Channel Modeling and Estimation for OFDM Systems in High-speed Trains Scenarios

Yuming Bi*, Jianhua Zhang*,Ming Zeng*, Xiaodong Xu*
*Key Laboratory of Universal Wireless Communications, Ministry of Education Beijing University of
Posts and Telecommunications, Mailbox NO.92, China, 100876
Email:yumingbi@bupt.edu.cn

Abstract—**With the fast development of high-speed trains (HST) globally, some related technical issues of HST communications should be addressed, e.g. time-variant (TV) radio channel modeling and channel estimation. In this paper, we propose a more generic TV channel model based on Qian Liu's research. Moreover, a generalized subspace-based channel estimation algorithm with the aid of an improved basic expansion model (BEM) is proposed for orthogonal frequency-division multiplexing (OFDM) systems in high-speed mobile environments. The proposed channel model is based on the sum-of-sinusoids (SoS) method and accounts for the fast fading characteristics of HST channels due to the terminal's movement. The statistical analysis shows that the proposed model can be considered as a more general channel for TV-HST scenarios. Additionally, to accurately estimate the TV-HST channels, the improved BEM is applied to reconstruct the model by using the subspace projection method. Finally, numerical simulation illustrates that the accuracy of our channel estimation algorithm.**

I. INTRODUCTION

With the rapid development of HST and the growing train speed, the current HST communication systems are facing many challenges and some related technical issues of them, e.g. TV channel modeling and channel estimation should be addressed. As one of the indispensable parts in the communication systems, HST channels have been widely studied [1]. However, due to the TV characteristics of HST channels, it is very difficult to design an accurate and concise simulation channel to describe the actual scenarios, and in turn it leads to a significant degradation in the performance of the channel estimation algorithm [2]. Therefore, a perfect HST communication system should be designed by jointly considering TV channel modeling and estimation algorithm [3] [4].

On the one hand, a number of channel models for HST environments are available in the literatures [3-5]. Among them, the well known Xiao's model [5], a modified simulation model of Jake's model [6], is the most widely accepted one. Furthermore based on this model, Qian Liu has proposed his HST channel model in [7]. However, since both the Xiao's model and Qian Liu's model have been simulated using a fixed maximum Doppler frequency which is time-invariant. Thus they are not suitable for the TV-HST scenarios. Meanwhile, according to the theory of geometrically-based stochastic models (GBSM), the MIMO non-stationary channel models characterized by the cluster dynamic evolution on both transmitter and receiver has been widely investigated in [8], [9] and [10]. In [11] [12], Bo Ai and Ruisi He have devoted their measurement data to the researches on the HST channel modeling. In addition, some special technologies have been utilized in the field of HST channel modeling, such as RIMAX and Ray-tracing [13] [14].

On the other hand, although many methods are able to estimate the general channel precisely under low-speed movement, because the estimation performance of the receiver decreases a lot due to the the fast varying channel within one OFDM symbol subframe's period under high-speed mobile environments. To overcome this shortcoming, some works resort to basis expansion model (BEM), which can be employed by the different basis expansion function to reconstruct the real-world channels [15]. In [16] [17], the discrete Karhuen-Loeve BEM (DKL-BEM), which is optimal from the BEM has been discussed. In [18], the authors have proposed a complex-exponential BEM (CE-BEM) for channel estimation. Recently, a great deal of attention has been paid to the discrete prolate spheroidal BEM (DPS-BEM), which is more convenient to estimate the channel coefficients via rectangular spectrum [19].

In this paper, we propose a generic TV channel model which improved by the Qian Liu's works. Considering the changes of the velocity numerical magnitudes and directions, this new model can simulate a variety of fast fading channels with different mobile motion states, such as static, motion with the constant speed and acceleration. Then a generalized subspace-based channel estimation algorithm with the aid of BEM is proposed for OFDM systems in high-speed mobile environments. Finally, we apply our proposed channel estimation algorithm to reconstruct our proposed TV-HST channels for MIMO-OFDM systems. Comparing the reconstructed model and proposed model, the result indicates the precision and efficiency of the proposed estimation algorithm and channel model for TV-HST environments. The paper is organized as follows: the proposed TV-HST channel model is described in Section II. The system model is discussed in Section III. Then the proposed BEM and subspace-based channel estimation algorithm are provided in Section IV. The related simulation and numerical analysis are presented in Section V. At last, the conclusion is drawn in Section VI.

II. TV-HST CHANNEL MODELS

In this section, we firstly introduce the Xiao's model which is mainly referred to as a mathematical reference. When the

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signal is sampled by discrete times, the channel impulse response (CIR) of the normalized Xiao's two dimensional (2D) isotropic scattering fading model is given by

$$X(i) = X_c(i) + jX_s(i) \tag{1}$$

$$X_c(i) = \sqrt{\frac{1}{N}} \sum_{n=0}^{N} \cos[2\pi f_d T_s (i+1) \cos(\frac{2\pi n + \theta}{N}) + \phi_n] \tag{2}$$

$$X_s(i) = \sqrt{\frac{1}{N}} \sum_{n=0}^{N} \sin[2\pi f_d T_s (i+1) \cos(\frac{2\pi n + \theta}{N}) + \phi_n] \tag{3}$$

where $n = 1, 2, \ldots, N$, N is the number of propagation paths, f_d is the maximum Doppler frequency. θ and ϕ_n are the angle of arrival (AoA) and initial phase of the n-th propagation path respectively. Both of them are uniformly distributed over $[-\pi, \pi)$ for all θ and mutually independent. T_s means sampling interval over the whole fading process.

Assume that the base station (BS) is static while the HST is moving with a velocity varying in magnitude and direction. If $\mathbf{v_i}$ means the velocity of the train at the i-th instant, then it can be expressed as

$$\mathbf{v_i} = |v_i|\, e^{j\theta_{v_i}} \tag{4}$$

where θ_{v_i} and v_i are the direction and magnitude of the velocity at the i-th instant. As shown in Fig.1, β_i is the AoA of the i-th plane wave. Then based on the Qian Liu's model, the channel fading process becomes

$$Y(i) = Y_c(i) + jY_s(i) \tag{5}$$

$$Y_c(i) = \sqrt{\frac{1}{N}} \sum_{n=0}^{N} \cos[\frac{2\pi}{\lambda} T_s \sum_{k=1}^{i} |v_k| \cos\alpha_{nk} + \phi_n] \tag{6}$$

$$Y_s(i) = \sqrt{\frac{1}{N}} \sum_{n=0}^{N} \sin[\frac{2\pi}{\lambda} T_s \sum_{k=1}^{i} |v_k| \cos\alpha_{nk} + \phi_n] \tag{7}$$

where $\alpha_{nk} = \beta_n - \theta_{v_k}, \beta_n = \frac{2\pi n + \theta_n}{N}, n = 1, 2, \ldots, N$ Similar to the Xiao's model, θ_n and ϕ_n are the AoA and initial phase of the i-th propagation path respectively. And they are uniformly distributed over $[-\pi, \pi)$ for all n and mutually independent.

When the line of sight (LOS) case and the stationarity intervals of TV-HST channel [20] are considered in the HST channel model, we get the improved model:

$$Z(i) = Z_c(i) + jZ_s(i) \tag{8}$$

$$Z_c(i) = \frac{[Y_c(i) + \sqrt{K} \cos(\frac{2\pi}{\lambda} \widetilde{T_s} \sum\limits_{k=1}^{i} |v_k| \cos\theta_0 + \phi_0)]}{\sqrt{1+K}} \tag{9}$$

$$Z_s(i) = \frac{[Y_s(i) + \sqrt{K} \sin(\frac{2\pi}{\lambda} \widetilde{T_s} \sum\limits_{k=1}^{i} |v_k| \cos\theta_0 + \phi_0)]}{\sqrt{1+K}} \tag{10}$$

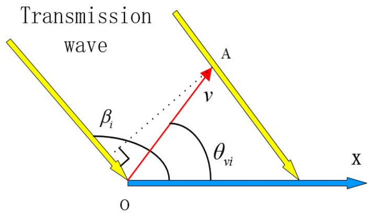

Fig. 1: The geometrical relationship between transmission wave and moving direction.

where θ_0 means the AoA of the LOS propagation path and $\widetilde{T_s}$ indicates the stationarity intervals of TV-HST channel, which can be acquired from the measurement campaign. In terms of other parameters, they are the same as those in Qian Liu's model. Furthermore, the new channel model can also be used to simulate the MIMO fading processes. Let $Z_l(i)$ be the l-th propagation factor in the i-th instant of time given by

$$\begin{aligned} Z_l(i) = &\sqrt{\frac{1}{N}} \sum_{n=0}^{N} \cos[\frac{2\pi}{\lambda} \widetilde{T_s} \sum_{k=1}^{i} (|v_k| \cos\alpha_{nk,l}) + \phi_{n,l}] + \\ &j\sqrt{\frac{1}{N}} \sum_{n=0}^{N} \sin[\frac{2\pi}{\lambda} \widetilde{T_s} \sum_{k=1}^{i} (|v_k| \cos\alpha_{nk,l}) + \phi_{n,l}] \end{aligned} \tag{11}$$

where $\alpha_{nk,l} = \beta_{n,l} - \theta_{v_k}, \beta_{n,l} = \frac{2\pi n + \theta_{n,l}}{N}, n = 1, 2, \ldots, N$ the values of $\theta_{n,l}$ and $\phi_{n,k}$ are statistically independent and follow uniformly distributed over $[-\pi, \pi)$ for all n.

III. SYSTEM MODEL

Consider a general MIMO-OFDM system with N_t transmitting antennas and N_r receiving antennas. Each OFDM symbol is modulated by N carriers. Let us now define the transmitted signals as

$$\mathbf{S} = [\mathbf{S_1^T}, \mathbf{S_2^T}, \ldots \mathbf{S_{nt}^T}, \ldots, \mathbf{S_{N_t}^T}]^\mathbf{T} \tag{12}$$

where $\mathbf{S_{nt}} = [\mathbf{S_{nt}(0)}, \mathbf{S_{nt}(1)}, \ldots, \mathbf{S_{nt}(N-1)}]^\mathbf{T}$. The received signals sequences are denoted as $\mathbf{Y} = [\mathbf{Y_1^T}, \mathbf{Y_2^T}, \ldots \mathbf{Y_{nr}^T}, \ldots, \mathbf{Y_{N_r}^T}]^\mathbf{T}$, where $\mathbf{Y_{nr}} = [\mathbf{Y_{nr}(0)}, \mathbf{Y_{nr}(1)}, \ldots, \mathbf{Y_{nr}(N-1)}]^\mathbf{T}$. The received signal stream within one OFDM system duration can be summarized as

$$\mathbf{Y} = (\mathbf{E_{N_r}} \otimes \mathbf{F_N})\mathbf{H}(\mathbf{E_{N_t}} \otimes \mathbf{F_N^H})\mathbf{S} + \mathbf{W} \tag{13}$$

where $\mathbf{E_{N_r}}$and $\mathbf{E_{N_t}}$ are $\mathbf{N_R} \times \mathbf{N_R}$ and $\mathbf{N_T} \times \mathbf{N_T}$ identity matrix respectively, $\mathbf{F_N}$ is an $\mathbf{N} \times \mathbf{N}$ unity discrete Fourier transform (DFT). $(\cdot)_H$ means Hermitian transpose, $\mathbf{W}$ denotes the additive white Gaussian noise matrix. And we can express the CIR matrix as

$$\mathbf{H} = \begin{bmatrix} \mathbf{H^{1,1}}, & \ldots & \mathbf{H^{1,N_T}} \\ \ldots & \mathbf{H^{n_r,n_t}}, & \ldots \\ \mathbf{H^{N_R,1}} & \ldots & \mathbf{H^{N_R,N_T}} \end{bmatrix} \tag{14}$$

where the sub-element H^{n_r,n_t} is presented as

$$\mathbf{H^{n_r,n_t}} = \begin{cases} [\mathbf{h^{n_r,n_t}}]_{\mathbf{i},\ \mathrm{mod}\ (\mathbf{i-j,N})}, 0 \leq \mathrm{mod}(i-j,N) \leq U \\ 0, \qquad\qquad \text{others} \end{cases} \tag{15}$$

where U represents the number of effective channel taps and $\mathbf{h^{n_r,n_t}}$ is the CIR between transmit antennas n_t and receive antennas n_r.

According to the theory of BEM, $\mathbf{h^{n_r,n_t}}$ can be represented as

$$\mathbf{h^{n_r,n_t}} = \mathbf{Bg^{n_r,n_t}} + \varepsilon \tag{16}$$

where $\mathbf{B}$ is the subspace basis function matrix. Q denotes the number of basis functions, which are related to the channel properties. $\mathbf{g^{n_r,n_t}}$ is the basis coefficient matrix and ε stands for the estimation error. Substituting equation (16) into (15), the sub-element $\mathbf{H^{n_r,n_t}}$ can be derived as

$$\mathbf{H^{n_r,n_t}} = \sum_{\mathbf{q=0}}^{\mathbf{Q}} \mathbf{diag(B(:,q))[F_N^H diag(F_L[g^{n_r,n_t,}(q,:)]^T)F_N]} \tag{17}$$

where $\mathbf{F_L}$ means the first columns of the matrix $\sqrt{N}\mathbf{F_N}$. To sum up the above arguments, Equation (13) can be rewritten as

$$\mathbf{Y^{n_r,n_t}} = \sum_{\mathbf{q=0}}^{\mathbf{Q}} \mathbf{O_q\{[E_{N_R} \otimes [S(E_{N_T} \otimes F_L)]\}g_q + W} \tag{18}$$

where $\mathbf{O_q} = \mathbf{(E_{N_R} \otimes F_N)(E_{N_R} \otimes diag(B(:,q))(E_{N_R} \otimes F_N^H)}$, $\mathbf{S} = \mathbf{[diag(S_1), ..., diag(S_{N_T})]}$, $\mathbf{g_q} = \mathbf{[g^{1,1}(q,:), ..., g^{N_R,N_T}(q,:)]^T}$.

IV. THE PROPOSED BEM AND CHANNEL ESTIMATION ALGORITHM

As mentioned in Section III, the key to estimate the TV-HST channel effectively comes from two aspects: the choice of BEM and the estimation algorithm. Therefore, in this section, we will discuss the modified BEM and our proposed subspace-based channel estimation algorithm.

A. The proposed BEM

As we all know, the BEM is designed to use less coefficients to characterize the TV channel and reconstruct the real channels by the linear combination of a series of basis functions. The general mathematical representation equation is given by (16). To explain BEM conveniently and clearly, the equation (16) can be simplified as

$$h(n,l) = \sum_{q=0}^{Q} g_q(\lfloor {}^{n}/_{N} \rfloor, l) p_{q,n}, l \in [0, L-1] \tag{19}$$

where $p_{q,n}$ is the q-th basis function, $g_q(\lfloor {}^{n}/_{N} \rfloor, l)$ denotes the q-th BEM coefficient of the l-th path, and Q represents the order of BEM.

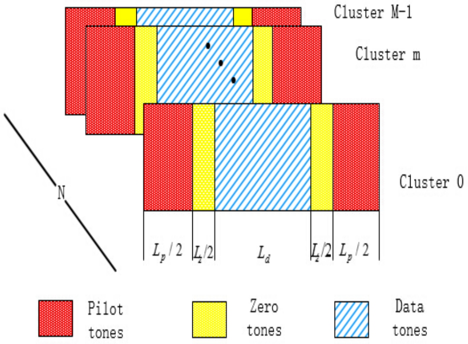

Fig. 2: Pilots program scheme in an OFDM symbol

B. The Proposed Channel Estimation Algorithm

Various traditional BEM designs have been proposed in[14-17], i.e. the CE-BEM [15], GCE-BEM [18] , P-BEM [16] and DKL-BEM [17]. The most commonly used model among them is the CE-BEM, which can be calculated as

$$h(n,l) = \sum_{q=0}^{Q} g_{q,l} e^{jW_q n} \tag{20}$$

Different from the equation (19), the basis function is $e^{jW_q n}$ and $w_q = 2\pi(q-Q/2)/N$. In order to eliminate the estimation errors, reference [16] developed the generalized Complex-Exponential BEM (GCE-BEM) by utilizing K times sampling:

$$h(n,l) = \sum_{q=0}^{Q} g_{q,l} e^{j2\pi \frac{q-Q/2}{KN} W_q n} \tag{21}$$

However, if we apply this model to estimate TV-channel, the finite basis functions will cause greater error between actual channel and reconstructed channel due to the fact that it is only designed by the normalization of the Doppler frequency. Accordingly, reference [19] provided a modified CE-BEM (MCE-BEM) to avoid this problem, which can be express as

$$h(n,l) = \sum_{q=0}^{Q} g_{q,l} e^{\frac{j2v_D N_S}{Q} W_q n} \tag{22}$$

where $w_q = 2\pi(q-Q/2)/N, w_q = 2\pi(-Q/2 + (q-1)Q/(Q-1))$, N_S is the number of OFDM symbol within one block. v_D is the normalized Doppler frequency. Because the factor $2v_D N_S/Q$ ensures that the highest frequency in this model is equivalent to the maximum Doppler frequency, the performance of BEM is significantly improved. But this model doesn't solved the effect of Gibbs, which is another cause to generate errors. Consequently, we propose an improved CE-BEM as

$$h(n,l) = \sum_{q=0}^{Q} g_{q,l} \{ e^{\frac{j2v_D N_S}{Q} W_q n} + A_q n \} \tag{23}$$

$$A_q = \frac{\{ e^{\frac{j2v_D N_S}{Q} W_q (N-B)} - e^{\frac{j2v_D N_S}{Q} W_q B} \}}{(N-2B)} \tag{24}$$

$$w_q = \frac{2\pi(-Q/2 + (q-1)Q)}{N} \tag{25}$$

where B is the simulation parameter (generally$B$=6). We assume that there are N subcarriers, which are divided into M clusters (as shown in Fig.2). Each cluster consists of L_p pilot tones, L_z zeros tones and L_d data tones. Especially, the pilot tones and zeros tones are interleaved with the data symbols. In this paper, we suppose that the zero tones are part of the pilots zones. Then the received signals are stacked together form the nonzero pilots as $Y^p = [(Y_0^p)^T, (Y_1^p)^T, ..., (Y_{M-1}^p)^T]^T$, where Y_m^P is the nonzero pilots sequences in clusters m. According to (18), we obtain

$$\mathbf{Y^P} = (\mathbf{A^P}\boldsymbol{\Delta}^\mathbf{P} + \mathbf{A^d}\boldsymbol{\Delta}^\mathbf{d})\mathbf{g} + \mathbf{W^P} \tag{26}$$

where

$$\mathbf{A^P} = \begin{pmatrix} \mathbf{a^P_{1,0}} & \cdots & \mathbf{a^P_{Q,0}} \\ \vdots & \ddots & \vdots \\ \mathbf{a^P_{1,M-1}} & \cdots & \mathbf{a^P_{Q,M-1}} \end{pmatrix} \tag{27}$$

$$\mathbf{A^d} = \begin{pmatrix} \mathbf{a^d_{1,0}} & \cdots & \mathbf{a^d_{Q,0}} \\ \vdots & \ddots & \vdots \\ \mathbf{a^d_{1,M-1}} & \cdots & \mathbf{a^d_{Q,M-1}} \end{pmatrix} \tag{28}$$

$$\boldsymbol{\Delta}^\mathbf{P} = \mathbf{E_Q} \otimes \{\mathbf{E_{N_R}} \otimes [\mathbf{S^P}(\mathbf{E_{N_T}} \otimes \mathbf{F^P_L})]\} \tag{29}$$

$$\boldsymbol{\Delta}^\mathbf{d} = \mathbf{E_Q} \otimes \{\mathbf{E_{N_R}} \otimes [\mathbf{S^d}(\mathbf{E_{N_T}} \otimes \mathbf{F^P_L})]\} \tag{30}$$

And g is denoted as $g = [g_1^T, g_2^T, ..., g_Q^T]^T$. For a certain OFDM pilot placement scheme, the equation (22) can be simplified as

$$\mathbf{Y^P} = \mathbf{D}\mathbf{g} + \mathbf{d} + \mathbf{W} \tag{31}$$

where $\mathbf{D} = \mathbf{A^P}\boldsymbol{\Delta}^\mathbf{P}, \mathbf{d} = \mathbf{A^d}\boldsymbol{\Delta}^\mathbf{d}\mathbf{g}$. Then we will obtain the basis coefficients matrix by using LMMSE estimation:

$$\widetilde{\mathbf{g}} = \mathbf{R_g}\mathbf{D^H}/(\mathbf{D}\mathbf{R_g}\mathbf{D^H} + \mathbf{R_d} + \mathbf{R_{W^P}}) \tag{32}$$

where $\mathbf{R_g}, \mathbf{R_d}$ and $\mathbf{R_w}$ denote the autocorrelation matrix of coefficients $\mathbf{g}$, inter-carrier interference (ICI) sequences $\mathbf{d}$ and noise matrix $\mathbf{w^P}$ respectively. And they are calculated as following:

$$\mathbf{R_g} = (\mathbf{B}^+\mathbf{E}\{\mathbf{h_{i,l}}\mathbf{h_{j,l}}\}\mathbf{B}^{+\mathbf{H}}) \otimes \frac{\mathbf{E}\{\mathbf{h_{n,i}}\mathbf{h_{n,j}}\}}{\mathbf{E}\{\left|\mathbf{h^{n_r,n_t}_{n,l}}\right|^\mathbf{2}\}} \otimes \mathbf{R_{MIMO}} \tag{33}$$

$$\mathbf{R_d} = \mathbf{E}\{(\mathbf{A^d}\boldsymbol{\Delta}^\mathbf{d}\mathbf{g})(\mathbf{A^d}\boldsymbol{\Delta}^\mathbf{d}\mathbf{g})^\mathbf{H}\} \tag{34}$$

$$\mathbf{R_{W^P}} = \sigma^\mathbf{2}\mathbf{E_{MN_RL_p}} \tag{35}$$

where $\mathbf{R_{MIMO}}$ is a spatial correlation matrix which computed by the Kronecker product between transmit correlation matrix and receive correlation matrix.

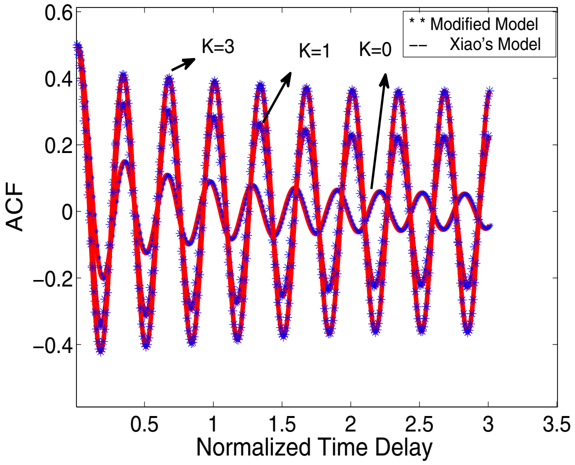

Fig. 3: The ACF of Xiao's model and modified model

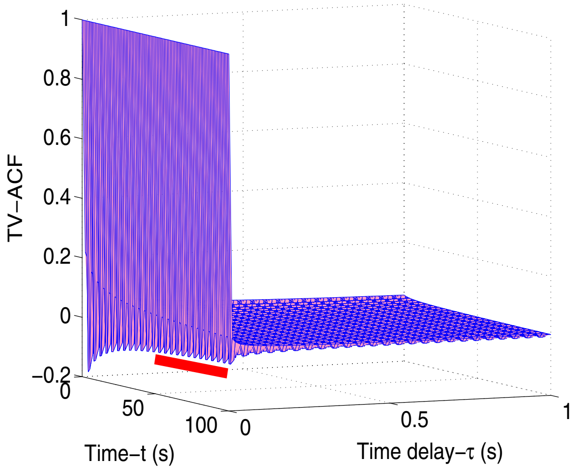

Fig. 4: The TV ACF of acceleration process

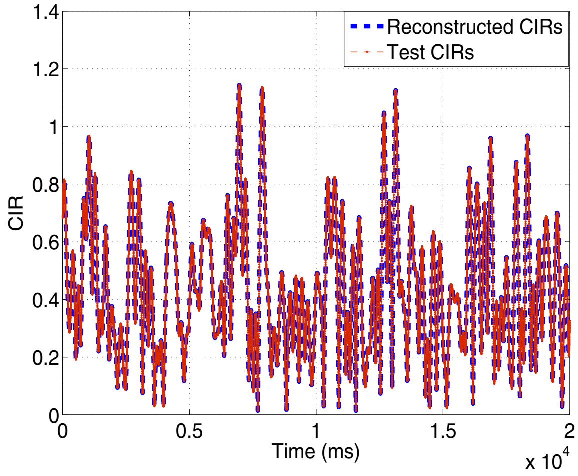

Fig. 5: The fading envelop of test CIRs and reconstructed CIRs with speed at 200 km/h

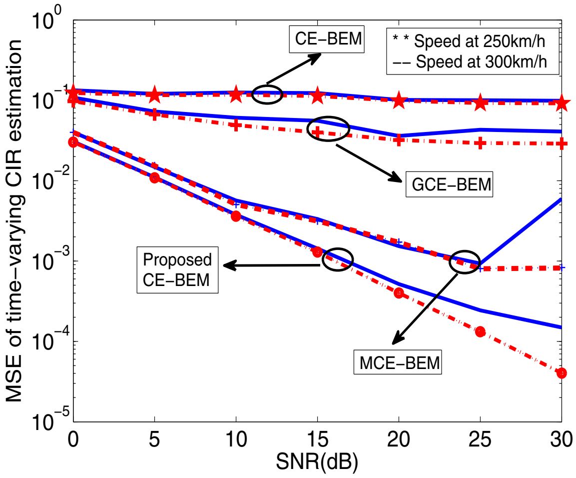

Fig. 6: The MSE of TV CIR estimation for a range of SNRs with different moving speeds

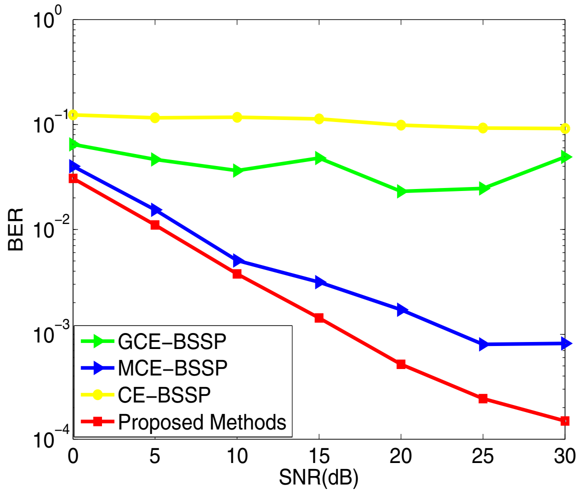

Fig. 7: BER performance versus the SNR for various estimation methods

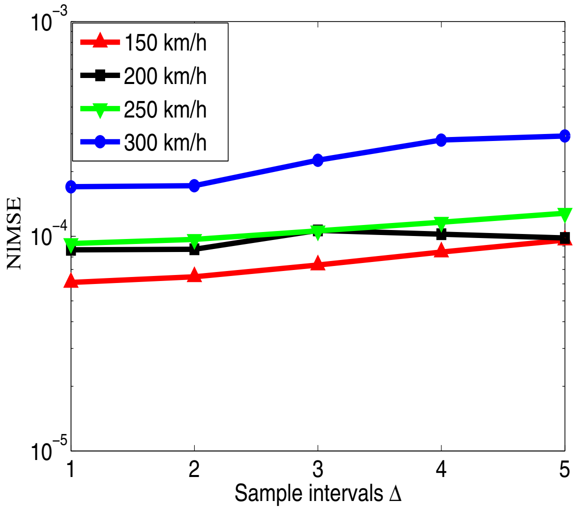

Fig. 8: NMSE performance with different moving speeds and sample intervals

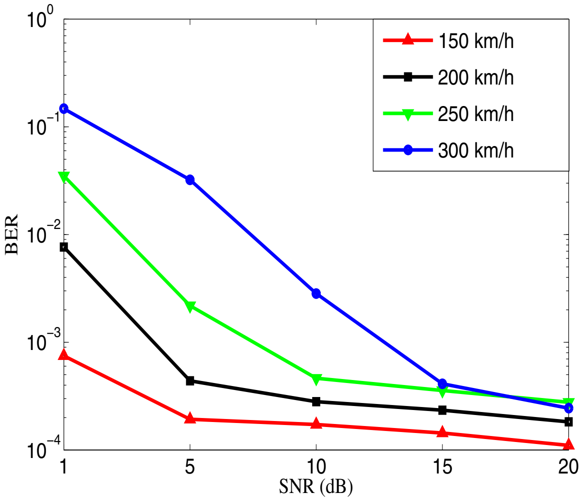

Fig. 9: BER performance with different moving speeds

V. NUMERICAL RESULTS AND ANALYSIS

Firstly, we consider the performance of the proposed TV-HST channel model. The center frequency of the considered communication system is assumed to be 1GHz. After 1000 Monte Carlo simulations have been carried out, Fig. 3 shows the properties of autocorrelation function (ACF) between the proposed model and Xiao's model when the speed of train is 350 km/h. It is not difficult to find that the ACFs of proposed model match the Xiao's model very well under the different values of Rice-K factor. In addition, Fig. 4 describes the TV-ACF of acceleration process when the train is moving with the acceleration process from 200 km/h to 350 km/h. It clearly indicates that the ACF of our proposed model converges to the Xiao's model with a constant speed (350 km/h) when the train is reaching the same speed (as shown on red line in the Fig. 4). It also means our proposed channel model is appropriate for the TV-HST scenarios.

Secondly, we concentrate on the properties of TV channels and BEM. Specially, the functions and performances of our proposed estimation algorithm and BEM can be analysed by reconstructing the test CIR which is generated from the proposed TV-HST channel model. By comparing the fading envelop in Fig. 5, it is manifest that the reconstructed channel has attained an ideal effect for the TV-HST channels when applying our proposed estimation algorithm and BEM.

Finally, we consider a 2×2 MIMO-OFDM system with N=256 subcarriers containing 16 clusters. To approximate the TV-HST channel by a subspace, we set Q=4 and the lengths of tones are assumed as following: L_p=2, L_d=8,L_g=2. In order to develop a convenient and feasible simulation system, the correlation between different sub-channels is not considered.

In Fig. 6, we compare the channel tracking performance of the CE, GCE, MCE and proposed BEM for a range of signal-noise ratios (SNR) with the moving speeds at 250km/h and

300km/h respectively. According to the simulation results, it can be observed that the CE and GCE have a similar modeling error at low speeds but lose track if the channel varies faster. Therefore, it is obvious that the proposed BEM performs better than other BEMs. Fig. 7 plots the bit error ratios (BER) performance versus the SNR for different estimation methods (i.e., CE-BSSP, GCE-BSSP, and MCE-BSSP) with the moving speed at 250 km/h. It is evident that the proposed estimation scheme can attain a lower BER with classic Doppler spectrum in the various SNR. Fig. 8 describes the normalized MSE (NMSE) properties under various moving speeds and sample intervals. It can be seen that the properties of NMSE over the whole range speeds degrade significantly because the increased sample intervals lead to the increased estimation difference. And Fig. 9 depicts the BER properties under various moving speeds. It is quite clear that the BER is less than $10^{-3}$when the speed is under 300 km/h and the SNR is more than 15dB, which meets the basic requirements of communication system in HST.

VI. CONCLUSION

In this paper, a novel channel model for TV-HST scenarios is proposed. The numerical simulation proves the model is better than the others in TV statistical properties. Furthermore, a generalized subspace-based channel estimation algorithm and BEM are also developed for the MIMO-OFDM systems. The simulation results validate that our proposed methods can provide better performance than the traditional ones. Meanwhile, the overall proposed scheme is quite suitable for the simulation of OFDM system in TV-HST scenarios.

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