A New Clipping Function for PAPR Mitigation: The Gaussian Clipping Function

Jacques Palicot*, Yves Louet* and Désiré Guel†
*CentraleSupélec/IETR, CentraleSupélec Campus de Rennes, 35510 Cesson-Sévigné, France
{email: jacques.palicot, yves.louet}@centralesupelec.fr
† Nokia Networks 7 Route de Villejust, 91620 Nozay, France
email: desire.guel@nokia.com

***Abstract*—A Gaussian function to clip multicarrier modulation is presented in this paper in order to decrease their Peak to Average Power Ratio (PAPR). The Gaussian clipping (GC) function is a soft non-linear function which keeps constant the average power of the signal, what is a characteristic of great importance in real transmission. The characteristics and performance of this GC is analysed both theoretically and by simulation. Furthermore, in the application context of Wifi IEEE802.11, this GC is compared to several other clipping functions well known in the literature such as hard clipping, smooth clipping and deep clipping. The results prove that the average power is kept constant which was our objective.**

***Index Terms*— PAPR; Clipping; Gaussian Clipping.**

I. Introduction

Orthogonal Frequency Division Multiplexing (OFDM), although used in standards such as IEEE 802.11a/g, IEEE 802.16, Digital Video Broadcasting (DVB) [1] and 4G cellular system/LTE suffers from the high Peak-to-Average Power Ratio (PAPR). Large PAPR requires a linear High Power Amplifier (HPA), which is inefficient. Moreover, the combination of a non linear HPA and large PAPR leads to in-band distortion and out-of-band radiation [2]. Several PAPR reduction techniques have been proposed [3]- [6]. The simplest way to reduce PAPR is to deliberately clip and filter the OFDM signal before amplification. However, clipping is a nonlinear process and may cause significant distortions that degrade the Bit Error Rate (BER) and increase adjacent out-of-band carriers [7]. The contribution of this paper is the following:

- first, we propose a new clipping function, the Gaussian Clipping (GC) function, which has the main advantage, compared to other clipping functions from the literature, to keep constant the average power.
- second, we analyse both theoretically and by simulation the performance of this GC, in terms of PAPR reduction gain and average power variation.
- third, we compare the GC performance with those of other clipping functions in terms of average power and BER in a Wireless Local Area Network (WLAN) context.

In Section II, we recall the basic idea of OFDM and we describe the PAPR problem. In Section III, we first describe clipping functions already known in the literature. In Section IV, we present our Gaussian clipping function. Section V deals with theoretical performance. Then Section VI presents some results and performance comparison, in a WLAN environment, of different clipping functions and finally we conclude the paper.

II. OFDM Systems and PAPR Issue

The basic idea underlying OFDM systems is the division of the available frequency spectrum into several subcarriers. To obtain a high spectral efficiency, the frequency responses of the subcarriers overlapp in an orthogonal way, hence the name OFDM. This orthogonality can be completely maintained with a small price in SNR degradation, even though the signal passes through a time dispersive fading channel, by introducing a Cyclic Prefix (CP).

The continuous-time baseband representation of an OFDM symbol is given by

$$x\left(t\right)=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}X_k e^{j2\pi f_k t},\quad 0\leq t\leq T_s\ , \qquad (1)$$

where N data symbols X_k form an OFDM symbol $\mathbf{X}=[X_0,\cdots,X_{N-1}]$, $f_k=\frac{k}{T_s}$ and T_s is the time duration of the OFDM symbol.

In practice, OFDM signals are typically generated by using an Inverse Discrete Fourier Transform (IDFT).

The OFDM symbol represented by the vector $\mathbf{X}=[X_0\cdots X_{N-1}]^T$ is transformed via IDFT into T_s/N-spaced discrete-time vector $\mathbf{x}=x\left[n\right]=\left[x_0\cdots x_{N-1}\right]^T$, i.e.

$$x_n=\frac{1}{\sqrt{N}}\sum_{k=0}^{N-1}X_k e^{j2\pi\frac{n}{N}k},\quad 0\leq n\leq N-1\ . \qquad (2)$$

In this paper, the discrete-time indexing $[n]$ denotes Nyquist Rate samples. Since oversampling may be needed in practical designs, we will introduce the notation $x\left[n/L\right]$ to denote oversampling by L. Different oversampling strategies of $x\left[n/L\right]$ can be defined. From now on, the oversampled IDFT output will refer to oversample of equation (2), which is expressed as follows:

$$x\left[n/L\right]\ =\frac{1}{\sqrt{N}}\sum_{k=0}^{NL-1}X_k e^{j2\pi\frac{n}{NL}k},\ 0\leq n\leq NL-1\ . \qquad (3)$$

The above expression (3) can be implemented by using a length-(NL) IDFT operation with the input vector

$$\mathbf{X}^{(\mathrm{L})} = \left[\begin{array}{ccc} X_0, \cdots, X_{\frac{N}{2}-1}, & \underbrace{0, \cdots, 0}_{(L-1)N \text{ zeros}} & X_{\frac{N}{2}}, \cdots, X_{N-1} \end{array} \right].$$

Thus, $\mathbf{X}^{(\mathrm{L})}$ is extended from $\mathbf{X}$ by using the so-called zero-padding scheme, i.e., by inserting $(L-1)N$ zeros in the middle of $\mathbf{X}$, i.e.,

$$X_k^{(\mathrm{L})} = \begin{cases} X_k, & k \in \mathcal{S}_1 \\ 0, & k \in \mathcal{S}_2 \end{cases},$$

where $\mathcal{S}_1$ and $\mathcal{S}_2$ are the set of In-Band (IB) indices and Out-Of-Band (OOB) indices respectively.

In the literature, the envelope variations of $x[n/L]$ are often described in terms of the crest-factor (CF), peak-to-mean envelope power ratio (PMEPR) or simply peak-to-average power ratio (PAPR). In this paper, we adopt the term PAPR to quantify the envelope excursions of the signal. The PAPR of the signal $x(t)$ may be defined as

$$\mathrm{PAPR}_{[x]} \triangleq \frac{\max\limits_{t\in[0,T_s]} |x(t)|^2}{\mathcal{P}_x}, \tag{4}$$

where $\mathcal{P}_x = E\left\{|x(t)|^2\right\}$ is the average signal power and $E\{.\}$ is the statistical expectation operator. Note that, in order to avoid aliasing distortion into the data bearing subcarriers and in order to accurately describe the PAPR, an oversampling factor $L \geq 4$ is required.

In the literature, it is customary to use the Complementary Cumulative Distribution Function (CCDF) of the PAPR as a performance criterion. It is denoted as

$$\mathrm{CCDF}_{[x]}(\psi) \triangleq \Pr\left\{\mathrm{PAPR}_{[x]} \geq \psi\right\}.$$

If N is large enough, based on the central limit theorem, the real and imaginary parts of OFDM $x(t)$ follow a Gaussian distribution and its envelope will follow a Rayleigh distribution. This implies a large PAPR.

III. Some Clipping Functions

In this Section, we present the clipping techniques family [8]. Whatever the clipping technique to reduce OFDM PAPR is, the output signal y_n, in terms of the input signal x_n is given as follows:

$$y_n = f(|x_n|)\, e^{j\varphi_n}, \tag{5}$$

where φ_n is the x_n phase and $f(.)$ is the clipping function.

A. Classical Clipping (CC) technique

The Classical Clipping (CC) proposed in [7] is one of the most popular clipping technique for PAPR reduction known in the literature [9] [7]. It is sometimes called hard clipping or soft clipping. To avoid any confusion, it is called Classical Clipping (CC) in this paper. In [7], its effects on the performance of OFDM, including the power spectral density, the PAPR and BER are evaluated. The function-based clipping used for CC technique is defined below and depicted in Figure 1 (a).

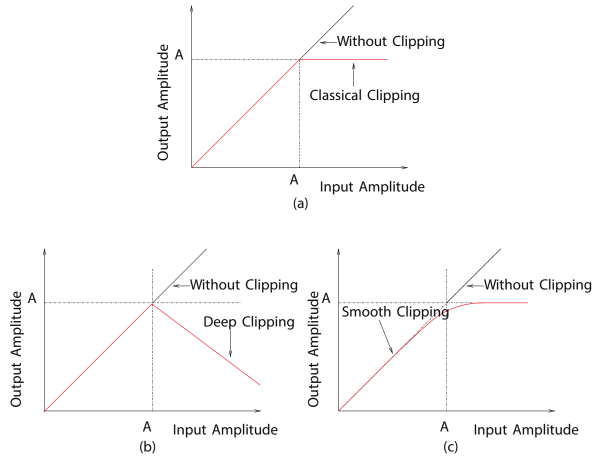

Fig. 1: Functions-based clipping for PAPR reduction

$$f(r) = \begin{cases} r, & r \leq \mathrm{A} \\ A, & r > \mathrm{A} \end{cases}, \tag{6}$$

where A is the clipping threshold.

B. Deep Clipping (DC) technique

Deep Clipping has been proposed in [10] to solve the peaks regrowth problem due to the out-of-band filtering of the classical clipping and filtering method. So, in DC technique, the clipping function is modified in order to "deeply" clip the high amplitude peaks. A parameter called clipping depth factor has been introduced in order to control the depth of the clipping. The function-based clipping used for DC technique is defined below and depicted in Figure 1 (b).

$$f(r) = \begin{cases} r & , \quad r \leq \mathrm{A} \\ \mathrm{A} - \beta(r - \mathrm{A}) & , \quad \mathrm{A} < r \leq \frac{1+\beta}{\beta}\mathrm{A} \\ 0 & , \quad r > \frac{1+\beta}{\beta}\mathrm{A} \end{cases},$$

where β is called the clipping depth factor.

C. Smooth Clipping (SC) technique

In [11], a Smooth Clipping technique is used to reduce the OFDM PAPR. In this paper, the function based-clipping for SC technique is defined below and depicted in Figure 1 (c).

$$f(r) = \begin{cases} r - \frac{1}{b}r^3, & r \leq \frac{3}{2}\mathrm{A} \\ \\ \mathrm{A}, & r > \frac{3}{2}\mathrm{A} \end{cases},$$

where $b = \frac{27}{4}\mathrm{A}^2$.
These three clipping functions are drawn on Figure 1 and have been completely studied and compared in [8]. We may notice that the 'invertible clipping' of [12] is a variant of SC.

To the best of authors's knowledge, since 2008 with the DC [10], no new clipping functions has been proposed in the literature. Of course, a lot of papers deal with OFDM clipping but from many others point of view as the threshold computation of the CC [16] [17] [18] or the mitigation of the clipping noise [10] [13] [15] ..., but no new clipping function have been proposed so far as the GC introduced in this paper.

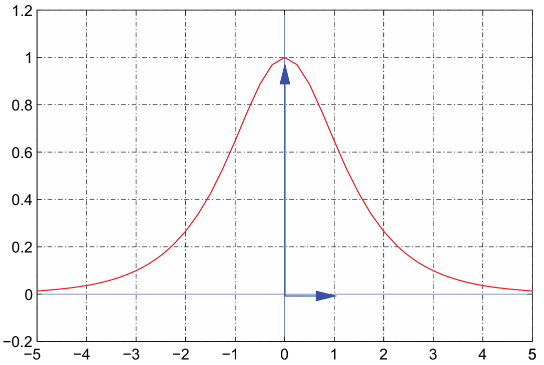

Fig. 2: Gaussian function curve.

IV. Gaussian Clipping

In this Section, we present the Gaussain clipping for PAPR reduction. We start from the Gaussian function, which is drawn in Figure 2. It will act on the multicarrier signal amplitude in order to decrease its PAPR. In this context, only positive values are taken into account, because the signal modulus is always a positive value.

The Gaussian Clipping function $f(.)$, associated to this Gaussian function, is expressed as:

$$f(r) = Ae^{-(\eta r)^2}, \quad r \geq 0. \tag{7}$$

The parameters A and η control the performance of the method (the transmitted mean power variation and the PAPR reduction capability).

The GC technique equation (7) can reduce the OFDM PAPR by increasing low amplitudes samples and by decreasing high amplitudes samples, as illustrated in Figure 3.

Figure 3 shows that for samples r_n such that $r_n = |x_n| \leq r^{(\text{threshold})}$, the signal is amplified whereas for samples r_n such that $r_n \geq r^{(\text{threshold})}$ the signal is attenuated.

The threshold value, $r^{(\text{threshold})}$, which corresponds to the threshold between amplification and reduction of the signal is obtained by solving equation (8) and is given by equation (9).

$$f[r] = Ae^{-(\eta r)^2} = Ar. \tag{8}$$

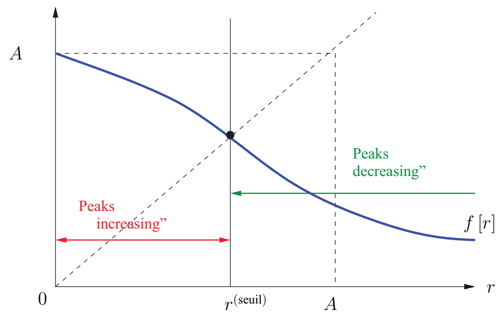

Fig. 3: Behavior of Attenuation/Amplification of the signal versus A and r parameters.

What gives:

$$r^{(\text{threshold})} = \sqrt{\frac{W(2\eta)}{2\eta}}, \tag{9}$$

where W is the Lambert function. The equation (9) shows that $r^{(\text{threshold})}$ depends only on the η parameter of the GC (see equation (7)). It is therefore clear that $r^{(\text{threshold})}$ and consequently η, drives the PAPR reduction gain of the GC.

We will now explain the influence of A in the PAPR reduction gain. We remind that one of our main objective is to keep constant the average power between the input and the output of clipping. Therefore, we would like to have $\mathcal{P}_y = \mathcal{P}_x$, where $\mathcal{P}_x$ is the average power of the signal before and $\mathcal{P}_y$ is the average power of the signal after the PAPR mitigation technique. Considering equation (7) $\mathcal{P}_y$ is given by equation (10):

$$\mathcal{P}_y = \int_0^\infty f(r)^2 p(r) dr = A^2 \int_0^\infty \mathrm{e}^{-2(\eta r)^2} p(r) dr. \tag{10}$$

Therefore, the ratio γ between the two powers $\mathcal{P}_x$ and $\mathcal{P}_y$ is expressed as follows:

$$\gamma = \frac{\mathcal{P}_y}{\mathcal{P}_x} = \frac{A^2}{\mathcal{P}_x} \int_0^\infty \mathrm{e}^{-2(\eta r)^2} p(r) dr. \tag{11}$$

As shown by equation (11), A and η influence the ratio γ. This means that it is possible to drive the ratio γ between the two powers thanks to parameter A without modifying the PAPR reduction gain, for a given η . In fact we schowed that the PAPR reduction gain only depends on η parameter.

The A parameter value which give $\mathcal{P}_y = \mathcal{P}_x$ is given by the equation (12)

$$A^{(\text{opt})} = \frac{\sqrt{\mathcal{P}_x}}{\left[\int_0^\infty \mathrm{e}^{-2(\eta r)^2} p(r) dr\right]^{\frac{1}{2}}}. \tag{12}$$

To summarize, we have shown, theoretically, that η parameter drives the PAPR reduction gain whereas A parameter drives the average power variation for a given η.

V. Theoretical Study of Gaussian Clipping

In this Section, we analyse theoretically the behavior of the GC function. We focus (subsectionV-A) on the average power variation given by the following equations:

$$\gamma = \frac{\mathcal{P}_y}{\mathcal{P}_x}, \tag{13}$$

$$\Delta \mathrm{E} = 10 \log_{10}(\gamma). \ [\mathrm{dB}] \tag{14}$$

In subsectionV-B, we focus on the PAPR CCDF at the ouput of the GC function. We are interested in the PAPR reduction gain ΔPAPR for a CCDF value of 10^{-2} before and after clipping.

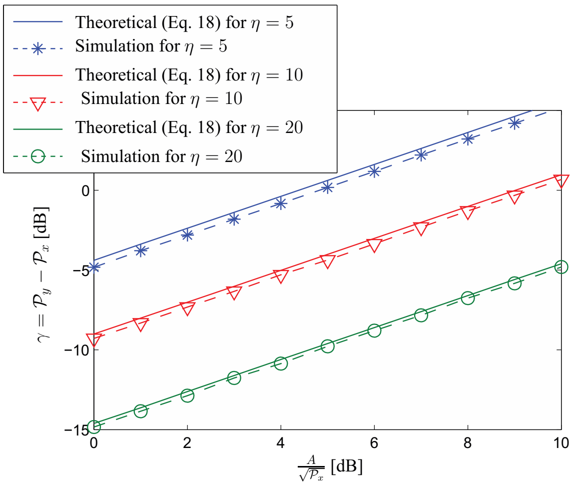

Fig. 4: Theoretical and simulation average power variation comparison for several values of $\frac{A}{\sqrt{\mathcal{P}_x}}$.

A. Average power variation analysis

The expression of the transmitted mean power $\mathcal{P}_y$ as a function of the OFDM mean power $\mathcal{P}_x$ can be expressed as:

$$\mathcal{P}_y = \int_0^{+\infty} [f(r)]^2 p_x(r)\, dr, \tag{15}$$

where $p_x(r)$ is the probability density function (PDF) of the OFDM envelope and can be assimilated to a Rayleigh distribution for a large number of OFDM subcarriers:

$$p_x(r) = \tfrac{2r}{\mathcal{P}_x} e^{-\frac{r^2}{\mathcal{P}_x}}, \quad r \geq 0. \tag{16}$$

By sustituting the expression of $p_x(r)$ in equation (15), the expression of the transmitted mean power $\mathcal{P}_y$ is given by:

$$\mathcal{P}_y = \int_0^{+\infty} \left[A e^{-(\eta r)^2} \right]^2 \frac{2r}{\mathcal{P}_x} e^{-\frac{r^2}{\mathcal{P}_x}} = \frac{A^2}{1 + 2\eta^2 \mathcal{P}_x}. \tag{17}$$

Let us consider γ the output-to-input mean power ratio; using equation (17), γ is expressed below

$$\gamma \triangleq \frac{\mathcal{P}_y}{\mathcal{P}_x} = \frac{A^2}{(1 + 2\eta^2 \mathcal{P}_x)\, \mathcal{P}_x}. \tag{18}$$

From equation (18), it is easy to compute the value of $A^{(\text{opt})}$ (that means the A value which gives $\mathcal{P}_y = \mathcal{P}_x$) and is expressed as:

$$A^{(\text{opt})} = \left[\left(1 + 2\eta^2 \mathcal{P}_x\right) \mathcal{P}_x \right]^{\frac{1}{2}}. \tag{19}$$

Equation (19) shows that $A^{(\text{opt})}$ depends on η (which controls the PAPR reduction gain) and the average power of input signal.

The average power variation related to the Gaussian clipping given by equation (14) is compared with simulation results in Figure 4. Results show a good match between theory (equation 18) and simulation. For a given η value, the average power is a linear function of A. Therefore, for a given η parameter value, it is possible to find the value of $A^{(\text{opt})}$ which keeps constant the average power (our objective). This is given by the value $\gamma = 0$.

B. PAPR distribution analysis

In this subsection, the PAPR CCDF is derived anatyically for the ouput signal. To perform this analysis, like in [19] for the classical OFDM PAPR analysis, we assume that the signals $x(t)$ and $y(t)$ (input and ouput of the Gaussian clipping respectively) are sampled at the Nyquist rate (that means, oversampling factor $L = 1$). Therefore, input and output samples x_n and y_n are respectively given by:

$$\begin{aligned} x_n &= x\left(\frac{n}{N} T_s\right), \\ y_n &= y\left(\frac{n}{N} T_s\right), \end{aligned} \tag{20}$$

where $0 \leq n < N$, N is the subcarriers number and T_s the OFDM symbol period. The signals x_n and y_n may also be written:

$$\begin{aligned} x_n &= r_n e^{j\phi_n}, \\ y_n &= f[r_n]\, e^{j\phi_n} = v_n e^{j\phi_n}, \end{aligned} \tag{21}$$

where r_n is the amplitude of x_n et ϕ_n its phase; $v_n = f(r_n)$ is the amplitude of y_n.

The PAPR of y_n is defined as:

$$\text{PAPR}_{[y]} = \frac{\max\limits_{0 \leq n < N} |y_n|^2}{\mathcal{P}_y} = \frac{\max\limits_{0 \leq n < N} {v_n}^2}{\mathcal{P}_y}. \tag{22}$$

By applying the same development as in [19], and by assuming v_n independance values we derive:

$$\begin{aligned} \text{CCDF}_{[y]}\left(\tilde{\psi}\right) &= \Pr\left[\text{PAPR}_{[y]} \geq \tilde{\psi}\right] = \Pr\left[\frac{\max\limits_n v_n^2}{\mathcal{P}_y} \geq \tilde{\psi}\right] \\ &\simeq 1 - \prod_{n=0}^{N-1} \left\{ \Pr\left[\frac{f(r_n)^2}{\mathcal{P}_y} \leq \tilde{\psi}\right] \right\}, \\ &\simeq 1 - \prod_{n=0}^{N-1} \left\{ \Pr\left[f(r_n) \leq \sqrt{\tilde{\psi}\mathcal{P}_y}\right] \right\} \end{aligned} \tag{23}$$

where $f[r]$ is the Gaussian clipping function given by equation (7).

By using equation (7), we get:

$$\text{CCDF}_{[y]}\left(\tilde{\psi}\right) \simeq 1 - \prod_{n=0}^{N-1} \left\{ \Pr\left[r_n \geq \frac{1}{\eta} \left[\ln\left(\frac{A}{\sqrt{\tilde{\psi}\mathcal{P}_y}} \right) \right]^{\frac{1}{2}} \right] \right\}. \tag{24}$$

As r_n is a Rayleigh i.i.d random variable whose probability density function is given by equation (16), equation (24) becomes:

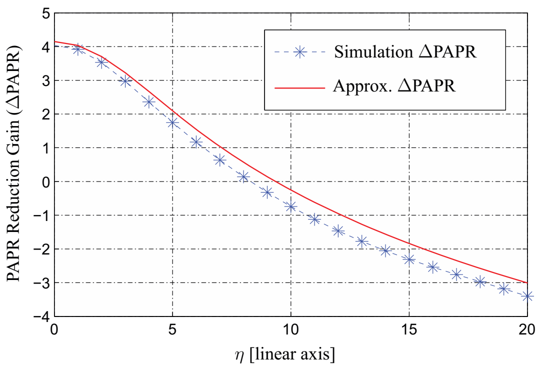

Fig. 5: PAPR reduction gain comparison between theoretical and simulation results versus η parameter for $\frac{A}{\sqrt{\mathcal{P}_x}} = 3$ dB.

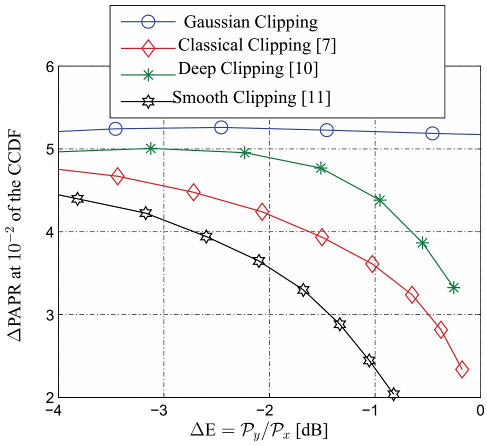

Fig. 6: PAPR reduction gain versus ΔE for the four clipping techniques.

$$\mathrm{CCDF}_{[y]}\left(\tilde{\psi}\right) \simeq 1 - \prod_{n=0}^{N-1}\left[e^{-\frac{\ln\left(\frac{A}{\sqrt{\tilde{\psi}\gamma\mathcal{P}_x}}\right)}{\eta^2\mathcal{P}_x}}\right], \qquad (25)$$

$$\simeq 1 - e^{-N\frac{\ln\left(\frac{A}{\sqrt{\tilde{\psi}\gamma\mathcal{P}_x}}\right)}{\eta^2\mathcal{P}_x}},$$

where $\mathcal{P}_x$ is the OFDM average power and γ the ratio between output average power and input average power given by equation (18).

The PAPR reduction gain is compared to simulations results and is presented in Figure 5 for several values of η parameter. It shows that the theoretical approximation of equation (25) is very close to simulation results.

The PAPR reduction gain decreases when η parameter increases. This result provides us an upper bound of η. In fact it should be smaller then 8 to have a positive PAPR reduction gain, which, of course, is our objective.

VI. Comparative Results Study with Other Clipping Functions

In this subsection, GC performance are compared with classical clipping [7] (Section III-A), Deep clipping [10] (Section III-B) and Smooth clipping [11] (Section III-C) performance. This comparative study is performed in the context of the WLAN standard IEEE 802.11 a/g, whose parameters are given in Table I.

TABLE I: SIMULATION PARAMETERS.

Paramètres du Système	Valeurs
Modulation type	16-QAM
Carriers number	N = 64
Data sub carriers number	48
Pilots number	4
Oversampling factor	$L = 4$
Channel type	AWGN

In Figure 6 the PAPR reduction gain, ΔPAPR, is analysed for the four clipping techniques in function of the average power variation ΔE. For the Classical, Deep and Smooth clipping functions, the PAPR gain decreases with ΔE and becomes very small for ΔE $\simeq 0$ dB. At the opposite, this PAPR gain with the GC is quasi constant in function of ΔE. In fact, whatever the value of ΔE is, the PAPR gain ΔPAPR of GC is equal to around 5.2 dB. This result is the great advantage of the GC, because it offers a PAPR reduction of 5.2 dB without modifying the average power. To reach this result it is necessary to set $\frac{A}{\sqrt{\mathcal{P}_x}}$ at 0.45 dB as it is shown in Figure 7. In this figure, the influence of the A parameter is presented. The results show that parameter A could control the average power variation without modifying the PAPR reduction gain. This result is very important. In fact, it is possible to choose A in such a way that $\mathcal{P}_y = \mathcal{P}_x$ without modifying the PAPR reduction gain. In other words, with the GC function it is possible to reach a PAPR reduction gain of 5 dB with an average power variation $\Delta\mathrm{E} = 0$.

Figure 8 presents the BER for the four clipping techniques. As expected, these techniques degrade the BER. In fact the signal resulting from clipping functions is useful for PAPR reduction but is also the interferer signal which deteriorates the signal both in band and out of band. Generally out of band

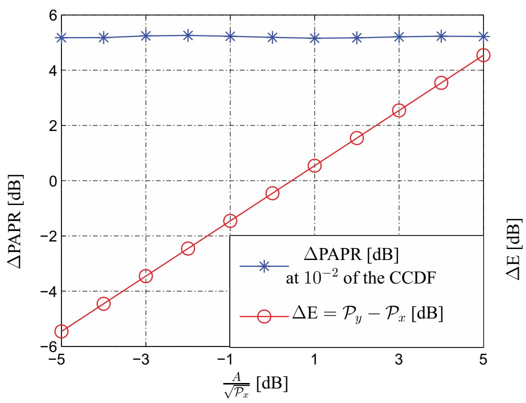

Fig. 7: PAPR reduction gain and average power variation of the GC function versus parameter A .

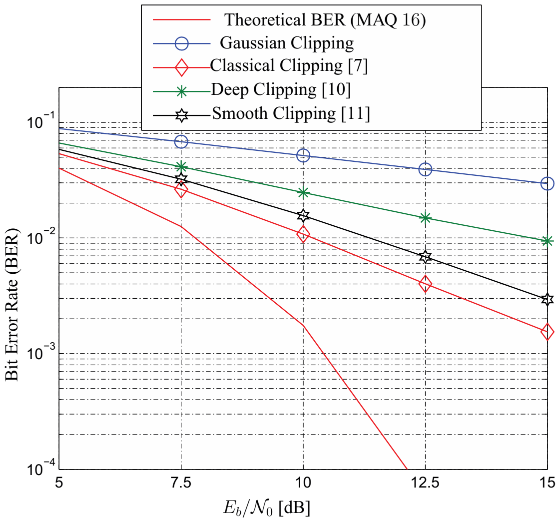

Fig. 8: BER Comparison for the four clipping techniques.

degradation is suppressed by filtering (it is why clipping techniques are generally named clipping and filtering techniques). As it could be seen GC is the one which degrades the most the BER. This was expected because the PAPR reduction was the greatest. That means that GC (as every clipping function) could not be used without BER improvement. To improve BER degradation due to clipping noise (whatever the clipping function is), several techniques could be performed:

- by inverting the clipping function or by iterative subtraction of the noise regenerated with the clipping function at the receiver [13]. Iterative methods to substract the estimated noise have been proposed in [14] and in [15]. The main drawbacks, in our point of view, is that these techniques become no more backward compatible and add complexity at the receiver side. Furthermore, the OOB noise, will degrade the signal in the ajacent band (the so called shoulders), which is, of course, not acceptable.
- Another alternative consists in turning the clipping method into a Tone Reservation (TR) method. By principle TR does not deteriorated the BER. This technique has several advantages: i) to be performed at the transmitter side, ii) to be backward compatible, iii) to be very simple to realize. It is this technique we have studied in [20] .

VII. Conclusion

Gaussian Clipping has been proposed in this paper. We studied its theoretical performance in terms of PAPR reduction and average power variation. These performances were also evaluated trough simulations and compared to other clipping techniques. The main conclusion is that the proposed GC is a very interesting clipping method when keeping the average power constant is a strong requirement.

Of course, as all the other clipping functions, due to their non linear characters, GC degrades the BER, which means that this clipping technique should be used in addition with filtering and/or with TR method.

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