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Analysis of MIMO-OFDM Signals with Optimum Equalization

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***Abstract*:-Multi-input multi-output orthogonal frequency division multiplexing (MIMO-OFDM) has become a promising candidate for next generation broadband wireless communications. However, like a single-input single-output (SISO)-OFDM, one main disadvantage of the MIMO-OFDM is the high peak to-average power ratio (PAPR), which can be reduced either by using an amplitude clipping or through CI-OFDM (carrier interferometry orthogonal frequency division multiplexing). The proposed methods are based on the technique called iterative amplitude reconstruction (IAR). Further, we theoretically analyze the performance of IAR with optimum equalization, and also provide highly accurate channel estimation of the OFDM. Simulation results show that the proposed receivers effectively recover contaminated OFDM signals with a moderate computational complexity.**

Keywords:* Multi-input multi-output (MIMO), orthogonal frequency division multiplexing (OFDM), space-frequency block code (SFBC), iterative amplitude reconstruction (IAR), CI (carrier interferometry), PAPR (peak to average power ratio*).

I. INTRODUCTION

OFDM (orthogonal frequency division multiplexing) technique has been adopted as the standards in the several high data rate applications, such as Europe DAB/DVB (digital audio and video broadcasting) system, high-rate WLAN (wireless local area networks) such as IEEE802.11x, HIPERLAN II and MMAC (multimedia mobile access communications), and terrestrial DMB (digital multimedia broadcasting) system.
OFDM system transmits information data by many sub-carriers, where sub-carriers are orthogonal to each other and sub-channels are overlapped so that the spectrum efficiency may be enhanced. OFDM can be easily implemented by the IFFT (inverse fast Fourier transform) and FFT (fast Fourier Transform) process in digital domain, and has the property of high-speed broadband transmission and robustness to multi-path interference, frequency selective fading.
However, OFDM signal has high PAPR (peak to average power ratio) because of the superimposition of multi-carrier signals with large number of sub-carriers. The high PAPR makes the signal more sensitive to the nonlinearities of the HPA (high power amplifier) and result in signal distortion when the peak power exceeds the dynamic range of the amplifier. The non-linear distortions in the transmitted signal will lead to both in-band and out-of band emissions. The former provokes BER degradation whereas the later results in spectral spreading.
A number of approaches have been proposed to cope with the MIMO-OFDM PAPR problem. First, techniques based on the channel coding transmit only the code words with low PAPR [1], [2]. Such coding techniques offer good PAPR reduction and coding gain. The critical problem of coding approach is that for an OFDM system with large number of subcarriers, either it encounters design difficulties or the coding rate becomes prohibitively low. Phase rotation is another approach to reduce PAPR, including selective mapping (SLM) [3], and partial transmit sequence (PTS) [4]. It generates a set of sufficiently different candidate data blocks, all representing the same information as the original data block, and selects the most favorable block for transmission. Although the phase rotation works with an arbitrary number of subcarriers and any modulation schemes, it requires high computational complexity and side information. Deliberate amplitude clipping [5]-[10] may be one of the most effective solutions when the number of subcarriers is large. It clearly removes the amplitude peak, and does not introduce redundancy and power increase. Clipping, however, causes distortion that degrades the system performance. To mitigate the harmful effects of the clipping, a few techniques have been proposed. In [11], a scheme called decision aided reconstruction (DAR) is proposed, and oversampled signals are used to compensate for the SNR degradation due to the clipping [12]. However, the techniques in [11] and [12] work well at high clipping ratio (CR) values, and the technique in [12] needs bandwidth expansions.
Recently, a new kind of technique called CI-OFDM has been

widely studied [13-16]. In the CI-OFDM technique, each information symbol is sent simultaneously over all carriers and the each carrier for the symbol is assigned a corresponding orthogonal CI spreading code. This CI/OFDM system not only can reduce PAPR problem significantly but also achieve frequency diversity gains without any loss in throughput.

In this paper, we evaluate the performance of MIMO OFDM and MIMO CIOFDM system on the basis of MIMO technique theoretical analysis when HPA nonlinearity or NBI (narrow band interference) are existed. SFBC coding is applied in both MIMO OFDM system and MIMO CI-OFDM system. For CIOFDM realization, digital implemented CI-OFDM structure is used in which CI code spreading operation and carrier allocation are separately processed by simple IFFT type operation [17]. As results, MIMO CI-OFDM system outperforms MIMO OFDM significantly in the existence of both HPA nonlinearity and NBI. In this paper we have also proposed a technique called iterative amplitude reconstruction (IAR) for coded OFDM, which recovers clipped signals by comparing the estimates of clipped and non clipped OFDM samples [18].Since the IAR is an iterative technique, the reliability of initial estimates and the propagation error to the next iteration considerably affect the overall system performance.

II. ITERATIVE AMPLITUDE RECONSTRUCTION (IAR) WITH OPTIMUM EQUALIZATION

In this section, we present the IAR for the clipped signal reconstruction.

A. *The Procedure of IAR*

The procedure of IAR is explained as follows with reference to the Fig. 1.

1) Frequency domain channel observation, $R_m[n]$, is obtained by performing FFT on the discrete received samples, $\{r_m[k]\}_{k=0}^{N-1}$

2) Then, the estimate of the clipped sample, $\hat{x}_m[k]$, is obtained and stored in memory by performing IFFT $\{\hat{X}_m^{n}[n]\}_{n=0}^{N-1}$ where

$$\bar{\hat{X}}_m[n]=\frac{H_m^*[n]}{|H_m[n]|^2+N_0/P_{out}}R_m[n]$$
$$=W_m[n]R_m[n], \qquad 0\le n\le N-1$$

$W_m[n]$ is an MMSE (minimum mean square error) equalizer tap coefficient for the nth subcarrier, and (.)* denotes the complex-conjugate operation. Estimate the transmitted symbols $\{\hat{X}_m^{(I)}[n]\}_{n=0}^{N-1}$ from the $\{C_m[n]R_m[n]\}_{n=0}^{N-1}$ where I represents the iteration number and starts with an initial value of I=0. $Cm[n]$ is an optimum equalizer tap coefficient to get the non clipped OFDM samples from the received signals, which will be detailed at the end of this subsection.

3) IFFT is performed on the decisions in Step 3 to obtain the estimates of the samples before the clipping, thus yielding $\hat{x}_m^{(I)}[k]$.

4) Clipped signals are detected by comparing the amplitude of $\hat{x}_m^{(I)}[k]$ to A. Then, the amplitude of clipped signals is reconstructed, and a new sequence $\{y_m^{(I)}[k]\}_{k=0}^{N-1}$ is generated as

$$y_m^{(I)}[k]=\begin{cases}\bar{\hat{x}}_m[k], & |\hat{x}_m^{(I)}[k]|\le A,\\ |\hat{x}_m^{(I)}[k]|\exp\{\arg(\bar{\hat{x}}_m[k])\}, & |\hat{x}_m^{(I)}[k]|> A,\\ & 0\le k\le N-1|\end{cases}$$

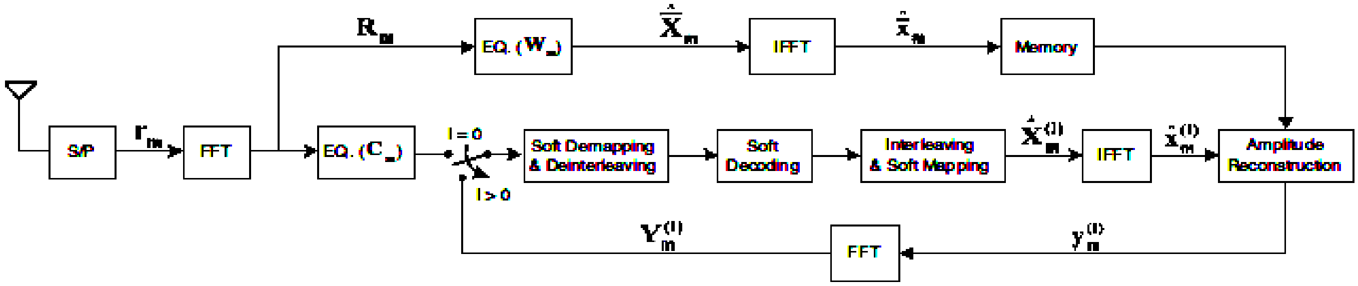
where

$$\bar{x}_m[k]=\begin{cases}x_m[k],| & x_m[k]|\le A,\\ A\exp\{\arg(x_m[k])\}, & |x_m[k]|> A,\\ & 0\le k\le JN-1\end{cases}$$

6) The sequence $\{y_m^{(I)}[k]\}_{k=0}^{N-1}$ is converted to the frequency domain, yielding $Y_m^{(I)}[n]$, and the transmitted signals $\{\hat{X}_m^{(I+1)}[n]\}_{n=0}^{N-1}$ are estimated.

7) This completes the Ith iteration, and for more iterations, go back to Step 4 with I=I+1.

Since the clipping affects not the phase but the amplitude of OFDM signals, IAR replaces only the amplitude of detected samples in Step 5. This is why we call this algorithm as iterative amplitude reconstruction. From Fig. 1 and the discussion above, each iteration of the amplitude reconstruction requires a



single pair of N-point IFFT/FFT operations

Fig. 1. The receiver structure with iterative amplitude reconstruction (IAR).

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III. IAR FOR SFBC MIMO-OFDMS

In this section, we extend the IAR into the two-by-two MIMO-OFDMs with spatial diversity, such as SFBC.
The SFBC encoder generates the coded symbol $X_{p,\,m}[n]$ for the neighboring subcarriers, $n = 2v$, $2v + 1$ & $v = 0, 1, .., N/2 - 1$, as follows

$$\begin{bmatrix} X_{1,m}[2v] X_{1,m}[2v+1] \\ X_{2,m}[2v] X_{2,m}[2v+1] \end{bmatrix} \overset{\Delta}{=} \begin{bmatrix} S_m[2v] & S_m[2v+1] \\ S_m^*[2v+1] & -S_m^*[2v] \end{bmatrix}$$

$\{x_{p,m}[k]\}_{k=0}^{N-1}$ are obtained by taking N-point IFFT on the symbols $\{X_{p,m}[n]\}_{n=0}^{N-1}$ Assuming that the Channel Frequency Responses between adjacent subcarriers are approximately constant, i.e., $H_{pq,m}[2v] \approx H_{pq,m}[2v + 1]$, the SFBC combined signals at the receiver are obtained as

$$\begin{bmatrix} \tilde{S}_m[2v] \\ \tilde{S}_m[2v+1] \end{bmatrix} = \begin{bmatrix} \sum_{q=1}^{2} (H_{1q,m}^*[2v] R_{q,m}[2v] - H_{2q,m}[2v] R_{q,m}^*[2v+1]) \\ \sum_{q=1}^{2} (H_{2q,m}^*[2v] R_{q,m}^*[2v] + H_{1q,m}^*[2v] R_{q,m}[2v+1]) \end{bmatrix}$$

$$0 \le v \le N/2-1$$

In order to recover the clipped signals of each transmit antenna, $X_{1,m}[n]$ and $X_{2,m}[n]$ are required at the receiver. However, it is difficult to derive those signals from the $\{S_m[2v], S_m[2v+1]\}$ because $S_m[2v]$ and $S_m[2v + 1]$ are mixed with the clipping noises from each transmit antenna. Therefore, it is difficult to obtain robust estimates of the clipped and non clipped OFDM samples for recovering the amplitude of clipped signals. So, we propose a new SFBC transmitter for clipped OFDM and its signal reconstruction method at the receiver.

A. *Proposed Clipping for SFBC-OFDM*

By separating $\{X_{1,m}[n]\}_{n=0}^{N-1}$ of the SFBC-OFDM into even and odd elements, time domain signals of the first antenna can be written as

$$x_{1,m}[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_{1,m}[n] W_N^{-nk}$$

$$= \frac{1}{\sqrt{N}} \sum_{v=0}^{(N/2)-1} (S_m[2v] + W_N^{-k} S_m[2v+1]) W_{(N/2)}^{-vk}$$

$$= \frac{1}{\sqrt{2}} (s_m^e[k] + W_N^{-k} s_m^0[k]), 0 \le k \le N-1$$

where $s_m^e[k]$ *and* $s_m^0[k]$ are represented as

$$s_m^e[k] = \sqrt{\frac{2}{N}} \sum_{v=0}^{(N/2)-1} s_m[2v] W_{N/2}^{-vk},$$

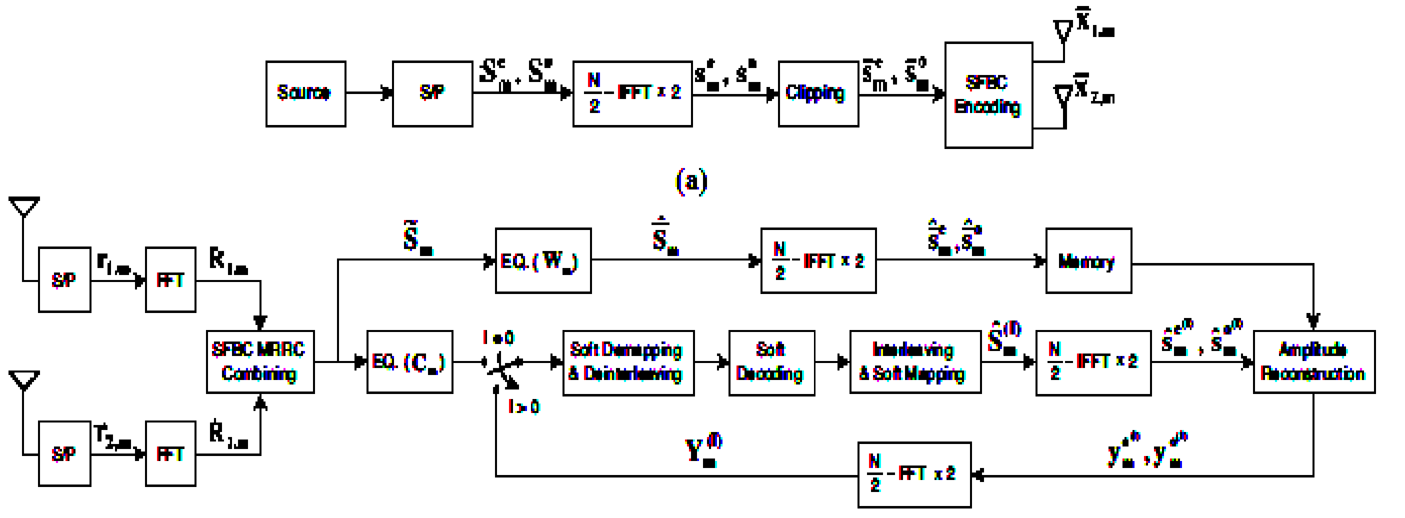
$$s_m^0[k] = \sqrt{\frac{2}{N}} \sum_{v=0}^{(N/2)-1} s_m[2v+1] W_{N/2}^{-vk}.$$

Since $s_m^e[k]$ *and* $s_m^0[k]$ are periodic in k with period $N/2$, we can replace them with $s_m^e[(k)_{(N/2)}]$ *and* $s_m^o[(k)_{N/2}]$. Here, we clip $s_m^e[(k)_{(N/2)}]$ *and* $s_m^o[(k)_{N/2}]$. instead of $x_{1,m}[k]$.Then, the transmitted signals of the first antenna can be written as

$$\bar{x}_{1,m}[k] = \frac{1}{\sqrt{2}} \left(\bar{s}_m^{e}[(k)_{(N/2)}] + W_N^{-k} \bar{s}_m^{o}[(k)_{(N/2)}] \right),$$

$$0 \le k \le N-1.$$

Because conjugate operation does not change the PAPR properties, transmitted signals of the second antenna can be derived from the first antenna signals using the SFBC and discrete Fourier transform (DFT) symmetry property i.e., $S^*[-k]_N \Leftrightarrow S^*[n]$, k, $n = 0, 1, ..., N - 1$, as follows



(b)

$$\bar{x}_{2,m}[k]=\frac{1}{\sqrt{2}}\left(s_m^{-o*}[(-k)_{(N/2)}]-W_N^{-k}s_m^{-c*}[(k)_{(N/2)}]\right),$$

$$0\le k\le N-1.$$

This implies that only $N/2$ multiplications and N additions are required to derive $\bar{x}_{2,m}[k]$ *from* $\bar{x}_{1,m}[k]$.The block

Fig.2. The transceiver structure for the proposed SFBC-OFDM with amplitude clipping. (a) Transmitter. (b) Receiver.

diagram of proposed SFBC transmitter for clipped OFDM is shown in Fig. 2 (a), which has approximately half the computational complexity of the conventional SFBC-OFDM transmitter, especially when the number of subcarriers is large. Note that the proposed clipping preserves the orthogonality of transmitted signals, and the clipped signals, $s_m^c[k]$ *and* $s_m^0[k]$ can be completely separated after the SFBC combining at the receiver, while the addition of separately clipped signals increases the PAPR of transmitted signals. Fig. 3 shows the PAPR complementary cumulative distribution functions (CCDFs) of clipped OFDM and proposed clipped SFBCOFDM with N=128 at CR=0 dB. The ideal band limited analog OFDM signals are approximated by oversampling the discrete signals by a factor of sixteen. It is observed that the PAPR of proposed SFBC-OFDM is approximately 1 dB higher than the clipped OFDM at CCDF = 10^{-3}.

B. IAR for Proposed SFBC-OFDM

In this subsection, we describe the IAR for the proposed SFBC-OFDM. The procedure of the proposed receiver is explained as follows with reference to Fig. 2 (b).

1) From the received signals, SFBC combined signals are obtained as

$$\begin{bmatrix}\tilde{S}_m[2v]\\ \tilde{S}_m[2v+1]\end{bmatrix}=\begin{bmatrix}\sum_{q=1}^{2}(H_{1q,m}^*[2v]R_{q,m}[2v]-H_{2q,m}[2v]R_{q,m}^*[2v+1])\\ \sum_{q=1}^{2}(H_{2q,m}[2v]R_{q,m}^*[2v]+H_{1q,m}^*[2v]R_{q,m}[2v+1])\end{bmatrix}$$

$$\approx\begin{bmatrix}\Lambda_m[2v]\bar{S}_m[2v]+Z_m'[2v]\\ \Lambda_m[2v+1]\bar{S}_m[2v+1]+Z_m'[2v+1]\end{bmatrix},\quad 0\le v\le N/2-1$$

$$where\ \ \Lambda m[2v]=\Lambda_m[2v+1]$$

$$=\sum_{p=1}^{2}\sum_{q=1}^{2}|H_{pq,m}[2v]|^2, Z_m'[2v]$$

$$=\sum_{q=1}^{2}(H_{1q,m}^*[2v]Zq,m[2v]-H_{2q,m}[2v]$$

$$Z_{q,m}^*[2v+1]),\ and$$

$$Z_m'[2v+1]=\sum_{q=1}^{2}(H_{2q,m}[2v]$$

$$Z_{q,m}^*[2v]+H_{1q,m}^*[2v]Z_{q,m}[2v+1])$$

2) Estimates of the clipped samples, $s_m^e[k]$ *and* $s_m^0[k]$ are obtained and stored in memory by performing two $N/2$-point IFFTs on even and odd elements of the $\{\tilde{S}_m[n]\}_{n=0}^{N-1}$, where

$$\hat{\bar{S}}_m[n] = \frac{\sigma_s^2}{\Lambda_m[n]\sigma_s^2 + N_0}\tilde{S}_m[n] = W_m[n]\tilde{S}_m[n], \\ = 0 \le n \le N-1.$$

$\{W_m[n]\}_{n=0}^{N-1}$ are the MMSE equalizer tap coefficients for the clipped OFDM samples.

3) Estimate the channel coded OFDM symbol vector $\{\hat{S}_m^{(I)}[n]\}_{n=0}^{N-1}$ from the $\{C_m[n]S_m[\tilde{n}]\}_{n=0}^{N-1}$, where $C_m[n]$ is the optimum MMSE equalizer tap coefficient for the non clipped OFDM samples, and the number of iteration starts with $I = 0$. Since the SFBC combined signals in Step 1 can be represented as

$$\tilde{S}_m[n] \approx \Lambda_m[n]\tilde{S}_m[n] + Z'_m[n] \\ = \Lambda_m[n](\alpha S_m[n] + D_m[n]) + Z'_m[n] \\ = 0 \le n \le N-1$$

and $Sm[n]$, $Dm[n]$, and $Zm[n]$ are uncorrelated, $\{C_m[n]\}_{n=0}^{N-1}$ are obtained as follows

$$C_m[n] = \frac{\alpha\Lambda_m[n]\sigma_s^2}{(\alpha\Lambda_m[n])^2\sigma_s^2 + E[|\Lambda_m[n]D_m[n]|^2] + E[|Z'_m[n]|^2]} \\ = \frac{\alpha\Lambda_m[n]\sigma_s^2}{(\alpha\Lambda_m[n])^2\sigma_s^2 + \Lambda_m^2[n]\{(1-e^{-\gamma^2} - \alpha^2)\sigma_s^2\} + \Lambda_m[n]N_0} \\ = \frac{\alpha\sigma_s^2}{(1-e^{-\gamma^2})\Lambda_m[n]\sigma_s^2 + N_0}, \qquad 0 \le n \le N-1$$

where $\sigma_s^2 (\underset{=}{\Delta} E[|S_m[n]|^2])$ is the power of $Sm[n]$.

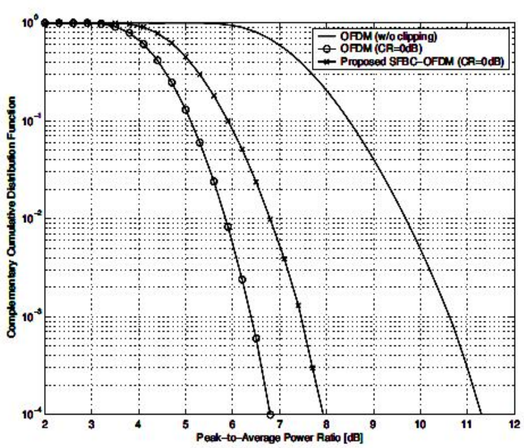


Fig.3. PAPR CCDFs of the clipped OFDM and proposed SFBCOFDM.

4) $\hat{S}_m^{e(I)}[k]$ and $\hat{S}_m^{o(i)}[k]$ are obtained by performing two $N/2$-point IFFTs on even and odd elements of the $\hat{S}_m^{(I)[n]}$ and the clipped signals are detected by comparing the amplitude of $\hat{S}_m^{e(I)}[k]$ and $\hat{S}_m^{o(i)}[k]$ to A. Then, the clipped signals are reconstructed, and two new sequences with block length $N/2$, $\{y_m^{c(I)}[k]\}_{k=0}^{N/2-I}$ and $\{y_m^{o(I)}[k]\}_{k=0}^{N/2-I}$, are generated as

$$\{y_m^{c(o)^{(I)}}[k]\} = \begin{cases} \hat{\bar{S}}_m^{c(0)}[k], & |\hat{\bar{S}}_m^{c(0)}[k]| \le A, \\ |\hat{S}_m^{c(0)}[k]|\exp\{\arg(\hat{\bar{S}}_m^{c(0)}[k])\}, & |\hat{\bar{S}}_m^{c(0)^{(I)}}[k]| > A, \end{cases} \\ 0 \le k \le N/2-1.$$

5) $y_m^{e(I)}[k]$ and $y_m^{o(I)}[k]$ are converted to the frequency domain, yielding $Y_m^I[2v]$ and $Y_m^{(I)}[2v+1]$, and the channel coded OFDM symbol vector $\{\hat{S}_m^{(I+1)}[n]\}_{n=0}^{N-1}$ is estimated.

6) This completes the Ith iteration, and for more iterations, go back to Step 4 with $I=I+1$.

From Fig. 2 (b) and the discussion above, each iteration requires two pairs of $N/2$-point IFFT/FFT operations, which has even less computational complexity than that of the IAR for SISO-OFDM. Note that we have derived the $\{C_m[n]\}_{n=0}^{N-1}$ in Step 3 to get reliable estimates of the nonclipped OFDM samples before the first iteration, and replace only the amplitude of clipped samples in Step 4, since the clipping affects not the phase but the amplitude of the signals.

IV. MIMO SFBC CI-OFDM SYSTEM

In the MIMO SFBC CI-OFDM system, the CI spreading process can be expressed as follows.

$$C_i(t) = \sum_{k=0}^{N-1} e^{j2\pi k\Delta ft}.e^{jk\Delta\theta_i} \qquad \Delta\theta_i = \frac{2\pi}{N}i, \quad i = 0,1,.....N-1$$

where , $j = \sqrt{-1}$, N is the total number of sub-carriers,

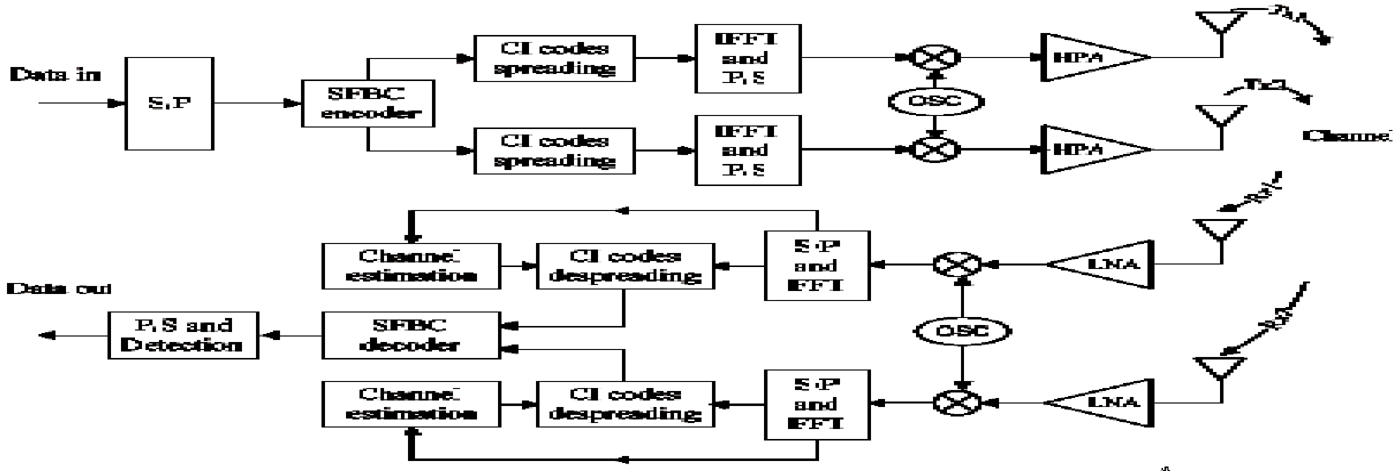


Fig.4. MIMO SFBC CI-OFDM Transceiver diagram (2Tx-2Rx).

Δf means the carrier spacing and $\Delta\theta_i$ is the assigned base spreading phase offset for the i th parallel data. Here, for general expression, we define CI spreading sequence series for the i th parallel data as

$$[C_i] = \{C_i^0, C_i^1, \ldots\ldots, C_i^{N-1}\} = \{e^{j\frac{2\pi}{N}i.0}, e^{j\frac{2\pi}{N}i.1}, \ldots\ldots, e^{j\frac{2\pi}{N}i.(N-1)}\}$$

Before passing through nonlinear HPA, the lth Tx transmitted signal for one entire MIMO SFBC CI-OFDM symbol is as follows.

$$S^l(t) = \sum_{k=0}^{N-1}\sum_{i=0}^{N-1} x_k^l . e^{j2\pi k\Delta ft} . e^{jk\Delta\theta_i} . e^{j2\pi f_0 t} . p(t)$$

$$= e^{j2\pi f_c t} . \sum_{k=0}^{N-1} S_k^l . e^{j2\pi k\Delta ft}$$

where, x_k^l is the time domain SFBC coded data on the k th carrier and lth Tx antenna, f_c is the center frequency and $P(t)$ is the pulse shaping for the bit duration T_b . Besides, here $\sum_{i=0}^{N-1} x_t^l . e^{jk\Delta\theta_i}$ is defined as s_k^I ,

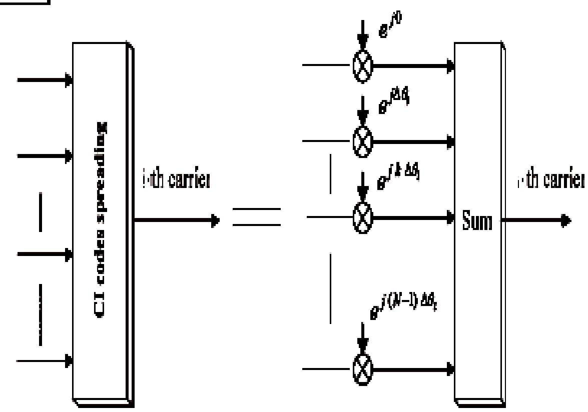


Fig. 5. CI codes spreading block

Theoretically, in the MIMO SFBC CI-OFDM receiver side, the jth Rx received signal can be expressed as follows.

$$R^j(t) = e^{j2\pi f_c t} . \sum_{l=1}^{L}\sum_{k=0}^{N-1} h_k^{lj} . s_{k.}^l e^{j2\pi k\Delta ft} + n^{lj}(t)$$

$$= e^{j2\pi f_c t} . \sum_{l=1}^{L}\sum_{k=0}^{N-1} \alpha_k^{lj} . s_{k.}^l e^{j2\pi k\Delta ft} . e^{j\phi_k^{lj}} + n^{lj}(t)$$

$$= e^{j2\pi f_c t} . \sum_{l=1}^{L}\sum_{k=0}^{N-1} \alpha_k^{lj} . e^{j2\pi k\Delta ft} . e^{j\phi_k^{lj}} . \sum_{i=0}^{N-1} x_k^l . e^{jk\Delta\theta_i} + n^{lj}(t)$$

$$= \sum_{l=1}^{L}\sum_{k=0}^{N-1}\sum_{i=0}^{N-1} \alpha_k^{lj} . x_k^l . e^{j2\pi k\Delta ft} . e^{jk\Delta\theta_i} . e^{j2\pi f_c t} . e^{j\phi_k^{lj}} + n^{lj}(t)$$

where L is the total transmit antenna number . $R^j(t)$ is the jth Rx antenna received signal, h_k^{lj} is the time domain channel response of the kth carrier from lth Tx antenna to jth Rx antenna when channel is frequency selective fading channel,

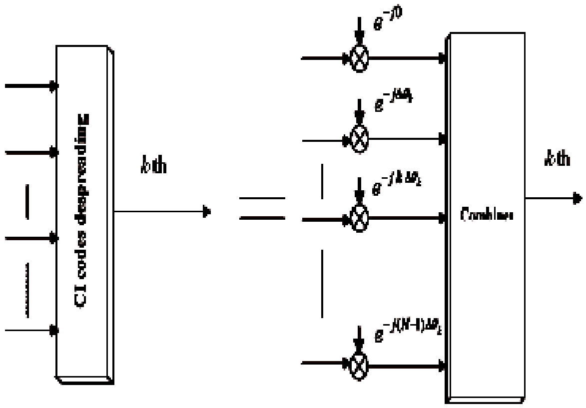


Fig.6. CI codes despreading block..

α_k^{lj} *and* ϕ_k^{lj} are the fade parameter and phase offset of h_k^{lj} and $n^{lj}(t)$ is the AWGN (additive white Gaussian noise).

The above received signal is separated into its N orthogonal sub-carriers through FFT process. After channel state estimation, each symbol stream's phase offset due to spreading is removed from each carrier by CI codes despreading. The obtained vectors from each carrier are then combined by certain combining strategy. The combining strategy is employed to help restore orthogonality between symbol streams, maximize frequency diversity benefits, and minimize interference and noise. In AWGN or flat fading channel, EGC (equal gain combining) can be used. In the frequency selective channel, MMSEC (minimum mean-square error combining) can be used to minimize inter-symbol interference from other spreading codes and noise.

V. SIMULATION RESULTS

Based on the above theoretical analysis, in order to compare the transmission performance both in the MIMO SFBC OFDM and MIMO SFBC CI-OFDM system, we evaluate the PAPR and BER of MIMO SFBC OFDM and MIMO SFBC CI-OFDM when SSPA(solid state power amplifier) is used as each transmitter's HPA or NBI is inserted to the data carriers. HPA backoff is supposed to 2, 3and 6. Total sub-carrier number interrupted by NBI is defined as p and supposed p=8, besides, JSR of NBI is supposed to 0dB or 1dB. AWGN channel is considered through the whole evaluation. The total sub carrier number is supposed to 1024 and 16QAM modulation is used in the whole simulation.

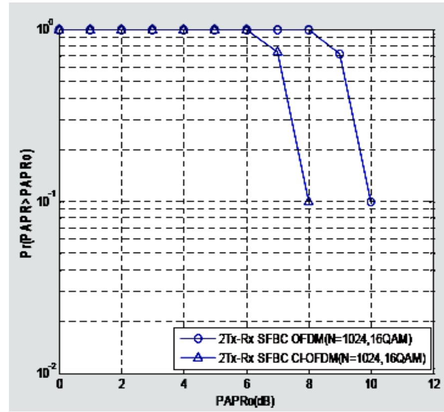


Fig.7. PAPR in MIMO SFBC OFDM and MIMO SFBC CI-OFDM (N=1024, 16QAM).

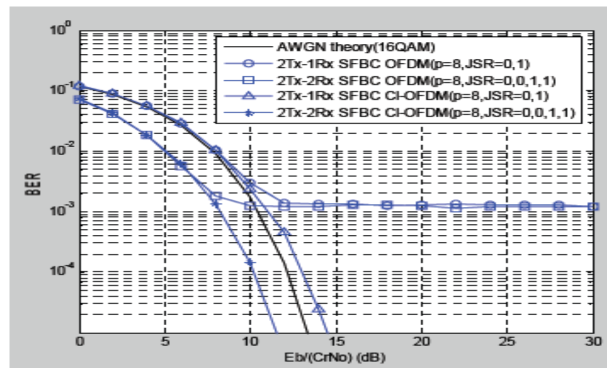


Fig8. BER in MIMO SFBC OFDM and MIMO SFBC CI-OFDM with NBI (N=1024, 16QAM, AWGN channel

Fig.7 is the PAPR of MIMO SFBC OFDM and MIMO SFBC CI-OFDM system As seen from Fig.7, there is about 2dB PAPR gain at 10^{-1} in the MIMO SFBC CI-OFDM system compared with MIMO SFBC OFDM system when total sub-carrier number is supposed to 1024.

Fig.8 is the BER of MIMO SFBC OFDM and MIMO SFBC CI-OFDM when NBI is inserted to the data carriers. As seen from the figure, MIMO SFBC CI-OFDM system can nearly compensate all the NBI affect when p is equal to 8 and JSR is 0 or 1 respectively. Only About 1dB SNR penalty is observed at BER of 10^{-4} in the 2Tx-1Rx SFBC CI-OFDM system, and even if 2dB SNR gain is observed in the 2Tx-2Rx SFBC CI-OFDM system compared with theory without NBI. However, worse than 10^{-3} of BER are achieved in the 2Tx-1Rx SFBC OFDM and 2Tx-2Rx SFBC OFDM system, and error floors occur in the both of MIMO SFBC OFDM systems.

VI. CONCLUSION

We first proposed clipped signal reconstruction methods for MIMO-OFDMs based on the IAR. Since the IAR is an iterative technique, the performance of IAR largely depends on the reliability of initial estimates and the propagation error to the next iteration. Theoretical analysis showed that the optimum equalization of IAR increases the reliability of initial estimates and its amplitude reconstruction halves the power of propagation error to the next iteration. Further, we showed that the IAR can not be directly employed for the SFBC-OFDM due to the dependency of sequences over each transmit antenna. So, we proposed a new SFBC transmitter for clipped OFDM, which has approximately half the computational complexity of conventional SFBC-OFDM and can also be applied to nonclipped OFDM. The proposed clipping preserves the orthogonality of transmitted signals, and the clipped signals were iteratively recovered at the receiver.

Then, we evaluate the system performance of MIMO SFBC OFDM and MIMO SFBC CIOFDM system on the basis of MIMO technique theoretical analysis. SFBC coding is applied in both MIMO OFDM system and MIMO CI-OFDM system.

1) From the simulation results, it is found that MIMO SFBC CI-OFDM reduces PAPR significantly compared with MIMO SFBC-OFDM system. The carefully selected CI codes result in one symbol stream's power reaching a maximum, when the powers of the remaining *N-1* symbol streams are at a minimum. Therefore, a more stable envelope, average PAPR and standard deviation of PAPR far smaller than traditional schemes can be achieved.

2) When the Narrow band interference exists, MIMO SFBC CI-OFDM system achieves considerable BER improvement compared with the MIMO SFBC-OFDM system in which error floor occurs even in high SNR. It is because that CIOFDM method has frequency diversity benefit so that it brings robustness to the narrow band interference.

3) Much better system performance can be expected by using MIMO SFBC CI-OFDM method than MIMO SFBCOFDM in the situation of existing both nonlinear HPA and NBI.

Overall, MIMO SFBC CI-OFDM system outperforms MIMO SFBC OFDM significantly when system is interrupted by the HPA nonlinearity or NBI. Therefore, the MIMO SFBC CI-OFDM method can be further applicable to the any kinds of MIMO type multi-carrier communication systems with many sub-carriers.

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