



# Linearization of a Transmitter using an IC Digital/Analog Cartesian Feedback in CMOS 65nm for Advanced Communication Standards

Nicolas Delaunay

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présentée à  
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Ecole Doctorale des Sciences Physiques et de l'Ingénieur

par **Nicolas Delaunay**  
POUR OBTENIR LE GRADE DE  
DOCTEUR  
SPÉCIALITÉ : ÉLECTRONIQUE

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**Linearization of a Transmitter using an IC Digital/Analog Cartesian  
Feedback in CMOS 65nm for Advanced Communication Standards**

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# Remerciements

*A faire plus tard*

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# Introduction

Since the first generation of mobile phones, a lot of functions, standards and tools have been integrated on handsets. Twenty years ago, consumers could use their phone only to call and to send messages. Nowadays, internet access, radio, cameras, games and music are included and available as options for each personal mobile phone.

All of these new services make the cost of production for a cellular phone more expensive. Despite that, industry has to find a solution to maintain their products the most attractive as possible including the large range of integrated functions.

Thus, compromises have to be made in order to integrate increasing numbers of chips on a mobile phone. In fact, new standards impose specifications that make the design challenging in terms of linearity, the output power, or even the power injected on the adjacent channel. These specifications make the transmitter less efficient, or less linear. A lower efficiency will increase the power consumption of the chip, and a lower linearity will lead to a lower data rate.

In the context of interaction with other standards, the aim of this thesis is to design and implement a chipset able to improve the linearity of a transmitter for third generation mobile phones, using both digital and analog technologies. For this purpose, the study will focus on the improvement of the linearity, keeping the consumption and the die area of the circuit as small as possible. We will prove that linearization on an integrated circuit is possible with almost the same consumption and die area occupation compared to a classic transmitter. Most importantly, even if a large literature about linearization techniques can be found, only a few show the entire integration in a single technology, as we will show.

The first chapter will present the different architectures used for a transmitter and various linearization techniques with their advantages and drawbacks. Some metrics are also presented in order to evaluate these architectures. The goal of this part is to choose a linearization technique associated to a transmitter in order to fit with our application and constraints.



The second chapter will explain the complete system, digital and analog parts, with theoretical studies. We will start by detailing the constraints and precautions that must be taken into account by the designer to study the instability and the noise generated by the transmitter. We will describe how two algorithms back signal corrections. Then, in the last part we will show system level simulations of the Cartesian feedback using, first, an ideal power amplifier (PA) of ADS, second, a PA in BiCMOS technology, and third and finally, a PA in CMOS technology that will be used for the final integrated circuit.

Finally, the third and last chapter shows the digital synthesis in CMOS technology of the two algorithms previously mentioned, considering all steps, from the VHDL code until the layout of the digital part. We will describe and simulate each analog building block of the Cartesian Feedback, and some of them will have their performance measured.

Each chapter will be working towards the goal of this study, demonstrated in the final part: to make an integrated circuit, with its complete solution and simulations.

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# Linearization of a Transmitter

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Wireless communication standards are continually becoming more sophisticated because of the increasing data rate. It remains a delicate operation to transmit data at a high rate without creating distortions through the antenna with a transmitter. That is why the choice of the transmitter architecture is very important for the quality of the communication, a choice that is directly linked to the modulation or the standard used. In the case of the W-CDMA standard, a modulation with non-constant envelope is employed, the HPSK (Hyper Phase Shift Keying) modulation.

It generates a signal varying in phase and amplitude. It is used to reduce the peak-to-average ratio of the signal avoiding the zero crossings and the zero degree phase transitions. As a consequence, special care has to be taken to avoid any distortions on the main signal, which creates unwanted information on the adjacent channel reducing the data rate of the communication.

In this chapter, different transmitter architectures are studied in order to choose the one which will fit better with the integration of radio-frequency components for our application, and also, that will reduce most of the die area and consumption of the whole solution. To improve performance of this transmitter and keep the minimum consumption and die area at the same time, a linearization technique has been chosen among many techniques that we will see below.

## 1.1 Mobile Communication Standards

Cellular communications have evolved for thirty years in terms of the modulation employed. The first generation used an analog modulation whereas the second generation used a digital modulation. For the next generations of standards, the modulation became more complex with the transmission of images or data. Indeed, from the first generation in Japan (with the standard called DoCoMo) to the third generation of cellular technology, wireless standards have had to handle internet access and other services, such as video streaming, making more complex the modulation employed.

The widespread use of mobile phones all over the world requires good management of the spectrum to avoid interferences. With added economic issues, the establishment of new standards is essential in defining the specifications for manufacturers and operators.

	NMT [1]	GSM/ DCS [2]	GPRS/ EDGE[3]	UMTS [4]	HSDPA/ HSUPA[4]	HSPA+ [5]	LTE- Advanced [6]
Generation	1G	2G	2.5/2.75G	3G	3.5/3.75G or 3G+/3G++	Evolution 3G++	4G
Approximate Years	1981	1993	2001/03	2004	2007/08	2008/09	2009/10
Frequency bands (Hz)	0.45G- 0.9G	0.9/1.8G	0.9G	1.92G/ 1.98G	1.92G/ 1.98G	1.92G/ 1.98G	2.57G / 2.62G
Carriers Spacing (Hz)	12.5/ 25k	200k	200k	5M	5M	5M	15k (sub- carrier)
Access Methodology	-	F/TDMA	F/TDMA	W- CDMA	W-CDMA	MIMO CDMA	OFDMA/ SC-FDMA
Modulation	FM	GMSK	8PSK	HPSK/ QPSK	QPSK/ 16QAM	QPSK/ 16QAM/ 64QAM	QPSK/ 16QAM/ 64QAM
Data Rates (bps)	-	270k	812.5k	0.8-2M	14M	28M	100M

Table 1.1: Standard evolution for mobile phone communications

In Table 1.1, a non-exhaustive list of main standards is given from the first generation of mobile until the fourth generation in Europe. This comparison table shows main features that are more relevant for our study (frequency bands, modulation used, carrier spacing, access methodology).

### From 1G to 4G:

In the 1970s/1980s, the first analog wireless terminals started to appear in cars and in the business world. These mobiles had almost the size of a suitcase and did not cover a lot of territory.

This generation did not guarantee the confidentiality of communications and became saturated using a modulation technique similar to the FM radio. The development of this first generation was limited by the analog modulation used, short spectral resources, low battery life and cost of mobile terminals.

Thus, the second generation became digital in order to increase the capacity and the data rate of wireless communications. This generation insured a better sound quality and confidentiality and reduced the size of terminals. At the beginning of the 1990s, the GSM standard for wireless communication (Global System for Mobile Communication) was adopted in Europe. Since this

date, the GSM operates around the world except for Japan, South America and North America (which have another standard).

In 2001, an important evolution of the GSM Standard was the GPRS (General Packet Radio Service, more commonly called 2.5G). Concretely, the GPRS standard is an improvement of the GSM; the voice information remained on the GSM network, whereas data used the GPRS standard for a lighter internet access, as an e-mail without an attachment. The GPRS standard allowed a more comfortable connection with the WAP (internet access); however users of mobile phones were not convinced regarding the price and a limited internet connection.

Right after the GPRS, the GSM standard was developed towards the EDGE standard (Enhanced Data rate for GSM Evolution, also called 2.75G). Compared to the GPRS standard, the EDGE is able to send and receive data faster between a mobile phone and a base station. It can be considered as a complement of the future standard UMTS (Universal Mobile Telecommunication System), or a direct competitor of the 3G for a less expensive cost.

The UMTS standard exploits a new protocol of communication, called W-CDMA (Wide-Code Division Multiple Access), and new frequencies between 1900 and 2200MHz. This standard made it possible to reach high data rates for multimedia applications. Moreover, the standard made it possible to send parallel data with a higher data rate, comparing to the GSM and the GPRS. It is not a radical change compared to the second generation but it changed significantly the use of a mobile phone: high speed wireless internet, videophone and video streaming in the mobile. In theory, the UMTS can reach 2Mbps from a fixed location and 384kbps from a moving point.

After many delays of the development of the 3G, its deployment has finally started in Europe at 2004. The UMTS/W-CDMA standard has been selected in Europe by the 3GPP Association (Third Generation Partnership Project) grouping main telecommunications companies. It required each interested operator to purchase a license issued by its state, and, in parallel, to develop the infrastructure for the GSM/GPRS network. In France in 2004, according to an analysis of users of the 2G by the Telecommunications Regulatory Authority (*Autorité de Régulation des Télécommunications - ART*) more than 70% of French people had a mobile phone.

The worldwide industry of telecommunication is currently developing the fourth generation of mobile phones called LTE-Advanced (Long Term Evolution). This new generation will probably

use the OFDM technology (Orthogonal Frequency Division Multiplexing) reaching data rates up to 300Mbps. In order to prepare this generation, specialists developed the evolution of UMTS, called HSDPA (High Speed Downlink Package Access) and HSUPA (High Speed Uplink Package Access). The evolution of the UMTS offered a data rate of 14Mbps and the deployment in Europe began in 2006.

## 1.2 Transmitter Architectures

Transmitters are mainly composed of converters, filters, mixers, local oscillators and a power amplifier. Their goal is to modulate and up-convert the baseband signal into a higher frequency. The resulting signal is then strengthened by the power amplifier before driving the antenna [7]. A band-pass filter (or a multiplexer) is used to filter the desired channel and send it through the antenna. Nevertheless, it is complicated to determine the architecture of the transmitter without taking into account the architecture of the receiver. Indeed, unwanted frequencies and noise that are generated by transmitter are injected into the receiver. Ideally, the transmitter and the receiver architectures should be considered at the same time, in order to optimize performance, die area and consumption of the RF circuit.

In this section, main architectures are evaluated. Among the most classic ones, we can find architectures based on a simple or on a double mixing frequency, as homodyne or heterodyne, or otherwise, transmitters based on a PLL (Phase-Locked Loop).

### 1.2.1 Heterodyne transmitter architecture

The heterodyne transmitter architecture is described in Figure 1.1.

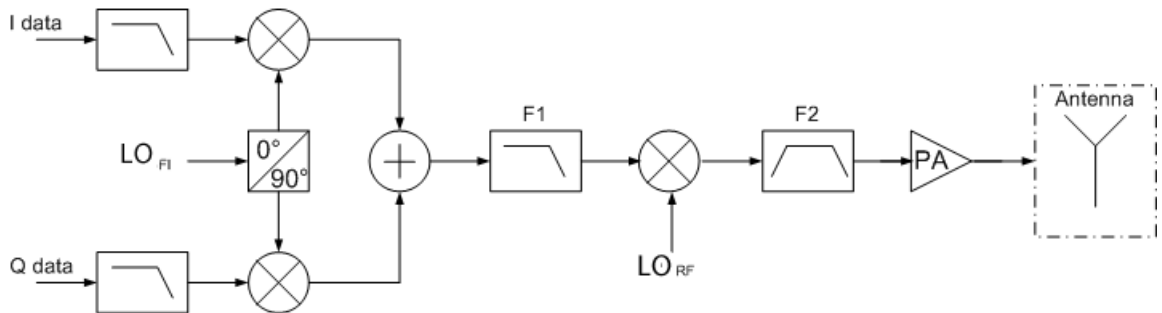


Figure 1.1: Heterodyne transmitter architecture

Analog baseband signals I and Q are up-converted to an intermediate frequency ( $f_{IF}$ ) by two mixers in quadrature. The signal is then sent to a low-pass filter (F1) which deletes harmonics

coming from the  $f_{IF}$  frequency. In order to avoid any other distortion coming from the baseband domain, it remains important to realize this operation before to transpose the baseband signal to another frequency ( $f_{IF} + f_{RF}$ ) given by the mixer M2 ( $LO_{RF}$ ).

The mixer M2 will produce two inter-modulations tones at  $f_{IF} + f_{RF}$  and  $f_{IF} - f_{RF}$  frequencies. As the frequency  $f_{IF} - f_{RF}$  is a non-desired tone, a band-pass filter (F2) has to be used to remove it. The resulting signal is finally amplified and transmitted through the antenna.

This architecture works with constant and non-constant envelop modulations and the modulation of signals is made at a low  $f_{IF}$  frequency (which is about a few MHz). This architecture, thanks to its two stages of frequency conversion, obtains good results in terms of sensitivity, selectivity and linearity according to the modulation used; however, it uses two local oscillators which increase the power consumption of the transmitter.

Moreover, as the  $f_{IF} - f_{RF}$  tone has the same power as the  $f_{IF} + f_{RF}$  tone, the band-pass filter of the second stage of conversion has to reject the signal of  $\sim 40\text{dB}$  at least. This filter has to work at high frequency, which complicates the result of getting a good quality factor. In general, an external filter is used as a SAW filter (Surface Acoustic Wave). This filter adds area and cost that made this kind of architecture very expensive.

### 1.2.2 Direct conversion transmitter architecture

The direct conversion transmitter architecture, also called homodyne or Zero-IF (Zero-Intermediate Frequency, meaning without intermediate frequency), uses a single stage of frequency conversion as shown in Figure 1.2.

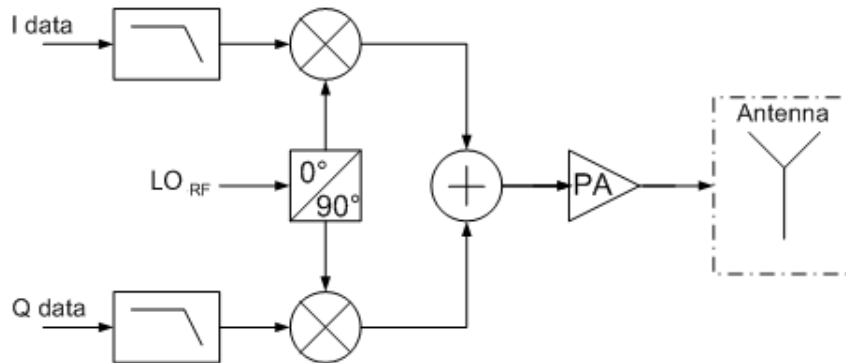


Figure 1.2: Direct conversion transmitter architecture

Like the heterodyne architecture, analog baseband signals are modulated and up-converted to

RF frequency by mixers on quadrature, a PLL with a LO working at the RF frequency. In this case, the frequency of the resulting signal is directly transposed to the frequency of the standard.

This architecture takes less area than the heterodyne architecture because it uses fewer components reducing thereby the consumption. Moreover, the image rejection on the main channel is realized dynamically by summing the output of mixers (Cf. Equation 1.1).

$$(\cos \omega_{BB}t * \cos \omega_{RF}t) + (\sin \omega_{BB}t * \sin \omega_{RF}t) = \cos(\omega_{RF}t - \omega_{BB}t) \quad (1.1)$$

As a consequence, we do not need the band-pass filter of the intermediate stage and release some constraints of the external filter. (Trigonometric functions removed automatically the  $\omega_{RF}t + \omega_{BB}t$  term).

Such architecture is compatible with a complex I/Q modulation as HPSK, but has some drawbacks concerning the mismatch in phase and amplitude between I and Q. The major issue is to realize a good quadrature generation at the RF frequency. If a weak quadrature takes place, we will have a bad rejection of the image frequency [8].

One of the major drawbacks for the direct conversion transmitter is the pulling phenomenon [9] [10]. In other words, when two oscillating components are close to each other, they are oscillating at the same frequency due to the coupling phenomena. In the case of the Zero-IF frequency, the local oscillator and the PA works at frequencies very close to each other and can be coupled by the substrate. The RF signal at the output of the amplifier is much greater than the output power of the LO, so the working frequency of the LO will be shifted into the RF frequency. The isolation needed remains hard to quantify without considering the receiver associated with the Zero-IF transmitter [11].

Moreover, the direct conversion is sensitive to DC parasites. Indeed, parasites are mixed with the main signal for the receiver. In W-CDMA the baseband signal is spread around 3.84MHz. A high pass filter with a cut-off frequency lower or equal to 10 kHz can removed all of these DC parasites keeping the integrity of the W-CDMA signal.

### 1.2.3 Low-IF transmitter architecture

The pulling phenomenon disappears if we use two LO frequencies (frequencies of  $LO_{\omega_1}$  and  $LO_{\omega_2}$ ). The Figure 1.3 shows the architecture of the low-IF frequency.



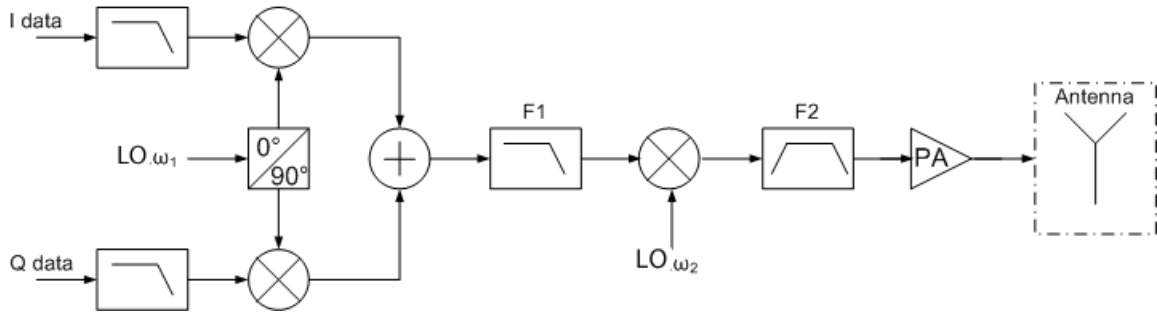


Figure 1.3: Low-IF transmitter architecture

The low-IF transmitter architecture is based on the same principle of a heterodyne transmitter, except that the  $LO_{\omega_1}$  frequency is obtained by mixing and filtering two lower frequencies. By consequence, the first mixing stage in the IF frequency can be realized on the digital circuit. This method has the same advantages as the direct conversion transmitter, except that this transmitter is not sensitive for the pulling if the IF frequency is not too low.

Furthermore, an incorrect choice of LO frequencies produces harmonics at the output. A low-pass filter which follows the VCO (Voltage-Controlled Oscillator) has to be very selective in order to keep a good quality of the output signal.

#### 1.2.4 Offset-PLL transmitter architecture

This technique was developed to answer the strong constraints of the GSM standard for the noise in the reception band. The architecture of the Offset-PLL transmitter [7] is presented in Figure 1.4.

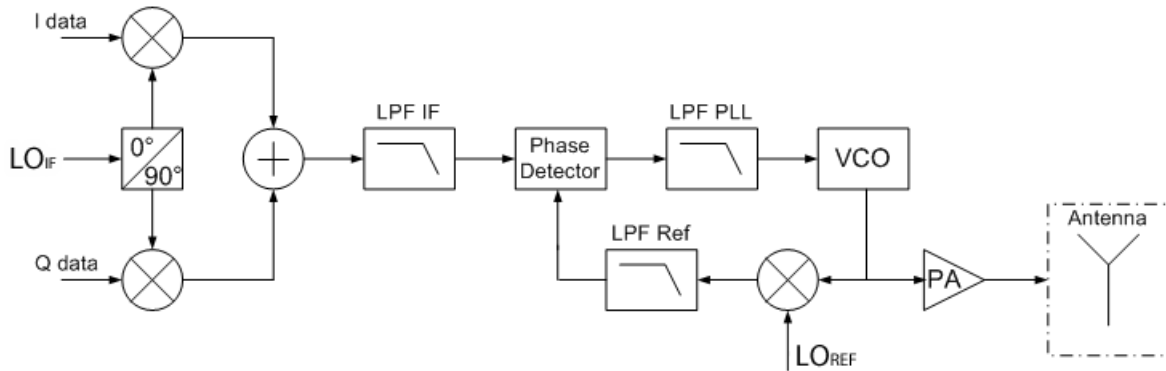


Figure 1.4: Offset-PLL transmitter architecture

The offset-PLL transmitter [12] has the objective to mix the signal from the local oscillator

with the output signal from the VCO (Voltage Control Oscillator). The resulting signal is then injected into a phase comparator to generate an error signal which, after being filtered, controls the VCO. This will work at the RF frequency. A current offset generator is needed to obtain a fast locking of the loop. Indeed, the output current of the phase comparator does not allow the locking of the loop in a reasonable time. The time-locked can be improved by 26% [13]. Nonetheless, without adjustment of the current, the loop will become unstable.

The offset-PLL transmitter has a good IC integration because it does not use mixers to convert the baseband frequency to the RF frequency, neither external filter. Moreover, it does not consume a lot of power, which makes a very attractive solution for the cost reduction. On the other side, this solution works with constant envelop modulation and does not fit with our modulation which is the HPSK. This architecture also has the drawback of suffering from the pulling, so a good isolation between the VCO and the PA has to be realized.

### 1.2.5 Summary of transmitter architectures

For the wireless communication in the context of mobile devices with the W-CDMA standard, it is very important to get a good integration level on the Silicon area and a low consumption of our chip, as both reduce the cost of production. Furthermore, the architecture has to work with constant and non-constant envelope modulation considering the modulation of the W-CDMA used.

The Table 1.2 makes a comparison of these last transmitters according to these criteria.

Architecture	Integration	Consumption	Modulation
Heterodyne	Low	Medium	All
Homodyne	Good	Low	All
Low-IF	Medium	Medium	All
Offset-PLL	Good	Low	Constant envelope

Table 1.2: Comparison of transmitter architectures

Despite serious issues imposed by the Zero-IF architecture explained at the topic 1.2.2 and according to the previous table, this transmitter is the most appropriated to reduce consumption and die area for W-CDMA communications. These criteria remain our first objective and it is for this reason that the Cartesian Feedback has been implemented on a Zero-IF transmitter.

### 1.3 Nonlinear Behavior of a RF Transmitter

To understand an RF transmitter, we must have the previous knowledge that it is composed of RF transistors.

In 1948[14], W. Shockley describes for the first time the principle of the bipolar junction transistor. However, it is only at the end of the 1950s that the first transistor able to amplify signals up to 1GHz was developed.

The term Radio Frequency (RF) refers to the electromagnetic waves with frequencies ranging from Gigahertz or higher. Until the 1980s, most applications for RF transistors were used only by the military. Since then, the appearance of wireless communication systems (Wi-Fi, Bluetooth and WLAN) for the costumers has completely changed the sales for mobile phones.

The conception of RF transmitters starts with the choice of the type of transistor depending of the targeted application. For our study, which is to improve the linearity of the transmitter, we opted for the characterization of a transmitter (e.g. power amplifier), using nonlinear model of transistor. This part shows the nonlinear behavior of a transistor (or components using transistors) and the metrics to measure it.

Furthermore, each of the RF transmitter architectures can be evaluated according to its integration on the Silicon area, as well as the power consumption or the modulation used. Therefore, in order to validate the RF transmitter some specifications are given by the standard communication.

#### 1.3.1 AM/AM and AM/PM conversion

The power amplifier is a very important component of a transmitter and defines the consumption and the linearity of the whole system. The behavior of the PA acts also directly on the receiver and can increase the bit error rate (BER) of the transceiver.

An ideal power amplifier has a linear response. In other words, the output signal is proportional to the input signal.

$$V_{out}(t) = a_0 + a_1.V_{in}(t) \quad (1.2)$$

Where  $a_1$  is the voltage gain of the power amplifier.

For a low input voltage, the functioning of the power amplifier can be assimilated to an ideal power amplifier. However, for strong input voltage, the output signals suffer from distortions due to amplitude (AM/AM) and phase (AM/PM) conversions. Current sources are the first elements able to cause amplitude variation of the output signal according to the input signal. Moreover, nonlinear behavior gives an additional shifting of the output signal, which is also linked with the input signal from the power amplifier.

The Figure 1.5 shows these conversions:

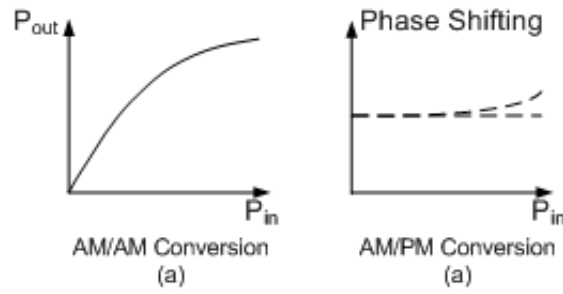


Figure 1.5: AM/AM and AM/PM conversion

Let's consider an input sinusoidal signal with an amplitude and phase modulation

$$V_{in} = A(t) \cdot \cos[\omega_0 t + \phi(t)] \quad (1.3)$$

At the output of the power amplifier, we obtained

$$V_{out} = A'(t) \cdot \cos[\omega_0 t + \phi(t) + \phi'(t)] \quad (1.4)$$

With

$$A'(t) = F_{NL}[A(t)] \quad (1.5)$$

And

$$\phi'(t) = G_{NL}[A(t)] \quad (1.6)$$

Where  $F_{NL}$  and  $G_{NL}$  are nonlinear functions.

The deflection of AM/AM and AM/PM curves, for strong input level power, permits to understand the nonlinear behavior of the power amplifier.

However the efficiency  $\eta$  is defined as the ratio between the output power (in Watts) and the power supply consumed (in Watts). It will be maximal when the amplifier is used with high level signal (Cf. Figure 1.6). Thus, a nonlinear function and generation of spurious tones are produced when we will try to improve the saturation power or the efficiency of the power amplifier.

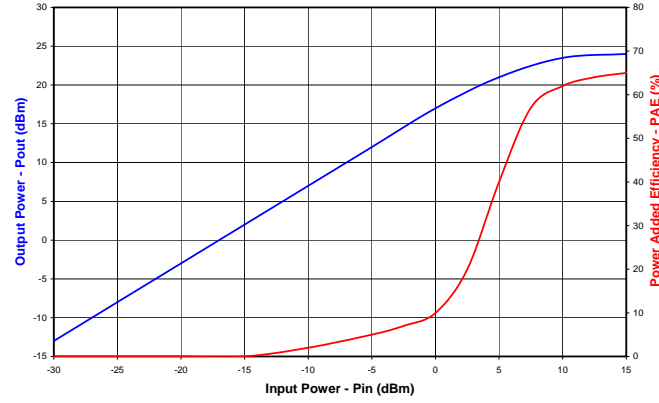


Figure 1.6: Output power and efficiency of a PA versus input power

### 1.3.2 Carrier to Intermodulation ratio (C/In)

The analysis with two carriers is a simple method to evaluate the linearity of an amplifier and to show, at the same time, amplitude and phase distortions of the amplifier. This method focuses on the variation of the envelope's signal, using the amplifier over its entire dynamic range. It is a very severe test for the amplifier because the envelope ranges from zero to a maximum level, as shown in Figure 1.7(a).

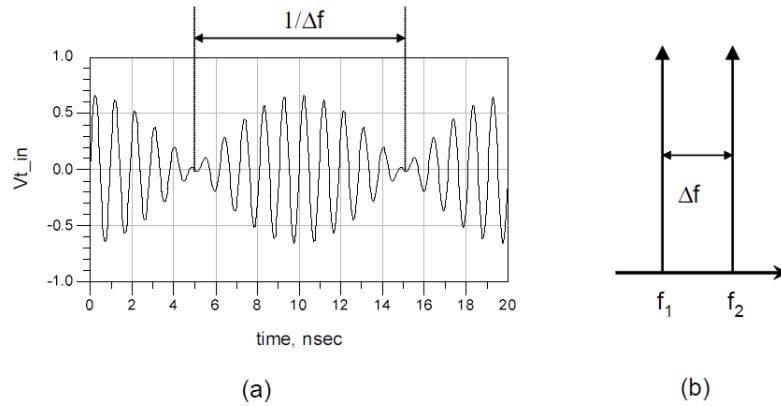


Figure 1.7: Signal with two carriers in timed (a) and frequency (b) domains

Two signals at the frequency  $f_1$  and  $f_2$ , with the same amplitude and non-modulated with  $\Delta f = f_2 - f_1$ , are applied at the input of the amplifier. On the time domain, a signal with two carriers is represented in Figure 1.7 (a). The equivalent spectrum is given in Figure 1.7 (b).

As shown in the previous part, such signal going through the nonlinear amplifier will generate intermodulation products.

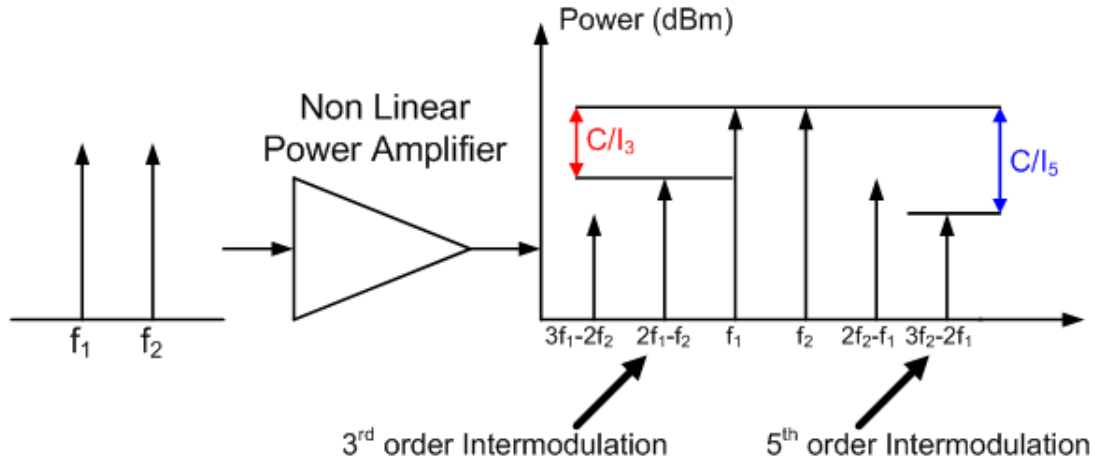


Figure 1.8: Carrier to Intermodulation ratio

The 3<sup>rd</sup> order ratio C/I (Carrier to Intermodulation ratio) is defined as being the difference between the power in dBm, of the carrier  $f_1$  or  $f_2$ , and the intermodulation tone respectively at  $2.f_1 - f_2$  or  $2.f_2 - f_1$  frequency (Cf. Figure 1.8). In general, we define the  $n^{th}$  order ratio C/I as the difference between the power of the carrier  $f_1$  or  $f_2$  and the intermodulation tone at  $p.f_1 + q.f_2$  with  $p + q = n$ .

### 1.3.3 Adjacent Channel Power Ratio (ACPR)

The ACPR measures the power level at the output of the power amplifier, between the adjacent channel and the main channel. It uses a modulated signal of the same type as the one that must be amplified for standard chosen. It is defined as the ratio between the average powers (in Watts) contained in a frequency band  $B_1$  and  $B_2$  to an offset  $f_0$  of the carrier frequency  $f_c$  of the channel, and the average power (in Watts) contained in the frequency band  $B_0$  around the carrier frequency  $f_c$  of the standard. It leads to the following expression:

The concept of this measure is shown in Figure 1.9. Its expression is the following:

$$ACPR_{dB} = \left( \frac{2 \cdot \int_{B_0} P(f) \cdot df}{\int_{B_1} P(f) \cdot df + \int_{B_2} P(f) \cdot df} \right) \quad (1.7)$$

The output power according to the frequency is shown in Figure 1.9.

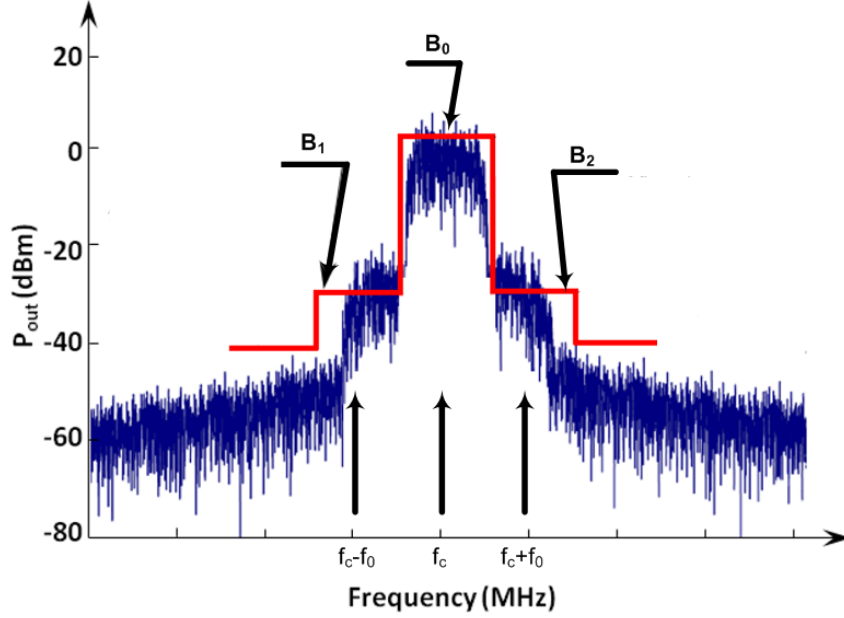


Figure 1.9: Adjacent Channel Power Ratio spectrum

Frequency bands  $B_0$ ,  $B_1$  and  $B_2$  and the offset  $f_0$  depend on the chosen standard of communication.

### 1.3.4 Noise Power Ratio (NPR)

The NPR is a power measurement of distortions on the channel caused by nonlinearities of the power amplifier. It can be evaluated by removing a frequency band of the input signal of the channel, and measuring the output level of distortion in the frequency space left free (SLF).

The NPR is defined as the ratio between the maximum power of the signal on the channel and the power of distortions, where both are expressed in Watts. Figure 1.10 illustrates the measure of the NPR on the modulated signal after the power amplifier.

$$NPR(dB) = 10 * \log \left( \frac{\int_{SLF} P(f) \cdot df}{\int_{Carriers} P(f) \cdot df} \cdot \frac{BW_{Carriers}}{BW_{SLF}} \right) \quad (1.8)$$

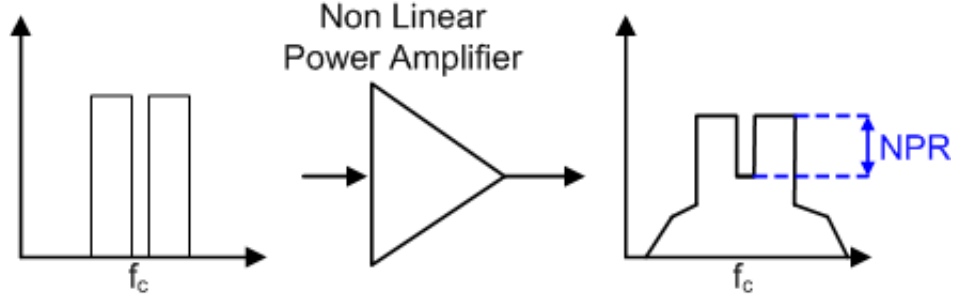


Figure 1.10: Noise Power Ratio

### 1.3.5 Error Vector Magnitude (EVM)

The EVM represents the difference between the effective demodulated signal after the transmission or amplification, and the signal as it should be if the system was perfectly linear. This measurement gives the information about the error of the phase and amplitude of the received signal [15]. Figure 1.11 represents the EVM definition for a modulated signal by two signals in quadrature I and Q.

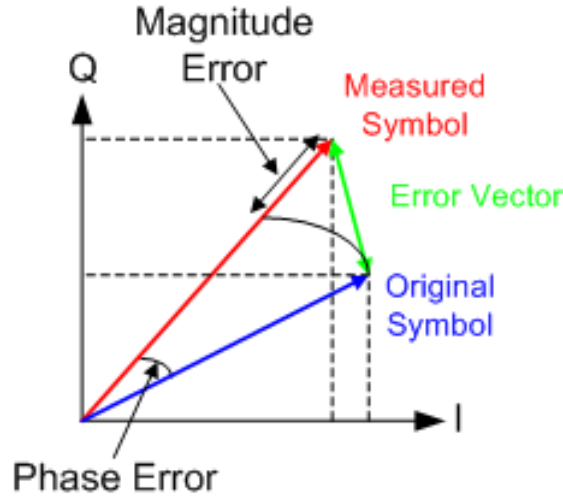


Figure 1.11: Error Vector Magnitude

The EVM represents the module of the error vector as followed:

$$\text{Magnitude Error} = \sqrt{I_{\text{Measured}}^2 + Q_{\text{Measured}}^2} - \sqrt{I_{\text{Original}}^2 + Q_{\text{Original}}^2} \quad (1.9)$$

$$\text{Phase Error} = \arctan\left(\frac{Q_{\text{measured}}}{I_{\text{Measured}}}\right) - \arctan\left(\frac{Q_{\text{Original}}}{I_{\text{Original}}}\right) \quad (1.10)$$



$$EVM = \sqrt{(I_{Measured} - I_{Original})^2 + (Q_{Measured} - Q_{Original})^2} \quad (1.11)$$

This criterion is generally used for signal processing to characterize the whole system of a couple modulator/demodulator. It helps to quantify the effect of the additive noise on the distortion of the transmitted signal's constellation.

## 1.4 Linearization Techniques

The HPSK modulation has a good spectral efficiency, but the power efficiency remains poor with linear power amplifiers. For mobile applications, a low efficiency increases the consumption of the transmitter, decreasing the life time of mobile phone battery.

To obtain, for the same device, a good linearity and spectral efficiency, we use linearization techniques such as EER (Envelope Elimination Restoration), Feed-forward, Polar loop or Cartesian Feedback. On the other hand, a good linearity for components like an RF modulator or a power amplifier can be obtained with overconsumption. A linearization technique offers the possibility to use nonlinear components in order to reduce the consumption of the transmitter.

We have two categories of linearization techniques: those which improve the linearity of the transmitter and those which improve the efficiency of the whole system.

### 1.4.1 Tradeoffs between linearity and efficiency

The challenge for the design of a power amplifier can be simplified by a system of two variables in which we have to reach an output power imposed by the wireless communication standard while optimizing its efficiency. This must be done with respect to the given linearity constraints. Nevertheless, a compromise has to be made between the maximum output power and the efficiency of the power amplifier. According to specifications given by the standard, we must distinguish among many topologies of power amplifiers.

#### 1.4.1.1 Classes of power amplifier

In literature, classes of power amplifiers can be divided into two categories [16]:

- The first class is defined by the polarization given to the transistor and is called the “linear class”. They are designated by: Classes A, B, C, and AB.
- The other class is the commutation class because transistors work like switches. It is called the “highly efficient class”. They are designated by: Classes D, E, F and S.

*Linear class of power amplifier:*

For power amplifiers taken from the linear class, the transistor behavior can be assimilated as a current source. The output power is proportional to the input power. The gain is constant on a wide band of power. This kind of class is well known and is required when a complex modulation is used (non-constant envelop, i.e. HPSK) in order to preserve the information contained into the envelope.

*Commutation class of power amplifier:*

For power amplifiers of the commutation class, the transistor works like a switch with the output non-proportional to the input. This kind of power amplifiers is really efficient for constant envelope modulation (like GMSK used for the GSM standard). The consumption of the circuit is considerably reduced and the efficiency remains very high. Indeed, in theory, when the level of the input voltage of the transistor is high, the current is equal to zero and reciprocally. This advantage increases the time-to-talk of batteries implanted into a mobile device.

The most important parameters able to define performances of a power amplifier are the efficiency and the nonlinearities. The values of these parameters are linked to the classes of the chosen power amplifier. So it is important to define their characteristics to understand how to correct or improve them.

**1.4.1.2 Efficiency**

The efficiency of a power amplifier can be evaluated as a drain efficiency (if the transistor is a field-effect transistor, otherwise, it is a collector efficiency for a bipolar junction transistor), a power added efficiency (PAE), or a whole efficiency.

The drain efficiency is defined as the ratio between the fundamental output power ( $P_{out}$ ) and the power delivered by the continuous alimentation ( $P_{DC}$ ).

$$\eta_{DE} = \frac{P_{out}}{P_{DC}} \quad (1.12)$$

The power added efficiency of a power amplifier is defined by the Equation 1.13. It represents the ability of a transistor to transform the energy delivered by continuous alimentation in a microwave energy. This efficiency can be calculated as the ratio between the added sine wave power of the power amplifier and the supply voltage applied to it. The added sine wave power

is defined as the difference between the input power ( $P_{in}$ ) and the output power ( $P_{out}$ ) of the power amplifier.

$$PAE = \frac{P_{out} - P_{in}}{P_{DC}} \quad (1.13)$$

Finally, the whole efficiency represents the ratio between the energy available at the output of the power amplifier and the energy provided to the amplifier in order to insure the output power.

This efficiency is defined by the Equation:

$$\eta = \frac{P_{out}}{P_{DC} + P_{in}} \quad (1.14)$$

#### 1.4.1.3 Nonlinearity

Power amplifiers suffer from nonlinearities, or distortions, when a certain level of the input signal is applied.

The amplitude distortions are divided in two categories:

- distortions making a compression effect on the power of the signal (compression point at 1dB);  
and
- distortions due to interferences between two signals which have frequencies close to the fundamental frequency.

The combination of the different frequencies creates intermodulation products. Distortions due to interferences have impacts only on the receiver. These two types of nonlinearity are described in the following section.

##### Compression Point at 1dB:

The sinusoidal signal at the input of the power amplifier is defined by the Equation 1.15

$$V_{in} = A \cdot \cos(\omega t) \quad (1.15)$$

Where  $A$  is the maximum amplitude of the input signal and  $\omega$  is the operating frequency.

Considering the output voltage as a temporal function of the input voltage, it is possible to express this voltage in polynomial form until the  $3^{rd}$  order nonlinearities term. The expression of the output voltage will be:

$$V_{out} = a_1 V_{in} + a_2 V_{in}^2 + a_3 V_{in}^3 \quad (1.16)$$

After the development of the Equation 1.16, the expression of the fundamental will be:

$$a_1 A \left[ 1 + \frac{3 a_3}{4 a_1} A^2 \right]$$

The coefficient  $a_1 A$  represents the linear gain of the system. If  $a_3 > 0$ , the power amplifier works on the linear region, but if  $a_3 < 0$ , the power amplifier suffers of the compression phenomenon at the fundamental frequency. The output power of the amplifier is often given at 1dB of compression, meaning that the gain is 1dB under the linear zone. The compression point is described in Figure 1.12.

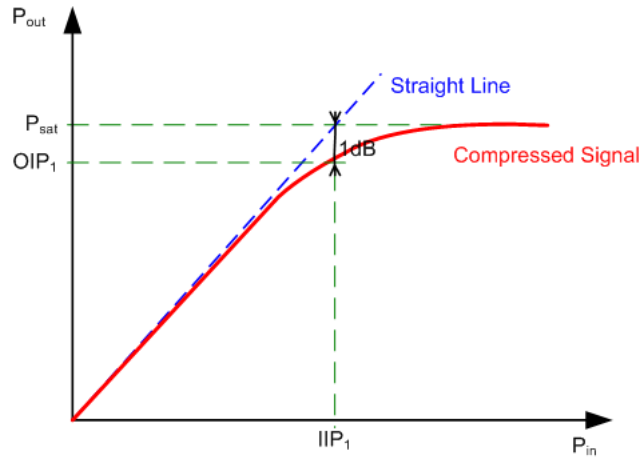


Figure 1.12: Definition of the compression point

Where OCP1 is the first order Output Compression Point and ICP1 is the first order Input Compression Point.

Evaluating the 1dB compression point is not enough to quantify the nonlinearity of a power amplifier used on a communication system. For this reason, we must also study the intermodulation products in order to evaluate distortions of the same function. The knowledge of the 1dB compression point is important to determine the back-off level to take into account when complex modulation is brought into play (as the OFDM modulation).

#### Intermodulation products:

Intermodulation products represent a combination of fundamental frequencies. Spurious signals

can disturb the wireless communication system if they are within the bandwidth of the system.

In order to characterize intermodulation products, we must consider the input signal ( $V_{in}$ ) - Cf. Equation 1.17 - composed of two sinusoidal signals with the same amplitude  $A$  but with different frequencies ( $f_1$  and  $f_2$ ).

$$V_{in} = V [\cos(\omega_1 t) + \cos(\omega_2 t)] \quad (1.17)$$

The output signal of the power amplifier is then defined in Equation 1.18.

$$\begin{aligned} V_{out} = & a_1 V [\cos(\omega_1 t) + \cos(\omega_2 t)] + a_2 V^2 [\cos(\omega_1 t) + \cos(\omega_2 t)]^2 \\ & + a_3 V [\cos(\omega_1 t) + \cos(\omega_2 t)]^3 \end{aligned} \quad (1.18)$$

The output signal  $V_{out}$  has term in  $n\omega_1 \pm m\omega_2$ ,  $n$  and  $m$  being integers. The order of the intermodulation products is defined by the relation:

$$i = |n| + |m| \quad (1.19)$$

Second order intermodulation products are thus represented by tones at frequencies  $f_1 + f_2$  and  $f_2 - f_1$ , and these of the third order by frequencies  $2f_1 + f_2$ ,  $2f_1 - f_2$ ,  $2f_2 + f_1$ , and  $2f_2 - f_1$ .

Some of intermodulation products with an odd order ( $i$  being an odd number) are, as represented in Figure 1.13, within the operating bandwidth of the system, perturbing the transmitted signal.

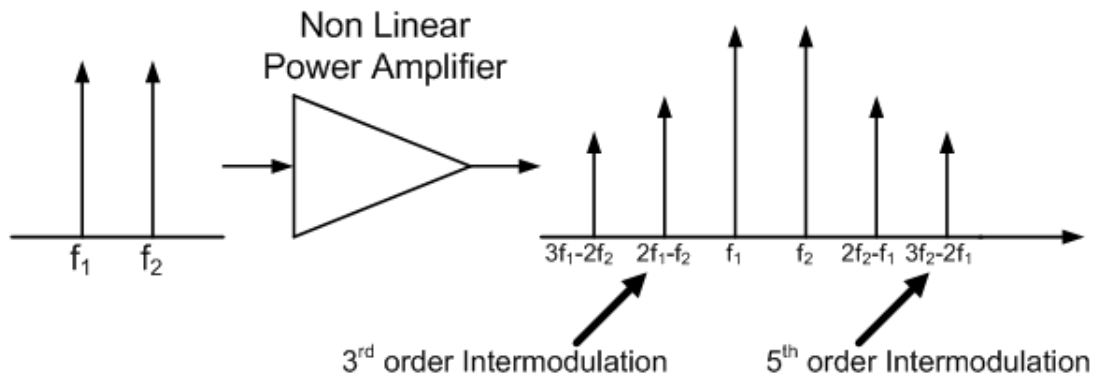


Figure 1.13: Input and output spectrum of a nonlinear amplifier after compression

The third order intermodulation product is represented in Figure 1.14.

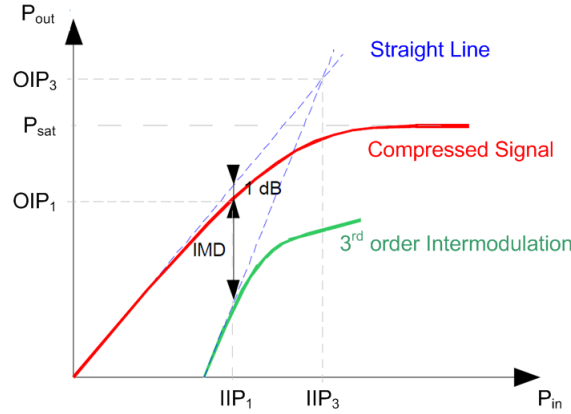


Figure 1.14: Definition of the third order intermodulation product

In this graphic, we can deduce the third order interception point (IP3) at the input (IIP3) and at the output (OIP3). This point can be determined by calculating the intersection between the linear line of the fundamental power and the line of the third order products intermodulation.

The Figure 1.14 represents also intermodulation products, called IMD (InterModulation Distortion) and is expressed in dBc. This distortion is defined by the Equation 1.20 knowing that the power of the fundamental is defined at the 1dB compression point:

$$IMD_3 = P_{2\omega_2 - \omega_1} - P_{\omega_1} (dBc) \quad (1.20)$$

The process is similar to determine  $i^{th}$  order intermodulation distortions. This time, it can be calculated by the Equation 1.21.

$$IMD_i = P_{m\omega_2 - n\omega_1} - P_{\omega_1} \quad (1.21)$$

Where  $m$ ,  $n$ , and  $i$  are integers numbers and linked by this expression:  $i = m + n$ .

Third and fifth intermodulation products are, in most of cases, within the bandwidth of the system and have sufficient power in order to have an influence on the output signal of the power amplifier.

The knowledge of the third and fifth intermodulation products is valuable for communication system using a complex modulation, as the HPSK for the UMTS Standard.

### 1.4.2 Power back-off

The easiest way to improve the linearity of a power amplifier is to use it in a Class A operation, reducing the power level needed by the amplifier (back-off) until that the desired level of linearity is reached. For a Class A power amplifier,  $3^{rd}$  order intermodulation products increase according to a factor of 3 (Cf. Figure 1.14), and those of  $5^{th}$  order by a factor 5. A back-off of the input power can significantly reduce the intermodulation products, and especially if these intermodulation products have a high order.

An alternative of using a Class A power amplifier is employing a Class AB power amplifier, where the polarization level is typically about 10 to 50% of the Class A level. The linearity in this case depends largely on the characteristics of the RF component technology. Some technologies, such as the LDMOS, have also very good linearity performances for a Class AB [17].

The drawback of these very simple methods is the low efficiency and the need to have a power amplifier oversized for the application. Therefore, other methods are applied to increase the linearity without deteriorate the whole efficiency of the system.

### 1.4.3 Efficiency improvement

#### 1.4.3.1 Envelope Tracking (ET)

Envelope tracking has an objective to keep the linearity of the power amplifier, and at the same time, increase the efficiency. The principle of this topology is to modify the biasing according to the envelope of the input signal. The envelope tracking system imposes the envelope signal at the drain of the power amplifier, and the PA is consequently working in a high efficiency region.

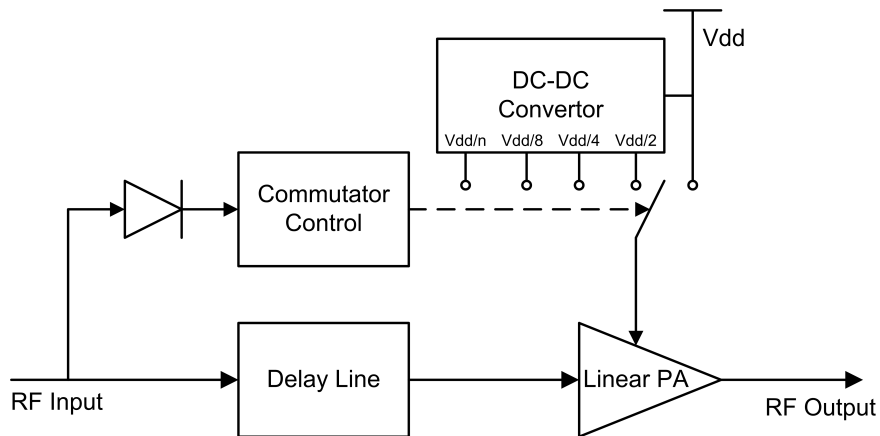


Figure 1.15: Discrete Envelope Tracking

The control of the polarization can be seen as a discrete (Cf. Figure 1.15) or continuous (Cf. Figure 1.16) DC-DC converter. In the first case when the envelope of the signal is greater or lower than a certain value, the polarization voltage can take a value previously fixed. In the other case, the ET works on the biasing, which makes a continuous adjustment of the amplitude according to the envelope signal in real time.

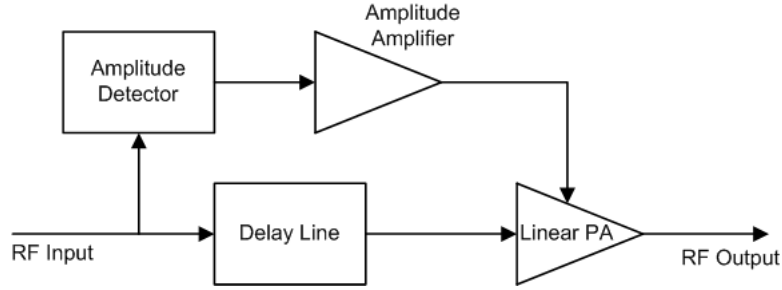


Figure 1.16: Continuous Envelope Tracking

The efficiency reached by the discrete or continuous ET is greater than the efficiency of a power amplifier with a fixed polarization. The current consumption of DC-DC converters and others control components reduce performances in the case of low signal envelope. It is a discrete control, easier to implement and taking less area than a converter can be used, but it does not reach an improvement like continuous correction. Moreover, the generation of different DC signals remains complex for an integrated solution. Nevertheless, dynamic biasing offers full use of the linearity performance of a Class A, while improving significantly the maximum efficiency, which is an attractive solution.

#### 1.4.3.2 Envelope Elimination and Restoration (EER)

The Envelope Elimination and Restoration (EER) in Figure 1.17 has been developed by Leonard E. Khan in 1952 [18]. The goal of this technique is to amplify a non-constant envelope modulation with a nonlinear power amplifier. The input modulated signal at an intermediate frequency (IF) is decomposed in phase and amplitude through an envelope detector and a limiter. The constant envelope signal in amplitude is up-converted to the RF frequency in the mixer. The resulting signal is then amplified by the nonlinear power amplifier in saturation mode. The envelope of the signal is sent back modulating the voltage supply of the PA by the phase of the input signal.



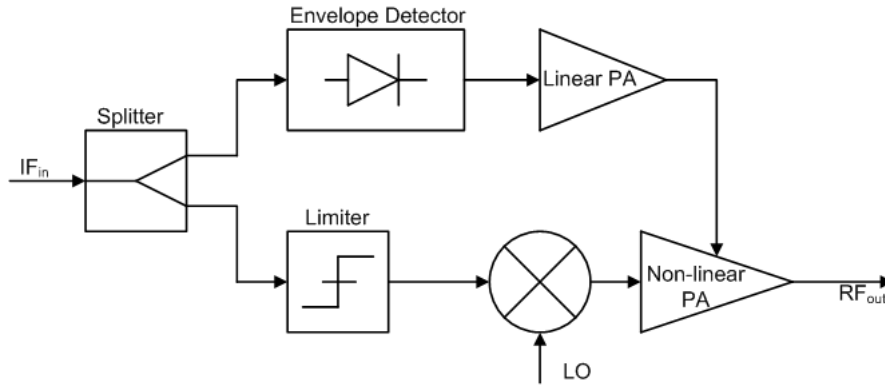


Figure 1.17: Envelope Elimination and Restoration

The advantage of the EER is to improve the linearity and the efficiency of the power amplifier. This technique remains a complex solution to implement for two reasons. First, it requires controlling perfectly the envelope power amplifier linearity. Secondly, the phase alignment between envelope and phase signals has to be guaranteed. Linearization performance of the EER is linked to the bandwidth of the envelope, but also the phase shifting between the envelope modulation and the phase modulation. It is therefore crucial to control the variation of delay in both paths of the architecture, as well in temperature and for the process used for a good recombination of the output signal. In recent application of the EER [19], the DSP (Digital Signal Processor) splits the signal in phase and amplitude. This solution does not fit with the criteria of a system integrated on a single technology without using external devices.

#### 1.4.3.3 Linear amplification with Nonlinear Components(LINC) / Combined Analogue Locked Loop Universal Modulator (CALLUM)

This technique uses the property of the decomposition of a non-constant envelope signal into two constant envelope signals, thereby splitting the input signal into two distinct parts. Each of these parts is separately amplified on the nonlinear region and they permit to get a high efficiency and a high output power before a recombination at the output (Cf. Figure 1.18(b)) [20, 21].

The LINC technique may use a DSP to generate two baseband signals modulated in phase and with constant amplitude envelope (Cf. Figure 1.18(a))

Thus an input signal  $s(t) = a(t) \cos[2\pi \cdot f_c \cdot t + \phi(t)]$

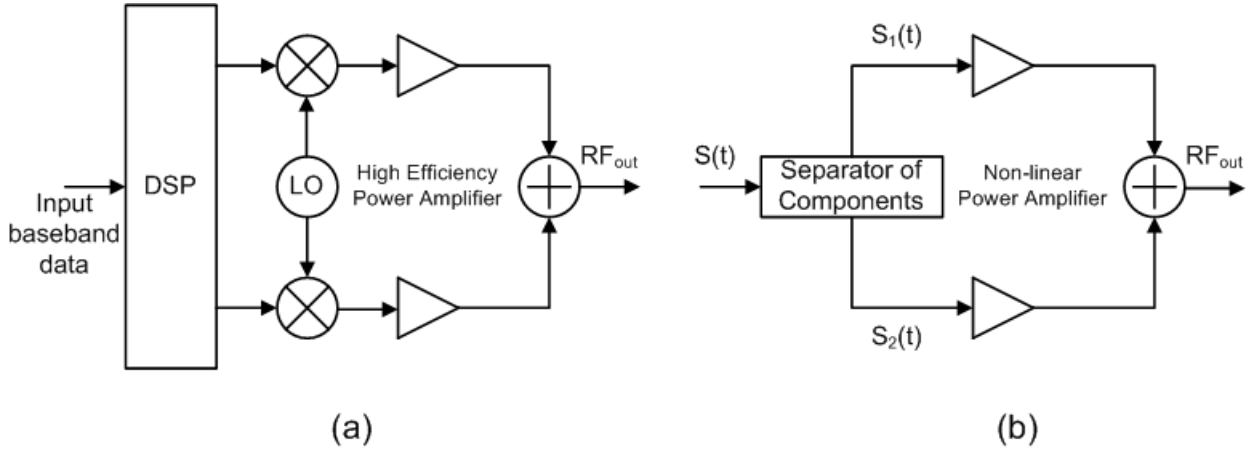


Figure 1.18: Principle of LINC technique

This signal can be described by two signals  $s_1(t)$  and  $s_2(t)$  with a constant amplitude such as:

$$s_1(t) = a_{max} \cdot \cos[2\pi \cdot f_c \cdot t + \phi(t) + \alpha(t)] \quad (1.22)$$

And

$$s_2(t) = a_{max} \cdot \cos[2\pi \cdot f_c \cdot t + \phi(t) - \alpha(t)] \quad (1.23)$$

With

$$2 \cdot s(t) = s_1(t) + s_2(t) \quad (1.24)$$

And

$$\alpha(t) = \cos^{-1} \left( \frac{a(t)}{a_{max}} \right) \quad (1.25)$$

These two signals are separately amplified by high efficiency power amplifiers, and then recombined to make an amplified copy of the input signal. Any nonlinear behavior appearing in the signal will be found in phase opposition and equal amplitude before the summation. Each nonlinear part will be canceled after summing. This method can use amplifiers in high efficiency commutation to approach an efficiency of 100%.

The two most important difficulties of the method are: to adjust both paths in phase and amplitude in order to obtain the elimination of any intermodulation; and to work with identical power amplifiers [22, 23, 24].

The CALLUM is a feedback technique with the objective to cancel problems in phase and gain of the LINC [25].

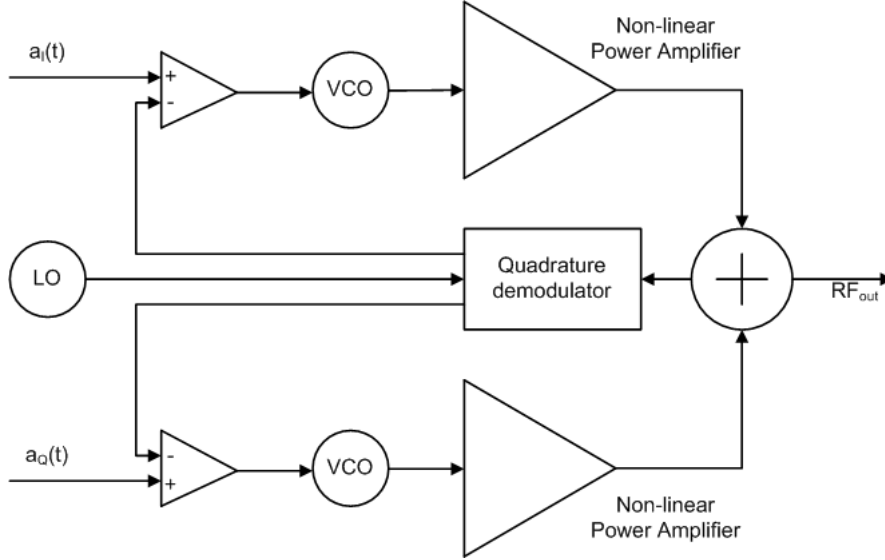


Figure 1.19: CALLUM technique

A Cartesian decomposition on the feedback path is used with the CALLUM method. The output signal is demodulated into two quadrature signals. These signals are compared in baseband domain with the two input signals and used to generate error signals. The main issue of this method is the stability of the system. For this reason, this method can be used only for straight bandwidth modulation.

#### 1.4.3.4 Doherty technique

Performances of the power amplifier are mainly determined by the load impedance at the output of the transistor [26]. The Doherty technique allows the load impedance to be modified according to the input power thanks to the combination of two power amplifiers: a main and an auxiliary; and also with two lines  $\lambda/4$  which  $Z_C$  and  $Z_{Cin}$  represent the impedance of these lines. (Cf. Figure 1.20)

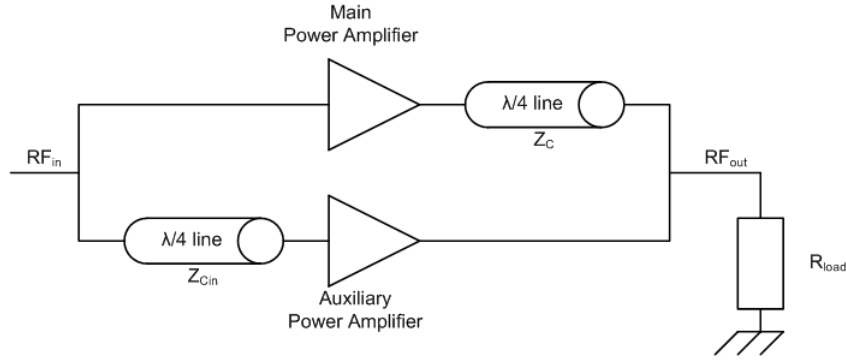


Figure 1.20: Doherty technique

The  $\lambda/4$  line placed at the output of the main power amplifier realizes the impedance transformation needed to maintain the efficiency over an important variation of the output power. This line will, however, add a phase shift of  $90^\circ$ .

The second  $\lambda/4$  line allows compensation of this phase shift between signals from power amplifiers. Thereby, the in-phase recombination of the output signal is possible.

The functioning of the Doherty technique is as follows:

- For low input power, only the main power amplifier delivers an output signal. The auxiliary power amplifier is turned off.
- For an intermediate power level, the auxiliary power amplifier begins to deliver output power, while the main power amplifier starts to saturate. The fact that the auxiliary power amplifier is turned-on leads to the modification of the load impedance presented by the main amplifier.
- For strong power level, both amplifiers are saturated.

Assuming the same size of the two power amplifiers and the characteristic impedance of the  $\lambda/4$  line  $Z_C = 2.R_{load}$ , load impedances of power amplifiers will vary according to the input power as follows:

- For a low power level, the auxiliary power amplifier is clamped. So its drain current is then equal to zero. In this case, the impedance at the output of the main power amplifier is equal to  $4.R_{load}$ , and the one that is seen at the auxiliary is equal to an open circuit.

- For a high power level, both power amplifiers are saturated and its drain currents are equal. Both amplifiers will have the same load impedance equal to  $2.R_{load}$ .

The transition power  $P_\alpha$  is the output power from which the auxiliary amplifier begins to operate. The Table 1.3 summarizes the Doherty amplifier principle:

Power level	Main power amplifier mode	Auxiliary power amplifier Mode
Power $< P_\alpha$	Current sources controlled	Turned off
$P_\alpha < \text{Power} < P_{Max}$	Saturated	Current sources controlled
Power = $P_{Max}$	Saturated	Saturated

Table 1.3: Doherty amplifier principle

The Figure 1.21 represents the evolution of the efficiency, in cases where the main and the auxiliary power amplifier have the same size, versus the ratio between the load voltage and the voltage supply ( $V_{load}/V_{dd}$ ).

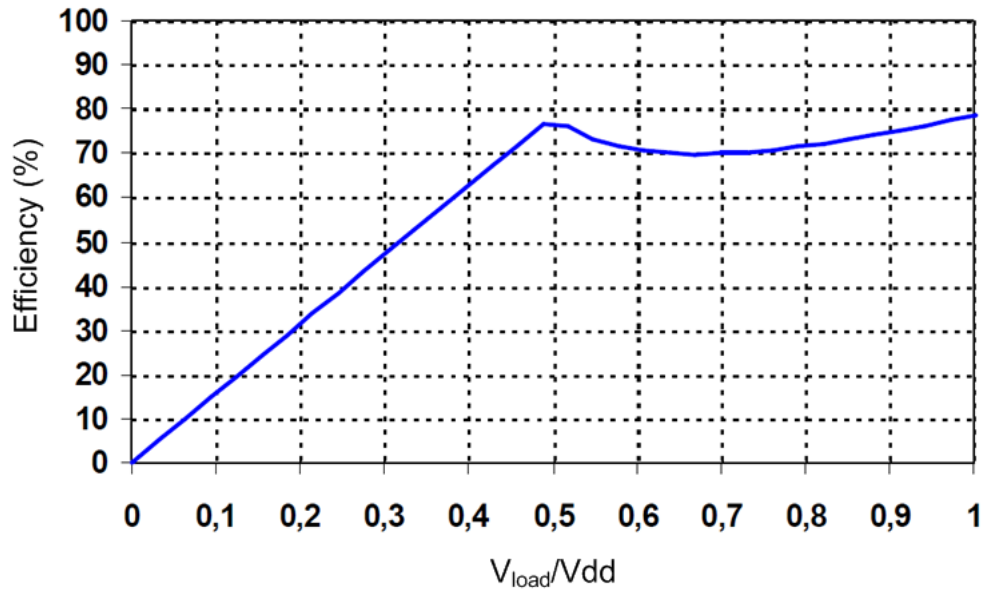


Figure 1.21: Evolution of the efficiency for a two stage Doherty amplifier

The efficiency is evolving linearly to the point corresponding to the transition power  $P_\alpha$ . Then, it decreases when the auxiliary power amplifier starts to operate. After that, it increases again

until it reaches the maximum efficiency at the maximum power.

The size of the auxiliary power amplifier depends on the curve variation of the efficiency according to the power level. It imposes different shapes and moves the transition point. The choice of this stage size is related to the application and the shape of the signal to amplify.

This technique can also be realized with several auxiliary power amplifiers. As a consequence, different levels of transition can be obtained, being able to achieve and to maintain a strong efficiency over a wide dynamic output power [27].

The operation class of the auxiliary power amplifier has a strong influence on the power added efficiency of the whole amplifier [27]. An adjustment of the polarization of the auxiliary amplifier is then needed for each output power. A possible solution is to use a variable variation of the auxiliary power amplifier according to the power level available at the input of the amplifier [28].

#### 1.4.4 Linearity improvement

##### 1.4.4.1 Predistortion / Postdistortion

Predistortion technique deals with the RF signal before the power amplifier (Cf. Figure 1.22), and inversely to the postdistortion technique, which deals with the signal after the power amplifier. They can be used for wideband systems and for multi-carriers as amplifiers used for satellite communication.

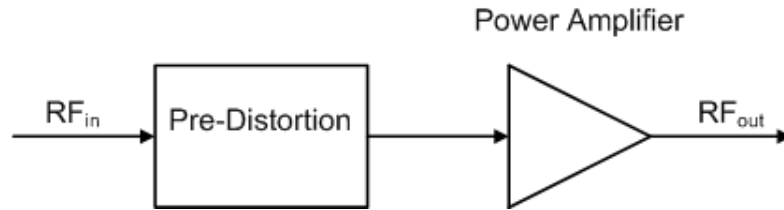


Figure 1.22: Predistortion technique

This approach involves the creation of a complementary function of distortion according to those of the power amplifier. The resulting system from the serialization of the predistortion (or postdistortion) will present few or no distortion at the output.

In the ideal case, the predistortion component generates a distorted signal which is the opposite of those created by the power amplifier (Cf. Figure 1.23). When this generator and the amplifier are cascaded together, distortion is cancelled [29].

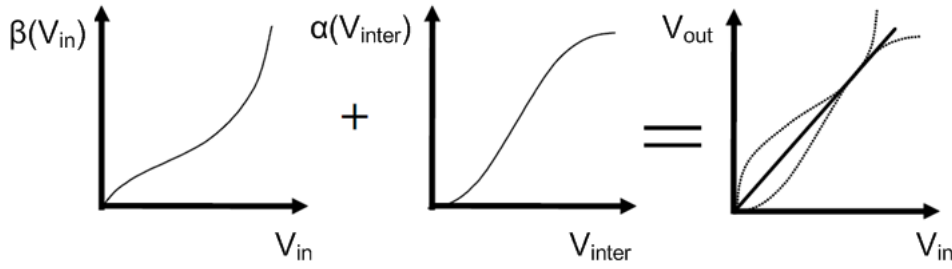


Figure 1.23: Signal before and after predistortion technique

The difficulty of this technique is the realization of a circuit able to provide a precise inverse distortion in phase and amplitude of the distortions given by the power amplifier. Meanwhile, due to the relative simplicity of implementation, this technique is interesting for power amplifiers used for mobile system when the size, the cost and a low consumption are the most important parameters.

Different systems are based on a predistortion of the input signal and can be divided into three categories:

- High Frequency (HF) Predistortion: the component of predistortion works at the frequency of the carrier
- Intermediate frequency (IF) Predistortion: the component of predistortion works at an intermediate frequency, allowing the design of a single predistortion for different carrier frequencies. Simply varying the frequency of the local oscillator for the element of predistortion can be used with multiple carriers.
- Baseband Predistortion: before the appearance of DSP, this approach did not give a lot of advantages compared to an IF predistortion. It now becomes a very powerful tool. Predistortion characteristics are saved on a DSP in order to correct the baseband signal.

There is no feedback on this technique as the system works as an open loop. This means that the distortion has to be known in advance and performances of this technique can be sensitive to variation due to the process variation, temperature, aging etc.

The improvement of the linearity obtained from the realistic case depends on many considerations, and more particularly on the behavior of the power amplifier. In general, the improvement of the linearity remains low compared to a Feedforward or a Cartesian Feedback. However, this

improvement is satisfying for a lot of applications such as satellite communication systems.

These techniques do not make a dynamic correction according to the changes of the power amplifier behavior. For this reason a feedback path is added to the predistortion technique in order to make a dynamic correction.

#### 1.4.4.2 Adaptive predistortion

The Figure 1.24 shows an example of adaptive predistortion: it is the logical evolution of the classical predistortion due to a more often utilization of DSP. The idea comes from the main drawbacks of the predistortion technique, an inability to correct dynamically distortions of the power amplifier. The system shown in Figure 1.24 [30] keeps a dynamic correction of the power amplifier. A predistortion table is computed based on the parameters of this model and from the comparison between the output signal from the PA and transmitted signal from the input of the whole system. This system is able to react to any variations by periodically updating the predistortion table.

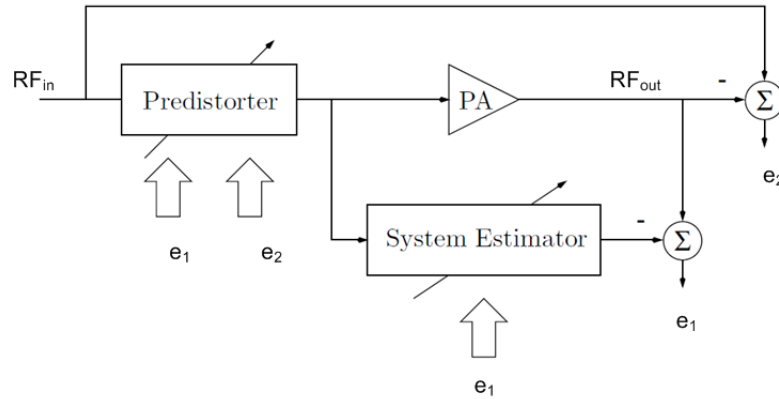


Figure 1.24: Example of an adaptive predistortion

The main drawback is the complexity of the adaptive part. Usually, this system requires high computing capacities, like a DSP. The power required for the DSP can exclude the adaptive predistortion for mobile applications.

#### 1.4.4.3 Feedforward

The popularity of this technique is due to its linearization potential of signals on multiple carriers and on a wide bandwidth application. The concept of this technique is simple but the implementation of it can be expensive. On the simplest configuration of this technique, two



paths can be distinguished (Cf. Figure 1.25): a main RF path and a cancellation path. A hybrid divider is able to split the input signal and send it into the both paths.

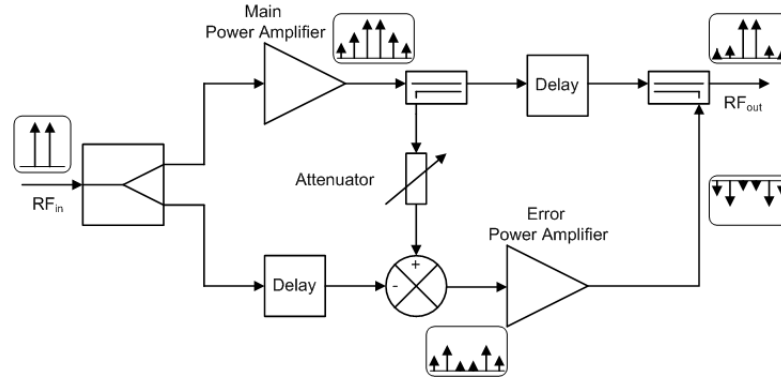


Figure 1.25: Feedforward technique

Part of the signal is delayed into a line in order to compensate the delay given by the power amplifier. The other part of the RF input power is amplified by a main power amplifier. Another part of the output signal is taken from a coupler, attenuated and sent into a hybrid combiner where it is compared with the signal coming from the delay line. With a variable attenuator to adapt the levels of RF signals, we recover only at the output of the hybrid combiner distortions generated by the power amplifier. These are then amplified in a highly linear Class A power amplifier. The signal coming from the error amplifier is ideally formed by the distortions created by the main power amplifier. The delay line of the main path is implemented to adjust the delay introduced by the second path; at the output of the last coupler, only the useful signal is conserved if distortions generated by the main amplifier are well compensated.

The most critical component in the Feedforward technique is the error amplifier. It does not help itself to generate intermodulation products on one hand and have an important gain on the other. So, the delay and the gain are parameters that have to be carefully adjusted in this technique for a significant reduction of the intermodulation, and controls have to be implemented to realize these functions.

The technique has proven its efficiency for wide frequency bands and with multicarrier signals like the CDMA or W-CDMA. Used in the base stations and for satellites systems, this solution is not selected for portable systems, mobile, or low weight, die area and consumption.



adapted for severe bandwidth modulation, regarding the polar representation of the signal, which involves an extension of the bandwidth.

#### 1.4.4.5 Cartesian Feedback

Unlike the polar loop, the Cartesian Feedback works with signals decomposed in the Cartesian plan on the I and Q axis.

The linearization technique using a Cartesian Feedback was discovered by D. C. Cox in 1975 [32], then by V. Petrovic in 1979 [31]. Theoretically, this technique can reach an improvement of 40 dB on the adjacent channels.

According to the Figure 1.27,  $I_{BB}$  and  $Q_{BB}$  baseband signals of the direct path are directly up-converted to the RF frequency. This frequency transposition is performed by quadrature mixers.

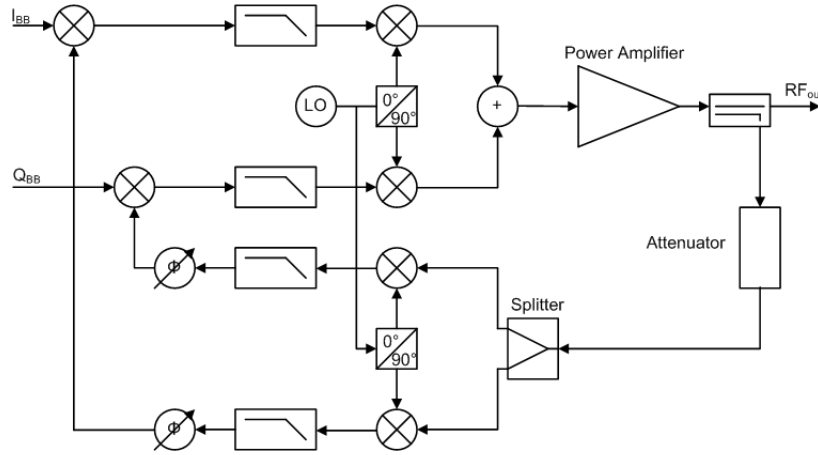


Figure 1.27: Cartesian Feedback technique

The RF signal is then amplified through a PA to reach the output power requested. The output signal of the power amplifier is attenuated and demodulated in quadrature (via down mixers) on the feedback path. Intermodulation products (at  $2 * \omega_{LO}$ ), generated by the down conversion, are removed by low pass filters. Demodulated and filtered signals of the feedback path have been shifted in phase due to components of the direct path. A phase correction has to be applied on the baseband signals to compensate phase distortions. These corrected signals are then subtracted from the original baseband signals resulting, error signals for I and Q baseband signals. Error signals carry information of the nonlinearities of the transmitter, and by loop effects the error signal will be reevaluate according to the response of the direct path. Before low

pass filters of the direct path, input signals are predistorted signals. The result signal after the power amplifier is a linearized signal due to the compensation effect between the distortion of the amplifier and the predistorted signal.

The main issue of this architecture is to maintain its stability. To insure this, phase correction can be used guaranteeing an enough phase margin.

Compared to previous techniques explained before, the Cartesian Feedback linearizes simultaneously the amplitude and the phase of the RF signals, using a single feedback path. This technique fits well with the Zero-IF architecture. It allows linearizing not only the PA, but also the RF modulator.

However the linearization works correctly only within the bandwidth of the Cartesian Feedback. The system can be optimized in order to increase the bandwidth of the Cartesian Feedback. It corresponds to the bandwidth of the useful signal with its adjacent channels.

With only few more components, the Cartesian Feedback remains relatively easy to implement. This technique is very effective because it can work with all types of modulation and offers good linearization performances.

Compared to other techniques presented, the Cartesian Feedback is the one which offers most assets in term of cost (die area) and the consumption of the whole solution.

## 1.5 Conclusion

As demonstrated in this chapter, standards used for mobiles have been getting more and more complex; adding functions, decreasing the size and increasing the battery life time.

This evolution has been made with the interest of suppliers and sellers to decrease not only the size, but the cost to produce it, without forgetting that the cellular phone had to be fashionable enough to increase the sales for all consumers, and not only businessmen. A mobile able to please either the buyer who searches for capacity and efficiency to make phone calls and to text, or the purchaser who is looking for an accessory able to reflect prosperity and modernity even if it is not so efficient.

The development of cellular communications has several architectures of a transmitter with their

pros and cons as announced on the Table 1.2. As our goal was to focus on the integration, the consumption and the nature of the modulation, the homodyne architecture was highlighted as it matches with these criteria. It is an architecture which uses simple and few analog components, resulting in an integration that does not need a lot of silicon area and, thereby, does not need a high consumption of energy. Also, this topology can work with constant and non-constant envelopes modulation, in other words, with any existent modulation on telecommunication standards nowadays.

Besides this, all transmitters in certain levels of output power will create interferences that must be linearized with a technique. Some linearization techniques will increase the efficiency of the transmitter and others will increase the linearity of it:

- Increasing the efficiency of the transmitter but without the improvement of the linearity, and a high or medium consumption and die area occupation: EER and LINC techniques; and
- Increasing the linearity of the transmitter but with different results regarding the flexibility, simplicity, consumption and die area occupation:

	Flexibility	Simplicity	Consumption	Die Area Occupation
Predistortion	No	Yes	Low	Low
Adaptive Predistortion	Yes	No	Medium	Low*
Feedforward	Yes	No	High	High
Polar Feedback	Yes	No	Medium	Medium
Cartesian Feedback	Yes	Yes	Low	Low

\*Needs an external component as a DSP.

Table 1.4: Comparison of linearization techniques

Please note that as mentioned in this chapter, our study considers only the most cited techniques in the literature, the ones above.

We opted for a linearization technique as we did to select the transmitter architecture: the one that fitted with our criteria and goal. Ergo, the Cartesian Feedback remains a technique with a lower consumption and takes the less die area. Moreover, this technique has as advantages to be flexible and simple to implement, as we will demonstrate in the following chapters.

# Analysis and Specifications of the Cartesian Feedback Loop

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Every mobile communication system has to be led by a normalized standard able to define a lot of characteristics as power, linearity and noise generated at the output of the RF transmitter or at the input of the receiver. Unfortunately the standard does not determine the architecture that the designer has to use on his chip and does not give any specification about all the building blocks included in a transmitter. Hence, theory and simulations on system levels have to be realized before the design of the circuit and be in consistence with the standard used, in our case the W-CDMA.

In order to get an optimal solution, minimizing the die area and the consumption of the circuit, the study must consider the whole architecture of the transmitter, identifying and resolving problems in the architectural design, from the digital process in baseband domain to the analog RF components. A model of the Cartesian Feedback will be explained in order to understand stability and noise issues capable of making the oscillation of the system.

From the perspective of the integration of our transmitter in CMOS 65nm technology, a mixed architecture has been developed [33] improving the linearity in real time. The digital part will help the system correct distortions on the signal whatever the result at the output of the transmitter is.

## 2.1 Digital/Analog Partitioning Architecture of the Transmitter

A Cartesian Feedback able to linearize a power amplifier with only an analog circuit was designed few years ago [34, 35, 36]. In order to reduce the consumption and increase the level of integration of the solution, a previous study with STMicroelectronics proposed to move some functions to the digital domain [37]. The operations of phase correction and subtraction can be implemented in VHDL using DSP (Digital Signal Processing) [38]. The phase corrector and the subtractor are components of the feedback path and have to be the most linear as possible. Realizing these two functions with analog circuitry resulted in increased power consumption and increased die area. Moreover, the main advantage of the mixed solution is to correct any

signal from the output of the power amplifier. The output signal of the power amplifier can change depending on aging or temperature, but if the output signal will change, the signal will be consequently different and a correction will have to be done. This will make an attractive solution with low cost and low power consumption, which has been the current trend in the past years for the industry.

A conversion stage is then added to the architecture of the Cartesian Feedback. This stage is composed by two digital-to-analog converters (DAC) for the direct path and two analog-to-digital converters (ADC) for the feedback path. The increased power consumption induced by these converters will be compensated by the improvement of the linearity in comparison with an analog solution. This architecture is presented on figure 2.1 [39].

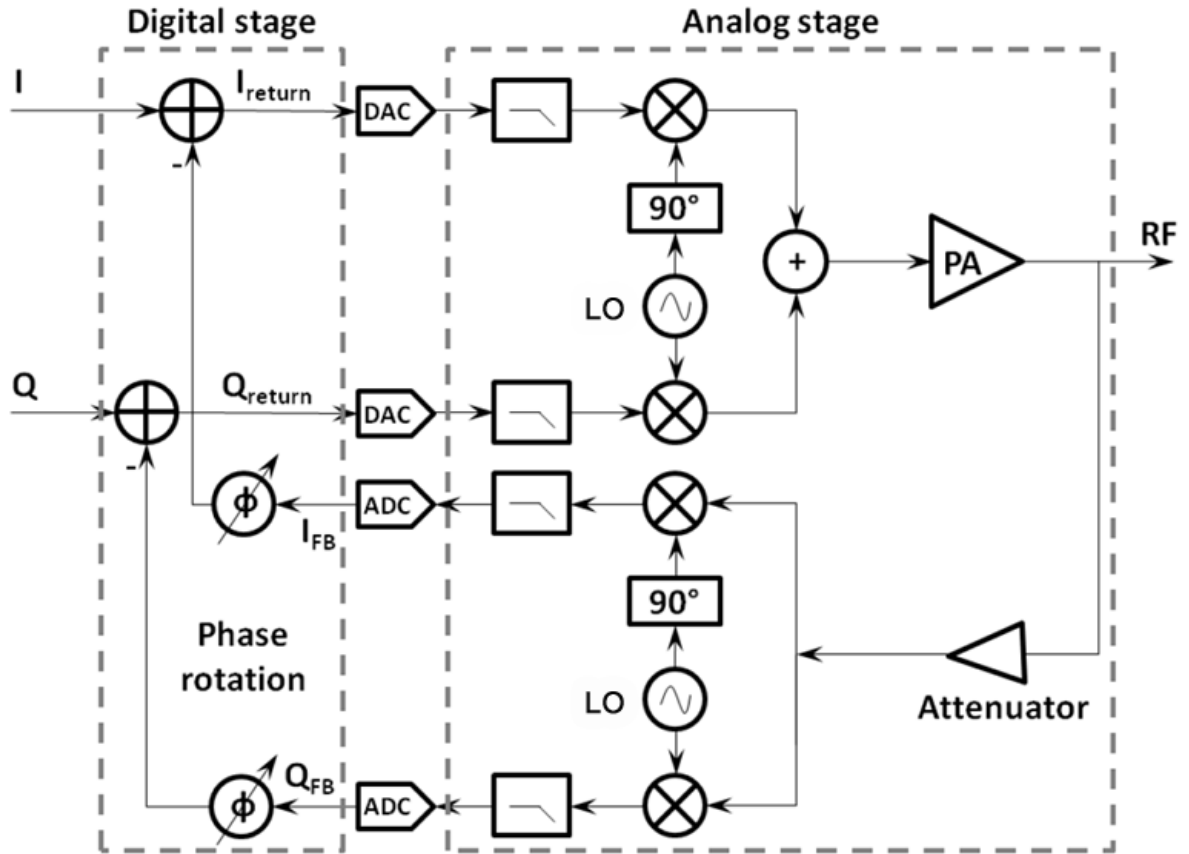


Figure 2.1: Mixed architecture of the Cartesian Feedback

In this architecture, low pass filters have two objectives. They have to filter frequency image of all harmonics from the sampling frequency, and remove the  $2f_{LO}$  harmonics introduced by the down conversion in the feedback path.



These harmonics make the second order term at the output of the RF modulator able to unbalance and deteriorate the image rejection of the modulated signal.

Low-pass filters of the feedback path remove the  $2.f_{LO}$  intermodulation products given by the RF demodulator, and avoid the folding spectrum during the conversion of the analog signal to the digital signal.

## 2.2 Mathematical Analysis of the Cartesian Feedback Loop

The problem of the stability appears with the feedback path. If the system is unstable the given correction will not be efficient and the system will oscillate. Therefore, a study about the stability, the noise and the effects of external or internal components has to be developed.

### 2.2.1 Stability study

It remains an important part to explain the phenomena of stability, noise and linearity capable to make the system oscillate. The study and the feasibility of the Cartesian Feedback have been proved by C. Tassin [33]. This study is based on the modeling of the Cartesian Feedback by using the Volterra Series.

#### 2.2.1.1 Modeling of the Cartesian Feedback

The studies of M. Briffa and M. Faulkner [40, 41] propose a detailed analysis of the Cartesian Feedback. They present the evaluation of the stability through the response of the power amplifier for a little perturbation  $\Delta V_i$  around a functioning point fixed by an input vector  $V_i$ . A simplified model of the transmitter in open loop is presented in the figure 2.2.

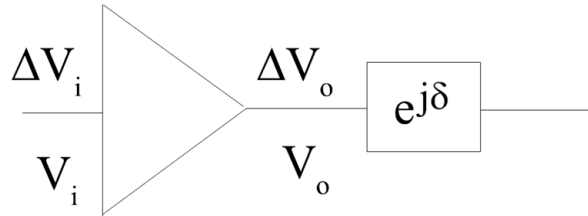


Figure 2.2: Simplified model of an open loop transmitter

The phase shifting between the direct and the feedback path is given by  $e^{j\delta}$ .

The large signal gain  $g$  is defined by the equation 2.1:

$$g = \frac{\Delta V_o}{\Delta V_i} \quad (2.1)$$

As the Cartesian Feedback has multiple inputs and outputs, it can be considered as a MIMO system (Multiple-Input and Multiple-Output), with signals  $I_{in}$  and  $Q_{in}$  for the input path and  $I_{out}$  and  $Q_{out}$  for the feedback path (Cf. figure 2.3).

$G(s)$  represents the gain of the loop and  $\delta$  is the phase shifting of the system.

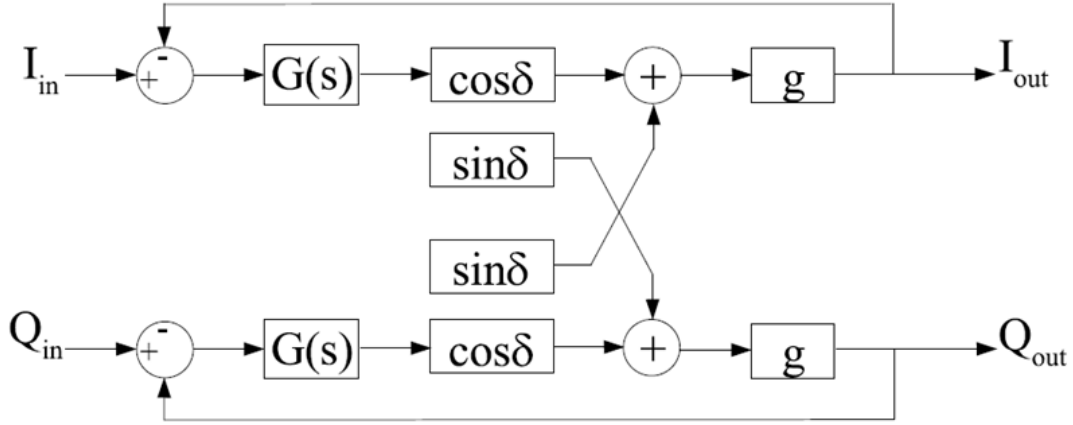


Figure 2.3: MIMO model of the Cartesian Feedback

The open-loop transfer function (OLTF) of the previous model is given by the equation 2.2:

$$OLTF = \begin{bmatrix} gG(s) \cos \delta & -gG(s) \sin \delta \\ gG(s) \sin \delta & gG(s) \cos \delta \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix} \quad (2.2)$$

The model of the system becomes unstable on this equation when the gain margin is zero. The characteristics of the stability can be deduced from the equation 2.3, where  $Id$  is the identity matrix and the absolute value represents the determinant.

$$\left| Id + \begin{bmatrix} gG(s) \cos \delta & -gG(s) \sin \delta \\ gG(s) \sin \delta & gG(s) \cos \delta \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix} \right| = 0 \quad (2.3)$$

Solutions of this stability equation are given by the equation 2.4:

$$gG(s) = -\cos \delta \pm j \sin \delta \quad (2.4)$$

Thanks to the previous equation, we can deduce that the phase margin of the system is equal to  $|\delta|$ .

### 2.2.1.2 Influence of phase shifting on the stability

The main problem of the angular distortion is a classic issue of all systems using a feedback path. The phase shifting can be translated as a time delay which reduces the phase margin and can lead to the instability of the loop. In our case, signals from the feedback path are at high frequencies compared to the baseband signals and can cause a mismatch between I and Q signals. J.L. Dawson [34] has mathematically expressed the RF phase shifting using the schematic illustrated in figure 2.4.

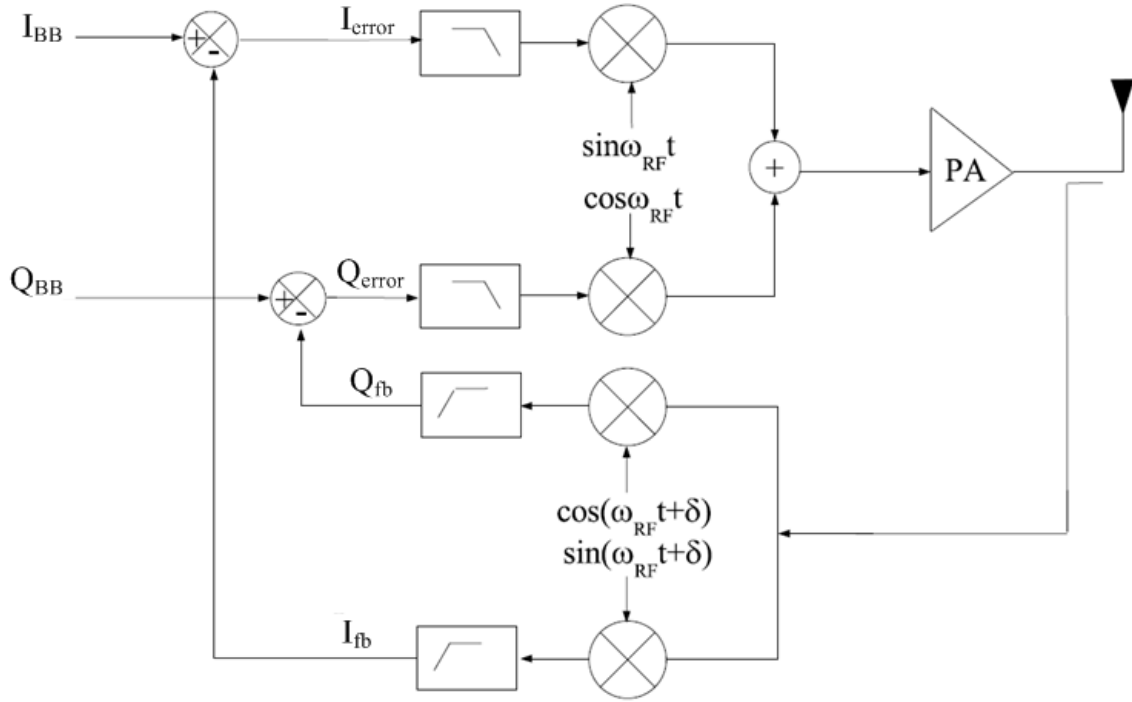


Figure 2.4: Coupling of the I and Q paths by phase shifting

Mathematical expressions of I and Q path at the input of the system are:

$$I_{BB} = I \cdot \sin(\omega_{BB}t) \quad (2.5)$$

$$Q_{BB} = Q \cdot \cos(\omega_{BB}t) \quad (2.6)$$

After the RF modulation, the RF demodulation and the attenuation, we can observe the following expressions of I and Q feedback signals ( $I_{fb}$  and  $Q_{fb}$ ).

$$I_{fb} = \frac{I}{2} \sin(\omega_{BB} + \delta) + j \frac{Q}{2} \cos(\omega_{BB} + \delta) \quad (2.7)$$

$$Q_{fb} = \frac{I}{2} \cos(\omega_{BB} + \delta) + j \frac{Q}{2} \sin(\omega_{BB} + \delta) \quad (2.8)$$

Finally, the subtraction between the original signals and feedback signals is:

$$I_{error} = \frac{I}{2} \sin(\delta) + j \frac{Q}{2} \cos(\delta) \quad (2.9)$$

$$Q_{error} = \frac{I}{2} \cos(\delta) + j \frac{Q}{2} \sin(\delta) \quad (2.10)$$

Mathematical expressions of error signals show the effect of the RF distortion. Each mathematical component is described as a function of the other one. I and Q signals are coupled and modified the initial gain ( $G_{fb}$ ) margin of the Cartesian Feedback as shown on the equation 2.11.

$$G_{fb\_coupling} = G_{fb} \cdot \sin(\omega_{BB}t + \delta) - \frac{jG_{fb}^2}{2 \cdot (1 + jG_{fb} \cdot \cos(\omega_{BB}t + \delta))} \quad (2.11)$$

If  $\delta = \pi$  then there is an inversion of I and Q channels, the response of the loop is greater than zero and the system becomes unstable.

If  $\delta = \pi/2$ , then  $G_{fb\_coupling} = G_{fb}^2$  and the coupling is maximal.

### 2.2.1.3 Graphical criteria of the stability

The RF phase shifting is mainly introduced by the AM/PM conversion of the power amplifier. The increase of the gain amplifies the phenomenon of the AM/AM conversion and reduces, therefore, the phase margin of the system. When the RF distortion is equal to the phase margin, the system becomes unstable.

The graphical explanation is illustrated on the figure 2.5 and expresses the RF phase shifting in function of the gain margin specifying the instability areas.

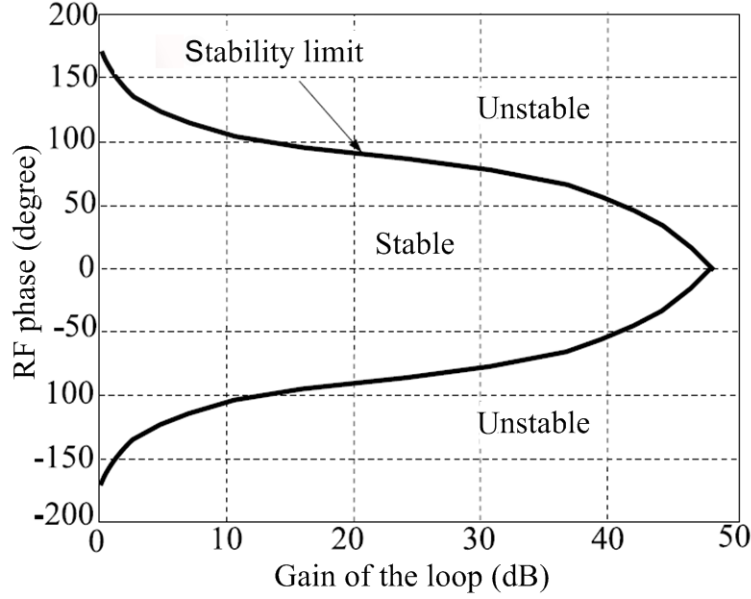


Figure 2.5: Graphical criteria of the instability

### 2.2.2 Noise study

For any communication system, we need a transmitter and a receiver. The receiver has to be as noiseless as possible in order to retrieve the data. However, the transmitter can not be noisy because the carried data will be harder to retrieve. So noise performances of a transmitter have to be checked for the design. Strong levels of noise will degrade the sensitivity of the receiver and will lead to loss useful information.

A noise study of the Cartesian Feedback has been made by P. B. Kenington [42]. This research uses an equivalent model of the Cartesian Feedback represented on the figure 2.6.

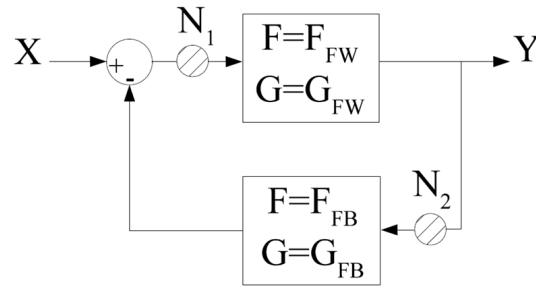


Figure 2.6: Noise model of the Cartesian Feedback

The noise level of the direct and the feedback path are defined as  $N_1$  and  $N_2$ , respectively, where

$k$  represents the Boltzmann constant equal to  $1,38.10^{-23} J/K$ ,  $T_0$  is the reference temperature in *Kelvin* (297 K, 24°C) and  $B$  represents the bandwidth of the noise in *Hertz*.

$$N_1 = F_{FW}kT_0B \quad (2.12)$$

$$N_2 = F_{FB}kT_0B \quad (2.13)$$

The output value  $Y$  in function of the input value  $X$  and the noise values  $N_1$  and  $N_2$  are given by the equation 2.14.

$$Y = \frac{X}{G_{FB}} + \frac{N_1}{G_{FB}} - N_2 \quad (2.14)$$

The equation 2.15 shows the output value  $Y$  in function of the input value  $X$  and the noise factors  $F_{FW}$  and  $F_{FB}$ .

$$Y = \frac{X}{G_{FB}} + \frac{F_{FW}kT_0B}{G_{FB}} - F_{FB}kT_0B \quad (2.15)$$

This model realized by Kenington shows that the noise value of the direct path is attenuated by a factor equal to the gain of the loop. However, the noise of the feedback path is added directly to the noise level of the Cartesian Feedback. This conclusion is only valid within the bandwidth of the system and can not be the same out of band. The noise outside of the bandwidth is generated by RF components working at the frequency of the RF signals, such as mixers or power amplifiers.

### 2.2.3 Linearity

The linearization of a transmitter treats the linearity characteristics of its constituents. As the noise can degrade the performance of the transmitter, the linearity can also deteriorate the carried data and the sensitivity of the receiver. If the levels of harmonic distortions are too high, the receiver will not be able to distinguish the real information in the signal. Therefore, it remains essential to analyze the linearity contribution of each block making the Cartesian Feedback.

For the linearity analysis of our system, we can use the Volterra series [43, 44] which form the model of a system using a feedback path according to the frequency used. The analysis using the Volterra model is strictly used for systems reasonably nonlinear with low-level signals, in order to stay on the linear region. It seems complicated to measure the efficiency of the Volterra

Series, so an analysis using a harmonic balance has to be done at the same time. This kind of simulation aims to equilibrate iteratively harmonics of a system until the convergence criteria is achieved.

The decomposition in Volterra series gives  $n^{th}$  order of the transfer function which are the  $n^{th}$  order of Volterra kernel. The figure 2.7 helps to calculate Volterra kernel of the Cartesian Feedback.

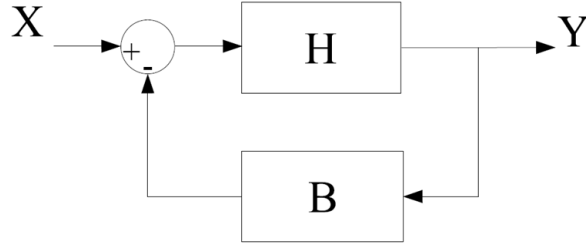


Figure 2.7: Linearity model of the Cartesian Feedback

From [43, 44], we can define the expression of  $i^{th}$  order of Volterra kernel of the Cartesian Feedback  $G_i$  according to  $i^{th}$  order kernels of the direct path  $H_i$  and of the feedback path  $B_i$ . The matrix identity is represented by  $Id$ .

$$G_i(X) = H_i(Id - B_i(G_i(X))) \quad (2.16)$$

For example,  $G_1(f_1)$  is the voltage gain of the Cartesian Feedback and  $G_3(f_1, f_1, f_1)$  represents the  $3^{rd}$  order kernel of harmonics distortions. In order to simplify the study, we will only consider the  $3^{rd}$  order nonlinearities and we will exclude the  $2^{nd}$  order nonlinearities which do not effect our useful channel.

Equations 2.17 and 2.18 can be found with the development of the previous equation until the third order of kernels.  $G_1(f_1)$  and  $G_3(f_1, f_1, f_1)$  represent, respectively, the  $1^{st}$  and the  $3^{rd}$  order of kernels for the Cartesian Feedback.

$$G_1(f_1) = \frac{H_1(f_1)}{1 + B_1(f_1) \cdot H_1(f_1)} \quad (2.17)$$

$$G_3(f_1, f_1, f_1) = \frac{H_3(f_1, f_1, f_1) - H_1(f_1)^4 B_3(f_1, f_1, f_1)}{(1 + H_1(f_1) B_1(f_1))^4} \quad (2.18)$$

The output spectrum of a Zero-IF transmitter has a useful component which is  $f_{LO} - f_{BB}$ , a

LO signal and a  $3^{rd}$  order harmonic distortion (HD3). The figure 2.8 shows these signals, where  $f_{BB}$  is the baseband frequency corresponding to the frequency  $f_1$ .

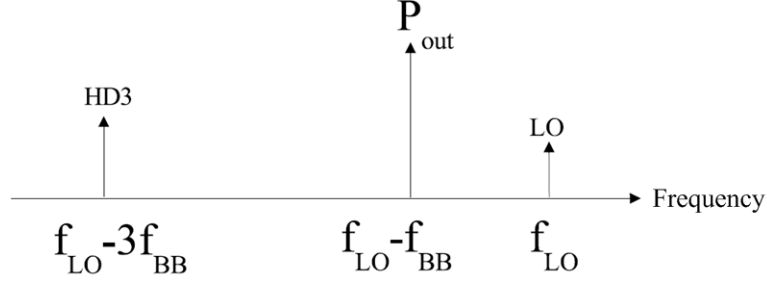


Figure 2.8: Output transmitted spectrum

The HD3 at the output of the Cartesian Feedback is defined by the equation 2.19 [45], where  $V_{in}$  is the level of the input signal.

$$HD3_G = \frac{1}{4} G_3(f_1, f_1, f_1) \cdot V_{in}^3 \quad (2.19)$$

The expression of  $G_3$  (Cf. equation 2.18) shows the linearization of the direct path using a feedback loop. The  $3^{rd}$  order Volterra kernel of the direct path  $H_3$  is largely attenuated by a factor equal to the gain of the loop exponent 4. This equation shows also that the value of the  $3^{rd}$  order kernel for the feedback path  $B_3$  has to be weak in order to maintain the linearization.

#### 2.2.4 Bandwidth of the loop

The W-CDMA standard has specifications according to the ACPR at 5MHz from the carrier but also at 10MHz. For this reason, the Cartesian Feedback has to linearize both channels. The bandwidth of the main channel  $B_{channel}$  of the W-CDMA signals is 3.84MHz and, in order to cover the required predistortion on the input signal, the bandwidth of the Cartesian Feedback ( $B_{CFB}$ ) has to be 1.8 times larger (which represents  $B_{channel}/2 + 5MHz = 6.92MHz$ )[29]. The figure 2.9 shows the signal of the W-CDMA standard with its useful channel and the bandwidth of the Cartesian Feedback.



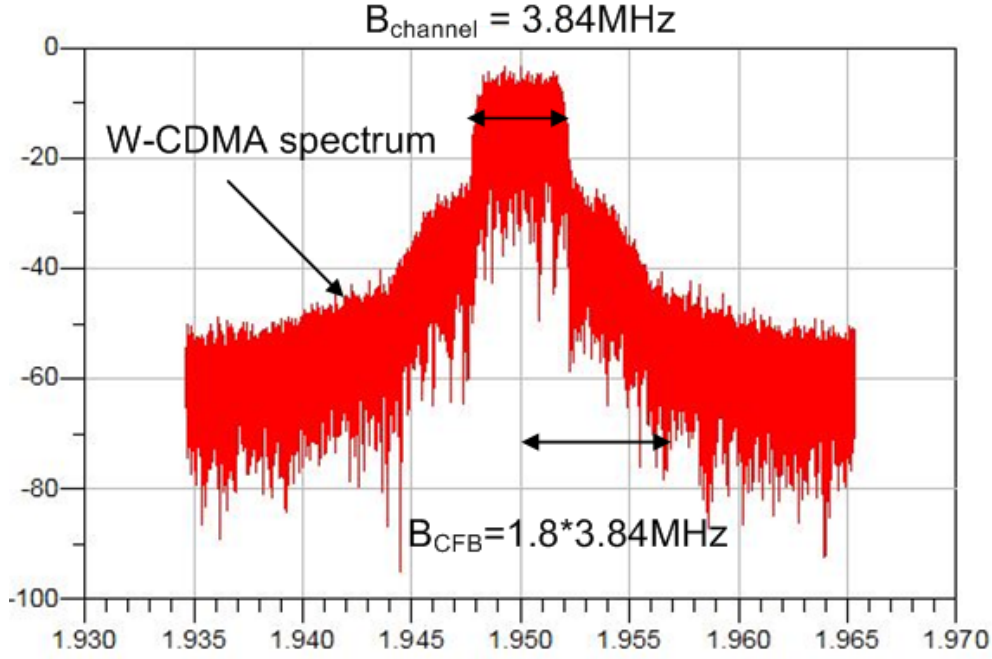


Figure 2.9: Bandwidth of the Cartesian Feedback

The transmitter using the Cartesian Feedback with filters for the direct and the feedback paths will have a bandwidth controlled by the bandwidth of the feedback filter.

### 2.2.5 LO pulling on Zero-IF architecture

For the Zero-IF architecture, using only one stage for the frequency conversion from the baseband signals until the RF signal, some phenomena can interfere between internal components. At the output of the PA, the RF signal can disturb the local oscillator as it is shown in figure 2.10.

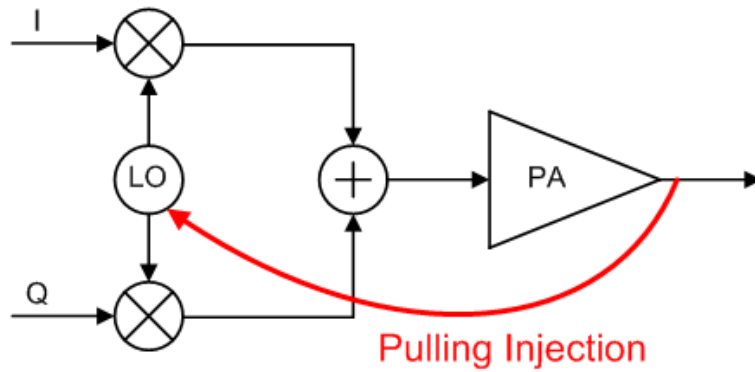


Figure 2.10: Pulling injection for a Zero-IF transmitter

As a consequence, the frequency of the local oscillator will be switched at the frequency of the output signal. Then, the system will be deteriorated and will lose data.

In our case, we propose to place the local oscillator off-chip in order to observe the improvement of the linearization on an integrated circuit.

### 2.2.6 DC offset

Another disturbance can be observed and is linked to the process technology. As a matter of fact, the mismatch of components can deliver a DC offset associated to the baseband signals with the useful information. This offset is transposed into the LO frequency in the RF modulator and represents the DC leakage as it is shown in figure 2.11.

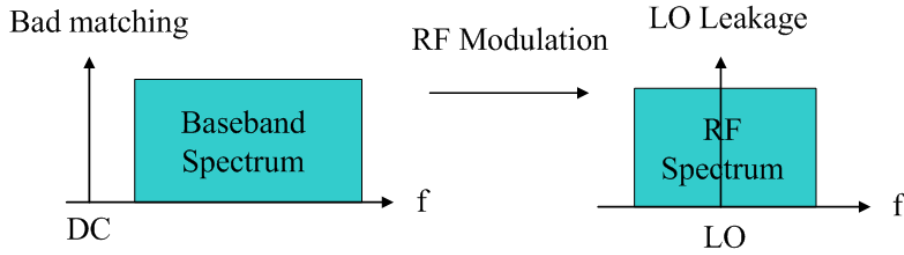


Figure 2.11: LO leakage for a RF transmitter

This leakage of the LO is impossible to cancel, because it is located on the main channel and can deteriorate the sensitivity of the receiver. One solution is to avoid this leakage during the design of the IC layout. If we design all critical paths (e.g. RF signals) the most symmetrical as possible, we can minimize this phenomenon inside an IC.

## 2.3 Digital Part

The digital part remains a critical section in order to insure the stability of the system. After a theoretical study, two solutions have been proposed to make the signal correction of the Cartesian Feedback, using LUTs (Look-Up Table) and a CORDIC (COordinate Rotation Digital Computer) algorithm [46].

### 2.3.1 Context and constraints

The main objective of the digital part is to put signals in phase like the synchronization between the transmitter and the receiver. The direct path can be assimilated as the transmitter and the

feedback path as the receiver. Then, the digital part performs a subtraction operation in order to get the corrected signal. These operations must be the most accurate as possible in order to maintain the stability of the system. Indeed, the delay of the digital part has to be controlled to respect real time constraints of the application on one hand, and keep the system stable on the other.

For W-CDMA standard, the data rate is equal to 3.84Mcps. So the entire delay can not be greater than the delay between two frames which is 260ns. The delay for the analog part is essentially caused by low-pass filters and can be estimated by simulation to 32ns and 45ns for the direct and the feedback filters respectively. From these results, the whole delay of the analog part can not exceed 77ns (32ns + 45ns) and, as a consequence, gives 183ns (260ns - 77ns) for the digital part. This constraint has to correspond to the sampling frequency fixed by converters (ADC and DAC) which is 240MHz.

It is from the digital delay of 183ns that the designer will choose the best solution for the implementation of the digital part on ASIC technology.

### 2.3.2 CORDIC implementation

The CORDIC algorithm is an iterative algorithm developed to make basic operations such as rotation on a vector. This algorithm was improved in order to process trigonometric functions (i.e.: exponential, cosines, sines. . .) by setting its inputs variables.

The figure 2.12 represents the different blocks to consider when the CORDIC algorithm is used.

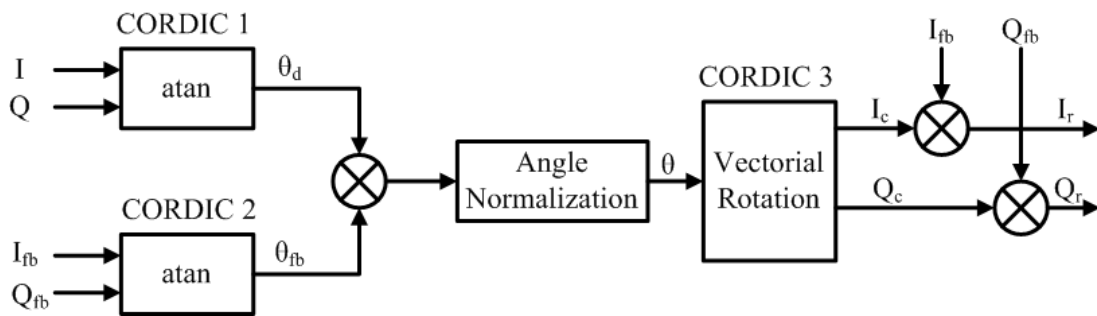


Figure 2.12: Solution based on CORDIC algorithm

### 2.3.2.1 CORDIC algorithm

Electronic computers have experienced a considerable expansion since the 1950s. They began to reach more diverse areas, such as aerospace, and the need for real time calculations became stronger, especially for trigonometric calculations. In 1959, Jack E. Volder created an algorithm which led to approximate trigonometric functions from basic operations (addition, subtraction and multiplication) [47]. This algorithm is called “CORDIC” and is based on the calculation of a vector which is applied in the considered rotation. Thanks to its material integrability (the ingenuity of the algorithm allows a simple electronic connection), the algorithm is used for many applications such as mathematical co-processors and scientific calculators [48].

The CORDIC algorithm is based on Givens rotation [49] as we can see on the equation 2.20.

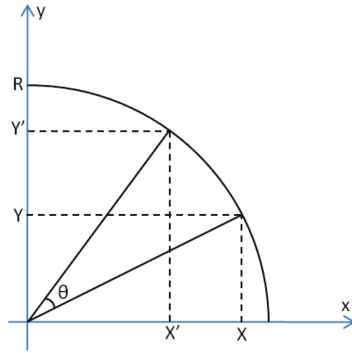
$$R = \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \sin(\theta) & \cos(\theta) \end{bmatrix} \begin{bmatrix} X \\ Y \end{bmatrix} \quad (2.20)$$

This rotation is an iterative rotation with specific angle  $\theta_i = \arctan(2^{-i})$  as it will be shown further. This operation has as a consequence to produce a pseudorotation where the matrix is defined by the equation 2.21. The pseudorotation appears only when  $\theta$  is a sum of  $\theta_i = \arctan(2^{-i})$ .

$$R_c = \begin{bmatrix} 1 & -\tan(\theta) \\ \tan(\theta) & 1 \end{bmatrix} \quad (2.21)$$

As previously mentioned, the CORDIC algorithm uses the vector rotation where the relation is given by the following equation:

$$\begin{bmatrix} X' \\ Y' \end{bmatrix} = \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \sin(\theta) & \cos(\theta) \end{bmatrix} \begin{bmatrix} X \\ Y \end{bmatrix} \quad (2.22)$$

Figure 2.13: Vectorial rotation by a  $\theta$  angle

The idea is to apply  $n$  angle rotations on a vector of coordinates  $\begin{bmatrix} X \\ Y \end{bmatrix}$ . For each iteration, the value of angle decreases until to be equal to 0. At the end of the  $n$  iterations, the sum of all angles is equal to  $\theta$ .

If the initial vector is chosen such as  $\begin{bmatrix} X_0 \\ Y_0 \end{bmatrix} = \begin{bmatrix} X_0 \\ 0 \end{bmatrix}$ , the quotient  $\left[ \frac{Y_n}{X_n} \right]$  clearly represents the approximation of the  $\tan(\theta)$ .

Moreover, if  $X_0 = 1$  then  $X_n$  and  $Y_n$  are approximations of  $\cos(\theta)$  and  $\sin(\theta)$ , respectively.

$\theta$  can be decomposed in a sum of  $\theta_1 + \theta_2 + \dots + \theta_n$ , where  $\theta_i = \arctan(2^{-i})$  is small enough to insure the convergence and it can be used more than once on this decomposition. The figure 2.14 represents iterations that the CORDIC algorithm has to follow in order to converge into the  $\theta$ .

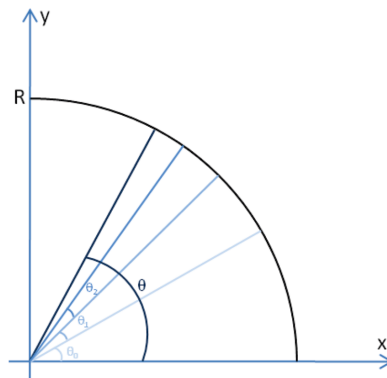


Figure 2.14: Iterative convergence of CORDIC algorithm

Subsequently, we have to calculate  $n$  intermediate coordinates, but in order to calculate them, we have to know how to calculate results of  $\cos(\theta_k)$  and  $\sin(\theta_k)$ . This calculation introduces a

problem because the algorithm has been made to calculate these latters. To avoid this issue, we transform these relations in order to bring up tangents and thus factorize  $\cos(\theta_k)$  as it is described in the following equations:

$$\begin{bmatrix} X_{n+1} \\ Y_{n+1} \end{bmatrix} = \begin{bmatrix} \cos(\theta_n) & -\sin(\theta_n) \\ \sin(\theta_n) & \cos(\theta_n) \end{bmatrix} \begin{bmatrix} \cos(\theta_{n-1}) & -\sin(\theta_{n-1}) \\ \sin(\theta_{n-1}) & \cos(\theta_{n-1}) \end{bmatrix} \dots \begin{bmatrix} \cos(\theta_0) & -\sin(\theta_0) \\ \sin(\theta_0) & \cos(\theta_0) \end{bmatrix} \begin{bmatrix} X_0 \\ Y_0 \end{bmatrix} \quad (2.23)$$

$$\begin{bmatrix} X_{n+1} \\ Y_{n+1} \end{bmatrix} = \prod_{k=0}^n \cos(\theta_k) \begin{bmatrix} 1 & -\tan(\theta_k) \\ \tan(\theta_k) & 1 \end{bmatrix} \begin{bmatrix} X_0 \\ Y_0 \end{bmatrix} \quad (2.24)$$

$$= \prod_{k=0}^n \cos(\theta_k) R_c \begin{bmatrix} X_0 \\ Y_0 \end{bmatrix} \quad (2.25)$$

Where  $R_c$  is a pseudorotation matrix.

The development of the equation 2.24 can be expressed as:

$$X_{n+1} = k * [X_n - \tan(\theta_n) * Y_n] \quad (2.26)$$

$$Y_{n+1} = k * [Y_n + \tan(\theta_n) * X_n] \quad (2.27)$$

Considering only elementary angles, we can replace  $\tan(\theta)$  by  $\pm 2^{-n}$ .

$$X_{n+1} = X_n - d_i * 2^{-n} * Y_n \quad (2.28)$$

$$Y_{n+1} = Y_n + d_i * 2^{-n} * X_n \quad (2.29)$$

where  $d_i$  represents the decision function driven by the sign of registers.

In this case, we can ignore the  $k$  factor. Actually, the amplitude error can be rectified by multiplying only the result of the last iteration by the inverse of the coefficient  $\prod_{k=0}^n \cos(\theta_k)$ . In order to know the direction and the angle of the rotation related to the actual iteration, we use another equation which gives a new value of the rotation angle as it is described on the next equation:

$$Z_{n+1} = Z_n - d_i * \arctan(2^{-i}) \quad (2.30)$$

The algorithm converges by setting the auxiliary variable  $Z_n$  from the start angle  $\theta$  to zero. The  $Z_n$  is updated according to the direction of the rotation. The choice of the value  $d_i$  makes us consider the actual angle of the vector  $Z_{n+1}$  that we subtracted from the wanted angle. We test the result, and, if it is positive, the rotation is in the direction of clockwise, and if it is negative, the rotation is in the direction of counterclockwise (Cf. figure 2.15).

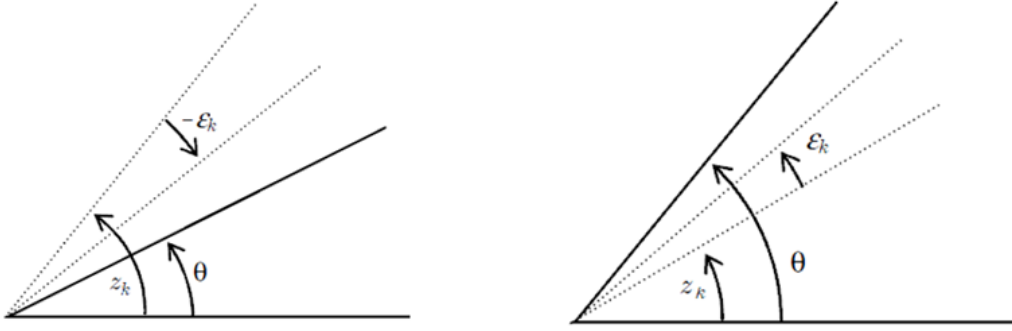


Figure 2.15: Updating of  $Z_k$  value

The arctangent such as  $\theta = \arctan\left(\frac{Y}{X}\right)$ , can be directly calculated using the vectoring mode according to [50]. The CORDIC rotation has two functioning modes: the circular mode and the vector mode.

The vectoring mode allows to perform computation of inverse trigonometric functions such as the arctangent, arccosine, arcsine.  $X$  and  $Y$  are initialized respectively by coordinates  $x$  and  $y$  of the desired point. The variable  $Z$  is then set the 0. The algorithm converges when the value of  $Y$  is equal to 0. The direction of the rotation is defined by  $Y$  [50, 47].

The circular mode (or rotation) represents the conventional functioning mode of the algorithm.  $X$  and  $Y$  are initially set to the coordinates of the point that we want to rotate and  $Z$  is set to  $\theta$ . The algorithm then converges when the value of  $Z$  is equal to zero. Value of  $X$  and  $Y$  are initialized respectively to the coordinate  $(x, y)$  of the targeted point and  $Z$  is set to zero.

The implementation of this algorithm can be realized in several ways depending on the constraints. A compromise between the working frequency and the die area has to be done according to the application targeted. For our case, the phase calculation has to be done in real time. The iterative solution, which uses a register at the output for one iteration, and then re-uses

the same resource, does not fit with our applications. As this architecture is slow, a parallel CORDIC with pipelined register is used as it shown in the figure 2.16 [50].

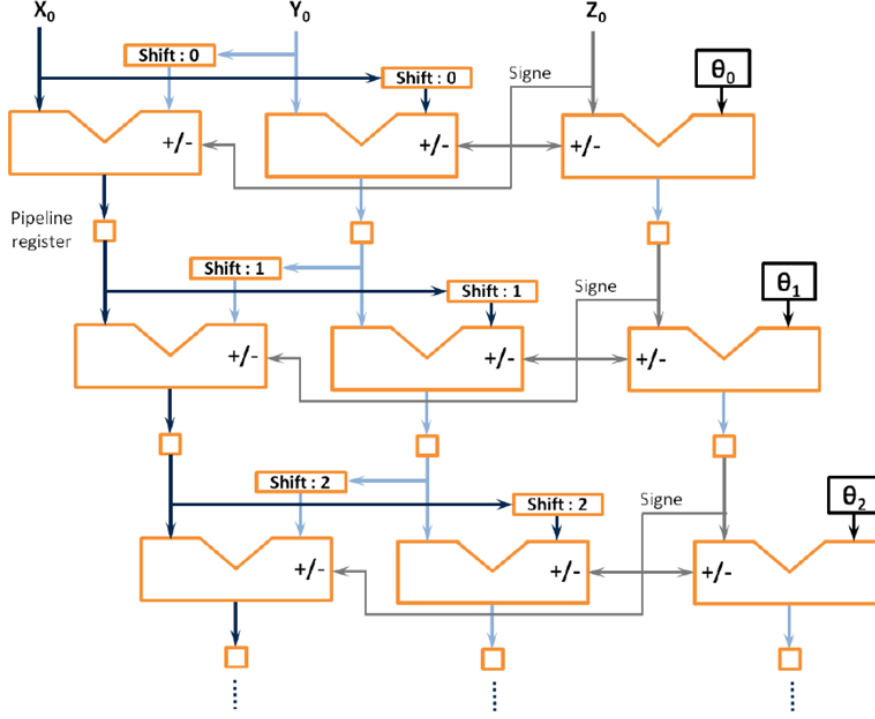


Figure 2.16: Parallel CORDIC with pipelined register

### 2.3.2.2 Normalization of the angle $\theta$

After the calculation of the phase of the direct path ( $I/Q$ ) and the feedback path ( $I_{fb}/Q_{fb}$ ), we have to perform the subtraction of these two angles to find the desired phase shifting.

$$\theta = \theta_d - \theta_{fb} \quad (2.31)$$

Once this subtraction is done, we need to normalize the angle  $\theta_k$  in order to have an angle modulo  $2\pi$ . The implementation of this function in VHDL is realized with successive tests and rotations. Despite the fact that this function does not take up a lot of silicon area, the time-response can be long for our application.

### 2.3.2.3 Subtraction

The digital subtraction is simple for the implementation in VHDL. It is sufficient to calculate the two complements of the second operand and then use an adder to get the result of this operation.



### 2.3.3 LUTs implementation

#### 2.3.3.1 LUTs optimization

As it is illustrated in the figure 2.17, values of sine, cosine and arc-tangent used for the phase correction are stored in interpolation tables (LUT – Look-Up Table). In order to minimize the size of these tables, symmetrical properties are used (Cf. equations 2.32, 2.33, 2.34).

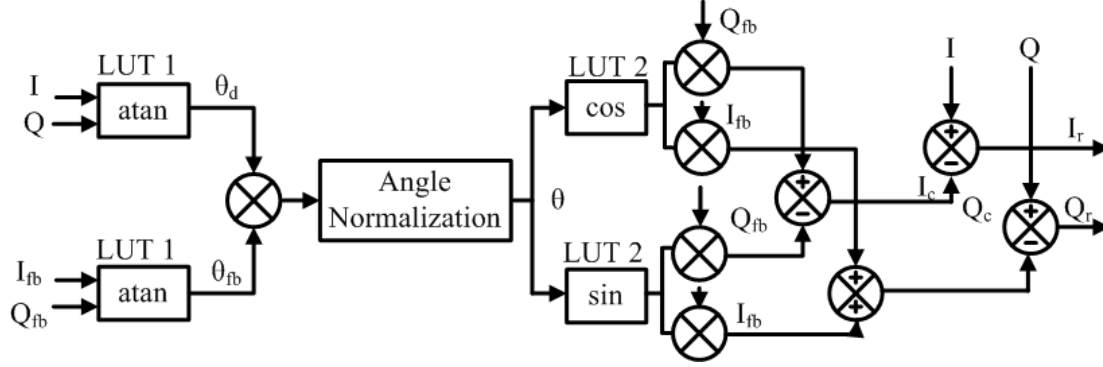


Figure 2.17: Solution based on LUT

$$\cos(-x) = \cos(x) \quad (2.32)$$

$$\sin(-x) = -\sin(x) \quad (2.33)$$

$$\arctan(-x) = -\arctan(x) \quad (2.34)$$

Thereby only values corresponding to positives angles are stored on LUTs. To avoid any overflow and keep an accurate precision ( $\sim 1^\circ$ ) on the phase correction, the floating format used is 12bits including 11bits for the fractional part.

The LUT solution is implemented in VHDL as a ROM (Read-Only Memory). The VHDL syntax is a one-dimensional table which can be read by a simple value.

#### 2.3.3.2 The Restoring Division Algorithm (RDA) divider

The phase error calculation, before the rotation of the signals, uses dividers (Cf. figure 2.17). The division of two signed operands is realized in simulation owing to the operator “/”. But this operation does not work with synthesis tools in VHDL, hence the need to create a division algorithm.

In the literature, we have found algorithms to implement a digital divider [51, 52]. The simplest solution is to use the algorithm of RDA (Restoring Division Algorithm). This algorithm forces into a value of operands strictly positives and also, a value of numerator greater than or equal to the denominator. A division by zero represents a saturation of the maximum value. In order to make the same operations for negative operands or numerators lower than the denominators, a special treatment has to be developed.

The algorithm used is iterative and has as objective to store the dividend in a register A, and the divider in a register B (note that the other registers are set to zero), and further make  $n$  iterations ( $n$  is the length of the quotient).

For each iteration:

- Shift the register (P, A) one bit to the left
- Subtract the value of B from P, and put the new result in P
- If the result is positive, put the most significant bit (MSB) of the register A to 1
- If the result is negative, put the less significant bit (LSB) of the register A to 0, and then restore the old value of the register P adding the value of B to P.

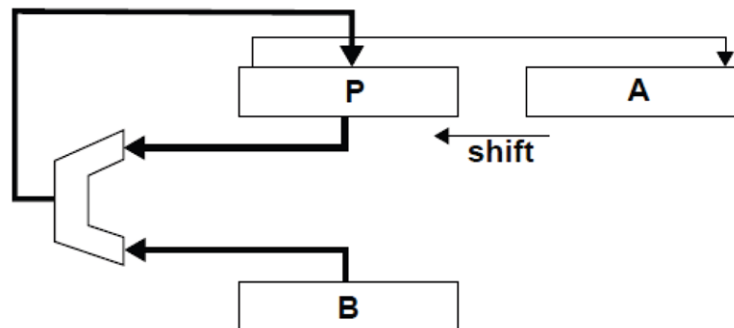


Figure 2.18: Restoring division algorithm

### 2.3.3.3 Conclusion of the digital part

The main parameter that the designer can work with is the frequency of the algorithm. The frequency has to be as high as possible to respect the real time constraint of the application.

In this part, we investigated two solutions to work with the frequency to find which one will respect this constraint better. The first solution is based on the CORDIC algorithm and the second uses LUTs. A theoretical synthesis of both has been done and will be explained further

in the next chapter. This study will give us some ideas of the maximum frequency that each solution can reach for the digital part.

We could think that the CORDIC solution needs less area than the LUT solution as this latter uses look-up table taking up area on the circuit. But, unfortunately, the normalization function of the angle puts limits on the working frequency of the solution. Simulation using ModelSim software with a synthesized circuit gave us a clock 8 times slower than the constraints defined by the standard. On that account, we can be confident of the realization of these techniques and they will be explained deeper with a complete synthesis.

## 2.4 Integration of the Cartesian Feedback on Silicon

The main objective of our study is to realize a single chip with the technology delivered by STMicroelectronics. It remains a good challenge for an IC to consider the stability issue without increasing consumption or area, advantages given by the Zero-IF architecture.

The two following examples taken from recent study [53, 54] illustrate the improvement of the Cartesian Feedback technique.

### Automatic Phase Correction

On his thesis [53], J. L. Dawson realized a Cartesian Feedback in CMOS 0.25 $\mu\text{m}$  technology from National Semiconductor. This solution has an automatic phase corrector which is presented in figure 2.19.

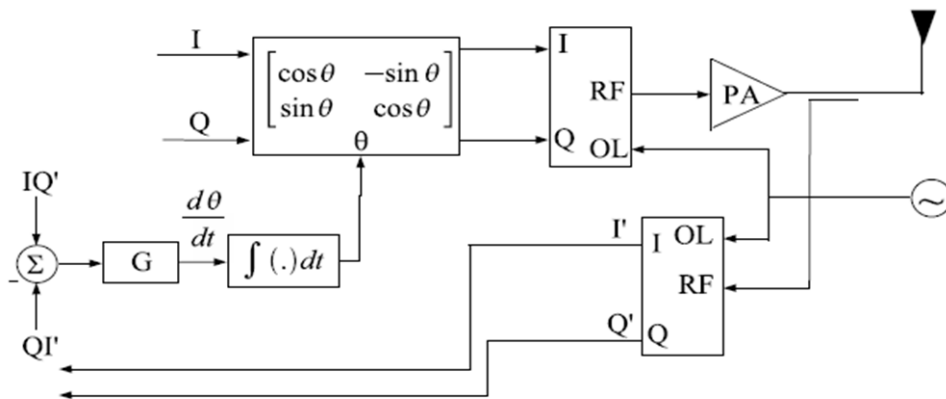


Figure 2.19: Automatic phase correction using a Cartesian Feedback

The goal of phase correction is to multiply I and Q signals by a matrix of circular rotation with an angle  $\theta$ . The phase error  $\theta$  is defined by the successive integration of the term  $IQ' - QI'$ , where I and Q are quadrature signals of the direct path and I' and Q' are quadrature signals of the feedback path. This correction requires the knowledge of the magnitude of the signal in order to adjust the gain of the integrator.

Moreover, if the gain of the integrator is fixed, the method will work only with constant envelope modulation.

### WCDMA Transmitter from STMicroelectronics

The Cartesian Feedback is applied to the last stage of the heterodyne transmitter for the WCDMA standard [54]. The circuit was realized in BiCMOS 0.8 $\mu$ m technology from STMicroelectronics and has a die area of 4mm<sup>2</sup>. This work presents results in a linearization bandwidth of 60 MHz, and an ACPR improvement of 24dB at 5MHz from the RF carrier.

The phase shifting introduced by components of the loop is compensated by adjustments on the LO signals ( $LO_I^*$  and  $LO_Q^*$ ) of the feedback demodulator (Cf. figure 2.20).

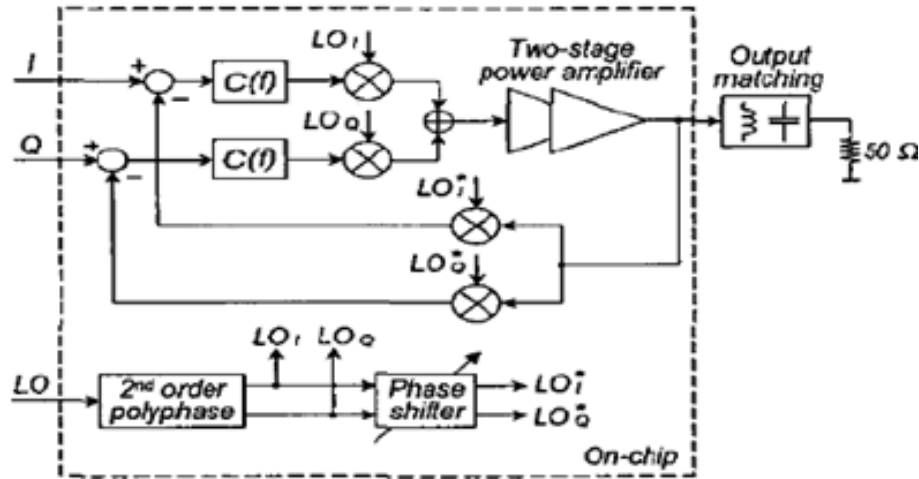


Figure 2.20: Heterodyne transmitter using Cartesian Feedback loop for W-CDMA standard

The settings of  $LO_I^*$  and  $LO_Q^*$  signals are performed using external voltage supply as shown on the equations 2.35 and 2.36.

$$LO_I^* = k \cdot (V_A \cdot LO_I + V_B \cdot LO_Q) \quad (2.35)$$

$$LO_Q^* = k.(-V_B \cdot LO_I + V_A \cdot LO_Q) \quad (2.36)$$

These examples demonstrate how the Cartesian Feedback can improve the linearity with different architectures. Our study presents a different approach of the linearity correction using the same principle.

## 2.5 System Level Behavior

The design of any electronic circuit has to pass through a first step called system level behavior. This step consists essentially in to fix characteristics of all building blocks of the circuit (transmitter, receiver or transceiver). ADS software from Agilent was used in order to have a first idea of the components' properties in this study of the Cartesian Feedback.

We performed simulations on the system level with different power amplifier designs to perform entire simulations of the mixed system at a transistor level.

The first analysis was a simulation using an ideal power amplifier from the library of ADS to validate our innovative approach. Its properties have a gain of 15dB for a saturation output power of 23dB and a compression power of 21dB. This last value has been chosen with the purpose to be close to the saturation value in order to get more interference.

The second experiment was the test of this ideal Cartesian Feedback based on a power amplifier designed in BiCMOS7RF from IMS Laboratory [55]. This power amplifier has been simulated and measured with a gain of 25dB and a saturation power of 33dBm. The compression point is equal to 31dBm for a corresponding PAE of 20%.

Finally the third and last analysis used the previous power amplifier in 65nm CMOS technology [56], that is the same technology that our Cartesian Feedback must be manufactured.

All these simulations will be explained on the following part to show the evolution of the parameters of each block for the improvement of the Cartesian Feedback. A study that remains critical as it will determine the architecture to be used before the design of the CMOS chip.

### 2.5.1 System Level Behavior of the Cartesian Feedback According to the State-of-the-Art

The figure 2.21 shows the simulated system behavior performed with ADS of the Cartesian Feedback. This schematic is composed of low-pass filters, up-conversion mixers, a power amplifier and a coupler for the direct path, and an attenuator, a splitter, down-conversion mixers, and low-pass filters for the feedback path. The correction of the system is made by an ideal phase shifter and a subtractor.

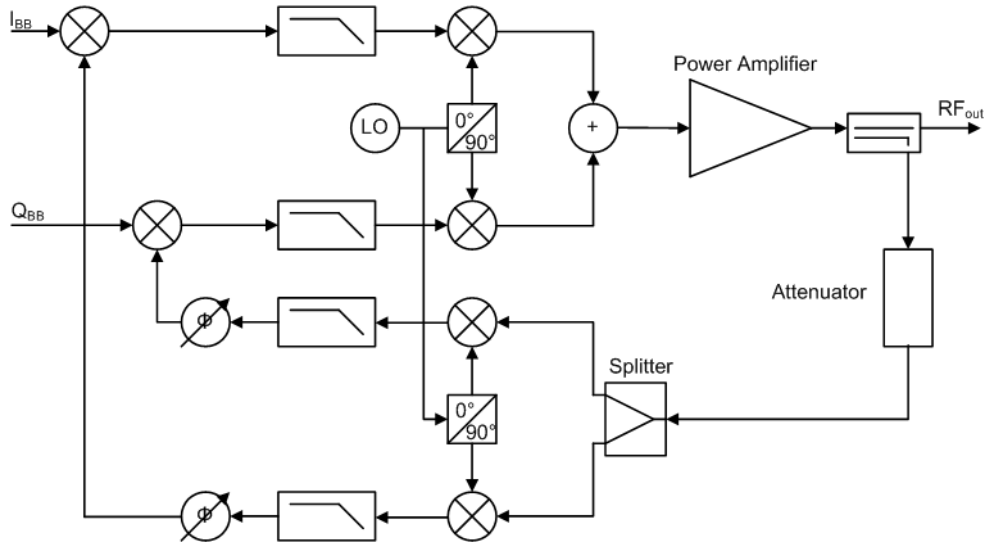


Figure 2.21: Cartesian Feedback technique

The table 2.1 shows parameters that we used to perform the simulation with ADS. In order to perform it, we used the parameters found in the state-of-the-art for each building block. As it is noticeable, the specification is not very aggressive but helps to give us a first idea of the improvement that the system will be able to reach.

Building Blocks	Simulation Parameters
Active up-converter Mixer	Gain = 14dB & NF = 6.4 dB
Passive down-converter Mixer	Gain = -8 dB & NF = 8.4 dB
Active filter of the direct path	2nd order Butterworth low pass, Cut-off frequency = 7 MHz
Passive filter of the feedback path	3rd order Butterworth low pass, Cut-off frequency = 7 MHz
Attenuator	Insertion loss = 6 dB
Power Amplifier in CMOS 65nm	Gain = 15 dB Psat = 23 dBm & OCP1 = 21 dBm

Table 2.1: Parameters of each block making the Cartesian Feedback

This first experiment gave the output spectrum of the power amplifier without and with the linearity correction using the Cartesian Feedback (illustrated in figure 2.22).

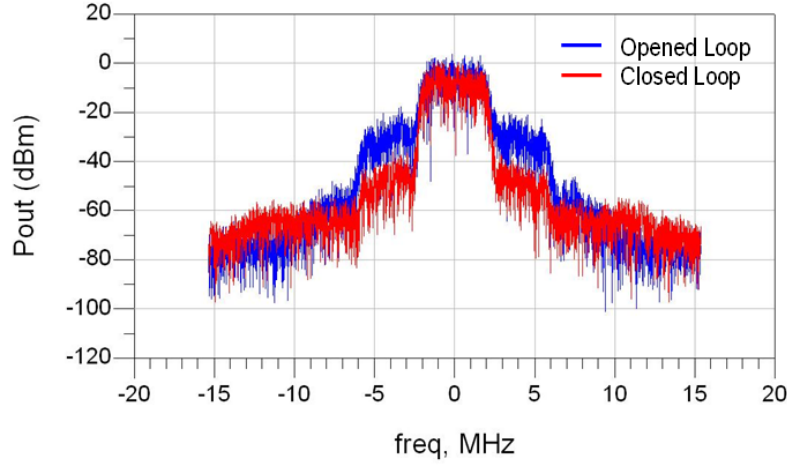


Figure 2.22: Spectrum with and without linearization

	ACLR @ +/- 5MHz from the carrier
Open Loop	- 25.1 / - 26.1 [dBc]
Closed Loop	- 40.7 / - 41.8 [dBc]

Table 2.2: Power on the adjacent channel with and without Cartesian Feedback

A W-CDMA signal was used for almost all simulations showing the behavior of the Cartesian Feedback with a real signal. Moreover, this modulated signal gave us results as we saw previously like the ACPR, the EVM, or the constellation of the standard. The improvement at 5MHz from the carrier is almost 16dB for an output power of 22dBm from the power amplifier, as it is illustrated on the table 2.2.

### 2.5.2 System Level Behavior of the Cartesian Feedback Using a Power Amplifier in BiCMOS7RF technology

After encouraging results, we decided to simulate the system with a PA designed at the IMS Laboratory of Bordeaux [57]. The technology is different from our final integration objective in 65nm CMOS technology, but it helped us to observe more phenomena like the constellation, the phase shifting and the ACPR curves.

The power amplifier was designed in BiCMOS7RF from STMicroelectronics (figure 2.23). This technology is mainly dedicated to RF applications (cellular and wireless device; ie WiFi) because

of its performance and its cost.

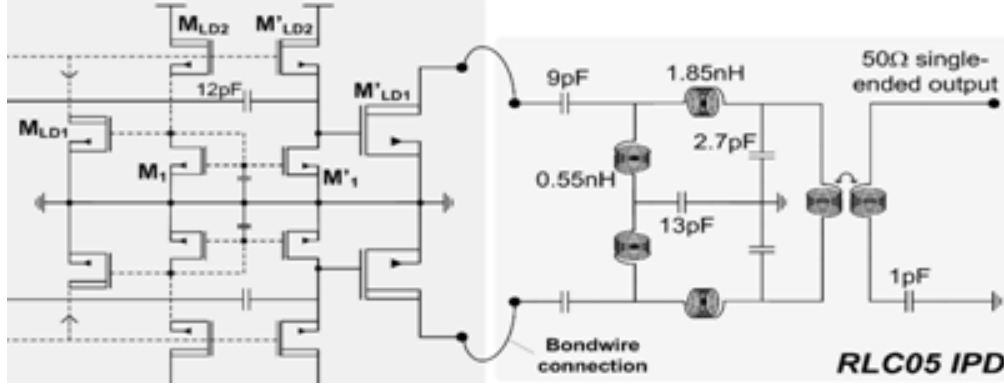


Figure 2.23: Schematic of the BiCMOS7RF power amplifier

For transmitter or receiver applications, this technology permits the use of components such as MOS/LDMOS and heterojunction bipolar transistor (HBT). Therefore, this technology offers advantages of two transistor's type:

- High operating frequencies and low noise factor for bipolar transistors; and
- A strong integration capability with a low consumption of MOS transistors. Moreover, LD-MOS maintains easily a high output power on the operating frequency band.

The power amplifier is composed of two differential stages with nLDMA (laterally doped MOS).

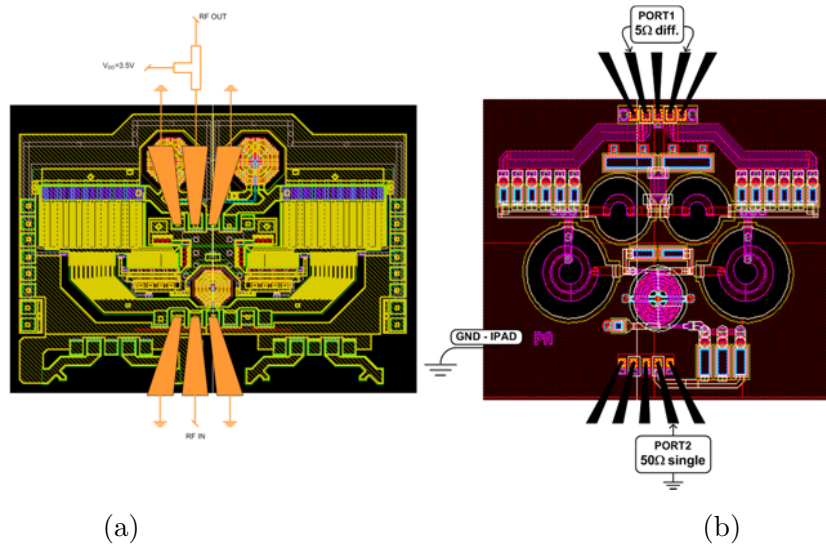


Figure 2.24: Layout of (a) the power amplifier and its (b) IPD associated

The matching network of the PA is off chip and is implemented on a glass substrate (IPD



technology process of STMicroelectronics, Cf. figure 2.24). This particular substrate has as an advantage of limiting its conversion loss, leading to a better efficiency.

This PA has already good linearity performances because it is able to dynamically adjust its output power according to its input power. This adjustment has an objective to decrease the consumption of the power amplifier and enhance the PAE at low and medium power levels.

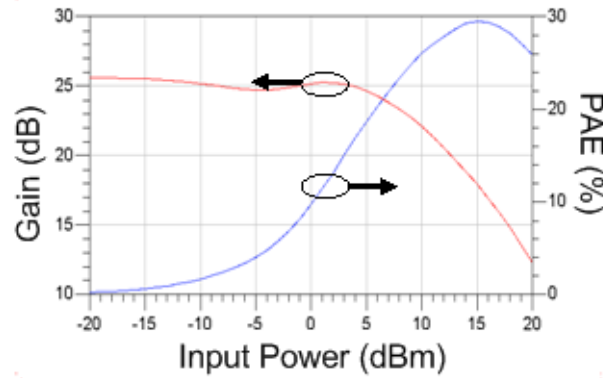


Figure 2.25: Gain and efficiency of the PA according to the input power

The power amplifier at the operating frequency has a gain of 25dB. Figure 2.25 shows the evolution of the gain and the efficiency of the power amplifier according to the input power. The saturation power of the PA is 33dBm for a compression point of 31dBm. As a result, we can observe the spectrum with and without linearization technique on the figure 2.26. This result has already been observed in the previous part and confirms the theoretical study of the Cartesian Feedback.

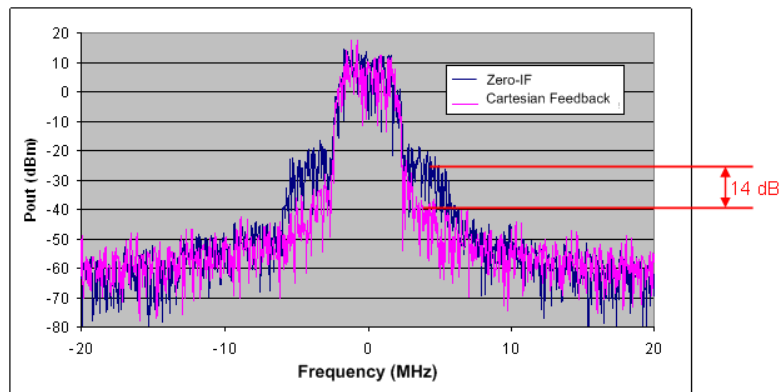


Figure 2.26: Spectrum with and without CFB

The improvement of the linearity can be also observed with the EVM of the output signals. As explained before, the signal at the input of the power amplifier is predistorted. Nonlinearities given by the PA added with the errors signals (predistorted signals) are compensated and produce a linearize signal after the power amplifier. This phenomenon is illustrated in the figure 2.27 with the input signal of the power amplifier (left constellation) and with the output signal of the power amplifier (right constellation). The improvement of the EVM is equivalent to the ACPR improvement previously mentioned, where the right constellation represents a linear behavior of the power amplifier. From this result, the linearization of a power amplifier designed in transistor level is proved and can lead to a new simulation with another PA designed on our target technology.

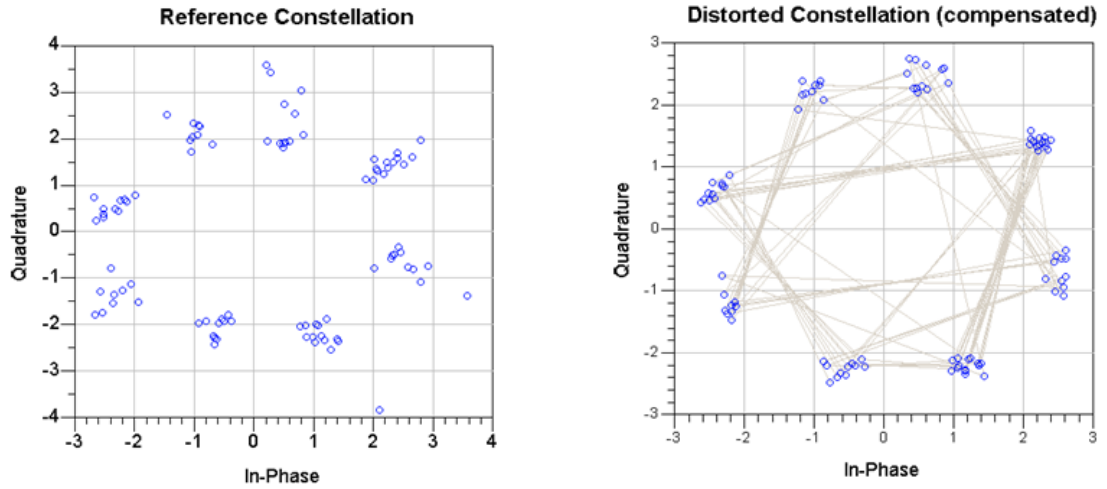


Figure 2.27: Input and output constellation of the power amplifier

### 2.5.3 System Level Behavior of the Cartesian Feedback Using a Power Amplifier in 65nm CMOS technology

In this experimentation, a power amplifier designed in 65nm CMOS technology from STMicroelectronics was used (Cf. figure 2.28).

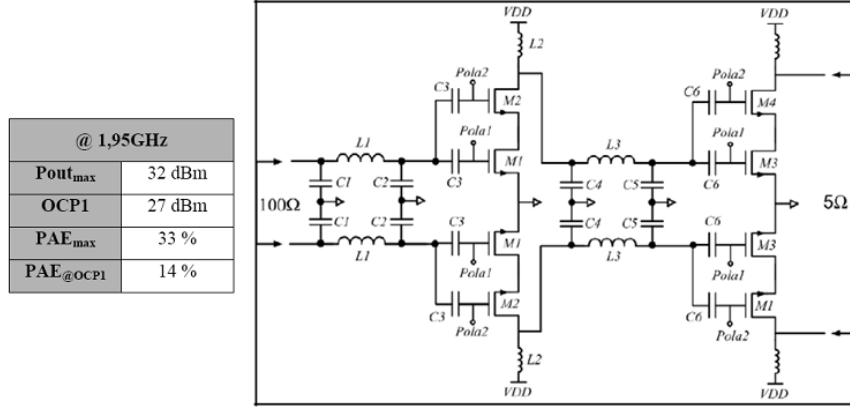


Figure 2.28: Schematic of the 65nm CMOS power amplifier

The power amplifier has been designed at the IMS Laboratory of Bordeaux in 65nm CMOS technology [56]. A simplified topology of the half Stacked Fully Differential Structure (SFFDS) is presented in figure 2.29, which is inspired by a push-pull structure. A high output power will result from this structure, even if a miniaturized technology is used. The two stacked transistors (M1 and M2) are in phase but biased with a different voltage. The drain degeneration is ensured by bonding and MIM capacitors (C1 and C2) are employed for coupling.

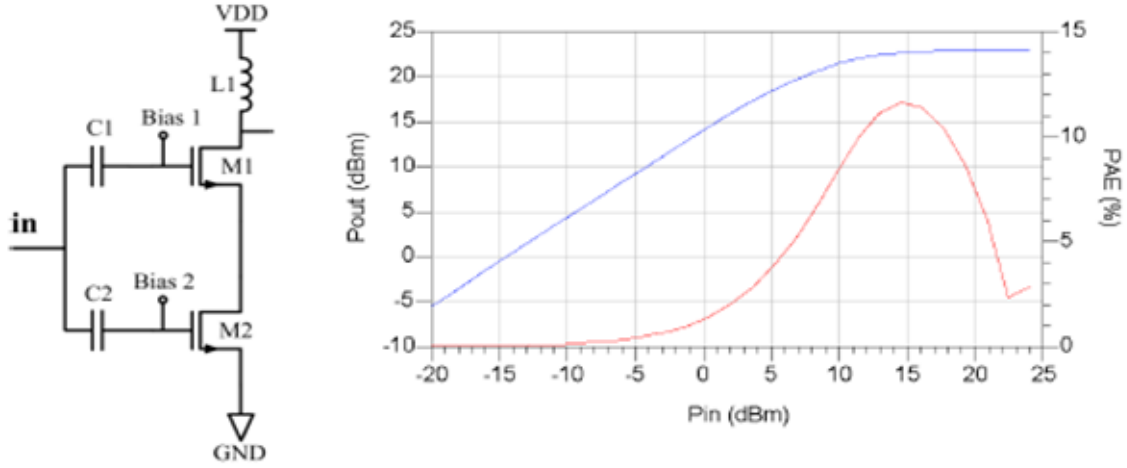


Figure 2.29: SFFDS structure and PA performance associated

At 1.95GHz, the power gain of the power amplifier is equal to 15dB. Figure 2.29 depicts the power gain and the PAE according to the input power. It offers a saturation power of 23dBm with an output compression point (OCP1) of 21dBm. These signal characteristics fulfill the W-CDMA standard (21dBm at 1.95GHz – Class 4) in terms of output power and operating

frequency.

The table 2.3 shows the improvement of parameters for each building block used from ADS simulation to Cadence simulation. These new parameters will lead naturally on new results in ACLR improvement, gain and architecture that will be used.

Components	ADS simulation	Cadence simulation
Active up-converter Mixer	Gain = 14dB & NF = 6.4dB	Gain = 16.3dB & NF = 14.2dB
Passive down-converter Mixer	Gain = -8dB & NF = 8.4dB	Gain = -4dB & NF = 14.6dB
Active filter of the direct path	2nd order Butterworth low pass, Cut-off frequency = 7MHz	3rd order Leap-frog low pass, Cut-off frequency = 8MHz
Passive filter of the feedback path	3rd order Butterworth low pass, Cut-off frequency = 7MHz	1st order Bypass, Cut-off frequency = 8MHz
Attenuator	Gain = -6dB	Gain = -12dB (easily variable)
Power Amplifier in CMOS 65nm	Gain = 15dB Psat = 23dBm & OCP1 = 21dBm	Gain = 13.2dB Psat = 32dBm & OCP1 = 27dBm

Table 2.3: Parameters of each block making the Cartesian Feedback

The ADS simulation part is taken from the previous table made in the theoretical part according to the state-of-the-art. From this previous table, we can notice changes on several components. Simulations performed with Cadence on components taken from the 65nm CMOS library of STMicroelectronics show different results in the architecture and its performance. It brings to our study new results and the adjustment of parameters. Moreover, the attenuator becomes now variable giving a dynamic feedback path.

As results, a simulation of the power of the adjacent channel is shown in figure 2.30. In order to make the comparison and evaluate the efficiency of the technique, ACLR simulation has been done with and without Cartesian Feedback. The figure 2.30 shows that the power amplifier is out of the specifications given by the standard for an output power greater than 28dBm. However with the Cartesian Feedback, the maximum output power reached by the system is 31dBm. The gain of this output power can be translated as lighter specifications on the power amplifier to get a power amplifier available for all classes of the specified standard [58].

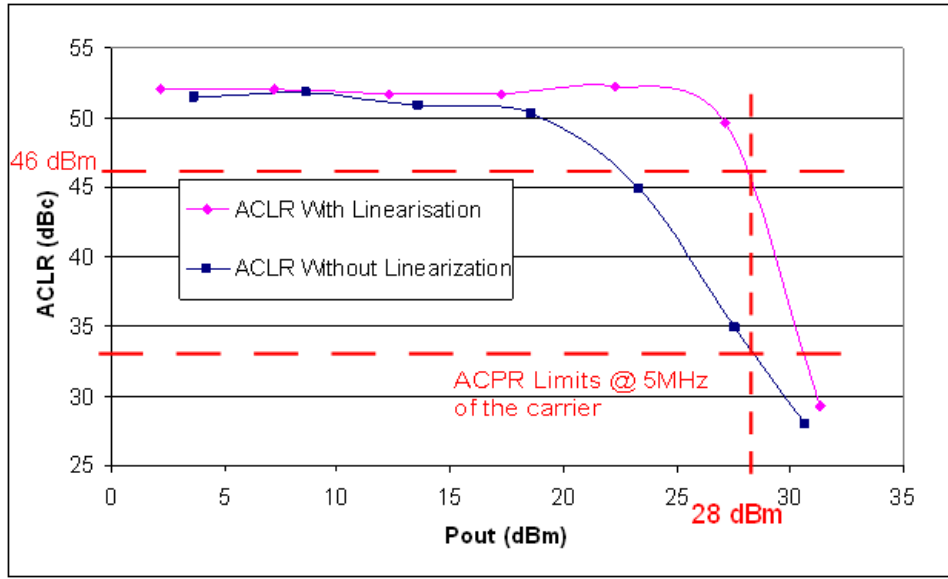


Figure 2.30: ACLR improvement with and without Cartesian Feedback

## 2.6 Conclusion

In the literature we can find the Cartesian Feedback using exclusively analog circuits [32, 31]. In our case, we proposed to split this architecture in two parts: analog and digital. The analog part allows us to decrease the consumption and the die area, to realize the frequency conversion and the amplification of a signal. Moreover, the digital part realizes the phase shifting and the subtraction of signals coming from the direct and the feedback paths, with the main advantage that the correction is dynamical whatever is the response of the output of the transmitter.

Before the realization of the integrated circuit, we have to be aware of limitations and cares that must be taken into account when using the Cartesian Feedback. We must be certain that the system is stable, furthermore, that this architecture will not generate noise and will not degrade the linearity of the circuit. Concerning the restraints of the Cartesian Feedback, we have to deal with a limited bandwidth and the implicit effects of the pulling and the DC offsets. Therefore, mathematical representations were used to illustrate these limitations and cares.

Another point clarified was the real time constraint of the digital part linked to a frequency higher as possible. We explained two solutions for it, the CORDIC algorithm and the LUT with their advantages and drawbacks.

For the digital part, we opted to use an ASIC integration whilst different solutions were illustrated on [53, 54]. These two publications exploited two ways to improve the linearity of the Cartesian Feedback. One applied a more mathematical approach with CMOS technology using only constant envelope modulation but our targeted application uses a non-constant envelope modulation. The other employed an analog technique using the Bi-CMOS technology with our targeted standard with a die area of  $4\text{mm}^2$  and an ACPR improvement of 24dB at 5Mhz from the RF carrier.

Nevertheless, before the realization of a circuit we must follow its characterization through different phases, from an ideal system to a more realistic transmitter. This evolution of stages will allow us to obtain all the properties of the components, to optimize the values of the blocks used in our system, establishing their parameters and providing us the next step, the realization of the transmitter with the Cartesian Feedback.



# A Fully Integrated Transmitter Using a Digital/Analog Cartesian Feedback

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This chapter presents the integration of the analog and digital Cartesian Feedback described previously in a 65nm CMOS from STMicroelectronics.

First, the digital part generating the phase correction and subtraction will be shown in ASIC technology, with a CORDIC algorithm to reduce its consumption and size. The choice of this algorithm will be explained regarding the advantage of being more flexible than a classical algorithm based on LUT (Look up Table). Moreover, constraints (as frequency, latency...) will be described and the implementation of the digital part in VHDL (Very High Speed Integrated Circuit Hardware Description Language) from its description until the layout will be presented.

Secondly, the architecture and each specification simulated of building blocks will be shown. In our case, as presented at the first chapter, the direct path is composed of filters, RF modulator and a Power Amplifier. Our objective is to design these three components to minimize the consumption and the Silicon area of the integrated circuit. Moreover, we do not need to take into account the constraints in noise and in the linearity of the building blocks of the direct path because they will be improved by the Cartesian Feedback itself for the whole system. Nevertheless, for the feedback path, precautions have to be taken regarding the linearity and the noise of this path to prevent distortion of the feedback signal. Moreover, unlike a classic system as the Zero-IF transmitter, our architecture has a feedback path which will not increase greatly the consumption and the Silicon area, as it will be shown.

Finally, system level simulations will be presented using the ADS (Advanced Design Software) from Agilent for the analog part. Co-simulations have been done to analyze the whole system, with SystemVue for the digital part. The simulations using ADS will provide the performance of every building block on the transistors level. However, some hypotheses had to be considered in order to perform on the closed loop mode where only converters (ADC and DAC) are not included in such co-simulations to reduce the time of simulations.

Also, a comparison with the state-of-the-art will be shown answering the needs of manufacturers in term of consumption and Silicon area improvement of the linearity.

### 3.1 A Digital Solution for the Phase Alignment and Subtraction

In this part, we will focus on the synthesis algorithm and the ASIC integration of the digital part (cf. figure 3.1) for the Cartesian Feedback transmitter. Different steps have been realized from the RTL code to the final layout and were applied both to the LUT and CORDIC architectures.

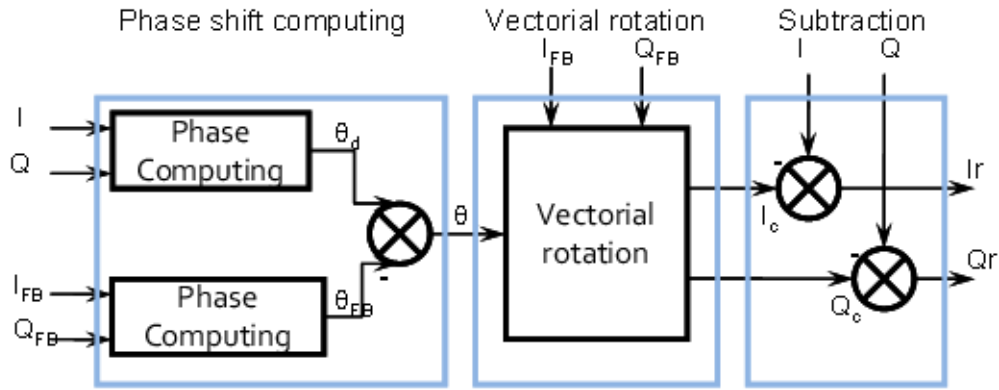


Figure 3.1: Synoptic of the Digital Part

#### 3.1.1 Synthesis and Implementation on a 65nm CMOS technology

To be able to get the final synthesis of the digital part, three points have to be considered beforehand to accomplish the final circuit on ASIC technology (Cf. figure 3.2):

- To check the HDL code to synthesis. This code has to be optimized and can not have heavy tables (e.g. LUTs technique). The number of registers used has to be controlled in order to not increase the dynamic consumption of the solution.
- To choose the technology responding to the constraints given by the client. In our case, the technology is the 65nm CMOS technology. The selected technology has to insure that the technology fits with constraints given by the HDL code.
- To specify input/output delays, raising/falling times given like this all constraints allowing the synthesis of the HDL code.

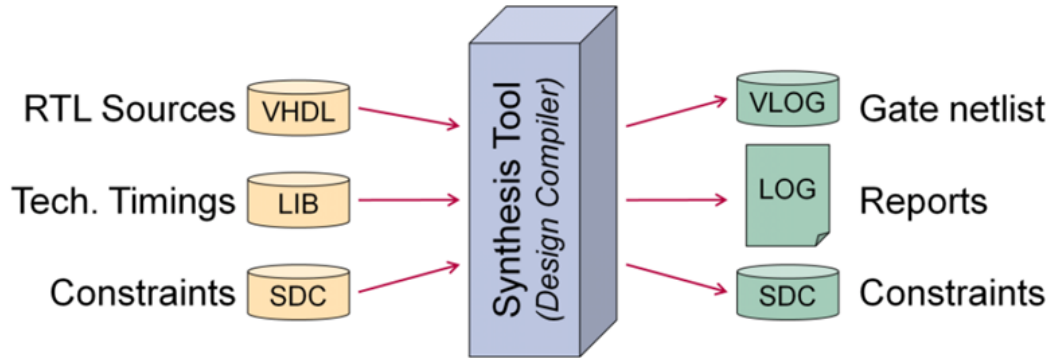


Figure 3.2: Graphical Illustration of the Synthesis Tool

The ASIC synthesis and the mapping have been made with, respectively, Design Compiler tool from Synopsys and Cadence Encounter tool. The synthesis of the digital part can be done in three steps. First, an analysis of the digital code is made with relaxed constraints where the time period is great in order to check if the design is synthesizable or not. For the second step, an incremental synthesis is used, taking the previous result into account. This second analysis has as an objective to optimize the first step in order to get the optimal time period of the digital code. The final step performed a new incremental synthesis with severe constraints and an optimal period is found according to the second step.

#### 3.1.1.1 Synthesis with Relaxed Constraints

The analysis with relaxed constraints has been performed with Synopsys software. This tool generates first Boolean equations from the RTL program, describing the design on the register levels (cf. figure 3.3).

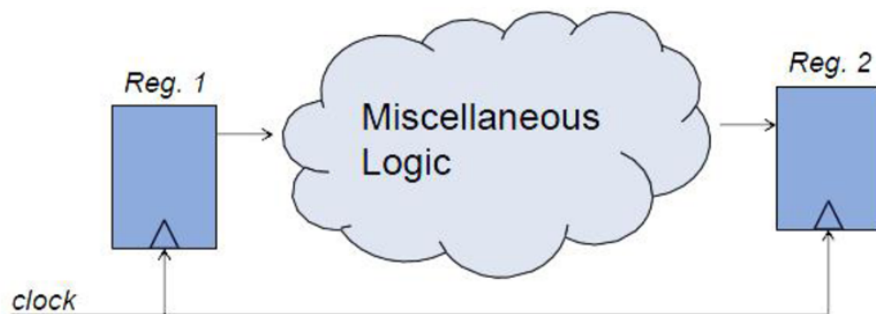


Figure 3.3: Register Level Descriptions

Hence, the compiler translates these Booleans equations in a logical circuit according to the technology used. The gate of transistors used is not ideal and needs to include a response delay in order to approximate the behavior of an ASIC circuit.

In order to have a synthesizable design, the sum of delays generated by each logical gate on the critical path has to be less than the specified period for the synthesis (cf. figure 3.4).

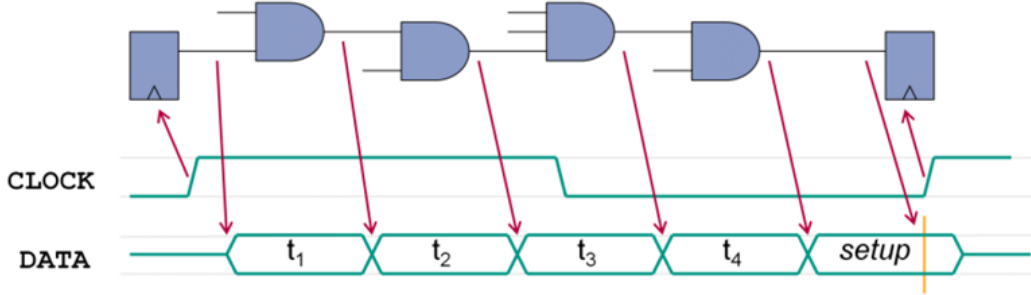


Figure 3.4: Time-Driving Step

At the end of the synthesis, an information file is generated on the SlackTime of the system. If the value is negative, the design or the constraints have to be modified. A positive value enables us to optimize the design using high frequencies constraints.

The synthesis with relaxed constraints consists of using, at first, a chosen period (in our case, the period is 100ns). The maximal period reached by the system is the difference between the specified period and the SlackTime.

	Area ( $\mu\text{m}^2$ )	Whole consumption ( $\mu\text{W}/\text{MHz}$ )	$T_{min}(\text{ns})$	$F_{max}(\text{MHz})$
LUT based	32100	130.9	63.5	15.7
Mixed CORDIC	30800	7.49	8.3	120.5
Pipelined CORDIC	33900	5.92	5.6	177.9
Non-pipelined CORDIC	17400	8.46	18.6	53.8

Table 3.1: Results of the Relaxed Constraints Synthesis

According to the Table 3.1, we observe that the synthesis needs 63.5ns for the solution based on LUTs. This does not allow us to have a high frequency compared to the pipelined solution. In fact, the pipelined solution can reach high-operating frequency, but it occupies the largest die area of the Table 3.1. The mixed solution is able to reach a frequency of 120.5MHz, for a die area inferior to the solution based on LUT.

The synthesis tool does not optimize all of the design because the specified period is pretty high (100ns), but in the following point, the incremental synthesis will present results with a specified period of 15.4ns (reduction of 84.6% on the synthesis period).

### 3.1.1.2 Incremental Synthesis

Results of the incremental synthesis are summarized in Table 3.2. The LUT-based solution remains synthesizable with a period of 15.4ns. Maximal frequency can be reached until 178MHz for the mixed solution and 291MHz for the pipelined solution.

	Area ( $\mu\text{m}^2$ )	Whole consumption ( $\mu\text{W}/\text{MHz}$ )	$T_{min}(\text{ns})$	$F_{max}(\text{MHz})$
LUT based	53500	32.9	15.4	64.9
Mixed CORDIC	37800	17.6	5.6	178
Pipelined CORDIC	35900	23.1	3.4	291
Non-pipelined CORDIC	41900	26.4	5.8	172

Table 3.2: Results of the Incremental Synthesis

The only drawback of the incremental synthesis is that the compiler may no longer understand the logic circuit generated. Every time that we increase the frequency, the incremental synthesis optimizes the circuit generated from the previous synthesis and updates the report. In order to obtain an interpretable result, the synthesis will have to be made again with same constraints used for the incremental synthesis.

### 3.1.1.3 Synthesis with Severe Constraints

Increasing the operating frequency one more time, the LUT-based solution became non-synthesizable. The maximal frequency that the system reached is 65MHz. However, the mixed CORDIC solution gives a frequency up to 200MHz with a die area smaller than the LUT-based solution and the Pipelined CORDIC solution. These results are shown in Table 3.3.

	Area ( $\mu\text{m}^2$ )	Whole consumption ( $\mu\text{W}/\text{MHz}$ )	$T_{min}(\text{ns})$	$F_{max}(\text{MHz})$
LUT based	53500	32.9	15.4	64.9
Mixed CORDIC	44600	19.2	4.3	233
Pipelined CORDIC	51500	28.6	2.5	400
Non-pipelined CORDIC	42500	26.8	5.8	172

Table 3.3: Results of the Severe Constraints Synthesis

Based on these results, we can assume that a study targeting a FPGA device will be inconclusive. In fact, tables used for the LUT solution will be replaced by logical connections. Increasing these connections, the number of transistors will be increased and, by consequence, the die area of the system and the consumption will be also significantly increased. The phenomena will be invisible considering only a FPGA device.

For our application, the mixed solution has to be used for three reasons. First, this architecture does not need pre-computed tables and thus the consumption will be increased according to the operating frequency desired. Then, the solution can reach a maximal frequency of 232.6MHz which fits better with the chosen architecture using ADC and DAC working at 240MHz. Third and last, the die area of the mixed solution remains small if we compare to the LUT-based solution: this last occupies more die area for a frequency lower than 100MHz using the CMOS 65nm technology.

### 3.1.2 Place & Route (mapping) of the digital part

After the validation of the previous step, software called Design Compiler is used. It takes the source code of the LIB file and the result file of the synthesis constraints at the input. At the end of the synthesis, Design Compiler delivers a logic circuit, a report with errors and new constraints given by the software making an optimized place & route (cf. figure 3.2).

We can distinguish five steps to make an efficient place & route of the logical circuit generated at the end of the ASIC synthesis:

*Floorplan:* This first step consists of specifying the die area of the digital part, new constraints, the numbers of layers that the circuit will have and the localization of the input/output pins.

*Power Grid:* A grid has to be placed above the circuit to supply all transistors of the chip.

*Place Design:* According to the die area given previously, this step has to put all the design of the specified area during the floorplan step.

*Tree Synthesis:* After the design and the supply of the circuit were put in place, the clock tree can be placed.

*Route Design:* It is the final step which connects all devices of the entire design together .

### 3.1.3 Die Area of the Digital Part

After the mapping operation, we can observe the layout of the digital part on the figure 3.5. The results of all these operations can be observed for the LUT and CORDIC techniques.

Considering that we choose the same die area for both, we can find the best technique observing the operating frequency and the consumption of the solution (Cf. Table 3.3).

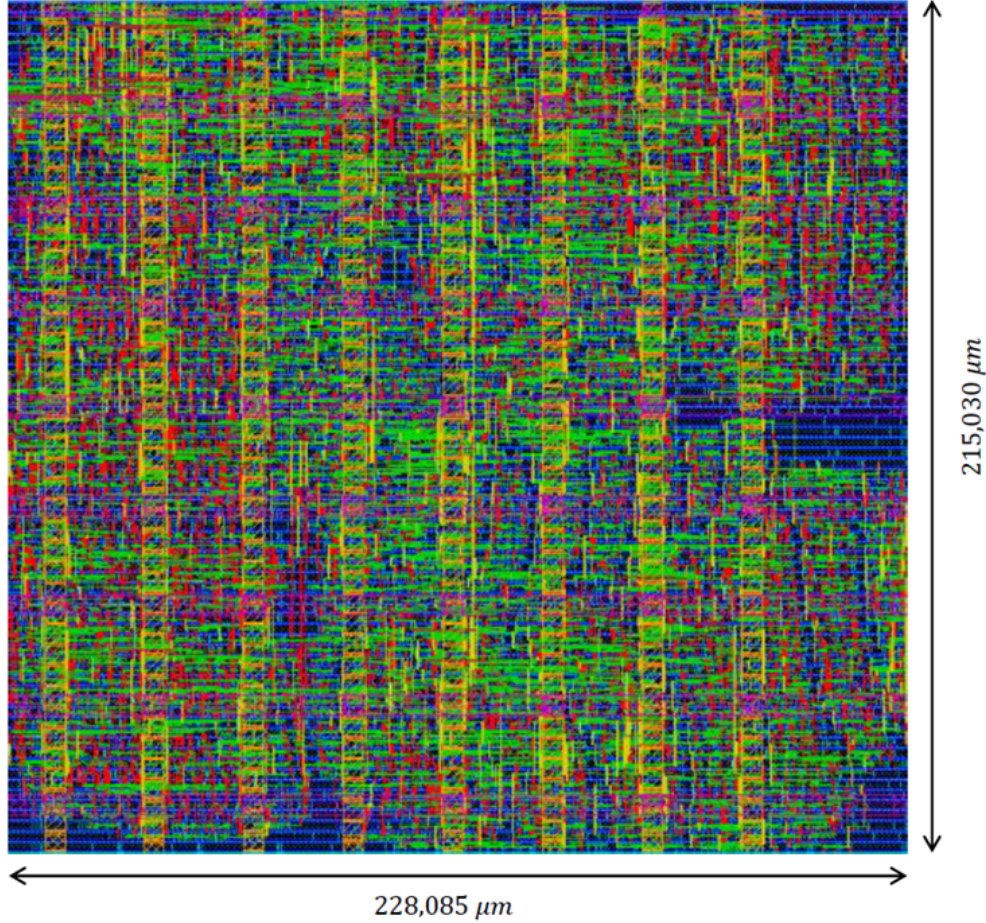


Figure 3.5: Die Area of the Mixed CORDIC Solution

### 3.1.4 Conclusion of the Digital Part

Several steps were described to select for the best architecture for the ASIC implementation of the digital part. The choice of the divider was validated comparing different digital architectures in accordance with the consumption and the die area. Between the LUT and the CORDIC techniques, a detailed architecture of the CORDIC solution has been presented and selected in conformity with the consumption and the area of the chip reduced by the modified technique. The comparison of these architectures shows that the mixed CORDIC solution fits better with the RF linearization technique of a power amplifier.

## 3.2 Details of the Analog Circuit on CMOS 65nm from STMicroelectronics

As illustrated earlier, the analog part of the Cartesian Feedback has two paths, the direct path and the feedback. Converters are used in order to make the interface between the analog and the digital part. A digital-to-analog converter is using in the direct path and the analog-to-digital converter for the feedback path.

The direct path is composed of:

- Filters used for the direct path are a Gm-C filter with a variable gain and cut-off frequency.
- Mixers are based a Gilbert Cell architecture. They are used to perform the up-conversion and the modulation of the I/Q baseband signals.
- Before the coupler, a power amplifier is used, delivering a maximum output power of 27dBm, which permits it to fit with 3 of 4 classes specified by the W-CDMA standard.

The feedback path is composed of:

- After the coupler in the feedback path, a variable attenuator taken from [59] has been optimized for our application.
- A passive ring mixer is used in order to save power consumption and keep a high bandwidth.
- After the demodulator, filters are used to remove intermodulation products resulting from the down-conversion.

This part will detail every blocs making the Cartesian Feedback for the direct and the feedback path. It will show the most significant parameters and results that we have to take into account for simulations and the realization of the whole system. All components presented were designed in CMOS 65nm from STMicroelectronics. It is important to note that critical components such as the power amplifier or converters have been already measured before the integration in the Cartesian Feedback.

### 3.2.1 Digital-to-Analog Converter (DAC)

The architecture used for the digital-to-analog converter is current-steering, widely used for high-speed applications. It has as its advantages easy implementation, a good matching, and a fast response of the conversion [60, 61]. The figure 3.6 shows the classical current-steering DAC architecture.



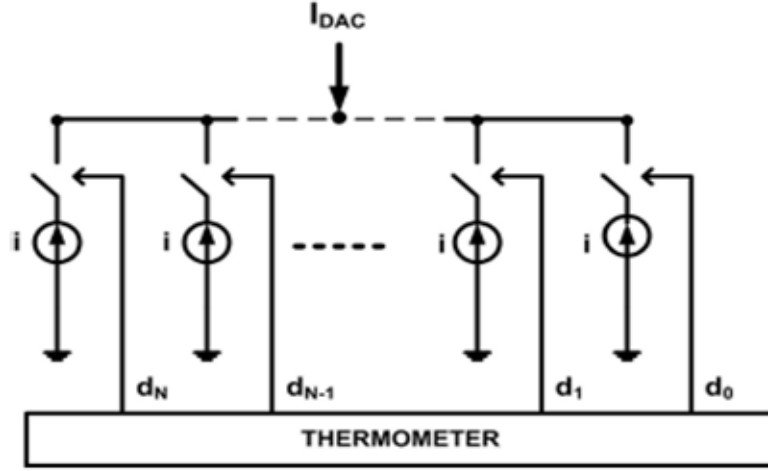


Figure 3.6: General Unit-Element Current Steering DAC

The converter is composed of an array of unit-element reference current sources. Each branch of the DAC has the same reference source, and each branch current is switched ON or OFF depending on the input code. The main drawback of this architecture is the die area. In fact, for a unit-element DAC,  $2n$  current sources are required to generate an accurate output signal. To resolve this limitation, a combination of binary and unit-element current-steering has been used in order to increase the resolution and the linearity.

Table 3.4 shows performances of the converter which has been designed and measured by STMicroelectronics. The converter has differential current output having a full scale current from 6mA to 10mA by setting a proper reference resistor.

	Measurements
Technology	CMOS 65nm
DAC resolution	12 bit
INL	+/- 2.5 LSB
DNL	+/-1.8 LSB
Output bandwidth	80MHz
Frequency sample	242MHz
Area	0.476mm <sup>2</sup>
Consumption	Analog: 26mW   Digital: 2.5mW

Table 3.4: DAC Measured Performances

### 3.2.2 Active Filter

The filter used is composed of three stages (cf. figure 3.7) and it has been designed and measured by STEricsson.

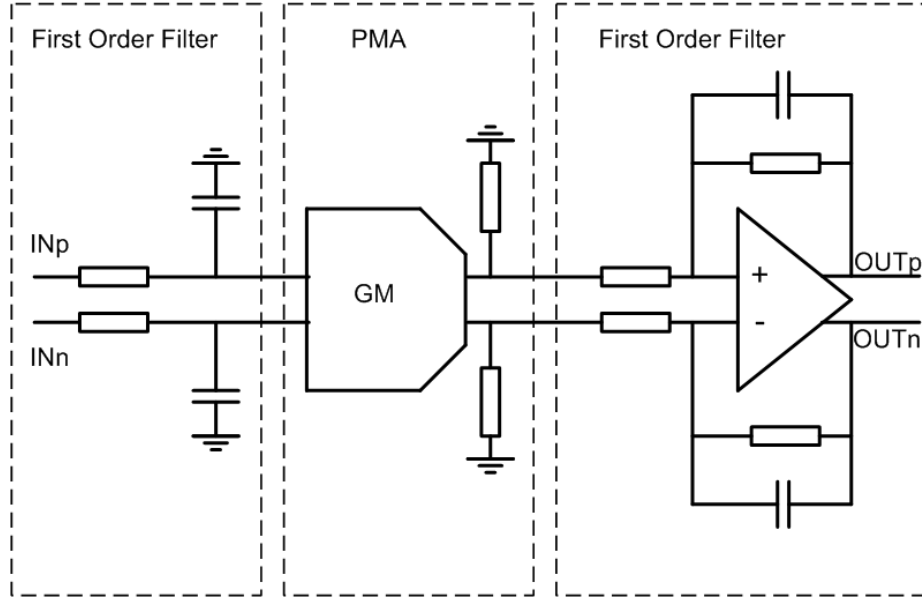


Figure 3.7: Architecture of the Active Filter

The input stage is a first-order passive filter without tuning frequency capability. This filter has a 3dB cut-off frequency equal to 10MHz. This stage is used to adapt the impedance between the block placed before the filter (the converter or the passive mixer) and the PMA (Pulse with Modulation Amplifier).

The second stage (PMA) is an amplifier designed with linear transconductance in order to increase the in-band linearity. It is the most constraining design because of the linearity and noise constraints (applied to the feedback path).

Finally, the last stage is a first-order active filter with a 3dB cut-off frequency equal to 8MHz. This cut-off frequency can be tuned at  $\pm 10\%$  in order to ensure the useful signal.

#### *PMA design:*

In the analog baseband (ABB) structure, two different transistors are used: the lvtlp (low  $V_t$  low power) and the hvtp (high  $V_t$  low power). The inconvenience of the lvtlp is the inability to obtain a large gain of the design as we are using a small  $R_{DS}$  value for transistors. Consequently,

it is very difficult to obtain a large gain in the design. Moreover, according to the design rule manual (DRM) of STMicroelectronics, these transistors have a very bad matching for the gate length higher than  $0.2\mu\text{m}$ . For these reasons, we have used these transistors only when it was necessary. Thus, for the PMA design (as shown in figure 3.83.8), all transistors has been chosen in hvtlp excepted two transistors, the M2 and M3 for stacking reasons.

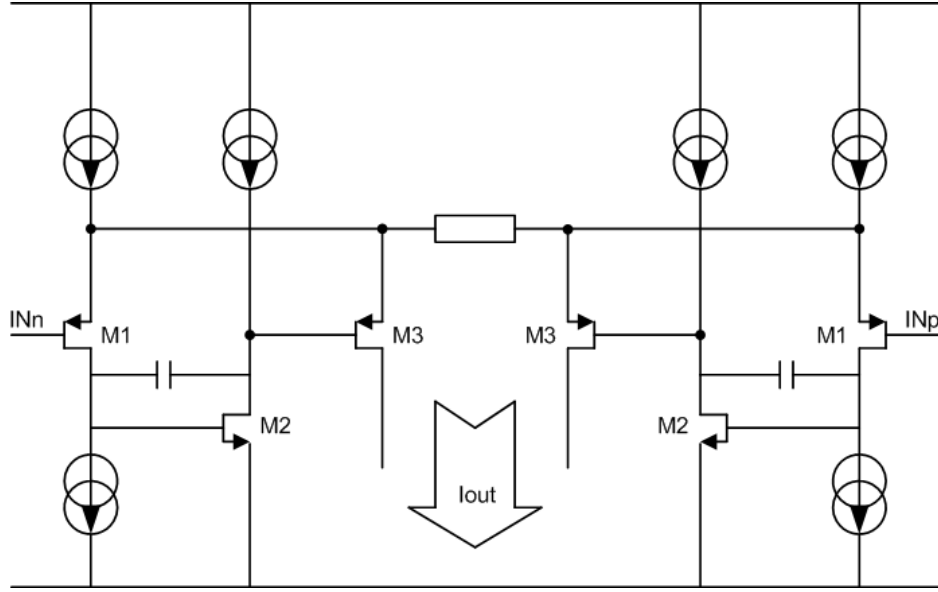


Figure 3.8: Schematic of the PMA

The loop carried out by transistors M1, M2 and M3 are used to control the current through the input transistor M1. This current is the first contributor for the non-linearity of the PMA. Thereby, in order to insure the current control for frequencies higher than 20MHz, this loop must have both a large gain and a large bandwidth (BW=80MHz).

#### *First order active filter:*

This filter is a classical active filter which is used, on one hand, to select the useful channel and, on the other hand, to cancel the adjacent channel (higher than 10MHz) before the ADC.

For linearity and accuracy gain of the filter, the amplifier must have an efficient unit gain bandwidth (BW=150MHz) and a DC gain (50dB). Moreover, as explained before, the 3dB cut-off frequency can be adjusted at  $\pm 10\%$  by switching resistors (Rc1, Rc2, Rd1 and Rd2 in the same time).

The figure 3.9 below shows different gain delivered by the filter. The gain of the filter can be

controlled by value given at the input of the PMA. The filter can have a gain of 8dB, 18dB or 28dB according to the standard targeted. The cut-off frequency can be seen on the graphic for each gain value, and validate its value compared to our application, which is 8MHz.

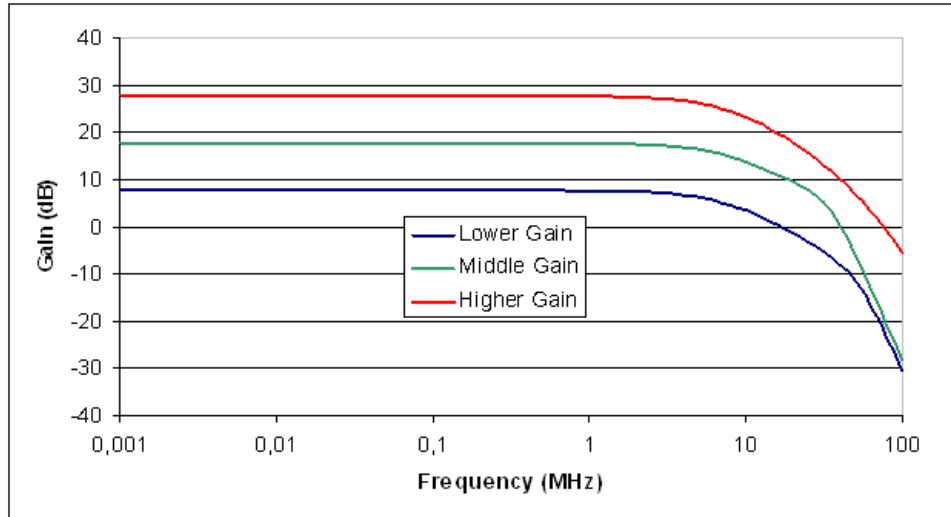


Figure 3.9: Gain of the Filter According the PMA's Functioning Mode

The figure 3.10 shows the die area of the active filter used for the direct and the feedback path which is  $360 \times 360 \mu\text{m}^2$ .

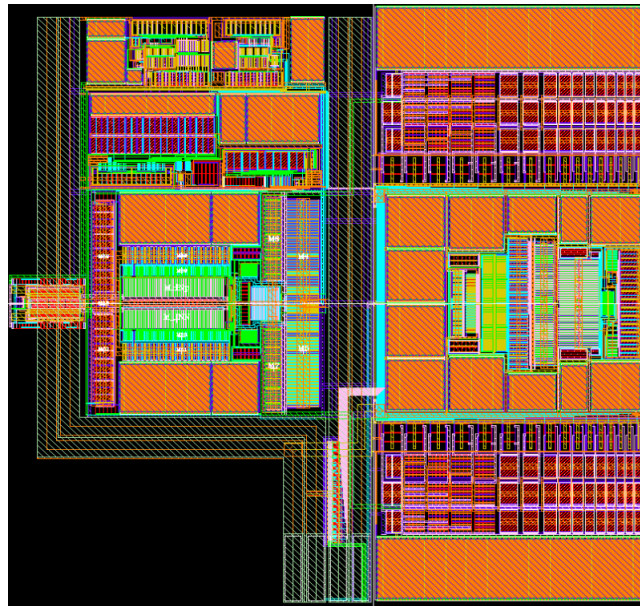


Figure 3.10: Die Area of the Active Filter

### 3.2.3 Active Up-Conversion Mixer

The up-converter mixer is used to transpose the information from a low frequency to a higher frequency. In our case of a Zero-IF architecture, this mixer shifts the information from the baseband to the RF frequency specified by the W-CDMA Standard.

Topologies of mixer can be distinguished as passive or active ones. Passive mixers are used for applications which require good linearity and dissipate no DC current. However, they have no gain and need a large LO signal.

For our application, an active mixer using the Gilbert Cell topology was chosen. This architecture provides a gain to amplify the error signal. The linearity of the mixer is not an issue because the system of Cartesian Feedback will correct non-linearities.

The figure 3.11 shows the Gilbert Cell mixer designed in CMOS 65nm.

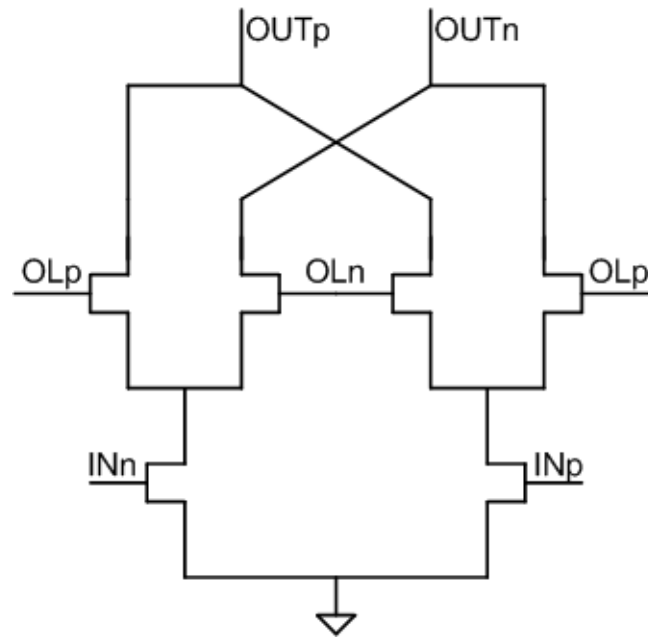


Figure 3.11: Schematic of the Gilbert Cell

The figure 3.12 (on the left) shows the evolution of the gain conversion of the mixer according to the power delivered by the local oscillator. This power has been fixed to 0dBm according to the recent literature given a gain of 16dB for the active mixer. Linearity simulation has been done on the figure 3.12 (on the right). Thereby, the mixer gives a gain of 16dB for a maximum output power of 10dBm and a compression point of -17dBm.

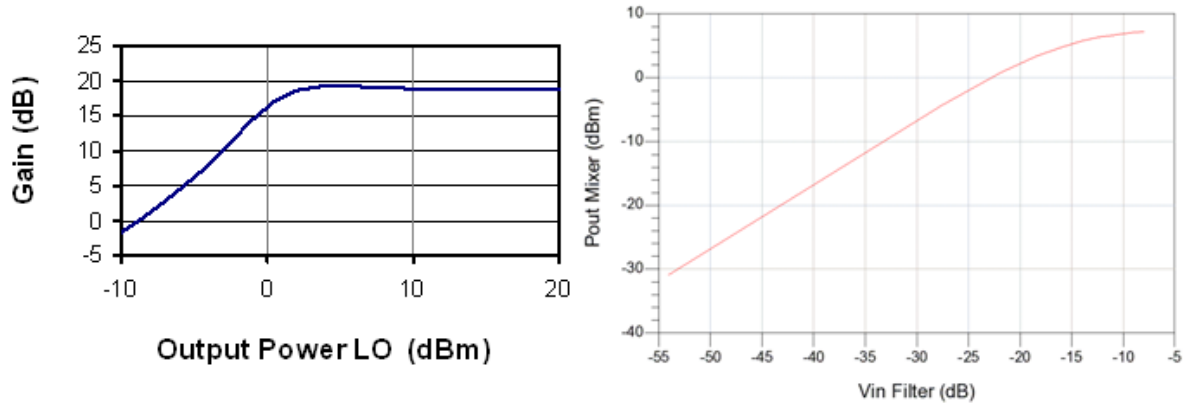


Figure 3.12: Gain Conversion and Linearity of the Active Mixer

As Gilbert Cells give very homogeneous features, they fit perfectly with our application. Its architecture is easy to design and is also reliable as it has been shown previously.

The figure 3.13 shows the die area of the active mixer which is  $270 \times 380 \mu\text{m}^2$ .

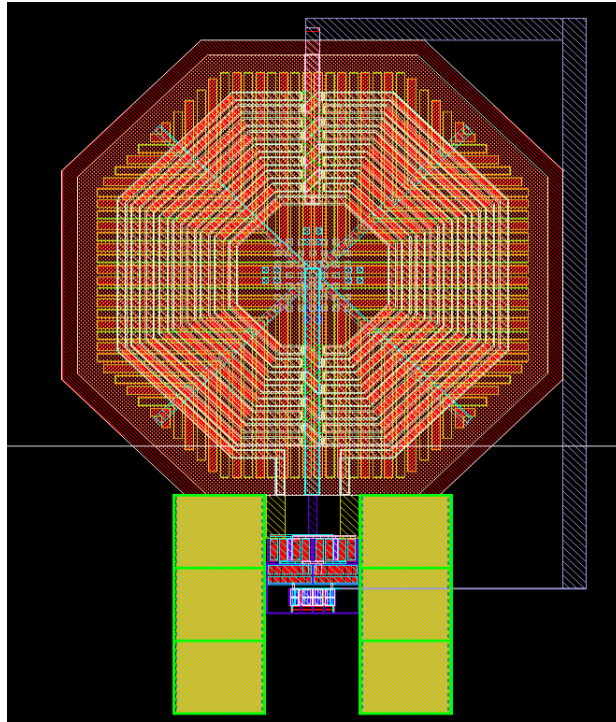


Figure 3.13: Die Area of the Up-Conversion Mixer

### 3.2.4 Power Amplifier (PA)

In order to polarize and match the input and the output of the circuit, the power amplifier is made of active and passive components. The active ones, mainly transistors, determine the power amplifier class of operation.

A simplified architecture of the power amplifier is shown in the figure 3.14. The PA relies on two stages, a driver stage and a power stage, based on cascode and differential structures. The power amplifier has been designed and measured by STEricsson.

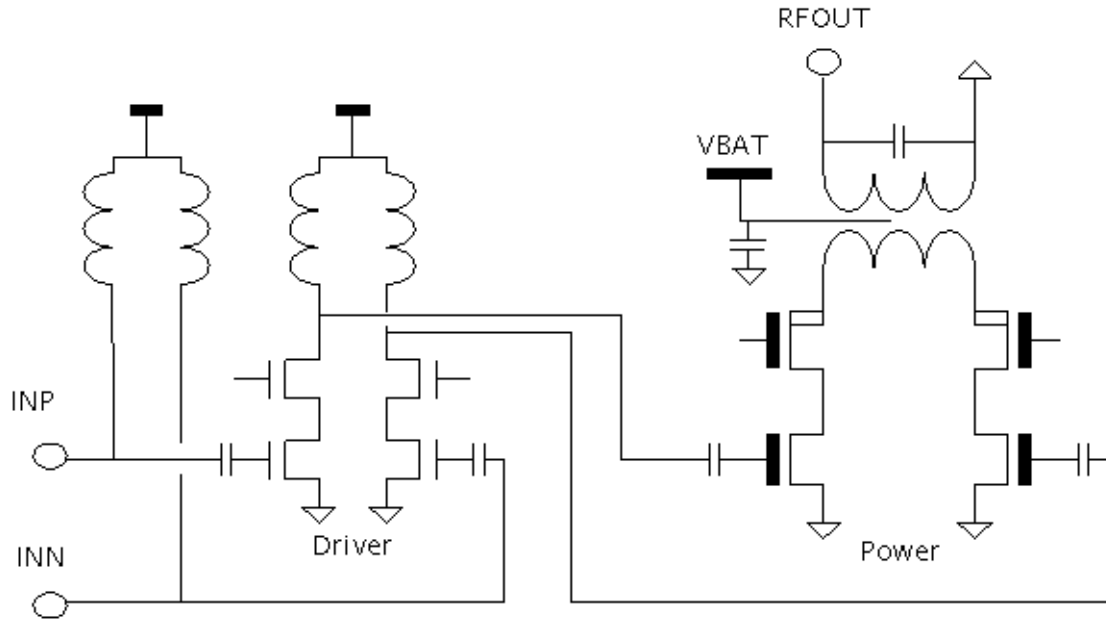


Figure 3.14: Topology of the Power Amplifier

The figure 3.15 shows the gain and the output power of the PA according to the input power. We can notice that the gain delivered by the PA is 24dB that it is much more than we expected from the previous study. This major information will show us after that the higher gain will have an effect to make the correction of our circuit more efficient. In fact, the error signal will be more amplified and thus the non-linearity of the signal will be compensated for easily.

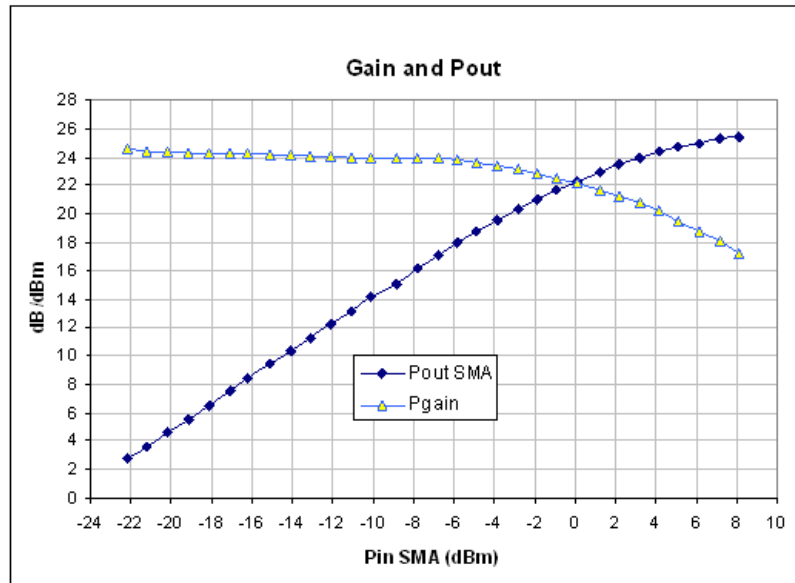


Figure 3.15: Gain and Output Power of the Power Amplifier

The consumption of the power amplifier is a crucial factor for the appreciation of the whole solution. In fact, we will observe that the consumption of the power amplifier will decide if the transmitter is relevant to make it into a product or not. In our case, we can observe the consumption of the PA in figure 3.16. We differentiate the consumption of the driver stage and the power stage called *ldrv* (for 1.2V supply voltage) and *Ibat* (for 3.6V supply voltage) respectively.

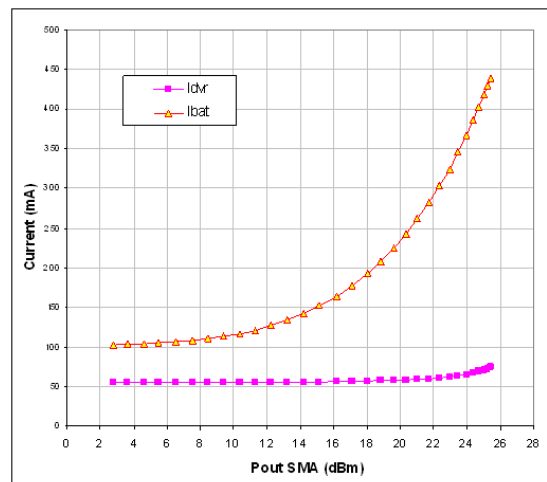


Figure 3.16: Current Consumption of the PA

The total die area of the power amplifier is  $1.3\text{mm}^2$  for a consumption of 658.4mW.



The output matching of the power amplifier has been removed from the original circuit to cancel any mismatching phenomena and to reduce the die area of the transmitter. So the output matching has been realized on the PCB (Printed Circuit Board) as shown in figure 3.17.

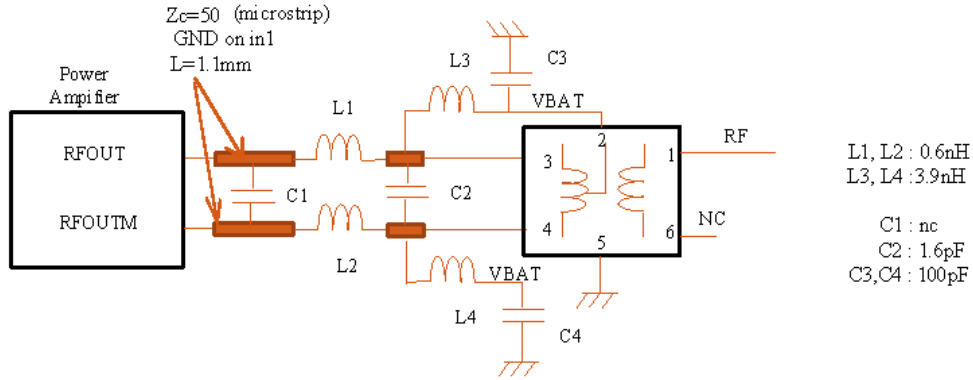


Figure 3.17: Schematic of PCB for the Output of the PA

To perform the measurement of the circuit, we had to change the matching of the power amplifier from a differential output to a single output. As illustrated on the figure 3.18 the matching study was performed with Agilent software. The PCB shows that the differential output of the PA is coming from the left side of the power amplifier. Moreover we can see microstrip lines and the place for balun, capacitors and inductors.

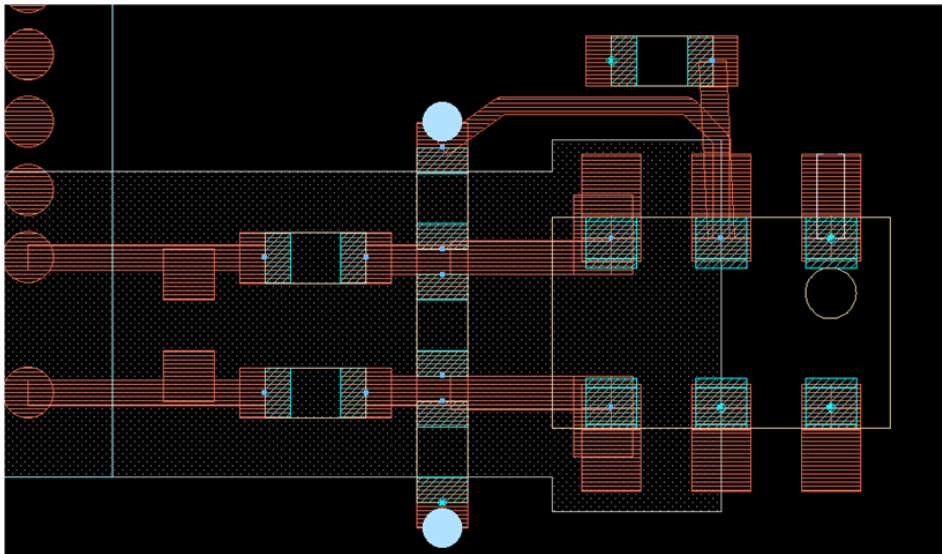


Figure 3.18: Realization of the PCB on ADS

### 3.2.5 Attenuator

A variable attenuator has been chosen in order to insure linearity and stability of the feedback path. The figure 3.19 represents the architecture of the variable attenuator [59].  $V_{ctrl}$  and  $V_{match}$  are used respectively to control the attenuation level and the input/output matching with the help of a DC feedback loop. The variable attenuator is a two stage  $50\Omega$  equivalent  $\pi$ -attenuator.

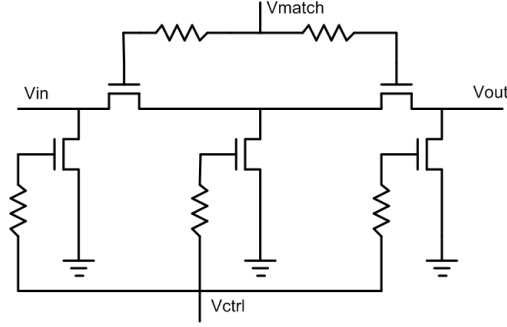


Figure 3.19: Schematic of the Designed Attenuator

The MOS transistors are used in their resistive region to realize voltage controllable resistors. Resistors at the gate of transistors are used in order to modulate the gate with the average of drain and source signal amplitudes. This has a consequence of creating a more uniform channel resistance resulting in less distortion [62, 63]. Two  $\pi$ -stages are cascaded to achieve more attenuation for our frequency, which is 1.95GHz.

The attenuator previously described is replicated in this structure and compared to a resistor of  $50\Omega$  illustrated in figure 3.20.

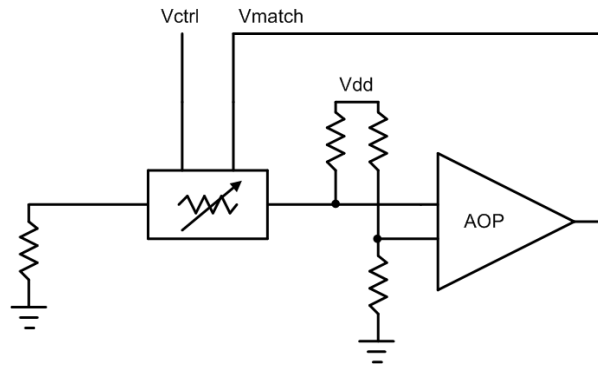


Figure 3.20: Feedback Loop for Impedance Matching

Resistors connected to the gate are removed in order to reduce the die area of the circuit. As

the  $V_{ctrl}$  is varied, the equivalence impedance will be no longer  $50\Omega$ . Then the feedback path will force series devices to increase or decrease their resistance value. This will happen in order to compensate the change in the equivalent impedance of the attenuator. The figure 3.21 shows the attenuation varying at the output while the input power stays almost the same according to  $V_{ctrl}$ .

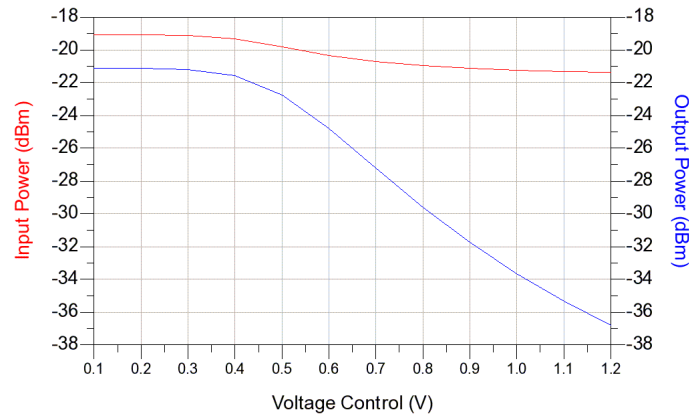


Figure 3.21: Input and Output Signal Variation of the Attenuator

The gain of the attenuator can be observed in figure 3.22 and shows the variation range according to the voltage control. The circuit delivers attenuation going from -2dB to -16dB that will set the attenuation of the whole feedback path and may control the instability of the whole circuit.

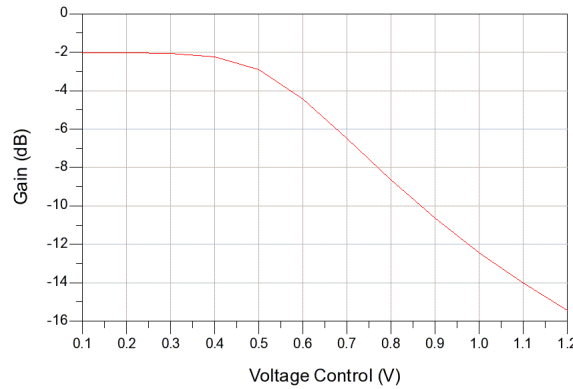


Figure 3.22: Gain According Voltage Control of the Attenuator

Parallel devices are chosen to give a value of  $50\Omega$  when the control voltage of the attenuator is set to 1.2V. The voltage control drives the shunt devices (transistors) which are completely turned off for the minimum attenuation value. It improves the minimum insertion loss of the attenuator. The gain of the attenuator is given by:

$$A = ((R_{shunt} - 50) / (R_{shunt} + 50))^2 \quad (3.1)$$

where  $R_{shunt}$  is inversely proportional to the voltage control value (Vctrl).

The figure 3.23 shows the block diagram of the attenuator.

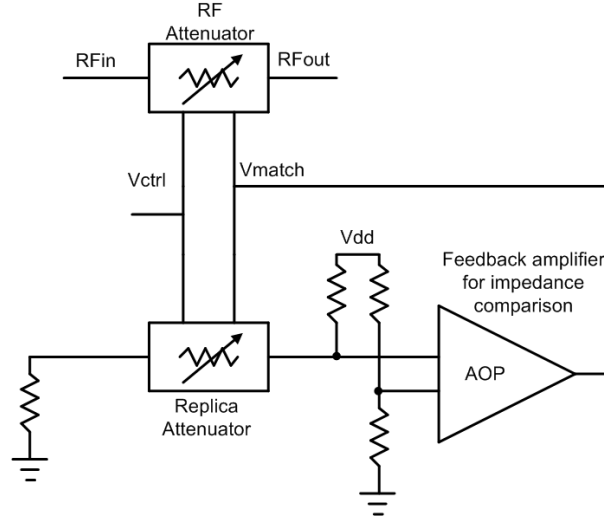


Figure 3.23: Whole Topology of the Attenuator

Finally, the figure 3.24 shows the layout of the attenuator with a die area equal to  $71 \times 63 \mu\text{m}^2$ .

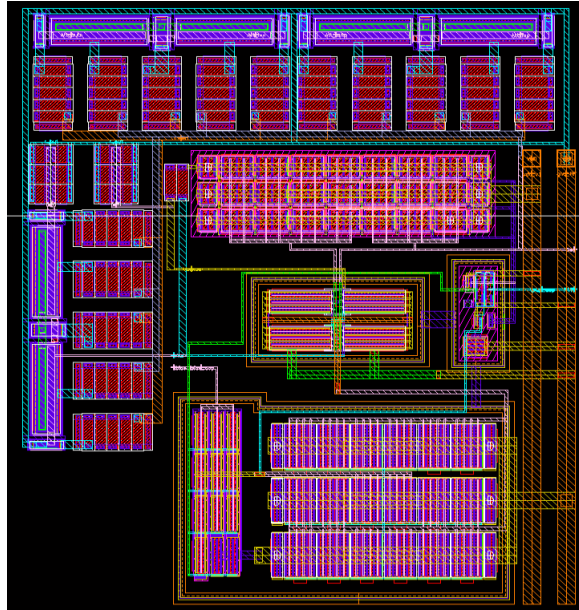


Figure 3.24: Die Area of the Attenuator

The table 3.5 resumes performances of the attenuator.

Operating frequency	1.95GHz
Minimum attenuation	-2dB
Attenuation range	16dB
Insertion loss at the input	2.4dB
Maximum input power	12dBm
Supply voltage	1.2V
Power consumption	1.25mW

Table 3.5: Simulated Performances of the Attenuator

### 3.2.6 Passive down-conversion mixer

Passive mixer of the feedback path is described in figure 3.25. The topology chosen is the quadrature ring mixer with advantages to have a large bandwidth and work in the voltage domain.

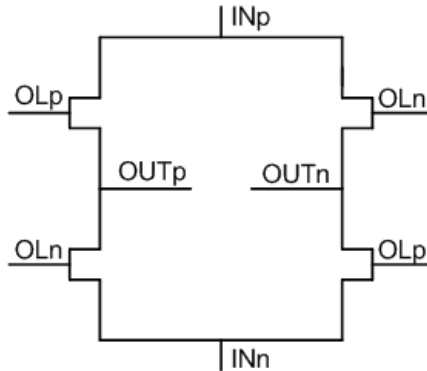


Figure 3.25: Topology of the Passive Mixer

Transistors switch from the active state (ON) to the passive state (OFF), according to the high and low LO level. When the LO signal is positive (maximum voltage of the LO), the output of the mixer is positive and, when the LO signal is negative, the output of the mixer is also negative. This topology does not consume because it does not involve current supply (or outside polarization of transistors). The linearity of this mixer is high according the constant gain for a large dynamic of the output signal.

As explained in the attenuator part, the maximum output power of the attenuator is -4dBm. Thereby the mixer has to be linear for an input power of -7dBm minus 3dB because of the

splitting of the signal towards the I and Q paths. This mixer has a conversion loss [64, 65] because the topology used is passive and is defined by the equation 3.2:

$$G_c = \frac{2}{\pi} \quad (3.2)$$

The linearity of the mixer is related to its output impedance. In our case, the input impedance of the baseband filter is an open impedance which insures a good linearity of the mixer.

The figure 3.26 shows the die area of the passive down-mixer and measures  $5.5 \times 24 \mu\text{m}^2$ .

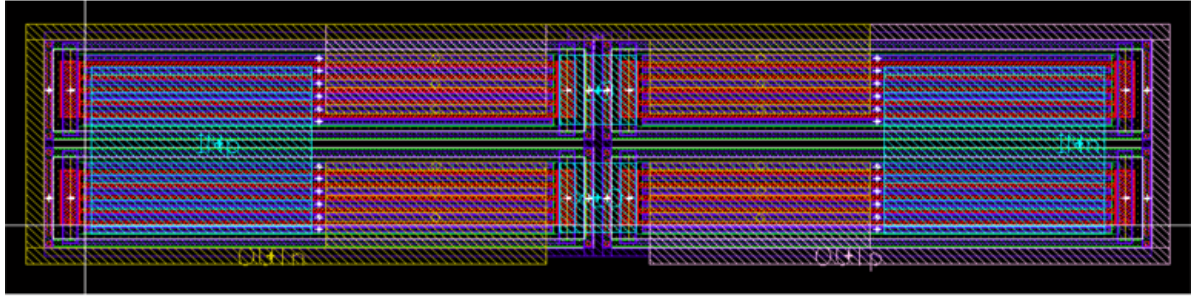


Figure 3.26: Die Area of the Down-Mixer

The filter of the feedback path is not described here because we are using the same filter that the direct path previously described.

### 3.2.7 Analog-to-Digital converter (ADC)

The architecture of the analog-to-digital converter is a pipeline converter. The principle of the pipeline converter is that for each rising clock, the converter makes  $n$  parallel conversions, where each conversion is a part of the output code. Figure 3.27 describes the architecture of the parallel converter. For each clock period a new digital code is performed according to  $n$  voltage values. As a consequence this digital code has to be delayed in order to make the correct output code according to the number of stages of the converter.

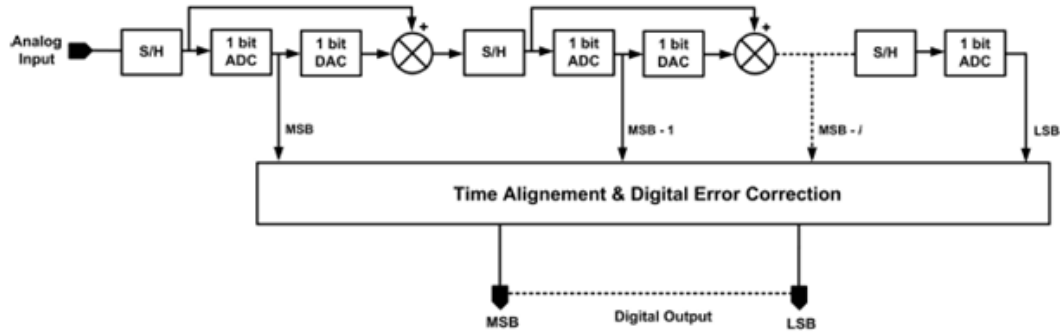


Figure 3.27: Classical architecture of a pipelined converter

The ADC is a parallel pipeline architecture with two identical internal pipeline units. Each internal pipeline unit is based on a double sampling architecture.

The first stage is a 1.5bit stage for high analog bandwidth constraint (130MHz). It is followed by four stages of 2.5bit per stage resolution and finally a 3bit flash ADC.

A flash converter (cf. figure 3.28) is composed of parallel comparators. A  $n$  bits converter needs  $2n-1$  comparators and  $2n$  resistors. The conversion is performed for each clock period. This architecture has a drawback: occupy a large die area because of the number of comparators. But the main drawback of this architecture is the consumption, as for each supplementary bit the consumption is doubled. It is for this reason that a 3bit is employed, making a good compromise between performance and consumption.

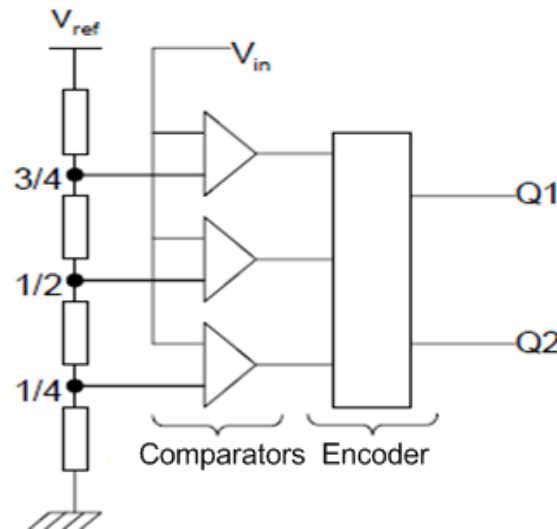


Figure 3.28: Flash Converter

A biasing is used to reduce power consumption adapting the consumption to a wide sampling frequency range from 81MHz to 324MHz. Table 3.6 presents performances of the converters which has been designed and measured by STMicroelectronics.

DAC resolution	12 bit
INL	+/- 2LSB
DNL	+/- 1LSB
Input bandwidth	130MHz
Fs	324MHz
Area	0.3mm <sup>2</sup>
Consumption	Analog: 90mW/Digital 1mW

Table 3.6: Measured Performances of the ADC

### 3.2.8 Circuit assembly

The schematic of the whole architecture is illustrated in figure 3.29. The Cartesian Feedback shows all the analog building blocks without converters.

Converters are not implemented on the chip because supply options (GO2 are respectively equal to 1.8 and 2.5V) of them are different and thus it is impossible to put two types of supply on the same chip.

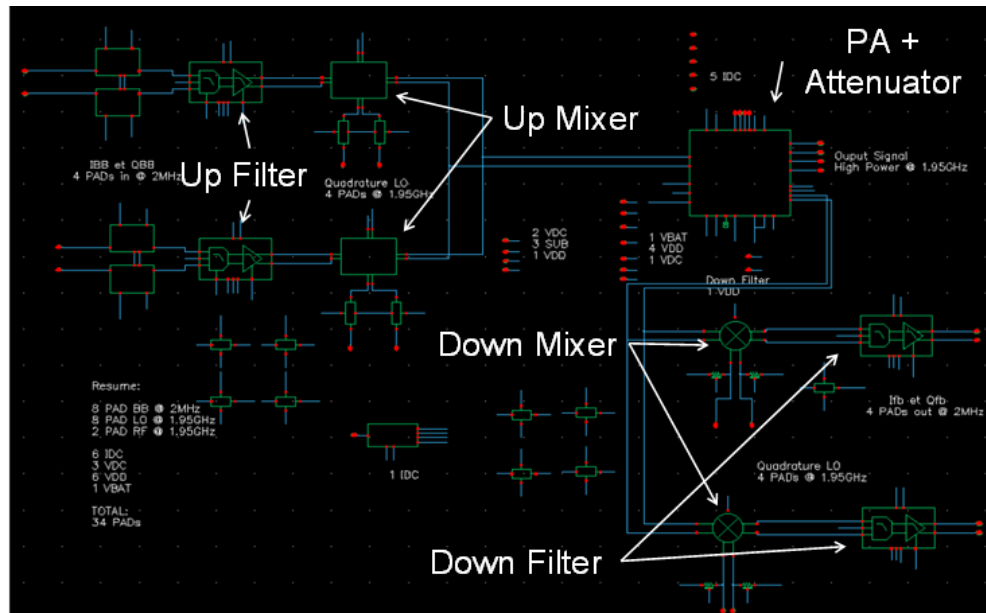


Figure 3.29: Whole Schematic of the Analog Transmitter



The layout of the analog transmitter is presented in figure 3.30. The IC solution includes the PAD ring with ESD (ElectroStatic Discharge) protection. As a precaution, we split the voltage supply and the ground according to the frequency domain of the circuit. For example, the base-band feedback and direct path have separate supplies to reduce the voltage drop. By the same way, the supply and the ground of the power amplifier are separated from the others to insure an accurate voltage or current going in the main component of the chip.

The total analog transmitter die area is equal to  $2 \times 2 \text{ mm}^2$  for a DC/RF consumption of 480/765mW respectively.

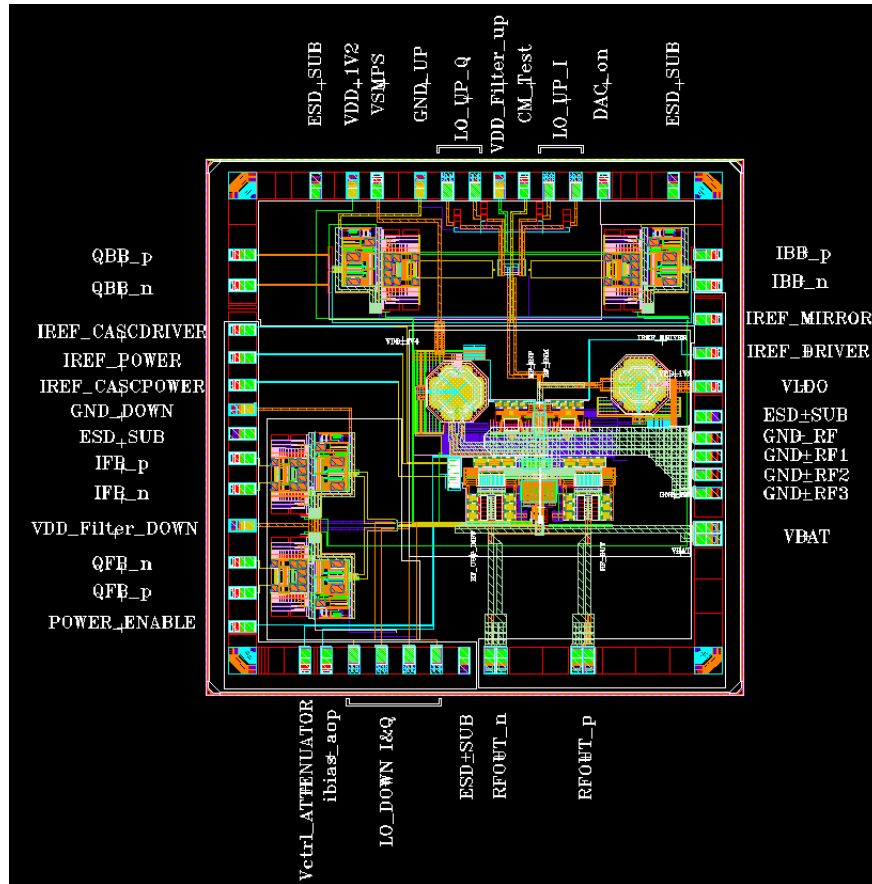


Figure 3.30: Die Area of the Analog Transmitter

Table 3.7 and 3.8 summarize the die area and the DC/RF consumption of each component, making the linearization of a transmitter for the third generation of mobile communications.

COMPONENTS	DC consumption	RF consumption @ $P_{out} = 23dBm$
Up Filter	20.5mW	20.5mW
Up Mixer	63.3mW	67.8mW
Power Amplifier	370mW	658mW
TOTAL	453.8mW	746.3mW

Table 3.7: Consumption of each building block of the direct path

COMPONENTS	DC consumption	RF consumption @ $P_{out} = 23dBm$
Attenuator	1.25mW	1.25mW
Down Mixer	-	-
Down Filter	24.9mW	24.9mW
TOTAL	26.15mW	26.15mW

Table 3.8: Consumption of each building block of the feedback path

The figure 3.31 shows the photography of the realized IC and can be compared to the layout shown in figure 3.30.

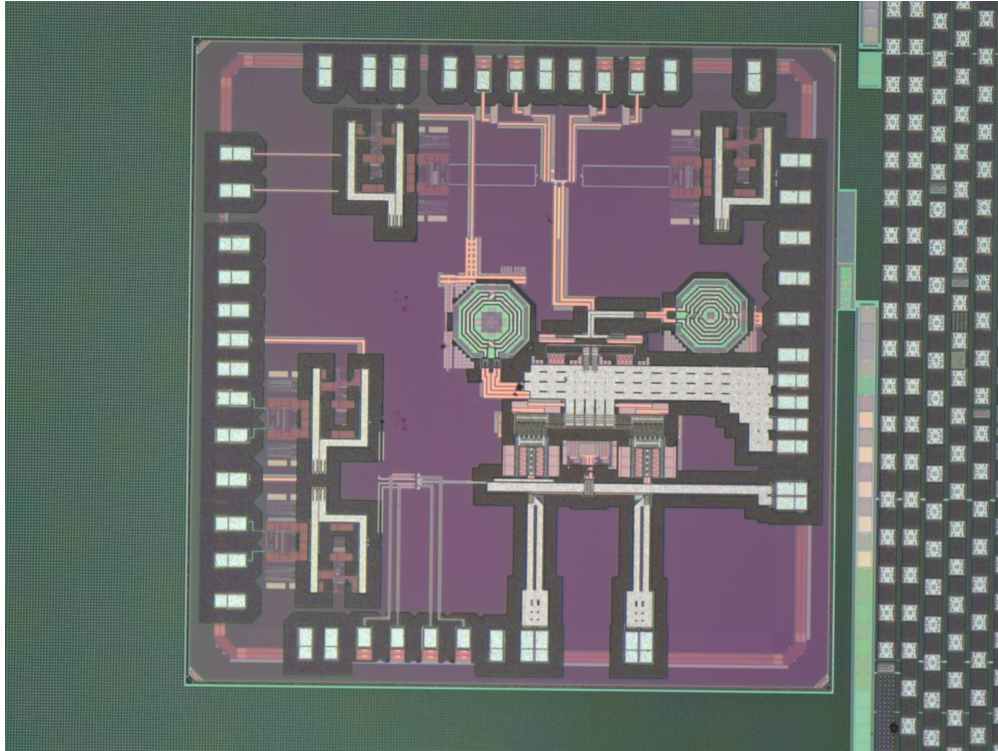


Figure 3.31: Photography of the IC Cartesian Feedback

### 3.3 Linearization Results

All RF systems have to pass through several steps before their integration on cell phones as a finalized product. In this part, simulation of the whole system will be demonstrated with digital, analog and mixed simulations.

First, the architecture of the transmitter has to be determined according to the wireless standard targeted (cf. Chapter 2) which is, in our case, the W-CDMA used for the third generation of wireless communication. Taking in consideration the architecture adopted, the Cartesian Feedback technique has been chosen according to the criteria as the consumption, the size of the chip and the complexity of the solution which has to be integrated in CMOS 65nm technology.

Every transmitter has to be studied on a system level before beginning the design of each analog component. This step remains very important as all further decisions will result from it. The system level study has for objective to specify characteristics of each building block of the transmitter, in our case, the Cartesian Feedback.

Once those parameters of each component are chosen, the design on the transistor level can be performed respecting specifications given by the previous study. In the same time, the layout of each analog circuit is designed. The final step is to make the entire layout of the transmitter with the PAD ring associated in order to measure it after.

This part will illustrate linearization results of the circuit from simulation behavior using ADS software; to the transistors level using RFDE or GoldenGate software. To validate our system, which includes a digital and an analog part, simulations using SystemVue (with ModelSim) and ADS (via Dynamic Link) have been made showing the improvement given by the mixed IC Cartesian Feedback.

#### 3.3.1 Methodology of Co-Simulations Experiments

To complete the study, co-simulation between the analog and the digital parts has been performed. Several Agilent's softwares were used in order to get complete results. SystemVue and ADS were used for the digital part and the analog part, respectively.

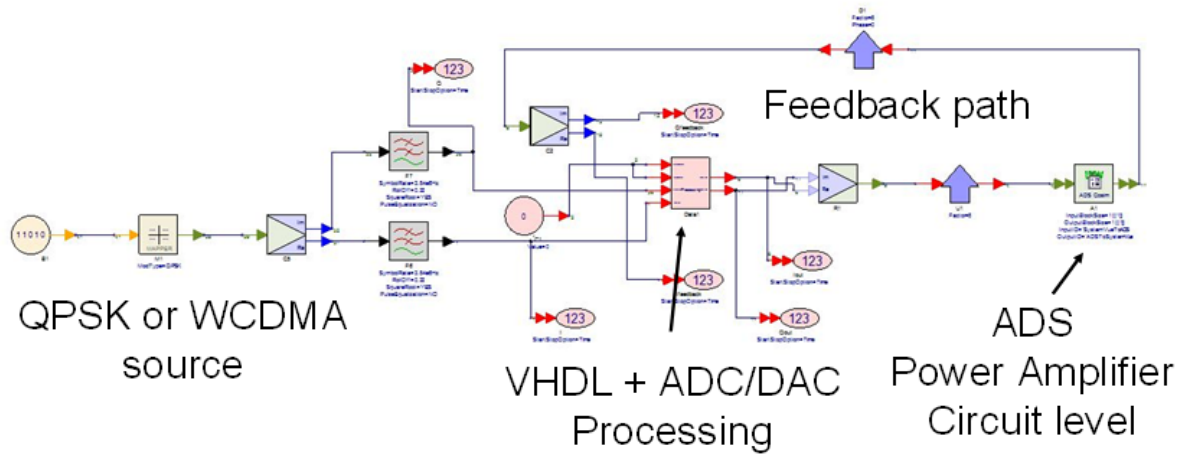


Figure 3.32: Whole Co-Simulation Overview

The whole simulation is depicted in figure 3.32 (co-simulation). The simulation is led by SystemVue according to results coming from ModelSim (for the VHDL code) and ADS (Ptolemy + AnalogRF for the entire circuit).

Focusing on the digital part, the objective was to validate the behavior of the VHDL code with a modulated signal as the W-CDMA Standard. The figure 3.33 shows the simulation of the digital part using SystemVue from Agilent.

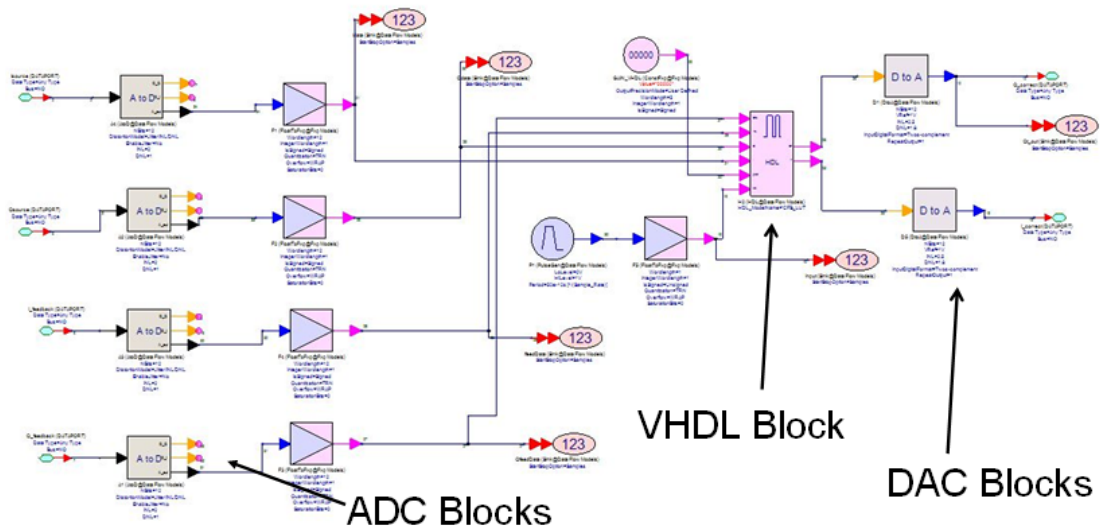


Figure 3.33: Simulation's Bench of the Digital Part

The simulation of the digital part allows to observe the behavior of the VHDL code. From this simulation, we were able to interpret results and some limitation of the code as it is represented in figure 3.34. The right figure shows that the correction is effective when the delay is less than 10% to the period of the signal, but has issue to make the correction in the other case (Cf. left figure).

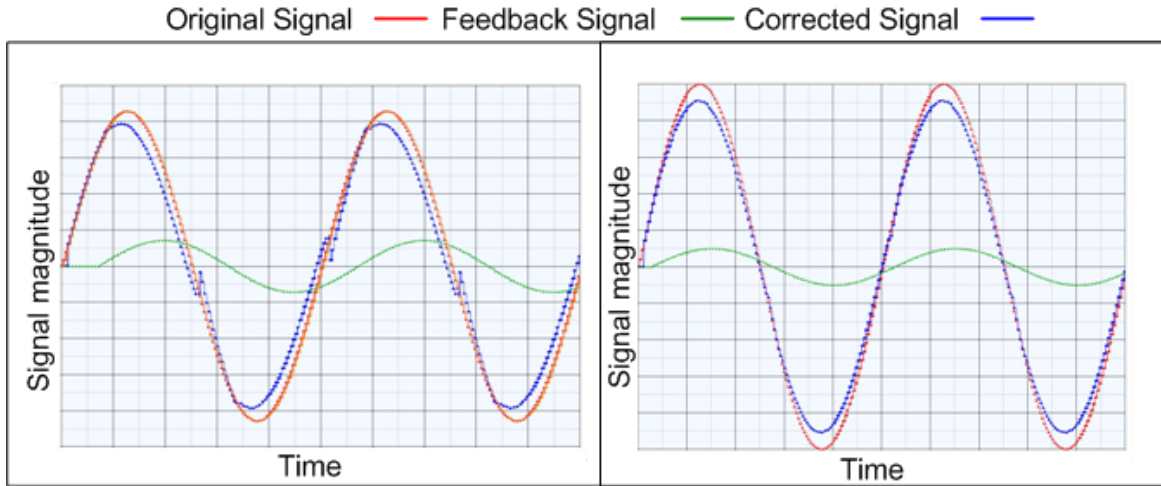


Figure 3.34: Digital Signals Simulated with SystemVue

This digital simulation is composed of converters (analog-to-digital and digital-to-analog) and a VHDL block. In order to reduce the simulation time, converters blocks with parameters taken from measurements of them are used.

The VHDL code is implemented inside the VHDL block. It works with the simulator called ModelSim (simulator specialized in digital simulation).

On the left in figure 3.34, original and feedback signals are coming, whereas corrected signals are coming out from the VHDL block (right side).

The analog part, as it is shown in figure 3.35, is simulated by ADS from Agilent. Two domains (analog and digital) exist for this new co-simulation. The digital part is working with Ptolemy, making the interface between the circuit realized on CMOS and the digital part.

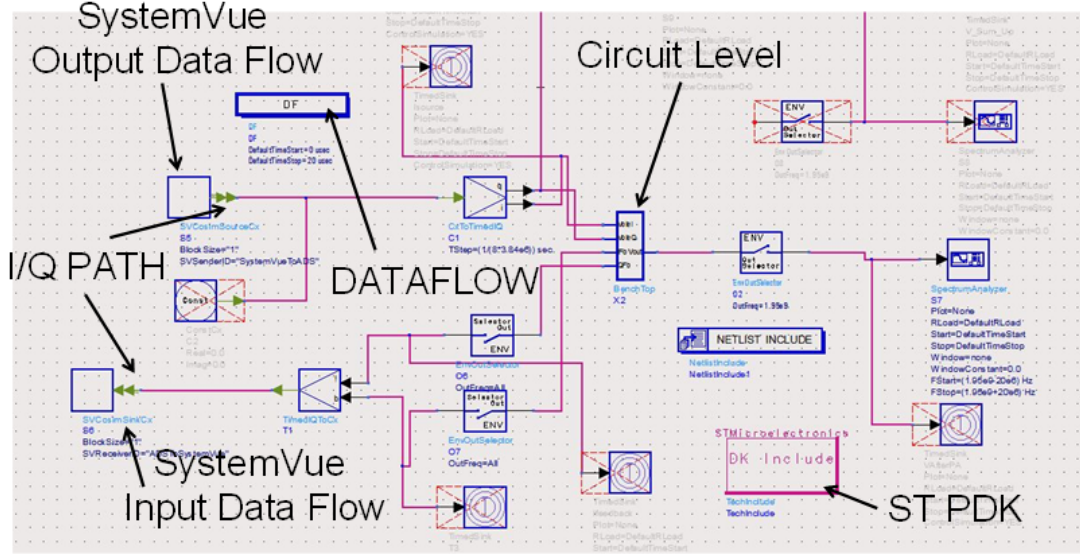


Figure 3.35: Simulation's Bench of the Analog Part

The circuit level part is the same that was presented in the Part 3.2.8 - Circuit Assembly. The circuit includes all components and the PAD ring associated. As a consequence, the simulation of the entire circuit will show the whole behavior of the system.

### 3.3.2 Co-simulation results

From the methodology of simulations previously described, several simulations can be performed in order to evaluate the whole circuit. We propose to show system level simulations results from different metrics presented on the second chapter as the spectrum, ACLR before and after linearization, or with and without linearization for the EVM.

The figure 3.36 presents the spectrum from the input and the output of the power amplifier without linearization.



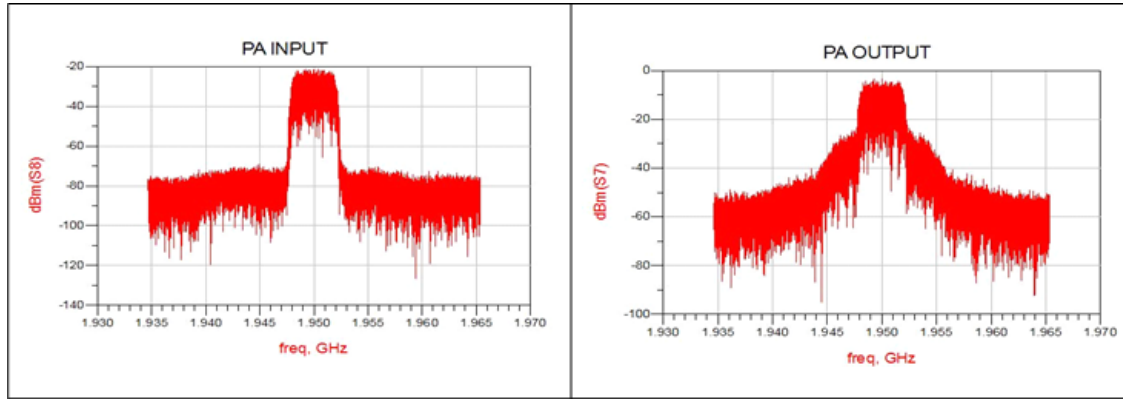


Figure 3.36: Input and Output Spectrum without Linearization

As illustrated above (Cf. Chapter 1.3: Nonlinear Behavior of a RF Transmitter) the output spectrum of the power amplifier working in the non-linear region shows the appearance of non-desired power on the adjacent channels. The goal of the simulations is to reduce the power on the adjacent channels and can be observed in figure 3.37. The figure on the left shows the phenomena of signal pre-distorted at the input of the power amplifier.

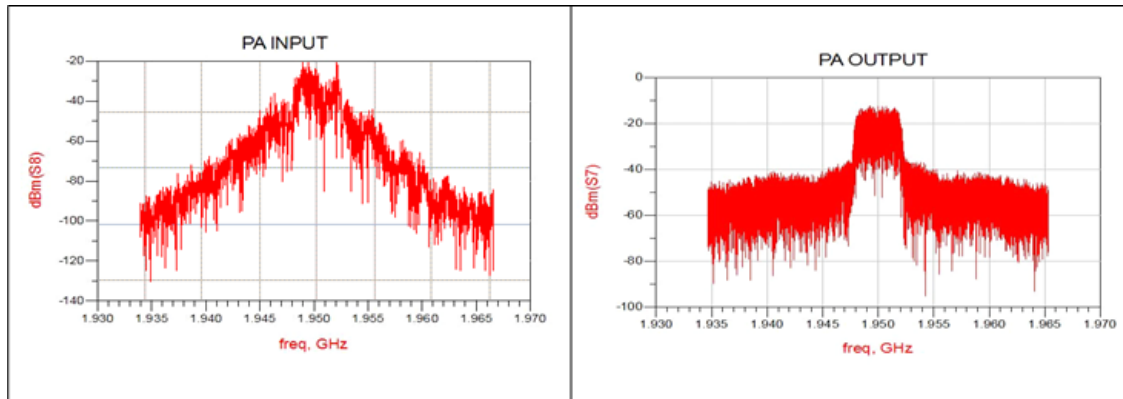


Figure 3.37: Input and Output Spectrum with Linearization

From this spectrum, it looks difficult to provide a metric to this linearity's improvement, so we propose to show the decreasing of the power on the adjacent channels in a different way. The figure 3.38 illustrates the ACLR level according to the output power of the power amplifier.

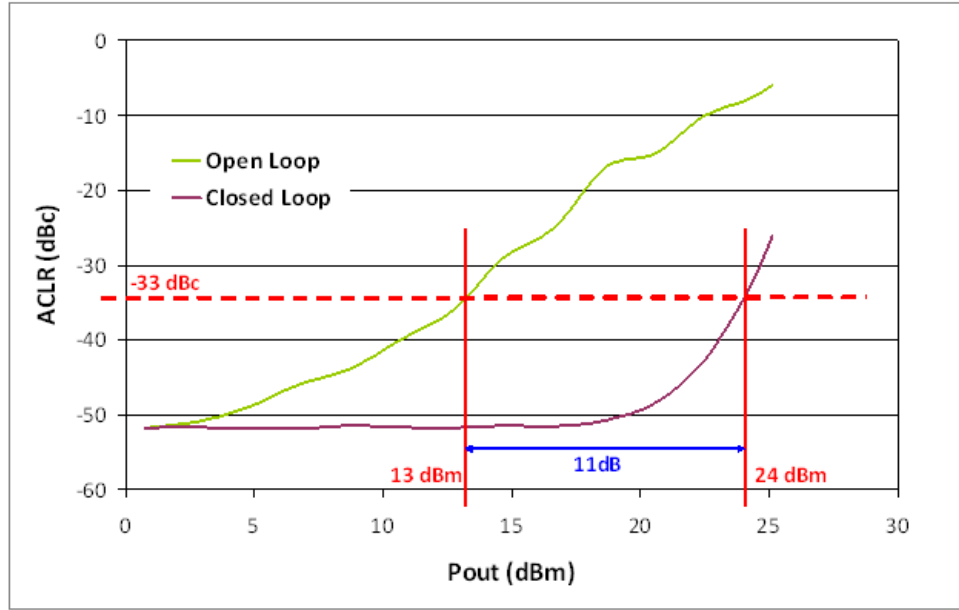


Figure 3.38: ACLR Improvement with and without Cartesian Feedback

The open and closed loops are using the whole circuit designed in CMOS technology. The high level of the ACLR of the open loop can be explained from equation of the second chapter (IMD<sub>x</sub>). In fact the non-linearity of the up-converter added with the non-linearity of the power amplifier gives a supplementary effect that the designer has to take into account for the design of a stand-alone transmitter. In our case, it is not necessary to consider the non-linearity behavior of the direct path because the whole system (more accurately the feedback path) will correct these non-linearities. From this simulation, the most important information can be seen as the improvement of the ACLR according to a fixed output power. But also, it can be the increasing output power that the system can reach staying in conformity with the targeted standard. In our case, if we fix the simulation to the specifications (ACLR @5MHz from the carrier = -33dBc) given by the standard, the output power of the chip is going from 13dBm to 24dBm. Let's remember that the transmitter is highly non-linear to prove our theory explained on the previous parts.

Another result of the pre-distortion of signals can be elucidated with the constellation at the input and output of the power amplifier. This phenomenon is presented in figure 3.39 and has been simulated with Agilent tools.



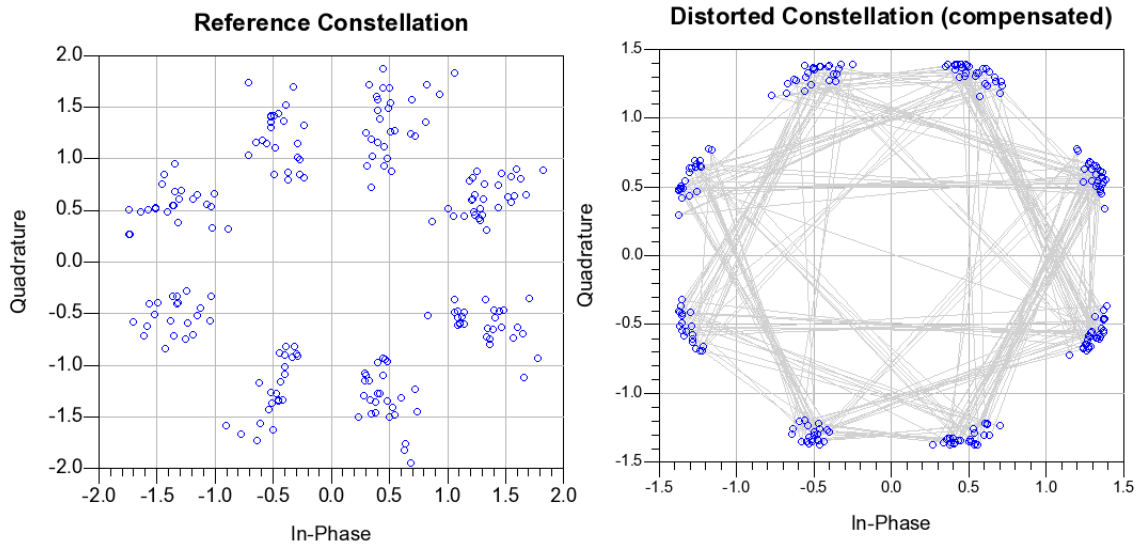


Figure 3.39: Input and Output Constellations of the Power Amplifier

The constellation on the left figure shows the non-linearity of the signal at the input of the transmitter, whereas the constellation on the right figure shows that the non-linearity of the signals added to the non-linearity given by the direct path of the transmitter. These latter are compensated and resulted as a linear signal or spectrum.

### 3.4 Comparison with the State-of-the-Art

In order to conclude the study of our transmitter, a comparison can be performed taking into account some precautions about the technology, the standard targeted, and the level of integration of the system on Silicon.

All circuits presented in this part are only considering the RF part of the transmitter (without the frequency synthesis).

Reference	[66]	[38]	[67]	[33]	This Work
Technology	LDMOSFET	Mini-circuit	HSB0.3 $\mu$ m	CMOS 65nm	CMOS 65nm
PA Class operation	AB	A	AB	PPA	AB
Frequency (GHz)	2.14	0.9	1.95	1.95	1.95
Standard / Modulation	W-CDMA	16-QAM	W-CDMA	W-CDMA	W-CDMA
ACLR improvement	15.3dB	10dB	24dB	27dB	35dB
Consumption	NC	NC	NC	45mW (w/o PA)	772mW (114mW w/o PA)
Die Area	NC	NC	2.9x1.5 mm <sup>2</sup>	0.53mm <sup>2</sup> (w/o PA)	2 x 2 mm <sup>2</sup>

Table 3.9: Comparison with the Literature

It is important to note that it seems hard to find the perfect literature for a realistic comparison with our system. In fact, some publications are enunciated for the theoretical comparison and others for the consumption and the die area.

Table 3.9 presents a comparison of our system with a recent state-of-the-art. As we can see, references [66] and [38] are also using Cartesian Feedback correction. For the same PA class operation and the same standard, the ACLR improvement of our system is 7dB better than the reference [66]. Moreover, our work is a single circuit that offers a compact and cheaper system than other circuits earlier mentioned.

Finally to evaluate the die area and the consumption of our whole solution, we decided to compare it to [67] and [33]. The latter reference is a theoretical study based on prevision for each block contrary to our system which is an effective layout. We can see that our circuit offers a greater ACLR improvement for a similar die area than the reference [67].



# Conclusion

This study presents the integration on Silicon of an entire linearized transmitter using a Cartesian Feedback for the W-CDMA standard.

In the second chapter, architectures of transmitter and different techniques to improve the linearity or the efficiency have been presented. Moreover, metrics to evaluate transmitters has been explained in frequency and time domain. An architecture with a linearization technique has been chosen according the standard specifications and the reducing of cost area and consumption that we imposed.

In the third chapter, theoretical study has been illustrated according linearity and noise behavior making the transmitter unstable. Different digital architectures have been proposed and evaluated in the digital domain with theoretical results. Simulations performed with ADS have been shown in order to determine the value of each component making the final transmitter.

In the fourth chapter, each building block has been explained from the digital domain for the signals corrections, and to the analog domain for the direct and feedback path. Then we add the final circuit with PAD ring to make whole simulations in order to evaluate our system. Thanks to Agilent software, we were able to realize these simulations taking into account digital and analog specifications.

From theses simulations results, theoretical comparison has been made in order to appreciate the linearity improvement of the Cartesian Feedback. Each analog building block have been designed and simulated in CMOS 65nm technology. Whole simulations have been realized from these simulations and/or measurements results, like ADC and DAC components. The die area and the consumption are not presented on the comparison because it remains hard to find an integrated circuit able to linearize the entire transmitter.

Finally, a patented architecture will be detailed in order to make a single chip addressing a lot of standards, mainly for the mobile communications. This solution proposes to adapt the current Cartesian Feedback with few changes to linearize the whole transmitter. The new architecture has for advantages to relax some constraints of the components from the direct

path. For example, the designer will focus on the maximum output power and the bandwidth of its components letting the Cartesian Feedback dealing with the linearity.

## Perspectives: Multi-Standard Solution using the Cartesian Feedback Technique

Nowadays one of the main concerns on mobile firm is to integrate always more function in the same chip. This concept is very attractive for a lot of companies and, in account of that, gives us the idea to adapt our system using the Cartesian Feedback for one standard to multi-standard applications. The proposed solution, which will be presented, has been patented in collaboration between the laboratory IMS and STMicroelectronics.

### Context

The solution previously presented works perfectly for the integration on a system for a mono-standard application. However, mobile phone regarding its localization can change the standard used. For example, in a city it seems easy to found W-CDMA standard but in the countryside the preferred standard is the GSM. The common solution found in the literature is to duplicate transceiver for each standard that the mobile can address (Cf. figure 3.40). This involves an over consumption in Silicon and by consequence a cost of production higher.

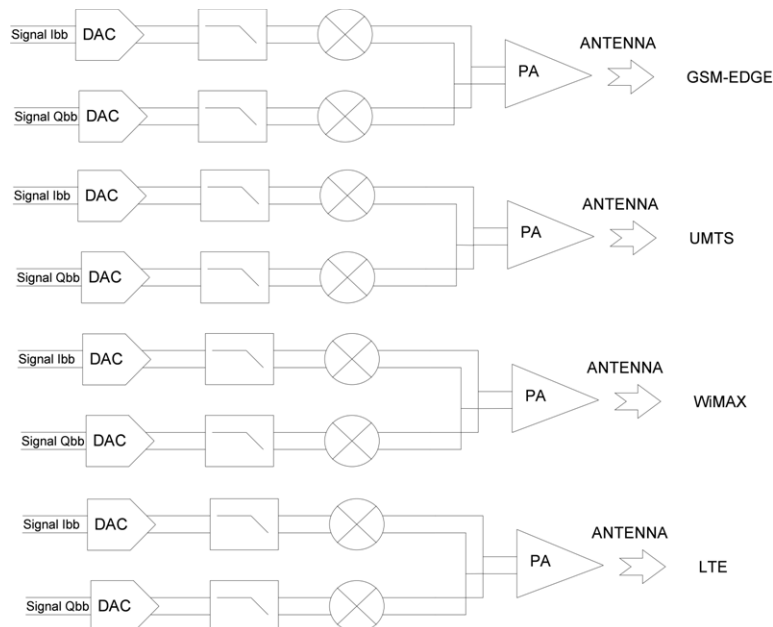


Figure 3.40: Parallel Transmitter Integrated on a Mobile

The idea is thus to apply our correction system of non-linearities for multi-standard applications.

Standards	Frequency (MHz)	Bandwidth	$P_{out}Max$	Data rate (bps)
GSM / GPRS	DCS1800: 1710 - 1785	200kHz	33dBm	9.6k
	PCS1900: 1850 - 1910			21.4 - 171.2k
UMTS	Band I: 1920 - 1980	5MHz	27dBm	144k - 2M
WiMAX	3500 (Europe)	3.5 - 10MHz	21dBm	70M
LTE	2570 - 2620	1.25 - 20MHz	27.5dBm	100M

Table 3.10: Specifications of Main Communications Standard

As a reminder, the table 3.10 shows the standard more usually used for integrated wireless system (or further integrated as LTE which is the next generation of communication standard also called 3.9G). These different standards were designed in order to meet an increasing demand of consumers involving higher data speeds.

### Proposed Solution

The solution to overcome this overhead of Silicon is to integrate all of these standards in one single system, as it is presented in the figure 3.41.

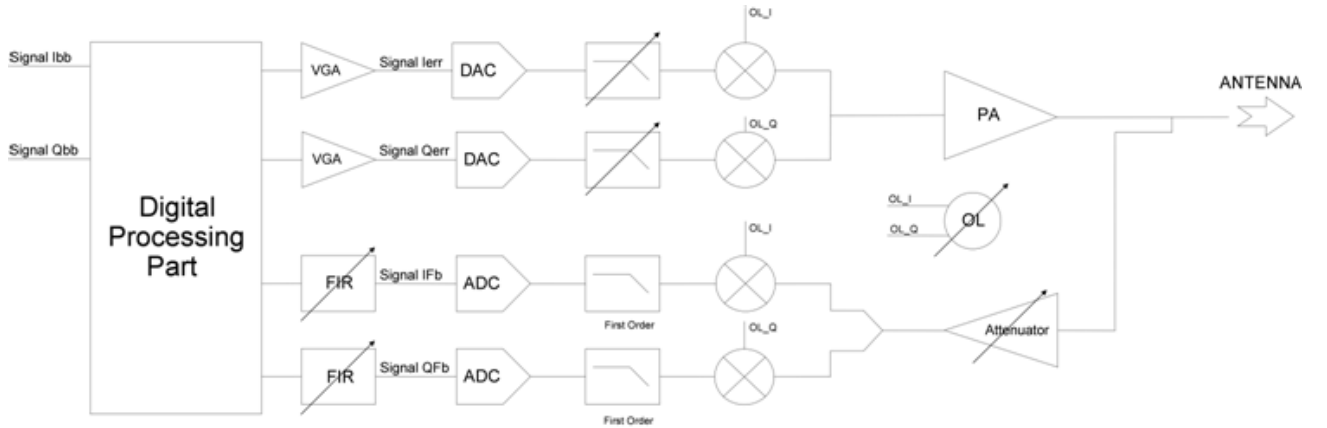


Figure 3.41: Architecture of the Multi-Standard Cartesian Feedback

From this new architecture we can notice that the main changes are on the filter of the direct and feedback path. In fact, to address different standards in a single chip, the system must be able to select channels of the wanted standard.

The functioning of this new architecture is the following:

- First the analog baseband filter of the direct path is replaced by an adaptive Gm-C filter [68]. The goal of this filter is the same than before expecting that the cut-off frequency can be change to select the right channel.
- Secondly the analog filter of the feedback path placed before the ADC has relaxed constraints regarding its order cut-off frequency. It makes a first filtering of the useful signal. Moreover, the filter must have a large cut-off frequency in order to pass the useful channel with its associated adjacent channel of the larger bandwidth standard. The adjacent channel should certainly not be cancelled because as it contains information of non-linearity of the transmitted signals.
- Finally the figure 3.42 shows the principle functioning of FIR (Finite Impulse Response) filter, which is the main components of this new architecture. This figure illustrates the extrema in terms of channels bandwidth that the system must deal with, i.e. GSM and LTE reference to the table 3.10. The FIR filter, thanks to its strict cut-off frequency, selects perfectly the desired channel and its adjacent channels providing us to correct distortions of the transmitter. The resulting spectrum will then be subtracted to the original signal in order to come back in the classic functioning of the Cartesian Feedback. The use of this filter allows us to remove a lot of constraints of the analog filter which is a higher order of filtering.

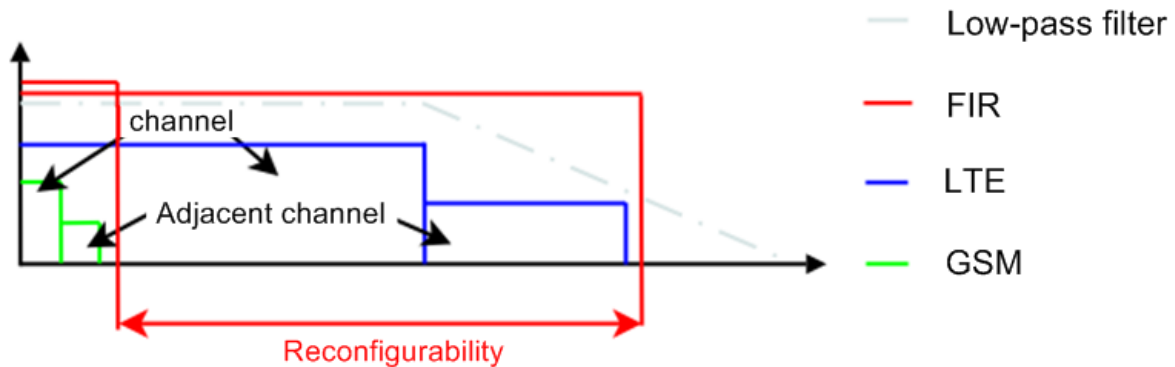


Figure 3.42: Principle functioning FIR on the new architecture

### Precautions and Specifications

In order to take advantages of the new architecture, we need to define specifications to the system:

- First the local oscillator has to be able to cover all RF frequency bands defined by all standards.
- A specific design has to be realized on mixers and the power amplifier to increase their bandwidth. Typically, these components have a variation of the gain on a frequency band. This variation can make the system unstable, so the designer has to check that mixers and power

amplifier have a minimum variation of their gain in the bandwidth of all targeted standard.

- In a context of multi-standard application, the frequency band used is important for the functioning of the Cartesian Feedback. This is not a problem for the feedback components, excluding the ADC, because they have been chosen to be the most linear as possible. In fact, as the components making of the feedback path are passive components, their bandwidths are large.





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