4.3. RR Architectures used in digital radio communications

4.3.1 Introduction

- Several types of RR architectures for digital radio transmissions can be identified.

- Some of the most widely spread architectures will be briefly discussed:
  - Super-heterodyne receivers;
  - Zero-IF receivers (or direct conversion receivers);
  - Very Low-IF receivers (VLIF);
  - Wideband double frequency change receivers.

- The analysis will consider the same structure presented in the introduction paragraph:

1. **Front end section** – the blocks which process the received RF signal in its original frequency band (RF tuned circuit, low noise amplifier (LNA), high frequency mixer, aso.)
2. **Intermediate frequency section (IF)** (if it exists) – includes the circuits which work at an intermediate frequency different from zero);
3. **Back end section** – which includes the circuits which process the signal at a lower frequency than the intermediate one (if the IF exists).

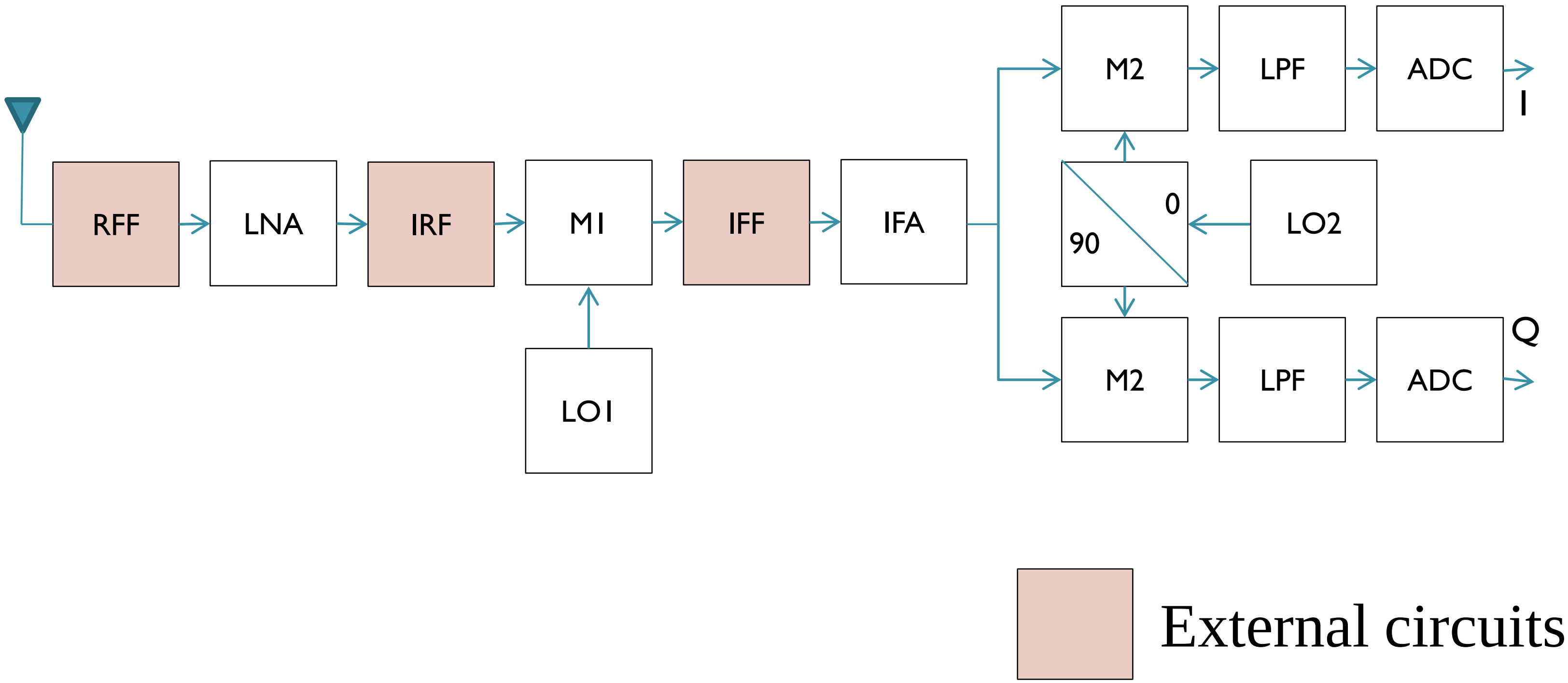
- We will focus on the factors that influence the SNR at the input of the detector or at the input of the analog to digital conversion section;
- The parameter gathers the effects of all perturbations: noise, in-band or out-of-band interference, together with the adverse effects of the RR processing: distortions, thermal noise, phase noise of the oscillators, aso.
- Finally the demodulator will convert the SNR into bit error rate (BER).

- Besides that, we will follow the behavior of the RR from the point of view of other parameters like: selectivity, DC offset, dynamic range, the effect of some signals that can block the RR aso.
- As it will be noticed, in case of digital transmissions, the final section uses a quadrature demodulator in order to deliver the I and Q signals towards the final baseband processing stage.

4.3.2 Super-heterodyne receivers

- This solution is suitable to be used in high quality RR for which performances of over **75 dB** from the gain point of view are required.

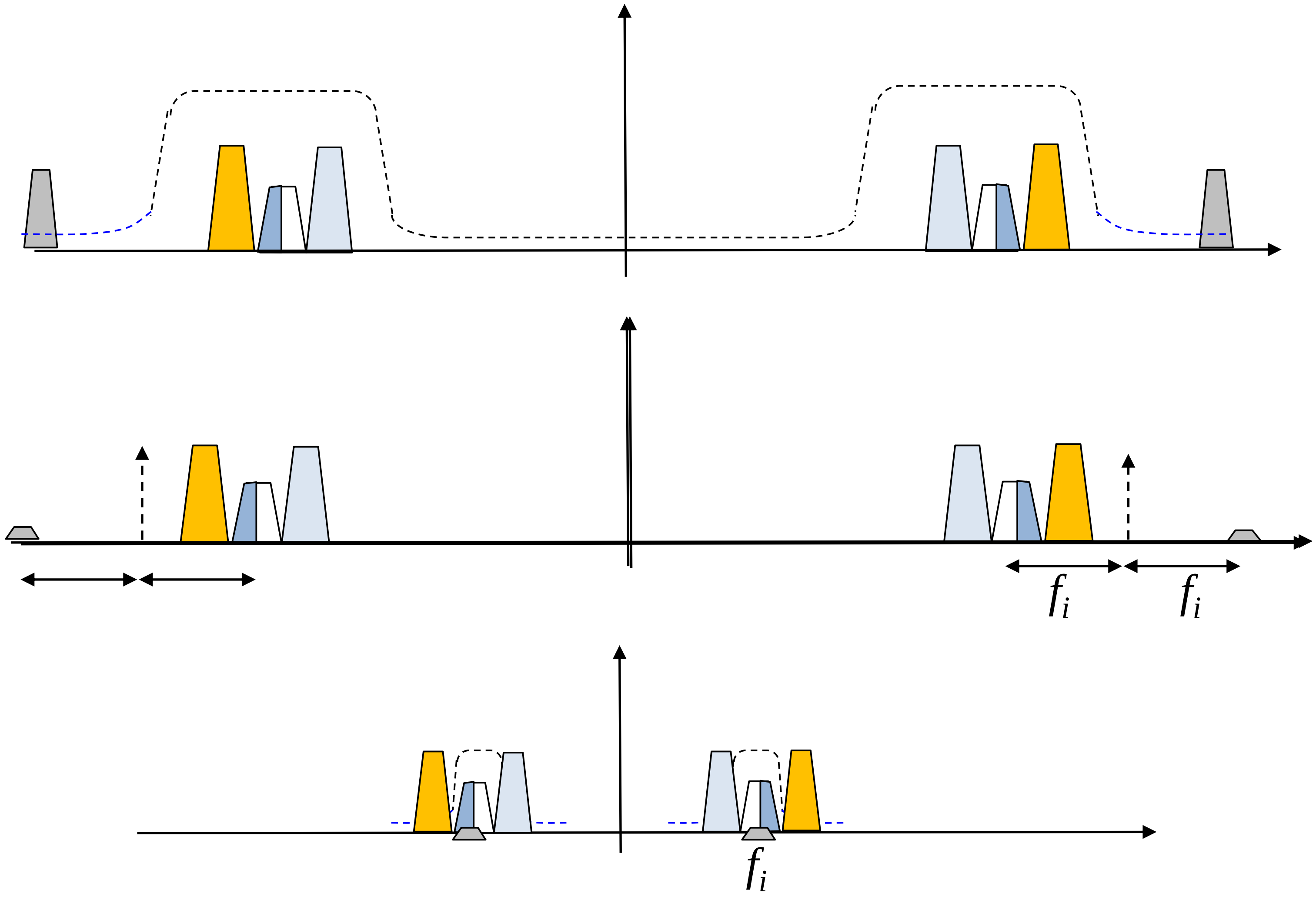
- In the next slide the block diagram of a super-heterodyne receiver with double frequency change is shown; this architecture was used in several devices: some in the 2.4GHz ISM band, others for communications in case of disasters, in the 860MHz frequency band.


M2
LPF
ADC
I
RFF
LNA
IRF
M1
IFF
IFA
0
90
LO2
Q
M2
LPF
ADC
LO1
External circuits

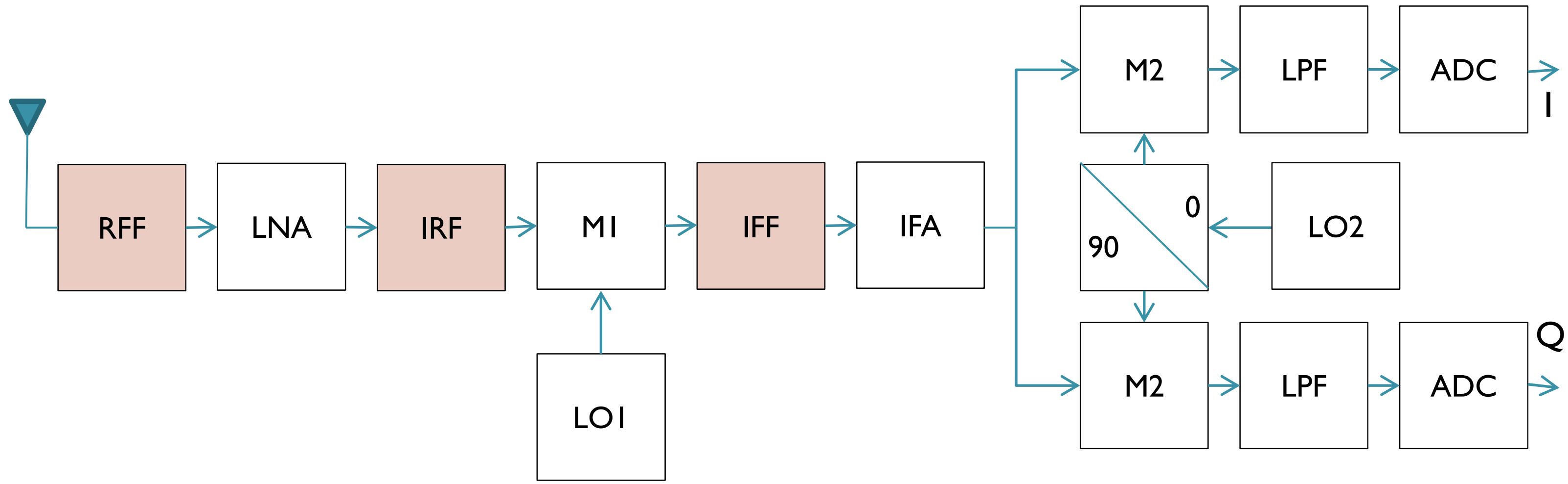
- A RF filter, which precedes the low-noise amplifier (LNA) attenuates both the image frequency and other out-of-band signals.

- The LNA ensures a gain in the order of 10-15 dB and a noise factor of about 2 dB;

- The image frequency is additionally attenuated using an image rejection filter before the first mixer (for the second application this is a BPF similar with the tuned circuit from the input).



f_i
f_i
f_i



External circuits

- The whole RF spectrum is down-converted to a fixed intermediate frequency (IF1) using a variable frequency oscillator.

- The channel selection is made using the first IF filter;

- In this way the perturbations are significantly reduced and the necessary dynamic range for the following blocks is also decreased.

- The selectivity and the sensitivity for this receiver are decisively determined by the value chosen for the intermediate frequency;

- The second frequency change is made using a quadrature mixer, in order to facilitate the digital processing of the I and Q signals.

- By processing the I and Q signals using a DSP, the final signal is obtained.

- It can be noticed the use of the I/Q conversion, necessary because one cannot process a signal with a mixed amplitude and phase modulation by using only one signal processing path;

- In the following we will review a series of specific parameters for this type of RR, using the RR used in case of disasters as example;

	First and second BPFs	LNA	First mixer	First LO	IF filter	Backend
Center frequency (MHz)	860	860	860	$f_R \pm f_{IF}$	73.35	0
Pass-band BW	19 MHz				18 kHz	2×9 kHz
Pass-band loss (dB)	1.5				3	1
Atten. @ $f_R \pm 2f_{IF}$ (dB)	35					
Atten. @ $f_R \pm 1/2 f_{IF}$ (dB)	10					
Atten. @ stop-band (dB)	45				35	60
Atten. @ >9 kHz off. (dB)					40	60
Atten. @ $f_{IF} \pm 500$ kHz (dB)					40	
Atten. @ $f_{IF} \pm 1000$ kHz (dB)					40	
NF (dB)		2	7			11
Gain (dB)		12	−7			70
IP2 (dBm)			28			
IP3 (dBm)		0	10			−20
SBN @ Adj. Ch.(dBc/Hz)				−125		−118
SBN floor (dBc/Hz)				−155		−145
IQ amplitude imbalance (dB)						0.5
IQ phase imbalance (deg)						3

Note: $f_{IF} = 73.35$ MHz; f_R = received channel frequency: 850.5 MHz $\leq f_R \leq$ 869.5 MHz.

- In order to emphasize the requirements that are imposed in the design process, we will review in the following the main elements of the quadrature demodulation;

Advantages of the I/Q processing technique:

1. The signal of band *B* is split into two signals of band less than B/2 each; in this way the two channels work at half of the processing speed (obviously the global band seen by the final section before the I/Q separation is B).

2. The IQ architecture is very suitable for generating and detecting modulated signals use for high speed data transmission.

3.The solutions allows the formal work with complex signals which makes possible the processing of single sideband signals.

- This helps a lot in understanding the process and simplifies the signal handling.

- Moreover, the treatments of arbitrary signals can be done in an unitary and efficient way.

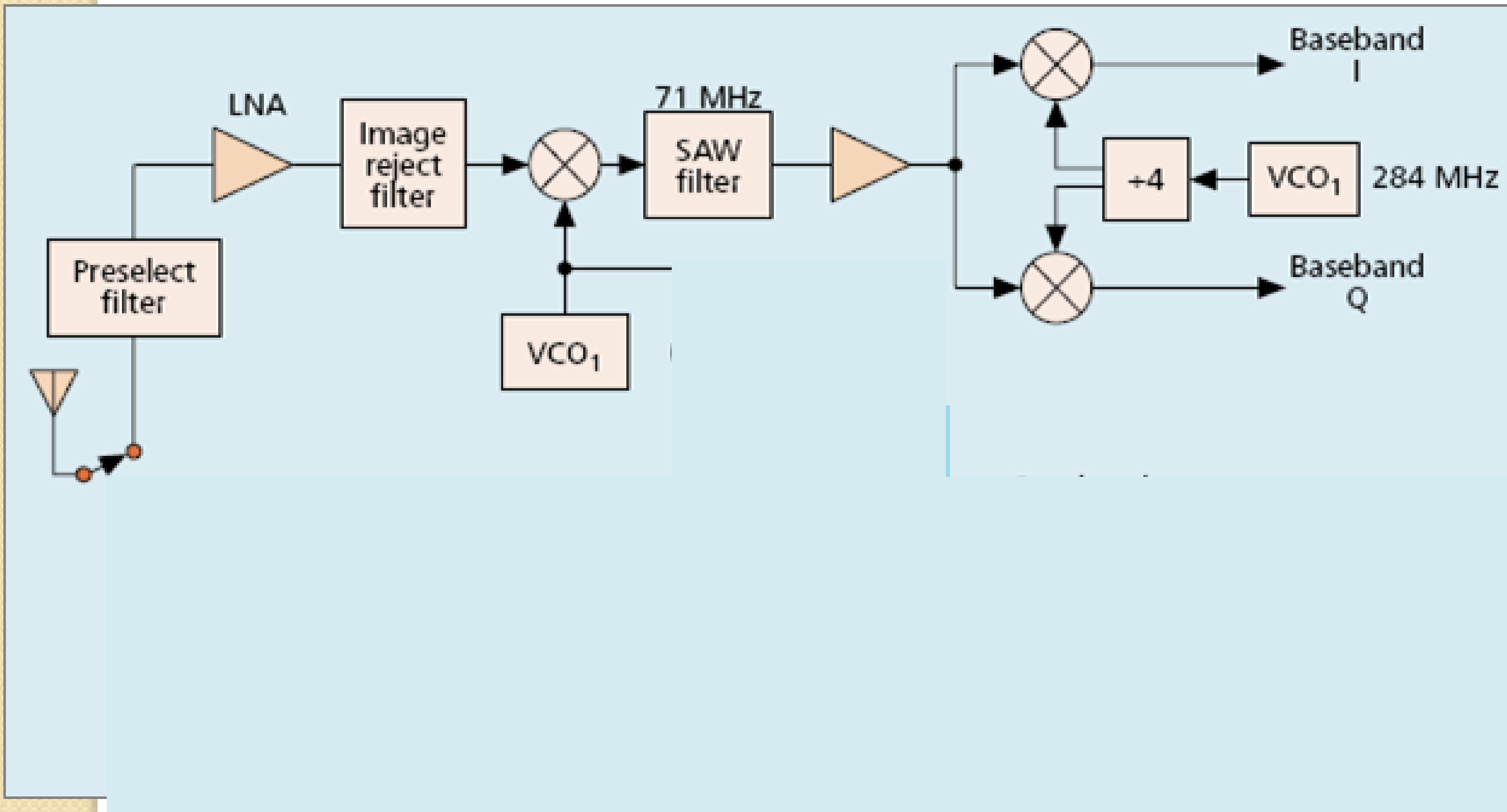
Examples of super-heterodyne RR

- The first example will be a solution used in **GSM equipment.**

- It is a solution that was used in the beginning of the evolution of the GSM technology (the mobile station variant is commented);

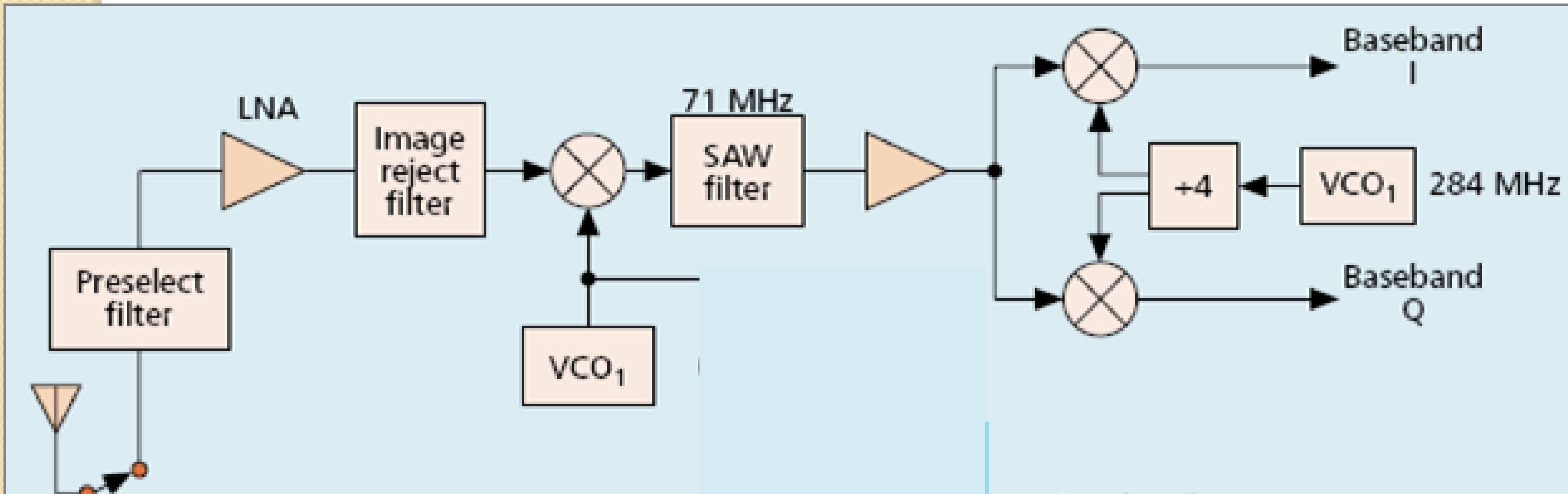
- A double frequency change can be noticed, with a first conversion from 935..960 MHz to an IF of 71 MHz.

- A second quadrature FC follows, transferring the signal into baseband and preparing the I and Q signals for the digital processing stage (back end).


Baseband
I
LNA
Image
reject
filter
71 MHz
SAW
filter
÷4
VCO1
284 MHz
Preselect
filter
VCO1
Baseband
Q

Frequency bands allocated for GSM

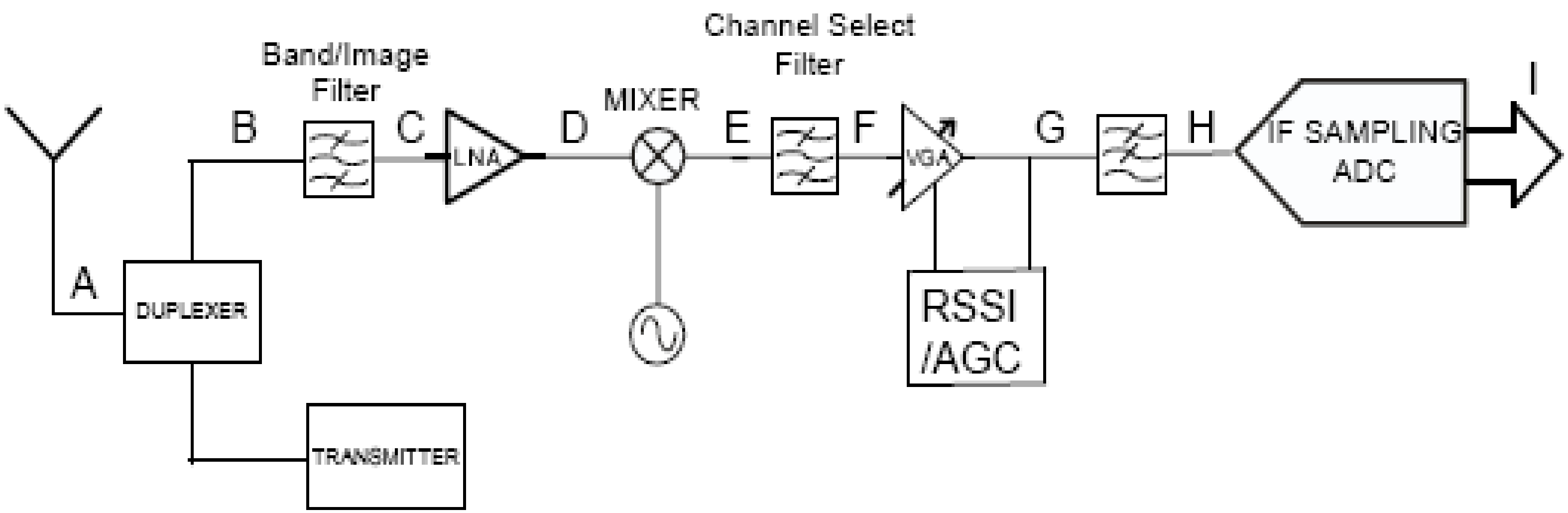
Allocated for standard GSM	Communication way
890 - 915 MHz	Mobile station → Base station
935 - 960 MHz	Base station → Mobile station



Base station → Mobile station: 935 - 960 MHz

LO1: 1006-1031 MHz

- The following type of receiver is one with only one frequency change;
- Is the most popular architecture for non-cellular communications;
- In order to suppress the signals that can block the receiver (blockers) or other unwanted components that are obtained during the mixing process, surface acoustic wave (SAW) filters are used;


Band/Image
Filter
Channel Select
Filter
MIXER
A
B
C
D
E
F
G
H
I
DUPLEXER
TRANSMITTER
LNA
VGA
RSSI
/AGC
IF SAMPLING
ADC

- It can be noticed that the power of the received signal is evaluated and a AGC system is implemented at IF level;

- The number of conversions is reduced by performing the quantization of the signal at IF level;

- The price that has to be paid: a high performance ADC;

- Concluding, the SH RR have the following characteristics:
- Are still currently considered as the most performant receiver topology from the sensitivity and selectivity points of view.

- For this architecture there are no problems related to DC offset and to the leakage of the signals produced by the local oscillator, as several conversions are used.

- Unfortunately there are problems related to cost and size.

- The external BPF with a high quality factor, necessary for the image frequency rejection and for channel selection have an important contribution from this point of view;

- As the channel selection is done at the first IF, the local oscillator has to be based on a reference oscillator with a phase noise as low as possible.

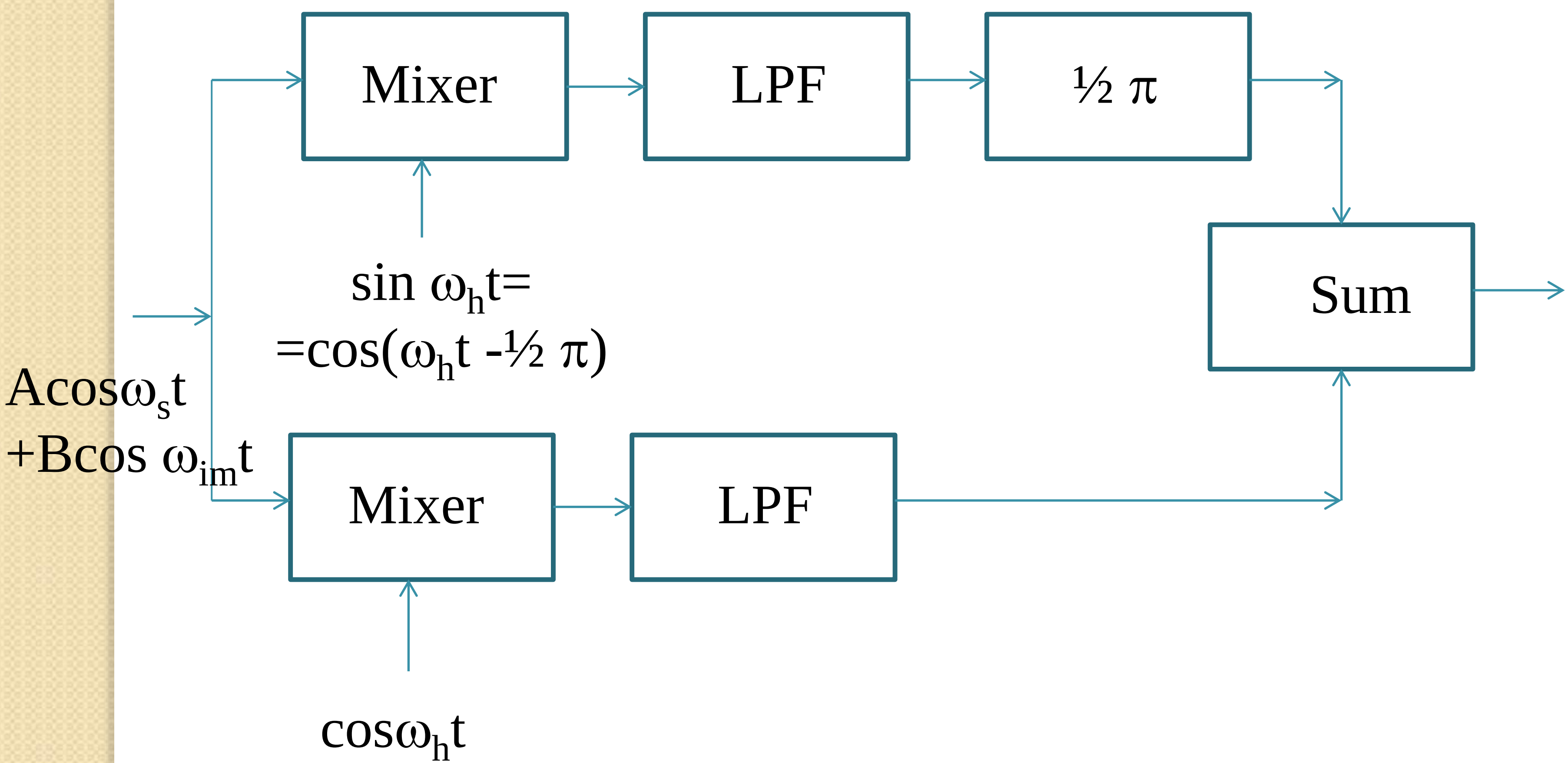
- Considering the fact that the front end section includes passive filters and the quadrature conversion is made at a fixed frequency value, excellent and reproducible performances can be obtained without imposing hard restrictions on the active components from the back end section;

- The obtaining of very good performances imply, on the other side, a high complexity, cost, number of components, size, energy consumption.

- Besides that, the integration of a transmit/receive equipment on one or even several chips is difficult.

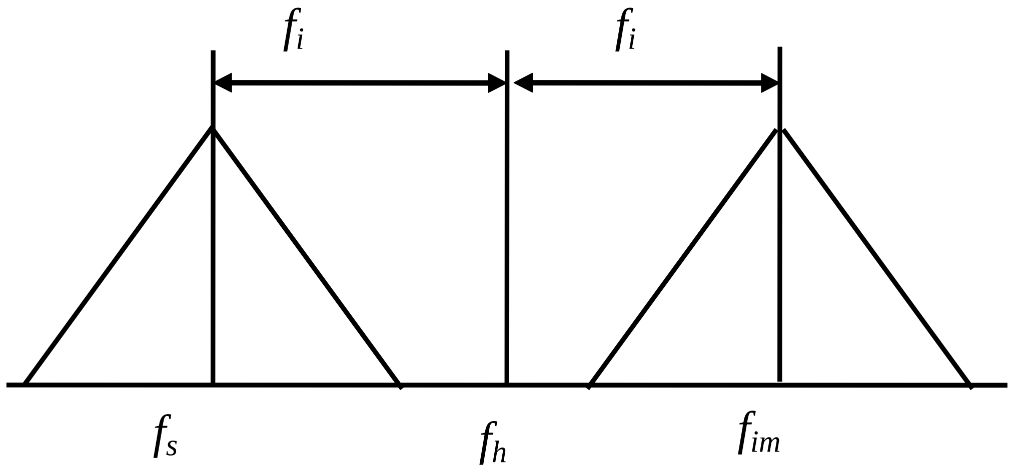
Applications: The rejection of the image frequency by use of complex mixers

1. Prove that the block diagram from the next slide can ensure the rejection of the image frequency.


Mixer
LPF
½ π
Sum
sin ω_ht=
=cos(ω_ht -½ π)
Acosω_st
+Bcos ω_{im}t
Mixer
LPF
cosω_ht

- $s(t) = A\cos\omega_s t + B\cos\omega_{im} t$ $u_h(t) = \cos\omega_h t$

$$\omega_s < \omega_h < \omega_{im}$$



$$s_1(t) = (A\cos\omega_s t + B\cos\omega_{im} t) * \cos\omega_h t =$$

$$= \frac{A}{2}[\cos(\omega_h - \omega_s)t + \cos(\omega_h + \omega_s)t] +$$

$$+ \frac{B}{2}[\cos(\omega_{im} - \omega_h)t + \cos(\omega_{im} + \omega_h)t]$$

$$s_2(t) = [A\cos\omega_s t + B\cos\omega_{im} t] * \sin\omega_h t =$$

$$= \frac{A}{2}\{\cos[(\omega_h - \omega_s)t - \pi/2] + \cos[(\omega_h + \omega_s)t - \pi/2]\} +$$

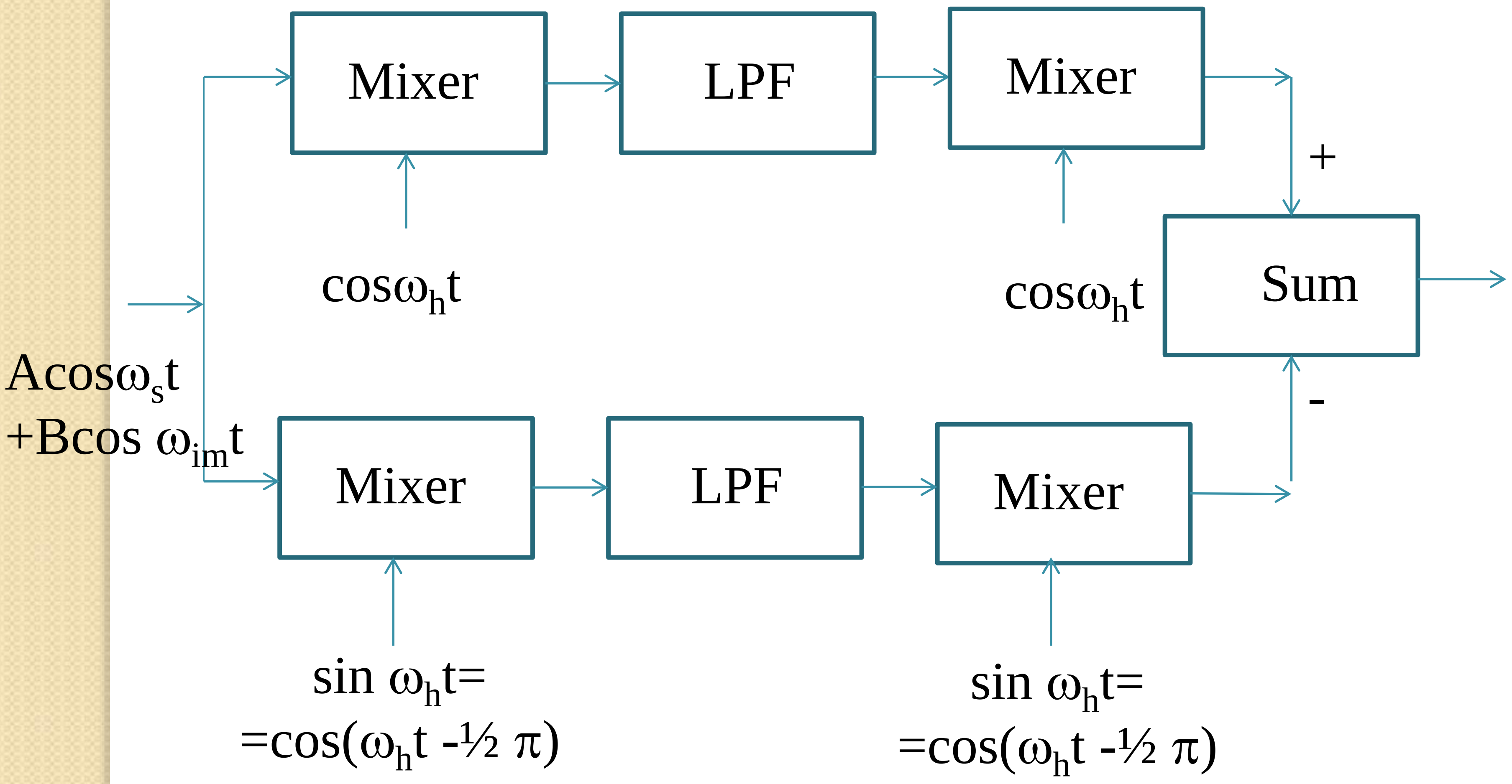
$$+ \frac{B}{2}\{\cos[(\omega_{im} - \omega_h)t + \pi/2] + \cos[(\omega_{im} + \omega_h)t - \pi/2]\}$$

$$s_{11}(t) = \frac{A}{2}[\cos(\omega_h - \omega_s)t] + \frac{B}{2}[\cos(\omega_{im} - \omega_h)t]$$

$$s_{21}(t) = \frac{A}{2}\cos[(\omega_h - \omega_s)t] + \frac{B}{2}\cos[(\omega_{im} - \omega_h)t + \pi]$$

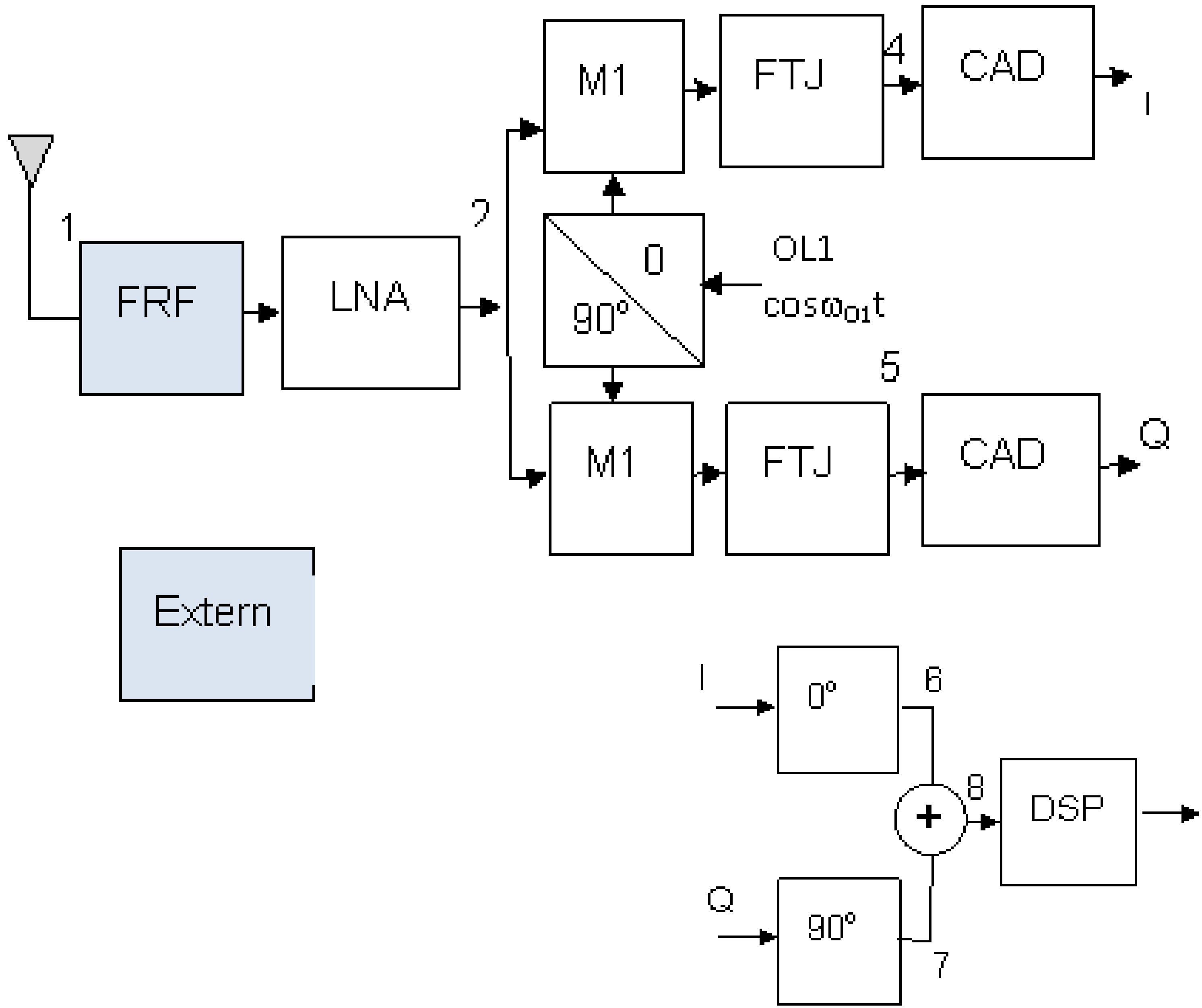
$$s_3(t) = A\cos[(\omega_h - \omega_s)t]$$

2. By using the same procedure like in the previous case, prove that the signal obtained at the output of the block diagram from the next slide doesn't contain a component on the image frequency.


Mixer
LPF
Mixer
+
$\cos\omega_h t$
$\cos\omega_h t$
Sum
$A\cos\omega_s t$
$+B\cos\omega_{im} t$
-
Mixer
LPF
Mixer
$\sin\omega_h t=$
$=\cos(\omega_h t - ½\pi)$
$\sin\omega_h t=$
$=\cos(\omega_h t - ½\pi)$

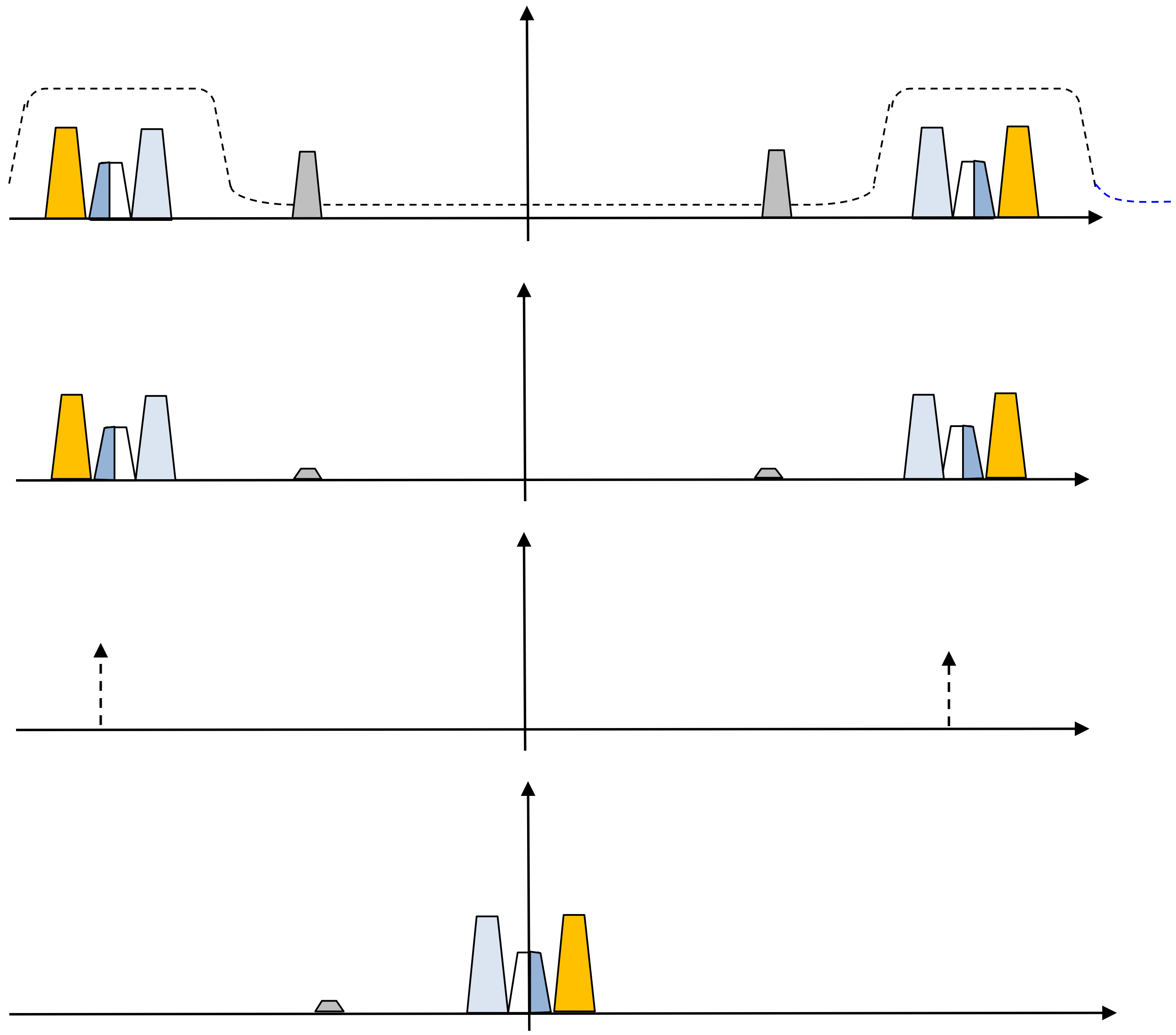
4.3.3 Zero-IF Receivers (Direct conversion receivers)

- The study and development of this architecture was imposed by the desire to eliminate off-chip components.
- A block diagram is given on the following slide.
- The diagram corresponds to communication systems that are used in unlicensed bands;

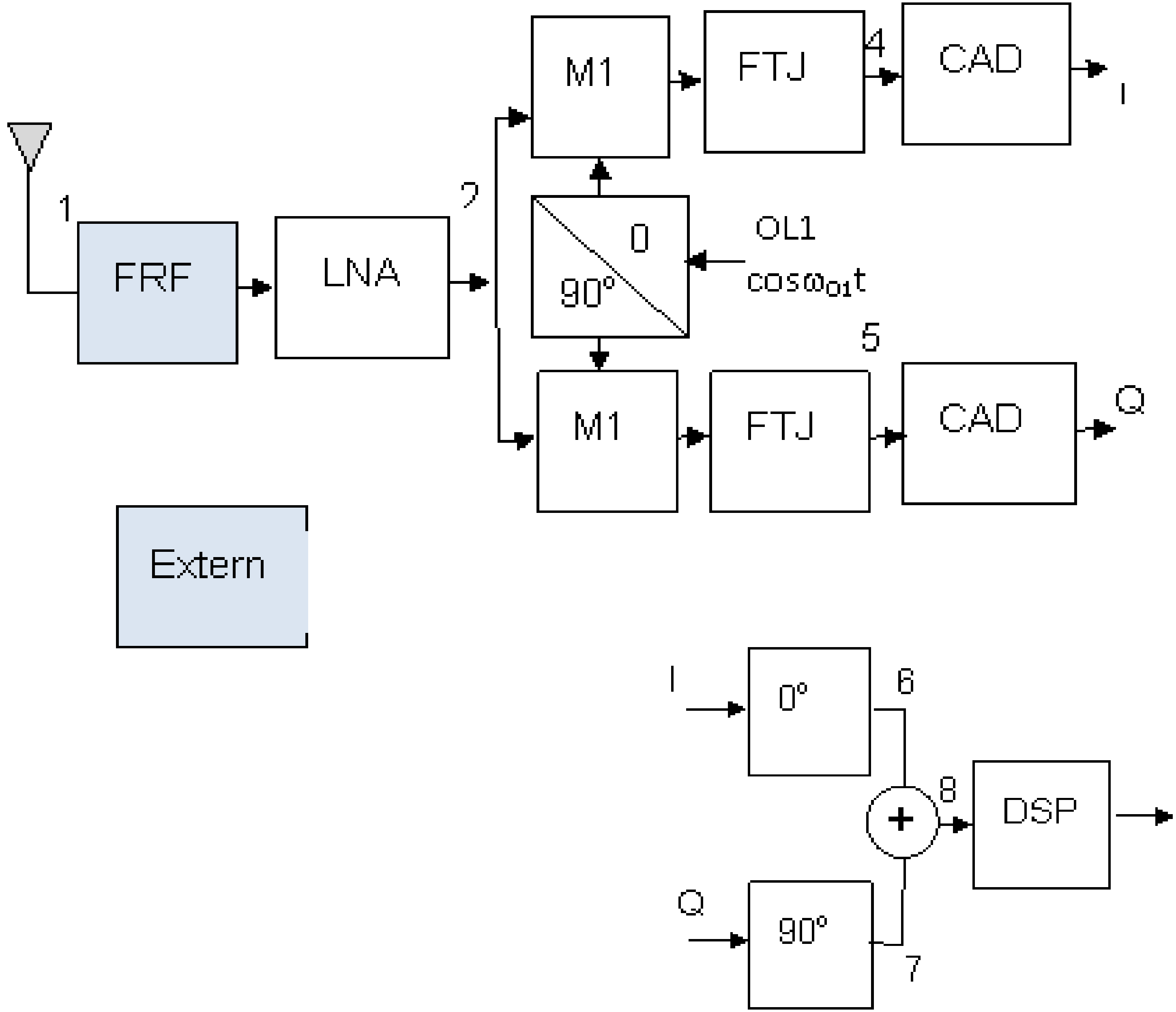


M1
FTJ
4
CAD
I
1
2
FRF
LNA
0
90°
OL1
cosω01t
5
M1
FTJ
CAD
Q
Extern
I
0°
6
8
DSP
Q
90°
7

- The RF spectrum, filtered using a band pass filter (FRF) and amplified by the LNA is down-converted to a intermediate frequency value equal with zero.
- The FRF has the purpose of eliminating the perturbation signals that, in the frequency domain, are far away from the desired signals;
- The presence of the LNA is compulsory, as the following section cannot ensure low noise;



- For channel selection, low pass filter with a steep attenuation slope are used (LPF).
- It has to be noticed that there are no problems regarding the rejection of the image frequency, so the presence of external filers with high quality factors can be avoided.
- The frequency change is implemented using a quadrature mixer (I/Q); otherwise, in many situations, the modulating signal cannot be recovered.


1
FRF
LNA
2
M1
FTJ
4
CAD
I
0
90°
OL1
cosω01t
5
M1
FTJ
CAD
Q
Extern
I
0°
6
8
DSP
Q
90°
7

- The zero-IF architecture allows the implementation of medium performance receivers;
- A set of typical values is given on next slide.
- These values are taken from the example of a RR for WiFi b/g standard, which works with 16 MHz frequency bandwidths and an interval of 20 MHz between successive channels.

Table 1.2 Typical subsystem values for a direct conversion receiver

	BPF	LNA	Backend
Center frequency (GHz)	2.45		0
Pass-band BW (MHz)	83		8
Pass-band loss (dB)	1.5		
Attenuation @ stop-band (dB)	45		70
Attenuation @ > 8 MHz offset (dB)			70
NF (dB)		2	11
Gain (dB)		12	70
IP2 (dBm)			50
IP3 (dBm)		0	12
SBN @ Adj. Ch.(dBc/Hz)			−135
SBN Floor (dBc/Hz)			−135
IQ amplitude imbalance (dB)			0.5
IQ phase imbalance (deg)			3
Flicker noise corner @ BJT (kHz)		3	
Flicker noise corner @ CMOS (kHz)		500	

Note: $f_{IF} = 0$ MHz; f_R = received channel frequency: 2.4 GHz $\leq f_R \leq$ 2.483 GHz.

- It can be noticed that many external components were eliminated, which makes this architecture suitable for integration;

- On the other side, the obtained performances are affected by:
  - Time variation of the DC component,
  - Signal leakages from the local oscillator (which has the same frequency with the received signal);
  - Fluctuation noise;

- These aspects can affect the signal detection process and imply strict requirements for the back end section compared to the previously discussed architecture.

- Besides that, a very good RF filtering cannot be performed before the frequency change, so in order to avoid the distortions that might result a mixer with a high linearity has to be used.

- The process of *self mixing* is reduced, as a single local oscillator is used for signal conversion.

- Similar to the super-heterodyne architectures, a high frequency local oscillator of variable frequency is necessary in order to perform the channels selection; because of that, the phase noise requirements are quite severe.

- The architecture is sensitive to second order intermodulation (IM2) because of the effect on the DC offset(to remark IIP2).
- The effects of the DC offset can be reduced by using an adequate DSP or implementing automatically return to zero functions;
- A comparison between the performances imposed for the SH and DC architectures will highlight that for the DC case the requirements for the back end section are much tougher;
- This aspect can be noticed by looking at the IIP3 and IIP2 values.

	First and second BPFs	LNA	First mixer	First LO	IF filter	Backend
Center frequency (MHz)	860	860	860	$f_R \pm f_{IF}$	73.35	0
Pass-band BW	19 MHz				18 kHz	2×9 kHz
Pass-band loss (dB)	1.5				3	1
Atten. @ $f_R \pm 2f_{IF}$ (dB)	35					
Atten. @ $f_R \pm 1/2 f_{IF}$ (dB)	10					
Atten. @ stop-band (dB)	45				35	60
Atten. @ >9 kHz off. (dB)					40	60
Atten. @ $f_{IF} \pm 500$ kHz (dB)					40	
Atten. @ $f_{IF} \pm 1000$ kHz (dB)					40	
NF (dB)		2	7			11
Gain (dB)		12	−7			70
IP2 (dBm)			28			
IP3 (dBm)		0	10			−20
SBN @ Adj. Ch.(dBc/Hz)				−125		−118
SBN floor (dBc/Hz)				−155		−145
IQ amplitude imbalance (dB)						0.5
IQ phase imbalance (deg)						3

Note: $f_{IF} = 73.35$ MHz; f_R = received channel frequency: 850.5 MHz $\leq f_R \leq$ 869.5 MHz.

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IQ amplitude imbalance (dB)			0.5
IQ phase imbalance (deg)			3
Flicker noise corner @ BJT (kHz)		3	
Flicker noise corner @ CMOS (kHz)		500	

Note: $f_{IF} = 0$ MHz; f_R = received channel frequency: 2.4 GHz $\leq f_R \leq$ 2.483 GHz.

Conclusion regarding the DC architecture

- The DC architecture is much simplified in comparison with the SH one: *a single LO is necessary, no IF chain is needed; therefore, cheap and small solutions can be implemented;*

- On the other side, the requirements regarding linearity, spectral purity and I/Q balance for the back end section are much tougher for the DC architecture in comparison to the SH one.

- The baseband frequencies that are close to zero are exposed to many types of perturbations like: 1/f noise, DC offset, self mixing and Doppler interferences caused by signal leakage from the LO;

- As a consequence, the baseband filters have to be designed in order to block very low frequencies;

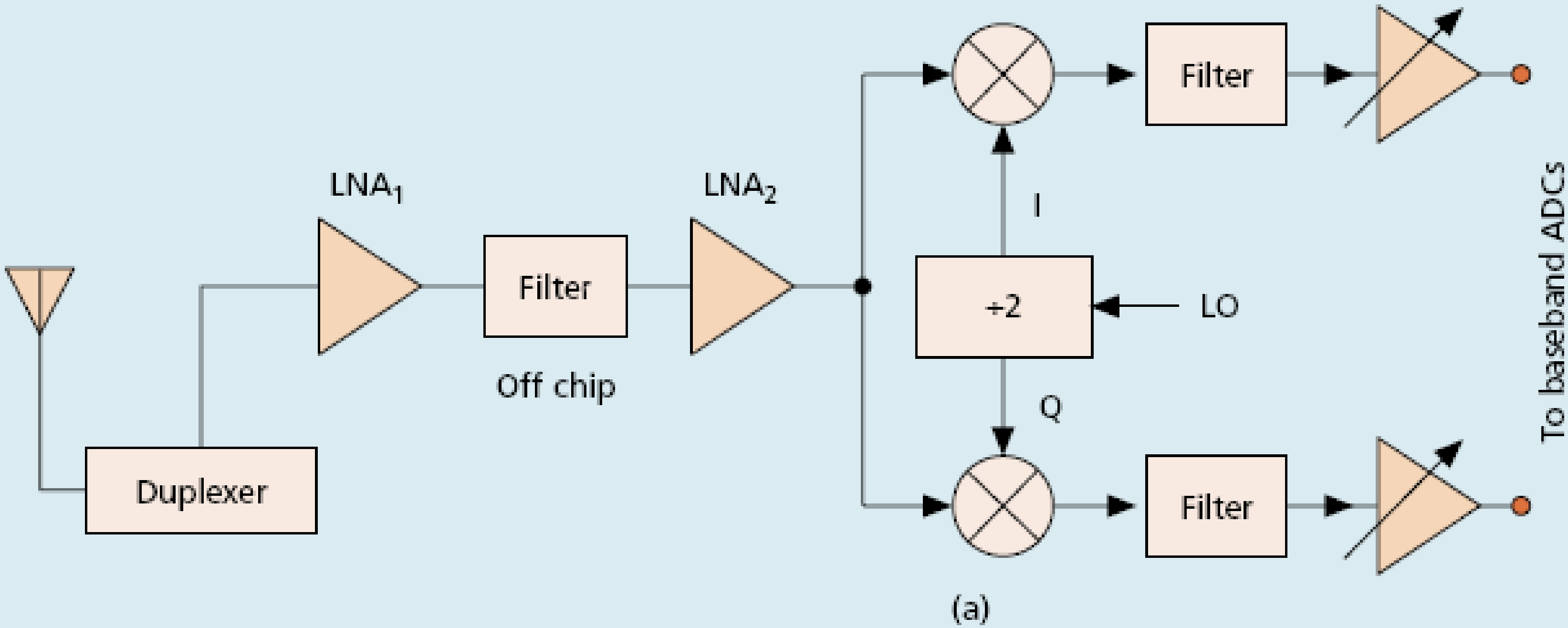
- This procedure is not critical for wideband transmissions, considering the fact that only a small fraction of the useful signal is affected, but can seriously affect narrowband transmissions;

- A series of functions that were until now implemented in hardware can be now implemented in software, so the dynamic adaptation to the frequency band of the channel and the implementation of multi-mode equipment becomes possible.

- The DC architecture was successfully used in paging systems using FSK modulation, case in which the energy consumption is low;

- It is also used in the receive part of WCDMA terminals and in IEEE 802.11 medium level modems;

- Finally, on the following slide the block diagram of a RR for WCDMA communications is given;



- It can be noticed that the LO works on a double frequency compared to the nominal one in order to avoid the frequency pulling by the transmitter;
- The obtained performances: NF= 6.5 dB (averaged on 10 kHz at 1.92 MHz), consumption 45 mW from an 1.5 V power supply, IIP3 = –8 dBm , variable gain between 16.5 and 87.5 dB; CMOS technology.

4.3.4 Very Low-IF receivers (VLIF)

- This architecture tends to combine the advantages of the two architectures discussed before (SH and DC), but it obviously has its own specific disadvantages;

- An IFA on a f_i frequency, having a value between ½Δf and *n*Δf , where Δf is the distance between two channels and *n* has small values (a few units);

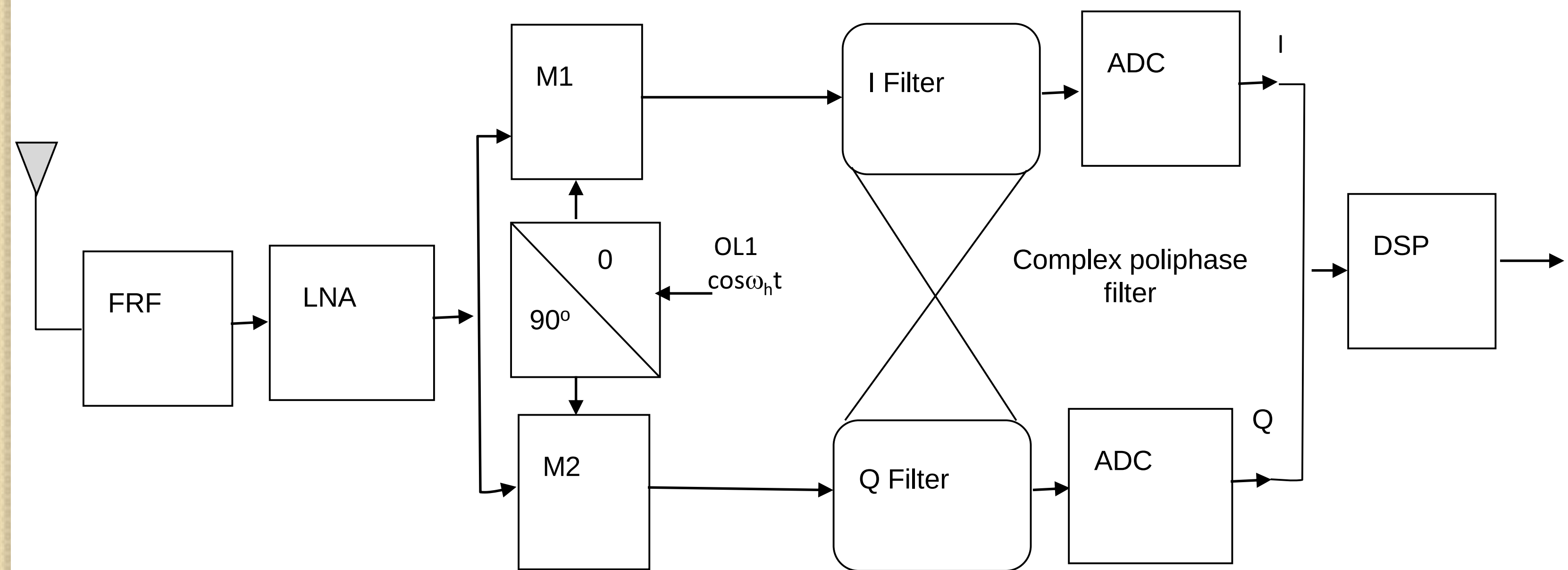
- In this way, the channels selection filter could have a lower quality factor (low-Q).

- This topology became more interesting with the development of on-chip pass band filters.

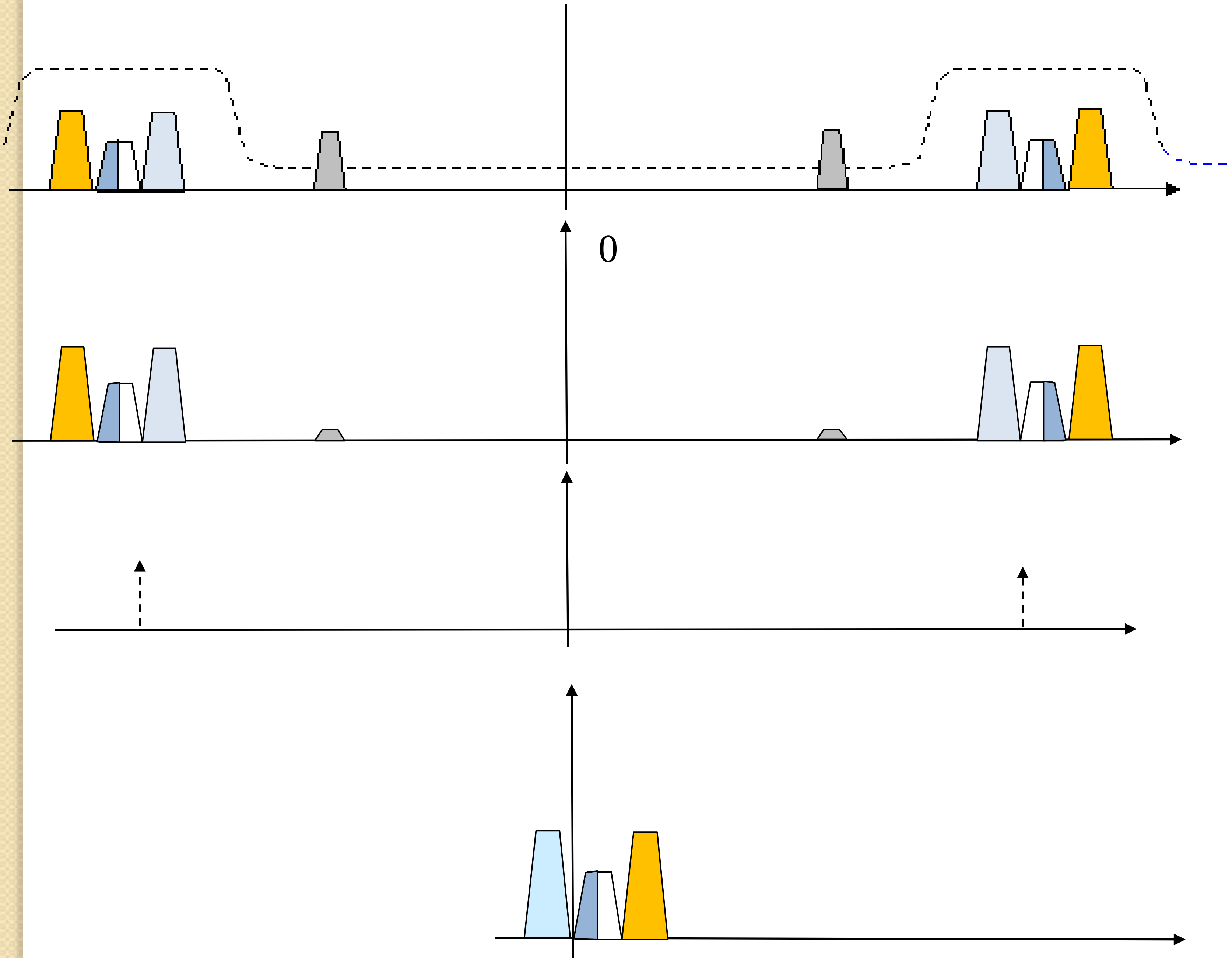
- The architecture is interesting especially in modern cellular communications, where is exploits the low level of the neighboring channel.

- The VLIF architecture was implemented in a great number of variants.

- We will briefly present a variant used in a series of GSM receivers.



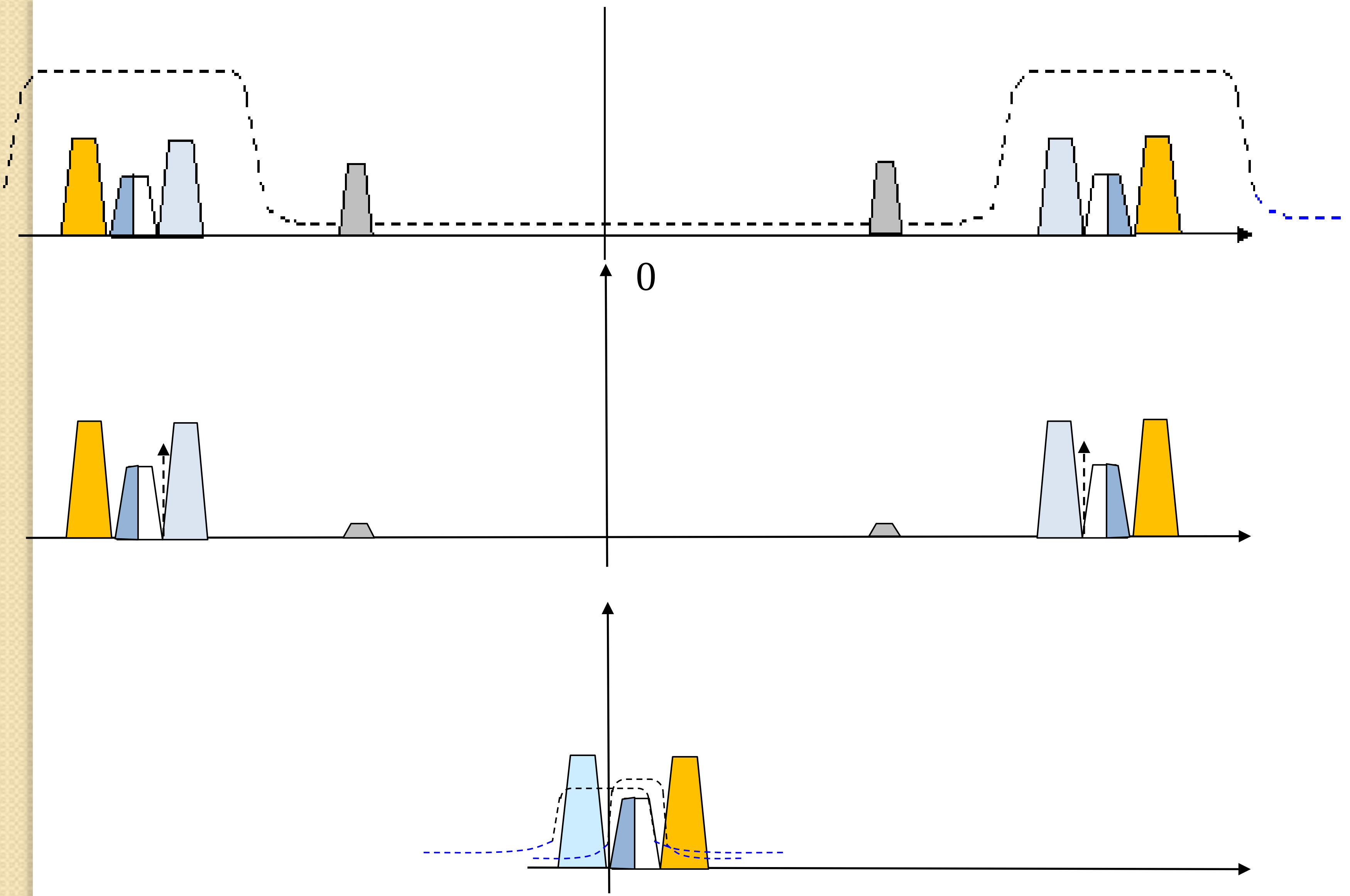
- For this example the receivers where f_i is equal with Δf/2 are considered;
- The spectral diagrams presented on the next slide are obtained;


0

- After an RF filtering and the low noise amplifying, the received signal is mixed in quadrature with a local signal $f_h = f_R - f_i$;

- After the mixing, the most dangerous interferences (one of the neighboring channels and the image channel) are shifted in frequency domains where no low pass filter can eliminate them (the neighboring channel will be located between -Δf and 0).

- *It will be noticed that this filtering cannot be performed with low pass filters located on the I and Q branches, like it was done in the DC architecture.*


0

- It can be proven that the filtering can be done by means of an interaction between the two processing paths: I and Q;

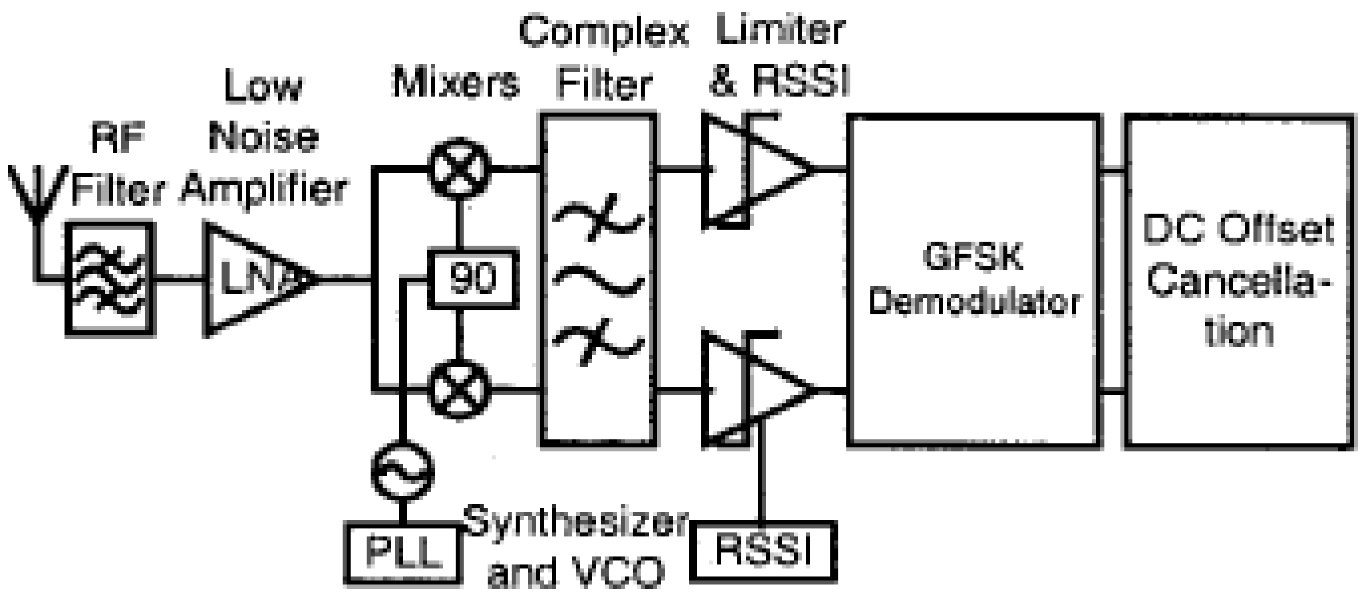
- This is equivalent with the implementation of a complex poliphase filter;

- It can be shown that this processing is equivalent with a band pass filtering which will select the desired signal and eliminate the unwanted ones.

- The demonstration is based on the analysis of the signals in the complex domain using the complex envelope, as it was defined in the previous paragraphs.

- Returning to the very low IF receiver;
- If was noticed that the polyphase filter can be also implemented entirely after the AD conversion on a DSP;
- In this case, the same architecture like the one used in case of DC receivers can be used, with only one change: the frequency value of the local oscillator.
- There are several situations where on the same chip, both the DC and the VLIF architectures are implemented.

- We will give an example for the VLIF architecture by discussing a Bluetooth receiver (2.4 GHz).
- An integrated version implemented in the laboratories of the Texas University is given in next slide.



- Channels of 1 MHz each, GFSK modulation;
- Intermediate frequency: 2MHz;
- Measured performances:
  sensitivity: -82dBm (for a BER of 10^{-3}), noise factor: 15dB, IIP3: -10dBm.

Conclusion regarding the RR based on this architecture:

- Opposite from the DC architecture (zero-IF), the VLIF architecture is not sensitive to DC offset, to signal leakage from the LO and to order 2 intermodulation.
- The obtained receivers can be integrated at the same level like the ones designed using the DC architecture;
- The requirements for the back end section are reduced compared to the case of DC;

- From the performance imposed to the functional sections point of view, the values are comparable to the ones from the DC case, with little changes in case of the attenuation of the back end section;

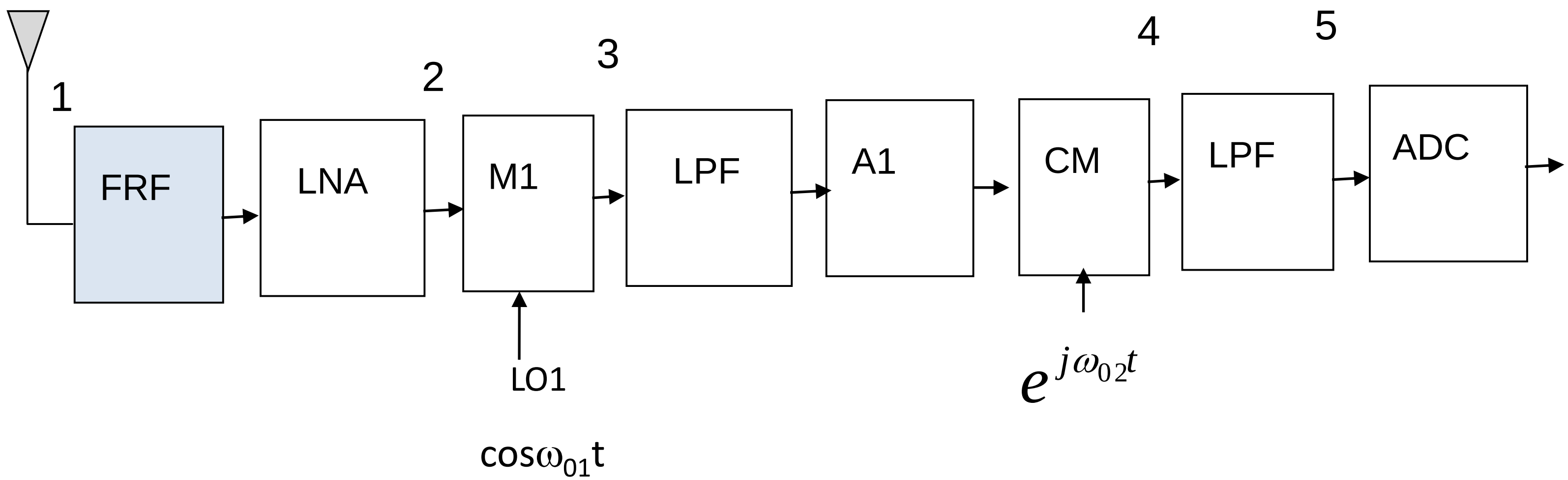
- The I and Q components have to be simultaneously processed as a complex signal using a polyphase filter.
- In applications which imply a low cost for the RR, analog polyphase filters are used, which will lead to more modest performances because of the limited precision for the balancing of the two paths (I and Q);
- As it was mentioned before, the image channel rejection in this case is limited to about 40dB.

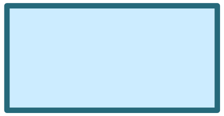
2.2.4 Wideband double frequency change receivers

- For wideband communications, a solution that combines the SH architecture with the DC one was proposed; this kind of solution allows an optimization of the power consumption and of the performances.

- The block diagram is presented on the following slide.

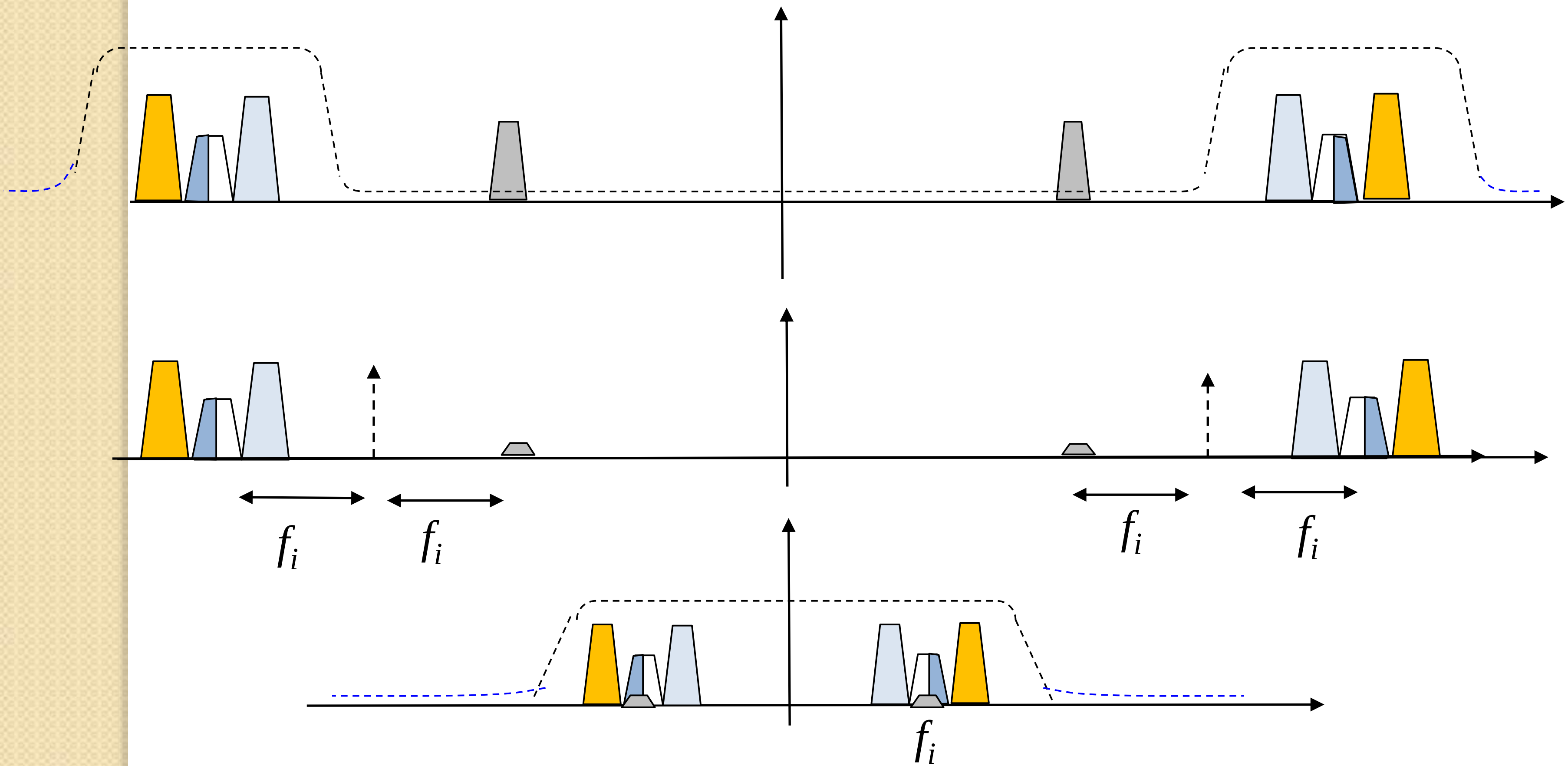
- It will be noticed that this solution is similar to the SH one, with the first IF value very high (hundreds of MHz).



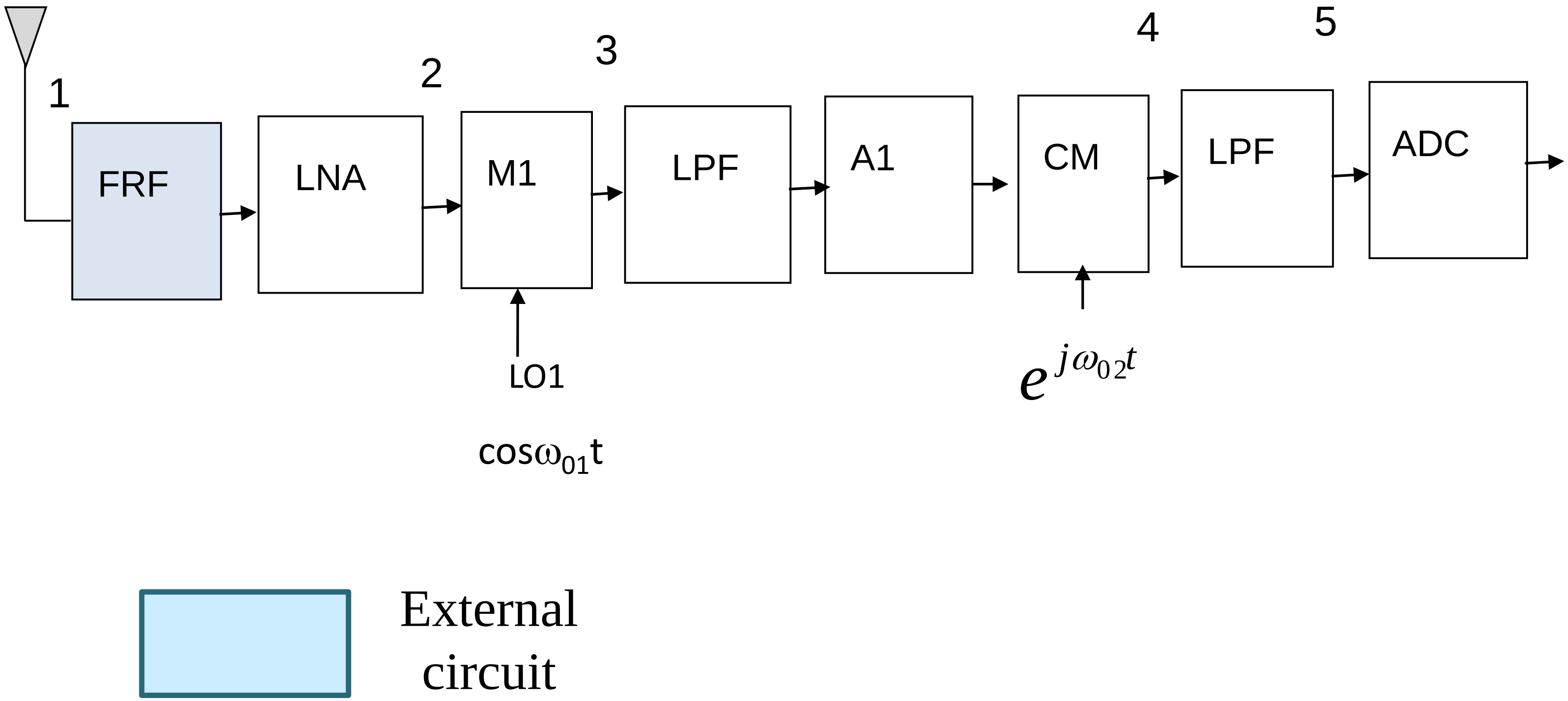


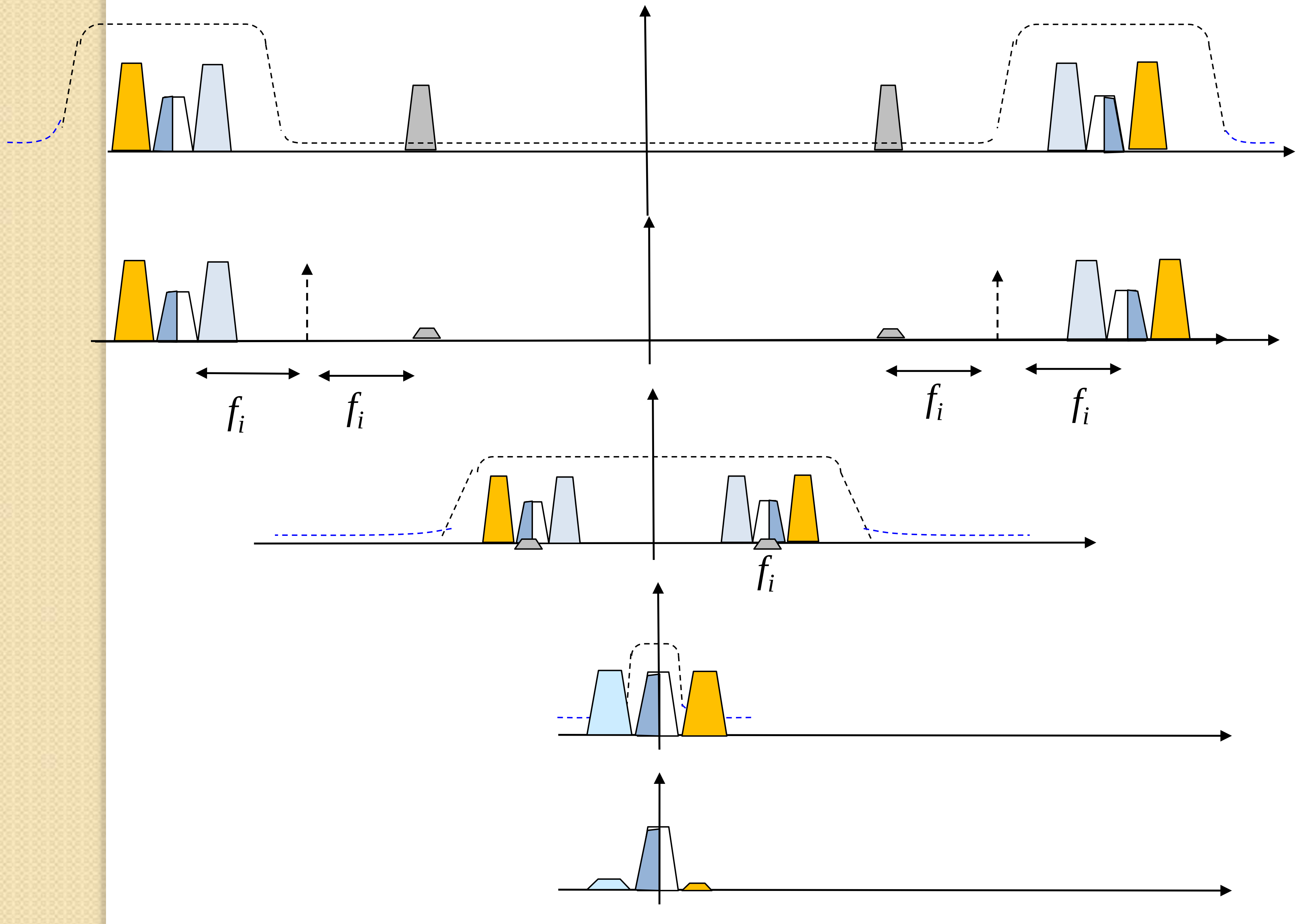
External circuit

- After an RF filtering, the first frequency change takes place, using a real mixer to downconvert the signal;
- As the intermediate frequency hase a large value, the RF filter can eliminate the image channels;
- Moreover, the perturbation signals (including the LO one) are located at frequencies quite far away towards higher frequency values, therefore a LPF can be used to select the desired band;


f_i
f_i
f_i
f_i
f_i

- Moreover, the **first LO has a fixed frequency**, so it can have good performance, having no major problems in designing and implementing low phase noise and high frequency synthesizers, if they work on fixed frequency.
- A second conversion towards baseband follows, using a **complex mixer** and a variable frequency local oscillator, which performs the tuning on the desired channel.
- The selection of the desired channel is performed using a LPF, which might have less power consumption than a BPF.


1
2
3
4
5
FRF
LNA
M1
LPF
A1
CM
LPF
ADC
LO1
$\cos\omega_{01}t$
$e^{j\omega_{02}t}$
External circuit



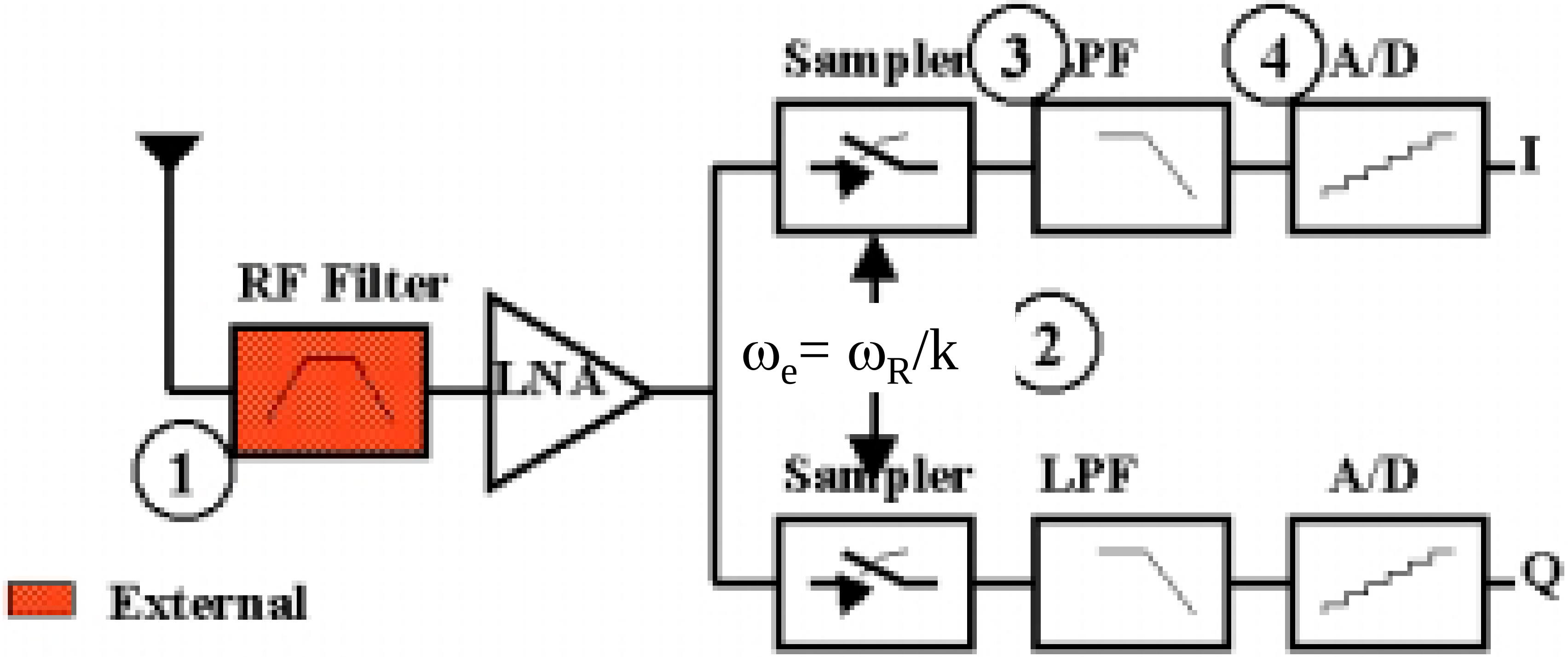
f_i
f_i
f_i
f_i
f_i

- The problem of signal leaks from the LO1 is not very serious, but the receiver is still sensitive (although in a lower degree) to the DC offset and to 2nd order intermodulation.
- The performances can also be affected by the diaphony which appears between the RF and IF paths.

- According to measurements, a rejection of image channels of 55 dB can be obtained by using this architecture;
- This is obtained not only by filtering, but also by using a great number of mixers which have be possess a good linearity in order to ensure the required dynamic range.
- In order to reduce the power consumption, at least some of them can be implemented using passive solutions.

2.2.5 Sub-sampling receivers

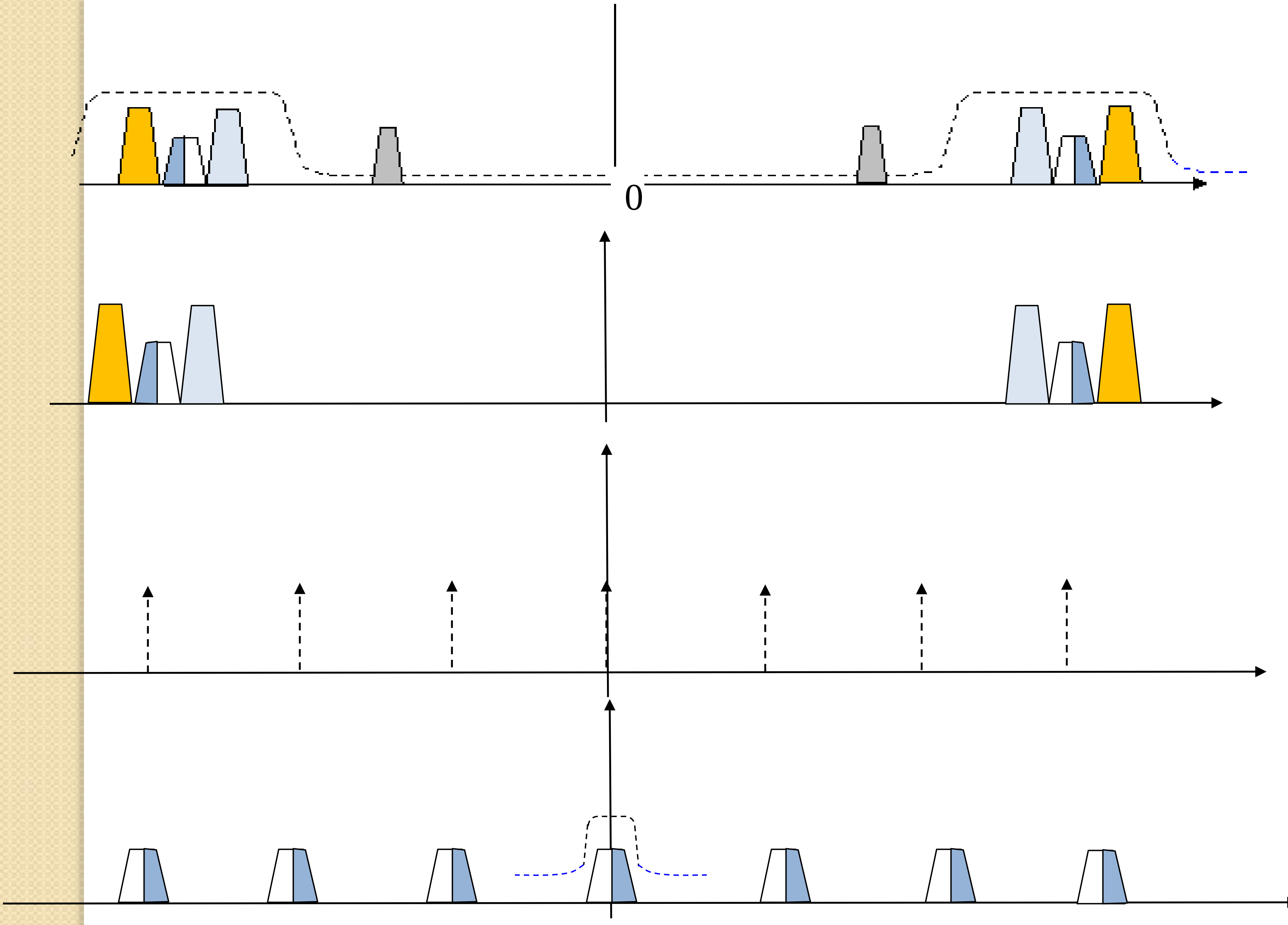
- Together with the development of the high-speed CMOS technology, several sub-sampling architectures were conceived and implemented;
- These are based on the band-pass sampling theorem;
- The sampling is performed at RF level, like it can be seen in the following slide.

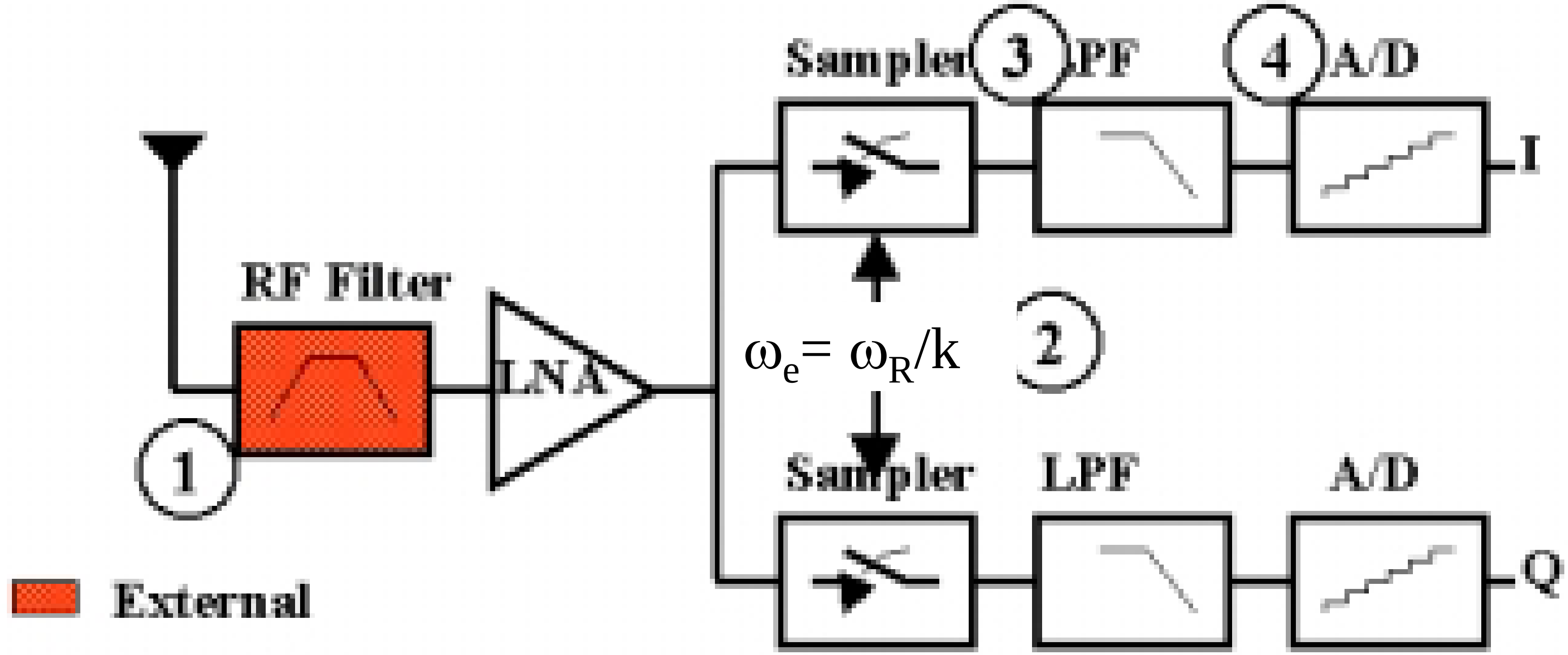

Sampler
3
LPF
4
A/D
I
RF Filter
LNA
$\omega_e = \omega_R / k$
2
1
Sampler
LPF
A/D
Q
External

- The sampling is performed using a sampling frequency equal with the one that would be used in the baseband.
- After the band-pass sampling, spectral images are obtained at locations given by the following expression:

$$\omega_i = k\,\omega_e \pm \omega_R \qquad (12)$$

- Where:
  - k is a constant integer;
  - ω_i corresponds to the center frequency of the spectral image ($\omega_i = 0$ corresponds to the sub-sampling architecture with the intermediate frequency equal with zero);
  - ω_e and ω_R correspond to the sampling frequency and to the carrier frequency.


0


RF Filter
LNA
Sampler
LPF
A/D
I
Q
$\omega_e = \omega_R / k$
1
2
3
4
External

- Concluding, the sub-sampling architecture is suitable for integration, especially in CMOS technology, because the complex process of downconversion is reduced to a sampling operation.
- The sampling operation performed at high frequencies makes the design of high speed switches an important matter.
- As the necessary sampling frequency is much lower than the carrier frequency, the requirements imposed for the clock oscillator are not too harsh and it can be implemented with low consumption.

- One of the main problems of this architecture is the phenomenon of noise aliasing.
- Noise power is increased with a $2*k$ factor, so is advisable to limit the noise bandwidth as much as possible before the sampling operation;
- In order to do this an external band pass filter has to be used.
- The fluctuation of the sampling clock is increased with a k^2 factor, which might lead to interferences in the desired channel.

- Another problem for this architecture is related to the parasite transmission of the clock oscillation and the incomplete stabilization of the working regime of the operational amplifiers.
- Because of these last aspects, it results an insufficient attenuation of the interferences, so it is necessary to use A/D converters with a high dynamic range.

Annex 4.3.1 Aspects regarding I/Q demodulation

- We will start the analysis with a very general expression of a signal having a mixed modulation with the envelope A(t) and phase $\phi(t)$,where $-\pi/2 \leq \phi(t) < \pi/2$:

$$s(t) = A(t)\cos(\omega t + \phi(t)) = Re[A(t)e^{j(\omega t+\phi(t))}]$$

- It can be noticed that:

$$s(t) = Re[S_c(t)] = s_r(t)\cos(\omega t) - s_i(t)\sin(\omega t)$$

$$S_c(t) = [A(t)e^{j\phi(t)}]e^{j\omega t} = [s_r(t) + js_i(t)]e^{j\omega t}$$

$$s_r(t) = A(t)\cos(\phi(t)),\ s_i(t) = A(t)\sin(\phi(t))$$

To be reminded that the expression: $A(t)e^{j\phi(t)} = s_r(t) + js_i(t)$ represents the complex envelope of the baseband signal;

- Reciprocally, if $s_r\ (t)$ and $s_i\ (t)$ are known, than $A(t)$ and $\phi(t)$ can be immediately deduced by using the known expressions:

$$A(t) = [s_r(t) + \mathrm{j}s_i(t)]^{1/2} \cdot sign[s_r\,(t)]\ )$$

$$\phi(t) = tan^{-1}[s_i\,(t)/s_r(t)] \in [-\pi/2, , \pi/2)$$

- The RF signal s(t) is mixed with two local signals having exactly the same amplitude and frequency,

- One of them (the I path) in phase with the carrier frequency and another one (the Q path) shifted with 90 from this one.

- It can be shown that a copy of the $s_r(t)$ signal is recovered at the output of the I path and a copy of the $s_i(t)$ signal at the output of the Q path.

- By using the equations for the s(t) signal, and by considering that the LO produces a local signal of frequency ω, the two outputs of the quadrature mixer can be deduced from the expressions:

$$I(t) + jQ(t) = \{s(t)[cos(\omega t) - j\,sin(\omega t)]\}|_{TJ}$$

$$= \{(Re[S_c(t)])e^{-j\omega t}\}|_{TJ} =$$

$$=\{½\,[S_c(t) + S^*_c\,(t)]\,e^{-j\omega t}\}|_{TJ}$$

$$= \{½\,[s_r\,(t) + js_i\,(t)] + ½\,[s_r(t) - js_i(t)]e^{-j2\omega t}\}|_{TJ}$$

$$= ½\,[s_r(t) + j\,s_i\,(t)] \quad (1)$$

- It can be noticed that if the frequencies are perfectly synchronized the two paths are perfectly isolated;
- If a perfect synchronization does not exist (between the local oscillation and the received carrier oscillation) a reciprocal coupling appears between the two paths; that means that a component depending on $s_r(t)$ is obtained on the Q path over $s_i(t)$ and the other way round.

- A similar effect will appear if the two local oscillations which attack the mixer are not shifted with exactly 90^{o} or if the amplitude gain for the two paths is not identical.

- In order to give an example, we will consider the simple but very frequent case when the local oscillator generates a signal with a small phase difference, φ, compared to the carrier of the received signal, *s(t)*.

- Because the offset φ is small, with a computation similar with the one from (1) and by Taylor series expansion the altered output signal can be written as:

$$\tilde{I} + j\tilde{Q} = \frac{1}{2} S_c(t) e^{-j(\omega t + \varphi)} \approx$$

$$\approx [\frac{1}{2} s_r(t) + j \frac{1}{2} s_i(t)](1 - j\varphi) \approx$$

$$\approx [I(t) + \frac{\varphi}{2} s_i(t)] + j[Q(t) - \frac{\varphi}{2} s_r(t)]$$

- It can be noticed that a coupling appears between the channels, dependent on φ because of the rotation of the axes (real and imaginary) with φ;

- This effect can be compensated by the inverse rotation of the axes, a posteriori, or by an offset introduced at the LO based on a known training sequence.

- In the general case when the local signals, in phase and quadrature, have arbitrary phase deviations θ and φ, and the I and Q channels have arbitrary amplitude gain deviations, ε and δ relative to the nominal values, the altered output signal can be written:

$$\tilde{I} + j\tilde{Q} = \{\frac{1}{2}[\mathrm{S_c}(\mathrm{t}) + \mathrm{S_c^*}(\mathrm{t})] \times$$

$$\times[(1+\varepsilon)\cos(\omega\mathrm{t}+\theta) - \mathrm{j}\,(1+\delta)\sin(\omega\mathrm{t}+\varphi)]\}\,|_{\mathrm{TJ}}$$

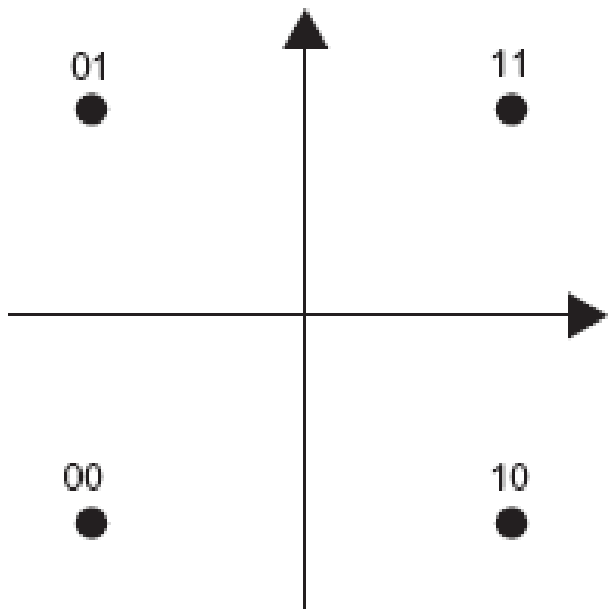
- Making a order one Taylor series expansion it results:

$$\tilde{I} + j\tilde{Q} \approx I(t) + jQ(t) + e(t)$$

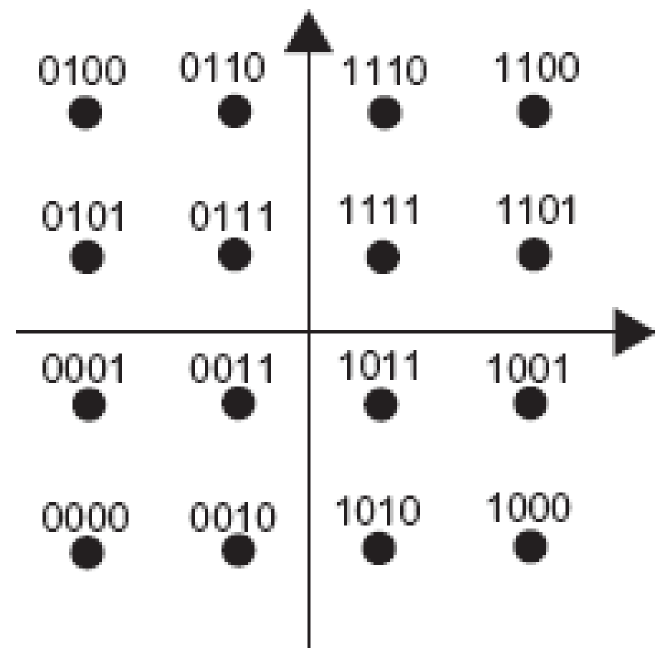
$$e(t) = \frac{1}{2}[\varepsilon\, s_r(\mathrm{t}) + \theta\, s_i(\mathrm{t})] + \mathrm{j}\frac{1}{2}[\delta\, s_i(\mathrm{t}) - \varphi\, s_r(\mathrm{t})]$$

- Without detailing, from the previous equation it is obvious that any deviation from the ideal balanced values introduce an error signal, dependent on a parameter $\lambda = max\{|\varepsilon|, |\delta|, |\theta|, |\varphi|\}$, *so it generates a noise dependent on* λ^2 .

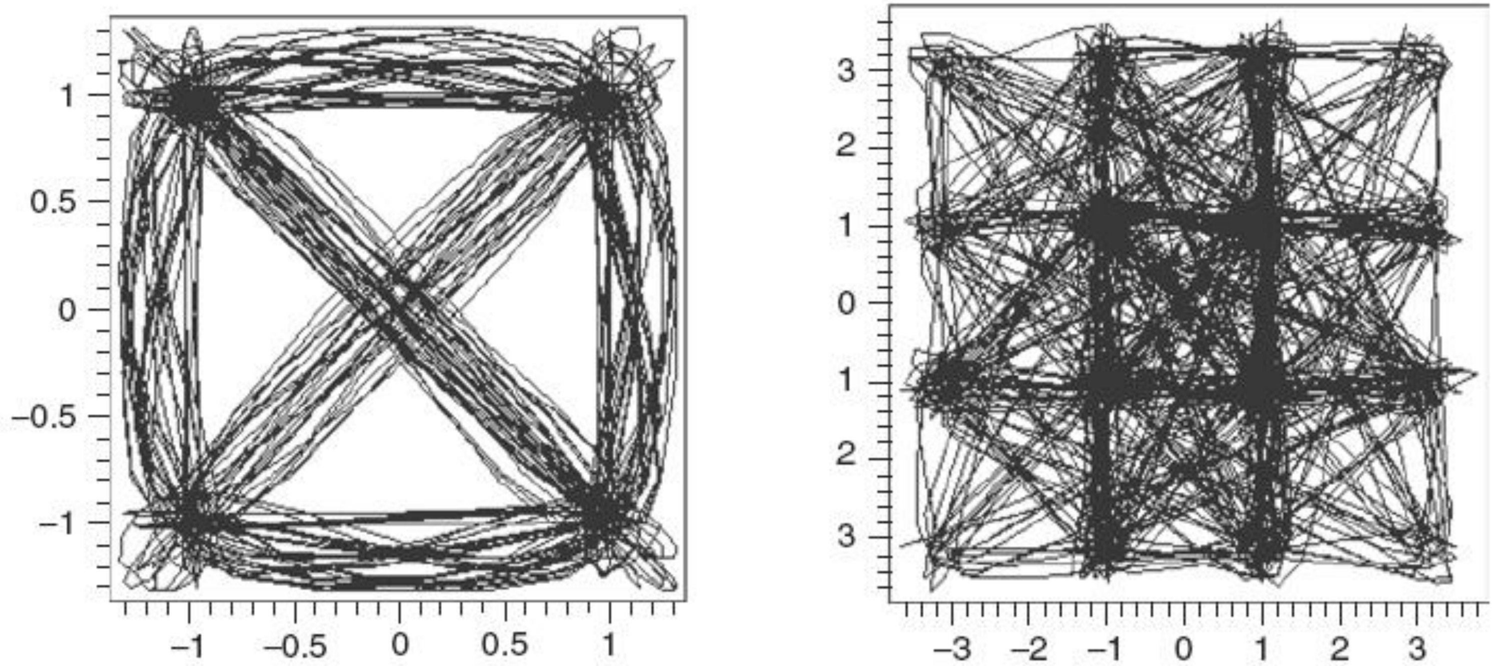
- We remind that in these conditions, the representation by using the signal constellation associated with the modulation scheme is useful.



(b) QPSK



(c) 16-QAM



(a) (b)

- During the detection process, the nearest symbol is chosen when performing the sampling, as it has the highest probability of being transmitted.

- From here resulted the conclusions regarding the effects of the transmission on the different modulation schemes.