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Block Pilot Based Channel Estimation and High-Accuracy Signal Detection for GSM-OFDM Systems on High-Speed Railways

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Abstract—In this paper, generalized spatial modulation-orthogonal frequency-division multiplexing is introduced to wireless communication system on high-speed railways for the first time. There are two main challenges to be tackled. On the one hand, channel estimation is difficult as small number of subcarriers are activated; on the other hand, as both the dimensions of the channel matrix and the number of nonzero elements of the unknown signals are large, the matching based and the compressive sensing based signal detectors may not work effectively. To overcome these problems, we first propose a new channel estimation scheme, in which the block pilot pattern instead of the comb pilot pattern is used and a novel interpolation method, which takes the time variation property in different symbols into consideration is adopted. Simulation results demonstrate that in comparison to the conventional interpolation methods, our proposed method achieves better normalized mean square error performance. Then, for the signal detection, we adopt the decomposition and iteration to reduce the dimension of the matrix to be processed. Based on the decomposed structure, the maximum likelihood method is conducted in the solution space to detect the signal. Simulation results demonstrate that the proposed detector presents higher accuracy compared with the existing signal detection schemes.

Index Terms—Generalized spatial modulation (GSM), orthogonal frequency-division multiplexing (OFDM), high-speed railways (HSR), channel estimation, signal detection.

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I. INTRODUCTION

RECENTLY, wireless communication systems on high speed railways (HSR) attracts a lot of research attention due to wide utilization of mobile terminals [1]–[3]. Existing works are concerned about how to improve the spectral efficiency of the wireless communication systems on HSR [3]. As a technology of high spectral efficiency exploited by many wireless communication systems, multiple input multiple output (MIMO)-orthogonal frequency-division multiplexing (OFDM) may be difficult to be used in HSR communication systems. This is because the high mobility leads to a large Doppler spread, which incurs the severe inter carrier interference (ICI). Besides, for conventional MIMO technique, there exists inter antenna interference (IAI). Due to the ICI and IAI, the MIMO-OFDM based wireless communication system on HSR will suffer from a prohibitive detection complexity at the receiver [4]. In addition, due to the unaffordable pilot overhead [5], the practical MIMO-OFDM based wireless communication system on HSR cannot support too many antennas. Thus, the application of MIMO-OFDM in wireless communication system on HSR may result in only limited spectral efficiency improvement.

Spatial modulation (SM)-OFDM [6] is an emerging multiple-antenna technique. Different from the conventional MIMO-OFDM techniques in which one subcarrier is activated in all the antennas, SM-OFDM activates a subcarrier only in one antenna. As a result, the information carried by each SM-OFDM subcarrier is comprised of two parts: the conventional modulated symbol and the information carried by the index of the antenna to which the subcarrier is attached. It can be observed that for SM-OFDM, there exists no IAI as one subcarrier is activated only in one antenna [7]. Thus, it can effectively reduce the detection complexity at the receiver. To further improve the system spectral efficiency, generalized SM (GSM)-OFDM scheme [8] in which several antennas instead of one antenna are activated for each subcarrier was proposed.

For better illustration, we compare the spectral efficiency and the system interference between MIMO-OFDM, SM-OFDM and GSM-OFDM, which is shown in Table I. From Table I, we can see that in comparison to MIMO-OFDM and SM-OFDM, GSM-OFDM offers a better trade-off among the spectral efficiency, cost of antennas and complexity of the receiver by adjusting the number of active antennas corresponding to one subcarrier.

TABLE I
COMPARISON OF THE SPECTRAL EFFICIENCY AND THE SYSTEM INTERFERENCE
BETWEEN SM-OFDM, GSM-OFDM, AND MIMO-OFDM

Technique	Spectral Efficiency (bps/Hz)	Interference
SM-OFDM	$\log_2(M) + \log_2(N_t)$	ICI
GSM-OFDM	$N_K \log_2 M + \log_2 C_{N_t}^{N_K}$	ICI and reduced IAI
MIMO-OFDM	$N_t \log_2(M)$	ICI and IAI

Although GSM-OFDM can reduce IAI of the wireless communication system on HSR, it makes the estimation of the rapidly varying channel a challenging problem due to limited number of activated subcarriers. For channel estimation, the literature [9] provides a review of conventional schemes. In [10] and [11], the spatial and temporal channel correlation is exploited for the channel estimation and differential modulation. However, the proposed schemes haven't considered the effect of ICI in the wireless communication system on HSR. To deal with ICI, the channel estimation schemes in [12]–[16] adopt comb-style pilot pattern, while the pilots of GSM-OFDM system on HSR cannot be placed as the same pilot pattern, since for a transmit receive antenna pair, only a small portion of subcarriers are activated. With block pilot pattern, the works [17]–[19] design channel estimation schemes for time domain sequence-orthogonal frequency division multiplexing (TDS-OFDM) systems, which use the time domain sequence to guarantee the channel estimation accuracy and cannot be applied in the cyclic prefix-orthogonal frequency division multiplexing (CP-OFDM) systems. Besides, for SM systems, the existing channel estimation methods [20], [21] are mainly based on the prediction of the channel coefficient of the next time slot. However, such schemes cannot be applied in wireless communication system on HSR since the ICI coefficients may not be predicted in a straightforward manner.

Beside channel estimation, signal detection of GSM-OFDM system on HSR is another challenging problem. Due to the ICI in high mobile scenario, GSM-OFDM cannot be regarded as N parallel GSM transmission. As far as we know, few works have been done in literature to design signal detection method for GSM-OFDM system on HSR. For conventional minimum mean-squared error (MMSE) and maximum likelihood (ML) based schemes [22], on the one hand, the detection complexity is too high to afford due to the high dimension of the full channel matrix; on the other hand, inaccurate localization of the nonzero elements, which correspond to the active subcarriers, may damage the effectiveness of these schemes. In [23], a compressive sensing based detection method was proposed. However, this scheme suffers from some performance degradation in GSM-OFDM system on HSR because it requires a high signal sparsity. e.g., there exists a large number of zero elements. Considering that the high complexity of conventional ML scheme [24] results from the fact that it traverses all the symbols in the solution space, the authors of [25] proposed a simplified scheme, named hard-limiter ML (HL-ML) detector, which only searches for the nonzero element positions in columns of the channel matrix. This scheme can effectively re-

duce the detection complexity. However, it also sacrifices the detection accuracy. Moreover, some other MMSE based detection schemes, including ordered block minimum mean-squared error (OB-MMSE) [26] and ordered nearest neighbor search minimum mean-squared error (O-NNS-MMSE) [27], are not suitable for the GSM-OFDM system on HSR. The reason is that high dimension of the channel matrix results in inaccurate nonzero elements localization.

Owing to the signal sparsity of GSM-OFDM, CS is an effective detection means for its low complexity. In existing literature, some generic CS algorithms [28], such as orthogonal matching pursuit (OMP), basic pursuit (BP), and compressive sampling matching pursuit (CoSaMP), have been applied for SM detection. However, conventional CS algorithms may be hard to achieve a reasonable performance due to large number of nonzero elements and high coherent measurement matrix in GSM-OFDM system on HSR. In [29], spatial modulation matching pursuit (SMMP) was proposed for multiple access channels, which located the active antennas for each user separately. A normalized CS (NCS) method in [30] was proposed for and could only be applied to the space shift keying (SSK) signal detection. In [22], it was extended to GSM signal detection. However, in GSM-OFDM system on HSR with high dimensional channel matrix, the algorithm in [22] presented high complexity and deteriorative performance.

In this paper, we introduce GSM-OFDM system on HSR due to its balance between receiver complexity and spectral efficiency improvement. For channel estimation in GSM-OFDM system on HSR, the comb-style pilot pattern designed for the conventional DS channel estimation can not be employed due to small number of active subcarriers. As such, we adopt block-style pilot pattern and propose a novel interpolation method to improve estimation accuracy. Specifically, a block is constituted by four OFDM symbols, where the first one is a pilot symbol and the remaining three are data symbols. For the pilot symbol, the estimated channel coefficients are linearly approximated due to linear property in delay domain. As for the first and the third data symbols, we obtain their estimated channel coefficients by linearly extending the channel coefficients of their adjacent pilot symbol. The channel coefficients of the second data symbol is obtained by linear interpolation of the first and the third symbols. Different from the existing interpolation methods, we take the channel variation property into consideration and it is shown that our proposed method can improve the estimation accuracy.

Based on the estimated channel coefficients, we propose a signal detection scheme for GSM-OFDM system on HSR. In our previous work [7], we have designed a pioneering scheme for SM-OFDM system on HSR with perfect channel state information (CSI). However, its straightforward extension in the GSM-OFDM system on HSR with our proposed channel estimation presents degraded performance due to channel error and increased number of nonzero elements. Motivated by this problem, we adopt decomposition and iteration to simplify the signal localization and recovery. On the basis of the decomposed structure, we solve each sub-problem with ML method. Simulation results indicate that our proposed algorithm achieves better

bit error rate (BER) performance in comparison to the existing schemes.

The main content of each section is summarized as follows. In Section II, we provide the basic knowledge of the GSM-OFDM and the model of the wireless communication system on HSR. Section III describes our proposed channel estimation scheme, which includes the pilot pattern design, the channel estimation scheme for the pilot symbol and interpolation method for the data symbol. A novel signal detection algorithm for the GSM-OFDM system on HSR is proposed in Section IV. In Section V, we analyze the complexity of the signal detection scheme and compare it with the MIMO-OFDM system with the same spectral efficiency. Simulation results are provided in Section VI to evaluate the performance of our proposed channel estimation and signal detection methods. The Section VII concludes the paper.

Notations: C_a^b stands for the number of the combinations for us to select b elements from a set with a different elements. $\log_a(b)$ denotes the logarithm b based on a . $\lfloor a \rfloor$ represents the largest integer less than a . The notation $\approx$ means approximately equal and $\sum$ denotes the operation of summing. $H(m, n)$ stands for the element of the matrix $\mathbf{H}$ on the m -th row and the n -th column. $a(n)$ represents the n -th element of the vector $\mathbf{a}$. The notation $\text{diag}(\mathbf{A})$ denotes the diagonal of the matrix $\mathbf{A}$. The superscript $(\cdot)^T$ stands for the operation of transposition. Finally, $\|\cdot\|_2$ stands for the Euclidean norm.

II. BACKGROUND AND SYSTEM MODEL

In this section, we briefly introduce the basic rule of GSM-OFDM and simply describe the wireless communication system on HSR.

A. GSM-OFDM

GSM is becoming an emerging multiple antenna technique because of its high spectral efficiency and low complexity. For a GSM system with N_t transmit antennas, only N_K ($N_K \ll N_t$) antennas are activated in each time slot and the remaining ones keep silent. The information carried by one GSM symbol consists of two parts, e.g., the M -quadrature amplitude modulation (QAM) symbol transmitted by each antenna and the information carried by the indexes of active antennas. For a GSM-OFDM system with M -QAM modulation, as each subcarrier is activated in N_K antennas, a number of $N_K \log_2 M$ bits are carried by one subcarrier. In addition, as there are $C_{N_t}^{N_K}$ active antenna combinations, the indexes of active antennas carry $\lfloor \log_2 C_{N_t}^{N_K} \rfloor$ bits of information. Hence, the total data rate of a GSM-OFDM

TABLE II
MAPPING RULE OF GSM-OFDM

Bits	Antenna Combination
00	Antenna 1 and Antenna 2
01	Antenna 1 and Antenna 3
10	Antenna 1 and Antenna 4
11	Antenna 2 and Antenna 3

system with N subcarriers is given by $T = NN_b$, in which, $N_b = N_K \log_2 M + \lfloor \log_2 C_{N_t}^{N_K} \rfloor$ bps/Hz.

Denote $\mathbf{D} \in \mathbb{Z}^{T \times 1}$ as the incoming bit stream. It is first reshaped to be a matrix $\mathbf{E} \in \mathbb{Z}^{N \times N_b}$ as shown by equation (1) at the bottom of this page. Each row of $\mathbf{E}$ corresponds to information carried by one subcarrier. We divide each row of $\mathbf{E}$ into $N_K + 1$ blocks, where the former N_K blocks, each of which consists of M bits, are mapped as the N_K M -QAM symbols and the last block, which is constituted by $\lfloor \log_2 C_{N_t}^{N_K} \rfloor$ bits is used to select the N_K antennas on which the subcarrier will be activated.

To clearly illustrate the modulation process, we give an example as follows. Assume that the number of the subcarriers is $N = 2$ and quadrature phase shift keying (QPSK) is adopted ($M = 2$ bits). Besides, the number of antennas is assumed to be $N_t = 4$ and the number of the active antennas is set as $N_K = 2$. Thus, there are $2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor} = 4$ active antenna combinations and indexes of each active antenna combination convey 2 bits of information. The mapping rule is presented in Table II and the bit matrix $\mathbf{E}$ is assumed to be

$$\mathbf{E} = \begin{bmatrix} 1 & 1 & 0 & 0 & 1 & 1 \\ 0 & 1 & 1 & 0 & 0 & 1 \\ \underbrace{\hspace{1.5cm}}_{\text{symbol 1}} & \underbrace{\hspace{1.5cm}}_{\text{symbol 2}} & \underbrace{\hspace{1.5cm}}_{\text{index}} \end{bmatrix}.$$

without loss of generality. According to the mapping rule of Table I, it can be seen that for the first subcarrier, Antenna 2 and Antenna 3 are active. The bits '11' and '00' are mapped as QPSK symbols $1 + i$ and $-1 - i$, and then transmitted by Antenna 2 and Antenna 3, respectively. Similarly, the second subcarrier is activated in Antenna 1 and Antenna 3 and two QPSK symbols $-1 + i$ and $1 - i$ are transmitted on these two antennas, respectively. In other words, the signal in frequency domain for the first antenna is $\{0, -1 + 1i\}$ and $\{1 + 1i, 0\}$, $\{-1 - 1i, 1 - 1i\}$, $\{0, 0\}$ for the second, third, fourth antennas. The example is depicted in Fig. 1.

For a GSM-OFDM with N_t transmit antennas and N_r receive antennas, the received signal can be represented as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{e}, \quad (2)$$

$$\mathbf{E} = \begin{bmatrix} E_{1,1} & \cdots & E_{1,\log_2 M} & E_{1,(N_K-1)\log_2 M+1} & \cdots & E_{1,N_K \log_2 M} & E_{1,N_K \log_2 M+1} & \cdots & E_{1,N_b} \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ E_{N,1} & \cdots & E_{N,\log_2 M} & E_{N,(N_K-1)\log_2 M+1} & \cdots & E_{N,N_K \log_2 M} & E_{N,N_K \log_2 M+1} & \cdots & E_{N,N_b} \\ \underbrace{\hspace{2.5cm}}_{\text{symbol 1}} & \underbrace{\hspace{2.5cm}}_{\text{symbol } N_K} & \underbrace{\hspace{2.5cm}}_{\text{index}} \end{bmatrix} \begin{matrix} \text{sub1} \\ \vdots \\ \text{subN} \end{matrix} \quad (1)$$

Fig. 1. The diagram of GSM-OFDM.

in which,

$$\mathbf{y} = [\mathbf{y}_1^T \cdots \mathbf{y}_{N_r}^T]^T,$$

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_{11} & \cdots & \mathbf{H}_{1N_t} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_{N_r 1} & \cdots & \mathbf{H}_{N_r N_t} \end{bmatrix},$$

$$\mathbf{x} = [\mathbf{x}_1^T \cdots \mathbf{x}_{N_t}^T]^T,$$

$$\mathbf{e} = [\mathbf{e}_1^T \cdots \mathbf{e}_{N_r}^T]^T,$$

where $\mathbf{x}_{n_t} \in \mathbb{C}^{N \times 1}$ denotes the transmitted signal of the n_t -th antenna, $n_t \in [1, N_t]$, $\mathbf{e}_{n_r}$ is the noise term, and $\mathbf{y}_{n_r} \in \mathbb{C}^{N \times 1}$ represents the received signal of the n_r -th antenna, $n_r \in [1, N_r]$. The matrix $\mathbf{H}_{n_r n_t} \in \mathbb{C}^{N \times N}$ stands for the channel matrix in frequency domain between the n_r -th receive antenna and the n_t -th transmit antenna,

$$\mathbf{H}_{n_r n_t} = \begin{bmatrix} H_{n_r n_t}(1, 1) & \cdots & H_{n_r n_t}(1, N) \\ \vdots & \ddots & \vdots \\ H_{n_r n_t}(N, 1) & \cdots & H_{n_r n_t}(N, N) \end{bmatrix}. \quad (3)$$

Based on equation (2), we can detect indexes of nonzero elements and the associated QAM symbols of $\mathbf{x}$ jointly. Following this, an inverse mapping process is performed to obtain the information bits.

B. The Wireless Communication System on HSR

A typical wireless communication system on HSR usually contains three elements: the base station, the communication relay nodes upon each carriage and the mobile terminals of the passengers [1], [2]. The communication link consists of two hops, including the hop between the base station and the relay and the hop from the relay to the mobile terminals inside the carriage. We focus on the first hop for its performance deterioration caused by the time varying channel. In wireless communication system on HSR, the high mobility incurs ICI, which results in the full channel matrix in frequency domain, i.e., $\mathbf{H}_{n_t, n_r}$ in (3) is full, $n_t \in [1, N_t]$, $n_r \in [1, N_r]$.

III. PROPOSED CHANNEL ESTIMATION SCHEME FOR GSM-OFDM SYSTEM ON HSR

In existing literature, little has been done on the channel estimation problem of GSM-OFDM system on HSR. The main challenge lays on the facts that only several subcarriers are activated in each time and there is severe ICI because of the high moving speed. For GSM-OFDM system on HSR, the conventional comb-style pilot based DS channel estimation schemes [31] are ineffective since the silent subcarriers cannot be used as pilot subcarriers. In this section, we use the block-style pilot structure and propose a novel interpolation method for GSM-OFDM system on HSR, which lays the foundation for the signal detection.

A. Block Pilot Pattern Based DS Channel Estimation Scheme

In MIMO-OFDM systems, the comb-style pilot pattern, e.g., in each OFDM symbol, pilot subcarriers are inserted among the data subcarriers, is usually exploited for DS channel estimation due to its high accuracy. However, this comb-style pilot pattern cannot be applied in GSM-OFDM systems as there are just several active subcarriers for each transmit receive antenna pair. Hence, we resort to the block-style pilot pattern, even though compared with the comb-style pilot pattern, it is less accurate with the same pilot overhead because of the time variation of the channel. For block-style pilot pattern, one whole OFDM symbol is regarded as the pilot symbol and several data symbols are inserted between two pilot symbols. The channel coefficients of all transmit receive antenna pairs are estimated in the pilot symbols and the channel coefficients for the data symbols are obtained by interpolation.

In the following, we briefly describe the block-style pilot pattern proposed in [32] and the corresponding channel estimation scheme, namely the complex exponential basis expansion model based least square (CE-BEM-LS) estimation [5]. In Fig. 2, the detailed pilot structure is depicted. Since it is regulated by International Telecommunication Union (ITU) that for wireless communication system with multiple antennas, the pilot overhead cannot exceed 25% [33], we assume a pilot overhead of 25% here by considering the difficulty in the channel estimation of GSM-OFDM system on HSR. We treat 4 successive OFDM symbols as a block, in which, the first OFDM symbol is the pilot symbol and the remaining three symbols carry GSM-OFDM data. For the pilot OFDM symbol, G groups are equally spaced on the N subcarriers. Each group accommodates the pilot subcarriers for N_t transmit antennas. The pilot subcarriers for different antennas are orthogonal with each other, e.g., the pilot subcarriers for one antenna are not activated on the remaining antennas, and placed sequentially in accordance with the antenna order. In each group, $R = 2Q + 1$ subcarriers [34] are allocated for each antenna, in which the subcarrier in the middle position carries nonzero pilot and the Q subcarriers on both sides carry zero pilots. The value of Q is jointly determined by the maximum Doppler shift, system bandwidth and the number of subcarriers [31].

Fig. 2. The pilot pattern with $Q = 2$.

Based on this pilot structure (as depicted in Fig. 2), CE-BEM-LS scheme is adopted to perform channel estimation. At the receiver side, the pilot subcarriers are first segregated for each transmit receive antenna pair due to the orthogonality of the pilot positions among different transmit antennas. The channel coefficients at the pilot symbol are approximated by CE-BEM method, which can effectively reduce the number of coefficients to be estimated. Following this, an LS based algorithm is exploited to estimate the channel coefficients of the pilot symbol.

B. Proposed Mixed Interpolation Method

In OFDM system, for block-style pilot pattern, the channel estimation of the data symbol is usually conducted by interpolation. The commonly used interpolation methods include linear interpolation, cubic interpolation, cubic spline interpolation and so on. However, the performances of these methods are unsatisfactory in HSR scenario due to the fast time variation property of the channel [5]. To overcome this problem, the comb-style pilot pattern is usually adopted. Nonetheless, due to the special signal structure, the comb-style pilot cannot be applied in GSM-OFDM systems. Thus, how to improve the channel estimation accuracy of the data symbols of GSM-OFDM system is a challenging work. In this part, based on the block-style pilot, we propose a mixed interpolation method, which achieves better accuracy than the existing schemes and lays the foundation for the signal detection in GSM-OFDM system on HSR.

Due to the fast time variation property of the channel in wireless communication system on HSR, the existing interpolation methods usually present performance deterioration. For better illustration, we take the linear interpolation as an example. Under the high signal-to-noise ratio (SNR), the linear interpolation presents degraded estimation accuracy and an obvious error floor since the effect of the channel variation is not properly considered. To deal with this problem, we design a novel interpolation method, referred as enhanced interpolation in the following. In this enhanced interpolation, three steps are performed sequentially: linearly approximating the estimated coefficients for each pilot symbol, predicting the channel coefficients of the first and third data symbol in each block by using the channel responses of their adjacent pilot symbols and estimating the channel coefficients of the second data symbol by using linear interpolation between the first and third data

symbol. Numerical results show that this enhanced interpolation method achieves obvious performance gain compared with the existing interpolation methods at high SNR region since we take the time variation property of the wireless channel into the consideration and introduce the nearest principle. However, the accuracy of the enhanced interpolation is worse than the linear interpolation in low SNR region due to the effect of the noise. Thus, a mixed interpolation method is proposed by combining the linear interpolation at low SNR and the enhanced interpolation at high SNR.¹ The details of the proposed mixed interpolation method is presented as follows.

In our proposed mixed interpolation algorithm, the SNR of the transmitted signals is first compared with a threshold SNR_{th} . If it satisfies $SNR < SNR_{th}$, the linear interpolation is directly applied to estimate the channel coefficients of the data symbols. Otherwise, Algorithm 1, referred as the enhanced interpolation, is adopted to improve the accuracy of channel estimation. In Algorithm 1, the estimated channel coefficients of the pilot symbols in all blocks and transmit-receive antenna pairs $\mathbf{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)}$, $b \in [1, B]$, $l \in [1, L]$, $n_r \in [1, N_r]$, $n_t \in [1, N_t]$ are used as the input signal. Here, B denotes the number of the blocks, L represents the number of the multiple paths and D stands for the number of the data symbols in each block. Without loss of generality, we assume $B = 2$ and $D = 3$ in this paper. The output signals $\mathbf{h}_{\mathcal{D}_{l,d}}^{(n_r, n_t)}$, $d \in [1, D]$, $l \in [1, L]$, $n_r \in [1, N_r]$, $n_t \in [1, N_t]$ of the enhanced interpolation are the estimated channel coefficients of the data symbols. It can be observed that the Algorithm 1 is mainly constituted by two parts: the linear approximation for the channel coefficients of the pilot symbols and the channel estimation process for the data symbols. In the first part, the estimated channel coefficients of the pilot symbols $\mathbf{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \in \mathbb{C}^{N \times 1}$ of each transmit receive antenna pair and each block are approximated by a straight line. First, the channel coefficient of the $N/4$ -th point is approximated as the mean value of the former $N/2$ points and the channel coefficient of the $3N/4$ -th point as the mean value of the latter $N/2$ points, e.g.,

$$\bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \left(\frac{N}{4} - 1 \right) \approx \frac{2}{N} \left(\sum_{n=0}^{N/2-1} h_{\mathcal{P}_{l,b}}^{(n_r, n_t)}(n) \right), \quad (4)$$

¹The threshold of the SNR can be obtained offline by simulation.

Algorithm 1: The Enhanced Interpolation.

Require:

$$\mathbf{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)}, b \in [1, B], l \in [1, L], n_r \in [1, N_r], \\ n_t \in [1, N_t]$$

Ensure:

$$\mathbf{h}_{\mathcal{P}_{l,d}}^{(n_r, n_t)}, d \in [1, D], l \in [1, L], n_r \in [1, N_r], \\ n_t \in [1, N_t]$$

Part I : Linear approximation for the pilot symbol

```
1: for  $n_r = 1$  to  $N_r$  do
2:   for  $n_t = 1$  to  $N_t$  do
3:     for  $b = 1$  to  $B$  do
4:       for  $l = 1$  to  $L$  do
5:         Get two special points, the  $N/4$ -th and
            $3N/4$ -th points, of  $\mathbf{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)}$  as (4) and (5).
6:         Calculate the slope of the line determined by
           the above two points,  $\beta_{l,b}^{(n_r, n_t)}$ , according
           to (6).
7:         The approximate channel  $\bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)}$  can be
           obtained as (7).
8:       end for
9:     end for
10:   end for
11: end for
Part II : Channel estimation for the data symbols
12: for  $n_r = 1$  to  $N_r$  do
13:   for  $n_t = 1$  to  $N_t$  do
14:     for  $l = 1$  to  $L$  do
15:       The channel coefficients of the first symbol in
         a block is obtained by (9).
16:       The channel estimation for the third symbol in
         a block is conducted as (11).
17:       The second symbol is performed according
         to (12).
18:     end for
19:   end for
20: end for
```

$$\bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \left(\frac{3N}{4} - 1 \right) \approx \frac{2}{N} \left(\sum_{n=N/2}^{N-1} h_{\mathcal{P}_{l,b}}^{(n_r, n_t)}(n) \right). \quad (5)$$

Then it can be seen that the approximated $N/4$ -th and the $3N/4$ -th points determine a straight line, whose slope can be calculated as

$$\beta_{l,b}^{(n_r, n_t)} = \frac{\bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \left(\frac{N}{4} - 1 \right) - \bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \left(\frac{3N}{4} - 1 \right)}{N/2}. \quad (6)$$

Thus, the complete approximate straight line can be expressed as

$$\bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)}(n) = \left(n + 1 - \frac{N}{4} \right) \beta_{l,b}^{(n_r, n_t)} + \bar{h}_{\mathcal{P}_{l,b}}^{(n_r, n_t)} \left(\frac{N}{4} - 1 \right). \quad (7)$$

In the second part, based on the approximate straight line of the pilot symbols, we conduct the channel estimation for the data symbols. Due to the fixed block structure, the channel estimation

for each data symbol in the block can be designed differently. For better illustration, we take block 1 as an example. The channel estimation for data 1, data 2, and data 3 in block 1 are obtained via the estimated channel coefficients of the pilot symbols in block 1 and block 2. In a data symbol, the channel coefficients are supposed to be linearly related to the sampling time. Thus, as to data 1 in block 1, the ending point of the linearly approximate pilot symbol in block 1 is regarded as the starting point of data 1 in block 1, i.e.,

$$h_{\mathcal{P}_{l,1}}^{(n_r, n_t)}(0) = \bar{h}_{\mathcal{P}_{l,1}}^{(n_r, n_t)}(N-1). \quad (8)$$

Besides, the slope of data 1 in block 1 is taken as the same as the slope of the linearly approximate pilot symbol in block 1. Thus, the relationship between the channel coefficients of data 1 in block 1 and the corresponding sampling points can be expressed as

$$h_{\mathcal{P}_{l,1}}^{(n_r, n_t)}(n) = n\beta_{l,1}^{(n_r, n_t)} + \bar{h}_{\mathcal{P}_{l,1}}^{(n_r, n_t)}(N-1). \quad (9)$$

For data 3 in block 1, the starting point of the linearly approximate pilot symbol in block 2 is treated as the ending point of data 3 in block 1, i.e.,

$$h_{\mathcal{P}_{l,3}}^{(n_r, n_t)}(N-1) = \bar{h}_{\mathcal{P}_{l,2}}^{(n_r, n_t)}(0). \quad (10)$$

The slope of data 3 in block 1 is set as the same as the linearly approximate slope of the pilot symbol in block 2. We can obtain the estimated channel coefficients of data 3 in block 1 as

$$h_{\mathcal{P}_{l,3}}^{(n_r, n_t)}(n) = -(N-n)\beta_{l,2}^{(n_r, n_t)} + \bar{h}_{\mathcal{P}_{l,3}}^{(n_r, n_t)}(0). \quad (11)$$

Finally, the estimation for the second data symbol in block 1 is performed by the linear interpolation of data 1 and data 3 in the same block, which is

$$h_{\mathcal{P}_{l,2}}^{(n_r, n_t)}(n) = \left(h_{\mathcal{P}_{l,1}}^{(n_r, n_t)}(n) + h_{\mathcal{P}_{l,3}}^{(n_r, n_t)}(n) \right) / 2. \quad (12)$$

IV. THE PROPOSED SIGNAL DETECTION SCHEME FOR GSM-OFDM SYSTEM ON HSR

Our previous work in [7], a signal detection scheme proposed for the scenario of SM-OFDM system on HSR under the perfect CSI, presents deteriorated BER performance in GSM-OFDM system on HSR due to the imperfect CSI and the increased number of the active subcarriers. In this section, we propose a novel signal detection scheme for GSM-OFDM system on HSR. The proposed scheme overcomes the BER performance deterioration since it adopts ML instead of CS algorithm based on the decomposed structure and diminishes the remaining term by utilizing the iteration process.

In our previous work [7], we have proposed a CS-based signal detection scheme for SM-OFDM system on HSR, which is under the assumption of the perfect CSI. It presents BER performance gain since the decomposed structure satisfies the restricted isometry property (RIP) condition. However, this previous signal detection scheme is ineffective in GSM-OFDM system on HSR under the estimated CSI. The challenges focus on two points: the increased number of the active antennas corresponding to one subcarrier, which results in a less sparse CS model, and channel error remained by channel estimation process. Both of them degrades the detection accuracy.

In this part, we propose a novel signal detection scheme for GSM-OFDM system on HSR under imperfect CSI. To obtain the $\mathbf{x}$ in the formula (2), a direct method is the zero forcing (ZF) algorithm. But the high dimensional full channel matrix results in extremely high complexity. Besides, the channel error leads to unsatisfactory BER performance of detection. Owing to the sparsity of the signal $\mathbf{x}$, CS has been introduced as detector with the assumption of $N_r < N_t$. However, the conventional CS algorithms are ineffective due to the high coherence of the measurement matrix. In our previous CS based detection scheme in [7], a decomposition and iteration process is designed to avoid treating the high coherent channel matrix as measurement matrix. In our proposed detection scheme for GSM-OFDM system on HSR, we remain such decomposition and iteration process to reduce the complexity of the receiver. Based on the decomposed structure, considering large value of the sparsity and channel error, ML is exploited for the sub-problem to improve the detection accuracy.

The decomposition of formula (2) is conducted as

$$\begin{aligned} \mathbf{y} &= \mathbf{H}\mathbf{x} + \mathbf{e} \\ &= (\mathcal{D}(\mathbf{H}) + \mathbf{H} - \mathcal{D}(\mathbf{H}))\mathbf{x} + \mathbf{e} \\ &= \mathcal{D}(\mathbf{H})\mathbf{x} + (\mathbf{H} - \mathcal{D}(\mathbf{H}))\mathbf{x} + \mathbf{e}, \end{aligned} \quad (13)$$

in which,

$$\mathcal{D}(\mathbf{H}) = \begin{bmatrix} \text{diag}(\mathbf{H}_{11}) & \cdots & \text{diag}(\mathbf{H}_{1N_t}) \\ \vdots & \ddots & \vdots \\ \text{diag}(\mathbf{H}_{N_r 1}) & \cdots & \text{diag}(\mathbf{H}_{N_r N_t}) \end{bmatrix}. \quad (14)$$

By adjusting the columns and rows order of $\mathcal{D}(\mathbf{H})$, we obtain a block diagonal matrix $\mathcal{D}(\mathbf{H})_{ad}$ as (23), shown at the bot-

tom of the page. The n -th diagonal elements of all the $\mathbf{H}_{n_r, n_t}$, $n_r \in [1, N_r]$, $n_t \in [1, N_t]$ are selected and arranged as a matrix $\mathbf{S}_n$ with the dimension $N_r \times N_t$, whose element in the n_r -th row and n_t -th column is $H_{n_r, n_t}(n, n)$. The formulated matrix $\mathbf{S}_n$ is placed as the n -th block of $\mathcal{D}(\mathbf{H})_{ad}$, $n \in [1, N]$. We rearrange the model (13) according to the sequence of the subcarrier index as

$$\mathbf{y}_{ad} = \mathcal{D}(\mathbf{H})_{ad}\mathbf{x}_{ad} + [\mathbf{H}_{ad} - \mathcal{D}(\mathbf{H})_{ad}]\mathbf{x}_{ad} + \mathbf{e}, \quad (15)$$

in which $\mathbf{H}_{ad}$ is represented in (24), shown at the bottom of the page,

$$\mathbf{y}_{ad} = [y_1(1), y_2(1), \dots, y_{N_r}(1), \dots, y_1(N), y_2(N), \dots, y_{N_r}(N)]^T \quad (16)$$

and

$$\mathbf{x}_{ad} = [x_1(1), x_2(1), \dots, x_{N_t}(1), \dots, x_1(N), x_2(N), \dots, x_{N_t}(N)]^T. \quad (17)$$

For the problem (15), the value of $\mathbf{x}_{ad}^{(i)}$ is obtained by

$$\mathbf{x}_{ad}^{(i+1)} = \text{SEARCH}_{alg}(\mathbf{y}_{it}^{(i)}, \mathcal{D}(\mathbf{H})_{ad}), \quad (18)$$

and

$$\mathbf{y}_{it}^{(i)} = \mathbf{y}_{ad} - [\mathbf{H}_{ad} - \mathcal{D}(\mathbf{H})_{ad}]\mathbf{x}_{ad}^{(i)}. \quad (19)$$

The iteration terminates when $\|\mathbf{y} - \mathbf{H}\mathbf{x}\|_2$, which is equal to $\|\mathbf{y}_{ad} - \mathbf{H}_{ad}\mathbf{x}_{ad}\|_2$, no longer decreases, $i \in [0, I - 1]$ denotes the index of the iteration and $\mathbf{x}_{ad}^{(0)} = \mathbf{0}$. SEARCH_{alg} means the searching method. For SEARCH_{alg} , assume that $n \in [1, N]$,

$$\mathbf{x}_{ad}^{(i)}[n] = [x_{ad}^{(i)}((n-1)N_t + 1), \dots, x_{ad}^{(i)}((n-1)N_t + N_t)]^T \quad (20)$$

$$\mathcal{D}(\mathbf{H})_{ad} = \begin{bmatrix} \mathbf{H}_{11}(1, 1) & \cdots & \mathbf{H}_{1N_t}(1, 1) \\ \vdots & \ddots & \vdots \\ \mathbf{H}_{N_r 1}(1, 1) & \cdots & \mathbf{H}_{N_r N_t}(1, 1) \\ \underbrace{\hspace{10em}}_{\mathbf{S}_1} & & \\ & \ddots & \\ & & \mathbf{H}_{11}(N, N) & \cdots & \mathbf{H}_{1N_t}(N, N) \\ & & \vdots & \ddots & \vdots \\ & & \mathbf{H}_{N_r 1}(N, N) & \cdots & \mathbf{H}_{N_r N_t}(N, N) \\ & & \underbrace{\hspace{10em}}_{\mathbf{S}_N} & & \end{bmatrix} \quad (23)$$

$$\mathbf{H}_{ad} = \begin{bmatrix} \mathbf{H}_{11}(1, 1) & \cdots & \mathbf{H}_{1N_t}(1, 1) & & \mathbf{H}_{11}(N, 1) & \cdots & \mathbf{H}_{1N_t}(N, 1) \\ \vdots & \ddots & \vdots & \cdots & \vdots & \ddots & \vdots \\ \mathbf{H}_{N_r 1}(1, 1) & \cdots & \mathbf{H}_{N_r N_t}(1, 1) & & \mathbf{H}_{N_r 1}(N, 1) & \cdots & \mathbf{H}_{N_r N_t}(N, 1) \\ & & \vdots & \ddots & & & \vdots \\ \mathbf{H}_{11}(N, 1) & \cdots & \mathbf{H}_{1N_t}(N, 1) & & \mathbf{H}_{11}(N, N) & \cdots & \mathbf{H}_{1N_t}(N, N) \\ \vdots & \ddots & \vdots & \cdots & \vdots & \ddots & \vdots \\ \mathbf{H}_{N_r 1}(N, 1) & \cdots & \mathbf{H}_{N_r N_t}(N, 1) & & \mathbf{H}_{N_r 1}(N, N) & \cdots & \mathbf{H}_{N_r N_t}(N, N) \end{bmatrix} \quad (24)$$

TABLE III
COMPARISON OF THE COMPLEXITY

Algorithm	Complexity	Example
SMMP	$\mathcal{O}(N_t N_r N^2)$	4.7×10^8
OMP	$\mathcal{O}(N_t N_r N^3)$	1.3×10^{10}
Scheme in [7]	$\mathcal{O}(N_t N_r N^2)$	1.6×10^8
Proposed	$\mathcal{O}(INN_t M^{N_K} 2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor})$	2×10^8

and

$$\mathbf{y}_{it}^{(i)}[n] = \left[y_{it}^{(i)}((n-1)N_r + 1), \dots, y_{it}^{(i)}((n-1)N_r + N_r) \right]^T. \quad (21)$$

Thus, $\mathbf{x}_{ad}^{(i)}$ can be regarded as the concatenate $\mathbf{x}_{ad}^{(i)}[n]$ with $n \in [1, N]$ and so as $\mathbf{y}_{it}^{(i)}$ and $\mathbf{y}_{it}^{(i)}[n]$. Moreover, $\mathbf{x}_{ad}^{(i)}[n]$, $n \in [1, N]$, can be acquired by solving the following underdetermined problem

$$\mathbf{y}_{ad}^{(i)}[n] = \mathbf{S}_n \mathbf{x}_{ad}^{(i)}[n]. \quad (22)$$

The $\mathbf{x}_{ad}^{(i)}[n] \in \mathbb{C}^{N_t}$ represents the signal of the n -th subcarrier from the N_t transmit antennas. It is found that $\mathbf{x}_{ad}^{(i)}[n]$ is an N_K sparsity vector. To guarantee the accuracy of the recovery, a search in the position code book and the set of the M -QAM is conducted for the solution $\tilde{\mathbf{x}}_{ad}^{(i)}[n]$ with a smaller $\|\mathbf{y}_{it}^{(i)}[n] - \mathbf{S}_n \tilde{\mathbf{x}}_{ad}^{(i)}[n]\|_2$. The indices of the N_K non-zero elements is denoted as $(n_t^1, \dots, n_t^{N_K})$. The code book for the position consists of $2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor}$ combinations with $n_t^{n_K} \in [1, N_t]$, $n_K \in [1, N_K]$ and $(n_t^1 \neq \dots \neq n_t^{N_K})$, which are determined in the transmitter design. The value of the non-zero element is represented by $a^{n_K} \in \{a_M\}$ with $n_K \in [1, N_K]$, in which, $\{a_M\}$ denotes the symbol set of the M -QAM. Finally, we obtain the $\mathbf{x}_{ad}^{(i)}[n]$ by performing the hard decision process. Thus, the $\mathbf{x}_{ad}^{(i)}$ in (13) is calculated according to (17) and (20). The detailed steps of the detection scheme are presented in Algorithm 2.

In comparison to our previous work in [7], the main difference of the proposed Algorithm 2 lies in that ML instead of CS algorithm is exploited for the subproblem. It combats the effect of large value of the sparsity and channel estimation error and improves the signal detection accuracy.

V. COMPLEXITY

In this section, we compare the complexity of our proposed Algorithm 2 and the typical signal detectors including SMMP [29], OMP [28], and the scheme in [7], as shown in Table III. In Algorithm 2, the number of multiplications for the searching process is $N_t M^{N_K} 2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor}$. Besides, taking account of the N sub-problems and I iterations, the total complexity for Algorithm 2 is $\mathcal{O}(INN_t M^{N_K} 2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor})$. Besides, the complexity of SMMP, OMP, and the scheme in [7] are $\mathcal{O}(N_t N_r N^2)$, $\mathcal{O}(N_t N_r N^3)$, and $\mathcal{O}(INN_t M^{N_K} 2^{\lfloor \log_2 C_{N_t}^{N_K} \rfloor})$.

For better illustration, we set an example as follows. The parameters for GSM-OFDM system are obtained from our simulation as $I = 2$, $N = 512$, $N_t = 12$, $N_K = 2$, and $N_r = 8$ with 16-QAM. The multiplication times for Algorithm 2 is 2×10^8 .

Algorithm 2: The Proposed Signal Detection Algorithm.

Require:

$\mathbf{y}, \mathbf{H}$

Ensure:

$\mathbf{x}$

```

1: Initialization:  $i \leftarrow 0$ ;  $\mathbf{x}_{ad}^{(0)} = \mathbf{0}$ ;
2: while halting criterion false, do
3:   for  $\{n$  traverses each element in  $[1, N]\}$ 
4:     Step I: Estimation of  $\mathbf{x}_{ad}^{(i)}[n]$ 
5:      $\tilde{\mathbf{x}}_{ad}^{(i)}[n] = \mathbf{0}$ 
6:     Step II: Search in the code book and the
       symbol set
7:     for  $\{(n_t^1, \dots, n_t^{N_K})$  traverses the code book $\}$ 
8:       for  $\{(a^1, \dots, a^{N_K})$  traverses the symbol
         set of the  $M$ -QAM $\}$ 
9:          $\tilde{\mathbf{x}}_{ad}^{(i)}[n] \leftarrow \mathbf{0}$ 
10:         $\tilde{\mathbf{x}}_{ad}^{(i)}[n](n_t^1, \dots, n_t^{N_K}) \leftarrow (a^1, \dots, a^{N_K})$ 
11:        if  $\|\mathbf{y}_{it}^{(i)}[n] - \mathbf{S}_n \tilde{\mathbf{x}}_{ad}^{(i)}[n]\|_2 <$ 
           $\|\mathbf{y}_{it}^{(i)}[n] - \mathbf{S}_n \tilde{\mathbf{x}}_{ad}^{(i)}[n]\|_2$ 
12:           $\tilde{\mathbf{x}}_{ad}^{(i)}[n] \leftarrow \tilde{\mathbf{x}}_{ad}^{(i)}[n]$ ;
13:        else
14:           $\tilde{\mathbf{x}}_{ad}^{(i)}[n] \leftarrow \tilde{\mathbf{x}}_{ad}^{(i)}[n]$ ;
15:        end if
16:      end for
17:    end for
18:    Step III: Hard decision of  $\mathbf{x}_{ad}^{(i)}[n]$ 
       $\mathbf{x}_{ad}^{(i)}[n] \leftarrow \text{dec}(\tilde{\mathbf{x}}_{ad}^{(i)}[n])$ 
19:  end for
20:  Obtain  $\mathbf{x}_{ad}^{(i)}$  by (17) and (20).
21:   $i \leftarrow i + 1$ 
22: end while

```

For SMMP, OMP, and scheme in [7], the multiplication times under the same condition are 4.7×10^8 , 1.3×10^{10} and 1.6×10^8 . It can be seen that the complexity of the proposed Algorithm 2 is lower than that of the conventional SMMP and OMP schemes and is affordable in the practical system design.

VI. SIMULATION

In this section, we conduct the simulation to verify the performance of the proposed channel estimation and signal detection schemes. The simulation results present that our proposed channel estimation and signal detection schemes outperform the existing ones.

A. Simulation Parameters

In this part, we introduce the compared channel estimation and signal detection algorithms. Then the parameters of the simulation system are illustrated. For the channel estimation process, considering that only several subcarriers are activated in GSM-OFDM systems, we adopt the block-style pilot pattern

Fig. 3. The channel estimation for the pilot symbol.

and the channel coefficients for the data symbols are obtained by interpolation. Our proposed mixed interpolation scheme is compared with the conventional polynomial interpolation, which is referred to as pchip, spline interpolation, and linear interpolation. Besides, we compare our proposed signal detection algorithm with LSQR [35], OMP, SMMP [29] and the scheme in [7]. Here OMP is the basic CS recovery algorithm, SMMP combines group localization with subspace pursuit (SP) algorithm and our previous scheme is a CS based decomposition and iteration process.

In our simulated GSM-OFDM system, the number of the transmit antennas $N_t = 12$, the number of the activated ones $N_K = 2$ and the number of the receive antennas $N_r = 8$. Besides, $N = 512$ subcarriers consist of an OFDM symbol. The channel model is assumed as Jake's model with 3 channel taps. The frequency of the central carrier is 2GHz and the subcarrier space is 10 KHz. In the pilot symbol, we set the number of the zero subcarriers on one side $Q = 2$ and the number of the groups $G = 10$.

B. Simulation Results

In Fig. 3, we present the channel estimation performance for the pilot symbol under the speed of 300 km/h and 500 km/h, which is obtained by the conventional CE-BEM-LS scheme [5]. It can be seen that the normalized mean square error (NMSE) for 500 km/h achieves -27 dB and the NMSE for 300 km/h is around -30 dB under high SNR. The higher speed decreases the accuracy of the channel estimation since the larger Doppler spread results in the larger ICI. Besides, referring to [33], we give the Cramer-Rao Lower Bound (CRLB) for the channel estimation of the pilot symbol. It can be seen that the estimation performance of the pilot symbol is close to the CRLB.

In Fig. 4–5, we compare the channel estimation performance for the data symbols with different interpolation methods under the speed of 300 km/h and 500 km/h. In Fig. 4, the performance of all the conventional interpolation methods, including pchip,

Fig. 4. Channel estimation for the data symbol with different interpolation methods at 300 km/h.

Fig. 5. Channel estimation for the data symbol with different interpolation methods at 500 km/h.

spline, and linear interpolation, are almost the same and overlapped with each other. Besides, their NMSE performance is not satisfactory and presents a floor under the high SNR resulting from the time variation property of the channel. It can be found that the conventional interpolation methods outperform the proposed enhanced interpolation method under 25 dB, which is considered as SNR_{th} . The enhanced interpolation method overcomes the NMSE floor and presents the performance gain under $SNR > 25$ dB. The channel error is mainly incurred by the noise under $SNR < SNR_{th}$ while it is mainly caused by the time variation under $SNR > SNR_{th}$. Under the condition of strong noise, the effect on the channel estimation

Fig. 6. BER performance with different interpolation methods and the proposed detector under QPSK at the speed of 500 km/h.

Fig. 7. BER performance with the mixed channel estimation scheme and different schemes for signal detection under QPSK and the speed of 300 km/h.

accuracy of the channel variation is weaker than the noise. The proposed enhanced interpolation method is designed according to the channel variation property. Thus, the performance gain of the enhanced interpolation appears under $\text{SNR} > \text{SNR}_{th}$. Since the value of the SNR_{th} is affected by the mobile speed and can be obtained by experiment offline, the proposed mixed interpolation method selects the linear interpolation method under $\text{SNR} < 25$ dB and the enhanced interpolation method under $\text{SNR} > 25$ dB. In Fig. 5, under the mobile speed of 500 km/h, the conventional interpolation methods also present almost the same performance and the SNR_{th} is 17 dB. In Fig. 6, taking advantage of the proposed Algorithm 2, we compare the BER between different interpolation methods with QPSK modulation under the speed of 500 km/h. We can see that the proposed

Fig. 8. BER performance with the mixed channel estimation scheme and different schemes for signal detection under QPSK and the speed of 500 km/h.

Fig. 9. BER performance with the mixed channel estimation scheme and different schemes for signal detection under 16QAM and the speed of 300 km/h.

mixed interpolation method presents the highest detection accuracy due to its outstanding channel estimation performance. Fig. 6 reflects the contribution of the proposed mixed interpolation method to the system BER.

In Fig. 7–Fig. 9, our proposed signal detection scheme, Algorithm 2, is compared with the existing schemes LSQR, OMP, SMMP, and the scheme in [7] with QPSK and 16QAM under the speed of 300 km/h and 500 km/h. Fig. 7 presents the BER performance with QPSK under 300 km/h, which is based on the estimated channel coefficients by the mixed interpolation. We can see that all the LSQR, OMP and SMMP are ineffective in our simulation. The LSQR presents degraded performance due to highly underdetermined matrix. The performance deterioration of OMP and SMMP results from large number of nonzero elements and high coherent measurement matrix. For the CS

based scheme in [7], the BER performance presents an obvious floor over high SNR due to large number of nonzero elements and channel error. In comparison to all the above methods, our proposed Algorithm 2 overcomes the floor and presents a satisfactory BER performance which results from the decomposition and iteration process and ML for the subproblem. In Fig. 8, several signal detection schemes are simulated under QPSK with the speed of 500 km/h. The BER performance of our proposed Algorithm 2 is a little worse than Fig. 7 since the higher speed leads to the larger channel error. Fig. 9 presents the BER performance under 16QAM with the speed of 300 km/h. For our proposed Algorithm 2, the BER performance is also worse than the Fig. 7 due to the higher dimensional modulation.

VII. CONCLUSIONS

In this paper, we introduce GSM-OFDM to wireless communication system on HSR, aiming at the balance between the receiver complexity and spectral efficiency improvement. For the channel estimation, we adopt block-style pilot pattern to suit the small number of activated subcarriers for each transmit receive antenna pair. Based on block-style pilot pattern, a novel interpolation method is proposed to improve the accuracy of the estimation, which considers the property of the channel variation. Next, a high accuracy signal detection algorithm is designed, performing a decomposition and iteration process to avoid the high dimensional channel matrix and then searching in the solution space to obtain the transmitted signal. Simulation results present that our proposed channel estimation scheme outperforms the algorithms in the existing literature. Besides, the proposed detector presents higher accuracy than the existing signal detection schemes.

REFERENCES

- [1] X. Ren, W. Chen, and M. Tao, "Position-based compressed channel estimation and pilot design for high-mobility OFDM systems," *IEEE Trans. Veh. Technol.*, vol. 64, no. 5, pp. 1918–1929, May 2015.
- [2] X. Ren, M. Tao, and W. Chen, "Compressed channel estimation with position-based ICI elimination for high-mobility SIMO-OFDM systems," *IEEE Trans. Veh. Technol.*, vol. 65, no. 8, pp. 6204–6216, Aug. 2016.
- [3] X. Ren, W. Chen, B. Gong, Q. Qin, and L. Gui, "Position-based interference elimination for high-mobility OFDM channel estimation in multi-cell systems," *IEEE Trans. Veh. Technol.*, vol. 66, no. 9, pp. 7986–8000, Sep. 2017.
- [4] S. Li, J. Xiong, P. Cheng, L. Gui, and Y. Xu, "Low-complexity ICI cancellation based on BEM for OFDM systems over doubly selective channels," *IEICE Trans. Commun.*, vol. E96.B, no. 6, pp. 1588–1596, Jun. 2013.
- [5] Z. Tang and G. Leus, "Identifying time-varying channels with aid of pilots for MIMO-OFDM," *EURASIP J. Adv. Signal Process.*, vol. 2011, no. 1, pp. 1–19, Sep. 2011.
- [6] R. Mesleh, H. Haas, C. W. Ahn, and S. Yun, "Spatial modulation-OFDM," *Proc. 11th Int. OFDM Workshop*, Aug. 2006, pp. 288–292.
- [7] B. Gong, L. Gui, Q. Qin, and X. Ren, "Compressive sensing-based detector design for SM-OFDM massive MIMO high speed train systems," *IEEE Trans. Broadcast.*, vol. 63, no. 4, pp. 714–726, Dec. 2017.
- [8] M. D. Renzo, H. Haas, A. Ghayeb, S. Sugiura, and L. Hanzo, "Spatial modulation for generalized MIMO: Challenges, opportunities, and implementation," *Proc. IEEE*, vol. 102, no. 1, pp. 56–103, Jan. 2014.
- [9] G. Wang, Q. Liu, R. He, F. Gao, and C. Tellambura, "Acquisition of channel state information in heterogeneous cloud radio access networks: Challenges and research directions," *IEEE Wireless Commun.*, vol. 22, no. 3, pp. 100–107, Jun. 2015.
- [10] Z. Zhang, C. Jiao, C. Zhong, H. Zhang, and Y. Zhang, "Differential modulation exploiting the spatial-temporal correlation of wireless channels with moving antenna array," *IEEE Trans. Commun.*, vol. 63, no. 12, pp. 4990–5001, Dec. 2015.
- [11] T. Li, X. Wang, P. Fan, and T. Riihonen, "Position-aided large-scale MIMO channel estimation for high-speed railway communication systems," *IEEE Trans. Veh. Technol.*, vol. 66, no. 10, pp. 8964–8978, Oct. 2017.
- [12] W. Ding, F. Yang, S. Liu, X. Wang, and J. Song, "Nonorthogonal time frequency training sequence based CSI acquisition for MIMO systems," *IEEE Trans. Veh. Technol.*, vol. 65, no. 7, pp. 5714–5719, Jul. 2016.
- [13] X. Ma, F. Yang, S. Liu, W. Ding, and J. Song, "Structured compressive sensing based channel estimation for time frequency training OFDM systems over doubly selective channel," *IEEE Wireless Commun. Lett.*, vol. 6, no. 2, pp. 266–269, Apr. 2017.
- [14] W. Ding, F. Yang, C. Pan, L. Dai, and J. Song, "Compressive sensing based channel estimation for OFDM systems under long delay channels," *IEEE Trans. Broadcast.*, vol. 60, no. 2, pp. 313–321, Jun. 2014.
- [15] P. Cheng *et al.*, "Channel estimation for OFDM systems over doubly selective channels: A distributed compressive sensing based approach," *IEEE Trans. Commun.*, vol. 61, no. 10, pp. 4173–4185, Oct. 2013.
- [16] X. Ma, F. Yang, S. Liu, J. Song, and Z. Han, "Sparse channel estimation for MIMO-OFDM systems in high-mobility situations," *IEEE Trans. Veh. Technol.*, vol. 67, no. 7, pp. 6113–6124, Jul. 2018.
- [17] R. Wang, J. Cai, X. Yu, and S. Duan, "Compressive channel estimation for universal filtered multi-carrier system in high-speed scenarios," *IET Commun.*, vol. 11, no. 15, pp. 2274–2281, 2017.
- [18] W. Hou and C. W. Lim, "Structured compressive channel estimation for large-scale MISO-OFDM systems," *IEEE Commun. Lett.*, vol. 18, no. 5, pp. 765–768, May 2014.
- [19] X. Ma, F. Yang, S. Liu, J. Song, and Z. Han, "Design and optimization on training sequence for mmWave communications: A new approach for sparse channel estimation in massive MIMO," *IEEE J. Sel. Areas Commun.*, vol. 35, no. 7, pp. 1486–1497, Jul. 2017.
- [20] Y. Acar, H. Dogan, and E. Panayirci, "On channel estimation for spatial modulated systems over time-varying channels," *Digit. Signal Process.*, vol. 37, pp. 43–52, Feb. 2015.
- [21] X. Wu, H. Claussen, M. D. Renzo, and H. Haas, "Channel estimation for spatial modulation," *IEEE Trans. Commun.*, vol. 62, no. 12, pp. 4362–4372, Dec. 2014.
- [22] L. Xiao *et al.*, "Efficient compressed sensing detectors for generalized spatial modulation systems," *IEEE Trans. Veh. Technol.*, vol. 66, no. 2, pp. 1284–1298, Feb. 2016.
- [23] Y. C. Eldar and G. Kutyniok, *Compressed Sensing: Theory and Applications*. Cambridge, U.K.: Cambridge Univ. Press, 2012.
- [24] J. Jeganathan, A. Ghayeb, and L. Szczecinski, "Spatial modulation: Optimal detection and performance analysis," *IEEE Commun. Lett.*, vol. 12, no. 8, pp. 545–547, Aug. 2008.
- [25] R. Rajashekar, K. V. S. Hari, and L. Hanzo, "Reduced-complexity ML detection and capacity-optimized training for spatial modulation systems," *IEEE Trans. Commun.*, vol. 62, no. 1, pp. 112–125, Jan. 2014.
- [26] Y. Xiao, Z. Yang, L. Dan, P. Yang, L. Yin, and W. Xiang, "Low-complexity signal detection for generalized spatial modulation," *IEEE Commun. Lett.*, vol. 18, no. 3, pp. 403–406, Mar. 2014.
- [27] J. Zheng, "Low-complexity detector for spatial modulation multiple access channels with a large number of receive antennas," *IEEE Commun. Lett.*, vol. 18, no. 11, pp. 2055–2058, Nov. 2014.
- [28] M. Duarte and Y. Eldar, "Structured compressed sensing: From theory to applications," *IEEE Trans. Signal Process.*, vol. 59, no. 9, pp. 4053–4085, Sep. 2011.
- [29] A. Garcia-Rodriguez and C. Masouros, "Low-complexity compressive sensing detection for spatial modulation in large-scale multiple access channels," *IEEE Trans. Commun.*, vol. 63, no. 7, pp. 2565–2579, Jul. 2015.
- [30] C. M. Yu *et al.*, "Compressed sensing detector design for space shift keying in MIMO systems," *IEEE Commun. Lett.*, vol. 16, no. 10, pp. 1556–1559, Oct. 2012.
- [31] Z. Tang, R. C. Cannizzaro, G. Leus, and P. Banelli, "Pilot-assisted time-varying channel estimation for OFDM systems," *IEEE Trans. Signal Process.*, vol. 55, no. 5, pp. 2226–2238, May 2007.
- [32] Y. Chen, H. Shao, W. Huang, and F. Yang, "Channel estimation based on block-type pilot in OFDM systems over doubly-selective channels," in *Proc. Int. Conf. Comput. Problem-Solving*, Oct. 2011, pp. 409–413.
- [33] L. Dai, Z. Wang, and Z. Yang, "Spectrally efficient time-frequency training OFDM for mobile large-scale MIMO systems," *IEEE J. Sel. Areas Commun.*, vol. 31, no. 2, pp. 251–263, Feb. 2013.

- [34] 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.
- [35] T. Hrycak, S. Das, G. Matz, and H. G. Feichtinger, "Low complexity equalization for doubly selective channels modeled by a basis expansion," *IEEE Trans. Signal Process.*, vol. 58, no. 11, pp. 5706–5719, Nov. 2010.

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