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PERFORMANCE OF CP-FREE ST-BC MIMO-OFDM SYSTEM UNDER IQ-IMBALANCE IN MULTIPATH CHANNEL

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ABSTRACT

The CP-free MIMO-OFDM system is considered to be an alternative to improve the spectral efficiency of the CP-MIMO-OFDM system. In this paper, we investigate the performance of the CP-free ST-BC MIMO-OFDM system under the effects of non-linear distortion due to HPA in the transmitter, and IQ-imbalance in the receiver. We first show that with the ST-BC encoder the constraint of the CP-free MIMO-OFDM system can be removed and back to the basic requirement of MIMO system. After compensation of the IQ-imbalance in the receiver, an equalizer under the framework of generalized sidelobe canceller (GSC) is derived for interferences suppression, and the partially adaptive (PA) array scheme is applied for complexity reduction. Simulation results show that the proposed scheme can perform very close to that with the idea CP-based ST-BC MIMO-OFDM system, and outperforms the one without compensation.

1. INTRODUCTION

The multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) system with cyclic-prefix (CP) has been widely adopted for high-speed wireless communications, due to its robustness of multipath propagation [1][2]. To increase the effective data rate, recently, CP-free single/multi-carrier systems were proposed on condition that the number of receive antennas is greater than the number of transmit antennas to provide sufficient degrees of freedom [3][4]. However, a drawback of these multicarrier signals is that it may produce high peak-to-average-power ratio (PAPR), hence introduce non-linear distortion due to high power amplifier (HPA) in the transmitter. On the other hand, MIMO systems are often realized with low cost front-end quadrature mixer architecture [5] as the zero-IF architecture. But, the quadrature mixers can be introduced into amplitude mismatches and phase mismatches, known as In-phase/Quadrature-phase (IQ) imbalance [6].

The CP-free MIMO-OFDM system, with space-time block-code (ST-BC), can provide diversity and coding gains [7]–[9]. In this paper, we first show that with the ST-BC encoder the above-mentioned constraint of the CP-free MIMO-OFDM system [3] can be removed and back to the basic requirement of MIMO system. Also, the effects of PAPR in the transmitter and IQ imbalance in the receiver are considered. To solve the problems, the adaptive Volterra predistorter [10] and the blind adaptive filtering approach with the nonlinear parameters estimation and compensation along with the power measurement [11] are employed, respectively. After the compensator in the receiver, an equalizer under the framework of generalized sidelobe canceller (GSC) [12] is derived for interferences suppression, and the partially adaptive (PA) array scheme [13] is applied for complexity reduction.

2. SYSTEM MODEL DESCRIPTION

Let us consider the CP-free ST-BC MIMO-OFDM systems with I (I=2) transmit-antennas and M receive-antennas in multipath and time varying channel with order L. Here, the block representation of OFDM signals addressed in [1] is adopted. Two N-dimensional transmitted QPSK symbol blocks $\mathbf{s}(k)$ and $\mathbf{s}(k+1)$, as depicted in Fig.1, are encoded with ST-BC encoder, where $\mathbf{s}(k)=[s(kN)\ \ s(kN+1)\dots s(kN+N-1)]^T$ and $\mathbf{s}(k+1)=[s((k+1)N)\ \ s((k+1)N+1)\dots s((k+1)N+N-1)]^T$. At time k, the time domain OFDM symbol blocks to be send from the transmit antenna is expressed as

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$$\mathbf{x}(k) = \mathbf{F}^H \mathbf{s}(k) \quad (1)$$

and

$$\mathbf{x}^{\#}(k) = \mathbf{F}^H \mathbf{s}^{*}(k) \quad (2)$$

where $\mathbf{x}(k) = [x(kN)\ x(kN+1) \ldots x(kN+Q-1) \ldots x(kN+N-1)]^T$ and $\mathbf{F}^H$ is the N-point inverse Fast Fourier transform (IFFT) matrix. Let $h_l^{(m,i)}(k)$, $l \in [0,L]$ be the lth channel impulse response between the ith transmit antenna and mth receive antenna. After demodulation, the baseband synchronized frequency-domain model at the mth receive antenna for $1 \le m \le M$, can be expressed as

$$\mathbf{z}^{(m)}(k) = \mathbf{F}\mathbf{H}_0^{(m,1)}(k)\mathbf{x}(k) + \left\{\mathbf{F}\mathbf{H}_0^{(m,2)}(k)\mathbf{x}(k+1) + \mathbf{F}\mathbf{H}_1^{(m,1)}(k-1)(-\mathbf{x}^{\#}(k-1)) + \mathbf{F}\mathbf{H}_1^{(m,2)}(k-1)\mathbf{x}^{\#}(k-2)\right\} + \mathbf{F}\mathbf{n}^{(m)}(k) \quad (3)$$

where channel matrices $\mathbf{H}_0^{(m,i)}(n)$ and $\mathbf{H}_1^{(m,i)}(n-1)$ are denoted as the upper triangular Toeplitz matrix and lower triangular matrix, respectively. The second term, on the right-hand side of (3), contains the ISI, whereas the first term is desired signal model which is a mixture of the desired tone-by-tone signals and channel tone matrix. The $N\times 1$ noise vector $\mathbf{n}^{(m)}(k)$ is with its each entry be modeled as AWGN with power σ_n^2. Since $\mathbf{H}_0^{(m,i)}(k)+\mathbf{H}_1^{(m,i)}(k)$ is a circulant matrix, (3) becomes

$$\begin{aligned}\mathbf{z}^{(m)}(k) &= \mathbf{D}^{(m,1)}(k)\mathbf{s}(k) + \mathbf{D}^{(m,2)}(k)\mathbf{s}(k+1) - \mathbf{F}\mathbf{H}_1^{(m,1)}(k)\mathbf{F}^H\mathbf{s}(k) \\ &\quad - \mathbf{F}\mathbf{H}_1^{(m,2)}(k)\mathbf{F}^H\mathbf{s}(k+1) + \mathbf{F}\mathbf{H}_1^{(m,1)}(k-1)\mathbf{F}^H\left(-\mathbf{s}^*(k-1)\right) \\ &\quad + \mathbf{F}\mathbf{H}_1^{(m,2)}(k-1)\mathbf{F}^H\mathbf{s}^*(k-2) + \mathbf{F}\mathbf{n}^{(m)}(k) \end{aligned} \quad (4)$$

In which $\mathbf{D}^{(m,i)}(k) = \mathbf{F}(\mathbf{H}_0^{(m,i)}(k) + \mathbf{H}_1^{(m,i)}(k))\mathbf{F}^H$ is a diagonal matrix with the nth entry being denoted as the channel frequency response of the nth subcarrier between the ith transmit-antenna and mth receive-antenna.

3. GSC-BASED EQUALIZER OF CP-FREE ST-BC MIMO-OFDM SYSTEM

To derive the equalizer based on the GSC structure, we denote the $MN\times 1$ multi-channel data model as $\mathbf{z}(k) = [\mathbf{z}^{(1)T}(k)\ \mathbf{z}^{(2)T}(k) \ldots \mathbf{z}^{(M)T}(k)]^T$, which is obtained by stacking the frequency-domain received data from all receive antennas, that is

$$\mathbf{z}(k) = \begin{bmatrix}\mathbf{D}^{(1)}(k) & \mathbf{D}^{(2)}(k)\end{bmatrix}\begin{bmatrix}\mathbf{s}(k) \\ \mathbf{s}(k+1)\end{bmatrix} + \mathbf{F}_M\begin{bmatrix}\mathbf{H}_1^{(1)}(k-1) & \mathbf{H}_1^{(2)}(k-1) & -\mathbf{H}_1^{(1)}(k) & -\mathbf{H}_1^{(2)}(k)\end{bmatrix}\mathbf{F}_4^H\begin{bmatrix}-\mathbf{s}^*(k-1) \\ \mathbf{s}^*(k-2) \\ \mathbf{s}(k) \\ \mathbf{s}(k+1)\end{bmatrix} + \mathbf{F}_M\mathbf{n}(k) \quad (5)$$

The corresponding parameters are denoted as $\mathbf{D}^{(i)}(k) = [\mathbf{D}^{(1,i)}(k)\ \mathbf{D}^{(2,i)}(k) \ldots \mathbf{D}^{(M,i)}(k)]^T$, $\mathbf{H}_1^{(i)}(k) = [\mathbf{H}_1^{(1,i)T}(k)\ \mathbf{H}_1^{(2,i)T}(k) \ \ldots\ \mathbf{H}_1^{(M,i)T}(k)]^T$, and $\mathbf{n}(k) = [\mathbf{n}^{(1)T}(k)\ \mathbf{n}^{(2)T}(k) \ldots \mathbf{n}^{(M)T}(k)]^T$. where $\mathbf{F}_j = \mathbf{F} \otimes \mathbf{I}_j$ and $\otimes$ denotes Kronecker product. Similarly, frequency domain received signal the time $k+1$ can be express as

$$\mathbf{z}(k+1) = \begin{bmatrix}\mathbf{D}^{(1)}(k+1) & \mathbf{D}^{(2)}(k+1)\end{bmatrix}\begin{bmatrix}-\mathbf{s}^*(k+1) \\ \mathbf{s}^*(k)\end{bmatrix} + \mathbf{F}_M\begin{bmatrix}\mathbf{H}_1^{(1)}(k) & \mathbf{H}_1^{(2)}(k) & -\mathbf{H}_1^{(1)}(k+1) & -\mathbf{H}_1^{(2)}(k+1)\end{bmatrix}\mathbf{F}_4^H\begin{bmatrix}\mathbf{s}(k) \\ \mathbf{s}(k+1) \\ -\mathbf{s}^*(k+1) \\ \mathbf{s}^*(k)\end{bmatrix} + \mathbf{F}_M\mathbf{n}(k+1) \quad (6)$$

Finally, we may take the complex conjugate to (6) and stacking with (5) to obtain a vector-matrix form, i.e.,

$$\begin{aligned}\overline{\mathbf{z}}(k+1) &= \begin{bmatrix}\mathbf{z}(k) \\ \mathbf{z}^*(k+1)\end{bmatrix} \quad (7) \\ &= \tilde{\mathbf{D}}(k+1)\tilde{\mathbf{s}}(k+1) + \underbrace{\tilde{\mathbf{H}}_{ISI}(k+1)\mathbf{e}(k+1) + \mathbf{v}(k+1)}_{\mathbf{b}(k+1)}\end{aligned}$$

where

$$\tilde{\mathbf{D}}(k+1) = \begin{bmatrix}\mathbf{D}^{(1)}(k) & \mathbf{D}^{(2)}(k) \\ \mathbf{D}^{(2)*}(k+1) & -\mathbf{D}^{(1)*}(k+1)\end{bmatrix} \quad (8)$$

$$\tilde{\mathbf{H}}_{ISI}(k+1) = \begin{bmatrix}\mathbf{H}_{ISI}(k) \\ \mathbf{H}_{ISI}(k+1)\end{bmatrix} = \begin{bmatrix}\mathbf{F}_M\mathbf{H}_1(k)\mathbf{F}_6^H \\ \mathbf{F}_M^*\mathbf{H}_1(k+1)\mathbf{F}_6^T\end{bmatrix} \quad (9)$$

and

$$\mathbf{v}(k+1) = \begin{bmatrix}\mathbf{F}_M\mathbf{n}(k) \\ \mathbf{F}_M^*\mathbf{n}^*(k+1)\end{bmatrix} \quad (10)$$

Their corresponding parameters are denoted as $\mathbf{H}_1(k) = [\mathbf{H}_1^{(1)}(k-1)\ \mathbf{H}_1^{(2)}(k-1)\ -\mathbf{H}_1^{(1)}(k)\ -\mathbf{H}_1^{(2)}(k)\ \mathbf{0}\ \mathbf{0}]$, $\mathbf{H}_1(k+1) = [\mathbf{0}\ \ \mathbf{0}\ \ -\mathbf{H}_1^{(2)*}(k+1)\ \ \mathbf{H}_1^{(1)*}(k+1)\ \ -\mathbf{H}_1^{(2)}(k)\ \ \mathbf{H}_1^{(1)*}(k)]$, $\tilde{\mathbf{s}}(k+1) = [\mathbf{s}^T(k)\ \ \mathbf{s}^T(k+1)]^T$, and $\mathbf{e}(k+1) = [-\mathbf{s}^H(k-1)\ \mathbf{s}^H(k-2)\ \mathbf{s}^T(k)\ \mathbf{s}^T(k+1)\ -\mathbf{s}^H(k+1)\ \mathbf{s}^H(k)]^T$. From (5), we learn that the nonzero entries of the upper triangular matrices $\mathbf{H}_1^{(m,i)}(k)$ for $1 \le i \le 2$, $1 \le m \le M$, $1 \le u \le U$, are all in the last L columns and hence $\mathbf{H}_1^{(i)}(k)$ is with rank L. Since the columns between $\mathbf{H}_1^{(1)}(k)$ and $\mathbf{H}_1^{(2)}(k)$ are in general linearly independent and thus $\mathbf{H}_1(k)$ has rank of $4L$. Since $\mathbf{F}_M$ and $\mathbf{F}_4^H$ are unitary matrices, the ISI signature matrix $\mathbf{H}_{ISI}(k) = \mathbf{F}_M\mathbf{H}_1(k)\mathbf{F}_6^H$ is still with rank $4L$. Therefore, $\tilde{\mathbf{H}}_{ISI}(k+1)$ is of rank $8L$. Accordingly, if the dimension of the null space of the signal signature matrix $\tilde{\mathbf{D}}(k+1)$, which equals $2N(M-1)$ is larger than $8L$, we can exploit the additional degrees-of-freedom to completely suppress the ISI term in (5) and extract the composite desired symbol $\tilde{\mathbf{s}}(k+1)$. Since the number of subcarriers is typically much larger than the channel order, i.e., $N>>L$ the choice of $M\geq 2$ will satisfy the condition.

Assume that the channel is perfectly known at the receiver, we can design an equalizer with weight matrix $\mathbf{W}(k+1)$ to recover the desired symbol blocks, where the output signal vector of the equalizer is given by

$$\begin{aligned}\mathbf{y}(k+1) &= \mathbf{W}^H(k+1)\bar{\mathbf{z}}(k+1) \\ &= \mathbf{W}^H(k+1)\tilde{\mathbf{D}}(k+1)\tilde{\mathbf{s}}(k+1) + \mathbf{W}^H(k+1)\mathbf{b}(k+1) \\ &\approx \tilde{\mathbf{D}}^H(k+1)\tilde{\mathbf{D}}(k+1)\tilde{\mathbf{s}}(k+1) \qquad (11)\end{aligned}$$

where $\tilde{\mathbf{D}}^H(k+1)\tilde{\mathbf{D}}(k+1)$ is diagonal matrix. To derive the optimum weight matrix, we may minimize $\mathbf{W}(k+1)$ $\mathbf{b}(k+1)$:

$$\text{Min } \mathbf{W}(k+1)\, \mathbf{b}(k+1) \qquad (12)$$

for interference suppression, and subject to the following constraint, i.e.,

$$\mathbf{W}^H(k+1)\tilde{\mathbf{D}}(k+1) = \tilde{\mathbf{D}}^H(k+1)\tilde{\mathbf{D}}(k+1) \qquad (13)$$

An efficient approach for obtaining the optimum $\mathbf{W}(k+1)$ is to transform the constrained optimization problem into an unconstrained one via the GSC structure. A schematic description of the GSC-based equalizer is shown in Figure 2, where the weight matrix is decomposed into

$$\mathbf{W}(k+1) = \tilde{\mathbf{D}}(k+1) - \mathbf{B}(k+1)\mathbf{U}(k+1) \qquad (14)$$

In (14) $\mathbf{B}(k+1)$ is the blocking matrix, whose columns are chosen from the orthogonal basis for the left null space of the signal signature $\tilde{\mathbf{D}}(k+1)$ and weight matrix $\mathbf{U}(k+1)$ is determined by minimizing the cost function, i.e.,

$$J(k+1) = E\left[\left\|\mathbf{W}^H(k+1)\mathbf{b}(k+1)\right\|^2\right] \qquad (15)$$

Following the conventional minimum mean output energy (MOE) (or power) [4] approach, we get

$$\begin{aligned}\mathbf{U}(k+1) &= \left(\mathbf{B}^{\mathrm{H}}(k+1)\mathbf{R}_{\mathbf{b}}(k+1)\mathbf{B}(k+1)\right)^{-1} \\ &\quad \times \mathbf{B}^H(k+1)\mathbf{R}_{\mathbf{b}}(k+1)\tilde{\mathbf{D}}(k+1)\end{aligned} \qquad (16)$$

where the correlation matrix of $\mathbf{b}(k+1)$ is defined as

$$\mathbf{R}_{\mathbf{b}}(k+1) = \tilde{\mathbf{H}}_{ISI}(k+1)\tilde{\mathbf{H}}_{ISI}^H(k+1) + \sigma_n^2\mathbf{I}_{2MN} \qquad (17)$$

The solve (16) it involves the inversion of the $2N(M-1)\times 2N(M-1)$ matrix $\mathbf{B}^H(k+1)\mathbf{R}_{\mathbf{b}}(k+1)\mathbf{B}(k+1)$, this will lead to a high computational load and poor convergence for real-time implementation whenever N is large. To circumvent this problem, the partially adaptive (PA) array scheme [13] can be applied for complexity reduction. To do so, we insert a dimension reducing matrix $\mathbf{T}(k+1)$ following the blocking matrix $\mathbf{B}(k+1)$, the block diagram of the PA-GSC equalizer is illustrated in Figure 3.

4. COMPUTER SIMULATION RESULTS

To verify the advantage of the proposed CP-free ST-BC MIMO-OFDM scheme, with the GSC-based equalizer, computer simulations are carried out to evaluate the system performance in terms of ICI and IBI suppression capability. The CP-based ST-BC MIMO-OFDM system (without considering effects of HPA and IQ-imbalance) is chosen as the benchmark, for comparison. Also, the number of subcarriers and the length of CP are set to be 64 and 16, respectively. Two transmit antennas and two receive antennas are equipped in the considered system. The channel is modeled as the FIR filter with maximum length of L (e.g., L=8), and each tap is generated, independently, using the Jakes model, where the mobile speed is 100 km/hr and carrier frequency is 5.18GH. Perfect channel information is assumed to be available at the access point (AP), and remain constant during an OFDM block. Both adaptive predistorter and compensator are based on the RLS algorithm to adaptively adjust the coefficients. For fair comparison, the transmit power of the CP-based OFDM symbol is normalized to N.

In the first case, we would like to investigate the nonlinear distortion effect due to HPA in the transmitter, and there has no IQ-imbalance problem in the receiver. For L=8, the results given in Fig. 4 showed that the BER performance of the CP-free ST-BC MIMO-OFDM system, with the GSC-based equalizer and adaptive Volterra-based predistorter [10], is very close to that with the CP-based ST-BC MIMO-OFDM system (without considering the effect of HPA). As evident from Fig. 4, the proposed CP-free ST-BC MIMO-OFDM system outperformed the one without using the predistorter. Also, we found that with the PA approach the performance is identical to that with fully adaptation scheme. Next, we consider the effect due to IQ-imbalance, and similarly from Fig. 5 we learn that the BER performance using the proposed scheme is very close to that with the CP-based ST-BC MIMO-OFDM system (without considering the effects of HPA and IQ-imbalance), and outperforms the one without using the compensator.

5. CONCLUSION

In this paper, we investigated the performance of the CP-free ST-BC MIM)-OFDM systems, under the effects of HPA and IQ-imbalance. From simulation results, we showed that it could be applied to achieve desired performance under the HPA and IQ-imbalance and DC-offset effects, with the cost of more complex received design associated with the GSC-based equalizer implemented in the frequency domain.

6. ACKNOWLEDGEMENT

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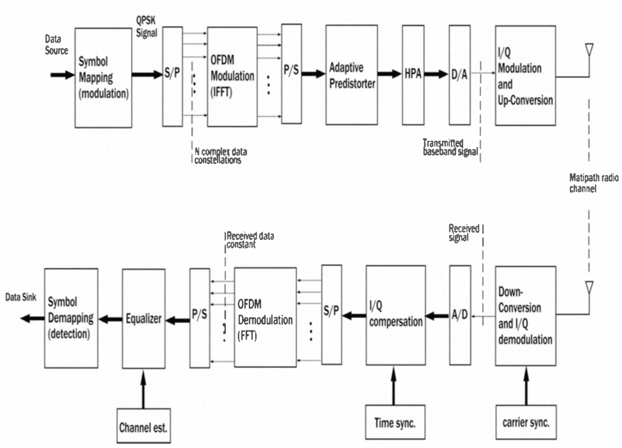

Figure 1 The CP-free ST-BC MIMO-OFDM systems under HPA and IQ-imbalance.

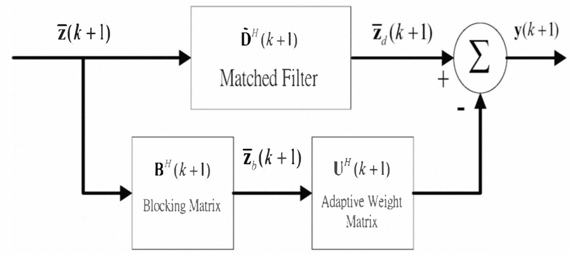

Figure 2 Block diagram of the GSC-based equalizer.

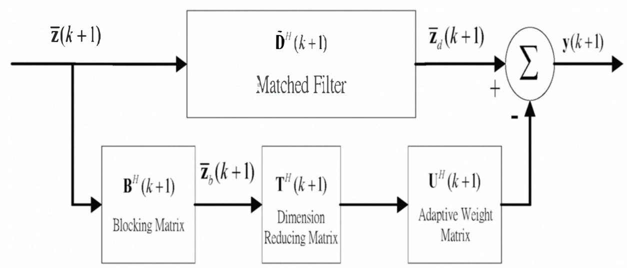

Figure 3 Block diagram of GSC-based equalize with PA.

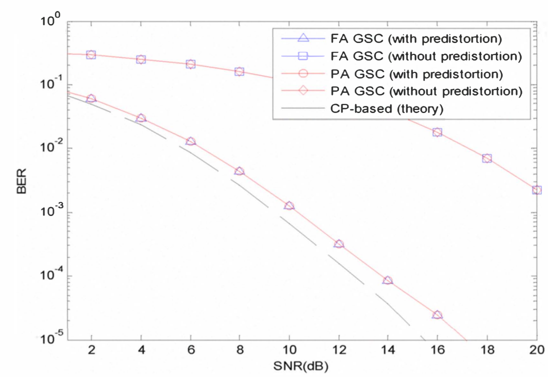

Figure 4 Performance comparisons of the CP-free ST-BC MIMO-OFDM system with and without adaptive predistorter.

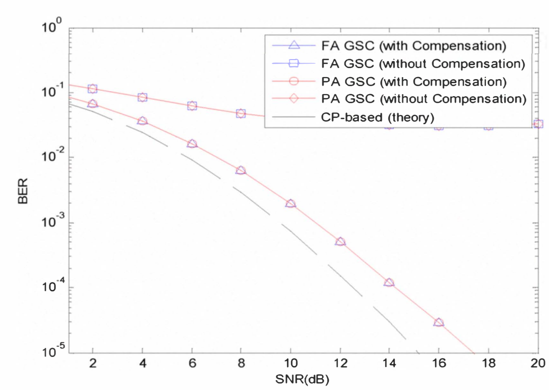

Figure 5 Performance comparisons of the CP-free ST-BC MIMO-OFDM system with and without adaptive compensator.