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# Clipping noise mitigation with capacity approaching FEC codes for PAPR reduction of OFDM signals

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**Abstract**—By virtue of its simplicity, clipping appears to be one of the most attractive methods to mitigate Peak to Average Power Ratio (PAPR) in OFDM systems. After positioning clipping in a new classification of PAPR reduction methods, this article focuses on the performance of clipped OFDM at low signal to noise ratio, typical of capacity approaching forward error correcting coded systems. A lower bound on the BER performance of clipped coded OFDM is given and shown to be consistent with the genie aided performance of the iterative clipping noise cancellation receiver. Finally, numerical results on the BER performance of a clipped LDPC coded OFDM system are provided.<sup>1</sup>

**Index Terms**—PAPR, classification, OFDM, clipping, iterative cancellation, LDPC

## I. INTRODUCTION

It is well known that Orthogonal Frequency Division Multiplexing (OFDM) is prone to large Peak to Average Power Ratio (PAPR) due to the summation of many carriers in parallel. In spite of large PAPR reduction gains, some PAPR mitigation methods are unfortunately shadowed by some deficiencies as complexity, bit error rate degradation, spectral efficiency loss, etc. This has motivated publication of PAPR reduction methods synthesis as in [1][2] aiming at gathering methods according to some key criteria as power increase, distortion, data rate loss, processing at the transmitter, etc. Nevertheless, some updates are necessary due to new PAPR reduction methods publication. In particular, [3] bridged clipping methods and adding signal methods thanks to a consequence of Busgang theorem. Thus the first part of this article proposes an updated tree classification of PAPR reduction methods. This classification results in a straightforward statement: regardless noise degradation, clipping appears as one of the most powerful PAPR reduction methods due to both its simplicity and PAPR reduction capacity. The second part of this article deals with an efficient coded receiver to mitigate the clipping noise at low signal to noise ratio (SNR), and, consequently, limit the BER degradation due to clipping. Based on the statistical model of the clipping device [22], a lower bound on the bit error rate (BER) performance of the clipped coded OFDM system is derived. Unlike [22], this bound is based on the assumption that clipping noise can be perfectly mitigated.

<sup>1</sup>The authors would like to thank Charbel Abdel Nour for providing the source code of the LDPC part used to obtain the simulation results and the Université Européenne de Bretagne for its financial support via the YGYC project.

Then we focus on the iterative clipping noise cancellation receiver [23] associated with capacity approaching codes, such as low density parity check codes (LDPC) or turbo codes. The BER performance of the iterative receiver is evaluated numerically and compared to the derived lower bound.

## II. CLASSIFICATION OF PAPR REDUCTION METHODS

### A. Discriminating PAPR reduction methods parameters

A huge quantity of PAPR reduction methods has been proposed since a dozen of years especially due to the overgrowing success of OFDM modulation. As usual, these methods are not equivalent regarding performance and some outstanding benefits may unfortunately shadow some major drawbacks. This implies the need of metrics definitions so as to classify PAPR reduction methods.

1) *PAPR reduction gain parameter*: This is the first parameter to consider. A PAPR reduction method is of interest if it is able to mitigate the original PAPR in a given ratio. This value is estimated at a probability level for which the PAPR exceeds a given threshold.

2) *Downward compatibility parameter*: The downward compatibility regards the receiver modification. A method has a downward compatibility if the receiver is not modified consequently when a PAPR reduction method is implemented. This is the case for instance for Active Constellation Extension or Tone Reservation methods. This characteristic is of utmost importance in broadcast or mobile communications if a method is implemented at the transmitter (base station).

3) *Bit error rate degradation parameter*: This key parameter drives a communication chain performance and has to be checked when implementing a PAPR reduction method. The reason is that some of these methods may degrade BER. For example clipping methods increase the bit error rate (BER).

4) *Power increase parameter*: Some methods imply a power increase of the signal after PAPR reduction. This is especially the case when considering adding signal methods formulation as Tone Reservation. This is viewed as a drawback.

5) *Data rate loss parameter*: Some methods need an increase of the bandwidth and consequently a decrease of the spectral efficiency. This is especially the case when some side information (SI) has to be transmitted like in Selected Mapping (SLM). If the bandwidth has to be kept constant, this implies a data rate loss and constitutes a drawback.

6) *Complexity parameter*: Even if a method has powerful characteristics, complexity has to be taken into account in case of implementation on real systems. In this case too much complex methods (complexity has to be defined clearly) will be removed.

7) *Synthesis*: Selecting a PAPR reduction method in a given context is the result of a complex trade off between these parameters and the constraints of the communication on which PAPR has to be mitigated. As a consequence, one needs to classify most popular PAPR reduction methods (not all because it would be impossible) according to these criteria. This will be detailed in next section.

### B. Proposed classification

The proposed classification follows [2] in its principle. The idea is to give the researchers and engineers a tool for selecting the most appropriate PAPR reduction method according to a set of criteria or parameters detailed in the section II-A. A new vision of distortion methods proposed in [3] paved the way to this classification. It has been shown, thanks to Bussgang theorem, that any distortion method can be written as an adding signal method. Consequently, following classifications given in [4] and [5], the three top categories are : adding signal methods, probabilistic methods and coding methods as shown in Fig.1

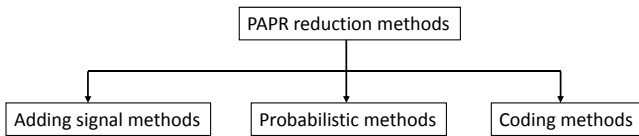


Fig. 1. The three PAPR reduction methods categories

1) *Adding signal methods category*: This category gathers all methods whose PAPR reduction can be formulated as  $PAPR(X + C) < PAPR(X)$  where  $X$  refers to the useful data (in time or frequency domain) and  $C$  the peak cancelling signal (in time or frequency domain) necessary to mitigate the original PAPR. The very first question is to know if the method has a downward compatibility or not. If no, one can find companding methods [6] and invertible clipping [7]. If yes, the following question is to know if there are some carriers that have been reserved to carry the corrective signal.

#### *Downward compatibility without reserved carriers*

When no carrier is reserved for PAPR reduction in the adding signal context, BER becomes a key parameter. Some methods do not imply bit error rate degradation as Active Constellation Extension (ACE) [8] or Tone Injection (TI) [9] whereas others as regular clipping [10], deep clipping [11] or geometric method [12] degrade BER.

#### *Downward compatibility with reserved carriers*

In this context, some carriers are dedicated for PAPR reduction and as a consequence BER is not affected. Two situations may happen: 1) peak reduction carriers are specially

dedicated (as methods using unused carriers as [13] or [14] in case of TR-clipping based method), 2) peak reduction carriers are taking among useful carriers (as regular Tone Reservation [9]). As a synthesis, Fig.2 sketches the positioning of all adding signal methods.

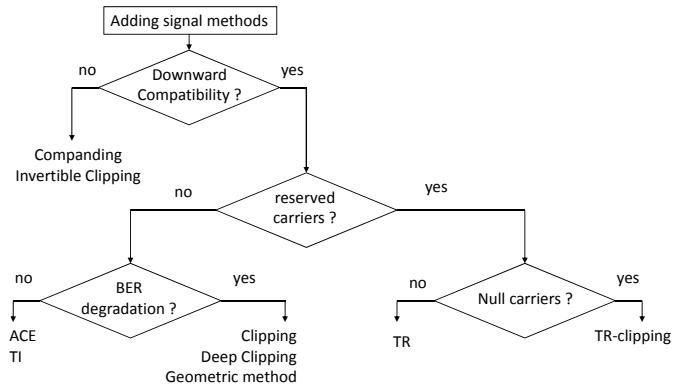


Fig. 2. Synthesis of Adding signal methods

2) *Probabilistic methods category*: By modifying the phase, amplitude and/or subcarrier position, the idea of these methods is to deliver several copies of the original OFDM symbols and to select the one whose PAPR is the lowest. As a consequence, side information (SI) has to be sent most of the time to the receiver so as to recover useful data. In this context, the BER is not modified since the useful data is totally recovered after SI processing. In these cases, the methods do not verify the downward compatibility. Among them one can find the Selected Mapping (SLM) [15] or Partial Transmit Sequences (PTS) [16]. Nevertheless in case of blind version of these methods, they become downward compatible as in [17]. Fig.3 gives a picture of these methods classification.

3) *Coding methods category*: This third category gathers all methods using coding techniques to mitigate the PAPR. It is straightforward that here the downward compatibility is not respected and the BER is not degraded. Reed-Muller codes generating constant PAPR Golay sequences [18] or Trellis shaping [19] fall under this category. Finally Fig.3 gives a picture of the proposed classification.

### III. CLIPPING METHOD AND CAPACITY APPROACHING ERROR CORRECTING CODES

In section II, the clipping method has been positioned in the classification tree. Clipping appears as one of the most powerful methods for its simplicity and its PAPR reduction capability. However, the BER degradation has to be mitigated so as this method can be implemented in real systems. The first part of this section describes the clipped LDPC coded OFDM system. In section III-B, assuming a statistical model of the clipping process and perfect mitigation of the clipping distortion term, a lower bound on the BER performance of the clipped coded OFDM system is produced. In section III-C, the iterative cancellation clipping noise receiver is briefly

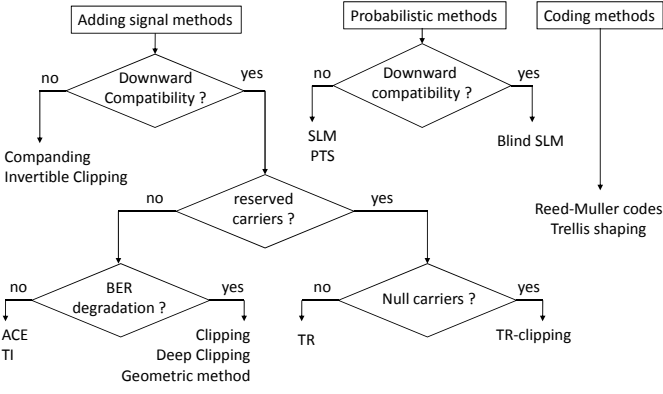


Fig. 3. The proposed classification

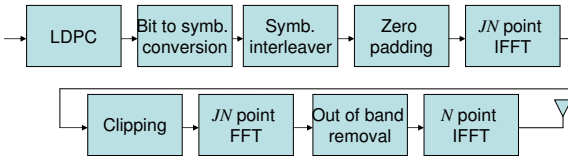


Fig. 4. Transmitter

reviewed and the lower bound is compared to the performance of the genie aided receiver. Finally, we provide simulated BER performance results of the clipped LDPC coded OFDM system and compare those results with the lower bound.

#### A. System model

Figure 4 shows the system model of the clipped coded discrete-time OFDM system with oversampling and filtering. Serial to parallel conversion and guard interval insertion are not shown. For oversampling factor  $J > 1$ , filtering is performed in the frequency domain (out-of-band removal). Data is first encoded by a 4/9 DVB-T2 LDPC. The frame of encoded bits of size 16200 is then converted to 16QAM complex symbols followed by a symbol interleaver. After serial to parallel conversion, the complex symbols in the frequency domain  $\mathbf{X} = [X_0, \dots, X_{JN-1}]$  are converted to the time domain through IDFT of size  $JN$ . The clipping operation  $g$  is applied to the time domain OFDM signal and in the case of a soft limiter, the signal after clipping is expressed as

$$x_j^c = g(x_j) = \begin{cases} x_j, & |x_j| \leq A \\ A \exp(\arg x_j), & |x_j| > A \end{cases} \quad (1)$$

where  $A$  is the maximum permissible amplitude. The Clipping Ratio (CR)  $\gamma$  is defined as  $\gamma = \left(\frac{A}{\sqrt{P_{in}}}\right)$ , where  $P_{in}$  is the average signal power before clipping, equal to  $\sigma_x^2$ . By assuming that the OFDM signal samples  $x_j$  are characterized by a discrete complex stationary Gaussian process, the clipping noise process at the output of the soft envelope limiter can be modelled statistically as an aggregate of an attenuation of the

input signal and clipping noise (Busgang's theorem)[24]

$$x_j^c = \alpha x_j + d_j \quad (2)$$

where the attenuation  $\alpha$  is given by  $\alpha = 1 - \exp(-\gamma^2) + \frac{\sqrt{\pi}}{2} \text{erfc}(\gamma) \cdot \gamma$  and the complex distortion term  $d_j$  is Gaussian and assumed to be uncorrelated with  $x_j$ . The average output signal power of the soft limiter is then

$$P_{out} = (1 - \exp(-\gamma^2)) P_{in} \quad (3)$$

The normalization factor  $K_\gamma$  is defined as [24]

$$K_\gamma = \alpha^2 P_{in} / P_{out} = \alpha^2 / (1 - \exp(-\gamma^2)) \quad (4)$$

In case of filtering ( $J > 1$ ), the output power is further reduced by a factor  $\beta$ [24]. We assume perfect synchronization, perfect carrier recovery and no residual ICI (Inter Carrier Interference) thanks to the guard interval. Finally the discrete-time received signal after DFT and symbol interleaving is modelled as

$$Y_k = H_k (\alpha X_k + D_k) + Z_k \quad (5)$$

where  $H_k$  is the complex channel gain corresponding to the  $k^{th}$  sub-carrier and  $Z_k$  is the complex AWGN noise, with zero mean and variance  $\sigma_z^2$ . The channel gain is assumed to be perfectly known. According to the statistical model, when  $N$  increases, the  $D_k$  samples approach i.i.d. complex Gaussian random variables with zero mean and variance  $\sigma_D^2 = (1 - K_\gamma) P_{out}$ . In this model,  $D_k$  is assumed to be uncorrelated with  $X_k$ . However, the particular frame  $\mathbf{D}_m = [D_{0,m}, \dots, D_{N-1,m}]$  is uniquely defined by the non linear clipping function  $g$  and the particular OFDM frame  $\mathbf{X}_m$  such that  $\mathbf{D}_m$  is given by

$$\mathbf{D}_m = \mathbf{X}_m^c - \alpha \mathbf{X}_m \quad (6)$$

where  $\mathbf{X}_m^c = [X_{0,m}^c, \dots, X_{N-1,m}^c]$  is the DFT of  $\mathbf{x}^c = [x_{0,m}^c, \dots, x_{N-1,m}^c]$ . The signal to noise ratio is given by

$$SNR_c = E_s / N_0 = \frac{\beta P_{out}}{P_Z} \quad (7)$$

where the noise power  $P_Z$  is equal to the complex AWGN variance  $\sigma_z^2$  and  $\beta = 1$  for  $J = 1$ . Then the  $E_b / N_0$  is given by

$$E_b / N_0 = \frac{\beta P_{out}}{P_Z} \cdot \frac{1}{R \cdot m} \quad (8)$$

where  $R$  is the code rate and  $m$  is the number of bits per symbol.

#### B. SNR loss due to clipping

Without any compensation of the clipping noise, and assuming Nyquist-rate sampling, the SNR loss due to clipping is given by [24]

$$\Delta_S = 10 \log_{10} \left[ \frac{1 - K_\gamma}{K_\gamma} SNR_c + \frac{1}{K_\gamma} \right] \quad (9)$$

Thanks to this SNR loss, it is then possible to derive an upper bound of the capacity of the clipped OFDM system [22]. In [25], a deterministic model of clipping is used to derive a tight lower bound on the system capacity.

| $\gamma$ | $\Delta_g$ (dB) |
|----------|-----------------|
| 1.6      | 0.04            |
| 1.2      | 0.15            |
| 1        | 0.26            |
| 0.8      | 0.41            |
| 0.5      | 0.70            |

TABLE I  
SNR LOSS  $\Delta_g$  IN DB VERSUS CLIPPING RATIOS

In this paper we derive another lower bound on the BER performance of clipped OFDM based on the statistical model expressed by (5). By assuming that the distortion term  $D_k$  can be suppressed perfectly whereas the other terms are kept the same, the corresponding signal-to-noise ratio of such an idealistic system is then given by  $SNR_g = \frac{\alpha^2 P_{in}}{P_z}$ . We define the SNR loss  $\Delta_g$  as

$$\begin{aligned}\Delta_g &= 10 \log_{10} \left( \frac{E_s/N_0}{SNR_g} \right) \\ &= 10 \log_{10} \left( \frac{P_{out}}{\alpha^2 P_{in}} \right) \\ &= -10 \log_{10}(K_\gamma)\end{aligned}\quad (10)$$

It is trivial to verify that  $\Delta_g$  can be equivalently obtained from (9) when  $SNR_c$  tends to zero

$$\Delta_g = \lim_{SNR_c \rightarrow 0} \Delta_S = -10 \log_{10}(K_\gamma) \quad (11)$$

Table I gives the SNR loss  $\Delta_g$  for several clipping ratios. The lower bound on the BER performance of clipped coded OFDM is then obtained by translating the BER performance of the unclipped coded system by the SNR loss.

### C. Iterative cancellation of clipping noise

In [23] the authors capitalize on the work of Tellado [20] and Chen [21] to propose a new version of the iterative clipping noise cancellation receiver. This version consists in: 1) applying a Maximum Likelihood (ML) demapping to the signals obtained after clipping noise cancellation in the frequency domain, 2) converting the *a posteriori* log likelihood ratio (LLR), provided by the channel decoder, into symbol estimates  $\tilde{X}_k$ , which are used to reconstruct an estimation of the clipping noise, thanks to a soft mapper. The iterative clipping noise cancellation receiver can be applied either to oversampled and filtered signals or to Nyquist-rate signals. The ML demapping of the received signal  $Y_k$  involves the computation of the LLR of the bit  $b_i$  belonging to  $X_k$  given the channel observation  $Y_k$  expressed as

$$\Lambda(b_i|Y_k) = \ln \left( \frac{\sum_{S \in \chi_1^i} \exp \left( \frac{-|Y_k - \alpha \cdot H_k \cdot S|^2}{\sigma_z^2} \right)}{\sum_{S \in \chi_0^i} \exp \left( \frac{-|Y_k - \alpha \cdot H_k \cdot S|^2}{\sigma_z^2} \right)} \right) \quad (12)$$

where  $\chi_a^i$  is the subset of the constellation alphabet  $\chi$  such that  $b_i$  is equal to  $a \in \{0, 1\}$ . It has been shown by simulations that for small clipping ratios, equal or lower than 1, the attenuation  $\alpha$  must be taken into account in the ML demapping otherwise

BER performance is strongly degraded. The iterative algorithm works as follows[23]:

Initialize  $\tilde{Y}_k^{(0)} = Y_k$ , and  $i = 1$ .

- 1) Perform ML demapping of  $\tilde{Y}_k^{(i-1)}$  based on equation (12).
- 2) Run the SISO (Soft Input Soft Output) decoding over the sequence of LLR  $\Lambda(b_i|Y_k)$ .
- 3) Compute the soft symbol estimates  $\tilde{X}_k^{(i)}$ .
- 4) Reconstruct the clipped signal  $\tilde{X}_k^{c(i)} = \alpha \tilde{X}_k^{(i)} + \tilde{D}_k^{(i)}$ , thanks to those sequential operations:  $JN$ -point IDFT, clipping,  $JN$ -point DFT, out-of-band removal,  $N$ -point IDFT.
- 5) Estimate the clipping noise in the frequency domain  $\tilde{D}_k^{(i)} = \tilde{X}_k^{c(i)} - \alpha \tilde{X}_k^{(i)}$ .
- 6) Cancel the clipping noise from  $Y_k$  to obtain  $\tilde{Y}_k^{(i)} = Y_k - H_k \cdot \tilde{D}_k^{(i)}$ .
- 7) Increment  $i$  and go to step 1).

By using equation (5), the signal after clipping noise cancellation at iteration  $i$  is given by

$$\tilde{Y}_k^{(i)} = \alpha H_k X_k + H_k (D_k - \tilde{D}_k^{(i)}) + Z_k \quad (13)$$

where  $(D_k - \tilde{D}_k^{(i)})$  is the residual clipping noise. If perfect clipping noise cancellation is obtained ( $D_k = \tilde{D}_k^{(i)}$ ) then the BER performance of the iterative receiver should achieve the lower bound defined in section III-B. Now we define the genie aided receiver as the receiver that knows perfectly the transmitted symbol  $X_k$ . By providing those symbols, instead of  $\tilde{X}_k^{(i)}$ , to the iterative algorithm, genie aided estimates  $\tilde{D}_k^{(g)}$  are computed. In figure 5, the lower bound defined in section III-B is plotted along with the BER performance (numerically evaluated) of the genie aided receiver. Simulation parameters are given in the caption. The BER performance of the unclipped system (circles) is also given for reference. The lower bound and the genie aided performance can not be distinguished in the case of  $\gamma$  equal to 1.2. For  $\gamma$  equal to 0.8, the genie aided performance diverges from the lower bound for  $E_b/N_0$  greater than 3.3 dB. Thus clipping noise can be cancelled by the genie aided receiver at low SNR.

### D. Simulation results

In order to evaluate the impact of the clipping process on the performance of the LDPC coded OFDM system for a practical - not genie - receiver, Monte Carlo simulations are performed. For the considered system, and unlike [23], iterations of the clipping noise cancellation do not bring any improvement thanks to the high error correction capability of the LDPC. Indeed, the performance of the genie aided receiver (not shown) is merged with the BER performance of the clipped LDPC coded OFDM system for the whole SNR range. Hence, in Fig. 6, simulation result of the BER performance over AWGN channel is only given for the *non* iterative receiver (no clipping noise cancellation, just soft demapping followed by LDPC decoding) for CR equal to 0.8 and 1.2. For reference, the BER performance of the unclipped system (circles) is also

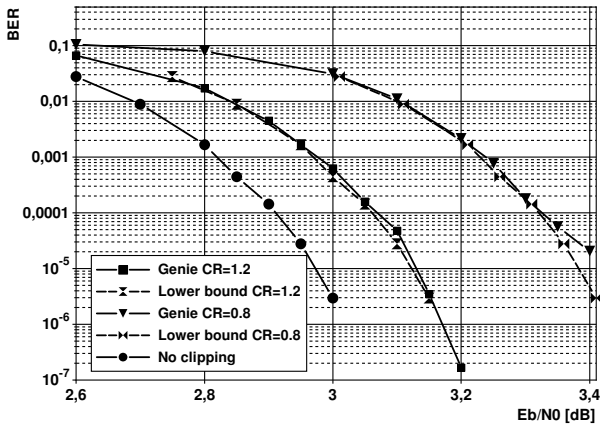


Fig. 5. BER performance of the genie aided receiver over AWGN channel ; oversampling rate equal to 4 ; 16QAM with Gray mapping, DFT size equal to 128, 32 symbols per frame, 16200 coded bits per frame ; 4/9 LDPC code, Sum Product decoding algorithm, Maximum number of LDPC decoding iterations equal to 50.

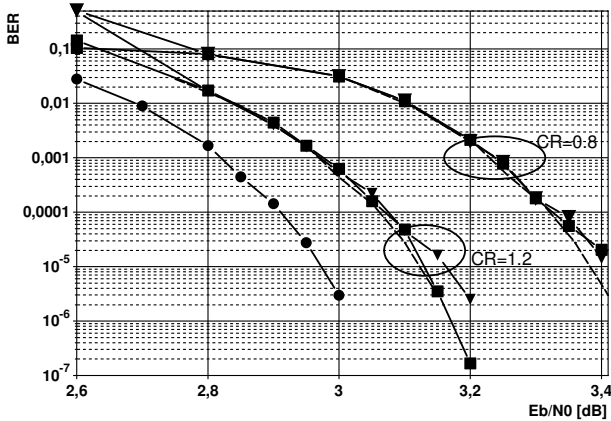


Fig. 6. BER performance of the clipped LDPC coded OFDM system over AWGN channel, oversampling rate equal to 1 (triangles) and 4 (squares) together with the unclipped coded OFDM system (circles) and the lower bounds (dashed); CR stands for  $\gamma$  ; other simulation parameters are the same as in figure 5.

shown as well as the lower bound (dashed). For a large SNR range, the performance of the clipped LDPC coded OFDM system, with ML demapping as the only mean of mitigate the clipping noise, is merged with the lower bound. The BER degradation of the clipping process is totally recovered except the power loss  $\Delta_g$ . Capacity approaching error correcting codes were already known to mitigate the clipping noise [24]. However, in this paper, a tight lower bound of the clipped LDPC coded system, valid at low SNR, is given. Then, for the considered simulation parameters, the BER performance of such a system can be kept within 0.4 dB from the performance of the unclipped system for clipping ratio greater than 0.8. Future work will consider the design of an iterative receiver based a deterministic model such as in [25].

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