NOVEL DETECTION METHODS FOR ZERO-PADDED SINGLE CARRIER SPATIAL MODULATION IN DOUBLY SELECTIVE CHANNELS

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ABSTRACT

In this paper, we present novel methods for signal detection in single carrier zero-padded spatial modulation under high mobility conditions. By expressing the doubly selective channel in terms of the basis expansion model (BEM), first a maximum likelihood (ML) method is presented as a processing framework. To reduce the complexity of the exhaustive ML search, two novel methods, respectively the BEM-based partial interference cancellation (BPIC) and BPIC with successive interference cancellation (BPIC-SIC), are then proposed. The complexity of the new methods are compared and their performance is evaluated by simulations in terms of bit error rate. The results indicate that the new schemes can remarkably improve the performance compared with the conventional methods.

***Index Terms*—** Spatial modulation, detection, high-mobility, interference cancellation, single carrier

1. INTRODUCTION

Spatial modulation (SM) has recently attracted significant attention from both academia and industry, as it can outperforms classical multiple-input multiple-output (MIMO) techniques in terms of spectral efficiency and transceiver complexity design [1,2]. SM is a special type of MIMO technique in which only one antenna is active during signal transmission, and data bits are transmitted through both signal constellation and the index of the active antenna. Thus, it allows utilization of a single radio frequency (RF) chain at the transmitter, and does not need antenna synchronization [3].

In the past few years, several new processing schemes have been presented for improving the performance of SM. In [4, 5], to increase the data rate, the authors proposed generalized SM where more than one antenna is active at each time interval. Quadrature SM was introduced in [3] to improve the spectral efficiency by in-phase and quadrature dimensions of the SM constellation. Moreover, line-of-sight SM for millimeter-wave communication and distributed SM for multirelay wireless networks were recently advanced in [6] and [7], respectively.

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The signal processing methods related to SM has been mostly developed for slow fading frequency flat channels. If these schemes shall be employed in frequency selective channels, the use of orthogonal frequency division multiplexing (OFDM), which can transform a frequency selective channel to several frequency flat channels, is inevitable. However, employment of OFDM negates many advantages of SM [8, 9] such as the possibility of using a single RF chain. Hence, recently, there has been increasing interest in single carrier spatial modulation (SC-SM) techniques.

Considering signal detection in zero-padded SC-SM systems, the authors in [10] have proposed partial interference cancellation (PIC) and PIC with successive interference cancellation (PIC-SIC) schemes for slow fading frequency selective channels. For a similar problem, reduced complexity methods exploiting the structure of the channel have been later presented in [11, 12]. Moreover, a tree search-based method has been developed in [13] for single-carrier generalized spatial modulation (SC-GSM) systems.

As mentioned above, slow fading frequency flat and frequency selective channels have been considered in previous works where it is assumed that the channel remains constant within blocks of data. However, many wireless communication channels experience both time and frequency selectivity, and are time varying within a block of data. Underwater acoustic communication, digital Video Broadcasting (DVB) and next generation cellular systems are examples of systems that should work in these doubly selective channels [14]. To the best of our knowledge, none of the previous works have addressed SC-SM signal detection in doubly selective channels.

In this paper, we propose new detection methods for SC-SM signal detection in doubly selective channels. We apply basis expansion model (BEM) [15] for representing the channel, and first present a maximum likelihood (ML) scheme. The ML method requires an exhaustive search over all possible values of the signals, and consequently is of a very high computational complexity. Aiming at reducing the complexity, we then present BEM based PIC (BPIC), and BPIC with successive interference cancellation (BPIC-SIC). We finally obtain the computational complexity of the new methods and assess the performance of the new schemes using simulations, demonstrating that the proposed schemes have significantly improved performance in comparison with the conventional methods.

2. SYSTEM MODEL

We consider a single carrier MIMO system with N_T transmit and N_R receive antennas. Block-based transmission is assumed where the number of consecutive SM symbols in each block is K. Each SM symbol conveys B information bits, $\log_2 N_T$ of which are transmitted using the active antenna index, and the remaining $\log_2 \zeta$ bits are transmitted using a symbol belonging to a signal constellation $\mathcal{A}$, where ζ is the size of $\mathcal{A}$. In view of this, the SM signal vector applied to the N_T transmit antennas at the kth time interval can be expressed in the following vector form,

$$\mathbf{s}_k = [\underbrace{0, \cdots, 0}_{n_k - 1}, s_k, \underbrace{0, \cdots, 0}_{N_T - n_k}]^T \in \mathbb{C}^{N_T \times 1} \quad (1)$$

where $s_k \in \mathcal{A}$ is the signal transmitted using n_kth antenna while the remaining transmit antennas are inactive. A block of information is then generated by assembling the SM signal vectors into a matrix as follows

$$\mathbf{S} = [\mathbf{s}_0, \mathbf{s}_1, \cdots, \mathbf{s}_{K-1}] \in \mathbb{C}^{N_T \times K}. \quad (2)$$

Subsequently, J zero vectors of size $N_T \times 1$ are appended in front of $\mathbf{S}$, and the resulting block is transmitted through a doubly selective channel. The kth received signal vector $\mathbf{y}_k \in \mathbb{C}^{N_R \times 1}$, after removing the first J zero vectors, can be written as

$$\mathbf{y}_k = \sum_{l=0}^{L_c - 1} \mathbf{H}_{k,l} \mathbf{s}_{k-l} + \mathbf{w}_k, \; k = 0, \cdots, N + L_c - 2 \quad (3)$$

where $\mathbf{w}_k \in \mathbb{C}^{N_R \times 1}$ is a complex additive white Gaussian noise vector with zero mean and covariance $\sigma^2 \mathbf{I}_{N_R}$ where $\mathbf{I}_N$ denotes an $N \times N$ identity matrix. Moreover, L_c is the number of channel taps, and $\mathbf{H}_{k,l} \in \mathbb{C}^{N_R \times N_T}$ represents the doubly selective MIMO channel matrix corresponding to the lth channel tap at the kth time interval.

2.1. BEM Representation of Channel

BEM provides the most accurate representation of a doubly selective channel [16]. Using BEM, $\mathbf{H}_{k,l}$ can be expressed as

$$\mathbf{H}_{k,l} = \mathbf{B}_k \mathbf{C}_l \quad (4)$$

where

$$\mathbf{B}_k = \begin{bmatrix} \mathbf{b}_k^T & \mathbf{0}_{1 \times Q} & \mathbf{0}_{1 \times Q} & \cdots & \mathbf{0}_{1 \times Q} \\ \mathbf{0}_{1 \times Q} & \mathbf{b}_k^T & \mathbf{0}_{1 \times Q} & \cdots & \mathbf{0}_{1 \times Q} \\ \vdots & \vdots & \ddots & & \vdots \\ \mathbf{0}_{1 \times Q} & \mathbf{0}_{1 \times Q} & \cdots & \mathbf{0}_{1 \times Q} & \mathbf{b}_k^T \end{bmatrix} \in \mathbb{C}^{N_R \times N_R Q}, \quad (5)$$

and

$$\mathbf{C}_l = \begin{bmatrix} \mathbf{c}_l^{0,0} & \mathbf{c}_l^{1,0} & \cdots & \mathbf{c}_l^{N_T-1,0} \\ \mathbf{c}_l^{0,1} & \mathbf{c}_l^{1,1} & \cdots & \mathbf{c}_l^{N_T-1,1} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{c}_l^{0,N_R-1} & \mathbf{c}_l^{1,N_R-1} & \cdots & \mathbf{c}_l^{N_T-1,N_R-1} \end{bmatrix} \in \mathbb{C}^{N_R Q \times N_T}. \quad (6)$$

In the above equations, $\mathbf{b}_k = [b_{k,0}, b_{k,1}, \cdots, b_{k,Q-1}]^T \in \mathbb{C}^{Q \times 1}$ denotes the basis function vector at kth time interval, and $\mathbf{c}_l^{m,n} = [c_{l,0}^{m,n}, c_{l,1}^{m,n}, \cdots, c_{l,Q-1}^{m,n}]^T \in \mathbb{C}^{Q \times 1}$ is the basis coefficient vector for the lth MIMO channel tap.

Now, (4) enables us to express the kth received signal in terms of BEM as

$$\mathbf{y}_k = \mathbf{B}_k \sum_{l=0}^{L_c-1} \mathbf{C}_l \mathbf{s}_{k-l} + \mathbf{w}_k, \quad (7)$$

which shows that, ignoring the noise, the received signal is the product of $\mathbf{B}_k$ and the convolution $\sum_{l=0}^{L_c-1} \mathbf{C}_l \mathbf{s}_{k-l}$. Finally, the received vectors can be represented using BEM as

$$\mathbf{y} = \mathbf{G}\mathbf{s} + \mathbf{w} \quad (8)$$

where $\mathbf{y} = [\mathbf{y}_0^T, \cdots, \mathbf{y}_{K'-1}^T]^T$, $\mathbf{w} = [\mathbf{w}_0^T, \cdots, \mathbf{w}_{K'-1}^T]^T$, $\mathbf{s} = [\mathbf{s}_0^T, \mathbf{s}_1^T, \cdots, \mathbf{s}_{K-1}^T]^T$, and

$$\mathbf{G} = \begin{bmatrix} \mathbf{B}_0\mathbf{C}_0 & \mathbf{0}_{N_R \times N_T} & \cdots & \mathbf{0}_{N_R \times N_T} \\ \mathbf{B}_1\mathbf{C}_1 & \mathbf{B}_1\mathbf{C}_0 & \cdots & \mathbf{0}_{N_R \times N_T} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{B}_{L_c-1}\mathbf{C}_{L_c-1} & \mathbf{B}_{L_c-1}\mathbf{C}_{L_c-2} & \cdots & \mathbf{0}_{N_R \times N_T} \\ \mathbf{0}_{N_R \times N_T} & \mathbf{B}_{L_c}\mathbf{C}_{L_c-1} & \cdots & \mathbf{0}_{N_R \times N_T} \\ \vdots & & \ddots & \vdots \\ \mathbf{0}_{N_R \times N_T} & \mathbf{0}_{N_R \times N_T} & \cdots & \mathbf{B}_{K'-L_c-2}\mathbf{C}_0 \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0}_{N_R \times N_T} & \mathbf{0}_{N_R \times N_T} & \cdots & \mathbf{B}_{K'-2}\mathbf{C}_{L_c-2} \\ \mathbf{0}_{N_R \times N_T} & \mathbf{0}_{N_R \times N_T} & \cdots & \mathbf{B}_{K'-1}\mathbf{C}_{L_c-1} \end{bmatrix}. \quad (9)$$

Note that $K' = K + L_c - 1$, $\mathbf{0}_{N_R \times N_T} \in \mathbb{C}^{N_R \times N_T}$ is an all zero matrix, and $(\cdot)^T$ denotes the transpose operation.

3. PROPOSED METHOD

Here, we first present an ML detection method, and next exploit the idea of PIC [10, 17] for development of new reduced-complexity signal detection schemes in doubly selective channels.

3.1. ML Method

Using (8), it is straightforward to show that the detected signals using the ML method can be represented as

$$\hat{\mathbf{s}} = \arg\min_{\tilde{\mathbf{s}} \in \mathcal{S}} \left\| \mathbf{y} - \mathbf{G}\tilde{\mathbf{s}} \right\|^2, \quad (10)$$

where $\hat{\mathbf{s}}$ and $\tilde{\mathbf{s}}$ are, respectively, the detected signal vector and the trial value of the signal vector corresponding to $\mathbf{s}$. Besides, $\mathcal{S}$ denotes the set of all valid transmit vectors with $\left(2^{\log_2 N_T+\log_2 \zeta}\right)^K$ elements. Since the ML method needs an exhaustive search over all possible signal vectors, it has a very high computational complexity, and is not practical. Hence, in the next subsections, we will introduce novel schemes that has significantly lower complexity.

3.2. BPIC Method

PIC which was presented in [17], and then used in [10] assuming that the channel remains constant within a block of data, can be tailored to cope with the time-variations of a doubly selective channel during the transmission of a block of data thanks to BEM. In this respect, in what follows, we propose the BPIC method.

Considering (8), the received signal can be equivalently expressed as

$$\mathbf{y} = \sum_{k=0}^{K-1} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_k} \mathbf{s}_{\mathcal{I}_k} + \mathbf{w}_k, \tag{11}$$

where

$$\bar{\mathbf{B}} = \begin{bmatrix} \mathbf{B}_0 & \mathbf{0}_{N_R \times N_R Q} & \cdots & \mathbf{0}_{N_R \times N_R Q} \\ \mathbf{0}_{N_R \times N_R Q} & \mathbf{B}_1 & \cdots & \mathbf{0}_{N_R \times N_R Q} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0}_{N_R \times N_R Q} & \cdots & \mathbf{0}_{N_R \times N_R Q} & \mathbf{B}_{k'-1} \end{bmatrix} \in \mathbb{C}^{N_R(K+L_c-1) \times N_R Q (K+L_c-1)}, \tag{12}$$

$\mathbf{C}_{\mathcal{I}_k}$ is a matrix composed of columns of $\mathbf{C}$ that are indexed by the elements of $\mathcal{I}_k$, and $\mathbf{s}_{\mathcal{I}_k}$ is a vector containing the elements of $\mathbf{s}$ which are indexed by the elements of $\mathcal{I}_k$. Moreover, $\mathcal{I}_k = \{N_T k, N_T k+1, \cdots, N_T k + N_T - 1\}$, and $\mathbf{C} \in \mathbb{C}^{N_R Q (K+L_c-1) \times N_T K}$ is a block Toeplitz matrix whose first column is given by $\left[\mathbf{C}_0^T, \mathbf{C}_1^T, \cdots, \mathbf{C}_{L_c-1}^T, \mathbf{0}^T, \cdots, \mathbf{0}^T\right]^T$ with $\mathbf{0} \in \mathbb{C}^{N_R Q \times N_T}$ being an all zero matrix. Next, for $m = 0, 1, \cdots, K-1$, we project $\mathbf{y}$ onto the orthogonal complement space of

$$\mathbf{T}_{\mathcal{I}_m} = \left[\bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_0}, \cdots, \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_{m-1}}, \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_{m+1}}, \cdots, \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_{K-1}}\right], \tag{13}$$

using orthogonal projection

$$\mathbf{\Pi}_{\mathcal{I}_m}^{\perp} = \mathbf{I} - \mathbf{T}_{\mathcal{I}_m} \left(\mathbf{T}_{\mathcal{I}_m}^H \mathbf{T}_{\mathcal{I}_m}\right)^{-1} \mathbf{T}_{\mathcal{I}_m}^H, \tag{14}$$

where $(\cdot)^H$ denotes the Hermitian operation. Thus, we have

$$\begin{aligned} \mathbf{u}_m = \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \mathbf{y} &= \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \sum_{k=0}^{k-1} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_k} \mathbf{s}_{\mathcal{I}_k} + \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \mathbf{w}_k \\ &= \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_m} \mathbf{s}_m + \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \mathbf{w}_k, \end{aligned} \tag{15}$$

Algorithm 1 BPIC Method

Inputs: $\mathbf{y}, \bar{\mathbf{B}}, \mathcal{I}_m, m = 0, \cdots, K-1$, and $\mathbf{C}$.
for m from 0 to $K-1$ **do**
  Generate the projection matrix for the mth signal vector as $\mathbf{\Pi}_{\mathcal{I}_m}^{\perp} = \mathbf{I} - \mathbf{T}_{\mathcal{I}_m} \left(\mathbf{T}_{\mathcal{I}_m}^H \mathbf{T}_{\mathcal{I}_m}\right)^{-1} \mathbf{T}_{\mathcal{I}_m}^H$ where $\mathbf{T}_{\mathcal{I}_m}$ is given in (13).
  Project $\mathbf{y}$ onto the orthogonal complement space of $\mathbf{T}_{\mathcal{I}_m}$ as $\mathbf{u}_m = \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \mathbf{y}$.
  Detect the mth transmitted signal vector as $\hat{\mathbf{s}}_m = \arg\min_{\tilde{\mathbf{s}}_m \in \mathcal{S}} \left\| \mathbf{u}_m - \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_m} \tilde{\mathbf{s}}_m \right\|^2$.
end for

where the last equality results from $\mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_k} = 0, \ k \neq m$.

Therefor, the signal at the mth time interval can be detected as

$$\hat{\mathbf{s}}_m = \arg \min_{\tilde{\mathbf{s}}_m \in \mathcal{S}_m} \left\| \mathbf{u}_m - \mathbf{\Pi}_{\mathcal{I}_m}^{\perp} \bar{\mathbf{B}} \mathbf{C}_{\mathcal{I}_m} \tilde{\mathbf{s}}_m \right\|^2, \tag{16}$$

where $\tilde{\mathbf{s}}_k$ is the trial value of $\mathbf{s}_k$, and $\mathcal{S}_m$ denotes the set including all valid transmit vectors whose number of elements is $\mu = 2^{\log_2 N_T + \log_2 \zeta}$. This method is summarized in Algorithm 1.

3.3. BPIC-SIC Method

One way of improving the performance of the BPIC method is exploiting the knowledge of the already detected signals in order to remove the interference caused by them [10]. Inspired by the work in [10], where the focus is on slowly fading channels, we introduce the BPIC-SIC method for doubly selective channels.

We begin with the detection of the first received signal vector using the BPIC scheme. Afterwards, for the detection of the remaining $K-1$ signal vectors, we first remove the interference terms resulting from the previously detected signals, and next apply projection onto the orthogonal complement space of the channel corresponding to the remaining undetected signals. This scheme is described in Algorithm 2.

4. PERFORMANCE EVALUATION

The computational complexities of the proposed schemes are summarized in Table 1 in terms of number of complex multiplications. In this table, we have provided a numerical example based on parameter values given in the paragraph below. According to this table, the proposed BPIC-SIC scheme has the lowest complexity, and complexity of both proposed BPIC and BPIC-SIC are significantly lower than that of the ML method.

We have also assessed the performances of the proposed methods using computer simulation. We consider a single

Table 1. Computational Complexity of Detection Methods

Method	Number of Multiplications	Example
ML	$(KL_c + L_c(L_c-1))QN_RN_T + \mu^K N_R(K+L_c-1)KN_T$	8×10^{81}
BPIC	$(KL_c + L_c(L_c-1))QN_RN_T + (N_RK')^2 + K\big((N_T(K-1))^3 + 2N_T^2N_RK'(K-1)^2 + N_TN_R^2K'^2(K-1) + \mu N_RK'N_T\big)$	4.4×10^9
BPIC-SIC	$(KL_c + L_c(L_c-1))QN_RN_T + \sum_{k=1}^{K}\big((N_Tk)^3 + 2N_T^2N_RK'k^2 + N_TN_R^2K'^2k + (N_RK')^2 + \mu N_RK'N_T\big)$	1.6×10^9

Algorithm 2 BPIC-SIC Method

1: Inputs: $\mathbf{y}, \bar{\mathbf{B}}, \mathcal{I}_i, i=0,\cdots,K-1$, and $\mathbf{C}$.
2: Detect the 0th transmit signal vector using the BPIC method, and put $\mathbf{y}^{(i)} = \mathbf{y}$.
3: **for** i from 0 to $K-2$ **do**
4: Remove the interfernce caused by the already detected signals as $\mathbf{y}^{(i+1)} = \mathbf{y}^{(i)} - \bar{\mathbf{B}}\mathbf{C}_{\mathcal{I}_i}\mathbf{s}_{\mathcal{I}_i}$.
5: Generate the projection matrix $\mathbf{\Pi}^{\perp}_{\mathcal{I}_{i+1}} = \mathbf{I} - \mathbf{T}_{\mathcal{I}_{i+1}}\big(\mathbf{T}^H_{\mathcal{I}_{i+1}}\mathbf{T}_{\mathcal{I}_{i+1}}\big)^{-1}\mathbf{T}^H_{\mathcal{I}_{i+1}}$ where

$$\mathbf{T}_{\mathcal{I}_{i+1}} = \big[\bar{\mathbf{B}}\mathbf{C}_{\mathcal{I}_{i+2}}, \bar{\mathbf{B}}\mathbf{C}_{\mathcal{I}_{i+3}}, \cdots, \bar{\mathbf{B}}\mathbf{C}_{\mathcal{I}_{K-1}}\big].$$

6: Project $\mathbf{y}^{(i+1)}$ onto the orthogonal complement space of $\mathbf{T}_{\mathcal{I}_{i+1}}$ as $\mathbf{u}_{i+1} = \mathbf{\Pi}^{\perp}_{\mathcal{I}_{i+1}}\mathbf{y}^{(i+1)}$.
7: Detect the $(i+1)$th transmitted vector as $\hat{\mathbf{s}}_{i+1} = \arg\min_{\tilde{\mathbf{s}}_{i+1}\in\mathcal{S}} \Big\|\mathbf{u}_{i+1} - \mathbf{\Pi}^{\perp}_{\mathcal{I}_{i+1}}\bar{\mathbf{B}}\mathbf{C}_{\mathcal{I}_{i+1}}\tilde{\mathbf{s}}_{i+1}\Big\|^2$.
8: **end for**

carrier SM MIMO system with $N_T = N_R = 4$, $K = 64$, and $J = 8$. The carrier frequency is 2 GHz and the sampling time is 1μs. The doubly selective channel is a multipath Rayleigh fading channel with Jake's spectrum, and has 5 taps ($l = 0,\cdots,4$) each with the delay of 1μs and exponential power delay profile of $\exp(-\frac{l}{10})$. The normalized Doppler frequency is $f_DT_s = \frac{0.12}{K}$ corresponding to a vehicle speed of 300 km/h, and generalized complex-exponential BEM [15] is used with $Q = 3$. Moreover, quadratic-amplitude modulation (QAM) with constellation sizes of 4 and 16 are used for SM signal generation. It is also assumed that the BEM coefficients are already estimated and known at the receiver.

Figs. 1 and 2 depict the bit error rate (BER) of different detection methods for 4-QAM and 16-QAM signal constellations, respectively, where the performance of PIC and PIC-SIC methods are shown with the assumption of having slow fading channels. It is evident that the new schemes, due to considering the time-variations of channel within a block of data, have significantly lower BER. The proposed BPIC-SIC method has also considerably lower BER than the proposed BPIC, owing to performing interference cancellation. Moreover, it is observed from comparing these two figures that the BER naturally increases with the number of constellation points. It is worth mentioning that we have not shown the performance of the ML scheme, since as shown in Table 1, it is too complex for simulation and impractical.

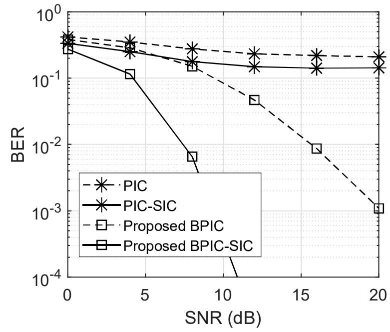

Fig. 1. Performance of different detection schemes for 4-QAM signal constellation.

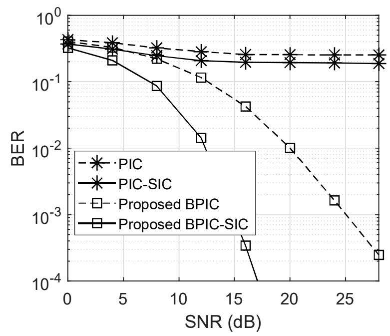

Fig. 2. Performance of different detection schemes for 16-QAM signal constellation.

5. CONCLUSION

This paper addressed zero-padded SC-SM signal detection in time-varying frequency selective channels. BEM were used for capturing time-variations of the channel, and an ML scheme followed by reduced complexity BPIC and BPIC-SIC methods were introduced. Simulation results show remarkable performance improvement in comparison with previous methods. Future works might include comprehensive performance investigation of the new schemes along with presentation of further reduced complexity methods.

6. REFERENCES

[1] L. Wang and Z. Chen, "Stacked alamouti based spatial modulation," *IEEE Trans. Commun.*, pp. 1–1, To be published 2018.

[2] M. D. Renzo, H. Haas, and P. M. Grant, "Spatial modulation for multiple-antenna wireless systems: a survey," *IEEE Commun. Mag.*, vol. 49, no. 12, pp. 182–191, Dec. 2011.

[3] R. Mesleh, S. S. Ikki, and H. M. Aggoune, "Quadrature spatial modulation," *IEEE Trans. Veh. Technol.*, vol. 64, no. 6, pp. 2738–2742, Jun. 2015.

[4] A. Younis, N. Serafimovski, R. Mesleh, and H. Haas, "Generalised spatial modulation," in *Conference Record of the Forty Fourth Asilomar Conference on Signals, Systems and Computers*, Nov 2010, pp. 1498–1502.

[5] J. Wang, S. Jia, and J. Song, "Generalized spatial modulation system with multiple active transmit antennas and low complexity detection scheme," *IEEE Trans. Wire. Commun.*, vol. 11, no. 4, pp. 1605–1615, April 2012.

[6] P. Liu, M. Di Renzo, and A. Springer, "Line-of-sight spatial modulation for indoor mmwave communication at 60 GHz," *IEEE Trans. Wire. Commun.*, vol. 15, no. 11, pp. 7373–7389, Nov. 2016.

[7] S. Narayanan, M. Di Renzo, F. Graziosi, and H. Haas, "Distributed spatial modulation: A cooperative diversity protocol for half-duplex relay-aided wireless networks," *IEEE Trans. Veh. Technol.*, vol. 65, no. 5, pp. 2947–2964, May 2016.

[8] S. Sugiura and L. Hanzo, "Single-RF spatial modulation requires single-carrier transmission: Frequency-domain turbo equalization for dispersive channels," *IEEE Trans. Veh. Technol.*, vol. 64, no. 10, pp. 4870–4875, Oct 2015.

[9] L. Xiao, Y. Xiao, Y. Zhao, P. Yang, M. Di Renzo, S. Li, and W. Xiang, "Time-domain turbo equalization for single-carrier generalized spatial modulation," *IEEE Trans. Wire. Commun.*, vol. 16, no. 9, pp. 5702–5716, Sept 2017.

[10] R. Rajashekar, K. V. S. Hari, and L. Hanzo, "Spatial modulation aided zero-padded single carrier transmission for dispersive channels," *IEEE Trans. Commun.*, vol. 61, no. 6, pp. 2318–2329, Jun. 2013.

[11] L. Xiao, L. Dan, Y. Zhang, Y. Xiao, P. Yang, and S. Li, "A low-complexity detection scheme for generalized spatial modulation aided single carrier systems," *IEEE Commun. Letters*, vol. 19, no. 6, pp. 1069–1072, June 2015.

[12] F. Yao, J. Zheng, and J. Chen, "Reduced-complexity partial interference cancellation receiver in spatial modulation aided zero-padded single carrier transmission," in *International Conference on Wireless Communications Signal Processing (WCSP)*, Oct 2015, pp. 1–5.

[13] L. Xiao, P. Yang, Y. Zhao, Y. Xiao, J. Liu, and S. Li, "Low-complexity tree search-based detection algorithms for generalized spatial modulation aided single carrier systems," in *IEEE International Conference on Communications (ICC)*, May 2016, pp. 1–6.

[14] Y. Li, X. Sha, F. Zheng, and K. Wang, "Low complexity equalization of hcm systems with dpfft demodulation over doubly-selective channels," *IEEE Sig. Process. Letters*, vol. 21, no. 7, pp. 862–865, Jul. 2014.

[15] Z. Tang, R. C. Cannizzaro, G. Leus, and P. Banelli, "Pilot-assisted time-varying channel estimation for ofdm systems," *IEEE Trans. Sig. Process.*, vol. 55, no. 5, pp. 2226–2238, May 2007.

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

[17] X. Guo and X. Xia, "On full diversity spacetime block codes with partial interference cancellation group decoding," *IEEE Trans. Information Theory*, vol. 55, no. 10, pp. 4366–4385, Oct 2009.