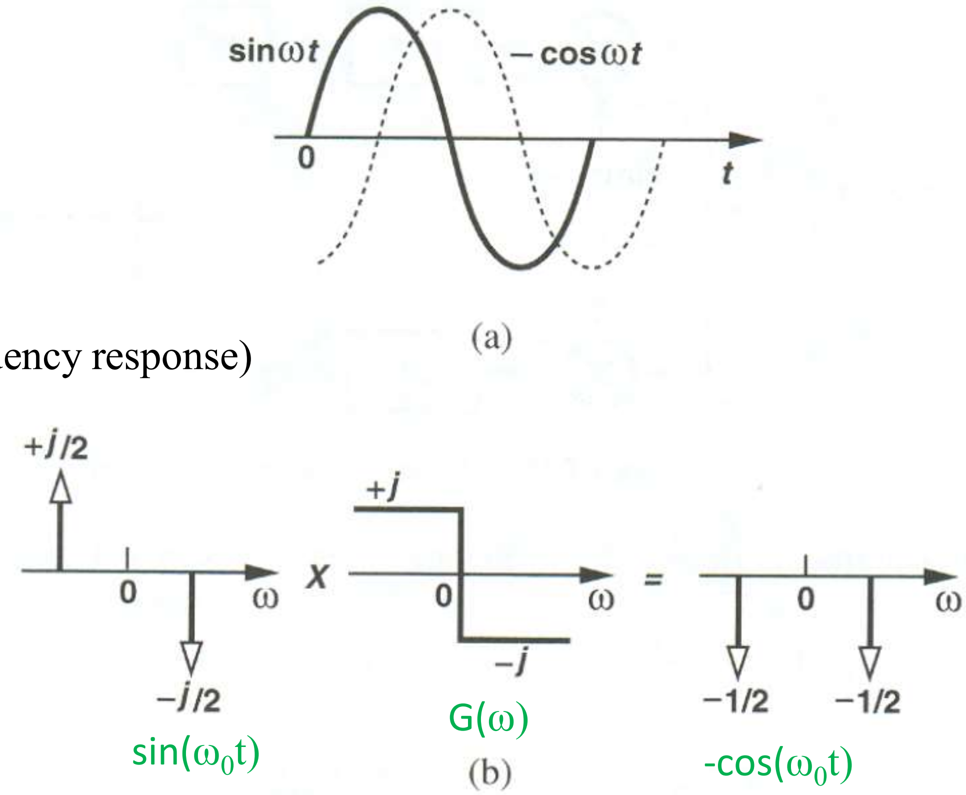
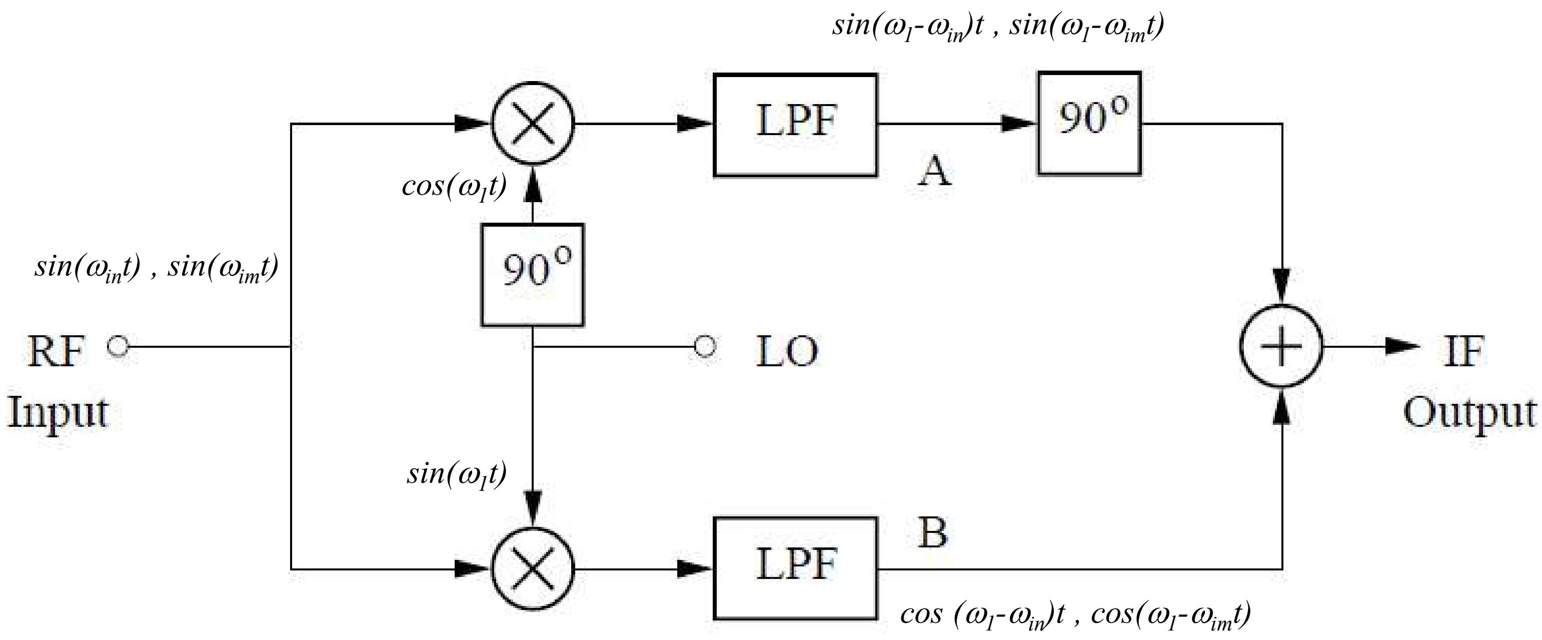


Figure 5.23 Shift by 90° in (a) time and (b) frequency domain.

Image Reject Receivers

Hartley image-reject receiver:



$$x_A = \frac{A_{in}}{2}\sin(\omega_1 - \omega_{in})t$$

$$+\frac{A_{im}}{2}\sin(\omega_1 - \omega_{im})t$$

$$= \frac{-jA_{in}}{2}\left[\frac{e^{j(\omega_1-\omega_{in})t} - e^{-j(\omega_1-\omega_{in})t}}{2}\right]$$

$$+\frac{-jA_{im}}{2}\left[\frac{e^{j(\omega_1-\omega_{im})t} - e^{-j(\omega_1-\omega_{im})t}}{2}\right]$$

$$x_B = \frac{A_{in}}{2}\cos(\omega_1 - \omega_{in})t$$

$$+\frac{A_{im}}{2}\cos(\omega_1 - \omega_{im})t$$

$$= \frac{A_{RF}}{2}\left[\frac{e^{j(\omega_1-\omega_{RF})t} + e^{-j(\omega_1-\omega_{RF})t}}{2}\right]$$

$$+\frac{A_{im}}{2}\left[\frac{e^{j(\omega_1-\omega_{im})t} + e^{-j(\omega_1-\omega_{im})t}}{2}\right]$$

$\omega_{LO} = \omega_1$

$\omega_{RF} = \omega_{in}$

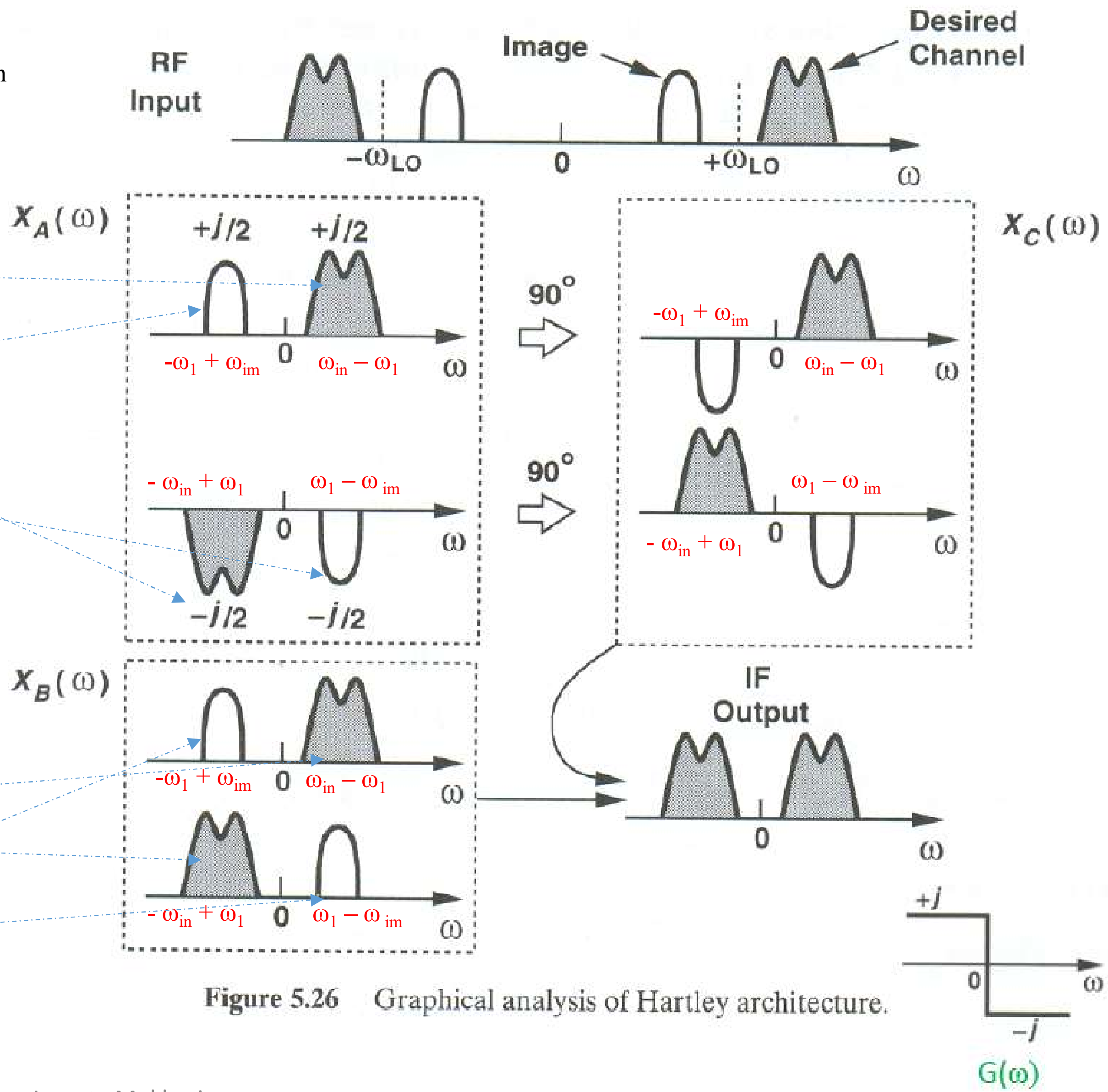


Figure 5.26 Graphical analysis of Hartley architecture.

Mismatch Analysis

In hartley architecture if the LO phases are not in exact quadrature or if gains or phase shifts of upper and lower paths are not identical, then image cancellation will be incomplete, image will still corrupt downconverted signal and image rejection filter will still be needed.

Assume the LO signal after 90° phase shift is $(A_{LO} + \varepsilon)\cos(\omega t + \theta)$.

The image rejection ratio is then evaluated (see textbook) for small mismatch.

$$IRR = \frac{\left(\dfrac{P_{im}}{P_{sig}}\right)}{\left(\dfrac{A_{image}^2}{A_{RF}^2}\right)} = \frac{(\varepsilon / A_{LO})^2 + \theta^2}{4}$$

Around 60 to 70 dB of image rejection is needed.

For the RC-CR 90° phase shifter the error at the optimal frequency is:

$$\frac{\varepsilon}{A_{LO}} \approx \frac{\Delta R}{R} + \frac{\Delta C}{C}$$

For example, $\Delta R/R = 20\%$ $(= 1/5)$ then IRR = -20dB $\Rightarrow$ only 20 dB image rejection.

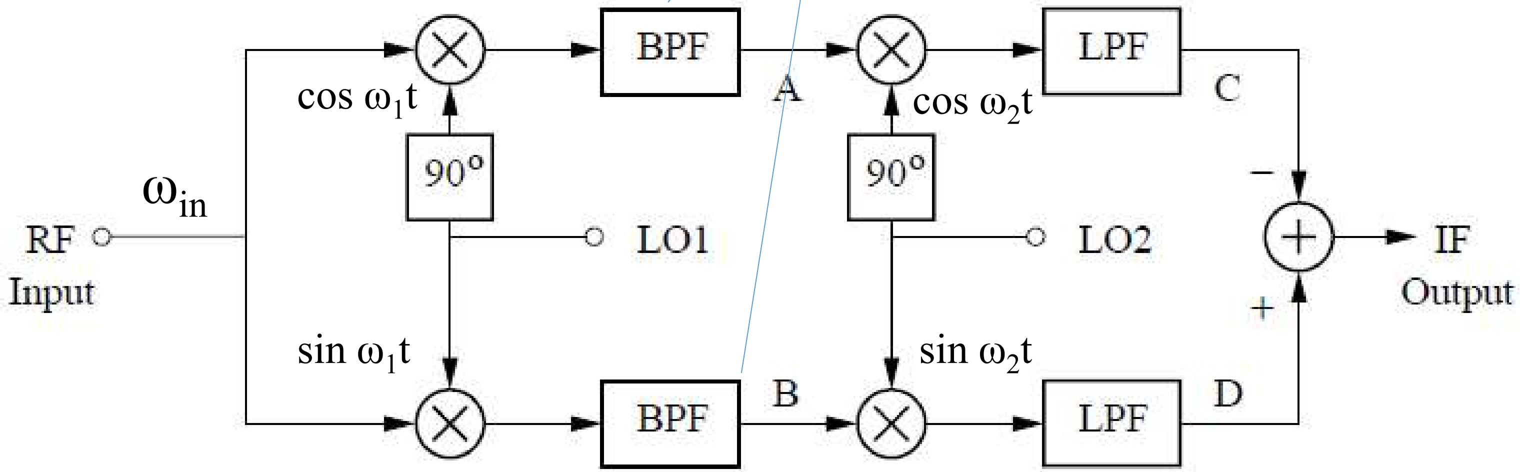
Note that the overall image rejection must be around 60 to 70 dB.

Hartley receiver(..Contd)

- Phase shifter provides 90 degree shift only at $\omega=1/RC$, for wider bandwidth the phase error may be high and IRR low.
- On chip implementation difficult as LPFs cannot fully remove interferers.
- Adder (see fig 4.66 in new edition) linearity critical (noise power trade off i.e to make it linear i/p power has to be low but then that will bring it close to noise floor).
- Loss and noise of phase shifter is significant.

Weaver Architecture

BPF instead of LPF to remove dc offsets
And second order beat frequencies



x_C (considering only ω_{in})

$$= \left[-\frac{A_{RF}}{2} \sin(\omega_{in} - \omega_1)t \right] \times \sin \omega_2 t$$

$$= -\frac{A_{RF}}{4} \left[\cos\left(\omega_{in} - \omega_1 - \omega_2 \right) t \right]$$

$$\underbrace{+ \frac{A_{RF}}{4} \left[\cos\left(\omega_{in} - \omega_1 + \omega_2 \right) t \right]}_{\text{filtered out by LPF}}$$

$$= -\frac{A_{RF}}{4} \left[\frac{e^{j(\omega_{in} - \omega_1 - \omega_2)t} + e^{-j(\omega_{in} - \omega_1 - \omega_2)t}}{2} \right]$$

Similar relations can be obtained for ω_{im} and also at D.

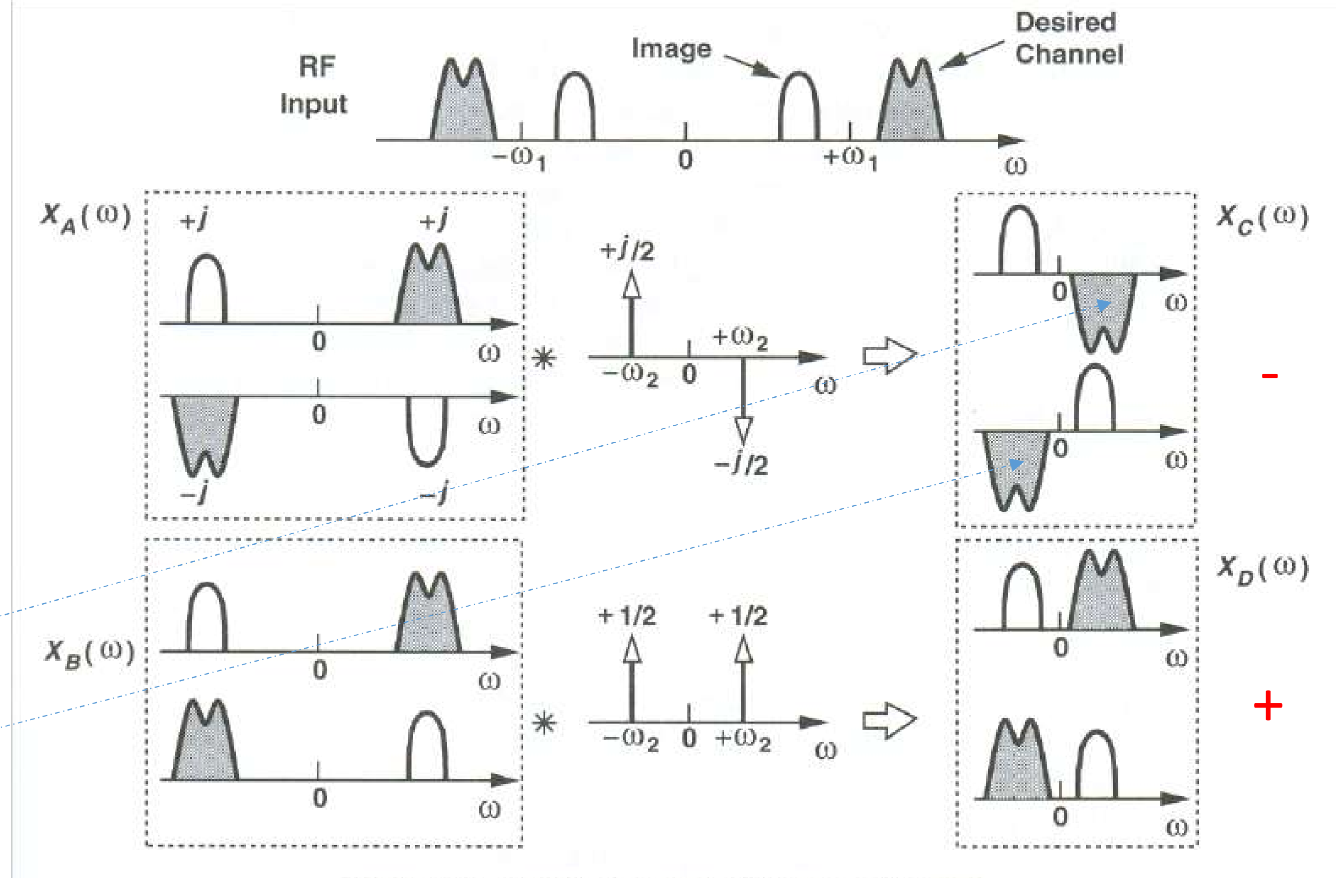


Figure 5.29 Graphical analysis of Weaver architecture.

Advantages and Drawbacks

·*Secondary Image Problem*

The input $2\omega_2\text{-}\omega_{\text{in}} + 2\omega_1$ is downconverted to $2\omega_2\text{-}\omega_{\text{in}} + \omega_1 = \omega_2 - \omega_{IF,2}$ which is the image of $\omega_{\text{in}}\text{-}\omega_1 = \omega_2 + \omega_{IF,2}$ (see Figure 5.30). A bandpass filter is required to eliminate it. Balance is also required to reject the secondary image.

–Problem can be eliminated by choosing $\omega_{\text{in}} = \omega_1 + \omega_2$ i.e. $\omega_{IF,2} = 0$

· mixer spurs - more mixing means additional mixing frequencies.

Advantages

· Offset due to self-mixing from first LO leakage and even-order distortion feedthrough are removed by the bandpass filter.

· 90° phase shifter in the RF path is not required so less chance of mismatch.

· 1/f noise problem is less of a problem due to prsence of BPF instead of LPF.

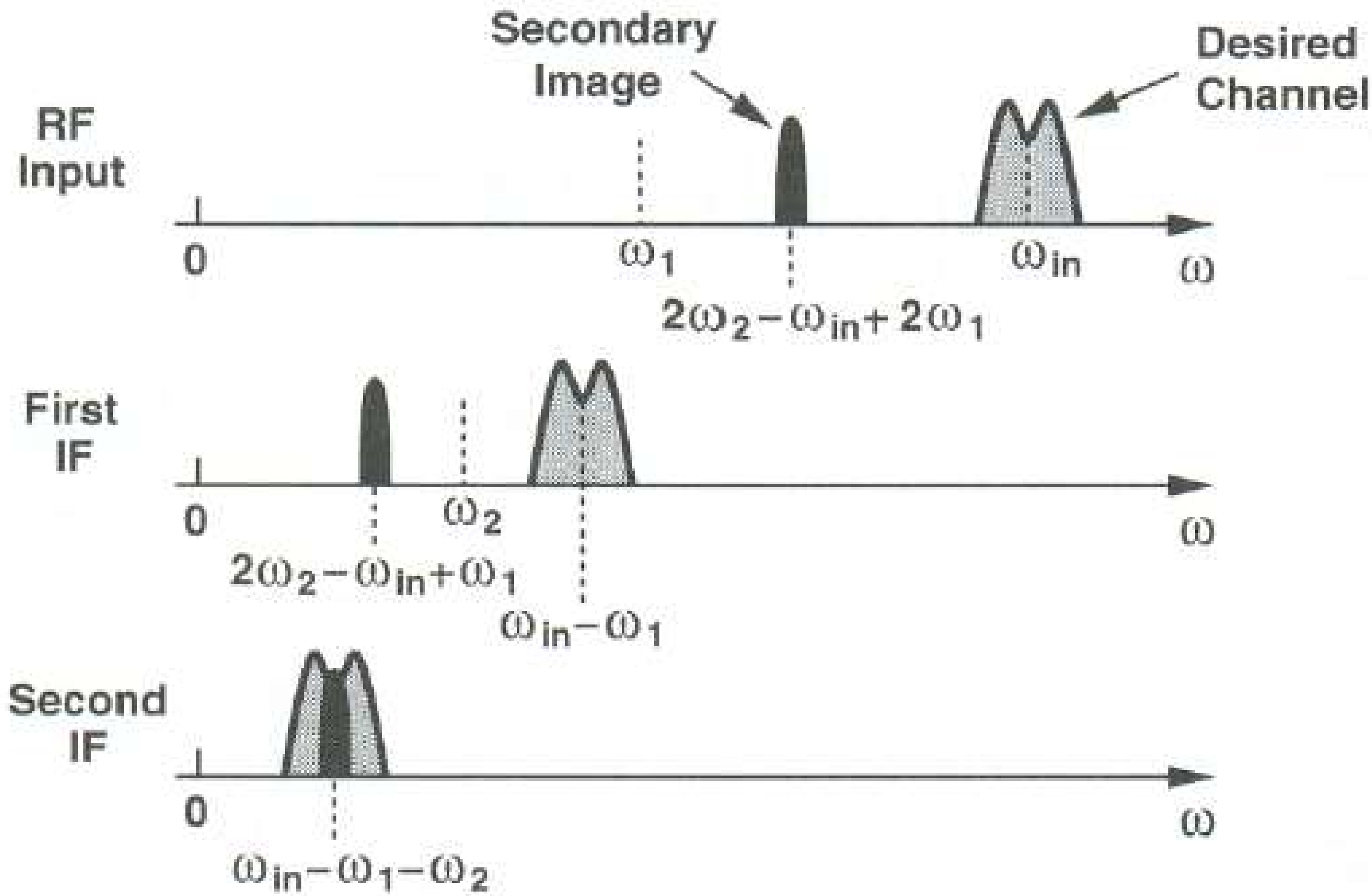


Figure 5.30 Problem of secondary image in Weaver architecture.