Iterative Detection Techniques for Clipped OFDM Systems

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Abstract—**This paper studies orthogonal frequency-division multiplexing (OFDM) systems employing coded modulation. Clipping is applied to reduce the peak-to-average-power-ratio (PAPR) of the transmit signal. A soft compensation method is proposed to combat the clipping effect. It is shown that the proposed method can outperform conventional clipping effect mitigation methods. The impact of signaling schemes of the coded modulation on system performance is investigated. The average variance analysis and numerical results demonstrate that superposition coded modulation (SCM) together with the proposed detection technique provides a simple and efficient solution to clipped OFDM transmission.**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) is an efficient multi-carrier transmission technique. It has attracted much attention due to its distinguished features of high spectral efficiency and low receiver complexity. However, as it is well known, the OFDM suffers a high peak-to-average power ratio (PAPR) [1] problem. This makes OFDM systems more prone to the non-linear effect of transmitter/receiver devices than the conventional single-carrier schemes. Pre-distortion methods such as deliberate clipping [2]-[8] can be used to reduce the PAPR, but they unfortunately induce in-band non-linear distortion and out-of-band radiation. The latter can be treated by band-pass filtering the clipped signal before transmission, which may incur re-growth of the PAPR of the transmitted signal. In order to further reduce PAPR, an iterative clipping and filtering process can be applied [4].

The in-band non-linear distortion would degrade the bit-error-rate (BER) performance. In low-rate cases, a system works at a relatively low signal-to-noise ratio (SNR). Hence, the channel noise may dominate the performance and forward error control coding (FEC) can ensure good results. In high-rate cases with coded modulation (such as the bit-interleaved coded modulation (BICM) [9]), the system works at a higher SNR. Thus, the non-linear distortion has a more significant impact and FEC alone may not be sufficient. To alleviate this problem, we propose a soft compensation method that can be incorporated into the iterative decoding process. We show that this method can effectively recover performance and outperform other conventional methods to treat clipping effect.

Simulation results show that the performance of soft compensation depends heavily on the signaling schemes employed in the coded modulation schemes. We will present an average variance analysis to explain this interesting observation. We show that superposition coded modulation (SCM) [10]-[12] is advantageous in terms of clipping effect mitigation. The overall system performance with different signaling schemes is investigated by *extrinsic* information transfer (EXIT) charts analysis and simulation studies. We show that SCM together with the soft compensation method provides an efficient solution to OFDM transmission with low PAPR.

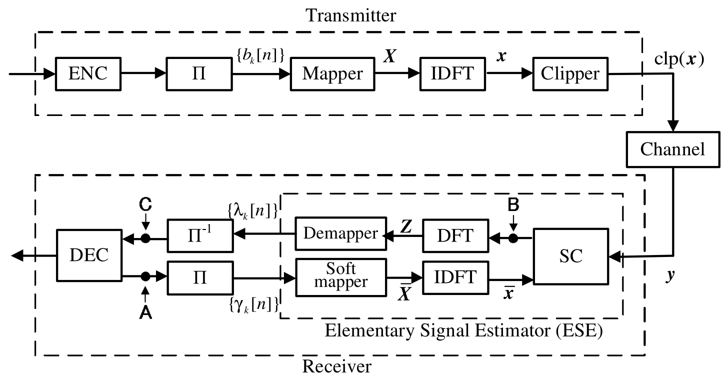

Fig. 1. Block diagram of coded OFDM systems with clipping, where Π denotes interleaver and Π^{-1} de-interleaver.

II. System Model

A. Transmitter Principles

Consider an OFDM system employing BICM. The transmitter structure is illustrated in the upper part of Fig. 1. The information bits are first encoded by a binary encoder (ENC). The resultant coded bits are then randomly interleaved and packed into groups $\boldsymbol{b}[n] = (b_1[n], b_2[n], \cdots, b_K[n])$ of K bits. The signal mapper maps each group $\boldsymbol{b}[n]$ to a symbol $X[n]$ to be carried by the nth sub-carrier.

Denote by N the number of OFDM sub-carriers and define a column vector $\boldsymbol{X} = [X[0], X[1], \cdots, X[N-1]]^{\mathrm{T}}$ of length N where $(\cdot)^{\mathrm{T}}$ denotes transpose. The time-domain signal $\boldsymbol{x}$ is produced by applying inverse discrete Fourier transform (IDFT) to $\boldsymbol{X}$,

$$\boldsymbol{x} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{X} \tag{1}$$

where $\boldsymbol{F}$ is the unitary discrete Fourier transform (DFT) matrix of size $N \times N$, and $(\cdot)^{\mathrm{H}}$ denotes the conjugate transpose. We will call $\boldsymbol{X}$ and $\boldsymbol{x}$ the frequency- and time-domain signals, respectively.

To reduce the PAPR, each entry $x[m]$ of $\boldsymbol{x}$ is deliberately

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clipped using the following clipping function,

$$\mathrm{clp}(x[m]) = \begin{cases} x[m], & |x[m]| \le A \\ Ax[m]/|x[m]|, & |x[m]| > A \end{cases} \tag{2}$$

where A is a clipping threshold and $|\cdot|$ the absolute value. The clipping ratio in decibel is given by $\mathrm{CR} = 10\log_{10}(A^2/\mathrm{E}[|x[m]|^2])$ where $\mathrm{E}[\cdot]$ denotes mathematical expectation. According to Price's theorem for Gaussian-input memoryless non-linear systems [13], we can approximately model the clipping operation as a linear process,

$$\mathrm{clp}(x[m]) = \alpha x[m] + \mathrm{clpn}(x[m]), \tag{3}$$

where $\alpha = \mathrm{E}[x^*[m]\mathrm{clp}(x[m])]/\mathrm{E}[|x[m]|^2]$ is a constant, $*$ denotes complex conjugate, and $\mathrm{clpn}(x[m]) \equiv \mathrm{clp}(x[m]) - \alpha x[m]$ is the clipping noise. The modeling in (3) is widely used to characterize the clipping effect in OFDM systems [3], [4], [6]. We will assume that the clipping distortion $\mathrm{clpn}(x[m])$ and the useful signal $x[m]$ are statistically uncorrelated when deriving the detection techniques in the next subsections.

With abuse of notations, we denote by $\mathrm{clp}(\boldsymbol{x})$ the clipped signal vector and by $\mathrm{clpn}(\boldsymbol{x}) = \mathrm{clp}(\boldsymbol{x}) - \alpha\boldsymbol{x}$ the corresponding clipping noise vector. The clipped signal $\mathrm{clp}(\boldsymbol{x})$ is then transmitted over an additive white Gaussian noise (AWGN) channel. For simplicity, we assume that the deliberate clipping is the only source of non-linear distortion in the whole system. This implies that linear power amplifiers are used. However, the derivations below can be extended to more general cases of non-linear distortions.

B. The Elementary Signal Estimator (ESE) Principle

The time-domain received signal $\boldsymbol{y}$ is given by

$$\boldsymbol{y} = \mathrm{clp}(\boldsymbol{x}) + \boldsymbol{w} = \alpha\boldsymbol{x} + \mathrm{clpn}(\boldsymbol{x}) + \boldsymbol{w} \tag{4}$$

where $\boldsymbol{w}$ is an AWGN vector. As shown in the lower part of Fig. 1, the receiver consists of an elementary signal estimator (ESE) and a decoder (DEC), connected by a pair of interleaver (Π) and de-interleaver (Π^{-1}). Let us begin with the "soft compensation" (SC) module in Fig. 1. Assume that we are given an estimate $\boldsymbol{c}$ of the clipping noise $\mathrm{clpn}(\boldsymbol{x})$. Then, the contribution of $\mathrm{clpn}(\boldsymbol{x})$ can be canceled from the received signal $\boldsymbol{y}$, resulting in

$$\boldsymbol{z} = \boldsymbol{y} - \boldsymbol{c}. \tag{5}$$

The frequency-domain counterpart $\boldsymbol{Z}$ of $\boldsymbol{z}$ is constructed by applying DFT to $\boldsymbol{z}$. From (1), (3) and (4),

$$\boldsymbol{Z} = \boldsymbol{F}\boldsymbol{z} = \alpha\boldsymbol{X} + \underbrace{\boldsymbol{F}(\mathrm{clpn}(\boldsymbol{x}) - \boldsymbol{c})}_{\boldsymbol{D}} + \boldsymbol{W}, \tag{6}$$

where $\boldsymbol{W} = \boldsymbol{F}\boldsymbol{w}$. Based on (6), the demapper (see Fig. 1) generates the *extrinsic* LLRs for each coded bit as

$$\lambda_k[n] = \ln\left(\frac{\Pr(\boldsymbol{Z}|b_k[n]=0)}{\Pr(\boldsymbol{Z}|b_k[n]=1)}\right), \forall k, \forall n. \tag{7}$$

Here (7) is evaluated ignoring the coding constraints. We adopt a simple strategy by treating $\boldsymbol{D}$ in (6) as an independent Gaussian noise. With this strategy, the demapping method used for BICM [9] can be applied here to evaluate (7).

At the start, without any information about $\mathrm{clpn}(\boldsymbol{x})$, we simply set $\boldsymbol{c} = \boldsymbol{0}$ in (6). This is a very rough starting point and the estimate $\boldsymbol{c}$ will be refined iteratively, as shown later.

C. The DEC Principle

The DEC takes the de-interleaved version of $\{\lambda_k[n]\}$ in (7) as the inputs and performs the standard *a posteriori* probability (APP) decoding. The *extrinsic* LLRs produced by the DEC (after interleaving) are denoted by

$$\gamma_k[n] = \ln\left(\frac{\Pr(b_k[n]=0)}{\Pr(b_k[n]=1)}\right), \forall k, \forall n. \tag{8}$$

In contrast to (7), (8) is evaluated based on coding constraints.

D. The Iterative Process

We now consider updating $\boldsymbol{c}$ using $\{\gamma_k[n]\}$. From (1), each entry $x[m]$ of $\boldsymbol{x}$ is a sum of N random variables. Following the central limit theorem, we model the entries of $\boldsymbol{x}$ as complex Gaussian random variables, i.e., $x[m] \sim \mathcal{CN}(\bar{x}[m], \mathrm{Var}[x[m]])$. The mean $\bar{x}[m]$ and variance $\mathrm{Var}[x[m]]$ can be estimated from the DEC feedbacks $\{\gamma_k[n]\}$ following the turbo processing principle. Each entry $c[m]$ of $\boldsymbol{c}$ can then be calculated as

$$c[m] = \mathrm{E}[\mathrm{clpn}(x[m])] = \frac{\int_{|x|\ge 0}(\mathrm{clp}(x) - \alpha x)e^{-|x-\bar{x}[m]|^2/\mathrm{Var}[x[m]]}dx}{\pi\mathrm{Var}[x[m]]}. \tag{9}$$

From [14], the method in (9) is optimal for minimizing the mean squared error $\mathrm{E}[||\mathrm{clpn}(\boldsymbol{x}) - \boldsymbol{c}||^2]$, where $||\cdot||$ denotes the Frobenius norm of a vector. In practice, (9) can be evaluated by using a look-up table method as introduced in [12].

Based on the above discussions, $\boldsymbol{c}$ is updated as follows.

(i) Generate the mean $\bar{X}[n]$ and variance $\mathrm{Var}[X[n]]$ of $X[n]$ from $\{\gamma_k[n]\}$. (We will discuss the details later.)

(ii) Generate the means and variances of the entries of $\boldsymbol{x}$ based on the relationship $\boldsymbol{x} = \boldsymbol{F}^{\mathrm{H}}\boldsymbol{X}$. More specifically, $\bar{\boldsymbol{x}} = \boldsymbol{F}^{\mathrm{H}}\bar{\boldsymbol{X}}$ where $\bar{\boldsymbol{x}}$ and $\bar{\boldsymbol{X}}$ are respectively the means of $\boldsymbol{x}$ and $\boldsymbol{X}$. The variance of $x[m]$ is approximated by $\mathrm{Var}[x[m]] \approx 1/N\sum_{n=0}^{N-1}\mathrm{Var}[X[n]]$.

(iii) Generate the soft estimate of the clipping noise using (9).

Steps (i)-(iii) are performed respectively by the soft mapper, IDFT and SC modules in the ESE. Then $\{\lambda_k[n]\}$ are re-estimated using (7) based on updated $\boldsymbol{Z}$ in (6). The ESE/DEC operations outlined above can be repeated for a number of iterations. If $\boldsymbol{c} \to \mathrm{clpn}(\boldsymbol{x})$, then the clipping effect can be mitigated. The above is referred to as soft compensation.

We now show the details of evaluating $\bar{X}[n]$ and $\mathrm{Var}[X[n]]$ in Step (i) above. For brevity, we ignore the symbol index n in the discussions below. Let $b = (b_1, b_2, \cdots, b_K)$ (with $b_k \in \{0,1\}$) be a binary K-tuple and s its image in the complex plane following a mapping rule. Let $\mathcal{B}$ and $\mathcal{S}$ be the set of b and s, respectively. Then the mapping rule can be denoted by

$\Phi : \mathcal{B} \rightarrow \mathcal{S}$. Clearly, the signal constellation $\mathcal{S}$ and mapping rule Φ jointly define the signaling scheme. Some examples of $(\Phi, \mathcal{S})$ for BICM can be found in [9]. Another example is SCM [10]-[12] that generates s as

$$s = \sum_{k=1}^{K} \beta_k (-1)^{b_k}, \tag{10}$$

where $\{\beta_k\}$ are complex constants. The operation in (10) will be referred to as the SCM signaling scheme in this paper. (Note that the SCM scheme can also be applied to BICM [15].) Given the *a priori* LLRs $\{\gamma_k\}$ from the DEC, the random variables $(b_1, b_2, \cdots, b_K)$ are characterized by

$$\Pr(b_k = 0) = 1 - \Pr(b_k = 1) = \frac{e^{\gamma_k}}{1 + e^{\gamma_k}}, \forall k. \tag{11}$$

Then the mean and variance of X can be computed as

$$\begin{aligned} \bar{X} &= \sum_{s \in \mathcal{S}} \Pr(s) s, \\ \mathrm{Var}[X] &= \sum_{s \in \mathcal{S}} \Pr(s) |s - \bar{X}|^2, \end{aligned} \tag{12}$$

where $\Pr(s) = \Pr(b_1)\Pr(b_2) \cdots \Pr(b_K)$ denotes the probability of the constellation point s.

III. Impact of Signaling Schemes

We observed from simulation results that the performance of the soft compensation scheme outlined in Section II depends heavily on the signaling methods used at the transmitter. We now explain this interesting observation based on variance analysis. We will show that SCM [10]-[12] is advantageous in terms of clipping effect mitigation.

A. Variance Analysis

Let us return to the detection based on (6). In order to evaluate (7), we treat the residual clipping noise $\boldsymbol{D} = \boldsymbol{F}(\mathrm{clpn}(\boldsymbol{x}) - \boldsymbol{c})$ as an additive noise. We want to minimize the residual clipping noise power $\mathrm{E}[||\boldsymbol{D}||^2]$. Recall from Section II-D that $\boldsymbol{c}$ is estimated from $\{\bar{X}, \mathrm{Var}[X]\}$, and, in turn, $\mathrm{Var}[X]$ is related to the signaling scheme $(\Phi, \mathcal{S})$ through (12). These connections indicate that the selection of $(\Phi, \mathcal{S})$ may influence system performance, which is indeed the case as discussed below.

We propose to use average variance defined by $\mathrm{E}_\gamma[\mathrm{Var}[X]]$ as a measure of the performance in estimating $\{X\}$, where the expectation is with respect to the distribution of the *a priori* LLRs $\{\gamma_k\}$. Clearly, a smaller $\mathrm{E}_\gamma[\mathrm{Var}[X]]$ indicates more accurate estimates of $\{X\}$ and so the clipping distortion. We now show some interesting properties related to the minimization of $\mathrm{E}_\gamma[\mathrm{Var}[X]]$.

Assumption 1: The signaling $(\Phi, \mathcal{S})$ is unbiased and with unit average power:

$$2^{-K} \sum_{s \in \mathcal{S}} s = 0, \quad 2^{-K} \sum_{s \in \mathcal{S}} |s|^2 = 1. \tag{13}$$

Assumption 2: The elements of $\{\gamma_k\}$ are independent, identically distributed (i.i.d.). The statistics of the *a priori* probabilities are symmetric:

$$\mathrm{E}_\gamma [\Pr(b_k = 0)] = \mathrm{E}_\gamma [\Pr(b_k = 1)] = 1/2, \forall k, \tag{14}$$

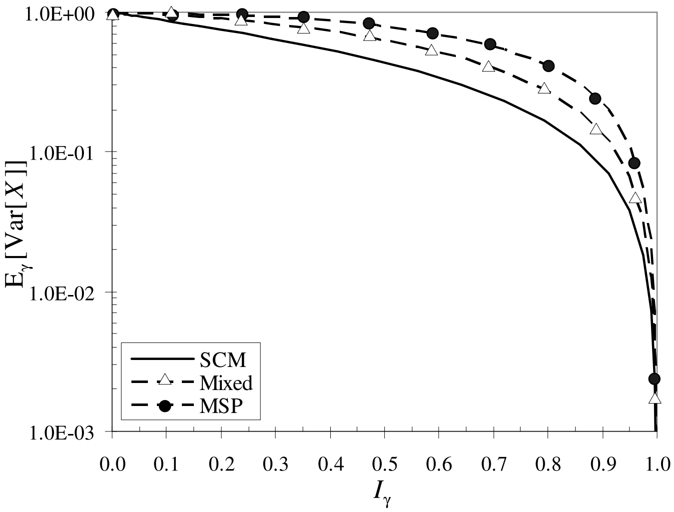

Fig. 2. Impact of signaling schemes on $\mathrm{E}_\gamma[\mathrm{Var}[X]]$. This figure is generated by modeling the *extrinsic* LLRs $\{\gamma_k\}$ as independent realizations of a consistent Gaussian random variable [16]. For the SCM scheme, $K = 4, \beta_1 = i\beta_2 = 1, \beta_3 = i\beta_4 = 1.05$, where $i = \sqrt{-1}$.

which also implies that there is a constant η such that

$$\mathrm{E}_\gamma[\Pr^2(b_k = 0)] = \mathrm{E}_\gamma[\Pr^2(b_k = 1)] = \eta, \forall k, \tag{15}$$

and the elements in $\mathcal{S}$ have equal occurrence probabilities

$$\mathrm{E}_\gamma[\Pr(s)] = 2^{-K}, \forall s \in \mathcal{S}. \tag{16}$$

The following two theorems state that SCM is optimal in the sense of minimizing $\mathrm{E}_\gamma[\mathrm{Var}[X]]$. (Due to space limitations, the proofs are not included here. Interested readers are referred to [17] for the proofs.)

Theorem 1: Under Assumption 1 and 2, the minimum average variance

$$\min \mathrm{E}_\gamma [\mathrm{Var}[X]] = 2 - 4\eta \tag{17}$$

where the minimization is over all possible $(\Phi, \mathcal{S})$.

Proof: See [17]. ■

Theorem 2: Under Assumption 1 and 2, for arbitrary K and arbitrary $\{\beta_k\}$, the SCM scheme defined by (10) achieves the minimum average variance.

Proof: See [17]. ■

Theorem 1 and 2 indicate that SCM leads to the most accurate estimate of the unclipped transmit signal X. Consequently, the minimum residual clipping noise power (after soft compensation) is achievable by SCM. The $\mathrm{E}_\gamma[\mathrm{Var}[X]]$ achieved by three 16-ary signaling schemes, namely, the 16-QAM with the Mixed and modified set-partitioning (MSP) mappings [9] and an SCM scheme with $K = 4$, are compared in Fig. 2. Clearly, the SCM scheme achieves the minimum average variance among the three schemes. From this, it can be expected that the soft compensation method outlined in Section II works best with the SCM scheme.

The above average variance analysis focuses on the clipping effect and the soft compensation performance. The situation can be different when we take into account the channel noise

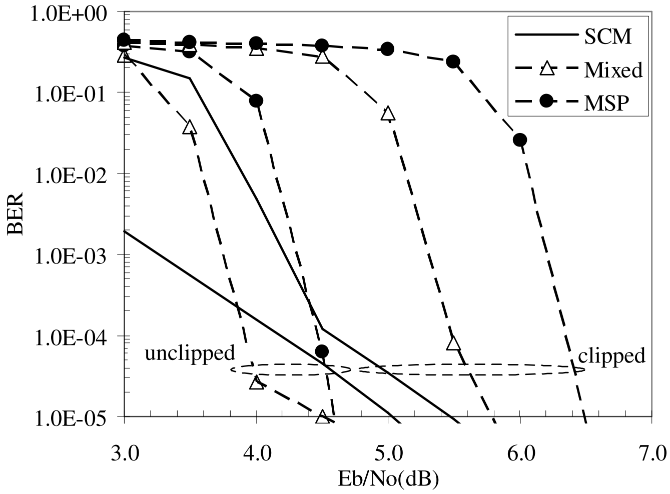

Fig. 3. Comparison of BICM-OFDM schemes with different signaling schemes when the soft compensation is used. The number of sub-carriers is $N = 256$. The clipping ratio is $\mathrm{CR} = 0$ dB. The rate-1/2 convolutional code $(23, 35)_8$ is used. The system throughput is $R = 2$ bits/symbol. The frame length is 4096. AWGN channels are assumed. The total number of iterations is 10 for all the curves.

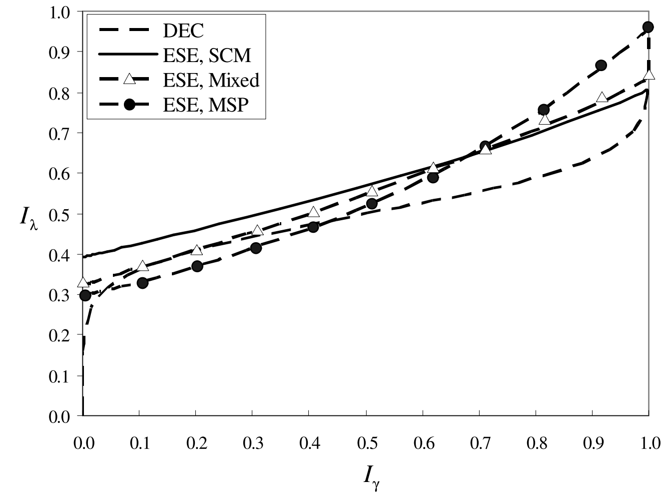

Fig. 4. EXIT chart for clipped BICM-OFDM with different signaling schemes and soft compensation. The parameters are the same as those in Fig. 3. AWGN channels with $\mathrm{E_b/N_0} = 4.5$ dB are assumed.

and consider the overall system performance. To elaborate this, we return to Fig. 1. The SCM scheme is optimal for the link from point A to point B in the iterative receiver. However, other parts of the receiver are also affected by the choice of signaling scheme for which the SCM scheme may not be optimal. For example, the SCM scheme may not be the optimal choice when the demapper is considered.

B. Examples

We now present simulation examples to verify the above analysis. Fig. 3 shows the impact of signaling schemes on BER. The BICM-OFDM schemes employing the three signaling schemes considered in Fig. 2 are simulated. From Fig. 3, the Mixed and MSP signalling schemes yield better unclipped performance at $\mathrm{BER} = 10^{-5}$ than the SCM scheme, implying that they are more powerful in combating the channel noise. However, their performance degrades significantly when the clipping effect is present. The SCM scheme provides more robustness against the clipping effect, which confirms the analysis in Section III-A.

We can study the overall convergence behavior of the iterative decoding through the EXIT chart technique [16]. For this purpose, we assume infinite interleaving lengths. Define the average mutual information between the coded bits $\{b_k\}$ and the *extrinsic* LLRs as $I_\lambda \equiv I(b_k, \lambda_k)$ and $I_\gamma \equiv I(b_k, \gamma_k)$, where $I(\cdot)$ represents the mutual information function. The EXIT functions $I_\lambda = \mathrm{T_{ESE}}(I_\gamma, \mathrm{E_b/N_0})$ and $I_\gamma = \mathrm{T_{DEC}}(I_\lambda)$ are then used to characterize the ESE and DEC, respectively, where $\mathrm{E_b/N_0}$ denotes the ratio of energy per bit to the noise power spectral density. They are generated by the Monte-Carlo simulation, similar to [16]. In general, a larger I_λ (or I_γ) implies that the *extrinsic* information produced by the ESE (or the DEC) are more reliable. Note that I_λ characterizes the joint effect of the soft compensation and demapping.

In Fig. 4, the EXIT curves are compared for the three signaling schemes considered in Fig. 3. The SCM scheme achieves the largest I_λ at low I_γ but the other two schemes take over at high I_γ. This observation can be explained as follows: Recall that soft compensation is used. Suppose that the DEC feedbacks are perfect. The clipping effect can then be fully mitigated. In this case, the performance comparison follows the unclipped scenario that the Mixed and MSP signaling schemes are optimized for. The SCM scheme is a special choice among all possible signaling schemes for BICM with iterative decoding, and so, it may not be the best one. However, when clipping noise is present, the SCM scheme can perform better, as demonstrated by the EXIT curves in Fig. 4 as well as by the simulation results in Fig. 3.

The observation that the SCM curve in Fig. 4 is relatively low at the high I_γ range also suggests a possible error floor problem. We have also observed that this problem is only noticeable in AWGN channels. In fading channels, relatively higher power is required to ensure satisfactory performance and errors are mostly caused by deep fades. In the EXIT chart, a deep fade can be viewed as an event of the ESE curve moving downwards, which causes the closure of the decoding tunnel between the ESE and DEC curves. Then the performance is mostly determined by the width of the decoding tunnel at relatively low I_γ values. From Fig. 4, we can see that the SCM scheme is advantageous in this aspect.

IV. Comparison of Detection Methods

Next we compare different clipping effect compensation strategies. The basic principle of the discussions so far is to estimate the mean of clipping distortion (denoted by $\boldsymbol{c}$) and cancel it as in (5). We proposed to generate $\boldsymbol{c}$ using (9) based on the distribution of clipping distortion. For this purpose, a look-up table can be used to characterize the relationship between $\mathrm{E}[\mathrm{clpn}(x[m])]$ and ($\bar{x}[m]$, $\mathrm{Var}[x[m]]$).

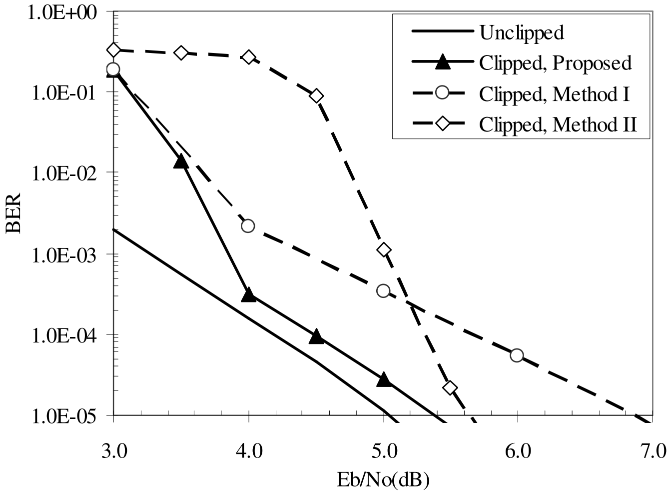

Fig. 5. Comparison of different detection methods for a BICM-OFDM system employing the SCM scheme. The number of sub-carriers is $N = 256$ and the clipping ratio $\mathrm{CR} = 1$ dB. Other parameters for simulations are the same as those in Fig. 3.

Alternatively, we can also use the following two simpler options:

- Method I: We simply treat the clipping noise $\mathrm{clpn}(\boldsymbol{x})$ in (4) as a Gaussian noise and set the clipping noise estimate $\boldsymbol{c} = \boldsymbol{0}$ in (5). (The variance of the clipping noise is a function of the clipping parameters and can be pre-calculated.) This method is similar to the approach used in [3].
- Method II: We adopt the following approximation:

$$\boldsymbol{c} = \mathrm{E}[\mathrm{clpn}(\boldsymbol{x})] \approx \mathrm{clpn}(\bar{\boldsymbol{x}}).$$

  This is similar to the soft-decision-aided clipping noise cancelation method in [8].

All these methods have roughly the same computational complexity except that the proposed method based on (9) requires more memory to store a look-up table. We will show below that this extra cost can be justified by considerable performance gain.

We now compare the performance of the above three methods. A clipped BICM-OFDM system employing the SCM scheme is used as an example. The size of the look-up table used in the proposed method is 20×20. The simulated performance with different detection methods is compared in Fig. 5. It is seen that the proposed method consistently outperforms the other two alternatives over the whole SNR range. This is because the proposed method estimates $\mathrm{clpn}(\boldsymbol{x})$ as its mean $\mathrm{E}[\mathrm{clpn}(\boldsymbol{x})]$, which is optimal when the *a priori* distribution of $\boldsymbol{x}$ is available. We can show that the advantage of the proposed method can be more evident when the clipping ratio is decreased or the transmission rate is increased. The proposed method is also advantageous when other signaling schemes are employed, but we will not present the results due to space limitations.

In this paper, only examples for AWGN channels are considered. Frequency-selective fading channels are of more practical interest. We can show that the proposed soft compensation method can be easily extended to the latter case. The advantage of the SCM scheme for clipping effect mitigation is also evident in frequency-selective fading channels. (See [17].)

V. Conclusion

We have studied coded OFDM systems involving clipping and iterative decoding. A soft compensation technique is developed to combat the clipping effect. The impact of signaling schemes on the performance is examined. The analysis and numerical results show that the proposed method can provide improved performance and the SCM scheme is more robust to clipping effect than other alternatives.

Acknowledgment

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