Statistics-Based Channel Estimation and ICI Mitigation in OFDM System over High Mobility Channel

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Abstract—In this paper, we propose a statistics-based channel estimation and inter-carriers interference (ICI) mitigation scheme for the orthogonal frequency division multiplexing (OFDM) transmission over high speed railway communication. The proposed scheme offers a new channel estimation method based on compressed sensing by utilizing the sparsity of high speed train (HST) in the time domain channel. The wireless channel is expressed by a weighted time-domain channel interpolation. The interpolation weights are designed according to the channel statistics, and the position of the selected parameter of each path is designed based on coherence of the model structure. Furthermore, by utilizing the channel statistics, we develop the receive post-processing in the receiver for ICI mitigation. Numerical and simulation results verify the effectiveness of the proposed schemes.

Keywords—Statistics channel, ICI mitigation, OFDM, Transmit post-processing, compressed sensing (CS).

I. INTRODUCTION

With the development of the the High Speed Train (HST), the speed of the train can reach up to 350Km/h, or even higher, so high data rate is demanded in wireless communication. Recently, orthogonal frequency division multiplexing (OFDM) has been adopted in this field due to high spectral efficiency and high data rate. However, the wireless channel becomes frequency-selective and time-varying within one OFDM symbol in the HST environment, which destroys the orthogonality and causes inter-carrier interference (ICI). As a result, a reliable channel estimation method should be investigated for the high mobility channel.

Channel estimation over high mobility channel has been considered in a number of recently papers, and various time-varying channel models have been established to reduce the number of parameters in the multipath channel [1]-[4], [1] and [2] assume the channel varies with time linearly in an OFDM symbol, however this model does not suit the high mobility channel which varies very fast. In [3] and [4], the doppler spread information is utilized to make a better approach to the channel model on the basis of statistics property, [5] proposed a channel estimation method based on compressive sensing (CS) by using multipath sparsity of the high mobility channel, however none of these papers exploit the statistics property to reduce the ICI and consider the line-of-sight (LOS) in our

HST model. In [6] , it proposes a statistics-based scheme to mitigate ICI, but it do not provide an effective channel estimation method.

In this paper, we propose a statistics-based channel estimation and ICI mitigation scheme in the OFDM system over high mobility channels. Firstly, we develop a receive post-processing in the receiver to mitigate the ICI by utilizing the channel statistics. Secondly, we will establish a new channel model based on the channel statistics. The time-domain channel coefficients are approximated by a weighted time-domain channel interpolation, where the interpolation weights are designed based on the Doppler and time-domain channel correlations, and the selected set of each path is designed in a way to improve the performance of proposed scheme. Then, we will transform the channel model to adapt the proposed CS based channel estimation scheme [7] combined with ICI mitigation. Finally, we will investigate the performance of the proposed scheme and show its performance gains compared with other existing schemes.

II. SYSTEM MODEL

Consider the wireless communication in the high speed train system shown in Fig. 1. The train travels in the rural area with the base station (BS) located away from the track. The receiver at the train is a relay station and it is placed on the top of the train. With the help of the GPS satellite, doppler frequency can be estimated conveniently from the speed and the position of the train. We describe the HST channel model at first.

A. Channel Model

Consider an OFDM system with N subcarriers with the sampling interval T_s, define $h_{n,l}$ as the lth channel tap at the nth time instant, and $l = 0, 1, ..., L - 1$. The first tap is assumed to be the LOS path and others correspond to the non-LOS (NLOS) paths. For the LOS path, we have

$$h_{0,n} = h_0 e^{j2\pi f_d n T_s}, 0 \le n \le N - 1, \quad (1)$$

where h_0 is the channel amplitude without considering the initial phase, and the power gain of the LOS path is $\xi_0 = |h_0|^2$, the Doppler frequency shift (DFS) is defined as $f_d = f_D \cos(\theta)$, in which f_D is the maximum Doppler frequency and the θ is the *angle of arrival* (AOA) of the LOS path. In
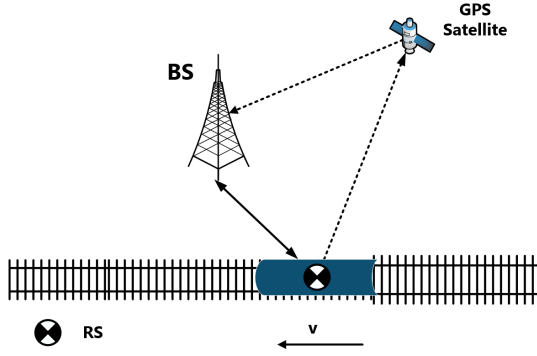
Fig. 1. The structure of the HST communication system

this paper, the DFS is defined as a constant which is dependent on the speed and location of the train within an OFDM symbol. For the NLOS paths, they are assumed to follow the characteristic of the Rayleigh channel. The correlation of the time-vary channel can be expressed as

$$E[h_{n,l}h_{m,l}] = \xi_l J_0(2\pi f_d(m-n)T_s) \triangleq \xi_l J(m-n), \quad (2)$$

in which $1 \le l \le L-1, 0 \le n, m \le N$, $J_0(\cdot)$ is the zero-order Bessel function and ξ_l is the average power gain over the lth channel tap, and $\sum_{l=0}^{L-1} \xi_l = 1$.

B. High-Mobility OFDM

Let us start with the channel input/output relationship in the discrete form. The transmitted signals in the frequency domain pass the serial to parallel converter (S/P) followed by N-point inverse DFT (IDFT) after modulation,

$$x(n) = \frac{1}{N} \sum_{k=0}^{N-1} X(k) e^{-j\frac{2\pi kn}{N}}, \quad (3)$$

where $X(k)$ is the data at the kth subcarrier of the OFDM symbol, and the $x(n)$ is the transmitted signal at the nth time instant. Let $y(n)$ represent the received signal at the nth time instant. It is given by

$$y(n) = \sum_{l=0}^{L-1} h_{l,n} x((n-l)_N) + w(n), \quad (4)$$

where $w(n)$ stands for the additive white Gaussian noise (AWGN) with a zero mean and variance σ_w^2, and $(\cdot)_N$ represents a cyclic shift in the base of N. After DFT, the receive signal in the frequency domain is

$$\mathbf{Y} = \mathbf{F}\mathbf{H}\mathbf{F}^H\mathbf{X} + \mathbf{W} = \mathbf{G}\mathbf{X} + \mathbf{W}, \quad (5)$$

where $\mathbf{F}$ and $\mathbf{F}^H$ are the DFT and IDFT matrix, and $\mathbf{H}$ and $\mathbf{G}$ denote the $N \times N$ time-domain channel matrix and frequency-domain channel matrix respectively. $\mathbf{H}$ is given by

$$H = \begin{pmatrix} h_{0,0} & 0 & \cdots & h_{2,0} & h_{1,0} \\ h_{1,1} & h_{0,1} & \cdots & h_{3,1} & h_{2,1} \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & h_{1,N-1} & h_{0,N-1} \end{pmatrix}. \quad (6)$$

If the channel is static or slowly changing, the time variation of the channel within an OFDM symbol can be neglected. Then $\mathbf{H}$ will be a circulant matrix and $\mathbf{G}$ will be a diagonal matrix. However, the channel is varying rapidly in the high mobility scene, this destroies the circularity of $\mathbf{H}$, and thus $\mathbf{G}$ is no longer a diagonal matrix, with the nonzero off-diagonal entries inducing ICI. The diagonal and off-diagonal elements of $\mathbf{G}$ are represented as

$$G(k_1, k_2) = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} h_{n,l} e^{-j\frac{2\pi}{N}lk_2} e^{-j\frac{2\pi}{N}(k_1-k_2)n}$$
$$, 0 \le k_1, k_2 \le N-1, k_1 \ne k_2,$$

$$G(k, k) = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} h_{n,l} e^{-j\frac{2\pi}{N}lk}. \quad (7)$$

Then the received signal can be expressed as

$$\mathbf{Y} = \mathbf{G}_{ICIfree}\mathbf{X} + \mathbf{G}_{ICI}\mathbf{X} + \mathbf{W}, \quad (8)$$

where $\mathbf{G}_{ICIfree}$ is a $N \times N$ diagonal matrix whose elements are the diagonal of $\mathbf{G}$, and $\mathbf{G}_{ICI}$ is a $N \times N$ matrix whose diagonal elements are zero and off-diagonal elements are $G(k_1, k_2)$ in (7).

III. RECEIVE POST-PROCESSING FOR ICI MITIGATION

As described in Section II-B, in the high mobility environment, the frequency-domain channel matrix $\mathbf{G}$ is not a diagonal matrix, therefore we can reduce the residual ICI by concentrating the power of $\mathbf{G}$ on its diagonal entries. In this section, we develop the *statistics-based receive post-processing* to perform at the receiver for ICI mitigation. With the post-processing, the actual received signal vector in frequency domain is $\mathbf{R} = \mathbf{V}\mathbf{Y}$ where $\mathbf{V}$ denote the $N \times N$ receive post-processing matrix. Eq. (5) is rewritten as follows

$$\mathbf{R} = \mathbf{V}\mathbf{Y} = \mathbf{V}\mathbf{F}\mathbf{H}\mathbf{F}^H\mathbf{X} + \mathbf{V}\mathbf{W} = \mathbf{D}\mathbf{X} + \mathbf{V}\mathbf{W}, \quad (9)$$

where $\mathbf{D} = \mathbf{V}\mathbf{G}$ is the equivalent channel frequency matrix. Since the transmit signal is independent with zero mean and unit variance, denote $P(i,j) = E\{|D(i,j)|^2\}$, then the post SIR at the ith subcarrier is given by

$$\mathbf{SIR}(i) = \frac{p(i,i)}{\sum_{j=0,j\ne i}^{N-1} p(i,j)}. \quad (10)$$

Our objective is to design the optimal matrix $\mathbf{V}$ to maximize the SIR. Consider the circular and symmetric structure of the OFDM modulation, it is expected that, with the optimal $\mathbf{V}$, the power gain only depends on the distance between the two involved subcarrier, i.e.,$P(i,j) = p(i-j)_N$, where $(\cdot)_N$ denotes the operation modulo N and $p(\cdot)$ is a one-dimensional function, and it is assumed that all the subcarriers have the same post SIR. Since $\mathbf{D} = \mathbf{V}\mathbf{G}$, it can be shown from (7)

that

$$P(i,j) = \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} V(i,k_1)V^*(i,k_2)E\{G(k_1,j)G^*(k_2,j)\}$$
$$= \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} V(i,k_1)V^*(i,k_2)f[(k_1-j)_N,(k_2-j)_N],$$

$$(11)$$

where function f is denoted as

$$f(s_1,s_2) = \frac{1}{N^2} \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} c(n_1-n_2)e^{\frac{j2\pi[s_2n_2-s_1n_1]}{N}} \quad (12)$$

$$c(n_1-n_2) = \xi_0 e^{j2\pi f_d(n_1-n_2)T_s} + \sum_{l=1}^{L-1} \xi_l J(n_1-n_2).$$

$$(13)$$

Let $\mathbf{V}$ be a circulant marix constructed from vector $v = (v_0,v_1,\cdots,v_{N-1},)^T$ with $\Sigma_{i=0}^{N-1}|v_i|^2 = 1$, i.e.,$V(i,j) = v_{(i-j)_N}$ to satisfy the equation as

$$V(i,k_1)V^*(i,k_2) = g[(k_1-i)_N,(k_2-i)_N], \quad (14)$$

and all the subcarriers have the same received signal power given by

$$P(0) = \sum_{m=0}^{N-1} P(0,m)$$
$$= \sum_{m=0}^{N-1} \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} g(k_1,k_2)f[(k_1-m)_N,k_2-m)_N],$$

$$(15)$$

and then we obtain

$$P(0) = c(0)\sum_{k_2=0}^{N-1} g(k_2,k_2) = (\xi_0 + \sum_{l=1}^{L-1} \xi_l)\sum_{k_2=0}^{N-1} g(k_2,k_2) = 1.$$

$$(16)$$

On the other hand, all the subcarriers have the same desired signal power which is given by

$$P_d = P(0,0)$$
$$= \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} V(0,k_1)V^*(0,k_2)f[(k_1-0)_N,(k_2-0)_N]$$
$$= \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} v_{k_1}f[k_1,k_2]v_{k_2}^*$$
$$= \mathbf{v}^T\mathbf{F_v}\mathbf{v}^*.$$

$$(17)$$

Then the post SIR can be expressed as

$$SIR(i) = \frac{P_d}{P(0)-P_d} = \frac{\mathbf{v}^T\mathbf{F_v}\mathbf{v}^*}{1-\mathbf{v}^T\mathbf{F_v}\mathbf{v}^*}. \quad (18)$$

Therefore, the optimal matrix $\mathbf{V}$ can be obtained by maximizing the $\mathbf{v}^T\mathbf{F_v}\mathbf{v}^*$, and the optimal $\mathbf{v}$ is the conjugate of the eigenvector of $\mathbf{F_v}$ associated with its largest eigenvalue $\lambda_{max}(\mathbf{F_v})$, then the maximum SIR is given by $\mathbf{SIR}_{max} = \frac{\lambda_{max}(\mathbf{F_v})}{1-\lambda_{max}(\mathbf{F_v})}$.

IV. PROPOSED SCHEME FOR CHANNEL ESTIMATION BASED ON COMPRESSED SENSING

In this section, we provide a novel channel estimation scheme based on CS for the communication in high speed train scene, which includes the LOS path and the NLOS paths.

A. Model Reconstruction

The Eq. (9) can be further written [8] as

$$\mathbf{R} = (\mathbf{V} \otimes \mathbf{X}^T)(\mathbf{F} \otimes (\mathbf{F}^H)^T)\mathbf{H}_{vec} + \mathbf{VW}, \quad (19)$$

where $\otimes$ is the Kronecker operator, and $\mathbf{H}_{vec}$ is a vector designed by stretching the $N \times N$ matrix $\mathbf{H}$ to a $N^2 \times 1$ vector, which is denoted as

$$\mathbf{H}_{vec} = [\mathbf{h}_0,\mathbf{h}_1,\cdots,\mathbf{h}_{N-1}]^T, \quad (20)$$

here $\mathbf{h}_n$ represents the impulse response vector of the time-domain channel at time n, which corresponds to the nth row of $\mathbf{H}$, and is equal to

$$\mathbf{h}_n = [h_{n,n},h_{n-1,n},\cdots,h_{0,n},0,\cdots,h_{n+1,n}]. \quad (21)$$

Let $\mathbf{U}$ stand for the preceding matrix of equation (19), that is $\mathbf{U} = (\mathbf{V} \otimes \mathbf{X}^T)(\mathbf{F} \otimes (\mathbf{F}^H)^T)$, so rewrite the equation (19)

$$\mathbf{R} = \mathbf{UH}_{vec} + \mathbf{VW}. \quad (22)$$

According to Eq. (6), $\mathbf{H}_{vec}$ has NL nonzero elements, therefore we can reduce the dimension of vector $\mathbf{H}_{vec}$ from $N^2 \times 1$ to $NL \times 1$. Meanwhile, as $\mathbf{H}_{vec}$ has a permutation decided by the sampling time, we rearrange it on the basis of channel path in order to make it simple to analyze in the following section, the new vector is given as

$$\mathbf{H}_{vp} = [h_{0,0},\cdots,h_{0,N-1},\cdots,h_{L-1,0},\cdots,h_{L-1,N-1}]^T.$$

$$(23)$$

Meanwhile, we transform the preceding matrix $\mathbf{U}$ as we do $\mathbf{H}_{vec}$. At first, we remove the column vectors which are corresponding to the zero parameters of $\mathbf{H}_{vec}$, and rearrange the position of matrix's column vector corresponding to the change of $\mathbf{H}_{vp}$. The new preceding matrix is denoted as $\mathbf{U}_N$, and the Eq. (22) can be expressed as

$$\mathbf{R} = \mathbf{U}_N\mathbf{H}_{vp} + \mathbf{VW}. \quad (24)$$

B. Time-domain Channel Matrix Reconstruction

As shown in the above sections, the number of coefficient for estimating in time-varying channel is generally much larger than the maximum number of observable samples, which is a NP-hard problem. To resolve this problem, we propose to reduce the number of non-zero channel coefficients that need to be estimated.

In the high speed train model, there are two cases in the channel propagation, that is LOS and NLOS [9]. As for the LOS path, the channel varies with time in a deterministic way for a given DFS. On the other hand, the NLOS path follows the Rayleigh fading channel and the Jacks model [10]. So we can find a way to reduce the number of these channel parameters and make a approach to the original channel.

If the DFS and the sampling interval T_s are given, we can obtain the whole elements of the LOS path only depend on a known parameter. Assuming that the position of the known parameter is m_0, then the time domain channel of the LOS path can be expressed as

$$[h_{0,0}, \cdots, h_{0,m_0}, \cdots, h_{0,N-1}]^T = \mathbf{C}_0 h_{0,m_0},$$
(25)

where $\mathbf{C}_0$ is a $N \times 1$ vector denoted as

$$C_0 = [e^{j2\pi f_d(-m_0)T_s}, \cdots, 1, \cdots, e^{j2\pi f_d(N-1-m_0)T_s}]^T.$$
(26)

For the NLOS paths, we utilize a small number of parameters to approach the whole channel path [4]. For example, we define $\mathbf{h}(l) = [h_{l,m(1)}, \cdots, h_{l,m(M)}]^T$ as the selected parameter vector of the lth path, where $M \ll N$. The impulse response of the lth path at time n can be expressed as

$$h_{l,n} = \mathbf{C}_{l,n}[h_{l,m(1)}, \cdots, h_{l,m(M)}]^T = \mathbf{C}_{l,n}\mathbf{h}(l),$$
(27)

where $\mathbf{C}_{l,n}$ is a $1 \times M$ vector designed to minimize the channel estimation error defined as $E[|h_{l,n} - \mathbf{C}_{l,n}\mathbf{h}(l)|^2]$. By using the orthogonality principle, we obtain

$$\mathbf{C}_{l,n} = E[h_{l,n}\mathbf{h}(l)^H]E[\mathbf{h}(l)\mathbf{h}(l)^H]^{-1}.$$
(28)

Considering (2), (28) can be further expressed as

$$\mathbf{C}_{l,n} = [J[m(1) - n], \cdots, J[m(M) - n]]$$
$$\cdot \begin{pmatrix} J[m(1) - m(1)] & \cdots & J[m(1) - m(M)] \\ \vdots & \ddots & \vdots \\ J[m(M) - m(1)] & \cdots & J[m(M) - m(M)] \end{pmatrix}^{-1}.$$
(29)

Note that if the positions of the M selected parameters are determined, then the vector $\mathbf{C}_{l,n}$ is determined, and the lth path can be expressed as

$$[h_{l,0}, \cdots, h_{l,N-1}]^T = \mathbf{C}_l[h_{l,m(1)}, \cdots, h_{l,m(M)}]^T = \mathbf{C}_l\mathbf{h}(l),$$
(30)

where $\mathbf{C}_l$ is a $N \times M$ matrix denoted as

$$\mathbf{C}_l = [\mathbf{C}_{l,0}^T, \mathbf{C}_{l,1}^T, \cdots, \mathbf{C}_{l,N-1}^T]^T.$$
(31)

Note that the coefficient matrix only depends on the location of the M selected parameters, and it has nothing to do with the path order l. Therefore, all the NLOS paths have a same coefficient matrix $\mathbf{C}_l$ if their selections are the same.

C. CS-based Channel estimation scheme

In this section, we regenerate the channel model (24) in term of the time-domain channel matrix reconstruction. Substitute (26) and (31) to (24), we obtain

$$\mathbf{R} = \underbrace{\mathbf{U}_N\mathbf{C}}_{\mathbf{U}_{NC}}\mathbf{h}_M + \mathbf{V}\mathbf{W},$$
(32)

where $\mathbf{C}$ is a block matrix of size $NL \times ((L-1)M+1)$, and is given as

$$\mathbf{C} = \begin{pmatrix} C_0 & 0 & \cdots & 0 \\ 0 & C_1 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & C_{L-1} \end{pmatrix},$$
(33)

and $\mathbf{h}_M$ is a $((L-1)M+1) \times 1$ vector, denoted as

$$\mathbf{h}_M = [h_{0,m_0}, \mathbf{h}(1)^T, \cdots, \mathbf{h}(L-1)^T]^T.$$
(34)

In this paper, we assume that the coefficients are constant within an OFDM symbol and different for different OFDM symbols. Note that many of practical wireless channels have sparse multipath structure, which means the number of non-zero channel taps, denoted by $\bar{L}$, is much lower than the maximum time delay L. Therefore, a large number of the elements of $\mathbf{h}_M$ are approximately zero and CS can be used to estimate the channel $\mathbf{H}_M$.

According to the CS theory, the purpose is to recover the vector $\mathbf{h}_M$ based on the knowledge of $\mathbf{R}$ and $\mathbf{U}_{NC}$, and $\mathbf{U}_{NC}$ should satisfy the restricted isometry property (RIP) to recover the vector $\mathbf{h}_M$ correctly. Therefore, as presented in [4] and [11], the number of pilot symbols at the receiver, denoted as N_p, is subject to

$$\mathbf{N}_p \gtrsim O(((\bar{L} - 1)M + 1)\log(N/\bar{L})).$$
(35)

Furthermore, a way to optimize the CS performance is to reduce the coherence between different normalized columns in $\mathbf{U}_{NC}$ [12]. As for the matrix $\mathbf{C}$, $C_1 = C_2 = \cdots = C_{L-1}$ if all the NLOS paths choose the same time-domain sample, which means the location of $m(1), m(2), \cdots, m(M)$ is unchanged in the NLOS paths. This scheme has a negative effect on the performance of the CS channel estimation method. Therefore, optimizing the location of NLOS paths is an achievable way to improve the performance of the proposed scheme. Considering the system model, the matrix $\mathbf{U}_N$ is a constant matrix if the DFS and pilot are known, and we can obtain it in advance, so a optimal matrix $\mathbf{C}$ can be obtained by optimization method to minimize the coherence in order to has a better performance. Meanwhile, all the processes are off-line which lead to a low complexity of channel estimation.

V. SIMULATION RESULTS AND DISCUSSION

In this section, we present numerical and simulation results to demonstrate the performance of the proposed scheme, and compare it with the schemes in [5] and [6]. The system parameters for the simulations are based on the LTE system parameters. In particular, we consider an OFDM system with 128 subcarriers and the number of pilots is 16. The carrier frequency is 2.0 **GHz** and sampling duration is $T_s = 1\mu s$. We define $F_d = NT_s f_d$ as the normalized DFS over the LOS path. Thus $F_d = 0.1$ when the Doppler frequency spread is $f_d = 0.781\mathbf{KHz}$ and the train travels at the speed of 420km/h. It is assumed that the maximum time delay of the channel L equals to 9 and only 3 taps are nonzero for each time n.
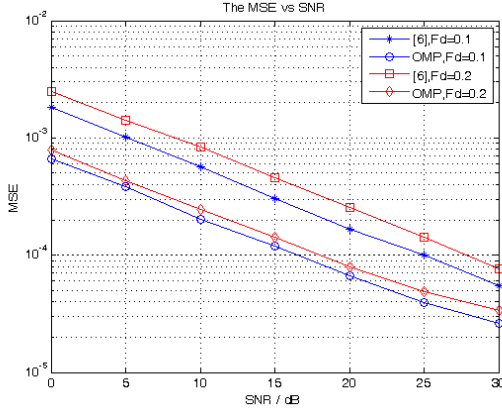
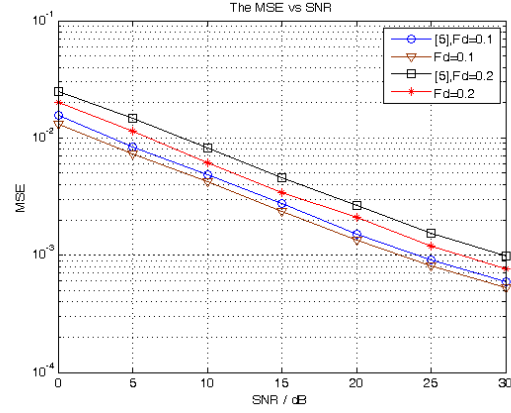Fig. 2.   MSE performance of LS and OMP estimation

Fig. 3.   MSE performance of general model and proposed model

Furthermore, we assume the Jacks Doppler spectrum applies to NLOS paths, and the pilot position are equidistant before data transmission.

Fig. 2 illustrates the comparison of the MSE performances of the proposed scheme and the scheme in [6] under different normalized Doppler spreads. Let $M = 3$ for the NLOS paths, and they have the same positions of the selected parameters in each path. It can be seen that the Orthogonal Matching Pursuit (OMP) channel estimators significantly improve the MSE performance by utilizing the inherent sparsity of the high mobility channels under the same ICI mitigation method. During the simulation, we find that our OMP preforms well and saves spectrum resources compared to the LS.

Fig. 3 depicts the comparison of the MSE performances of the proposed scheme without ICI mitigation and the scheme in [5] under different normalized Doppler spreads. As for the channel model in [5], they have the same time domain sampling position in all the paths, In our proposed scheme, they are different over each path in order to get the optimal matrix $\mathbf{C}$, which reduce the average coherence of the matrix $\mathbf{U}_{NC}$ to optimize the CS algorithm. It is an effective way to improve performance of channel estimation based on compressive sensing, and numerical results demonstrate that the proposed scheme gets a better channel estimation performance over high mobility channel.

VI. CONCLUSIONS

In this paper, we propose a new statistics-based channel estimation and ICI mitigation scheme for the OFDM transmission over a high-speed railway communication. In the proposed scheme, we develop the receive post-processing in the receiver for the ICI mitigation by utilizing the channel statistics, and a compressed sensing recovery method utilizing the sparsity of the high mobility system is proposed for the channel estimation. In addition, the time domain channel is modeled by some selected parameters of each path, and their coefficient matrices are conducive to CS method. Numerical and simulation results verify the effectiveness of the proposed schemes.

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