A Relay Assisted Low PAPR Technique for SFBC-OFDM Transmission

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ABSTRACT

The peak-to-average power ratio (PAPR) regrowth after the clipping is one main disadvantage of space-frequency block coded orthogonal frequency-division multiplexing (SFBC-OFDM). In this paper, we propose a relay assisted low PAPR technique for SFBC-OFDM transmission. For low PAPR at the source (mobile equipment), the relay processes SFBC encoding, which enables the source to transmit clipped single-input single-output (SISO)-OFDM signals without any increase of PAPR. Simulation results show that the clipped signal of proposed scheme is effectively recovered, and the proposed scheme achieves the diversity of SFBC without the complexity of multiple antennas at the source.

Key Words : OFDM, PAPR, clipping, SFBC, relay

Ⅰ. Introduction

Orthogonal frequency-division multiplexing (OFDM) has drawn significant interests as a candidate for broadband wireless communications, due to the high bandwidth efficiency and the robustness against multipath fading channels. However, high peak-to-average power ratio (PAPR) is an inherent drawback of OFDM especially in uplink transmissions. It requires large dynamic range of the transmit power amplifier and reduces the power efficiency, and thus the cost of mobile equipment is increased and the battery life time is decreased. A number of approaches have been proposed to cope with the OFDM PAPR problem[1]. Among those techniques, deliberate amplitude clipping may be one of the most effective solutions when the number of subcarriers is large[2]. Clipping, however, causes signal distortion that degrades the system performance.

Space-time and space-frequency block codes (STBC/SFBC) are simple and powerful transmit diversity techniques to combat the fading effect in wireless communications[3],[4]. Recently, clipped signal reconstruction methods for STBC/SFBC OFDMs were proposed, which is based on a scheme called iterative amplitude reconstruction (IAR) for single-input single output (SISO)-OFDM[5]. It is shown that the IAR can be easily employed for the STBC-OFDM, but it cannot be directly applied to the SFBC-OFDM because the simultaneously transmitted sequences over different antennas are dependent on each other. In order to preserve the orthogonality of transmitted sequences, a new clipped SFBC-OFDM transmitter was proposed in the paper. Although, the proposed transmitter allows the IAR to be applied to clipped SFBC-OFDM, it introduces considerable PAPR regrowth, since separately clipped signals are added before the transmission.

Employing multiple antennas in uplink transmissions is restricted, due to the limitation of size and complexity of the mobile equipment. Cooperative diversity overcomes these problems without

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additional complexity of multiple antennas and improves spectral and power efficiencies [6]. In this paper, we propose a relay assisted low PAPR technique for SFBC-OFDM transmission. In the proposed scheme, the SFBC encoding is performed at the relay, which enables the source (mobile equipment) to transmit clipped SISO-OFDM signals without any increase of PAPR and computational complexity. Simulation results show that the clipped signal of proposed scheme is effectively recovered at the destination, and the proposed scheme achieves the diversity of SFBC.

Ⅱ. Clipped SFBC-OFDM System

The SFBC-OFDM system codes QAM modulated sequence, $\{S[k]\}_{k=0}^{N-1}$, across two transmit antennas and over two adjacent subcarriers as[4]

$$\begin{bmatrix} X_1[2v] & X_1[2v+1] \\ X_2[2v] & X_2[2v+1] \end{bmatrix} = \begin{bmatrix} S[2v] & S[2v+1] \\ S^*[2v+1] & -S^*[2v] \end{bmatrix}, \quad v=0,1,\ldots,N/2-1 \qquad (1)$$

where $(\cdot)^*$ denotes the complex-conjugate operation. The discrete time OFDM samples for mth antenna, $\{x_m[n]\}_{n=0}^{N-1}$, m=1,2, are obtained by performing N-point inverse discrete Fourier transform (IDFT) on the SFBC encoded symbols $\{X_m[k]\}_{k=0}^{N-1}$. To reduce the PAPR of time-domain signals, deliberate clipping might be performed on $\{x_m[n]\}_{n=0}^{N-1}$. Then, the clipped OFDM samples for each antenna are given as

$$\bar{x}_m[n] := C_A(x_m[n]) = \begin{cases} x_m[n], & |x_m[n]| \le A, \\ Ae^{\arg x_m[n]}, & |x_m[n]| > A, \end{cases} \quad n=0,1,\ldots,N-1 \qquad (2)$$

where $C_A(\cdot)$ denotes the clipping operation with the clipping amplitude A. The clipping ratio (CR) is defined as $\gamma := A/\sqrt{P_{in}}$, and P_{in} is the average input power of $\{x_m[n]\}_{n=0}^{N-1}$. The clipping of $\{x_m[n]\}_{n=0}^{N-1}$, however, destroys the orthogonality of SFBC codeword, and thus makes it difficult to recover the clipped signals at the receiver. A

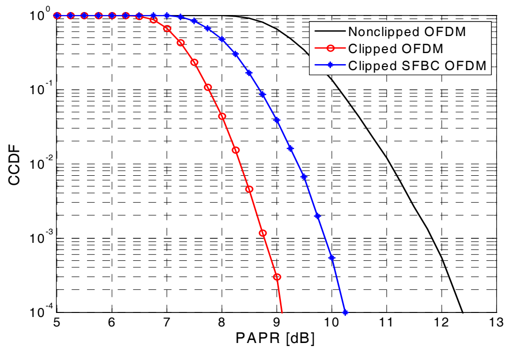

Fig. 1. The PAPR CCDF comparison of nonclipped OFDM, clipped OFDM, and clipped SFBC-OFDM (N = 1024, ϒ = 0 dB)

compromise solution is proposed in[6], where the clipping precedes the SFBC encoding as follows

$$\bar{x}_1[n] = \frac{1}{\sqrt{2}}\left(\bar{s}^e[(n)_{(N/2)}] + e^{j\frac{2\pi n}{N}}\bar{s}^o[(n)_{(N/2)}]\right)$$
$$\bar{x}_2[n] = \frac{1}{\sqrt{2}}\left(\bar{s}^{o*}[(-n)_{(N/2)}] - e^{j\frac{2\pi n}{N}}\bar{s}^{e*}[(-n)_{(N/2)}]\right) \quad n=0,1,\ldots,N-1. \qquad (3)$$

Here, $(p)_M$ represents p (modM) and $\bar{s}^e[n]$ $(\bar{s}^o[n])$ is the clipped samples of $s^e[n]$ $(s^o[n])$, which is obtained by taking N/2-point IDFT on $\{S[2v]\}_{v=0}^{N/2-1}$ $(\{S[2v+1]\}_{v=0}^{N/2-1})$. This scheme preserves the orthogonality of transmit signals $\bar{x}_1[n]$ and $\bar{x}_2[n]$. It is noted that the SFBC encoding at the transmitter introduces PAPR regrowth, due to the addition of separately clipped signals. In Fig. 1, we have compared the PAPR complementary cumulative distribution functions (CCDFs) of nonclipped OFDM, clipped OFDM, and clipped SFBC-OFDM with N=1024 at $\gamma = 0\,dB$. The ideal bandlimited analog OFDM signals are approximated by oversampling the discrete signals by a factor of sixteen. It is shown that the PAPR of clipped OFDM is 3.3 dB lower than that of nonclipped OFDM at CCDF = 10^{-4}. However, the PAPR of clipped SFBC-OFDM is 1.1 dB re-increased compared to clipped OFDM. This is a detrimental problem especially in uplink transmissions.

Ⅲ. Proposed Low PAPR Technique for SFBC-OFDM Transmission

In order to prevent the PAPR regrowth at the source (mobile equipment), the transmit signals should not be added after the clipping. By using a multi-node cooperation, we perform the SFBC encoding at the relay, which enables the source to transmit clipped SISO-OFDM signals without any increase of PAPR and computational complexity. Fig. 2 shows a conceptual block diagram of the proposed scheme. The clipping is performed at the source as $\bar{s}[n]=C_A(s[n])$, where $s[n]$ is obtained by taking N-point IDFT on the QAM modulated sequence $\{S[k]\}_{k=0}^{N-1}$. In the first time slot, the source transmits $\mathbf{x}_S^{(1)}=\bar{\mathbf{s}}$ $\left(:=\left[\bar{s}[0],\bar{s}[1],\ldots,\bar{s}[N-1]\right]^T\right)$ to the relay and destination simultaneously after appending a cyclic prefix (CP). At the relay and destination, removing the CP, the received signal is given by

$$\mathbf{r}_R=\sqrt{E_{SR}}\mathbf{H}_{SR}\mathbf{x}_S^{(1)}+\mathbf{n}_R \tag{4}$$

$$\mathbf{r}_D^{(1)}=\sqrt{E_{SD}}\mathbf{H}_{SD}\mathbf{x}_S^{(1)}+\mathbf{n}_D \tag{5}$$

where $\mathbf{n}_R$ and $\mathbf{n}_D$ are complex additive white Gaussian noise (AWGN) vectors with each entry having a zero mean and variance of $N_0/2$ per dimension. In this paper, subscripts S, R, and D stand for the source, relay, and destination nodes, respectively. E_{AB} represents the average energy available at the receiving node B. The channel impulse response (CIR) for the transmitting node A to receiving node B is given by an $N\times 1$ vector, $\mathbf{h}_{AB}=[h_{AB}[0], h_{AB}[1], \ldots, h_{AB}[L_{AB}], 0, \ldots, 0]^T$, where L_{AB} denotes the channel memory length. $\mathbf{H}_{AB}$ is an $N\times N$ circulant channel matrix with entries $[\mathbf{H}_{AB}]_{k,l}=\mathbf{r}_{AB}[(k-l)_N]$. The received signal, $\mathbf{r}_R$ is normalized as $\check{\mathbf{r}}_R:=\mathbf{r}_R/\sqrt{E_{SR}+N_0}:=\beta_R\mathbf{r}_R$ to ensure unit average energy, and the DFT of $\check{\mathbf{r}}_R$ is expressed as

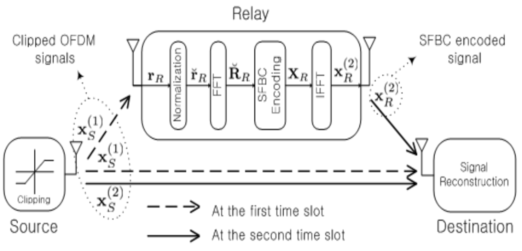

Fig. 2. The conceptual block diagram of proposed scheme

$$\check{\mathbf{R}}_R=\mathbf{F}\check{\mathbf{r}}_R=\beta_R\sqrt{E_{SR}}\Lambda_{SR}\bar{\mathbf{S}}+\beta_R\mathbf{N}_R \tag{6}$$

where $\mathbf{F}$ is a normalized DFT matrix, $\Lambda_{SR}:=\mathbf{F}\mathbf{H}_{SR}\mathbf{F}^{\mathbf{H}}$, and $\mathbf{N}_R=\mathbf{F}\,\mathbf{n}_R$. Λ_{SR} is an $N\times N$ diagonal matrix whose diagonal entries are the channel frequency response (CFR) between the nodes S and R. $(\cdot)^H$ denotes the complex -conjugate transpose.

Following [2], the clipped frequency domain signal, $\bar{\mathbf{S}}\,(=\mathbf{F}\bar{\mathbf{s}})$, can be modeled as the aggregate of an attenuated signal component and clipping noise

$$\bar{S}[k]=\alpha S[k]+D[k],\quad k=0,1,\ldots,N-1 \tag{7}$$

where the attenuation factor α is a function of γ, and given as

$$\alpha=1-e^{-\gamma^2}+\frac{\sqrt{\pi}\gamma}{2}\,\mathrm{erfc}(\gamma). \tag{8}$$

As the number of subcarriers increases and the CR value decreases, $\{D[k]\}_{k=0}^{N-1}$ approach complex Gaussian random variables with a zero mean and variance of $\sigma_D^2\left(=P_{in}(1-e^{-\gamma}-\alpha^2)\right)$. The SFBC encoding of $\check{\mathbf{R}}_R$ at the relay is performed as

$$\begin{bmatrix} X_R[2v] \\ X_R[2v+1] \end{bmatrix}=\begin{bmatrix} -\check{R}_R^*[2v+1] \\ \check{R}_R^*[2v] \end{bmatrix},\quad v=0,1,\ldots,\frac{N}{2}-1 \tag{9}$$

where $\mathbf{X}_R$ is the transmit signal of the relay in frequency domain. Then, the time domain signal can be represented as

$$\begin{aligned}\mathbf{x}_R^{(2)}&=\beta_R\sqrt{E_{SR}}\mathbf{F}^H\Lambda''_{SR}\bar{\mathbf{S}}'+\beta_R\mathbf{F}^H\mathbf{N}'_R\\&=\alpha\beta_R\sqrt{E_{SR}}\mathbf{F}^H\Lambda''_{SR}\mathbf{S}'\\&\quad+\beta_R\sqrt{E_{SR}}\mathbf{F}^H\Lambda''_{SR}\mathbf{D}'+\beta_R\mathbf{F}^H\mathbf{N}'_R\end{aligned} \tag{10}$$

where

$$\begin{aligned}
\Lambda''_{SR} &:= diag(\Lambda''^{*}_{SR}[1], \Lambda''^{*}_{SR}[0], \ldots, \Lambda''^{*}_{SR}[N-1], \\
&\quad \Lambda''^{*}_{SR}[N-2]), \\
\overline{\mathrm{S}}'[k] &:= \alpha \mathrm{S}' + \mathrm{D}', \\
\mathrm{S}' &:= \left[-S^{*}[1], S^{*}[0], \ldots, -S^{*}[N-1], S^{*}[N-2]\right]^{T}, \\
\mathrm{D}' &:= \left[-D^{*}[1], D^{*}[0], \ldots, -D^{*}[N-1], D^{*}[N-2]\right]^{T}, \\
\mathrm{N}'_{R} &:= \left[-N_{R}^{*}[1], N_{R}^{*}[0], \ldots, -N_{R}^{*}[N-1], N_{R}^{*}[N-2]\right]^{T}.
\end{aligned}$$

In the second time slot, the source and relay simultaneously transmit $\mathrm{x}_{S}^{(2)}\left(=\mathrm{x}_{S}^{(1)}\right)$ and $\mathrm{x}_{R}^{(2)}$ after appending CPs with length $L=\max\left(L_{SD,}\, L_{RD}\right)$, respectively. At the destination, removing the CP, the received signal is given by

$$\mathrm{r}_{D}^{(2)} = \sqrt{E_{SD}}\mathrm{H}_{SD}\mathrm{x}_{S}^{(2)} + \sqrt{E_{SR}}\mathrm{H}_{SR}\mathrm{x}_{R}^{(2)} + \mathrm{n}_{D} \quad (11)$$

where n_{D} is a complex AWGN vector with each entry having a zero-mean and variance of $N_0/2$ per dimension.

Ⅳ. Destination Structure of Proposed Scheme

The destination structure of proposed scheme is shown in Fig. 3. In the destination, the clipped OFDM samples are recovered by using the IAR for clipped SISO-OFDM, which reconstructs the amplitudes of clipped samples by comparing the estimates of clipped and nonclipped OFDM samples[5]. The recovered samples at the first time slot cannot achieve transmit diversity. However, the recovered samples at the second time slot were transmitted from the source and relay to the destination simultaneously with the diversity of SFBC. It is noted that the recovered samples at the second time slot should be decoded to obtain the information bits, and the recovered samples at the first time slot can be used to improve the performance by using the log-likelihood ratio (LLR) combining[7] and clipping noise cancellation[8]. Thus, the procedure of signal reconstruction at the first time slot is explain as follows

1) By using the $\tilde{\mathrm{S}}^{(1)}\left(=\mathrm{R}_{D}^{(1)}=\mathrm{Fr}_{D}^{(1)}\right)$, the estimate of the clipped sample, $\hat{\bar{s}}^{(1)}[k]$, is obtained by performing IDFT on $\left\{\hat{\bar{S}}^{(1)}[k]\right\}_{k=0}^{N-1}$

$$\begin{aligned}
\hat{\bar{S}}^{(1)}[k] &= W^{(1)}[k]\,\tilde{S}^{(1)}[k] \\
&= \frac{\left(\alpha^{2}P_{in}+\sigma_{D}^{2}\right)\Lambda_{SD}^{*}[k]}{\left(\alpha^{2}P_{in}+\sigma_{D}^{2}\right)\Lambda_{SD}[k]+\sigma_{\mathrm{N}_{D}}^{2}}\tilde{S}^{(1)}[k], \quad (12) \\
&\qquad k=0,1,\ldots,N-1
\end{aligned}$$

where $\sigma_{\mathrm{N}_{D}}^{2}$ is a variance of $\mathrm{N}_{D}\left(=\mathrm{Fn}_{D}^{(1)}\right)$, and $W^{(1)}[k]$ is a minimum mean -square error (MMSE) equalizer tap coefficient for clipped OFDM sample at the first time slot. Then, $\left\{\hat{\bar{s}}^{(1)}[n]\right\}_{n=0}^{N-1}$ are stored in memory and utilized for every iteration step.

2) To obtain the initial estimate of nonclipped samples, the output of MMSE equalizer with a tap coefficient vector $\mathrm{C}^{(1)}$ is soft-demapped, deinterleaved, and fed into a maximum a-posteriori (MAP) decoder. The MMSE equalizer tap coefficient for nonclipped OFDM sample at the first time slot, $C^{(1)}[k]$, is given by

$$C^{(1)}[k] = \frac{\alpha P_{in}\Lambda_{SD}^{*}[k]}{\left(\alpha^{2}P_{in}+\sigma_{D}^{2}\right)\Lambda_{SD}[k]+\sigma_{\mathrm{N}_{D}}^{2}}, \quad (13)$$
$$k=0,1,\ldots,N-1.$$

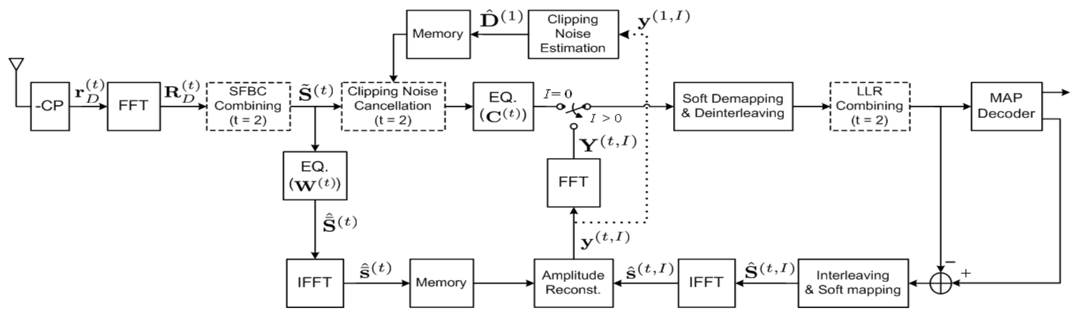

Fig. 3. The destination structure of proposed scheme

3) From the extrinsic information of MAP decoder output, estimate the nonclipped OFDM samples, $\{\hat{S}^{(t,I)}[k]\}_{k=0}^{N-1}$, $t=1,2,$ where I represents the iteration number and starts with an initial value of $I=0$.
4) IDFT is performed on $\{\hat{S}^{(t,I)}[k]\}_{k=0}^{N-1}$, $t=1,2,$ to obtain the estimates of the nonclipped samples, thus yielding $\hat{s}^{(t,I)}[n]$.
5) Clipped samples are detected by comparing the amplitude of $\hat{s}^{(t,I)}[n]$ to A in (2), Then, the amplitude of clipped samples are reconstructed, and a new sequence $\{y^{(t,I)}[n]\}_{n=0}^{N-1}$ is generated as

$$y^{(t,I)}[n]=\begin{cases}\hat{s}^{(t)}[n], & |\hat{s}^{(t,I)}[n]|\le A,\\ |\hat{s}^{(t,I)}[n]|e^{\arg\{\hat{s}^{(t)}[n]\}}. & |\hat{s}^{(t,I)}[n]|>A,\end{cases}\quad t=1,2\ \ n=0,1,\ldots,N-1. \tag{14}$$

6) The sequence, $\{y^{(t,I)}[n]\}_{n=0}^{N-1}$, is converted to the frequency domain, yielding $Y^{(t,I)}[k]$, and the LLR values as previous soft-information for the LLR combining are obtained by using the soft demapping and deinterleaving.
7) This completes the I th iteration, and for more iterations, go back to Step 3 with $I=I+1$.

In the procedure, the estimate of clipping noise in frequency domain, $\hat{\mathrm{D}}^{(1)}$, can be additionally obtained by using $\mathrm{y}^{(t,I)}$. Then, at the second time slot, the clipping noise can be cancelled by $\hat{\mathrm{D}}^{(1)}$ to estimate nonclipped samples precisely. Thus, $\hat{\mathrm{D}}^{(1)}$ is given by

$$\hat{\mathrm{D}}^{(1)}=\mathrm{F}\left(C_A\left(\mathrm{y}^{(1,I)}\right)-\alpha\mathrm{y}^{(1,I)}\right). \tag{15}$$

At the second time slot, the received signal in frequency domain, $\mathrm{R}_D^{(2)}\left(=\mathrm{Fr}_D^{(2)}\right)$ in (11), can be represented as

$$\mathrm{R}_D^{(2)}=\beta_1\Lambda_{SD}\overline{\mathrm{S}}+\beta_2\Lambda_{RD}\Lambda''_{SR}\overline{\mathrm{S}}'+\mathrm{Z} \tag{16}$$

where $\Lambda_{SD}\left(=\mathrm{FH}_{SD}\mathrm{F}^H\right)$ and $\Lambda_{RD}\left(=\mathrm{FH}_{RD}\mathrm{F}^H\right)$ are $N\times N$ diagonal matrices, $\beta_1:=\sqrt{E_{SD}}$, $\beta_2:=\beta_R\sqrt{E_{SR}\ E_{SD}}$, and $\mathrm{Z}:=\mathrm{Fn}_D+\beta_R\sqrt{E_{RD}}\Lambda_{RD}\mathrm{N}'_R$. Assuming that the CFRs between adjacent subcarriers are approximately constant, i.e. $\Lambda_{SD}[2v]\approx\Lambda_{SD}[2v+1]$, $\Lambda''_{SR}[2v]\approx\Lambda''_{SR}[2v+1]$, and $\Lambda_{RD}[2v]\approx\Lambda_{RD}[2v+1]$, the SFBC combined signals can be obtained as

$$\begin{aligned}S^{(2)}[2v]&=\beta_1\Lambda^*_{SD}R_D^{(2)}[2v]\\&\quad+\beta_2\Lambda_{RD}[2v]\Lambda''_{SR}[2v]R_D^{(2)*}[2v+1]\\&=\tilde{\Lambda}[2v]\overline{S}[2v]+\tilde{\mathrm{Z}}[2v]\\S^{(2)}[2v+1]&=\beta_1\Lambda^*_{SD}R_D^{(2)}[2v+1]\\&\quad-\beta_2\Lambda_{RD}[2v]\Lambda''_{SR}[2v]R_D^{(2)*}[2v]\\&=\tilde{\Lambda}[2v+1]\overline{S}[2v+1]+\tilde{\mathrm{Z}}[2v+1]\\&\qquad v=0,1,\ldots N/2-1\end{aligned} \tag{17}$$

where

$$\begin{aligned}&\tilde{\Lambda}[2v]=\tilde{\Lambda}[2v+1]=|\beta_1\Lambda_{SD}[2v]|^2+|\beta_2\Lambda_{RD}[2v]\Lambda''_{SR}[2v]|^2,\\&\tilde{Z}[2v]=\beta_1\Lambda^*_{SD}[2v]Z[2v]+\beta_2\Lambda_{RD}[2v]\Lambda''\\&\tilde{Z}[2v+1]=\beta_1\Lambda^*_{SD}[2v]Z[2v+1]+\beta_2\Lambda_{RD}[2v]\Lambda''_{SR}[2v]Z^*[2v].\end{aligned}$$

From (17), it is noted that $\tilde{S}^{(2)}[k]$ takes the form of a frequency domain channel observation of clipped SISO-OFDM with the CFR of $\tilde{\Lambda}[k]$ and additive noise $\tilde{Z}[k]$. Thus, the clipped OFDM samples can be recovered by proposed amplitude reconstruction, which is the same procedure at the first time slot except to clipping noise cancellation and LLR combining process. After the SFBC combining, the estimate of clipped sample at the second time slot, $\hat{\bar{s}}^{(2)}[n]$, is obtained from $\hat{\bar{S}}^{(2)}[k]$ as follows

$$\begin{aligned}\hat{\bar{S}}^{(2)}[k]&=W^{(2)}[k]\tilde{S}^{(2)}[k]\\&=\frac{\alpha^2P_{in}+\sigma_D^2}{\left(\alpha^2P_{in}+\sigma_D^2\right)\tilde{\Lambda}[k]+\sigma^2_{\tilde{\mathrm{Z}}}}\tilde{S}^{(2)}[k],\\&\qquad k=0,1,\ldots,N-1\end{aligned} \tag{18}$$

where $\sigma^2_{\tilde{\mathrm{Z}}}$ is a variance of $\tilde{\mathrm{Z}}$. Before estimating the nonclipped samples, the clipping noise of $\tilde{\mathrm{S}}^{(2)}$ is cancelled by the stored $\hat{\mathrm{D}}^{(1)}$ as

$$\begin{aligned}&\tilde{S}^{(2)}[k]-\tilde{\Lambda}[k]\hat{D}^{(1)}[k]\\&=\alpha\tilde{\Lambda}[k]S[k]+\tilde{\Lambda}[k]\left(D[k]-\hat{D}^{(1)}[k]\right)+\tilde{Z}[k]\\&\approx\alpha\tilde{\Lambda}[k]S[k]+\tilde{Z}[k]\\&\qquad\qquad k=0,1,\ldots,N-1.\end{aligned} \tag{19}$$

Then, the MMSE equalization for nonclipped samples, soft demapping, and deinterleaving processes are performed. The tap coefficient vector of MMSE, $\mathrm{C}^{(2)}$, is given by

$$C^{(2)}[k]=\frac{\alpha P_{in}}{\alpha^2 P_{in}\tilde{\Lambda}[k]+\sigma^2_{\tilde{Z}}},\ k=0,1,\ldots,N-1. \tag{20}$$

The LLR values of deinterleaved output are added with the stored LLR values at the first time slot for achieving additional gain, that is the LLR combining[7], and fed into a MAP decoder. From the extrinsic information of MAP decoder output, the nonclipped OFDM samples at the second time slot can be estimated. Using the estimates of clipped and nonclipped samples, the amplitude of clipped samples are reconstructed by (14). Then, the information bits are obtained from MAP decoder output in the Ith iteration.

Ⅴ. Simulation Results

We consider a coded OFDM with 1024 subcarriers $(N=1024)$, 16-QAM constellation, 5-MHz bandwidth, and a 1/2-rate convolutional code with constraint length of 3. All underlying links experience frequency-selective channels with the normalized Doppler frequency (f_dT_s) of 0.0001, where $S\rightarrow D$ and $S\rightarrow R$ links are modelled as six-tap typical urban (TU) channel, and $R\rightarrow D$ link is the 2-path channel with uniform power delay profile. We assume that the $S\rightarrow D$ and $R\rightarrow D$ links are balanced, i.e., perfect power control. In Fig. 4, the relay assisted low PAPR technique without iteration $(I=0)$ already outperforms the nonclipped SISO-OFDM. After the second iteration of reconstruction $(I=2)$, the proposed scheme approaches the relay assisted nonclipped SFBC-OFDM within 2 dB of SNR gap at BER=10^{-5}. It is also noted that the proposed scheme can achieve the diversity order of relay

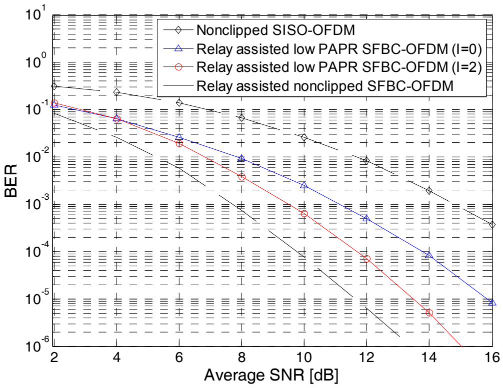

Fig. 4. BER performance of relay assisted low PAPR technique for SFBC-OFDM transmission (N = 1024, γ = 0 dB, 16-QAM, f_dT_s = 0.0001, E_{SR}/N_0 = 15 dB)

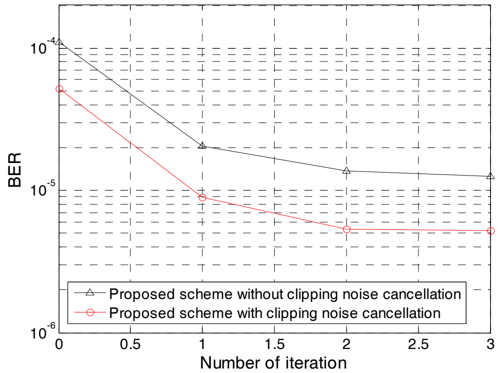

Fig. 5. Performance comparison of relay assisted low PAPR techniques with and without clipping noise cancellation (N = 1024, γ = 0 dB, 16-QAM, f_dT_s = 0.0001, E_{SR}/N_0 = 15 dB, E_{SD}/N_0 = 14 dB)

assisted nonclipped SFBC-OFDM. Fig. 5 compares the performance of relay assisted low PAPR schemes with and without clipping noise cancellation. In the second time slot, the clipping noise cancellation helps to improve the estimation of nonclipped samples, and thus achieves better BER performance at each iteration.

Ⅵ. Conclusions

This paper proposed a relay assisted low PAPR technique for SFBC-OFDM transmission. Owing to the SFBC encoding at the relay, the source of proposed scheme has the same PAPR characteristic of clipped SISO-OFDM. The corresponding destination structure was also proposed. Simulation results show that the clipped signal of proposed scheme is effectively recovered, and the proposed scheme achieves the diversity of SFBC without the complexity of multiple antennas at the source.

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