Pilot Symbol Assisted Channel Estimation for 4x4 Space Time Block Coded Spatial Modulation Systems

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***Abstract*— Recently, spatial modulation (SM) and space-time block coding (STBC) are combined to take advantage of the benefits of both while avoiding their drawbacks for multiple-input and multiple-output (MIMO) systems. The pioneering works on STBC-SM assume that perfect knowledge of the channel fading coefficients is available at the receiver. This work[1] addresses the challenging and timely problem of channel estimation for 4×4 STBC-SM systems in the presence of time-varying channels. In this paper, the estimation of channel at pilot durations is done by least square (LS) method and then the channel interpolation is performed by linear interpolation or nearest neighbor interpolation algorithms. Simulation results have demonstrated that the proposed channel estimation based on the linear interpolation offer substantial performance gains over the channel estimation based on the nearest neighbor interpolation. In particular, a savings of about $5dB$ is obtained at $BER = 10^{-5}$, as compared with the nearest interpolation based receiver at $120km/h$ for 4×4 STBC-SM systems with the binary phase shift keying (BPSK) modulation.**

Keywords*— Spatial Modulation; Space-Time Block Coding; Pilot Symbol Assisted Channel Estimation; Multiple-Input Multiple-Output; Time Varying Channels.

I. INTRODUCTION

The inter channel interference (ICI) and inter antenna synchronization (IAS) are traditional problems associated with practical multiple-input and multiple-output (MIMO) systems [1], [2]. Spatial modulation (SM) was proposed by Mesleh et al. [3] to overcome these problems associated with the conventional MIMO transmission schemes. The basic principle of the SM is to use the indices of transmit antennas to convey information in addition to the two-dimensional signal constellations. Hence, it adds a third dimension to the two-dimensional signal space to obtain three-dimensional signal space.

Space-time block code (STBC) is proposed by Alamouti [4] to increase the coding gain or diversity gain for MIMO systems. Recently, a new MIMO transmission scheme, called STBC-SM, is proposed by Basar et al. [5]. STBC-SM takes advantage of the benefits of STBC and SM while avoiding their drawbacks for MIMO systems. In the SM system, only one transmit antenna is active during each transmission interval, whereas STBC-SM uses the indices of the two transmit antennas employed for the transmission of the Alamouti-STBC. Therefore the transmitted information symbols are included not only to the space and time domains but also to the spatial (antenna) domain in STBC-SM scheme. In this manner both antenna indices and STBC carry information. It was shown that the STBC-SM scheme has significant performance advantages over the SM systems. It also provides high spectral efficiency with the low computational complexity due to low-complexity maximum likelihood (ML) decoder which profits from the orthogonality of the Alamouti code [5].

In literature, the pioneering works have assumed that STBC-SM has perfect channel state information (P-CSI) at the receiver. However, reliable coherent STBC-SM communication requires accurate estimation of the fading channel. It is clear that the quasi-static fading assumption is not reasonable for time-varying channels for spatial modulated systems [6]. One of the main challenges faced by high mobility communications is the fast time-varying fading caused by the Doppler effect [7]. Therefore, a very popular and widely accepted method to deal with the system design for fast time-varying fading is pilot-symbol-aided channel estimation (PSA-CE) with using interpolation [8]. In this paper, we assumed that channel changes rapidly for consecutive time slots and we propose PSA-CE for STBC-SM systems in such situations. Therefore we investigate nearest and linear interpolation techniques in PSA-CE for 4×4 STBC-SM systems. Finally, the PSA-CE methods have demonstrated their simplicity, effectiveness as well as efficiency.

Notation: Throughout the paper, bold and capital letters '**A**' denote matrices and bold and small letters '**a**' denote vectors. The notations, $(.)^*$, $(.)^T$, $(.)^\dagger$ and $\|.\|_F$ denote conjugate, transpose, Hermitian and Frobenius norm, of a matrix or a vector respectively.

II. 4×4 SPACE-TIME BLOCK CODED SPATIAL MODULATION (4×4 STBC-SM)

Let us consider an Alamouti 4×4 STBC-SM system. In Alamouti's STBC, two complex information symbols (x_1

[1]This work is supported in part by the Turkish Scientific and Technical Research Institute (TUBITAK) under Grant 114E011 and the Research Fund of Istanbul University under Projects Numbers: 44182 and 45861. This work is also supported by the 7th Frame European WIMAGIC Project.

TABLE I: Alamouti STBC-SM scheme's parameters with four transmit antennas and different modulation type

Total Pairs	BPSK	4-QAM
$l=0$	$(c_1,c_2)=(1,2),\theta_c=0$	$(c_1,c_2)=(1,2),\theta_c=0$
$l=1$	$(c_1,c_2)=(3,4),\theta_c=0$	$(c_1,c_2)=(3,4),\theta_c=0$
$l=2$	$(c_1,c_2)=(1,4),\theta_c=0$	$(c_1,c_2)=(1,4),\theta_c=0.61$
$l=3$	$(c_1,c_2)=(2,3),\theta_c=0$	$(c_1,c_2)=(2,3),\theta_c=0.61$

and x_2) drawn from an M-PSK or M-QAM constellation are transmitted from two transmit antennas in two symbol intervals in an orthogonal manner by the codeword

$$\mathbf{X}=(\mathbf{x}_1 \quad \mathbf{x}_2)=\begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix} \tag{1}$$

where columns and rows correspond to the transmit antennas and the symbol intervals, respectively.

For the STBC-SM scheme we extend the matrix in (1) to the antenna domain consisting of four transmit antennas in such a way that both STBC symbols and the indices of the transmit antennas from which these symbols are transmitted, carry information, as follows. Fist a pair of transmit antennas $c=(c_1,c_2)\in\{1,2,3,4\},(c_1\neq c_2)$ with a label $\ell\in\{1,2,..\}$ is chosen. Note that the total number of possible pairs, C, in such configuration is $C=6$ and a maximum 2 bits per label can be transmitted by selecting any of the 4 possible antenna-pair combinations. As to which combinations should be chosen is an another design problem and we do not consider it here. In our work,on the other hand, the four pairs and their labels are chosen according to the configuration considered in [5], as follows:

$$\begin{aligned} \ell=1 &\quad\Leftrightarrow\quad c=(1,2),\\ \ell=2 &\quad\Leftrightarrow\quad c=(3,4),\\ \ell=3 &\quad\Leftrightarrow\quad c=(1,4),\\ \ell=4 &\quad\Leftrightarrow\quad c=(2,3). \end{aligned}$$

Consequently, the following four codewords can be generated by the suggested 4×4 Alamouti STBC-SM scheme, arranged in two groups:

$$\begin{aligned} \{\mathbf{X}_{11},\mathbf{X}_{12}\} &= \left\{\begin{pmatrix} x_1 & x_2 & 0 & 0 \\ -x_2^* & x_1^* & 0 & 0 \end{pmatrix},\begin{pmatrix} 0 & 0 & x_1 & x_2 \\ 0 & 0 & -x_2^* & x_1^* \end{pmatrix}\right\} \\ \{\mathbf{X}_{21},\mathbf{X}_{22}\} &= \left\{\begin{pmatrix} 0 & x_1 & x_2 & 0 \\ 0 & -x_2^* & x_1^* & 0 \end{pmatrix},\begin{pmatrix} x_2 & 0 & 0 & x_1 \\ x_1^* & 0 & 0 & -x_2^* \end{pmatrix}\right\} e^{j\theta} \end{aligned} \tag{2}$$

where the two STBC-SM codewords $\mathbf{X}_{ij}, j=1,2$ in each groups above do not interfere to each other, satisfying $\mathbf{X}_{ij}\mathbf{X}_{ik}^H=\mathbf{0}_{2\times 2}, j\neq k$; that is they have no overlapping columns. In (2), θ is a rotation angle to be optimized for a given modulation format to ensure maximum diversity and coding gain at the expense of expansion of the signal constellation. In Table 1, the parameters $\ell, c=(c_1,c_2)$ and the optimal rotation angles θ are given for the codewords of (2) and for binary phase-shift keying (BPSK) and quadrature phase shift keying (QPSK) modulation formats.

We now formulate the ML decoder for the STBC-SM scheme. Due to the orthogonality of Alamouti's STBC scheme, we can expressed the received signals obtained at the output of each receive antenna $r=1,2,3,4$ in two consecutive discrete-time intervals n and $n+1$ as follows.

$$\begin{aligned} y_r(n) &= h_{r,c_1}(n)\psi_c x_1(n)+h_{r,c_2}(n)\psi_c x_2(n)+w_r(n)\\ y_r(n+1) &= -h_{r,c_1}(n+1)\psi_c x_2^*(n)+h_{r,c_2}(n+1)\psi_c x_1^*(n)\\ &\quad +w_r(n+1), \end{aligned} \tag{3}$$

where $\psi_c \triangleq e^{j\theta_c}$ and $w_r(n)$ is complex-valued, zero-mean additive white Gaussian noise (AWGN) with variance σ_w^2.

Under the assumption that the varying Rayleigh distributed channel coefficients between jth transmitter antenna and rth receiver antenna do not change along these intervals, that is $h_{r,j}(n+1)\approx h_{r,j}(n)$, the following equivalent observation model can be obtained from (3) and for $\ell\Leftrightarrow c=(c_1,c_2)$ in a vector form:

$$\mathbf{y}(n)=\mathrm{diag}(\boldsymbol{\psi}_l)\mathbf{H}_\ell\begin{bmatrix} x_1(n)\\ x_2(n)\end{bmatrix}+\mathbf{w}(n) \tag{4}$$

where $\mathbf{H}_\ell(\mathbf{n}) = [\boldsymbol{h}^{(1)}{}_\ell(n) \quad \boldsymbol{h}_\ell^{(2)}(n)]$. and $\boldsymbol{\psi}_l = [\psi_c,\psi_c^*,\psi_c,\psi_c^*,\psi_c,\psi_c^*,\psi_c,\psi_c^*]^T$

$$\boldsymbol{h}_\ell^{(1)}(n)=\begin{bmatrix} h_{1,c_1}(n)\\ h_{1,c_2}^*(n)\\ h_{2,c_1}(n)\\ h_{2,c_2}^*(n)\\ h_{3,c_1}(n)\\ h_{3,c_2}^*(n)\\ h_{4,c_1}(n)\\ h_{4,c_2}^*(n)\end{bmatrix},\quad \boldsymbol{h}_\ell^{(2)}(n)=\begin{bmatrix} h_{1,c_2}(n)\\ -h_{1,c_1}^*(n)\\ h_{2,c_2}(n)\\ -h_{2,c_1}^*(n)\\ h_{3,c_2}(n)\\ -h_{3,c_1}^*(n)\\ h_{4,c_2}(n)\\ -h_{4,c_1}^*(n)\end{bmatrix}.$$

Generally, we have C equivalent channel matrices $\mathbf{H}_\ell, \mathbf{0}\leq\ell\leq\mathbf{C-1}$, and for the ℓth combination, the receiver determines the ML estimates of x_1 and x_2 resulting from the orthogonality of $\boldsymbol{h}_\ell^{(1)}(n)$ and $\boldsymbol{h}_\ell^{(2)}(n)$:

$$\begin{aligned} \hat{x}_{1,\ell}(n) &= \arg\min_{x_1}\left\|\mathbf{y}(n)-\mathrm{diag}(\boldsymbol{\psi}_l)\boldsymbol{h}_\ell^{(1)}(n)\ x_1\right\|^2\\ \hat{x}_{2,\ell}(n) &= \arg\min_{x_2}\left\|\mathbf{y}(n)-\mathrm{diag}(\boldsymbol{\psi}_l)\boldsymbol{h}_\ell^{(2)}(n)\ x_2\right\|^2 \end{aligned} \tag{5}$$

Associated minimum ML metrics $m_{1,\ell}(n)$ and $m_{2,\ell}(n)$ for x_1 and x_2 are

$$\begin{aligned} m_{1,\ell}(n) &= \min_{x_1}\left\|\mathbf{y}(n)-\mathrm{diag}(\boldsymbol{\psi}_l)\boldsymbol{h}_\ell^{(1)}(n)\ x_1\right\|^2\\ m_{2,\ell}(n) &= \min_{x_2}\left\|\mathbf{y}(n)-\mathrm{diag}(\boldsymbol{\psi}_l)\boldsymbol{h}_\ell^{(2)}(n)\ x_2\right\|^2, \end{aligned} \tag{6}$$

respectively. Since $m_{1,\ell}(n)$ and $m_{2,\ell}(n)$ are calculated by the ML decoder for the ℓth combination, their summation $m_\ell(n)=m_{1,\ell}(n)+m_{2,\ell}(n), 0\leq\ell\leq C-1$ gives the total ML metric for the ℓth combination. Finally, the receiver makes a decision by choosing the minimum antenna combination metric as $\hat{\ell}(n)=\arg\min_{\ell} m_\ell(n)$ for which $(\widehat{x}_1(n),\widehat{x}_2(n))=(\widehat{x}_{1,\widehat{\ell}}(n),\widehat{x}_{2,\widehat{\ell}}(n))$. As a result, the total number of ML metric calculation is $2CM$, yielding a linear decoding complexity.

III. CHANNEL ESTIMATION FOR THE 4×4 STBC-SM SYSTEM

In the 4×4 STBC-SM system, the channel state information (CSI) is needed to detect modulated symbols,

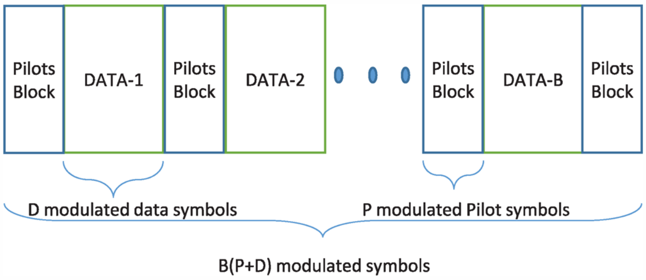

Fig. 1: Frame structure

Discrete Time / Antenna	n_p	n_p+1	n_p+2	n_p+3
Antenna #1	p_1	$-p_2^*$		
Antenna #2	p_2	p_1^*		
Antenna #3			p_3	$-p_4^*$
Antenna #4			p_4	p_3^*

Fig. 2: Pilots block structure, $p_i, i = 1, 2, 3, 4$ denotes pilot symbols

$(\widehat{x}_1(n), \widehat{x}_2(n))$, and transmit antenna pair number, $\widehat{l}(n)$. In practice, wireless systems encounter with the time-varying channel caused by the mobility. It is clear that the performance of conventional channel estimation based on quasi-static channel assumption will be degraded over the time-varying channels. Therefore, PSA-CE with interpolation is used for the proposed receiver structure under high-speed mobile communication environments. Fig. 1 shows the frame structure for the 4×4 STBC-SM systems. Pilot symbols are transmitted as shown in Fig. 2. In this work, the interpolator of proposed channel estimation method is obtained by the linear or nearest interpolation methods. Assuming that fading is constant across two consecutive symbols, we can write the observation model as follows:

$$\underbrace{\begin{bmatrix} y_1(n) \\ y_2(n) \\ y_3(n) \\ y_4(n) \\ y_1(n+1) \\ y_2(n+1) \\ y_3(n+1) \\ y_4(n+1) \end{bmatrix}}_{\mathcal{Y}(n)} = \underbrace{\begin{bmatrix} \mathcal{X}_1 & \mathcal{X}_2 \\ -\mathcal{X}_2^* & \mathcal{X}_1^* \end{bmatrix}}_{\mathcal{X}} \underbrace{\begin{bmatrix} h_{1,c_1}(n) \\ h_{2,c_1}(n) \\ h_{3,c_1}(n) \\ h_{4,c_1}(n) \\ h_{1,c_2}(n) \\ h_{2,c_2}(n) \\ h_{3,c_2}(n) \\ h_{4,c_2}(n) \end{bmatrix}}_{\mathcal{H}(n)} + \underbrace{\begin{bmatrix} w_1(n) \\ w_2(n) \\ w_3(n) \\ w_4(n) \\ w_1(n+1) \\ w_2(n+1) \\ w_3(n+1) \\ w_4(n+1) \end{bmatrix}}_{\mathcal{W}(n)} \quad (7)$$

where $\mathcal{X}_1$ and $\mathcal{X}_2$ are defined as follows:

$$\mathcal{X}_1 = \begin{bmatrix} x_1(n) & 0 & 0 & 0 \\ 0 & x_1(n) & 0 & 0 \\ 0 & 0 & x_1(n) & 0 \\ 0 & 0 & 0 & x_1(n) \end{bmatrix} \psi_c$$

$$\mathcal{X}_2 = \begin{bmatrix} x_2(n) & 0 & 0 & 0 \\ 0 & x_2(n) & 0 & 0 \\ 0 & 0 & x_2(n) & 0 \\ 0 & 0 & 0 & x_2(n) \end{bmatrix} \psi_c$$

The received signal can be writing more succinct form as follows:

$$\mathcal{Y}(n) = \mathcal{X}\mathcal{H}(n) + \mathcal{W}(n) \quad (8)$$

The least square (LS) solution of observation model by the known transmitted pilot symbols, (8), can be written as,

$$\mathcal{H}(n) = (\mathcal{X}^\dagger \mathcal{X})^{-1} \mathcal{X}^\dagger \mathcal{Y}(n) \quad (9)$$

where $(\mathcal{X}^\dagger \mathcal{X})^{-1} = \frac{1}{2E_{av}}\mathbf{I}$, E_{av} is average symbol energy and $\mathbf{I}$ is identity matrix.

One promising technique for time-varying Rayleigh fading channel is PSA-CE, because this method makes dynamic estimation for such channels. PSA-CE inserts known pilot symbols periodically in the time domain to track the time variation of the channel. Firstly, the received pilot blocks are operated with the known pilot blocks (by LS method given in (9)) for estimating the channel impulse responses (CIRs) on all pilot positions. The PSA-CE with interpolation widely used in new generation wireless communication systems. After the LS estimation, the estimated CIRs on all data positions are calculated by the linear or nearest interpolations. The linear and nearest channel interpolation algorithms are discussed in the following subsections.

A. *Piecewise Linear Interpolation*

It was shown that especially linear interpolation techniques are preferable due to their inherent simplicity and easy to implement [9] for pilot assisted channel estimation by means of interpolation techniques. However, the channel estimation with linear interpolation-based techniques and their performance analysis have not been investigated for the STBC-SM systems in the literature yet. In this work, time-varying, Rayleigh distributed channel coefficients are estimated by the one-dimensional linear interpolation technique in a frame structure, consisting of periodically inserted pilot blocks and the data blocks with certain lengths each. The linear piecewise interpoland can be expressed for $r = 1, 2, 3, 4$ and $p = 1, 2, \cdots, P$ as follows:

$$\begin{aligned} h_{r,j}(n) &= \hat{h}_{r,j}(n_p) + \Big(\hat{h}_{r,j}(n_{p+1}) - \hat{h}_{r,j}(n_p)\Big) \\ &\times \left(\frac{n - n_p}{n_{p+1} - n_p}\right), \text{for } j = 1, 2;\ n_p \le n \le n_{p+1} \\ h_{r,j}(n) &= \hat{h}_{r,j}(n_p + 2) + \Big(\hat{h}_{r,j}(n_{p+1} + 2) - \hat{h}_{r,j}(n_p + 2)\Big) \\ &\times \left(\frac{n - (n_p + 2)}{n_{p+1} - n_p}\right),\ \text{for } j = 3, 4;\ n_p + 2 \le n \le n_{p+1} + 2, \end{aligned} \quad (10)$$

where $\hat{h}_{r,j}(n_p)$ is estimated CIRs at pilot durations, $h_{r,j}(n)$ is estimated CIRs at all data positions and P is the total number of pilot symbols by which the LS channel coefficient estimates are obtained. The n_p and n_p+2 are the discrete times at which the pilot symbols located and usually called breakpoints.

B. Nearest Interpolation (Zero-Order Hold)

The nearest-neighbor interpolation technique interpolates between sample points by holding each sample value until the next sampling instant. In [9], it is shown that the performance of linear interpolation is better than the nearest neighbor interpolation. The zero-order hold interpolator firstly inserts $(V-1)$ zeros between successive samples of channel coefficients, $\hat{h}_{r,j}(n_p)$, as follows:

$$\tilde{h}_{r,j}(n) = \begin{cases} \hat{h}_{r,j}(n_p) & \mathrm{p}=1:\mathrm{P} \\ 0 & \text{otherwise} \end{cases} \tag{11}$$

where $V = P+D$ is the frame length. The interpolated channel samples $h_{r,j}(n)$ are obtained by convolving a zero-order hold interpolation pulse shape with the intermediate signal, $\tilde{h}_{r,j}(n)$ as:

$$h_{r,j}(n) = \sum_{n_p=-\infty}^{\infty} h(n-n_p)\tilde{h}_{r,j}(n_p) \tag{12}$$

where

$$h(n) = \begin{cases} 1 & 0 \le n < T_s \\ 0 & \text{otherwise} \end{cases} \tag{13}$$

Duration of the rectangular pulse is exactly equal to the sampling period.

IV. SIMULATION RESULTS

In this section, simulation results are presented to validate the proposed PSA-CE with interpolation method for the 4×4 STBC-SM systems. 4×4 STBC-SM block consists of $P+D=104$ samples $P=4$ of which constitutes the pilot symbols. Totally, $B=10$ block are transmitted in a frame. The signal to noise ratio (SNR) is defined as $\frac{E_s}{\sigma^2}$ where E_s is the energy per symbol and σ^2 is the noise power. The channel between transmitter and receiver is modeled as a time-varying Rayleigh fading channel having a Doppler shift determined by the terminal velocity and the carrier frequency chosen in the simulations. In our computer simulations, the mobile terminal is assumed to be moving with 120 km/h for a carrier frequency of 1.8 GHz and the symbol duration $T_s = 10\mu s$.

The effect of channel estimation on BER performance of the 4×4 STBC-SM with BPSK scheme is presented in Fig. 3. The results are shown for linear interpolation and nearest interpolation cases assuming we have perfect channel state information (P-CSI). It is observed that the linear interpolation exhibits about 5 dB detection gain over nearest interpolation at BER$=10^{-5}$ and the difference in BER performance between linear and nearest interpolation increases toward high SNR values. Moreover, the nearest interpolation method yields an error floor at high SNRs.

The results in Fig. 4 show that, with QPSK signaling employed, the linear interpolation technique outperforms the nearest interpolation. In particular, it is observed that a 5 dB gain is achieved at BER$=10^{-4}$, as compared with the nearest interpolation technique. It also is demonstrated that the nearest interpolation has irreducible error floors at high SNRs. Consequently, it is not surprising to see from the computer simulations that, the performance of nearest interpolation technique degrades especially at higher SNRs mainly due to the effect of the rapidly varying channel. Thus linear interpolation is preferable mainly due to its superior performance and easy implementation achieved by low computational complexity.

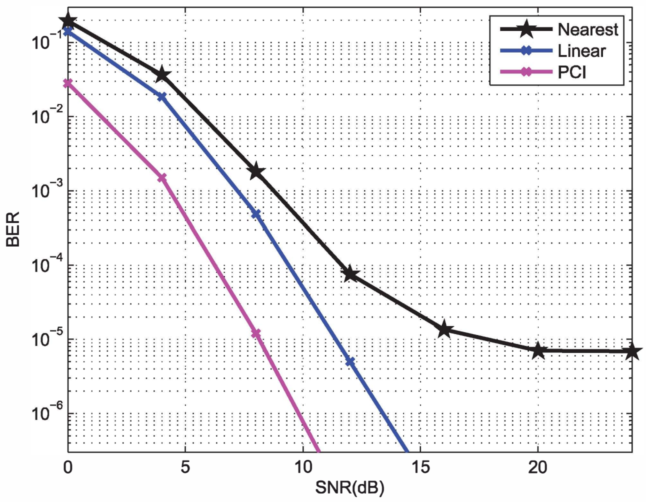

Fig. 3: The BER performance of 4×4 BPSK-STBC-SM with channel estimation at $V=120km/h$

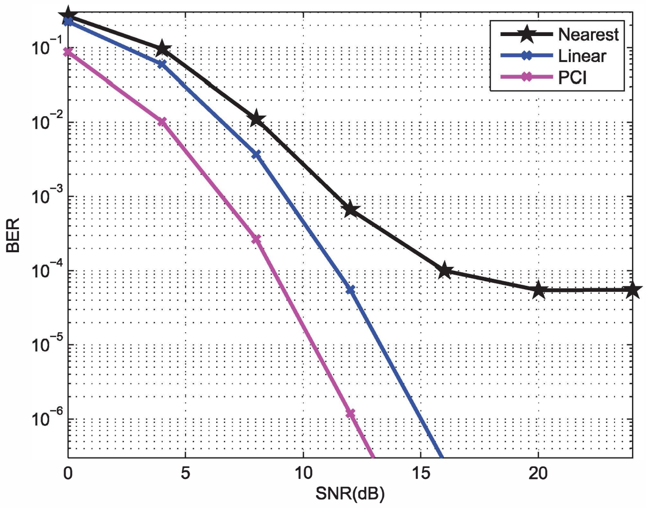

Fig. 4: The BER performance of 4×4 QPSK-STBC-SM with channel estimation at $V=120km/h$

V. CONCLUSIONS

The STBC-SM systems is one of the promising techniques for the next generation wireless communication systems and solving the channel estimation problem is of great interest in design and implementation of these systems. In this paper, we showed that the presence of channel estimation errors inevitably results in a performance degradation for STBC-SM systems and proposed a PSA-CE technique with linear interpolation to track the variations in time-domain over the time-varying Rayleigh fading channel. Linear interpolation based PSA-CE scheme was shown to be superior BER performance over the PSA-CE with the nearest interpolation technique, for a 4×4 STBC-SM system. The computer simulations indicated that the PSA-CE with nearest interpolation technique has an irreducible error floor at higher SNRs. As a conclusion, the PSA-CE scheme with linear interpolation is preferable mainly due to its performance and feasible implementation.

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