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중첩 부호의 반복 복호를 위한 개선된 클리핑 기법

(Modified Clipping for Iterative Decoding of Superposition Coding)

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요 약

본 논문에서는 클리핑 설계보다 파워를 덜 잃음으로서 중첩 코딩 시스템의 반복 복호를 위한 수정 클리핑 설계를 제안한다. 중첩코딩 시스템에서 제안한 설계는[1],[2] 특히 저 클리핑율의 경우에서 같은 클리핑 비율로 PAPR과 시스템 성능면에서 좋은 성능을 보였다. 마지막으로 클리핑 잡음 때문에 성능이 열화하는 것을 경감시키기 위해서 [1],[2]에서 제안한 반복 방법으로 SISO 복호알고리즘과 결합된 연 보상 알고리즘을 제안한다. 시뮬레이션 결과들은 제안한 설계로 대다수의 성능 손실이 회복될수 있다는 것을 보였다.

Abstract

In this paper, we propose a modified clipping scheme for iterative decoding of superposition coding system by losing less power than clipping scheme. Our proposed scheme in superposition coding systems[1]. [2] shows good performance in peak-to-average power ratio (PAPR) and system performance with the same Clipping Ratio especially in low Clipping Ratio case. Finally in order to alleviate the performance degradation due to clipping noises, we combine a soft compensation algorithm that is combined with soft-input-soft-output (SISO) decoding algorithms in an iterative manner proposed by [1][2]. Simulation results show that with the proposed scheme, most performance loss can be recovered.

Keywords: clipping, superposition, iterative decoding

I. Introduction

The traditional trellis coded modulation (TCM)[3] is uniformly distributed in a spaced constellation for every signaling point. In an additive white Gaussian noise (AWGN) channel, there is an asymptotical gap of about 1.53 dB (the so-called shaping gap) between the achievable performance of TCM (and other schemes based on uniform signaling[4~5]) and the channel capacity[6]. To narrow this gap, Gaussian signaling (that produces signals with Gaussian distribution) can be applied using shaping techniques, such as assigning non-uniform probabilities on different signaling points[6]. The related advantage is referred to as shaping gain[7]. 1Superposition coding is an alternative approach to bandwidth efficient coded modulation[8]. With superposition coding, several coded sequences (each referred to as a layer) are linearly superimposed before transmission. Consequently, the transmitted signal exhibits an approximately Gaussian distribution that matches to an AWGN channel. This provides anlternative means to achieve shaping gain. The work in [8] shows that such a concept is realizable with practical encoding and decoding methods. Simulation results show that a superposition coding scheme can operate within the shaping gap[8], surpassing the theoretical limit of the uniform signaling based methods. However, there is a practical concern on superposition coding: an ideal Gaussian distributed signal has an infinite peak

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power even its average power is finite. The resultant high peak-to-average power ratio (PAPR) may cause a problem for radio frequency amplifier efficiency[9]. Clipping is a straightforward technique for PAPR reduction but it may lead to performance degradation [8]. The same PAPR problem also exists in other schemes for achieving shaping gain[9]. In this paper, we will present a modified clipping issue that loss less than clipping scheme due to clipping Gaussian signals for superposition codes, that SNR(signal to noise ratio) can be improved. We will show by comparison of modified clipping and clipping of mutual information analysis that the theoretical penalty due to clipping is marginal for practical PAPR values. This demonstrates the theoretical advantage of the superposition codes. We will then combine a practical method to recover the performance loss due to clipping by exploiting the characteristics of clipping noises. It can be easily incorporated into the overall iterative receiver structure based on the low-cost multiuser detection principles developed in [10~11]. Compared with other clipping schemes, the proposed method can achieve noticeable performance improvement.

II. Transmitter and PAPR

1. Encoding

Here, we consider a $k-layer$ superposition coding system. The encoding scheme is shown in Figure.1. A binary sequence $u$is partitioned into K subsequence $\{u^{(k)}\}$. The kth subsequence $u^{(k)}$ is encoded by a binary encoder($ENC-K$) at the kth layer, resulting in a coded bit sequence $c^{(k)} = \{c_j^{(k)}\}$ of length J with $c^{(k)} = \{0,1\}$. An interleaver Π_k of length J is then applied to produce sequence $b^{(k)} = \{b_j^{(k)}\}$ which is subsequently transmitted through the channel after modulation. The randomly interleaved version $b^{(k)}$ of $c^{(k)}$, from Interleaver Π_k, is then mapped to a quadrature phase shifting keying(QPSK) sequence $x_j^{(k)} = x_{\mathrm{Re},j}^{(k)} + x_{\mathrm{Im},j}^{(k)}$where $i = \sqrt{-1}$. The subscripts

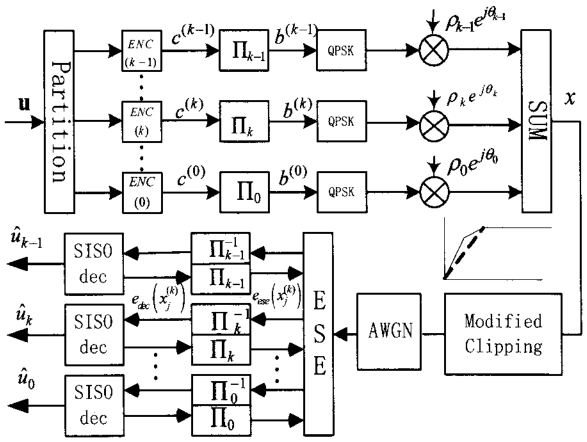

그림 1. 중첩 부호 시스템의 부호화기

Fig. 1. Encoder of a superposition coding system.

Re and Im are used to denote the real and imaginary parts of complex numbers, respectively. We will assume that $x_{\mathrm{Re},j}^{k} \in \{+1,-1\}$ and so does $x_{\mathrm{Im},j}^{k}$.

The output signal at time $j$is a linear superposition of independent coded symbols:

$$x_j = \sum_{k=0}^{K-1} \rho_k e^{i\theta_k} x_j^{(k)} \tag{1}$$

Where $\rho_k(\rho_k > 0)$ is anamplitude factor and $\theta_k(0 \le \theta_k < \pi/2)$ a rotation angle for layer-k. The overall rate is $R = 2\sum_{k=0}^{k-1} R_k$ in bits/symbol, where $R_k$is the rate of the kth binary component code. Appropriate $\rho_k's$, which are necessary to facilitate the detection at the receiver, can be determined by power allocation methods[8].

2. Peak-to-Average Power Ratio and clipping scheme

Let $E(\cdot)$denote the mathematical expectation and$|\cdot|$ the absolute value. The PAPR of x_j is defined as

$$PAPR = \frac{\max_j(|x_j|^2)}{E(|x_j|^2)} \tag{2}$$

We assume that all code bits $c_j^{(k)}$are independent, identically distributed (i.i.d) RV's (random variables). It can be verified that, the PAPR is maximized when all θ_k's are equal. As an example, for a 5-layer system with $\rho_k = \{0.1634, 0.2380, 0.3467, 0.5051, 0.7358\}$, the maximum PAPR=5.97dB is reached at $\theta_k = 0, \forall k$. On the other hand, if we set $\theta_k = k\pi/10, \forall k$, then the PAPR is reduced to 5.39dB. In order to further suppress PAPR, we clip x_j to $\overline{x_j}$ before transmission according to the rule:

$$\overline{x_j} \equiv \begin{cases} x_j, & |x_j| < A \\ Ax_j/|x_j|, & |x_j| > A \end{cases} \tag{3}$$

Where $A > 0$ is the clipping threshold (a real value). Then, the clipped signal $\overline{x_j}$ is transmitted over an AWGN channel, and the PAPR of the transmitted signal is $A^2/E(|\overline{x_j}|^2)$. The received signal can be written as $y_j = \overline{x_j} + w_j$.

Where w_j are samples of circularly symmetric complex Gaussian random process with zero mean and variance σ^2 per dimension. The ratio of energy per bit (E_b) to the noise power spectral density $(N_0 = 2\sigma^2)$ is given by $E_b/N_0 = E(|\overline{x}_j|^2)/2\sigma^2$.

3 Effect of Clipping on the Achievable Rates

We now investigate the impact of clipping on the performance limits of superposition coding systems.

The output signal of the superposition coding system is a Gussian process and its power is $E(|x_j|^2) = 2\sigma_s^2$.

The clipping ratio γ is defined as in [13].

$$\gamma @ \frac{A}{\sqrt{P_{in}}} = \frac{A}{\sqrt{2}\,\sigma_S}$$

Where $P_{in} = 2\sigma_s^2$ is the input power of the systems signal before clipping. To simplify our computation, its distribution of envelope of the signals (Gaussian RV's) that is Rayleigh distribution: $p(|x_j|) = |x_j|/\sigma_s^2 \exp(-|x_j|^2/2\sigma_s^2)$ the total output power P, which is the sum of the signal and distortion components, is given by [14]

$$\begin{aligned} P &= E_{\overline{x_j}}\left[\overline{x_j}^2\right] \\ &= \int_0^\infty \overline{x_j}^2 P\left(\overline{x_j}\right) d\overline{x_j} \\ &= \left(1-e^{-\gamma^2}\right)P_{in} = \left(1-e^{-\gamma^2}\right)\cdot 2\sigma_s^2 \end{aligned} \tag{4}$$

The relation of unclipped and clipped signals can be obtain by this formula, while the clipping ratio is very low then and shows very good performance in PAPR but the power of the clipped signal is in high-level loss.

Otherwise, while the clipping level is very high the PAPR performance is worse than low-level clipping but little loss in its power. Consider a complex AWGN channel $Y = X + W$, where X and Y are the input and output signal, respectively, and W an independent circularly symmetric complex Gaussian random variable with zero mean and variance σ^2 per dimension. For a given distribution of X, the maximum reliable transmission rate can be quantified by the mutual information [12],

$$I(X;\cdot\ Y) = H(Y) - H(Y|X) = -E(\log P(Y) - \log(2\pi e\sigma^2)$$

where $h(\cdot)$ is the entropy function.

$$I(X,\cdot Y) = \log(1 + E(|X|^2)/2\sigma^2) \tag{5}$$

For most cases of interest, $E(\log P(Y))$ (and hence $I(X,\cdot Y)$) can be computed by numerical integration methods. Here X is the clipped signal, the channel capacity can be calculated by (5).

$$I(X,\cdot Y) = \log(1 + (1 - e^{-\gamma^2}).SNR) \tag{6}$$

Where SNR(signal to noise ratio) is defined $SNR = \sigma_s^2/\sigma^2$.

Ⅲ. Modified clipping scheme

The power efficiency in different levels has been investigated in last section that the power of clipped Gaussian signals loss much in low γ. At first we denote that the variance of original signal x_j is $E(|x_j|^2) = 2\sigma_s^2$ and the variance of clipped signal $\overline{x_j}$

is $E(|\overline{x_j}|^2)=2\overline{\sigma_s^2}$. Obviously, the power of $\overline{x_j}$ is more lower than the power of x_j when the clipping level value A is very low. Then, we can denote that the mathematical formula of the relation between x_j and $\overline{x_j}$ is $2\overline{\sigma}_s^2 < 2\sigma_s^2$. To suppress PAPR and to improve the performance degradation due to sacrificing power of $\overline{\sigma}_s^2$, We now introduce our modified clipping scheme , that is shown in Fig.2, which enlarge the power of un-clipped signals which amplitude is lower than A. We can show the formula and Fig.2 that $E(|\overline{x}_2|^2) < E(|\overline{\overline{x}}_j|^2) \le 2\sigma_s^2$. That can be easily proved by the inequality (8).

Obviously, the useful variable for calculating the power of the transmitted signals is the envelope of the signals (Gaussian RV's) that is Rayleigh distribution. Where the distribution is given by $p(|x_j|)=|x_j|/\sigma_s^2 \exp(-|x_j|^2/2\sigma_s^2)$.

The modified clipping formula be shown below

$$|\overline{\overline{x}}_j| \equiv \begin{cases} f(|x_j|), & |x_j| \le A \\ A, & |x_j| > A \end{cases} \qquad (7)$$

$$|x_j| \le f(|x_j|) < A$$

In order to demonstrate our proposed scheme has efficiently improved performance in PAPR and mutual information closed to the channel capacity, we briefly prove our assumption using previous definitions and conclusions. Then, we can use the definition

$$PAPR = \frac{A^2}{\int_{|x|<A} |x|^2 p(x)dx + A^2 \int_{|x|<A} p(x)dx} \qquad (8)$$

Let $PAPR1$ be the mathematical equation of clipping scheme and $PAPR2$ be the mathematical equation of modified clipping scheme.

From the figure of our proposed scheme, we can obtain easily the power inequality between the clipping scheme and our modified scheme $|x|^2 < f^2(|x|) \le A^2$. By the conclusion in last section and this inequality, we can see that our proposed

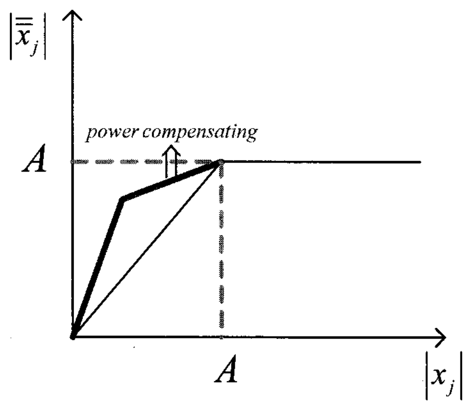

그림 2. 제안된 설계 함수
Fig. 2. Function of proposed scheme.

scheme that loss less than clipping scheme with the same clipping ratio and get the following equality:

$$\int_{|x|<A} |x|^2 p(x)dx < \int_{|x|<A} f^2(|x|)p(x)dx$$

As a comparison, we also examine the mutual information and get more improvement in it by the definition with the condition of $E(|\overline{\overline{x}}_j|^2) > E(|\overline{x}_2|^2)$.

1. Design Constraint Criterion

The achieved PAPR and mutual information for our proposed scheme that is shown in the simulations and proved is more improved. In the case of proposed scheme, we must follow power constraint at the transmitter $E(|\overline{\overline{x}}_j|^2) \le 2\sigma_s^2$, if not our design must be a failed scheme. The design criterion can be obtained by using (4)

$$\int_{|x|\le A} f^2(|x|)p(x)dx - \int_{|x|\le A} |x|^2 p(x)dx \le 2\sigma_s^2 e^{-\gamma^2} \qquad (9)$$

The optimum design can be deduced form the formula by choosing suitable value A and function $f(|x|)$ while $E(|\overline{\overline{x}}_j|^2) = 2\sigma_s^2$.As an example shown in section 1, for a 5-layer system with $\rho_i = \{0.1634, 0.2380, 0.3467, 0.5051, 0.7358\}$, the maximum PAPR=5.97dB is reached at $\theta_k = 0, \forall k$. We set A=1.26 mentioned in [1~2], with clipping scheme, and the resultant PAPR is 3.0dB. The optimum

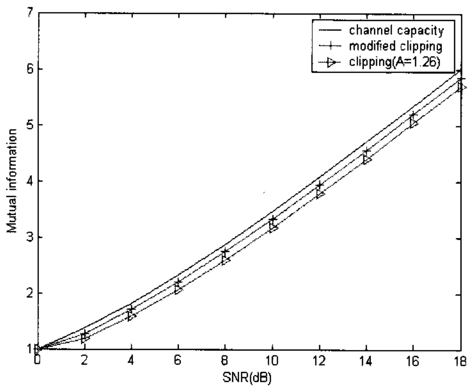

그림 3. A=1.26인 상호 정보의 비교
Fig. 3. Comparison of mutual information with A=1.26.

designed parameters and function can be get by [9] with A=1.26, and get the PAPR 1.09 dB.

Here, for our proposed scheme many function combinations can be proposed. In order to complement and compute, linear combination is introduced in our design and its linear combination function given here for a clipping level A. Here, a simple way to find the optimum clipping scheme, 1) Set a middle point $|x_j|= A/2$ between arrange from 0 to A. 2) With help of computer, a point $|\overline{\overline{x}}_j|$ can be calculated from $A/2$ to A with minimum interval $A/100$ increasing for satisfying (9). 3) Connect all points calculated by computer, following linear algebra give the function blew.

$$|\overline{\overline{x}}_j|_{opt} \equiv \begin{cases} 2|x_j| & ,0<|x_j|\le 0.63 \\ 0.33|x_j|+0.83, & 0.63<|x_j|\le 1.26 \\ 1.26 & ,|x_j|>1.26 \end{cases} \quad (10)$$

The simulation shows that we can get good mutual information performance almost closed to the channel capacity. We can construct lots of $f(|x|)$ functions following our modified scheme criterion but in order to simplify systematic complexity linear function should be proposed.

IV. Iterative Decoding

Now we turn our attention to practical superposition coding system. As discussed in previous sections, we have shown that, the performance incurred by clipping is not severe for high-level clipping threshold A. However, if the clipping effect is not treated, performance would degrade significantly even when A is moderate, as will be shown by simulation examples later. The following observations suggest possible approach toward a solution:

If $|y_j|$, the amplitude of received signal, is large, then the probability is high that x_j has been clipped.

If $|y_j|$, is small, then the clipping probability is small.

We may exploit this fact to compensate for the clipping effect. This is the underlying rationale for compensation algorithm (SCA) presented below.

1. The Basic Receiver Structure

The basic receiver structure has been introduced in Fig.1 without the compensation for clipping effect. We can treat the superposition coding system as a perfectly coordinated multiple-access system by viewing one layer as one user. The similar idea of multiuser detection principle can be applied. Specifically, due to the similarly between the superposition coding system, we employ a suboptimal iterative receiver similar to that in [10]. As illustrated in Fig.1, the receiver consist of one elementary signal estimator (ESE) and k soft-input soft-output (SISO) decoders (DECs). They are connected by the Π_k and Π_k^{-1}(de-Interleaver), operating iteratively. The messages passing between the ESE and the DECs are the so-called extrinsic information values. Since the SISO decoders perform standard a posteriori probability(APP) decoding, we will only focus on the ESE which performs SISO detection to generate the a posteriori log-likelihhood ration(LLRs) for all bits[10]. We first ignore the clipping effect and concentrate on the detection for layer-k. Rewrite (4) as

$$y_j = x_j + w_j = \rho e^{i\theta_k} x_j^{(k)} + \zeta_j^{(k)} \quad (11)$$

Where $\zeta_j^{(k)} = \sum_{k' \neq k} \rho e^{i\theta_k} x_j^{(k')} + w_j$, represent the interference-plus-noise with respect to $x_j^{(k)}$, we generate

$$\hat{y}_j^{(k)} = e^{-i\theta_k} y_j = \rho x_j^{(k)} + \hat{\zeta}_j^{(k)} \quad (12)$$

Where $\hat{\zeta}_j^{(k)} = e^{-i\theta_k}\zeta_j^{(k)}$. We approximate $\hat{\zeta}_j^{(k)}$ as an additive complex Gaussian variable. Then the output LLR($e_{ESE}(x_{\mathrm{Re},j}^{(k)})$ $x_{\mathrm{Re},j}^{(k)} \in \{+1, -1\}$) for can be computed as follows. ($x_{\mathrm{Im},j}^{(k)}$ can be computed in a similar way)

$$\begin{aligned} e_{ESE}(x_{\mathrm{Re},j}^{(k)}) &\equiv \ln\left(\frac{\Pr(\hat{y}_{\mathrm{Re},j}^{(k)} \mid x_{\mathrm{Re},j}^{(k)} = +1)}{\Pr(\hat{y}_{\mathrm{Re},j}^{(k)} \mid x_{\mathrm{Re},j}^{(k)} = -1)}\right) \\ &= 2\rho_k \frac{\hat{y}_{\mathrm{Re},j}^{(k)} - E(\hat{\zeta}_{\mathrm{Re},j}^{(k)})}{Var(\hat{\zeta}_{\mathrm{Re},j}^{(k)})} \end{aligned} \quad (13)$$

Where $Var(\cdot)$ denoted the variance function. The computational detail for $E(\hat{\zeta}_{\mathrm{Re},j}^{(k)}) = 0$ and $Var(\hat{\zeta}_{\mathrm{Re},j}^{(k)}) = I$ are shown in [10],[11].

2. Soft Compensation algorithm

Next we derive the SCA which performs a joint process to treat the inter-layer interference and clipping noise. Again, we focus on the layer-k. We rewrite (4) as

$$y_j = x_j + z_j + w_j = \rho_k e^{i\theta_k} x_j^{(k)} + \xi_j^{(k)} \quad (14)$$

Where $\xi_j^{(k)}$ denotes the distortion consisting of two terms, $\xi_j^{(k)} \equiv \zeta_j^{(k)} + z_j$ with $\zeta_j^{(k)}$ given by (13 and z_j the clipping noise. From (3) z_j is given by

$$z_j = \begin{cases} 0, & |x_j| \le A \\ A x_j / |x_j| - x_j, & |x_j| > A \end{cases} \quad (15)$$

Consider the detection of $x_{\mathrm{Re},j}^{(k)}$ based on (12) and (13). The detection procedure has an exponential complexity and hence is not practical (especially when k is large). Instead of we can adopt the following sub-optimal SCA with complexity only slightly higher than the basic detection algorithm in Section 4.1. As will be shown later, most the performance loss can be recover by SCA. The proposed algorithm is similar to (10), we generate

$$\hat{y}_j^{(k)} = e^{-i\theta_k} y_j = \rho_k x_j^{(k)} + \hat{\xi}_j^{(k)},$$

where $\hat{\xi}_j^{(k)} = e^{-i\theta_k}\xi_j^{(k)}$. We again approximate $\hat{\xi}_j^{(k)}$ by an additive complex Gaussian variable. Similar to (12, we have

$$\begin{aligned} e_{ESE}(x_{\mathrm{Re},j}^{(k)}) &\equiv \ln\left(\frac{\Pr(\hat{y}_{\mathrm{Re},j}^{(k)} \mid x_{\mathrm{Re},j}^{(k)} = +1)}{\Pr(\hat{y}_{\mathrm{Re},j}^{(k)} \mid x_{\mathrm{Re},j}^{(k)} = -1)}\right) \\ &= -\frac{1}{2}\ln\left(\frac{\left(Var^+(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)}{\left(Var^-(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)}\right) - \frac{\left(\hat{y}_{\mathrm{Re},j}^{(k)} - \rho_k - E^+(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)^2}{2\left(Var^+(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)} \\ &\quad + \frac{\left(\hat{y}_{\mathrm{Re},j}^{(k)} + \rho_k - E^-(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)^2}{2\left(Var^-(\hat{\xi}_{\mathrm{Re},j}^{(k)})\right)} \end{aligned} \quad (16)$$

Where $E^+(\hat{\xi}_{\mathrm{Re},j}^{(k)})$ and $E^-(\hat{\xi}_{\mathrm{Re},j}^{(k)})$ denotes the means of $\hat{\xi}_{\mathrm{Re},j}^{(k)}$ under the hypothesis that $x_{\mathrm{Re},j}^{(k)}$ is either +1 or −1, respectively. $Var^+(\hat{\xi}_{\mathrm{Re},j}^{(k)})$ and $Var^-(\hat{\xi}_{\mathrm{Re},j}^{(k)})$ are defines similarly [10~11].

Recalling (15) and assuming that A is known at the receiver, we can derive the following simple HCA:

(a) Initialization:

Set the estimation of z_j as $\tilde{z}_j = 0, \forall j$.

(b) Main iteration:

i) Taking $\{\tilde{y}_j = y_j - \tilde{z}_j\}$ as the received sequence, generate a hard estimation $\tilde{u}$ of u based on (10).

ii) Taking $\tilde{u}$ as the data sequence, construct as estimation $\{\tilde{x}_j\}$ of $\{x_j\}$ using the encoder shown in Fig.1

iii) Compute $\{\tilde{z}_j\}$ using (14), and then go back to step i.

Ⅴ. Numerical Results

In this section, simulation results are shown to demonstrate the performance of the modified clipped superposition coding system with k=5. The rare-1/2 doped code [12] with dada length 10^5 is closen as the component code for each layer. We set k=5. The corresponding total rate R=5 bits/symbol, and the PAPR with out clipping is 5.4 dB. We set A=1.26. Modified clipping shown in (15) for this system to reduce the PAPR to 1.09dB. The iteration Number in the component decoders is set to 200, and the number of iteration between detection and decoding is set to 6.

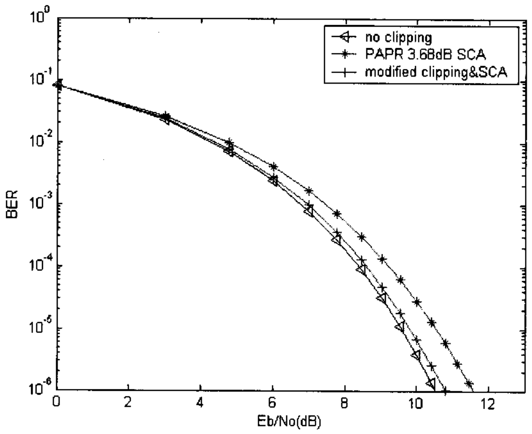

그림 4. 제안된 설계의 시뮬레이션 결과

Fig. 4. Simulation of our proposed scheme.

Ⅵ. Conclusions

We have proposed a modified clipping scheme for a peak-power-limited superposition coding system. Through computation of the mutual information closed to the power compensating in clipped input signal, we have shown that noticeable shapping gains can be achieved with reasonable clipping thresholds. In order to alleviate the clipping effect for practically coded, a soft compensation algorithm proposed in [2] is combined in our system. The simulation results shown that a good trade-off between PAPR and performance can be achieved with the proposed scheme.

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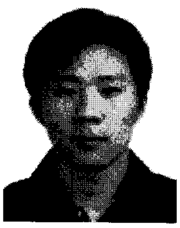

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