1

Characterization of Wireless Transmitter Distortions

1.1 Introduction

Wireless transmitters designed for modern communication systems are expected to handle wideband amplitude and phase modulated signals with three major performance metrics: linearity, bandwidth, and power efficiency. First, linearity requires the minimization of distortions mainly caused by the transmitter's radio frequency (RF) analog circuitry in order to preserve the quality of the transmitted signal and avoid any loss of information during the transmission process. Second, bandwidth is critical for multi-carrier and multi-band communication systems. Moreover, wider bandwidths are needed to accommodate higher data rates. Third, power efficiency is an important consideration that affects the deployment and operating costs of communication infrastructure as well as environmental impact.

In general, distortions refer to the alteration of the signal due to the imperfections of the transmitter's hardware. Distortions observed in wireless transmitters have various origins such as frequency response distortions, harmonic distortions, amplitude and phase distortions, and group delay distortions, in addition to modulator impairments (including direct current (DC) offset, gain, and phase imbalance), and so on. Among these distortions, the predominant ones are those due to the nonlinearity present in the transmitter's RF front end and mainly the RF power amplifier (PA). Indeed, wireless transmitters are made of a cascade of several stages including digital-to-analog conversion, modulation, frequency up-conversion, filtering, and amplification as illustrated in Figure 1.1. Among these subsystems, the PA is identified as the major source of nonlinear distortions. Thus, modeling and compensating for the transmitter nonlinear distortions is often trimmed down to the modeling and compensation of the PA's nonlinearity.

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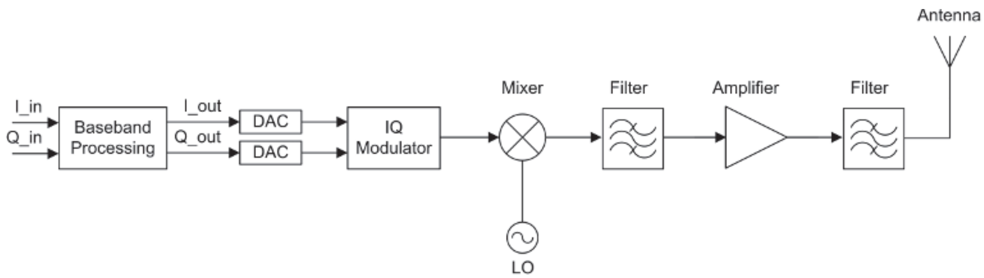

Figure 1.1 Simplified block diagram of a typical wireless transmitter

In the remainder of this chapter, the nonlinearity of RF PAs will be described and major metrics used to quantify nonlinear distortions will be presented.

1.1.1 RF Power Amplifier Nonlinearity

The nonlinearity of the PA depends mainly on its class of operation and topology. Classes of operation include the linear class A, the mildly nonlinear class AB, as well as highly nonlinear classes such as C, D, and E. The topology refers to whether the power amplification system is built using single-ended amplifiers or more advanced architectures such as Doherty, linear amplification using nonlinear components (LINC), envelope tracking, and so on. The design of power amplification systems is always subject to the unavoidable antagonism between linearity and power efficiency [1]. The objective is to design a power amplification system, or more generally, a transmitter that meets the linearity requirements with the highest possible power efficiency. The approach often consists of maximizing the power efficiency of the amplification stage while maintaining its distortions to a reasonable amount that can be compensated for at the system level using linearization techniques such as feedforward or predistortion [2]. Figure 1.2 shows the measured gain and power efficiency of a Gallium Nitride (GaN) based Doherty PA driven by a four-carrier wideband code division multiple access (WCDMA) signal and operating around a carrier frequency of 2140 MHz. This figure clearly illustrates the power efficiency versus linearity dilemma as low power efficiency is observed for low input power levels when the amplifier is operating in its linear region where the gain is constant. Conversely, higher power efficiency is obtained for large input power levels that drive the amplifier into its nonlinear region.

1.1.2 Inter-Modulation Distortion and Spectrum Regrowth

Transmitters' nonlinearity causes the appearance of unwanted frequency components at the output of the transmitter. To better understand the effects of the transmitter's nonlinearity on the transmitted signal, the case of a two-tone signal passing through

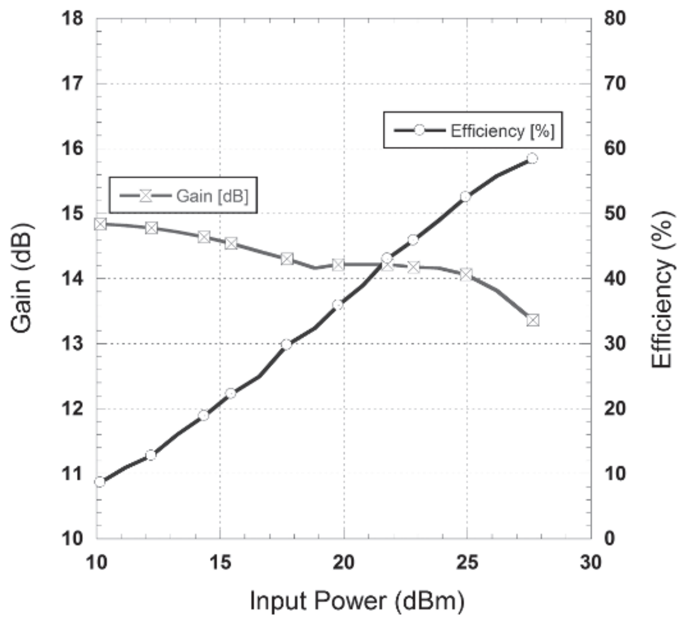

Figure 1.2 Gain and power efficiency characteristics of a power amplifier prototype

a third order memoryless nonlinear systems is considered in the example next. In this case:

- The transmitter's nonlinearity is modeled by a third order polynomial function according to the following equation:

$$x_{out\_Transmitter}(t) = a \cdot x_{in\_Transmitter}(t) + b \cdot x^2_{in\_Transmitter}(t) + c \cdot x^3_{in\_Transmitter}(t) \quad (1.1)$$

  where $x_{in\_Transmitter}$ and $x_{out\_Transmitter}$ are the time domain waveforms at the input and the output of the transmitter, respectively. a, b, and c are the model coefficients.
- The input signal $x_{in\_Transmitter}$ is a two-tone signal given by:

$$x_{in\_Transmitter}(t) = A_1 \cdot \cos(\omega_1 t) + A_2 \cdot \cos(\omega_2 t) \quad (1.2)$$

  where A_1 and A_2 are the magnitudes of each of the two tones, and ω_1 and ω_2 are their angular frequencies with $\omega_2 > \omega_1$.

By combining Equations 1.1 and 1.2, the transmitter's output for the two-tone input signal can be expressed as:

$$x_{out\_Transmitter}(t)$$
$$= \left[\frac{1}{2}bA_1^2 + \frac{1}{2}bA_2^2\right]$$

$$
\begin{aligned}
&+ \left[\left(a+\frac{3}{4}cA_1^2+\frac{3}{4}cA_2^2\right)\cdot A_1\cdot\cos(\omega_1 t)+\left(a+\frac{3}{4}cA_2^2+\frac{3}{4}cA_1^2\right)\cdot A_2\cdot\cos(\omega_2 t)\right]\\
&+ \left[\frac{1}{2}bA_1^2\cos(2\omega_1 t)+\frac{1}{2}bA_2^2\cos(2\omega_2 t)\right]+\left[\frac{1}{4}cA_1^3\cos(3\omega_1 t)+\frac{1}{4}cA_2^3\cos(3\omega_2 t)\right]\\
&+ [bA_1A_2\cos((\omega_2-\omega_1)t)+bA_1A_2\cos((\omega_2+\omega_1)t)]\\
&+ \left[\frac{3}{4}cA_1^2A_2\cos((2\omega_1-\omega_2)t)+\frac{3}{4}cA_1A_2^2\cos((2\omega_2-\omega_1)t)\right]\\
&+ \left[\frac{3}{4}cA_1^2A_2\cos((2\omega_1+\omega_2)t)+\frac{3}{4}cA_1A_2^2\cos((2\omega_2+\omega_1)t)\right] \qquad (1.3)
\end{aligned}
$$

To clearly separate the various frequency components present in the transmitter's output signal, Equation 1.3 can be re-arranged as:

$$
\begin{aligned}
x_{out\_Transmitter}(t) &= \{a\cdot[A_1\cdot\cos(\omega_1 t)+A_2\cdot\cos(\omega_2 t)]\}\\
&+ \left\{\left[\left(\frac{3}{4}cA_1^2+\frac{3}{4}cA_2^2\right)\cdot A_1\cdot\cos(\omega_1 t)\right]+\left[\left(\frac{3}{4}cA_2^2+\frac{3}{4}cA_1^2\right)\cdot A_2\cdot\cos(\omega_2 t)\right]\right\}\\
&+ \left\{\left[\frac{3}{4}cA_1^2A_2\cos((2\omega_1-\omega_2)t)\right]+\left[\frac{3}{4}cA_1A_2^2\cos((2\omega_2-\omega_1)t)\right]\right\}\\
&+ \left\{\left[\frac{1}{2}bA_1^2+\frac{1}{2}bA_2^2\right]+[bA_1A_2\cos((\omega_2-\omega_1)t)]\right\}\\
&+ \left\{\frac{1}{2}b\left[A_1^2\cos(2\omega_1 t)+A_2^2\cos(2\omega_2 t)\right]+[bA_1A_2\cos((\omega_2+\omega_1)t)]\right\}\\
&+ \left\{\frac{1}{4}c\left[A_1^3\cos(3\omega_1 t)+A_2^3\cos(3\omega_2 t)\right]\right.\\
&\left.+\frac{3}{4}cA_1A_2\left[A_1\cos((2\omega_1+\omega_2)t)+A_2\cos((2\omega_2+\omega_1)t)\right]\right\} \qquad (1.4)
\end{aligned}
$$

In this latter equation, the term between the first brackets ({}) in the right hand side represents the linearly amplified version of the input signal, while the second term corresponds to the distortions introduced by the transmitter's nonlinearity at the fundamental frequencies (these are the same as the input signal's frequencies). The remaining terms describe the mixing and harmonic frequency products that either fall in the close vicinity of the useful signal and thus cannot be removed by filtering, or are away from the useful signal (around DC or the harmonics). The latter are less critical as they can be removed by filtering the transmitter's output signal. The frequency domain representation of the transmitter's input and output signals given by Equations 1.2 and 1.4 are illustrated in Figure 1.3.

The frequency components present at the output of the nonlinear transmitter driven by a two-tone input signal are summarized in Table 1.1. These can be categorized in three groups:

- *The useful signal:* comprised of the linearly amplified fundamental frequency components.

- *The unwanted signals that can be removed by filtering:* these include the DC components, the second and third order harmonics, second order inter-modulation distortions, as well as out-of-band third order inter-modulation distortions.
- *The unwanted signals that cannot be filtered:* this includes the distortions that appear at the same frequencies as the input signal, and in-band third order inter-modulation products that are too close to the fundamental components to be filtered. For higher order nonlinear systems, additional even order in-band inter-modulation products are observed in the close vicinity of the useful signal.

The analysis presented here can be generalized to an *N*th order nonlinear model of the transmitter. In such case, up to the *N*th order harmonics and *N*th order mixing products will be generated at the output of the nonlinear transmitter [3, 4].

The study of PA and transmitter nonlinearities using two-tone and multi-tone signals is commonly used for understanding the origins of inter-modulation distortions for signals having discrete frequency spectrum components and can be used to derive closed form expressions of these distortions under two-tone or multi-tone input signals [5, 6]. Such results can be extrapolated to predict the behavior of the nonlinear system when driven by communications and broadcasting signals having characteristics comparable to that of synthetic multi-tone signals. However, when practical communication signals are used, the input signal's spectrum is continuous

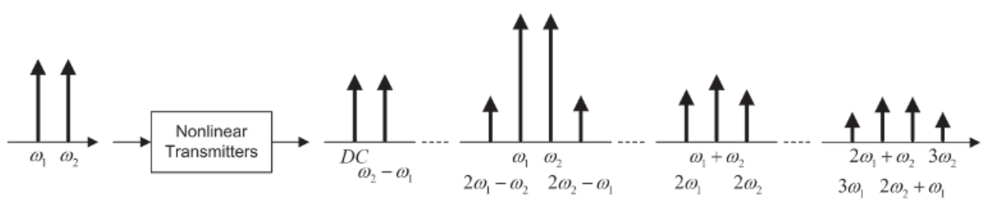

Figure 1.3 Frequency domain output of a nonlinear transmitter driven by a two-tone signal

Table 1.1 Frequency components at the output of a nonlinear transmitter for a two-tone input signal

Angular frequency	Designation
0	DC components
ω_1 and ω_2	Fundamental
$2\omega_1$ and $2\omega_2$	Second harmonics
$3\omega_1$ and $3\omega_2$	Third harmonics
$\omega_2-\omega_1$ and $\omega_2+\omega_1$	Second order inter-modulation products
$2\omega_1-\omega_2$ and $2\omega_2-\omega_1$	In-band third order inter-modulation products
$2\omega_1+\omega_2$ and $2\omega_2+\omega_1$	Out-of-band third order inter-modulation products

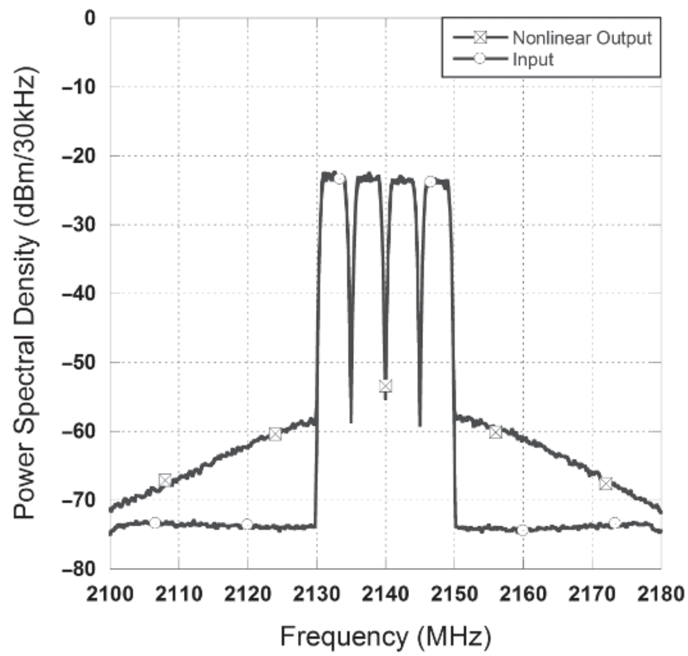

Figure 1.4 Output spectrum of a nonlinear transmitter driven by a multi-carrier WCDMA signal

and the inter-modulation distortions appear as a spectrum regrowth around the channel. Figure 1.4 presents the measured spectra at the output of a nonlinear transmitter driven by a four-carrier WCDMA signal having a total bandwidth of 20 MHz. This figure also reports the ideal output that would have been obtained if the transmitter were linear. This figure shows that there is significant spectrum regrowth that will create interferences with the adjacent channels. Such a transmitter does not meet the spectrum emission mask of the WCDMA standard and unavoidably requires linearization.

1.2 Impact of Distortions on Transmitter Performances

The nonlinearity of the PA depends on the input power level or equivalently on the input signal's amplitude. Thus, phase modulated signals having constant envelopes are not affected by the nonlinearity of the PA. Conversely, amplitude modulated signals are distorted by the nonlinearities. Almost all modern communication and broadcasting systems employ compact complex modulation schemes such as high order quadrature amplitude modulations (16QAM, 64QAM, etc.) and advanced multiplexing techniques, for example, orthogonal frequency division multiplexing (OFDM), and code division multiple access (CDMA), which result in amplitude modulated signals having strong envelope fluctuations. These signals are characterized by their peak-to-average power ratio (PAPR) that is given by:

$$PAPR_{\mathrm{dB}} = 10 \times \log_{10}\left(\frac{P_{\max,W}}{P_{avg,W}}\right) = P_{\max,\mathrm{dBm}} - P_{\mathrm{avg,dBm}} \tag{1.5}$$

where $PAPR_{\mathrm{dB}}$ is the signal's PAPR expressed in dB. $P_{\max,W}$ and $P_{avg,W}$ are the signal's maximum and average power levels expressed in watts, respectively. Similarly, $P_{\max,\mathrm{dBm}}$ and $P_{\mathrm{avg,dBm}}$ are the signal's maximum and average power levels expressed in dBm, respectively.

Typical PAPR values for modern communication systems are in the range of 10–13 dB. These can be reduced by several decibels using crest factor reduction (CFR) techniques [7–9]. The PAPR of the signal and its probability distribution functions are critical parameters that need to be considered when dealing with amplifier and transmitter nonlinearities. Indeed, to linearly amplify high PAPR signals without linearizing the amplifier, one must make sure that the maximum peak power of the input signal to be amplified remains within the linear region of the PA. This will impact the power efficiency of the system. To illustrate this concept of brute force linear amplification graphically, the gain and drain efficiency characteristics of a commercial PA are presented in Figure 1.5. In this figure, the gain and power efficiency are reported as a function of the output power back-off (OPBO) that is defined as:

$$OPBO_{\mathrm{dB}} = P_{\mathrm{out,dBm}} - P_{\mathrm{out,sat,dBm}} \tag{1.6}$$

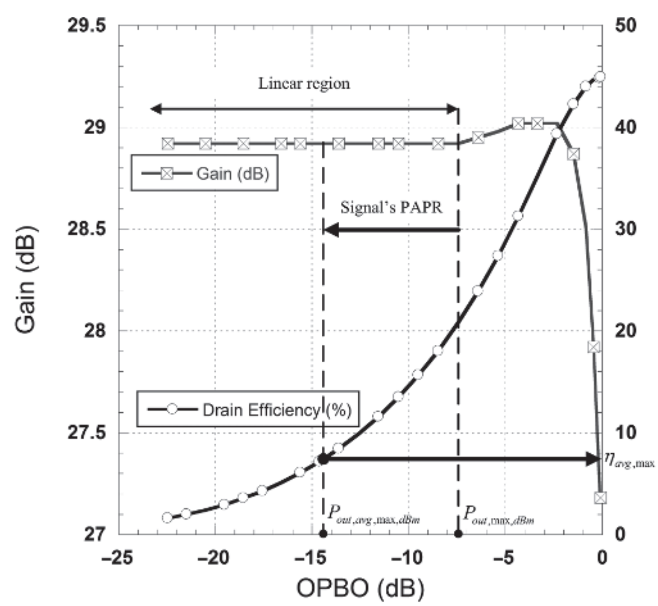

Figure 1.5 Gain and efficiency considerations in brute force linear amplification

where $OPBO_{dB}$ is the OPBO expressed in dB. $P_{out,dBm}$ and $P_{out,sat,dBm}$ refer to the dBm values of the amplifier's operating output power and the amplifier's output power at saturation, respectively.

According to the results of Figure 1.5, to ensure a linear behavior of the considered amplifier, the maximum peak output power ($P_{out,max,dBm}$) should not exceed −7 dB OPBO. Thus, the maximum average output power ($P_{out,avg,max,dBm}$) will be:

$$P_{out,avg,max,dBm} = P_{out,max,dBm} - PAPR_{dB} \tag{1.7}$$

In Equation 1.7, $PAPR_{dB}$ is the signal's PAPR expressed in dB. In this example, the signal's PAPR is assumed to be 7 dB.

Given this restriction on the maximum operating average power of the amplifier, the maximum average drain efficiency ($\eta_{avg,max}$) of the brute force linear amplifier will be less than 10%. This noticeably low power efficiency represents the maximum efficiency achievable from this amplifier if operated without a linearization technique. Conversely, if the same amplifier is used in conjunction with a linearization technique, for example, using digital predistortion (DPD), it will be able to operate linearly over its full output power range up to saturation. As graphically illustrated in Figure 1.6, the maximum average output power of the amplifier will be higher, which enables increased power efficiency. This example shows that by using linearization techniques, the maximum efficiency of the amplifier can be raised from 8 to 23%, which represents a substantial gain in power efficiency. It is worth mentioning that the amplifier used in this graphical analysis is optimized for linearity. Though, if a

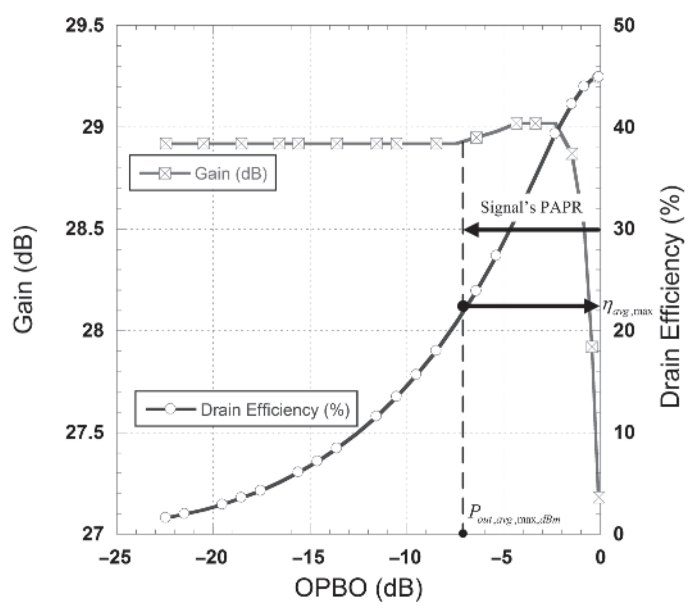

Figure 1.6 Gain and efficiency considerations in linearized power amplifiers

power efficient amplifier prototype is considered, a more important efficiency gain can be obtained with operating efficiencies of the PA in the range of 50%.

This brief discussion clearly shows the impact of distortions on the system efficiency as they constrain the brute force amplifier to work with large back-off levels to guarantee linear amplification. It also highlights the significant power efficiency improvement that can be obtained by using a DPD technique.

The cascade of the PA and the digital predistorter will behave as a linear amplification system whose gain can be set by controlling the small signal gain of the predistorter. It is a common misconception to think that the choice of the small signal gain of the predistortion and thus the gain of the linearized amplifier will influence the power efficiency performance of the linearized amplifier. Indeed, when seen as a function of the output power, the drain efficiency of the linearized amplifier will remain quasi unchanged [10]. The impact of the gain normalization on the DPD performance will be thoroughly discussed in Chapter 9.

These efficiency figures do not take into account the energy consumption of the linearization circuitry. Typical DPD circuitry has a power consumption in the range of a few watts. This power consumption needs to be taken into consideration when calculating the overall efficiency of the linearized amplifier. Obviously, the use of the predistortion technique for efficiency/linearity trade-off enhancement is a viable solution only when the predistorter's power consumption does not compromise the overall efficiency of the linearized amplifiers. Accordingly, and as rule of thumb, DPD is practically employed for PAs with output power that exceeds 10 W: Though accurate calculations can be made to decide on the suitability of DPD to improve the system performance compared to the case of a brute force amplifier topology based on the amplifier's power capability, its efficiency, and the predistorter's power consumption.

1.3 Output Power versus Input Power Characteristic

The output power versus input power (P_{out} vs. P_{in}) characteristic is commonly used to characterize the transfer function of amplifiers. This characteristic relates the input power of the device under test (DUT) at the fundamental frequency to its output power at the same frequency. When both power levels are expressed in watts, the slope of the P_{out} vs. P_{in} characteristic represents the linear gain of the system. Most commonly, the power levels are expressed in dBm. In such case, the slope of the P_{out} vs. P_{in} characteristic is equal to unity and the gain in dB corresponds to the y-intercept point (i.e., the value of the output power for a 0 dB m input power).

In the absence of memory effects, the P_{out} vs. P_{in} characteristic appears as a one to one mapping function that increases linearly with the input power. As the amplifier is driven into its nonlinear region, a gain compression appears as the actual output power becomes lower than the linearly amplified version of the input power. The amount of compression introduced by the amplifier increases until it reaches the saturation power. Figure 1.7 presents a sample P_{out} vs. P_{in} characteristic of an amplifier

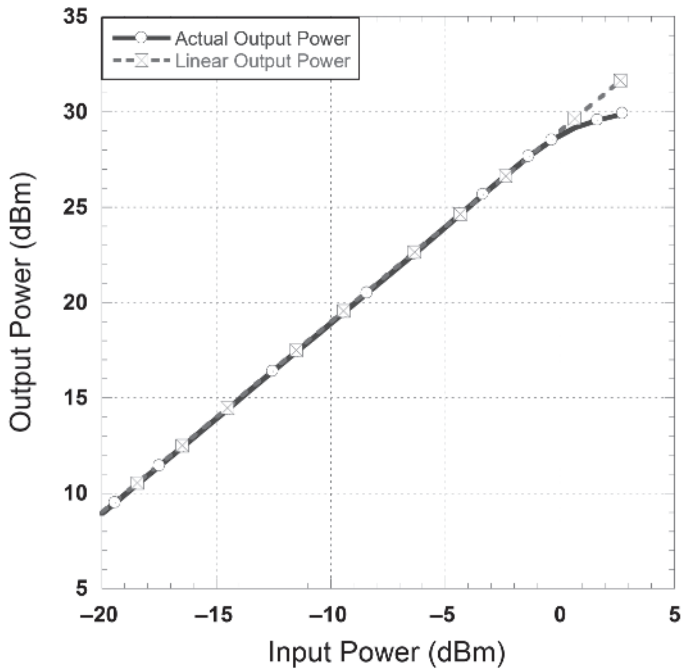

Figure 1.7 Sample output power versus input power characteristic

optimized for linearity. One can observe that the gain compression of the amplifier becomes noticeable only a few dBs before the saturation for input power levels beyond 0 dB m. With amplifiers optimized for efficiency, the gain compression is observed over a wider input power range starting from as early as 10 dB below the maximum input power.

The P_{out} vs. P_{in} characteristic is straightforward to derive as it only requires scalar measurements both at the input and output of the DUT. This can be performed using a network analyzer or a set up that comprises a signal generation instrument and a power measurement instrument such as a power meter or a spectrum analyzer. The P_{out} vs. P_{in} characteristic can be measured under a wide range of drive signals such as continuous wave (CW), multi-tone, or modulated signals.

1.4 AM/AM and AM/PM Characteristics

The P_{out} vs. P_{in} characteristic is a basic and incomplete means of characterizing nonlinear transmitters and PAs driven by modulated signals. Indeed, a more comprehensive representation that includes amplitude as well as phase information is needed. In the most general case, a dynamic nonlinear transmitter is fully described by a set of four characteristics, namely the amplitude modulation to amplitude modulation (AM/AM) characteristic, the amplitude modulation to phase modulation (AM/PM) characteristic, the phase modulation to phase modulation (PM/PM) characteristic, and the phase modulation to amplitude modulation (PM/AM) characteristic. PA distortions are amplitude dependant and phase modulated signals (having constant amplitudes) are not affected by the PA distortions. Thus, PAs are mainly characterized by their AM/AM and AM/PM characteristics. Conversely, transmitters might exhibit PM/AM and PM/PM distortions that are mainly due to the gain and phase imbalances in the frequency up-conversion stage and/or when the transmitter has a non-flat frequency response over a bandwidth equal to that of the input signal. Contrary to the AM/AM and AM/PM distortions generated by the unavoidably nonlinear behavior of the PA, the PM/AM and PM/PM distortions can be minimized by a careful design of the transmitter. So far, these have often been considered to have an insignificant impact on the performance of a behavioral model or a digital predistorter. With the adoption of multi-carriers and multi-band power amplification systems where the bandwidth of the signal to be transmitted is large enough to observe on a non-flat frequency response of the PA, the contribution of the PM/AM and PM/PM is becoming more significant and their inclusion in next generation behavioral models and predistorters is becoming inevitable.

Let's consider a DUT driven by a modulated input signal. x_{in} and x_{out} refer to the baseband complex waveforms corresponding to the DUT's input and output signals, respectively. The in-phase and quadrature components of the signals x_{in} and x_{out} are defined as:

$$\begin{cases} x_{in} = I_{in} + jQ_{in} \\ x_{out} = I_{out} + jQ_{out} \end{cases} \tag{1.8}$$

Under the assumption that this DUT, to be modeled or equivalently linearized, does not exhibit PM/AM and PM/PM distortions, its instantaneous complex gain, G, is solely a function of the input signal's magnitude and is given by:

$$G(|x_{in}|) = |G(|x_{in}|)| \cdot \underline{/G(|x_{in}|)} \tag{1.9}$$

where $|G(|x_{in}|)|$ and $\underline{/G(|x_{in}|)}$ represent the magnitude and phase of the instantaneous complex gain $G(|x_{in}|)$, respectively; and are expressed as a function of the input and output complex baseband waveforms according to:

$$|G(|x_{in}|)| = \frac{|x_{out}|^2}{|x_{in}|^2} = \frac{I_{out}^2 + Q_{out}^2}{I_{in}^2 + Q_{in}^2} \tag{1.10}$$

$$\underline{/G(|x_{in}|)} = \underline{/x_{out}} - \underline{/x_{in}} = \tan^{-1}\left(\frac{Q_{out}}{I_{out}}\right) - \tan^{-1}\left(\frac{Q_{in}}{I_{in}}\right) \tag{1.11}$$

The AM/AM characteristic of the DUT is obtained by plotting the magnitude of its instantaneous gain ($|G(|x_{in}|)|$), typically expressed in dB, as a function of the DUT's instantaneous input power. It is also possible, though less conventional to report the AM/AM characteristic as function of the DUT's output power. Similarly, the AM/PM

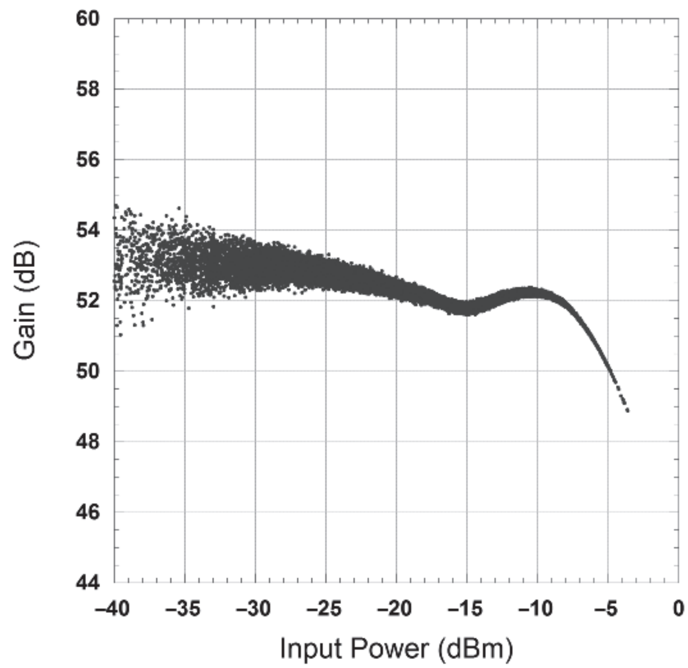

Figure 1.8 Sample AM/AM characteristic of a power amplifier

characteristic of the DUT is the one that reports the phase of the instantaneous gain ($\angle G(|x_{in}|)$), usually expressed in degrees, as a function of the DUT's input or output power. Sample AM/AM and AM/PM characteristics are reported in Figures 1.8 and 1.9, respectively. These figures provide insightful information about the nonlinear behavior of the DUT. In fact, the shape of the AM/AM and AM/PM characteristics provide information about how severe the nonlinearity of the DUT is. Similarly, the dispersion of these two characteristics is a qualitative indication about the memory effects of the device.

1.5 1 dB Compression Point

The 1 dB compression point is a figure of merit commonly used to characterize the power capabilities of PAs along with their linearity. In the P_{out} vs. P_{in} characteristic, the 1 dB compression point is the one for which the actual output power of the amplifier is 1 dB lower than what it would have been if the amplifier was linear (and having a gain equal to its small signal gain). This definition is illustrated graphically in Figure 1.10, which reports the P_{out} vs. P_{in} characteristics of the actual and ideal amplifier. The ideal amplifier characteristic represents the extrapolated version of the linear portion of the actual amplifier's P_{out} vs. P_{in} characteristic. This figure shows that the 1 dB compression point can be defined either with respect to the input power ($P_{\text{1dB},in}$ in Figure 1.10) or with reference to the output power ($P_{\text{1dB,out}}$ in Figure 1.10).

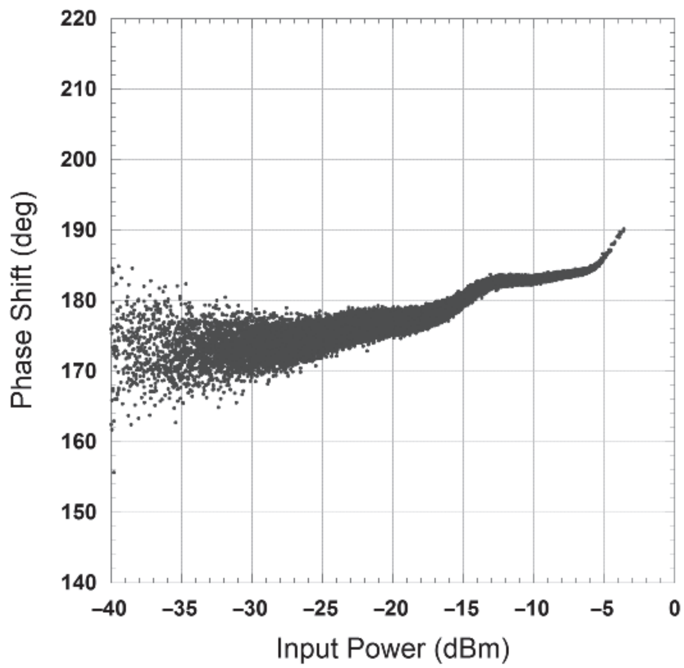

Figure 1.9 Sample AM/PM characteristic of a power amplifier

Though, the 1 dB compression point is commonly reported with respect to the output power of the device.

Similarly, the 1 dB compression point can be defined from the AM/AM characteristic. In this case, it corresponds to the power level for which the gain of the amplifier is 1 dB lower than its small signal linear value. From the AM/AM characteristic, the $P_{1dB,in}$ can be graphically determined as illustrated in Figure 1.11. If the small signal gain of the amplifier is denoted as G_{SS}, then the 1 dB compression point output power ($P_{1dB,out}$) can be obtained according to:

$$P_{1dB,out} = P_{1dB,in} + (G_{SS} - 1) \tag{1.12}$$

For a given DUT, the 1 dB compression point can vary depending on the test signal (CW versus modulated signals). The 1 dB compression point concept can be extended to the X-dB compression point. The X-dB compression point is defined in a way similar to that of the 1-dB compression point but for a gain compression of X-dB rather than 1 dB. Thus, the 3-dB compression point is the point of the P_{out} vs. P_{in} characteristic for which the actual output power of the amplifier is 3 dB less than what it would have been if the amplifier was linear; it is also the point of the AM/AM characteristic for which the gain of the device is 3 dB lower than its small signal value. The X-dB compression point can be used for the system level design of power amplification stages as well as building equation-based behavioral models in simulation software.

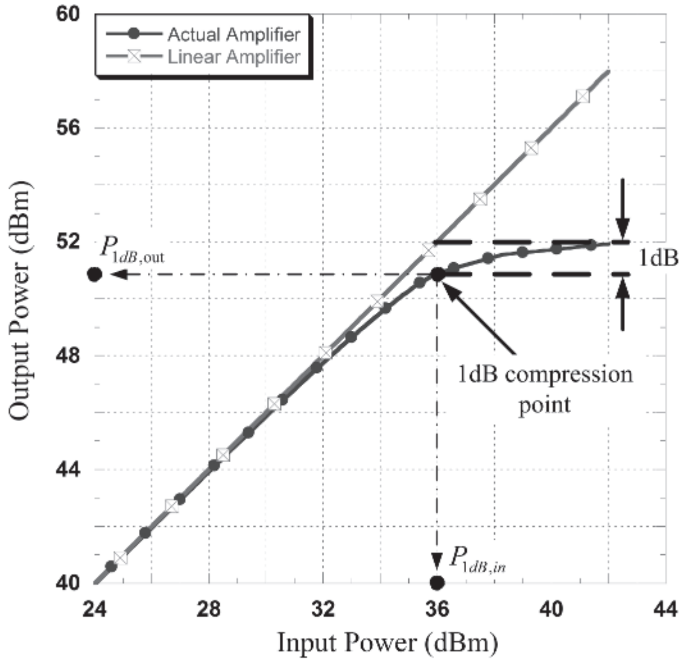

Figure 1.10 Graphical definition of the 1 dB compression point from P_{out} vs. P_{in} characteristic

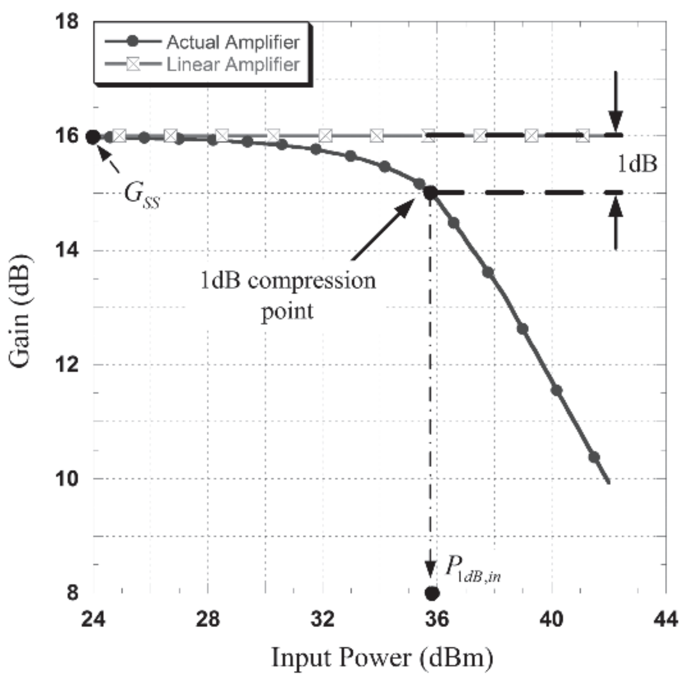

Figure 1.11 Graphical definition of the 1 dB compression point from the AM/AM characteristic

1.6 Third and Fifth Order Intercept Points

The 1 dB compression point characterizes the nonlinear behavior of PAs by only considering the power at the fundamental frequency. However, as amplifiers are driven deeper into their nonlinear regions, the amount of power generated at harmonic and inter-modulation frequencies becomes more significant. The intercept points are defined for odd order harmonics under a single-tone drive signal and odd order inter-modulation products under a multi-tone drive signal as these odd order harmonics and inter-modulation products fall within the close vicinity of the fundamental signal frequency. Third order intercept points are commonly used while fifth order intercepts points are used to a lesser extent. Higher order intercept points are seldom used since the power level generated at their corresponding frequencies is usually too low to have any significant impact on the behavior of the PA.

Figure 1.12 reports, for a sample amplifier driven by a two-tone test signal at frequencies f_1 and f_2 (with $f_1 < f_2$) and having equal amplitudes, the output power at the fundamental frequency (P_{out,f_1}) as a function of the total input power (P_{in}). In this same figure, the output power of the lower third order inter-modulation product ($P_{out,2f_1-f_2}$) is also plotted as a function of the total input power. The linear portion of the P_{out,f_1} vs. P_{in} characteristic has a 1 : 1 slope. However, the linear portion of the $P_{out,2f_1-f_2}$ vs. P_{in} characteristic has a 3 : 1 slope as it can be deduced from Equation 1.4.

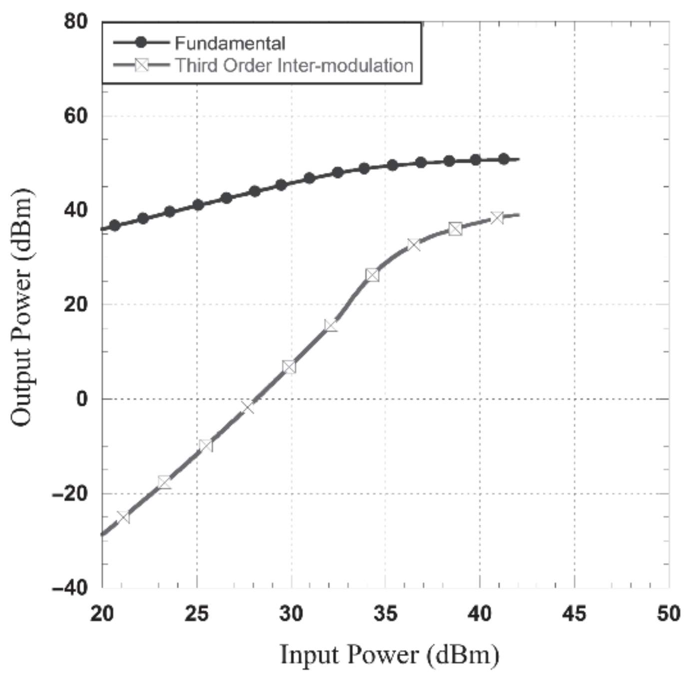

Figure 1.12 Output power characteristics at the fundamental and third order inter-modulation frequencies

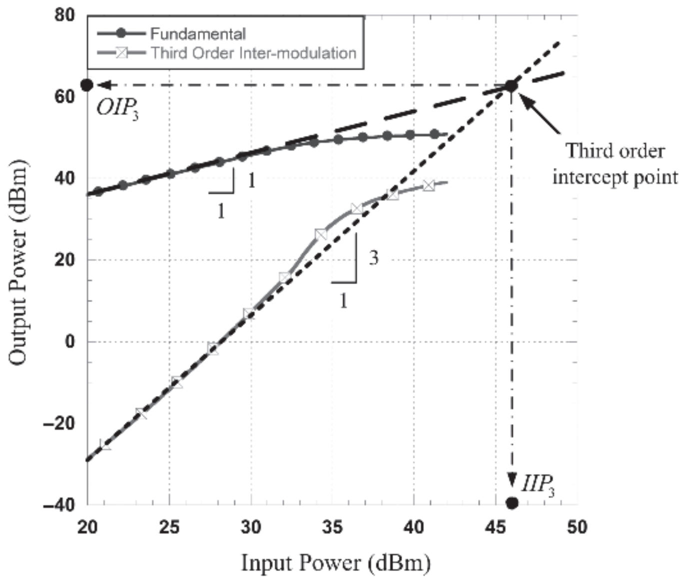

Figure 1.13 Graphical definition of the third order intercept point

The third order intercept point is defined as the intersection locus of the extrapolated linear portion of the P_{out,f_1} vs. P_{in} and the $P_{out,2f_1-f_2}$ vs. P_{in} characteristics as illustrated in Figure 1.13. When reported with respect to the input power, the third order intercept point is referred to as the third order input intercept point (IIP_3). Similarly, the third order intercept point can be reported with respect to the output power. In such a case, it is labeled as the third order output intercept point (OIP_3). For solid state PAs, the third order output intercept point is typically 10 dB higher than the output power at the 1 dB compression point.

The fifth order intercept point is defined in the same way by considering the extrapolated linear portions of the output power at the fundamental frequency and the output power at the frequency corresponding to the fifth order inter-modulation products (for example, $P_{out,3f_1-2f_2}$). In this case, the $P_{out,3f_1-2f_2}$ vs. P_{in} characteristic will have a 5 : 1 slope.

1.7 Carrier to Inter-Modulation Distortion Ratio

The 1 dB compression point and the intercept points characterize the nonlinear behavior of a PA without providing quantitative information about the amount of distortion it generates when operated at a given output power level. The carrier to inter-modulation distortion ratio (C/IMD) is a metric that quantifies the amount of distortion at the output of a PA driven by a two-tone, or in a more general case, a multi-tone test signal.

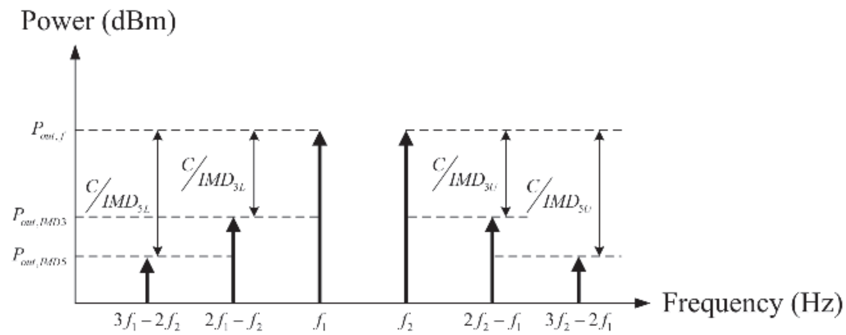

Figure 1.14 Graphical definition of the carrier to inter-modulation distortion ratio

It represents the ratio (in a linear scale) or equivalently the difference (in a logarithmic scale) between the power at the fundamental frequency (carrier) and the power generated at an inter-modulation frequency. The C/IMD is expressed in decibels relative to the carrier (dBc).

For an amplifier driven by a two-tone test signal at frequencies f_1 and f_2 (with $f_1 < f_2$), inter-modulation frequencies of interest commonly are the lower and upper third order inter-modulations ($2f_1 - f_2$ and $2f_2 - f_1$, respectively) and fifth order inter-modulations ($3f_1 - 2f_2$ and $3f_2 - 2f_1$, respectively). Figure 1.14 presents the power spectrum (in dBm) at the output of a memoryless PA having a fifth order nonlinearity and driven by a two-tone test signal at frequencies f_1 and f_2 (with $f_1 < f_2$). This figure graphically defines the lower and upper carrier to third order inter-modulation distortion ratios ($\frac{C}{IMD_{3L}}$ and $\frac{C}{IMD_{3U}}$, respectively) and those of the fifth order.

In memoryless PAs, the lower and upper C/IMD ratios are equal as reported in Figure 1.14. However, the stronger the memory effects of the amplifier are, the more significant the C/IMD asymmetry will be. The study of the asymmetry between the upper and lower C/IMDs provides an indication of the memory effects exhibited by the PA.

For a memoryless PA, it is possible to predict the third order C/IMD based on the operating output power and the third order intercept point of the device. Using the illustration of Figure 1.13, one can graphically determine that:

$$\frac{C}{IMD_3} = P_{\text{out},f} - P_{\text{out},IMD3} = 2 \times (OIP_3 - P_{\text{out},f}) \tag{1.13}$$

where $\frac{C}{IMD_3}$ is the carrier to third order inter-modulation distortion ratio and $P_{\text{out},f}$ and $P_{\text{out},IMD3}$ are the output power levels at the fundamental and third order inter-modulation frequencies, respectively. OIP_3 is the output power at the third order intercept point.

It is worth mentioning that the relation of Equation 1.13 is derived geometrically from the definition of the third order intercept point and assumes that the output power at both fundamental and third order inter-modulation frequencies is linear with respect to the input power. Thus, its accuracy will decrease as the amplifier is driven deeper into its nonlinear region where the output power characteristics at the fundamental and third order inter-modulation frequencies deviate from their linear approximations.

1.8 Adjacent Channel Leakage Ratio

The adjacent channel leakage ratio (ACLR) is used to quantify, in the frequency domain, the nonlinearity of PAs driven by modulated signals. It corresponds to the filtered ratio of the mean power in the main channel to the filtered mean power in an adjacent channel. This is a critical linearity parameter since the power generated by the nonlinear distortions in the adjacent channels cannot be eliminated by filtering and is perceived as interference when the adjacent channels are used for transmission. Thus, the power generated in the adjacent channels is considered as an unwanted emission that needs to be minimized and controlled. Accordingly, each communication standard stipulates, as part of the technical specifications of the transmitter characteristics, the ACLR threshold (also known as the spectrum emission mask) for base stations. A general illustration of the ACLR is illustrated in Figure 1.15, which

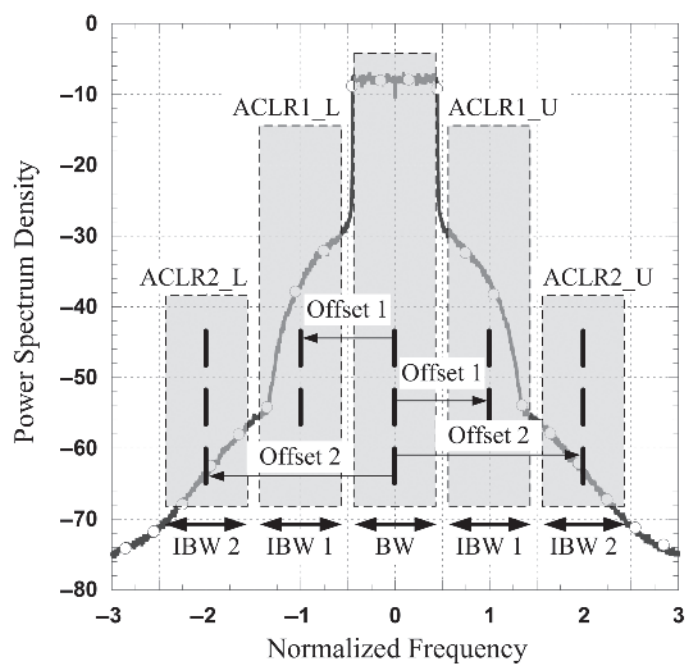

Figure 1.15 Graphical definition of the adjacent channel leakage ratio

reports a sample spectra at the output of the nonlinear transmitter as a function of the normalized frequency. The normalized frequency (f_n) is defined according to:

$$f_n = \frac{f - f_0}{BW} \tag{1.14}$$

where f and f_0 are the absolute frequency and the carrier frequency, respectively. BW represents the bandwidth of the signal.

Figure 1.15 shows that the channel power is calculated in a span that commonly equals the signal bandwidth (BW) and is centered around a normalized frequency of 0 (or equivalently an absolute frequency of f_0). It also shows the ACLR in the lower and upper first adjacent channels (ACLR1_L and ACLR1_U, respectively), and the ACLR in the lower and upper second adjacent channels (ACLR2_L and ACLR2_U, respectively). For each channel, the ACLR calculation requires the definition of the offset frequency that corresponds to the difference between the center of the main channel and that of the considered adjacent channel, as well as the integration bandwidth over which the power will be calculated in the considered adjacent channel.

The parameters used to calculate the ACLR are defined by the communication standards. These parameters include the main channel bandwidth, the adjacent channel parameters (offset frequency and integration bandwidth), and the type and parameters of the filter to be used to calculate the mean power.

1.9 Error Vector Magnitude

The error vector magnitude (EVM) is another measure used to quantify the nonlinear distortions of RF PAs and transmitters. The EVM is defined in the constellation domain and evaluates the deviation between the reference constellation point that should have been obtained in absence of distortions and the actual constellation point obtained in presence of distortions.

Transmitter distortions can be of three types: phase distortions, amplitude distortions, and in the more general cases, simultaneous phase and amplitude distortions. These three cases are illustrated in Figure 1.16 for the constellation diagram of a QPSK (Quadrature Phase Shift Keying) modulation scheme. Phase distortion appears as a rotation of the constellation points causing a phase error as shown in Figure 1.16a. Conversely, amplitude distortions will cause a magnitude error between the amplitudes of the vectors associated with the actual and reference constellation points as depicted in Figure 1.16b. Amplitude and phase distortions will result in an error on both the amplitude and phase of the vector associated with the demodulated constellation point. The effects of simultaneous phase and amplitude distortions on the constellation is illustrated in Figure 1.16c.

In the constellation domain, the error vector refers to the difference between the actual vector of the demodulated constellation point (S_n) and the reference vector associated with the corresponding reference constellation point ($S_{r,n}$) as shown in

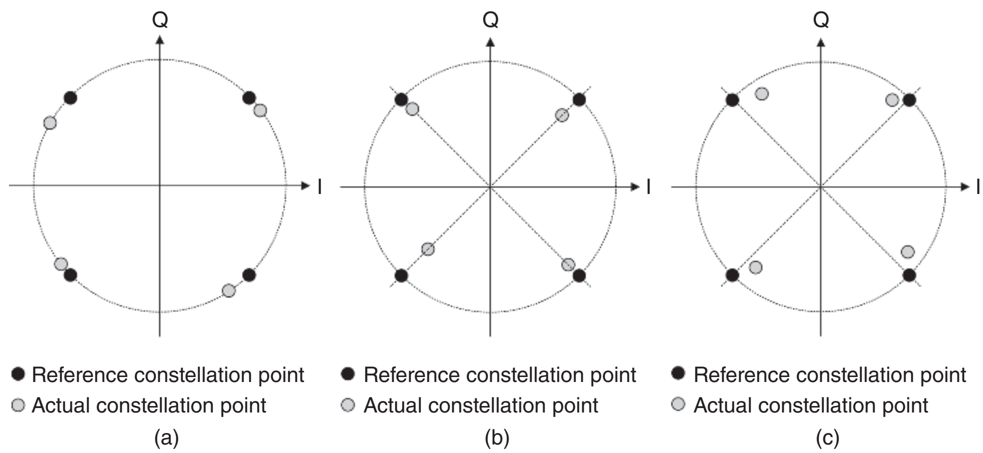

Figure 1.16 Effects of phase and amplitude distortions on the QPSK constellation. (a) Effects of phase distortions. (b) Effects of amplitude distortions. (c) Effects of phase and amplitude distortions

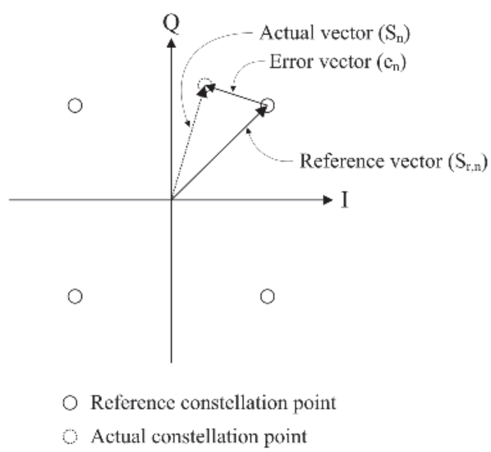

Figure 1.17 Graphical definition of the error vector

Figure 1.17. The EVM refers to the magnitude of the error vector, which is different from the error in the magnitudes except for the particular case where no phase distortions occur.

Threshold EVM values are specified for each communication standard and the latest technical specifications on the transmit modulation quality should be consulted. The EVM is typically expressed in percentage and calculated as the square root of the ratio of the mean power of the error vector to the mean reference power according to:

$$\text{EVM}\ (\%) = \sqrt{\frac{\frac{1}{N}\sum_{i=1}^{N}|e_i|^2}{\frac{1}{N}\sum_{i=1}^{N}|S_{r,i}|^2}} = \sqrt{\frac{\frac{1}{N}\sum_{i=1}^{N}|S_i - S_{r,i}|^2}{\frac{1}{N}\sum_{i=1}^{N}|S_{r,i}|^2}} \tag{1.15}$$

where N is the number of samples in the waveform. S_i and $S_{r,i}$ are vectors associated with the ith demodulated and reference constellation points, respectively. While e_i is the ith error vector between the demodulated and actual constellation points as defined in Figure 1.17.

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