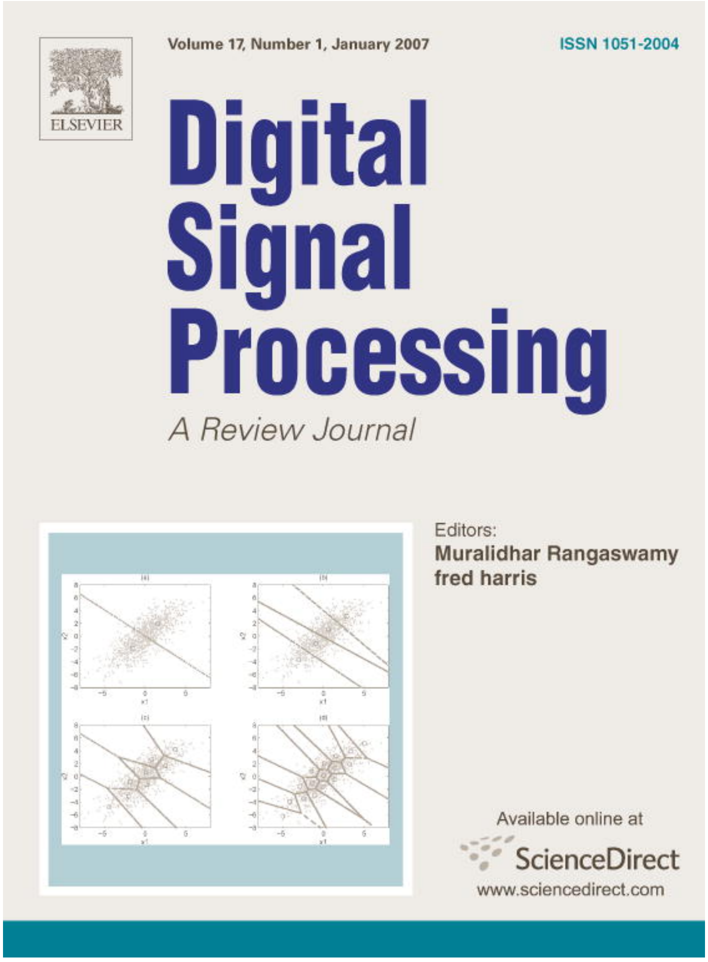
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Generalized multichannel amplitude-and-phase coded modulation for differential space–time communications ☆

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Abstract

We present a generalized multichannel amplitude-and-phase coded modulation scheme for differential space–time communications. Our scheme utilizes code matrices consisting of an amplitude and a phase component, which can be thought of as a space–time multichannel generalization of the scalar amplitude and phase shift keying (APSK) constellation. The amplitude component takes a scalar coefficient that controls the total transmission power, while the phase component is a unitary matrix formed from PSK symbols. Both the amplitude and phase components are differentially encoded and admit efficient differential decoding. We show that the maximum likelihood (ML) decoding of the amplitude coefficient and phase matrix is decoupled. Moreover, the phase matrix, when constructed from orthogonal designs, is amenable to decoupled differential decoding of the phase entries, which further simplifies the decoding complexity significantly. Simulation results show that the proposed amplitude–phase differential space–time coded modulation scheme achieves a performance close to its phase-only counterpart, while providing higher spectral efficiency offered by amplitude modulation.

Keywords: Differential space–time coded modulation; Amplitude-and-phase modulation; Maximum likelihood detection

1. Introduction

Utilizing multi-antenna transmission, space–time modulation/coding offers diversity and possibly coding gain to the receiver (e.g., [4] and references therein). Coherent detection of space–time codes requires channel estimation for multiple radio links, which is a costly and challenging task, especially in fading environments [11]. Differential or non-coherent space–time modulation/coding, on the other hand, circumvents this difficulty. A number of differential space–time modulation schemes have been proposed for both flat-fading [6,7,12,13,22] and frequency-selective fading [3,16–18] channels. Most of these schemes utilize unitary code matrices formed by phase-shift-keying (PSK) entries. These unitary code matrices can be thought of as multichannel extensions of the standard PSK constellations. Therefore, we may call these schemes as *generalized multichannel phase-only modulation* based differential space–time techniques.

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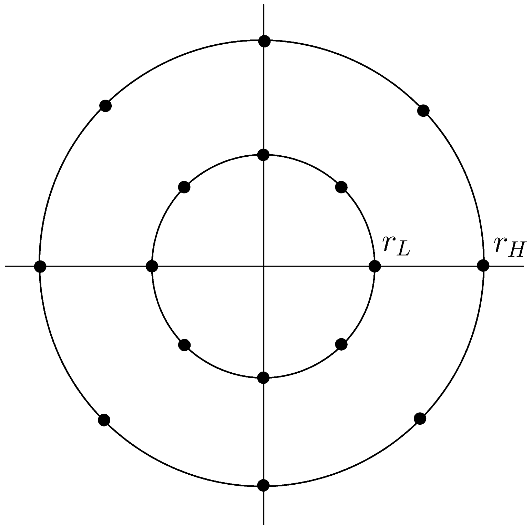

Fig. 1. 16-APSK constellation.

As is well known, PSK constellations are power inefficient when transmission rate is high (e.g., [19]). This motivated the use of multi-level constellations, such as amplitude-and-phase shift keying (APSK), for differential transmissions in single-antenna systems (see [1] and references therein). Differential space–time modulation using multi-level constellations were examined in several recent studies. Specifically, Tao and Cheng [21] proposed a scheme that forms space–time code matrices with multi-level entries from orthogonal designs [23]. Since the code matrix carries non-uniform energy (in terms of the Frobenius norm of the transmitted code matrix), their decoding technique requires an estimate of the energy of the previous code matrix in order to decode the current one. As a result, error propagation may occur. Another method introduced by Xia [25] utilizes APSK constellations for systems equipped with *two* transmit antennas. Specifically, the technique is to draw two APSK symbols at a time that are used to form an Alamouti code matrix [2]. Unlike the previous method, the code matrices so generated have constant energy due to a unique design constraint such that one of the symbol pair is always picked from the inner ring and the other from the outer ring of the APSK constellation (see Fig. 1 for an example of 16-APSK). The code matrix is differentially encoded, similarly to the differential Alamouti scheme [7]. In addition, a one-bit amplitude coefficient, which is differentially encoded by differential ASK [1], is used to control the overall energy transmitted from the two transmit antennas. Both the Alamouti code matrix and the amplitude coefficient can be differentially decoded, thus without incurring error propagation.

In this paper, we introduce a *generalized multichannel amplitude-and-phase coded modulation* scheme that is amenable for differential space–time transmissions. The proposed scheme utilizes space–time multichannel code matrices that can be expressed as a product of an amplitude and a phase component. Naturally, these space–time code matrices can be thought of as multichannel generalizations of the APSK constellations. The amplitude component is a scalar that controls the total transmission power, and the phase component is a unitary matrix formed from PSK symbols (note that [25] uses APSK symbols to construct the matrix). Both the amplitude and phase components are differentially encoded and allow efficient differential decoding. Whereas [25] was tailored for systems with *two* transmit antennas, our proposed coding scheme can accommodate systems with an *arbitrary* number of antennas. Section 3 contains some further discussions on the distinctions between the proposed and Xia's coding methods. We examine differential detection based on the maximum likelihood (ML) principle for the proposed space–time coded modulation scheme. We show that the ML detection of the amplitude coefficient and phase matrix is decoupled; furthermore, the phase code matrix, if constructed by orthogonal designs, offers decoupled differential decoding of the phase entries, thus further reducing the decoding complexity. The proposed scheme yields full spatial diversity. Simulation results show that the proposed amplitude–phase differential space–time coded modulation scheme achieves a performance close to its phase-only counterpart, while providing higher spectral efficiency offered by amplitude modulation.

The rest of the paper is organized as follows. Section 2 contains preliminaries on channel model, differential space–time coded modulation and APSK constellations. In Section 3, we introduce our generalized multichannel amplitude-and-phase coded modulation scheme, and discuss how it can be used for differential space–time communications. Differential detection based on the ML criterion for the proposed scheme is examined in Section 4. Numerical examples are presented in Section 5 to compare the proposed scheme and a standard differential space–time coded modulation technique that uses phase-only modulation. Finally, we summarize this work in Section 6.

2. Preliminaries

In this section, we discuss the channel fading model, and briefly review the phase-only differential space–time coded modulation and APSK constellations. First, we summarize notation used in this paper.

2.1. Notation

Vectors (matrices) are denoted by boldface lower (upper) case letters. All vectors are column vectors. Superscripts $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$ denote the complex conjugate, transpose, and conjugate transpose, respectively. $\Re(\cdot)$ takes the real part of the argument. $E\{\cdot\}$ denotes the statistical expectation. $\mathcal{CN}(0,1)$ denotes a zero-mean, unit-variance complex Gaussian random variable with independent real and imaginary parts, each having variance $1/2$. $\mathbf{I}_N$ denote the $N \times N$ identity matrix. $\otimes$ denotes the matrix Kronecker product [8]. Let $\mathbf{Y} \triangleq \{y_{m,n}\}$ denote an $M \times N$ matrix. Then, $\mathrm{vec}(\mathbf{Y})$ denotes an $MN \times 1$ vector formed by stacking the columns of $\mathbf{Y}$ on top of each other, and $\|\mathbf{Y}\|$ denotes the Frobenius norm of $\mathbf{Y}$ (e.g., [8]):

$$\|\mathbf{Y}\|^2 = \mathrm{tr}\big(\mathbf{Y}^H\mathbf{Y}\big) = \sum_{m,n} |y_{m,n}|^2, \tag{1}$$

where $\mathrm{tr}(\cdot)$ denotes the trace of a square matrix.

2.2. Channel model

We consider a wireless communication system equipped with N_T transmit antennas and N_R receive antennas. We assume that the underlying channels are frequency non-selective (frequency-selective fading can be dealt with by equalization or multicarrier signaling; see, e.g., [3,16]). Let $s_{\mu,t}$ denotes the symbol transmitted from the μth transmitted antenna during the tth symbol period. The baseband signal received at the νth receive antenna is given by

$$y_{\nu,t} = \sqrt{\rho}\sum_{\mu=1}^{N_T} h_{\mu,\nu,t}s_{\mu,t} + w_{\nu,t}, \quad t = 0, 1, \ldots;\ \nu = 1, \ldots, N_R, \tag{2}$$

where ρ is a power scaling factor to be explained shortly, $h_{\mu,\nu,t}$ denotes the complex-valued fading coefficients between the μth transmit and νth receive antenna at time t, and $w_{\nu,t}$ denotes the additive channel noise at receive antenna ν and time t, which is independently, identically distributed (i.i.d.) $\mathcal{CN}(0,1)$ with respect to both t and ν. We consider Rayleigh fading channels so that $\{h_{\mu,\nu,t}\}$ are i.i.d. (with respect to μ and ν) $\mathcal{CN}(0,1)$ random variables. The fading coefficients may change continuously accordingly to, e.g., the Jakes' model [14]. We assume that the fading rate is relatively slow so that the underlying channels remain approximately unchanged for $2N_T$ symbol periods to facilitate differential detection. To ensure that the total transmission power is properly scaled so that it is independent of the number of transmit antennas, we impose the constraint:

$$E\left\{\sum_{\mu=1}^{N_T} |s_{\mu,t}|^2\right\} = 1. \tag{3}$$

It follows that ρ in (2) is the average signal-to-noise ratio (SNR) per receive antenna.

Let

$$\mathbf{Y}_n = \begin{bmatrix} y_{1,nN_T} & \cdots & y_{1,(n+1)N_T-1} \\ \vdots & \ddots & \vdots \\ y_{N_R,nN_T} & \cdots & y_{N_R,(n+1)N_T-1} \end{bmatrix}_{N_R\times N_T}, \tag{4}$$

and note that subscript n denotes the code block index. Likewise, let $\mathbf{H} \in \mathbb{C}^{N_R \times N_T}$, $\mathbf{S}_n \in \mathbb{C}^{N_T \times N_T}$, and $\mathbf{W}_n \in \mathbb{C}^{N_R \times N_T}$ be matrices formed from $\{h_{\mu,\nu,t}\}$, $\{s_{\mu,t}\}$, and $\{w_{\nu,t}\}$, respectively. Then, (2) can be expressed as

$$\mathbf{Y}_n = \sqrt{\rho}\mathbf{H}_n\mathbf{S}_n + \mathbf{W}_n. \tag{5}$$

2.3. Phase-only differential space–time coded modulation

Differential unitary space–time coded modulation schemes considered in [12,13] can be thought of as multichannel phase-only techniques, since the unitary code matrices used therein can be interpreted as *phase matrices* and no amplitude modulation is involved. Let $\mathcal{C}$ be a codebook that contains a set of $N_T \times N_T$ unitary code matrices formed from PSK symbols. In particular, let $\mathbf{S}_0$ be any $N_T \times N_T$ unitary matrix. The differential space–time coded modulation proposed in [12,13] takes the following form:

$$\mathbf{S}_n = \mathbf{S}_{n-1}\mathbf{C}_n, \quad n = 1, 2, \dots, \tag{6}$$

where $\mathbf{C}_n \in \mathcal{C}$ which is selected according to a certain mapping rule that maps a set of information bits to a code matrix. Under the channel model described in the above, the ML differential detector for $\mathbf{C}_n$ is given by [13]

$$\hat{\mathbf{C}}_n = \arg\max_{\mathbf{C}_n \in \mathcal{C}} \Re\{\mathrm{tr}(\mathbf{C}_n\mathbf{Y}_n^H\mathbf{Y}_{n-1})\}. \tag{7}$$

2.4. APSK constellations

We finally review the notation associated with APSK constellations, which form the basis of the proposed amplitude–phase differential space–time coding scheme. Consider an $2M$-APSK constellation that consists of a combination of an independent M-PSK: $\exp\{j2\pi m/M\}$, $m = 0, 1, \dots, M-1$, and a binary ASK (2-ASK): r_L and r_H with

$$\frac{1}{2}(r_L^2 + r_H^2) = 1. \tag{8}$$

Let

$$\gamma \triangleq r_H/r_L. \tag{9}$$

Then, it is ready to show that

$$r_L = \sqrt{2/(\gamma^2+1)}. \tag{10}$$

Fig. 1 depicts an example of the 16-APSK constellation. Note that each $2M$-APSK symbol carries $\log_2 M + 1$ bits of information, with 1 bit carried by the 2-ASK while $\log_2(M)$ bits by the M-PSK.

Differential detection for scalar 16-APSK was considered in [1], with several improved versions reported later. Differential APSK is appealing to mobile communications due to its implementational simplicity and robustness against carrier phase variations [1].

3. Amplitude–phase differential space–time coded modulation

In this section, we present a new differential space–time coded modulation scheme based on generalized amplitude-and-phase multichannel modulation. For ease of exposition, we will focus on the case with $N_T = 2$ transmit antennas. The extension to an arbitrary N_T is straightforward, and is briefly discussed afterward. To initiate the transmission of the nth code block, the space–time encoder takes a total of $2\log_2 M + 1$ bits of information and map them to a 2×2 unitary matrix $\mathbf{C}_n$, formed from a pair of M-PSK symbols, and a one-bit coefficient

$$\alpha_n \in \{1,\ \gamma,\ 1/\gamma\}, \tag{11}$$

where γ is defined in (9). As we shall see, a bit "0" will be mapped to $\alpha_n = 1$ and a bit "1" be mapped to $\alpha = \gamma$ or $1/\gamma$, depending on the encoder state. The composite space–time code matrix $\alpha_n\mathbf{C}_n$ is a multichannel extension of the scaler APSK constellation, with α_n being denoted as the *amplitude coefficient* and $\mathbf{C}_n$ the *phase matrix*. The space–time code matrix $\alpha_n \mathbf{C}_n$ is then differentially modulated (to be specified shortly) and transmitted over a period of 2 symbol periods, thus yielding a spectral efficiency of $(\log_2 M + 0.5)$ bits/s/Hz.

The unitary phase matrix $\mathbf{C}_n$ can be formed in various ways. Examples include the diagonal cyclic space–time codes [12], unitary cyclic and bicyclic group codes [13], high-rate group codes [20], and Cayley transform based codes [10], among others. For efficient decoding, we consider the one based on orthogonal designs [23], which reduces to the Alamouti scheme for $N_T = 2$ [2]. In particular, we map the first $2\log_2 M$ bits of information to two M-PSK symbols $c_{n,1}$ and $c_{n,2}$. The phase code matrix is formed as follows:

$$\mathbf{C}_n = \frac{1}{\sqrt{2}} \begin{bmatrix} c_{n,1} & -c_{n,2}^* \\ c_{n,2} & c_{n,1}^* \end{bmatrix}, \tag{12}$$

where the scaling factor of $1/\sqrt{2}$ is to ensure that $\mathbf{C}_n$ is unitary and we assume that $c_{n,1}$ and $c_{n,2}$ are drawn from a unit-energy M-PSK constellation. The last information bit is mapped to the amplitude coefficient α_n as described below.

- *Initialization*:
  For $n = 0$, let
  $$\mathbf{D}_0 = \mathbf{I}_2, \tag{13}$$
  $$\beta_0 = r_L. \tag{14}$$
  The first transmitted space–time code matrix is
  $$\mathbf{S}_0 = \beta_0 \mathbf{D}_0. \tag{15}$$
  Note that $\mathbf{D}_0$ can be replaced by any unitary 2×2 matrix without affecting the performance. Likewise, we can also initialize $\beta_0 = r_H$ if (16) is modified accordingly to ensure proper transition of the amplitude.
- *Differential encoding*:
  For $n > 0$, the encoder takes a total of $2\log_2 M + 1$ bits of information, in which the first $2\log_2 M$ bits are mapped to $\mathbf{C}_n$, formed as in (12), and the last bit, denoted by b_n, is mapped to the amplitude coefficient α_n. The amplitude coefficient is assigned and differentially encoded according to the following rule:
  $$\alpha_n = \begin{cases} 1, & \text{if } b_n = 0, \\ \gamma, & \text{if } b_n = 1 \text{ and } \beta_{n-1} = r_L, \\ 1/\gamma, & \text{if } b_n = 1 \text{ and } \beta_{n-1} = r_H, \end{cases} \tag{16}$$
  $$\beta_n = \beta_{n-1}\alpha_n. \tag{17}$$
  The phase matrix is differentially encoded as follows:
  $$\mathbf{D}_n = \mathbf{D}_{n-1}\mathbf{C}_n. \tag{18}$$
  The transmitted space–time code matrix $\mathbf{S}_n$ is given by
  $$\mathbf{S}_n = \beta_n \mathbf{D}_n = \mathbf{S}_{n-1}(\alpha_n \mathbf{C}_n). \tag{19}$$

The above multichannel amplitude–phase modulation achieves full spatial diversity offered by orthogonal designs. Specifically, it is easy to show that the difference of two code matrices $(\mathbf{S}_i - \mathbf{S}_j)$, for $\mathbf{S}_i \neq \mathbf{S}_j$, is full rank, which ensures full diversity according to the rank criterion in [24]. Meanwhile, the use of an amplitude coefficient β_n provides additional power efficiency compared to phase-only modulation, which will be shown in Section 5.

It is straightforward to extend the scheme to the case with $N_T > 2$ transmit antennas. We may simply use orthogonal designs for arbitrary N_T to construct $\mathbf{C}_n$ (see [5,23]), which will retain full diversity and efficient decoding, but at the cost of losing rate for $N_T > 2$. Alternatively, we may also consider using other unitary space–time coded modulation schemes, such as [10,12,13,20], for $N_t > 2$, although the decoding complexity is in general higher, especially for large N_T and/or transmission rate.

Our coding scheme is closely related to Xia's method [25], but there are several major distinctions. Specifically, Xia's method was intended primarily for systems with only two antennas, whereas our coding scheme can be applied in systems with any number of transmit antennas. In particular, Xia's method divides the information bits into three streams, one is to select symbols from the inner ring of the 2-APSK, one is to select symbols from the outer ring, and the third is to select the amplitude bit. If more than two transmit antennas are involved, a direct extension would require APSK with more than two amplitude levels, which will create some difficulties, or the encoding process should be modified. In contrast, we approach breaks the encoding process into two steps, one for amplitude coding and one for (multichannel) phase coding. While the amplitude coding process is identical to Xia's method, our phase coding process is different and it directly encodes on PSK constellations, as opposed to APSK constellations in Xia's method. As a result, we do not have any problem extending to more than 2 transmit antennas, which is briefly discussed in the above.

Another distinction is that we have developed an ML differential detector, as shown in the next section, and proved that the ML detection of the amplitude bit and the phase matrix is decoupled. Our ML detection also differs from Xia's detection in that the phase matrix is decoded before the amplitude bit, whereas the reverse order is performed in [25].

4. ML differential detection

The signals received at the N_R receive antennas are described by (5). Substituting (19) into (5) leads to

$$\begin{aligned}\mathbf{Y}_n &= \sqrt{\rho}\mathbf{H}_n\mathbf{S}_{n-1}(\alpha_n\mathbf{C}_n) + \mathbf{W}_n \\ &= (\mathbf{Y}_{n-1} - \mathbf{W}_{n-1})(\alpha_n\mathbf{C}_n) + \mathbf{W}_n \\ &= \alpha_n\mathbf{Y}_{n-1}\mathbf{C}_n + \mathbf{V}_n,\end{aligned} \tag{20}$$

where we have assumed that $\mathbf{H}_n$ remains unchanged within two adjacent code blocks (see Section 2.2), and

$$\mathbf{V}_n \triangleq \mathbf{W}_n - \alpha_n\mathbf{W}_{n-1}\mathbf{C}_n. \tag{21}$$

We next determine the covariance matrix of $\mathbf{v}_n \triangleq \operatorname{vec}(\mathbf{V}_n)$. Let $\mathbf{w}_n \triangleq \operatorname{vec}(\mathbf{W}_n)$ and $\mathbf{w}_{n-1} \triangleq \operatorname{vec}(\mathbf{W}_{n-1})$. Then,

$$E\{\mathbf{v}_n\mathbf{v}_n^H\} = E\{\mathbf{w}_n\mathbf{w}_n^H\} + \alpha_n^2 E\{\operatorname{vec}(\mathbf{W}_{n-1}\mathbf{C}_n)\operatorname{vec}^H(\mathbf{W}_{n-1}\mathbf{C}_n)\} \tag{22}$$

$$= \mathbf{I}_{N_RN_T} + \alpha_n^2 E\{(\mathbf{C}_{n-1}^T \otimes \mathbf{I}_{N_T})\mathbf{w}_{n-1}\mathbf{w}_{n-1}^H(\mathbf{C}_{n-1}^T \otimes \mathbf{I}_{N_R})^H\} \tag{23}$$

$$= \mathbf{I}_{N_RN_T} + \alpha_n^2(\mathbf{C}_n^H\mathbf{C}_n)^T \otimes \mathbf{I}_{N_R} \tag{24}$$

$$= (1+\alpha_n^2)\mathbf{I}_{N_RN_T}, \tag{25}$$

where (22) is due to that $\mathbf{W}_n$ and $\mathbf{W}_{n-1}$ are independent of one another, in (23) and (24), we used the identities [9]

$$\operatorname{vec}(\mathbf{AB}) = (\mathbf{B}^T \otimes \mathbf{I})\operatorname{vec}(\mathbf{A}), \tag{26}$$

$$(\mathbf{A}\otimes\mathbf{C})(\mathbf{B}\otimes\mathbf{D}) = (\mathbf{AB})\otimes(\mathbf{CD}), \tag{27}$$

for arbitrary matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$ and $\mathbf{D}$ with compatible sizes, and in (25), we used the fact that $\mathbf{C}_n$ is unitary. The above calculation reveals that conditioned on $\mathbf{C}_n$, $\mathbf{V}_n$ consists of i.i.d. complex Gaussian entries with zero mean and variance $1+\alpha_n^2$.

In the sequel, we consider ML detection based on (20). The likelihood function of $\mathbf{Y}_n$, conditioned on α_n, $\mathbf{C}_n$ and $\mathbf{Y}_{n-1}$, is given by (e.g., [15])

$$p(\mathbf{Y}_n|\mathbf{Y}_{n-1},\alpha_n,\mathbf{C}_n) = \frac{1}{\pi^{N_RN_T}(1+\alpha_n^2)^{N_RN_T}}\exp\left\{-\frac{1}{1+\alpha_n^2}\|\mathbf{Y}_n - \alpha_n\mathbf{Y}_{n-1}\mathbf{C}_n\|^2\right\}. \tag{28}$$

Maximizing the likelihood function is equivalent to minimizing

$$f(\alpha_n,\mathbf{C}_n) \triangleq N_RN_T\log(1+\alpha_n^2) + \frac{\|\mathbf{Y}_n - \alpha_n\mathbf{Y}_{n-1}\mathbf{C}_n\|^2}{1+\alpha_n^2}. \tag{29}$$

Since $\mathbf{C}_n$ is unitary, we have

$$\|\mathbf{Y}_n - \alpha_n\mathbf{Y}_{n-1}\mathbf{C}_n\|^2 = \|\mathbf{Y}_n\|^2 + \alpha_n^2\|\mathbf{Y}_{n-1}\|^2 - 2\alpha_n\Re\{\operatorname{tr}(\mathbf{Y}_n^H\mathbf{Y}_{n-1}\mathbf{C}_n)\}. \tag{30}$$

Therefore, $f(\alpha_n, \mathbf{C}_n)$ can be expressed as

$$f(\alpha_n, \mathbf{C}_n) = f_1(\alpha_n) - f_2(\alpha_n) f_3(\mathbf{C}_n), \tag{31}$$

where

$$f_1(\alpha_n) \triangleq N_R N_T \log\left(1+\alpha_n^2\right) + \frac{\|\mathbf{Y}_n\|^2 + \alpha_n^2 \|\mathbf{Y}_{n-1}\|^2}{1+\alpha_n^2}, \tag{32}$$

$$f_2(\alpha_n) \triangleq \frac{2\alpha_n}{1+\alpha_n^2}, \tag{33}$$

$$f_3(\mathbf{C}_n) \triangleq \Re\left\{\operatorname{tr}\left(\mathbf{Y}_n^H \mathbf{Y}_{n-1} \mathbf{C}_n\right)\right\}. \tag{34}$$

Equation (31) indicates that the *decoding of* α_n *and* $\mathbf{C}_n$ *is decoupled*. In particular, we can first decode $\mathbf{C}_n$ by maximizing $f_3(\mathbf{C}_n)$, and then substitute the maximizing $\mathbf{C}_n$ back into (31) to decode α_n.

To proceed, we first decode the phase matrix by maximizing $f_3(\mathbf{C}_n)$ over all possible phase matrices. If $\mathbf{C}_n$ is obtained by orthogonal designs, the decoding of $\mathbf{C}_n$ can be further simplified and the associated decoding complexity is linear, which is shown next. As in Section 3, we assume $N_T = 2$ and the extension to $N_T \geqslant 2$ is straightforward. Specifically, it is easy to see that maximizing $f_3(\mathbf{C}_n)$ is equivalent to minimizing

$$f_3'(\mathbf{C}_n) = \|\mathbf{Y}_n - \mathbf{Y}_{n-1}\mathbf{C}_n\|^2. \tag{35}$$

Let $\mathbf{y}_{n,1}$ and $\mathbf{y}_{n,2}$ be the first and second column of $\mathbf{Y}_n$, and $\mathbf{y}_{n-1,1}$ and $\mathbf{y}_{n-1,2}$ are similarly defined for $\mathbf{Y}_{n-1}$. We can write f_3' as follows (see (12))

$$f_3'(c_{n,1}, c_{n,2}) = \|\mathbf{y}_{n,1} - c_{n,1}\mathbf{y}_{n-1,1} - c_{n,2}\mathbf{y}_{n-1,2}\|^2 + \|\mathbf{y}_{n,2} + c_{n,2}^*\mathbf{y}_{n-1,1} - c_{n,1}^*\mathbf{y}_{n-1,2}\|^2 \tag{36}$$

$$= \|\mathbf{y}_{n,1} - c_{n,1}\mathbf{y}_{n-1,1} - c_{n,2}\mathbf{y}_{n-1,2}\|^2 + \|\mathbf{y}_{n,2}^* - c_{n,1}\mathbf{y}_{n-1,2}^* + c_{n,2}\mathbf{y}_{n-1,1}^*\|^2 \tag{37}$$

$$= \|\tilde{\mathbf{y}}_n - \tilde{\mathbf{Y}}_{n-1}\mathbf{c}_n\|^2, \tag{38}$$

where in (37) we took the conjugation of the second term, which does not affect the norm, and in (38), we used the following definitions:

$$\tilde{\mathbf{y}}_n \triangleq \left[\mathbf{y}_{n,1}^T,\ \mathbf{y}_{n,2}^H\right]^T, \tag{39}$$

$$\mathbf{c}_n \triangleq [c_{n,1},\ c_{n,2}]^T, \tag{40}$$

$$\tilde{\mathbf{Y}}_{n-1} \triangleq \begin{bmatrix} \mathbf{y}_{n-1,1} & \mathbf{y}_{n-1,2} \\ \mathbf{y}_{n-1,2}^* & -\mathbf{y}_{n-1,1}^* \end{bmatrix}. \tag{41}$$

It is ready to verify that $\tilde{\mathbf{Y}}_{n-1}$ has orthogonal columns and

$$\tilde{\mathbf{Y}}_{n-1}^H \tilde{\mathbf{Y}}_{n-1} = \left(\|\mathbf{y}_{n-1,1}\|^2 + \|\mathbf{y}_{n-1,2}\|^2\right)\mathbf{I}_2. \tag{42}$$

Therefore, the phase angles of $\mathbf{c}_n$, i.e.,

$$\boldsymbol{\theta}_n \triangleq \arg(\mathbf{c}_n), \tag{43}$$

can be estimated by computing $\arg(\tilde{\mathbf{Y}}_{n-1}^H \tilde{\mathbf{y}}_n)$ followed by rounding to the nearest multiple of $2\pi/M$. Clearly, the decoding of $c_{n,1}$ and $c_{n,2}$ is decoupled, and the complexity is linear in N_T.

Once we have the decoded symbols $\hat{c}_{n,1}$ and $\hat{c}_{n,2}$, we use them to form $\hat{\mathbf{C}}_n$, substitute it back to (31) and decode α_n as follows:

$$\hat{\alpha}_n = \arg \min_{\alpha_n \in \{1, \gamma, 1/\gamma\}} f(\alpha_n, \hat{\mathbf{C}}_n). \tag{44}$$

Finally, $\hat{\alpha}_n$ is mapped a bit "0" if $\hat{\alpha}_n = 1$, and a bit "1" if $\hat{\alpha}_n = \gamma$ or $1/\gamma$.

Overall, the decoding complexity is only slightly higher than that of a phase-only counterpart scheme, e.g., [22], due to the extra minimization incurred in (44).

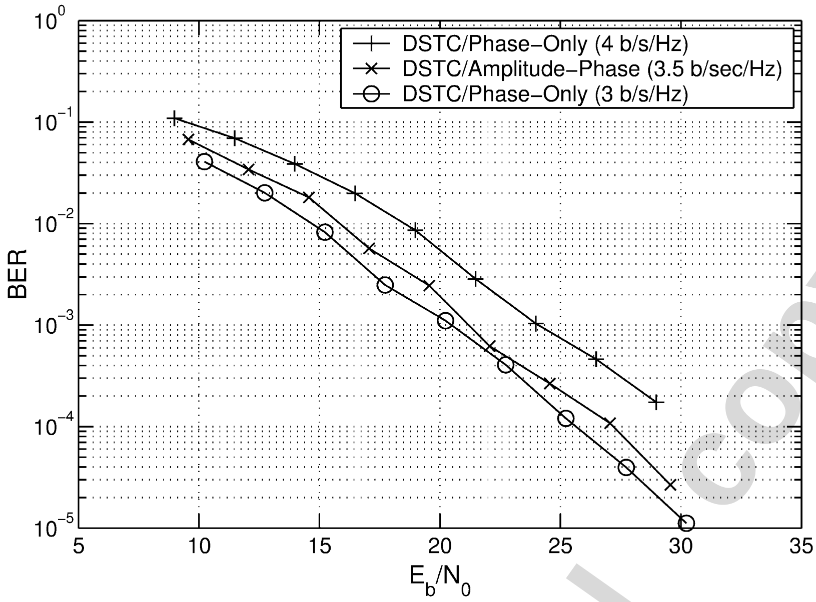

Fig. 2. Bit error rate (BER) as a function of E_b/N_0 for a phase-only differential space–time coding (DSTC) with 8PSK and 16PSK, and the proposed amplitude–phase DSTC scheme with 2ASK and 8PSK in Rayleigh fading channels when $N_T = 2$ and $N_R = 1$.

5. Numerical results

We consider a system equipped with $N_T = 2$ transmit antennas and $N_R = 1$ receive antenna. The underlying channel is flat Rayleigh fading, i.e., the channel coefficients contained in $\mathbf{H}$ are generated as i.i.d. $\mathcal{C}(0, 1)$ random variables, varying independently from trial to trial. We consider two differential space–time coding (DSTC) schemes, namely the proposed one based on generalized multichannel amplitude–phase modulation, which will henceforth be referred to as *DSTC/Amplitude–Phase* for brevity, and the one in [22], which can be interpreted as a multichannel phase-only scheme in this paper and, hence, is referred to as *DSTC/Phase-Only*. Due to the additional amplitude bit used in our scheme, we cannot match the data rate for both schemes exactly. Instead, we compare the two schemes for the nearest possible data rates. The performance measure is the bit error rate (BER) as a function of E_b/N_0, where E_b denotes the total energy per bit used in the transmission.

Fig. 2 depicts the BER of the DSTC/Phase-Only scheme with 8PSK and 16PSK constellations, respectively, with the associated data rate of 3 bits/s/Hz and 4 bits/s/Hz, respectively. Also shown there is the BER of the proposed DSTC/Amplitude–Phase scheme with 16APSK (i.e., 2ASK and 8PSK) with a rate of 3.5 bits/s/Hz. The 2ASK uses $\gamma = 1.6$, which was found to provide good performance for our scheme. In this and next examples, the value of γ was found by simulation. That is, we tried many values of γ with a range of SNR, and picked the one that gave the lowest BER in all simulations. The optimum γ is the one that minimizes the exact average BER and, in general, varies with SNR. However, an exact expression of the average BER (averaged with respect to the channel fading) is unavailable this time.

Fig. 3 depicts the BER of the DSTC/Phase-Only scheme with 16PSK (4 bits/s/Hz) and 32PSK (5 bits/s/Hz), respectively, and our DSTC/Amplitude–Phase scheme with 32APSK (4.5 bits/s/Hz) and $\gamma = 1.5$. Finally, Fig. 4 depicts the BER of the DSTC/Phase-Only scheme with 32PSK (5 bits/s/Hz) and 64PSK (6 bits/s/Hz), respectively, and our DSTC/Amplitude–Phase scheme with 64APSK (5.5 bits/s/Hz) and $\gamma = 1.3$.

It is seen from Figs. 2 to 4 that the proposed scheme achieves BER close to that of the lower-rate DSTC/Phase-Only scheme, and outperforms the higher-rate DSTC/Phase-Only scheme. This is particularly the case when the transmission rate increases, as shown in Figs. 3 and 4. Also noted is that all schemes yield full spatial diversity, which can be verified by examining the slope of the BER curves.

6. Concluding remarks

We have presented a differential space–time coded modulation scheme based on generalized multichannel amplitude and phase modulation. We have shown that the proposed scheme admits decoupled decoding of the amplitude

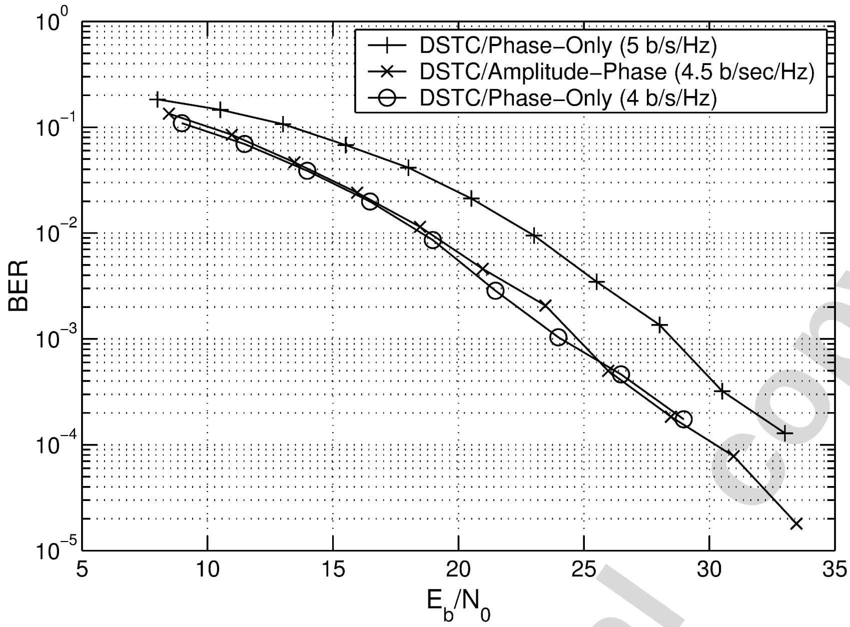

Fig. 3. Bit error rate (BER) as a function of E_b/N_0 for a phase-only differential space–time coding (DSTC) with 16PSK and 32PSK, and the proposed amplitude–phase DSTC scheme with 2ASK and 16PSK in Rayleigh fading channels when $N_T = 2$ and $N_R = 1$.

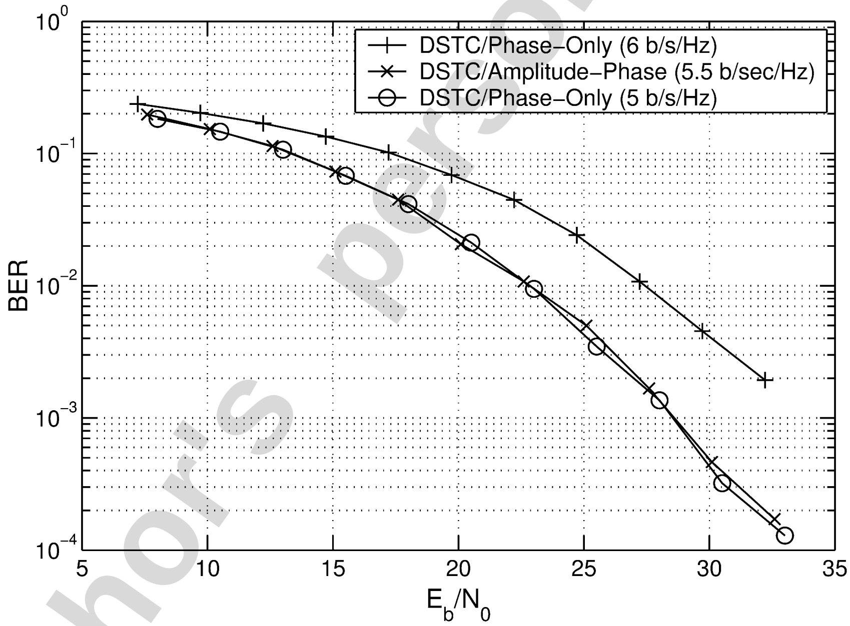

Fig. 4. Bit error rate (BER) as a function of E_b/N_0 for a phase-only differential space–time coding (DSTC) with 32PSK and 64PSK, and the proposed amplitude–phase DSTC scheme with 2ASK and 32PSK in Rayleigh fading channels when $N_T = 2$ and $N_R = 1$.

coefficient and phase matrix, as well as decoupled decoding of the phase entries of the phase matrix, given that the latter is formed by orthogonal designs. The proposed amplitude–phase differential space–time coding scheme achieves a BER performance close to its counterpart based on phase-only modulation, and offers higher spectral efficiency provided by amplitude modulation at the cost of slightly higher decoding complexity.

Our studies also open up several future research directions. First, as suggested by one of our reviewers, an alternative technique would be to consider a "hybrid" approach that utilizes differential modulation for the phase code matrix but non-coherent modulation for the amplitude coefficient. Such a hybrid scheme, however, appears more difficult to decode. A critical issue is to develop efficient decoding techniques. Second, there is an interest to extend our studies to include quadrature amplitude modulation (QAM) constellations, which offer better power efficiency than APSK at high data rates. In general, differential space–time designs with QAM will be less convenient since more than two amplitude levels may be involved in differential modulation and, furthermore, the number of constellation points associated with each amplitude level is not fixed. Finally, we may consider adaptive modulation in amplitude-and-phase coded multichannel systems by letting the number of amplitude levels adapt to the channel fading. Specifically, we can use more amplitude levels as the channel condition improves. As adaptive modulation usually requires channel feedback, this extension will be a further departure from the differential approach considered here which is primarily for open-loop applications.

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