

Position-based ICI Elimination with Compressed Channel Estimation for SIMO-OFDM High Speed Train Systems

Xiang Ren*, Wen Chen*[†], Diandian Ren*, Bo Gong*, Qibo Qin*, and Lin Gui*

*Department of Electronic Engineering, Shanghai Jiao Tong University, China

[†]School of Electronic Engineering and Automation, Guilin University of Electronic Technology, China
Email: {renx; wenchen; haisai; gongbo; qinqibo; guilin}@sjtu.edu.cn

Abstract—In this paper, we consider a typical high speed train (HST) communication system with single-input multiple-output (SIMO) orthogonal frequency-division multiplexing (OFDM). We show that the inter-carrier interference (ICI) caused by large Doppler shifts can be mitigated by exploiting the train position information as well as the sparsity of the basis expansion model (BEM) based channel model. For the complex-exponential BEM (CE-BEM) based channel model, we show that the ICI can be completely eliminated to get the ICI-free pilots at each receive antenna. In addition, we design the pilot pattern to reduce the system coherence so as to improve the compressed sensing (CS) based channel estimation accuracy. In specific, the optimal pilot pattern is independent of the number of receive antennas, the Doppler shifts, the train position, or the train speed. Simulation results confirm the effectiveness of the proposed scheme in high-mobility environments. The results also show that the proposed scheme is robust to the train moving speed.

I. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) has demonstrated great promise in achieving high data rate in stationary and low-mobility environment. In high speed trains (HST) environments, however, since the train travels at a speed more than 350km/h, the high Doppler shift destroys the orthogonality resulting in the inter-carrier interference (ICI) in OFDM systems. This directly degrades the channel estimation accuracy and significantly affects the overall system performance. It is thus necessary and important to investigate reliable channel estimation and ICI mitigation methods in high-mobility environments.

Channel estimation over high mobility channels has been a long-standing issue [1]. Many researches find that wireless channels tend to exhibit sparsity, where the channel properties are dominated by a relatively small number of dominant channel coefficients. Thus, to utilize the channel sparsity, several works studied the applications of compressed sensing (CS) in the channel estimation [2], [3]. Coherence is a critical metric in CS and a lower system coherence leads to a better recover

This work is supported by the National 973 Project #2012CB316106, by NSF China #61328101 and #61471236, by the STCSM Science and Technology Innovation Program #13510711200, and by the SEU National Key Lab on Mobile Communications #2013D11.

performance [4]. Previous works [5]-[7] proposed several pilot design methods to reduce the system coherence and hence to improve the CS-based channel estimation performance. However, none of them considered the ICI mitigation.

To combat the ICI, several techniques have been considered in [5]-[7]. The works [5] and [6] considered the ICI as Gaussian noise, which needs low complexity but the performance degrades quickly with large Doppler shift. The work [7] proposed the ICI mitigation method based on iterative process, which is effective for the mitigation of ICI but incurs high computational complexity and results in error propagation.

In this paper, based on the basis expansion model (BEM), we first show that the ICI caused by the large Doppler shift can be mitigated by exploiting the train position information. Then, considering the complex-exponential BEM (CE-BEM), we propose a new low complexity position-based ICI elimination method, by which we can get the ICI-free pilots at each receive antenna. After that, for the considered CS channel estimator, we design the pilot pattern to minimize the system coherence. Specifically, the optimal pilot pattern is independent of the train speed, the train position, the Doppler shift, or the number of receive antennas. Simulation results confirm the effectiveness and robustness of the proposed scheme in HST environments.

II. SYSTEM MODEL

A. HST wireless communication system

We consider a typical broadband wireless communication system for HST [7], as shown in Fig. 1. The communication between the base stations (BS) and the mobile users is conducted in a two-hop manner through a relay station (RS) deployed on the train. The RS is connected with several antennas evenly located on the top of the train to communicate with the BS. Moreover, the RS is also connected with multiple indoor antennas distributed in the train carriages to communicate with mobile users by existing technologies. The BSs are located along the railway at some intervals and connected with optical fibers. Here we assume that each BS is equipped with one antenna and has the same transmit power and coverage range

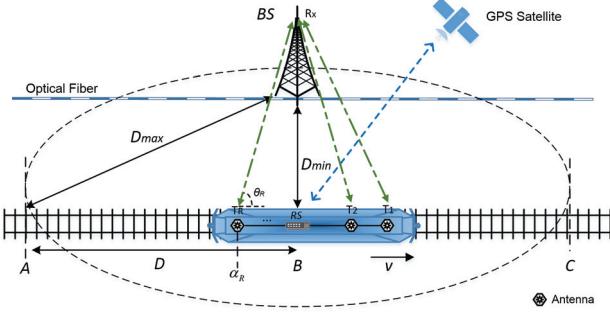


Fig. 1. A SIMO HST wireless communication system.

for simplicity. In addition, we assume that the HST is equipped with a global positioning system (GPS) which can estimate the HST's instant position and speed information perfectly and send them to the RS with no time delay [7].

In Fig. 1, we assume that the HST is traveling towards a fixed direction at a constant speed v . Let D_{max} denote the maximum distance from the BS to the railway (A and C to BS), D_{min} denote the minimum distance (B to BS), and D denote the distance between A and B . $\{T_r\}_{r=1}^R$ denotes the R receive antennas on the HST, and R_x denotes the antenna on the BS. In each cell, we define α_r as the distance between the r -th receive antenna and the position A , and define θ_r as the angle between the BS to T_r and the railway. For T_r at a certain position α_r , it suffers from a Doppler shift f_r , where f_r can be calculated by $f_r = \frac{v}{c} \cdot f_c \cos \theta_r$ with the carrier frequency f_c and the light speed c . We assume that f_r is constant within one OFDM symbol. In addition, considering the HST train channel exists a strong line-of-sight (LOS) propagation path [8], we here assume that all the paths of each receive antenna present identical Doppler shifts.

B. SIMO-OFDM system

In this paper, we only consider the first-hop communication in the HST system, i.e., from the BS to the RS. It is treated as a SIMO-OFDM system with one transmit antenna and R receive antennas. Suppose there are K subcarriers. The transmit signal at the k -th subcarrier during the n -th OFDM symbol is denoted as $X^n(k)$, for $n = 1, 2, \dots, N$ and $k = 1, 2, \dots, K$. At the r -th receive antenna, the received signals in the frequency domain are represented as

$$\mathbf{y}_r^n = \mathbf{H}_r^n \mathbf{x}^n + \mathbf{n}_r^n, \quad (1)$$

where $\mathbf{y}_r^n = [Y_r^n(1), Y_r^n(2), \dots, Y_r^n(K)]^T$ is the receive signal vector over all subcarriers of the n -th OFDM symbol, \mathbf{H}_r^n is the frequency domain channel matrix between the transmit antenna and the r -th receive antenna, $\mathbf{x}^n = [X^n(1), X^n(2), \dots, X^n(K)]^T$ is the transmitted signal vector, and $\mathbf{n}_r^n = [N_r^n(1), N_r^n(2), \dots, N_r^n(K)]^T$ denotes the noise vector, where $N_r^n(k)$ is the additive white Gaussian noise (AWGN) with a zero mean and σ_e^2 variance.

If the channel is time-invariant, \mathbf{H}_r^n is a diagonal matrix, which means the system is ICI-free. However, for high mobility channel, it varies rapidly in each OFDM symbol and thus

\mathbf{H}_r^n becomes a full matrix, resulting in ICI. Then, (1) can be rewritten as

$$\mathbf{y}_r^n = \mathbf{H}_{r,\text{free}}^n \mathbf{x}^n + \mathbf{H}_{r,\text{ICI}}^n \mathbf{x}^n + \mathbf{n}_r^n, \quad (2)$$

where $\mathbf{H}_{r,\text{free}}^n \triangleq \text{diag}\{[H_r^n(0,0), H_r^n(1,1), \dots, H_r^n(K-1, K-1)]\}$ denotes the ICI-free channel matrix, and $\mathbf{H}_{r,\text{ICI}}^n \triangleq \mathbf{H}_r^n - \mathbf{H}_{r,\text{free}}^n$ is the ICI part. As can be seen, in the presence of a high Doppler shift, the resulting ICI attenuates the desired signal and reduces channel estimation accuracy, resulting in degraded system performance. Thus, it is necessary to consider the ICI mitigation method in HST systems.

C. BEM-based Channel Model

Assume that the channel between the transmit antenna and each receive antenna consists of L paths. For each channel tap l , $0 \leq l \leq L-1$, we define $h_r^n(k, l)$ which collects the time variation of the channel tap at the k -th subcarrier within the n -th OFDM symbol of the channel of the r -th receive antenna. Denote f_{max} as the maximum Doppler shift, T as the symbol duration, and $Q = 2\lceil f_{max}T \rceil$ as the maximum number of the BEM order. Then, all the channel taps of the r -th receive antenna within the n -th OFDM symbol can be represented in one vector $\tilde{\mathbf{h}}_r^n = [h_r^n(0,0), \dots, h_r^n(0, L-1), \dots, h_r^n(K-1, 0), \dots, h_r^n(K-1, L-1)]^T$. Let $\mathbf{B} = [\mathbf{b}_0, \dots, \mathbf{b}_q, \dots, \mathbf{b}_Q]$ collects $Q+1$ basis functions as columns, \mathbf{b}_q denotes the q -th basis function ($q = 0, 1, \dots, Q$) whose expression is related to a specific BEM model. Then, we obtain

$$\tilde{\mathbf{h}}_r^n = (\mathbf{B} \otimes \mathbf{I}_L) \mathbf{c}_r^n + \boldsymbol{\epsilon}_r^n, \quad (3)$$

where we denote \mathbf{I}_L as the $L \times L$ identity matrix, $\mathbf{c}_r^n = [c_r^n(0,0), \dots, c_r^n(0, L-1), \dots, c_r^n(Q, 0), \dots, c_r^n(Q, L-1)]^T$ is the stacking coefficient vector, and $\boldsymbol{\epsilon}_r^n = [\epsilon_r^n(0,0), \dots, \epsilon_r^n(0, L-1), \dots, \epsilon_r^n(K-1, 0), \dots, \epsilon_r^n(K-1, L-1)]^T$ is the BEM modeling error. In the following, as our focus is to discuss the ICI eliminator and the channel estimator, we ignore $\boldsymbol{\epsilon}_r^n$ for convenience. We also assume that the coefficients are constant within one OFDM symbol.

Since we only consider the system in a single OFDM symbol in this paper, the symbol index n is omitted in the sequel for compactness. Then, substituting (3) into (1), we obtain

$$\mathbf{y}_r = \sum_{q=0}^Q \mathbf{D}_q \Delta_{r,q} \mathbf{x} + \mathbf{n}_r, \quad (4)$$

where $\mathbf{D}_q = \mathbf{F} \text{diag}\{\mathbf{b}_q\} \mathbf{F}^H$ denotes the q -th BEM basis function in the frequency domain, \mathbf{F} is the $K \times K$ DFT matrix, $\Delta_{r,q} = \text{diag}\{\mathbf{F}_L \mathbf{c}_{r,q}\}$ is a diagonal matrix whose diagonal entries are the frequency responses of $\mathbf{c}_{r,q}$, $\mathbf{c}_{r,q} = [c_r(q, 0), \dots, c_r(q, L-1)]^T$ denotes the BEM coefficients of all taps of the r -th receive antenna corresponding to the q -th basis function, and \mathbf{F}_L denotes the first L columns of $\sqrt{K}\mathbf{F}$.

III. POSITION-BASED ICI ELIMINATION

A. Exploiting the HST position information

We first give a definition of S -sparse channels based on the BEM channel model introduced in the previous section.

Definition 1 (S-sparse Channels [3]): For a BEM-based channel model given in (3), its dominant coefficients are defined as the coefficients which contribute significant powers, i.e., $|c_r(q, l)|^2 > \gamma$, where γ is a pre-fixed threshold. We say that the channel is S -sparse if the number of its dominant coefficients satisfies $S = \|\mathbf{c}_r\|_{\ell_0} \ll N_0 = L(Q + 1)$.

Then we give the following theorem, which reflects the position information of the HST system.

Theorem 1 (Position-based S-sparse channels): For a considered HST system, the high-mobility channel between the transmit antenna and each receive antenna at any given train position is S -sparse. For the r -th receive antenna at position α_r , its dominant channel coefficients locate in $\mathbf{c}_r^* = \mathbf{c}_{r,q|q=q_r^*} = [c_r(q_r^*, 0), c_r(q_r^*, 1), \dots, c_r(q_r^*, L - 1)]^T$, where q_r^* is its position-related dominant index.

Proof: The main idea is that, when the r -th receive antenna moves to α_r , as there exists a strong LOS, all its channel taps suffer from the same f_r . Thus, its dominant coefficients locate in \mathbf{c}_r^* with the q_r^* corresponding to α_r (also f_r). \mathbf{c}_r^* is sparse due to the multipath sparsity. Please refer to our technical report [9] for a complete proof. ■

Denote $F = Tf_{\max} = T_c^v \cdot f_c$. Then the relationship between q_r^* and α_r can be represented as

$$q_r^* = \begin{cases} \left\lceil F \cdot \frac{D - \alpha_r}{\sqrt{(D - \alpha_r)^2 + D_{min}^2}} \right\rceil + \frac{Q}{2}, & \alpha_r \in [0, D], \\ \left\lceil F \cdot \frac{D - \alpha_r}{\sqrt{(D - \alpha_r)^2 + D_{min}^2}} \right\rceil + \frac{Q}{2}, & \alpha_r \in (D, 2D], \end{cases} \quad (5)$$

where $\alpha_r \in [0, D]$ denotes the r -th receive antenna moving from A to B , and $\alpha_r \in (D, 2D]$ denotes moving from B to C .

From Theorem 1, we readily have the following corollary.

Corollary 1: For the considered HST system with any given train position, the high-mobility channel between the BS and the r -th receive antenna is S -sparse, and it can be modeled with its dominant coefficient vector \mathbf{c}_r^* and the dominant basis function \mathbf{D}_r^* , i.e., $\mathbf{H}_r = \mathbf{D}_r^* \Delta_r^*$, where $\Delta_r^* = \text{diag}\{\mathbf{F}_L \mathbf{c}_r^*\}$, and $\mathbf{D}_r^* = \mathbf{D}_{q|q=q_r^*}$. The relationship between the dominant index q_r^* and the antenna position α_r is given as (5).

According to Corollary 1, (4) can be simplified as

$$\mathbf{y}_r = \mathbf{D}_r^* \Delta_r^* \mathbf{x} + \mathbf{n}_r. \quad (6)$$

In this way, we exploit the position information of the BEM and utilize it to simply the required channel coefficients from $L(Q + 1)$ to L . Note that these analyses and conclusions are not restricted to any specific BEM.

B. Position-based ICI Elimination

In this subsection, we consider the CE-BEM [10] due to its independence of the channel statistics and it is strictly banded in the frequency domain. Specifically, for the CE-BEM, the q -th basis function \mathbf{b}_q can be represented as

$$\mathbf{b}_q = \left[1, \dots, e^{j \frac{2\pi}{K} k(q - \frac{Q}{2})}, \dots, e^{j \frac{2\pi}{K} (K-1)(q - \frac{Q}{2})} \right]^T. \quad (7)$$

Note that a large Q denotes a large Doppler shift caused by high mobility. Then, the \mathbf{D}_q can be written as

$$\mathbf{D}_q = \mathbf{F} \text{diag}\{\mathbf{b}_q\} \mathbf{F}^H = \mathbf{I}_K^{\langle q - \frac{Q}{2} \rangle} \mathbf{F} \mathbf{F}^H, \quad (8)$$

$$= \mathbf{I}_K^{\langle q - \frac{Q}{2} \rangle}, \quad (9)$$

where $\mathbf{I}_K^{\langle q - \frac{Q}{2} \rangle}$ denotes a matrix obtained from a $K \times K$ identity matrix \mathbf{I}_K with a permutation $q - Q/2$. Then, we have

$$\mathbf{H}_r = \sum_{q=0}^Q \mathbf{I}_K^{\langle q - \frac{Q}{2} \rangle} \Delta_{r,q}. \quad (10)$$

By detecting the matrix structure, we find that \mathbf{H}_r is strictly banded with the bandwidth $Q + 1$, which means that, suffering large Doppler shift, Q neighboring subcarriers give rise to interference, i.e., the desired signal suffers from the ICI caused by the Q neighboring subcarriers.

Let us consider Corollary 1, then the dominant basis \mathbf{D}_r^* for the CE-BEM can be rewritten as

$$\mathbf{D}_r^* = \mathbf{I}_K^{\langle q_r^* - \frac{Q}{2} \rangle}. \quad (11)$$

Similarly, we have

$$\mathbf{H}_r = \mathbf{I}_K^{\langle q_r^* - \frac{Q}{2} \rangle} \Delta_r^*. \quad (12)$$

where \mathbf{H}_r becomes a diagonal matrix with a permutation and its non-zero entries are corresponding to the dominant coefficients. From (12), we find that, with Corollary 1, the desired signal is free of ICI but with a permutation of the received subcarrier. This is reasonable because the dominant coefficients in \mathbf{c}_r^* , corresponding to f_r , describe the channel alone while the non-dominant ones can be ignored. Note that the conclusion that \mathbf{H}_r is a permuted diagonal matrix only holds for CE-BEM, since \mathbf{D}_q itself is a permuted identity matrix.

C. Channel Estimation with Position-based ICI Elimination

Assume that P ($P < K$) pilots are inserted in the frequency domain at the BS with the pilot pattern \mathbf{w} , where $\mathbf{w} = [w_1, w_2, \dots, w_P]$. Denote \mathbf{d} as the subcarrier pattern of the transmitted data. Assume \mathbf{w} is received at the r -th receive antenna with the pilot pattern $\mathbf{v}_r = [v_{r,1}, v_{r,2}, \dots, v_{r,P}]$. Then, with Corollary 1, the received pilots at the r -th receive antenna are represented as

$$\mathbf{y}_r(\mathbf{v}_r) = \mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \Delta_r^*(\mathbf{w}, \mathbf{w}) \mathbf{x}(\mathbf{w}) + \underbrace{\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{d}) \Delta_r^*(\mathbf{d}, \mathbf{d}) \mathbf{x}(\mathbf{d})}_{\mathbf{G}^*} + \mathbf{n}_r(\mathbf{v}_r), \quad (13)$$

$$= \mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \Delta_r^*(\mathbf{w}, \mathbf{w}) \mathbf{x}(\mathbf{w}) + \mathbf{n}_r(\mathbf{v}_r), \quad (14)$$

where $\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w})$ (or $\Delta_r^*(\mathbf{w}, \mathbf{w})$) represents the submatrix with row indices \mathbf{v}_r (or \mathbf{w}) and column indices \mathbf{w} , and $\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{d})$ (or $\Delta_r^*(\mathbf{d}, \mathbf{d})$) represents the submatrix with row indices \mathbf{v}_r (or \mathbf{d}) and column indices \mathbf{d} . In (13), the term \mathbf{G}^* denotes the ICI caused by the data, and we have $\mathbf{G}^* = \mathbf{0}$ due to its corresponding entries of the dominant basis are zero, i.e., $\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{d}) = \mathbf{0}$. Thus, it is easy to find the received

pilots are free of the ICI but with a permutation of the receive subcarriers. The relationship between \mathbf{v}_r and \mathbf{w} is given as

$$v_{r,p} = \left| w_p + \left(q_r^* - \frac{Q}{2} \right) \right|_K, \quad w_p \in \mathbf{w}, \quad v_{r,p} \in \mathbf{v}_r, \quad (15)$$

where $p = 1, 2, \dots, P$, and $|\cdot|_K$ denotes the mod K operator.

IV. PILOT DESIGN FOR COMPRESSED CHANNEL ESTIMATION

A. CS fundamentals

Considering an unknown signal $\hat{\mathbf{x}} \in \mathbb{C}^M$, suppose that $\hat{\mathbf{x}} = \Phi \mathbf{a}$, where $\Phi \in \mathbb{C}^{M \times U}$ denotes a known dictionary matrix and $\mathbf{a} \in \mathbb{C}^U$ denotes a S -sparse vector, i.e., $\|\mathbf{a}\|_{\ell_0} = S \ll U$. Then, CS considers the following problem

$$\hat{\mathbf{y}} = \Psi \hat{\mathbf{x}} + \boldsymbol{\eta} = \Psi \Phi \mathbf{a} + \boldsymbol{\eta}, \quad (16)$$

in which $\Psi \in \mathbb{C}^{V \times M}$ presents a known measurement matrix, $\hat{\mathbf{y}} \in \mathbb{C}^V$ presents the observed vector, and $\boldsymbol{\eta} \in \mathbb{C}^V$ is the noise vector. The objective of CS is to reconstruct \mathbf{a} accurately based on the knowledge of $\hat{\mathbf{y}}$, Ψ , and Φ .

Definition 2 (Average coherence [4]): Considering a matrix \mathbf{M} with the i -th column as \mathbf{g}_i , its average coherence is defined as the average of all absolute inner products between any two normalized columns in \mathbf{M} that are beyond a threshold δ , where $0 < \delta < 1$. Put formally

$$\mu_\delta\{\mathbf{M}\} = \frac{\sum_{i \neq j} (|g_{ij}| \geq \delta) \cdot |g_{ij}|}{\sum_{i \neq j} (|g_{ij}| \geq \delta)}, \quad (17)$$

where $g_{ij} = \tilde{\mathbf{g}}_i^H \tilde{\mathbf{g}}_j$, $\tilde{\mathbf{g}}_i = \mathbf{g}_i / \|\mathbf{g}_i\|_{\ell_2}$, and the operator is defined as $(x \geq y) = 1$ for $x \geq y$ or 0 for $x < y$.

It has been established in [4] that a smaller $\mu_\delta\{\Psi\Phi\}$ will lead to a more accurate recovery of \mathbf{a} . Thus, it can be expected that if Ψ is designed with a fixed Φ such that $\mu_\delta\{\Psi\Phi\}$ is as small as possible, then CS can get better recovery performance.

B. Position-based channel estimation and pilot design

In this paper, we assume that each receive antenna estimates its channel individually, and then sends the estimated channel to the RS for operation. Then, (14) can be rewritten as

$$\mathbf{y}_r(\mathbf{v}_r) = \mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \mathbf{S}(\mathbf{w}, :) \mathbf{c}_r^* + \mathbf{n}_r(\mathbf{v}_r), \quad (18)$$

where $\mathbf{S}(\mathbf{w}, :) = \text{diag}\{\mathbf{x}(\mathbf{w})\} \mathbf{F}_L(\mathbf{w}, :)$, and $\mathbf{S}(\mathbf{w}, :)$ and $\mathbf{F}_L(\mathbf{w}, :)$ denote their submatrices with row indices \mathbf{w} and all columns. In this way, the task of estimating the high-mobility channel \mathbf{H}_r in the frequency domain is converted to estimating the sparse dominant coefficient vector \mathbf{c}_r^* , which highly reduce the estimation complexity.

According to Definition 2, we aim to design the pilot pattern \mathbf{w} to minimize the average coherence and thus to improve the channel estimation performance. In this paper, we only design the pilot pattern and assume the pilot symbols are the same. Therefore, the global pilot pattern design problem can be formulated as

$$\mathbf{w}^* = \arg \min_{\mathbf{w}} \max_r \mu_\delta\{\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \mathbf{S}(\mathbf{w}, :)\}, \quad (19)$$

where \mathbf{w}^* denotes the optimal pilot pattern, and $r = 1, 2, \dots, R$. Note that for a given \mathbf{w} , its corresponding \mathbf{v}_r can be obtained by (15). Thus, \mathbf{w} is the only variable in this problem.

Taking the expression of \mathbf{D}_r^* into consideration, the objective function can be represented as

$$\begin{aligned} & \mu_\delta\{\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \mathbf{S}(\mathbf{w}, :)\} \\ &= \mu_\delta \left\{ \mathbf{I}_K^{\langle q_r^* - \frac{Q}{2} \rangle}(\mathbf{v}_r, \mathbf{w}) \text{diag}\{\mathbf{x}(\mathbf{w})\} \mathbf{F}_L(\mathbf{w}, :) \right\}, \end{aligned} \quad (20)$$

$$= \mu_\delta \{\text{diag}\{\mathbf{x}(\mathbf{w})\} \mathbf{F}_L(\mathbf{w}, :)\}, \quad (21)$$

where we have $\mathbf{I}_K^{\langle q_r^* - \frac{Q}{2} \rangle}(\mathbf{v}_r, \mathbf{w}) = \mathbf{I}_P$ for $r = 1, 2, \dots, R$, and \mathbf{I}_P denotes a $P \times P$ identity matrix.

Suppose that each pilot symbol has the same constant amplitude A , i.e.,

$$|X(w_p)|^2 = A, \quad \forall w_p \in \mathbf{w}. \quad (22)$$

According to Definition 2, it is not difficult to prove that the average coherence is independent of the constant amplitude. Thus, (21) can be further rewritten as

$$\mu_\delta\{\mathbf{D}_r^*(\mathbf{v}_r, \mathbf{w}) \mathbf{S}(\mathbf{w}, :)\} = \mu_\delta \{\mathbf{F}_L(\mathbf{w}, :)\}. \quad (23)$$

In this way, the problem (19) is simplified to the following optimization problem

$$\mathbf{w}^* = \arg \min_{\mathbf{w}} \mu_\delta \{\mathbf{F}_L(\mathbf{w}, :)\}. \quad (24)$$

From (24), we find that the optimal pattern \mathbf{w}^* is independent of the train speed v , the Doppler shift f_r , the antenna number R , or the antenna position α_r . Thus, for the given HST system, we can offline pre-design \mathbf{w}^* and then sends it to each receive antenna to estimate the channel during the whole system runs.

An intuitive idea to solve (24) is to perform the exhaustive search. However, this is impractical due to huge computational complexity. Following the same spirit of the discrete stochastic optimization (DSO) in [7], we propose a low complexity pilot pattern design algorithm to solve this problem. Due to space limit and we focus on the mobility robust property of the optimal pilot pattern, please refer to our technical report [9] for the detailed algorithm.

V. SIMULATION RESULTS

In this section, we present the performance of the proposed scheme by two typical compressed channel estimators, BP [11] and OMP [12]. We assume that the $R = 2$ receive antennas are equipped, one at the front and the other at the end of the HST, respectively. The HST system parameters are given in Table I. We consider a 512-subcarrier OFDM system with 40 pilot subcarriers. The bandwidth is set to be 5MHz, the packet duration is $T = 1.2\text{ms}$, and the modulation is 4-QAM. We consider the CE-BEM channel model and each channel has $L = 64$ taps, but only 5 taps are dominant ones with random positions. The speed of the HST is 500km/h, i.e., the maximum Doppler shift is $f_{max} = 1.087\text{KHz}$.

Fig. 2 presents the MSE performances of BP estimators versus the r -th receive antenna position at SNR = 15dB and 30dB. As a reference, we also plot the Doppler shift f_r versus

TABLE I
HST COMMUNICATION SYSTEM PARAMETERS

Parameters	Variables	Values
BS cover range	R_{BS}	1000 m
HST length	L_{hst}	200 m
Max distance of BS to railway	D_{max}	1000 m
Min distance of BS to railway	D_{min}	40 m
Carrier frequency	f_c	2.35 GHz

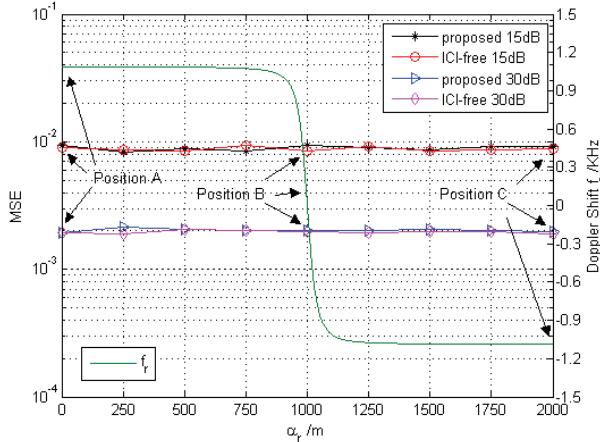


Fig. 2. MSE performances and the Doppler shift versus the antenna position.

the receive antenna position. “ICI-free” means that pilots are free of ICI, where the transmitted OFDM symbol is set as zero at the data subcarriers. All estimators are considered with the optimal pilot pattern designed in Section IV-B. It can be observed that the proposed method and the ICI-free one are almost superimposed at each position, which shows that the proposed method can effectively obtain the ICI-free pilots. In addition, although the HST suffers from large Doppler shift and f_r changes rapidly near B , we find that the MSE performances of the proposed method are stable, which shows that the proposed method is robust with respect to high mobility.

Fig. 3 shows the BER versus SNR of the 1×2 SIMO-OFDM system at the position A , where the Doppler shifts at the receive antennas are both 1.087KHz. In this figure, we compare three pilot pattern design methods: the equidistant pilot pattern [1] (equidi.), the exhaustive search based method [5] (exhaus.), and our designed pilot pattern (Alg.1). As can be seen, BP and OMP with Alg. 1 and the proposed ICI elimination method are closer to the one with the perfect CSI. It is also shown that “BP-Alg.1” with the proposed ICI elimination outperforms the one with the ICI mitigation method in [7] (BP-6 iterations) for effectively eliminating the ICI.

VI. CONCLUSION

In this paper, for the SIMO-OFDM HST system, we exploit the train position information and utilize it to mitigate the ICI caused by high mobility, especially for the CE-BEM. Then,

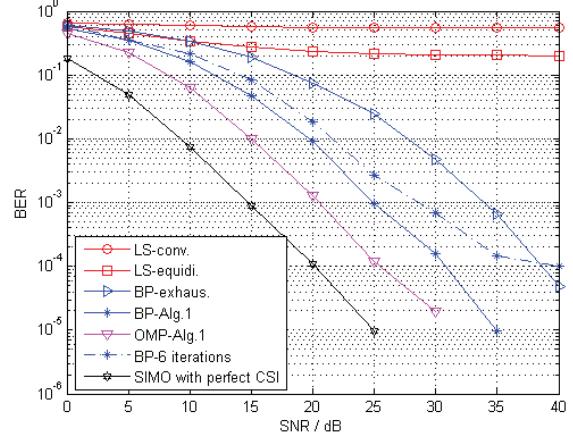


Fig. 3. BER performances of the 1×2 SIMO-OFDM system.

we design the pilot pattern to minimize the system coherence and hence can improve the CS-based channel estimation performance. Simulation results confirm the effectiveness and robustness of the proposed scheme in HST environments.

REFERENCES

- [1] X. Ma, G. Giannakis, and S. Ohno, “Optimal training for block transmissions over doubly-selective wireless fading channels,” *IEEE Trans. Signal Process.*, vol. 51, no. 5, pp. 1351-1366, May 2003.
- [2] G. Gui, W. Peng and F. Adachi, “High-resolution compressive channel estimation for broadband wireless communication systems,” *International Journal of Communication Systems*, vol. 27, no. 10, pp. 2396-2407, Dec. 2014.
- [3] W. U. Bajwa, J. Haupt, A. M. Sayeed, and R. Nowak, “Compressed channel sensing: a new approach to estimating sparse multipath channels,” *Proceedings of the IEEE*, vol. 98, no. 6, pp. 1058-1076, June 2010.
- [4] M. Elad, “Optimized projections for compressed sensing,” *IEEE Transaction on Signal Processing*, vol. 55, no. 12, pp. 5695-5702, Dec. 2007.
- [5] X. He and R. Song, “Pilot pattern optimization for compressed sensing based sparse channel estimation in OFDM systems,” *International Conference on Wireless Communications and Signal Processing (WCSP)*, pp. 1-5, Oct. 2010.
- [6] X. Ren, X. Shao, M. Tao, and W. Chen, “Compressed channel estimation for high-mobility OFDM systems: pilot symbol and pilot pattern design,” *IEEE International Conference on Communications (ICC)*, 2015.
- [7] X. Ren, W. Chen, and M. Tao, “Position-based compressed channel estimation and pilot design for high-mobility OFDM systems,” *IEEE Transactions on Vehicular Technology*, July 2014.
- [8] L. Liu, C. Tao, J. Qiu, H. Chen, L. Yu, W. Dong, and Y. Yuan, “Position-based modeling for wireless channel on high-speed railway under a viaduct at 2.35 GHz,” *IEEE Journal on Selected Areas in Communications*, vol. 30, no. 4, pp. 834-845, May 2012.
- [9] X. Ren, et al., “Compressed channel estimation with position-based ICI elimination for high-mobility SIMO-OFDM systems,” *Technical Report*, available at <https://drive.google.com/file/d/0B5hHqjlxFW8pMXU3X0dHeUg3Uk0/view?usp=sharing>
- [10] A. P. Kannu and P. Schniter, “MSE-optimal training for linear time-varying channels,” *Proc. IEEE Int. Conf. Acoust., Speech, Singal Process. (ICASSP)*, pp. 789-792, Mar. 2005.
- [11] S. S. Chen, D. L. Donoho, and M. A. Saunders, “Atomic decompositon by basis pursuit,” *SIAM Review*, vol. 43, pp. 129-159, 2001.
- [12] Y. C. Pati, R. Rezaifar, and P. S. Krishnaprasad, “Orthogonal matching pursuit: recursive function approximation with applications to wavelet decomposition,” *Proceedings of the 27th Annual Asilomar Conference on Signals, Systems and Computers*, vol. 1, pp. 40-44, Nov. 1993.