Blind Detection for 5G MIMO Short-Packet Communication System

Bo-Heng Yeh, Fang-Biau Ueng and Yun-Yu Li

***Abstract*—The upcoming fifth-generation (5G) systems will need to support novel traffic types that use short packets. The short-packet communication (SPC) is a concerned application in the subject about low-latency high-reliability. Due to the property of SPC, efficiently reducing the length of cyclic prefix (CP) and length of pilot is an important study. As far as we know, no research has been presented for blind receiver with joint channel shortening, carrier frequency offset (CFO) compensation and data detection for 5G SPC systems. Therefore, we employ the blind system in the receiver which does not need pilot and reduce the length of CP as short as possible. CFO estimation and channel shortening have been addressed individually. However, this leads to a circular set of assumptions, since standard CFO estimators perform poorly when the channel is longer than the guard interval, and standard channel shorteners assume that CFO estimation has already been perfectly performed. This paper proposes a blind receiver for uplink reception of a multi-input multi-output (MIMO) single-carrier interleaved frequency-division multiple-access (SC-IFDMA) system transmitting over a highly-dispersive channel, which is affected by both timing offsets (TOs) and frequency offsets (CFOs). When the length of CP is insufficient to compensate for channel dispersion and TOs, a common strategy is to shorten the channel by means of time-domain equalization, in order to restore CP properties and ease signal reception. This paper proposes a blind receiver for SPC systems that do not need pilot and can solve the interblock interference (IBI) caused by insufficient CP length. The receiver exhibits a four-stage structure: the first stage performs blind shortening of channel impulse responses (CIRs), without needing neither *a* priori knowledge of the CIRs to be shortened, nor preliminary compensation of the CFOs; the second stage performs joint estimation and compensation of the CFOs; the third stage implements signal-to-noise ratio (SNR) maximization, without requiring knowledge of the shortened CIRs; the fourth stage performs blind detection with fractional lower-order statistics constant modulus algorithm (FLOS-CMA). Some simulation examples are given to show the effectiveness and comparisons of the proposed receiver.**

***Index Terms*—short-packet communications, blind receiver.**

I. Introduction

THE vision of the Internet of Things promises to bring wireless connectivity. Each new generation of cellular systems has been mainly designed with the objective to provide a substantial gain in data rate over the previous generation. Fifth generation (5G) will depart from this scheme: its focus will not only be on enhanced broadband services. This is because the vast majority of wireless connections in 5G will most likely be originated by autonomous machines and devices rather than by the human-operated mobile terminals. 5G will address the specific needs of autonomous machines and devices by providing two novel wireless modes: ultrareliable communication (URC) and massive machine-to-machine communications (MM2M). With MM2M one refers to the scenario where a massive number of devices need to be supported within a given area. The data packets are short and reliability must be high to cope with critical events. Indeed, short packets are the typical form of traffic generated by sensors and exchanged in machine-type communications. Among these, short-packet communications (SPC) [1]-[2] is receiving a significant attention in the subject about low-latency high-reliability. Therefore, we employ the blind system in the receiver which does not need pilot and reduce the length of CP as short as possible. So, 5G will potentially require a transmission mode with very low air interface latency enabled by very short frames.

In the uplink, we employ MIMO SC-IFDMA [3] which can be considered as a precoded version of OFDMA [4]-[5]. One of the major drawbacks of OFDMA is due to the large fluctuations of the transmitted signal envelope [6], which dictates the use of expensive linear amplifiers in the uplink, as well as significant power back-off at the user transmitters, thus resulting in both undesirable additional costs and waste of battery power. Just like OFDMA, the performance of the SC-IFDMA uplink is susceptible to time offset (TO) and carrier frequency offsets (CFO) between the transmitters and the receiver. TO and CFO will cause both loss of orthogonality among subcarriers and IBI. To solve this question, synchronization/compensation originally designed for the OFDMA uplink [7] can be amended and applied to the SC-IFDMA uplink. MIMO [8]-[10] wireless communication which uses spatial multiplexing is able to raise channel capacity through added data, which need to be separated by a detection algorithm.

If the delay spread of the wireless channel is larger than the guard interval between blocks, then channel shortening is required for proper demodulation. Moreover, the vast majority of CFO estimation algorithms require that channel shortening has already been performed and in some cases CFO estimation techniques assume that the channel is an ideal impulse. However, fulfillment of the QS assumption might be impractical for systems operating over highly-dispersive multipath channels: in this case, joint timing and frequency synchronization at the BS is needed, which becomes a formidable task in the case of interleaved carrier assignment scheme (CAS), since the users cannot be separated in advance by simple bandpass filtering, as in subband CAS. In this paper, we consider the uplink MIMO SC-IFDMA systems affected by both TOs and CFOs,

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operating over a highly dispersive channel for which the QS assumption is violated, i.e., the CP is not long enough to compensate for channel dispersion plus the residual user TOs. The receiver of our study exhibits a four-stage structure: the first stage performs blind shortening of CIRs, without needing neither *a* priori knowledge of the CIRs to be shortened, nor preliminary compensation of the CFOs. In the general condition, the length of CP must be longer than channel path to effectively suppress IBI by CP removal. To overcome this problem, channel shortening techniques can make channel become an equivalent channel whose length is shortened. Several non-blind [11]–[15] as well as blind channel-shortening algorithms have been proposed for single-input single-output systems, some of which have also been extended to the MIMO case. The second stage performs joint estimation and compensation of the CFOs. Because SC-IFDMA is sensitive to CFO, it needs an accurate estimation of CFO to keep the orthogonality. In this paper, the CFO estimator is based on the algorithms of estimation of signal parameters via rotational invariance technique (ESPRIT). The number of parameters estimated using ESPRIT becomes equal to the number of unknown values, under the assumption that all the transmitted signals are independent. It can therefore be expected to result in an improved CFO estimation performance. ESPRIT has a much lower computational load. The third stage implements signal-to-noise ratio (SNR) maximization, without requiring knowledge of the shortened CIRs. The last stage performs blind data detection with FLOS-CMA. Due to using MIMO wireless communication, it needs to separate data received in different antenna by a detection algorithm. At the output of the antennas, the signals are interfered by both inter-symbol interference (ISI) and inter-user interference (IUI). To suppress the ISI, and also to separate different signals, a multi-user constant modulus algorithm (MU-CMA) was proposed. However, in many practical problems, the noise encountered is more impulsive in nature. In order to handle the realistic noise and interference in the data, we employ the FLOS-CMA.

II. System Model

Let us set that the uplink of a MIMO SC-IFDMA system with $k \le k_m$ active users, each has N_t-antenna transceiver, transmitting to BS equipped with N_r antennas. The transmitter is described as Figure 1. The system have total of *M* subcarriers, distributed in K_m disjoint sets, each consisting of M_u subcarriers, $M_u = M / K_m$. In MIMO SC-IFDMA system, the subcarriers are uniformly distributed to the signal band, i.e., let $i_{k,0} < i_{k,1} < \cdots < i_{k,M_u-1}$ denote the subcarriers distributed to user $k \in \{1, 2, \ldots, K\}$, one has

$$i_{k,l} = lK_m + \phi_k, \quad \textbf{for } l \in \{\mathbf{0}, \mathbf{1}, \ldots, M_u - 1\} \tag{1}$$

where $\phi_k \in \{0, 1, \ldots, K_m - 1\}$ denotes the index of the first subcarrier assigned to the k_{th} user.

Let $s_k(n) = [\, s_k^0(n), s_k^1(n), \cdots, s_k^{N_t-1}(n)\,]^T$, and $s_k^N(n) = [\, s_{k,0}^N(n), s_{k,1}^N(n), \cdots, s_{k,M_u-1}^N(n)\,]^T$ denote the n_{th} data block of the k_{th} user with N_{th} antenna. $\tilde{s}_k^N(n) = W_{dft} s_k^N(n) \in \mathrm{C}^{M_u}$, where W_{dft} denotes the M_u-point normalized DFT matrix, whose elements are $\{W_{dft}\}_{l_1, l_2} = M_u^{-1/2} e^{-j\frac{2\pi}{M_u} l_1 l_2}$, for $l_1, l_2 \in \{0, 1, \ldots, M_u - 1\}$. The block $\tilde{s}_k^N(n)$ is subject to SC-IFDMA subcarrier mapping through *M*-point IDFT. The length of CP is L_{CP}. Let $P = M + L_{CP}$, the block can be expressed as follows,

$$\begin{aligned} u_k^N(n) &= \left[\, u_{k,0}^N(n), u_{k,1}^N(n), \ldots, u_{k,P-1}^N(n)\,\right]^T \\ &= \mathbf{T_{cp}} \mathbf{W_k} \tilde{s}_k^N(n) \end{aligned} \tag{2}$$

where $\mathbf{T_{cp}}$ is CP insertion, $\mathbf{T_{cp}} = [\mathbf{I_{cp}^T}, \mathbf{I_M}]^T$, $\mathbf{W_k}$ is the *M*-point normalized IDFT matrix and elements are $\{\mathbf{W_k}\}_{l_1, l_2} = M^{-1/2} e^{j\frac{2\pi}{M} l_1 l_2}$, for $l_1 \in \{0, 1, \ldots, M-1\}$ and $l_2 \in \{0, 1, \ldots, M_u - 1\}$. The baseband received signal at the α_{th} antenna ($\alpha \in \{1, 2, \ldots, N_r\}$) can be expressed as follows,

$$r_\alpha(t) = \sum_{k=0}^{K-1} \sum_{N=0}^{N_t-1} e^{j2\pi\Delta f_k t} [u_k^N(t - \Delta\tau_k) * h_{k,\alpha}^N(t)] + \omega_\alpha(t) \tag{3}$$

where $h_{k,\alpha}^N(t)$ is the channel impulse responses between the k_{th} user transmitter at N_{th} transmit antenna and the α_{th} BS antenna receiver, Δf_k and $\Delta\tau_k$ are the remaining CFO and TO of the k_{th} user, and $\omega_\alpha(t)$ is the noise.

Assuming that $h_{k,\alpha}^N(t)$ spans L_k sampling periods, i.e., $h_{k,\alpha}^N(t) \equiv 0$ with $t \notin [0, L_k T_c]$ ($T_c = T / P$, where *T* is the symbol length). By assuming that user will try to adjust uplink synchronization parameters, the remaining TO will be either much smaller than L_k or, in the worst condition, reduces to the two-way propagation delay [7] between the k_{th} user and the BS, whereas the remaining CFO is limited to one-half of the subcarrier spacing $\Delta f_0 = 1/(MT_c)$, i.e., $|\Delta f_k| < \Delta f_0 / 2$. TO can be expressed as $\Delta\tau_k = \theta_k T_c + \xi_k$, for θ_k >0 and $\xi_k \in [0, T_c)$. Therefore, we denote the normalized CFO $\varepsilon_k^N = \Delta f_k^N / \Delta f_0^N$, and $|\varepsilon_k^N|$ will be smaller than 1/2. The signal $r_\alpha(t)$ is sampled with rate N_c / T_c, and $N_c \ge 1$ denoting the oversampling factor. The q_{th} sample of the received signal at the α_{th} antenna is given by

$$\begin{aligned} r_\alpha^{(q)}(m) &= r_\alpha(mT_c + q\frac{T_c}{N_c}) \\ &= \sum_{k=0}^{K-1} \sum_{N=0}^{N_t-1} e^{j\frac{2\pi}{M}\varepsilon_k^N m} \sum_{l=0}^{L_k+\theta_k} h_{k,\alpha}^{N\ (q)}(l - \theta_k) u_k^N(m - l) + \omega_\alpha^{(q)}(m) \end{aligned} \tag{4}$$

where

$$h_{k,\alpha}^{N\ (q)}(l) = e^{j\frac{2\pi}{MN_c}\varepsilon_k^N q} h_{k,\alpha}^N(lT_c + qT_c / N_c - \xi_k) \tag{5}$$

$$\omega_\alpha^{(q)}(m) = \omega_\alpha(mT_c + qT_c / N_c) \tag{6}$$

Set $Q = N_r N_c$ and by collecting the N_c data vectors resulting from oversampling over all the N_r receiving antennas into the *Q*-dimensional complex vector, we obtain

$$r(m) = [\{r^{(0)}(m)\}^T, \{r^{(1)}(m)\}^T, \ldots, \{r^{(N_c-1)}(m)\}^T]^T \tag{7}$$

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with $r^{(q)}(m)=[r_1^{(q)}(m),r_2^{(q)}(m),\ldots,r_{N_r}^{(q)}(m)]^T$

Thus, we can express the received signal as

$$r(m)=\sum_{k=0}^{K-1}\sum_{N=0}^{N_t-1}e^{j\frac{2\pi}{M}\varepsilon_k^N m}\sum_{l=0}^{L_k+\theta_k}h_k^{N(q)}(l)u_k^N(m-l)+\omega_\alpha(m) \quad (8)$$

where

$$h_k^N(l)=[\{h_k^{N(0)}(l)\}^T,\{h_k^{N(1)}(l)\}^T,\ldots,\{h_k^{N(N_c-1)}(l)\}^T]^T \quad (9)$$

$$h_k^{N(q)}(l)=[h_{k,1}^{N\,(q)}(l-\theta_k),h_{k,2}^{N\,(q)}(l-\theta_k),\ldots,h_{k,N_r}^{N\,(q)}(l-\theta_k)]^T \quad (10)$$

$$\omega(m)=[\{\omega^{(0)}(m)\}^T,\{\omega^{(1)}(m)\}^T,\ldots,\{\omega^{(N_c-1)}(m)\}^T]^T \quad (11)$$

$$\omega^{(q)}(m)=[\omega_1^{(q)}(m),\omega_2^{(q)}(m),\ldots,\omega_{N_r}^{(q)}(m)]^T \quad (12)$$

III. The Proposed Blind Receiver

The proposed blind receiver is described as Figure 2. The blind receiver includes four steps: blind channel shortening, CFO estimation and compensation, SNR maximization and blind data detection.

Blind Channel Shortening

When $L_{\max}=\max_{k\in\{1,2,\ldots,K\}}\{L_k+\theta_k\}$ and $L_{\max}>L_{cp}$, the effects of channel dispersion and TOs cannot be cancelled by removing CP. To solve this issue, we propose to incorporate in the receiver a TEQ, aimed at shortening the CIR to suppress IBI by CP removal. The proposed L_e -order finite-impulse response (FIR) TEQ acts on L_e+1 consecutive sample of $r(m)$ and can be expressed as follows,

$$\bar{r}(m)=[r^T(m),r^T(m-1),\ldots,r^T(m-L_e)]^T$$

$$=\sum_{k=0}^{K-1}\sum_{N=0}^{N_t-1}e^{j\frac{2\pi}{M}\varepsilon_k^N m}\Sigma_k^N H_k^N\bar{u}_k^N(m)+\bar{w}(m) \quad (13)$$

where

$$\Sigma_k^N=\mathrm{I}_Q\otimes diag(1,e^{-j\frac{2\pi}{M}\varepsilon_k^N},\ldots,e^{-j\frac{2\pi}{M}\varepsilon_k^N L_e}) \quad (14)$$

$$\bar{u}_k^N(m)=[u_k^N(m),u_k^N(m-1),\ldots,u_k^N(m-L_g)]^T \quad (15)$$

$$\bar{w}(m)=[w^T(m),w^T(m-1),\ldots,w^T(m-L_e)]^T \quad (16)$$

and block H_k^N is the Toeplitz channel matrix with $L_g=L_e+L_{\max}$

$$H_k^N=\begin{bmatrix} h_k^N(0) & h_k^N(1) & \cdots & h_k^N(L_{\max}) & 0_Q & \cdots & 0_Q \\ 0_Q & h_k^N(0) & h_k^N(1) & \cdots & h_k^N(L_{\max}) & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & 0_Q \\ 0_Q & \cdots & \vdots & h_k^N(0) & h_k^N(1) & \cdots & h_k^N(L_{\max}) \end{bmatrix}$$

(17) Because the combining MMOE TEQ depends on second order statistics of $r(m)$, it is investigated the properties of its covariance matrix. $R_{\bar{r}\bar{r}}(m)=\mathrm{E}[\bar{r}(m\bar{r}^H(m)]$ is periodically time varying in m with period P. After TEQ, the output is y_m, $y_m=f^H\bar{r}(m)$ where $f\in C^{Q(L_e+1)}$ is the weight vector of the TEQ.

$$y(m)=\sum_{k=0}^{K-1}\sum_{N=0}^{N_t-1}e^{j\frac{2\pi}{M}\varepsilon_k^N m}g_k^{N\,H}\bar{u}_k^N(m)+v(m) \quad (18)$$

and $g_k^N=H_k^{N\,H}\Sigma_k^{N\,H}f\in C^{L_g+1}$ is the combined L_g -order FIR channel-TEQ response. In this paper, we design the shortened CIR of the users as FIR channel of order L_{eff} and L_{eff} is a design parameter and equal or less than L_{cp}. The proposed desired window is $\{0,1,\cdots,L_{eff}\}$. According to the above equation, we can find that

$$\bar{u}_k^N(m)=[\bar{u}_{k,win}^N(m)^T,\bar{u}_{k,wall}^N(m)^T]^T \quad (19)$$

$$g_k^N=[g_{k,win}^{N\quad T},g_{k,wall}^{N\quad T}]^T \quad (20)$$

where $\bar{u}_{k,win}^N(m)\in C^{L_{eff}+1}$, $\bar{u}_{k,wall}^N(m)\in C^{L_g-L_{eff}}$, $g_{k,win}^N\in C^{L_{eff}+1}$, $g_{k,wall}^N\in C^{L_g-L_{eff}}$ denote as the signal and channel components at the TEQ output that are inside/outside desired window, and the TEQ output can be expressed as $y(m)=y_{win}(m)+y_{wall}(m)$, where

$$y_{win}(m)=\sum_{k=0}^{K-1}\sum_{N=0}^{N_t-1}e^{j\frac{2\pi}{M}\varepsilon_k^N m}g_{k,win}^{N\quad H}\bar{u}_{k,win}^N(m) \quad (21)$$

$$y_{wall}(m)=\sum_{k=0}^{K-1}\sum_{N=0}^{N_t-1}e^{j\frac{2\pi}{M}\varepsilon_k^N m}g_{k,wall}^{N\quad H}\bar{u}_{k,wall}^N(m)+v(m) \quad (22)$$

$y_{wall}(m)$ contains noise and the IBI component outside the desired window. This part cannot suppress IBI by CP removal. Assuming that $g_{k,wall}^N=0_{L_g-L_{eff}}$, $g_k^N=H_k^{N\,H}\Sigma_k^{N\,H}f=[g_{k,win}^{N\quad T},0_{L_g-L_{eff}}^{\quad T}]^T$. It can be solved for f (zero-forcing approach), which needs to require knowledge of H_k^N and Σ_k^N. But it cannot be called blind system. So we propose MMOE approach to make g_k^N become a blind manner, by generalizing the design to SC-IFDMA scenario. We can get the vector f by minimizing the mean output-energy $MOE(f)=\mathrm{E}[|y(m)|^2]=f^H R_{\bar{r}\bar{r}}f$ at the TEQ output, with appropriate limit aimed at preserving the desired component. Let $H_k^N=[H_{k,win}^N,H_{k,wall}^N]$ with

$$H_{k,win}^N=[h_{k,0}^N,h_{k,1}^N,\ldots,h_{k,L_{eff}}^N]\in C^{Q(L_e+1)\times(L_{eff}+1)},$$

$$H_{k,wall}^N=[h_{k,L_{eff}+1}^N,h_{k,L_{eff}+2}^N,\ldots,h_{k,L_g}^N]\in C^{Q(L_e+1)\times(L_g-L_{eff})}.$$

$g_{k,win}^N$ and $g_{k,wall}^N$ can be expressed as follows,

$$g_{k,win}^{N\quad H}=f^H\Sigma_k^N H_{k,win}^N$$

$$g_{k,wall}^{N\quad H}=f^H\Sigma_k^N H_{k,wall}^N.$$

Set $L_{eff}\in\{0,1,\ldots,\min(L_e,L_{\max})\}$. The $(\delta+1)_{th}$ column $h_{k,\delta}^N$ of $H_{k,win}^N$ with $\delta\in\{0,1,\ldots,L_{eff}\}$ can be rewritten as $h_{k,\delta}^N=\Theta_\delta\xi_{k,\delta}^N$ where $\Theta_\delta=[\mathrm{I}_{Q(\delta+1)},0_{Q(L_e-\delta)\times Q(\delta+1)}^T]^T$ is a known full column rank matrix, which not depend on k and N and contents $\Theta_\delta{}^T\Theta_\delta=\mathrm{I}_{Q(\delta+1)}$, whereas $\xi_{k,\delta}^N=[h_k^{N\,T}(\delta),h_k^{N\,T}(\delta-1),\ldots,h_k^{N\,T}(0)]$. Therefore, the $(\delta+1)_{th}$ column of $\Sigma_k^N H_k^N$ can be rewritten as

$$\Sigma_k^N H_k^N=\Sigma_k^N\Theta_\delta\xi_{k,\delta}^N=\Theta_\delta\bar{\Sigma}_{k,\delta}^N\xi_{k,\delta}^N$$

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where

$$\overline{\Sigma}_{k,\delta}^{N} = \mathrm{I}_Q \otimes diag(1, e^{-j\frac{2\pi}{M}\varepsilon_k^N}, \ldots, e^{-j\frac{2\pi}{M}\varepsilon_k^N \delta}).$$

Due to the recursive relation $\Theta_{\delta-1} = \Theta_\delta J_\delta$, with $J_\delta = [\mathrm{I}_{Q\delta}, 0_{Q\times Q\delta}^T]^T \in C^{Q(\delta+1)\times Q\delta}$, ${g_{k,win}^N}^H$ can be expressed as follows,

$${g_{k,win}^N}^H = f^H \Theta_{L_{eff}} [J_{L_{eff}} \cdots L_1 \overline{\Sigma}_{k,0}^N \xi_{k,0}^N, J_{L_{eff}} \cdots L_2 \overline{\Sigma}_{k,1}^N \xi_{k,1}^N, \ldots, \overline{\Sigma}_{k,L_{eff}}^N \xi_{k,L_{eff}}^N] \quad (23)$$

where $f^H \Theta_{L_{eff}} = \gamma^H$, with $\gamma \in C^{Q(L_{eff}+1)}$ a nonzero vector. The blind constrained minimization of MOE(f) can be written as

$$f_{mmoe} = \arg\min_f \{ f^H R_{\overline{r}\overline{r}} f \} \quad \text{subject to } f^H \Theta_{L_{eff}} = \gamma^H$$

(24) CFO Estimation and Compensation

We can find that it can be obtained a great shortening performance in medium-to-high SNR values so we choose $L_{eff} = L_{CP}$. Assuming in the high SNR and that great shortening condition is satisfied by MMOE shortening. Therefore

$$g_k^N = {H_k^N}^H {\Sigma_k^N}^H f_{mmoe} = [{g_{k,win}^N}^T, 0_{L_g - L_{cp}}^T]$$

where $g_{k,win}^N = G_{k,win}^N \gamma$ with $G_{k,win}^N = {H_{k,win}^N}^H {\Sigma_k^N}^H F_{mmoe}$ so the vector $z(m) = F_{mmoe}^H \overline{r}(m)$ can be expressed as

$$z(m) = \sum_{k=0}^{K-1} \sum_{N=0}^{N_t-1} e^{j\frac{2\pi}{M}\varepsilon_k^N m} {G_{k,win}^N}^H \overline{u}_{k,win}^N(m) + v(m) \quad (25)$$

where $v(m) = F_{mmoe}^H \overline{w}(m)$. After carrying out the poly-phase decomposition of $y_{mmoe}(m) = z^T(m)\gamma^*$ and removing CP to achieve IBI suppression, and the TEQ output vector $y_{mmoe}(n) = [y_{mmoe}(nP + L_{cp}), y_{mmoe}(nP + L_{cp} + 1), \ldots, y_{mmoe}(nP + P - 1)]^T$ of IBI free sample in the n_{th} symbol period is given by $y_{mmoe}(n) = Z(n)\gamma^*$ where $Z(n) = [z(nP + L_{cp}), \ldots, z(nP + P - 1)]^T$. According to [3] that $Z(n)$ can be rewritten as

$$Z(n) = \Psi A(n) + V(n) \quad (26)$$

Put $Z(n)$ into a $K_m N_t \times M_u Q(L_{cp} + 1)$ matrix V row by row and compute the estimated $K_m N_t \times K_m N_t$ covariance matrix

$$\Xi = \frac{1}{M_u Q(L_{cp} + 1)} V V^H \quad (27)$$

Let E denote the set of KN_t eigenvectors which correspond to the KN_t largest eigenvalues of Ξ. E spans the signal subspace, whereas the $(K_m - K)N_t$ smallest eigenvalues of Ξ are equal to σ^2. E_x is defined as the first $(K_m N_t - 1)$ row of E, E_y is defined as the last $(K_m N_t - 1)$ row of E. It can be verified that the nonsingular matrix $\overline{V}$ exists as $E_x \prod = E_y$. $\prod$ is obtained by the least square approach as $\prod = {E_x}^+ E_y$ where the superscript $(\cdot)^+$ denotes pseudoinverse. In addition, the set of KN_t eigenvalues of $\prod$ can be expressed as $\widetilde{\prod}_k^N = e^{j2\pi\hat{\varepsilon}_k^N}$ for $k = 0,1,\ldots,K-1$, $N = 0,1,\ldots,N_t - 1$, hence $\hat{\varepsilon}_k^N = \frac{\ln(\widetilde{\prod}_k^N)}{2\pi}$. Through taking $\hat{\varepsilon}_k^N$ into Ψ to implement CFO compensation, the estimate of $A(n)$ is given by

$$\hat{A}(n) = \Psi' Z(n) = A(n) + \Psi' V(n) \quad (28)$$

where $\Psi' = (\Psi^H \Psi)^{-1} \Psi^H$ is the Moore-Penrose inverse of Ψ.

Blind User SNR Maximization

Assuming that IBI has been removed and the CFOs have been perfectly compensated. Thus, the resulting k_{th} user data matrix $\hat{A}_k(n)$ can be express as

$$\hat{A}_k(n) = R_k A(n) \quad (29)$$

where $R_k = [0_{(k-1)M_u \times M_u}, \mathrm{I}_{M_u}, 0_{(K-k)M_u \times M_u}]$ is the extraction matrix. By recalling the eigenstructure properties of circulant matrices, we write $\widetilde{\zeta}_{k,i}^N = W_{dft}^H \Lambda_{k,i}^N W_{dft}$, where $\Lambda_{k,i}^N = diag(\lambda_{k,i}^N)$ and $\lambda_{k,i}^N$ can be written as

$$\lambda_{k,i}^N = {M_u}^{-1/2} W_{dft} C B_k [\{G_{k,win}^N\}_{0,i}, \ldots, \{G_{k,win}^N\}_{L_{cp},i}]^H \quad (30)$$

After DFT, it has

$$Q_k(n) = W_{dft} \hat{A}_k(n) \quad (31)$$

To take into account the constraint vector γ, we observe that the cascade of CFO compensation, user separation and DFT can be equivalently seen as a linear transformation of the TEQ output vector $y_{mmoe}(n) = Z(n)\gamma^*$ which allows one to write the frequency-domain vector $\hat{q}_k(n)$ as

$$\hat{q}_k(n) = W_{dft} R_k \Psi' y_{mmoe}(n) = Q_k(n)\gamma^* \quad (32)$$

Therefore, the SNR can be expressed as

$$SNR_k = \frac{E\left[\left\| \sum_{N=0}^{N_t-1} {M_u}^{-1/2} \widetilde{S}_k^N(n) W_{dft} C B_k {G_{k,win}^N}^* \gamma^* \right\|^2\right]}{E\left[\left\| W_{dft} N_k(n) F_{mmoe}^* \gamma^* \right\|^2\right]} = \frac{\gamma^H R_{Q_k Q_k}^* \gamma}{\gamma^H F_{mmoe}^H R_{N_k N_k}^* F_{mmoe} \gamma} - 1 \quad (33)$$

where

$$R_{Q_k Q_k} = E[Q_k^H(n) Q_k(n)] = \sigma_s^2 M_u {G_{k,win}^N}^T {G_{k,win}^N}^* + F_{mmoe}^T R_{N_k N_k} F_{mmoe}^*,$$

with $R_{N_k N_k} = E[N_k^H(n) N_k(n)]$. According to [3] that $R_{N_k N_k}$ is a scaled identity matrix, therefore maximizing SNR with respect to γ leads to

$$\gamma_{k,opt} = \arg\max_\gamma \{\gamma^H R_{Q_k Q_k}^* \gamma\} \quad (34)$$

subject to $\gamma^H F_{mmoe}^H F_{mmoe} \gamma = 1$. $\widetilde{\gamma}_{k,\max}$ is the eigenvector associated with the largest eigenvalue of matrix ${R_{mmoe}^{-1}}^H R_{Q_k Q_k}^* R_{mmoe}^{-1}$, and $F_{mmoe} = Q_{mmoe} R_{mmoe}$ is the QR decomposition of F_{mmoe}.

IV. Simulation Results

In this section, the performance of the proposed blind receiver for MIMO SC-IFDMA system is simulated. The MIMO technology is simulated with 2×2 configuration and the adopted channel models are the Rayleigh fading channel with path number 10 and the cost207 fading channel which is simulated over bad urban area (path number 21). The modulation format is QPSK. In this system, each user has two antennas for transmitter side and receiver side has two antennas. The DFT sizes are $M_u = 64, M = 128$ and $M_u = 32, M = 64$. The CFO is 0.15 subcarrier spacing and the Cyclic Prefix (CP) lengths are 6 and 1. The design parameter L_{eff} is equal to the CP length. The step size of the equalizer is 0.001. **Table 1** summarizes the simulation conditions.

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Table 1: The details of simulation conditions.

Item	Value
Number of antennas	2x2
Modulation formats	QPSK
DFT size	$M_u = 64, M = 128$; $M_u = 32, M = 64$
CFO	0.15 subcarrier spacing
CP length	6 and 1
Channel	Rayleigh, cost207
Channel path	Rayleigh is 10 and cost207 is 21
Design parameter	L_{eff} = CP length
initial condition of Equalizer	[0,0,0…1…. ,0,0], the length is equal to M
Step size of Equalizer	0.001

At first, we set the DFT size as $M_u = 64, M = 128$, and the results are shown in **Figure 3－5**. **Figure 3 and Figure 4** describe the BER performance comparisons between MIMO SC-FDMA and MIMO SC-IFDMA systems with CP length 6 in Rayleigh channel and cost207 channel, respectively. It is shown that the BER of MIMO SC-IFDMA system is better than that of MIMO SC-FDMA. **Figure 5** describes the BER performance comparisons of different CP lengths for SC-IFDMA systems in Rayleigh channel. **Figure 6** describes the BER performance comparisons between those with and without blind channel shortening, CP length is 6 in Rayleigh channel. The DFT size is $M_u = 32, M = 64$. The simulation conditions are demonstrated to show the performances of the proposed blind receiver for very short packet communication. Obviously, it is shown that carrying out blind channel shortening can suppress the inter-block interference caused by insufficient CP length. When compared with the performance at BER= 10^{-2}, carrying out blind channel shortening is better than that without blind channel shortening at BER=10^{-2}. The difference between the two conditions is about 2dB. **Figure 7** describes the BER performance comparisons between those with and without CFO compensation in cost207 channels without channel shortening. The DFT size is $M_u = 32, M = 64$. It is shown that carrying out blind CFO compensation can have better performance and the performance approaches to that with perfect compensation of CFO. When compared with the performance at BER=10^{-2}, we find that carrying out blind CFO compensation is better about 3dB than that without CFO compensation at BER=10^{-2}. It can be found that the blind CFO estimation is almost perfect in high SNR.

V. Conclusion

This paper proposes a blind receiver for uplink reception of a MIMO SC-IFDMA system transmitting over a highly-dispersive channel, which is affected by both timing offsets (TOs) and frequency offsets (CFOs). When the length of the cyclic prefix (CP) is insufficient to compensate for channel dispersion and TOs, a common strategy is to shorten the channel by means of time-domain equalization, in order to restore the CP properties and ease signal reception. This paper proposes a blind receiver for SPC systems that do not need pilot and can solve the interblock interference (IBI) caused by insufficient CP length, the blind CFO estimation and compensation overcomes the drawback of MIMO SC-IFDMA system that is sensitive to CFO, and the blind data detection is proposed for separating signal at different antenna. Moreover, the simulation results are shown that the performances of the proposed blind receiver are acceptable for different environments.

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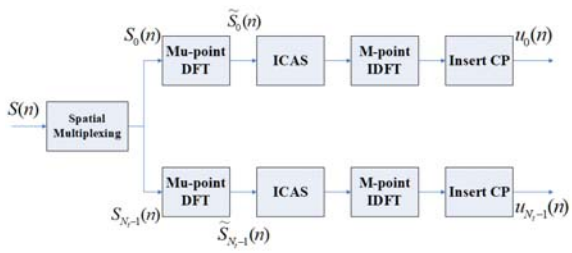

Figure 1: The transmitter of the MIMO SC-IFDMA system.

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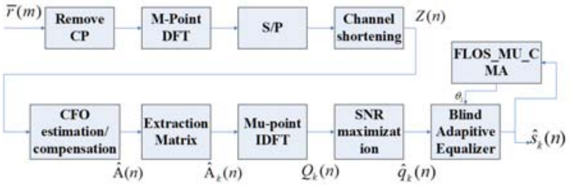

Figure 2: The proposed blind receiver.

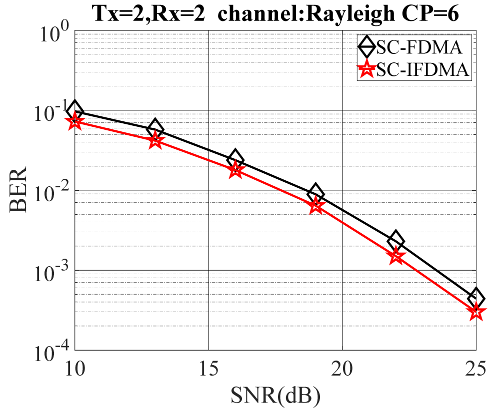

Figure 3: The BER performance comparisons between MIMO SC-FDMA and MIMO SC-IFDMA with CP length 6 in Rayleigh channel.

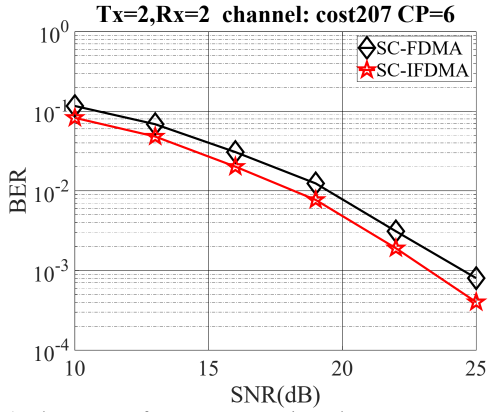

Figure 4: The BER performance comparisons between MIMO SC-FDMA and MIMO SC-IFDMA with CP length 6 in cost207 channel.

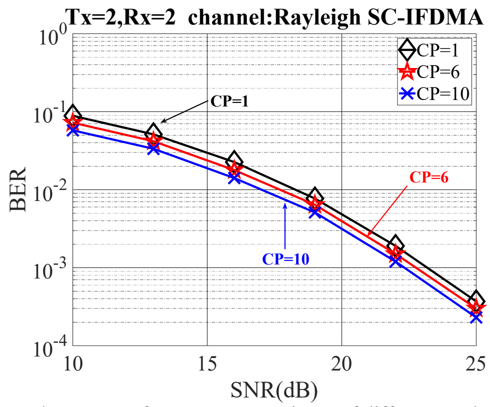

Figure 5: The BER performance comparisons of different CP length for MIMO SC-IFDMA systems in Rayleigh channel.

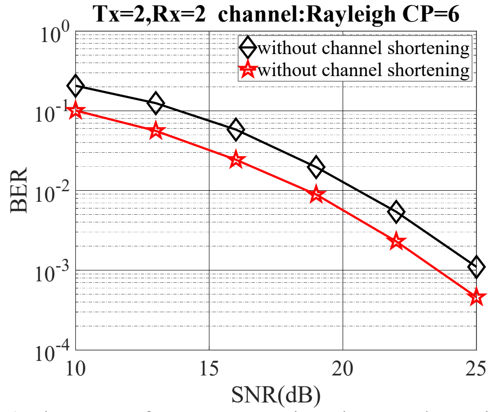

Figure 6: The BER performance comparisons between those with and without blind channel shortening, CP length is 6 in Rayleigh channel.

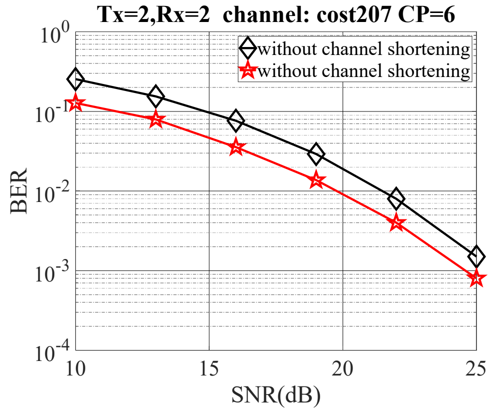

Figure 7: The BER performance comparisons between those with and without blind channel shortening, CP length is 6 in cost207 channel.