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Cooperative diversity with continuous phase modulation

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Abstract

In recent years, there has been a vast amount of research on cooperative diversity due to its potential capacity increase. Almost all of the research has been based on linear modulation formats such as phase-shift keying (PSK) or quadrature amplitude modulation (QAM). On the other hand, continuous phase modulation (CPM) promises impressive advantages relative to linear modulations for wireless communication systems. In this paper, a cooperative diversity scheme with space-time coded (STC) CPM is introduced. Both amplify-and-forward (AF) and decode-and-forward (DF) modes were considered for cooperation. Simulation results showed that the proposed system is very suitable for wireless fading channels, since it provides diversity gain and utilizes the advantages of CPM.

Key Words: *Continuous phase modulation, cooperative diversity*

1. Introduction

The signal fading effect is one of the main problems due to multipath propagation in wireless communication. This undesired effect can be mitigated by diversity techniques, such as antenna diversity [1]. Unfortunately, the use of multiple antennas might not be practical in some wireless systems as well as in wireless sensor networks (WSN) due to size and power constraints. To overcome this problem, cooperative diversity was proposed by Sendonaris et al. [2, 3] and Laneman et al. [4]. In cooperative diversity, the source terminal uses available mobile terminals as relays; forwarding the original message from the source to the intended destination. In this manner, the source and relays cooperate to form a virtual antenna array and improve spectral and power efficiency of the wireless networks without the additional complexity of multiple antennas. Recently, various cooperation modes have been proposed in the literature [4, 5]. AF and DF modes have received much attention among all of the proposed cooperation techniques. Here we focus on both AF and DF cooperation modes.

Space-time coding [6-8] for multiple transmit antennas has attracted considerable attention due to its potential capacity increase. In [9], Laneman et al. showed that space-time coding can be used in cooperative

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communication in a distributed fashion. After this pioneering work, many studies have been conducted on space-time coded cooperative diversity [10, 11]. However, all of these studies consider linear modulation formats. On the other hand, continuous phase modulation [12-14] is very attractive for wireless communication due to its spectral efficiency and wireless fading resistance. Furthermore, its constant envelope property allows use of inexpensive nonlinear amplifiers. Recently, some space-time coding schemes for CPM were introduced [15-19] and shown to have advantages relative to linear modulations.

In this paper, a cooperative diversity scheme with CPM is introduced both to achieve diversity gain and to utilize the advantages of continuous phase. A differential orthogonal STC technique, which was based on Alamouti's scheme and originally proposed in [19], was used to design STC-CPM codes. In this technique, as explained in Section 2, the space-time code design for CPM is very simple and it maintains the continuous phase constraint that is crucial for bandwidth efficiency. Moreover, this technique guarantees achievement of full diversity. Both AF and DF cooperation modes were considered to construct the system model. A special protocol, which was originally proposed by Laneman et al. in [4], was considered for cooperation. This protocol manages transmission within 2 phases, which will be explained in Section 3.

The rest of the paper is organized as follows. Section 2 describes the STC-CPM code design for multiple antenna transmission. Section 3 deals with the general system model. Simulation results and comments are presented in Section 4. Finally, Section 5 concludes the paper.

2. STC-CPM code design for multiple antenna transmission

CPM is a good choice for cooperative communication due to its spectral efficiency and resistance to wireless channel fading. However, the continuous phase constraint of CPM complicates the CPM code design for multiple antenna systems. Here, a differential approach [19] was used to design the STC-CPM signals. In cooperative transmission, N $=2^{\mu}$-1 relay terminals assist the source terminal, where μ is an integer and, without loss of generality, it is assumed that $\mu = 1$. The source terminal, destination terminal, and each relay terminal are equipped with only one transmitting and one receiving antenna. Here, M-ary continuous phase frequency shift keying (M-CPFSK), which is a special case of CPM, was considered for CPM signal design. The M-CPFSK signal with the information-bearing phase is given by

$$s(t,Y) = \sqrt{\frac{2E_s}{T}}\cos(2\pi f_1 t + \phi(t,Y) + \phi_0), \tag{1}$$

where f_1 is the asymmetric carrier frequency as $f_1 = f_c - h(\mathrm{M}-1)/2\mathrm{T}$, and f_c is the carrier frequency. E_s is the energy per channel symbol and ϕ_0 is the initial carrier phase. It is assumed that $f_1 T$ is an integer; this condition leads to a simplification when using the equivalent representation of the CPM waveform. The information-bearing phase can be expressed as

$$\phi(t,Y) = 4\pi h \sum_{i=0}^{\infty} Y_i \; q(t-iT), \tag{2}$$

where the modulation index h is equal to 1/M for full response CPM. Y is an input sequence of M-ary symbols, $\mathrm{Y}_i \in \{0, 1, 2, \ldots, \mathrm{M}-1\}$. T is the channel symbol period. The phase response function, $q(t)$, is a continuous and monotonically increasing function subject to the constraints

$$q(t) = \begin{cases} 0 & t \leq 0 \\ 1/2 & t \geq LT \end{cases} , \tag{3}$$

where L is an integer.

While designing space-time coded CPM signals for multiple antenna systems, the key requirement is that phase continuity must be maintained both within each space-time symbol and between space-time symbols. They can be designed in a differential way, as follows.

First, a differential unitary code is defined by a group of space-time data symbols, U. $U= \{\mathbf{U}_1,\mathbf{U}_2, \mathbf{U}_3,\ldots,\mathbf{U}_P\}$. Note that the proper selection of U is a crucial issue for the valid CPM phase transitions (for details, see [19]). The space-time code symbol set can be generated by using U and the space-time code seed matrix ($\mathbf{V}_1$). $V = U\mathbf{V}_1= \{\mathbf{V}_1, \mathbf{V}_2,\ldots, \mathbf{V}_P\}$. Finally, space-time code symbols $\{\mathbf{c}_n\}$ are constructed differentially by multiplying the space-time information bearing data symbol $\{\mathbf{d}_n\}$, which is chosen from U, and the previously transmitted symbol $\{\mathbf{c}_{n-1}\}$.

$$\mathbf{c}_n = \mathbf{d}_n\mathbf{c}_{n-1} \tag{4}$$

Here, $\mathbf{c}_0$ can be initialized with any element of the code seed matrix set. The structure of U guarantees to provide the property $\mathbf{c}_n \subset V$ for all n. Let $\mathbf{c}_n^{(a,b)}$ denote the $(a,b)^{th}$ element of the n^{th} space time code symbol. $\mathbf{c}_n^{(a,b)}$ represents the sample of the M-CPFSK signal transmitted from the b^{th} antenna, at the a^{th} subinterval of the n^{th} space-time interval. In this case, the space-time symbol period is $\mathrm{T}_{ST} = \mathrm{NT}$. For each T-spaced interval, the M-CPFSK state will alternate between 2 disjointed subsets, Q_0 and Q_1.$\mathrm{Q}_0 = \{\varphi_{2,2M}, \varphi_{4,2M}, \ldots, \varphi_{2M,2M}\}$ and $\mathrm{Q}_1 = \{\varphi_{1,2M}, \varphi_{3,2M}, \ldots, \varphi_{2M-1,2M}\}$ where $\varphi_{l,L} = \exp\{j2\pi l/L\}$. For 2-CPFSK and 4-CPFSK with a source and a relay terminal, the space-time code seed matrix $\mathbf{V}_1$ and the primitive element of U, Ω, are given in the Table. Here, the primitive element defines the underlying differential unitary code U, $U = \langle\Omega\rangle$. These parameters were first defined in [19] for space-time coding with 2 transmit antennas.

Table. Differential codes for 2-CPFSK and 4-CPFSK modulations.

Modulation	Code seed matrix V_1	Primitive element Ω	$\lvert U\rvert$
2-CPFSK	$\begin{bmatrix} \phi_{4,4} & \phi_{1,4} \\ \phi_{1,4} & \phi_{4,4} \end{bmatrix}$	$\begin{bmatrix} 0 & \phi_{1,4} \\ \phi_{1,4} & 0 \end{bmatrix}$	4
4-CPFSK	$\begin{bmatrix} \phi_{8,8} & \phi_{1,8} \\ \phi_{3,8} & \phi_{8,8} \end{bmatrix}$	$\begin{bmatrix} 0 & \phi_{1,8} \\ \phi_{1,8} & 0 \end{bmatrix}$	8

3. System model

A cooperative transmission model, in which a source sends information to a destination with the assistance of a relay according to cooperation protocol, was considered. The cooperation protocol manages the transmission as follows. The total transmission is divided into 2 phases: Phase 1 and Phase 2. During Phase 1, the source sends information to the relay but the destination terminal does not receive any information from the source.

The relay terminal amplifies the received noisy information in the same phase. Both the source and relay terminals send information to the destination terminal simultaneously during Phase 2. It is assumed that all of the terminals, S, R, and D, have only one antenna. Clearly, without loss of generality, the relay terminal transmits the first STC-CPM signal, indicated by the first row of the $\mathbf{c}_n$, and the source terminal transmits the second STC-CPM signal, indicated by the second row of the $\mathbf{c}_n$ in 2 symbol periods.

3.1. Amplify-and-forward mode

For the AF mode, the received signal at the relay terminal can be written as

$$r_R = \sqrt{E_{SR}} h_{SR} x_1 + n_R, \tag{5}$$

where x_1 is the transmitted signal from the source in Phase 1, $\sqrt{E_{SR}}$ is the average energy available at the relay terminal considering the possible path loss and the shadowing effects on the channel $(S \rightarrow R)$, n_R is the additive white Gaussian noise (AWGN) with zero mean and a variance of $N_0/2$ per dimension, and h_{SR} is the fading channel coefficients modeled as complex Gaussian random variables with a variance of 1/2 per dimension. In the AF mode, the relay terminal amplifies and transmits the received signal. It is assumed that the received signal is multiplied by a normalizing factor of $\sqrt{E\left\{|r_R|^2\right\}}$. Note that after the normalizing process, the average energy of the signal becomes unity. In Phase 2, the destination terminal receives normalized and corrupted (in the channel $R \rightarrow D$) versions of r_R, r_{D1}, from the relay terminal. In the same phase, the destination terminal also receives signal r_{D2} from the source. Clearly, the destination terminal receives a superposition of r_{D1} and r_{D2}, which is indicated as r_D^{AF}.

$$r_D^{AF} = \frac{\sqrt{E_{RD}} r_R}{\sqrt{E\left\{|r_R|^2\right\}}} h_{RD} + \sqrt{E_{SD}} h_{SD} x_2 + n_D, \tag{6}$$

where $\sqrt{E_{RD}}$ and $\sqrt{E_{SD}}$ are the average energies available at the destination terminal considering the possible path loss and the shadowing effects for r_{D1} and r_{D2}, respectively. h_{RD} and h_{SD} are the complex fading coefficients for the channels $R \rightarrow D$ and $S \rightarrow D$, respectively. x_2 is the transmitted signal from the source in Phase 2 and n_D is the AWGN with zero mean and a variance of $N_0/2$ per dimension.

At the receiver, first, the received signal is passed through an ideal antialiasing filter with bandwidth $f_s/2$ Hz and then sampled at a rate of f_s Hz. The sampled signal at the receiver can be written as

$$r_{D,k}^{AF} = \frac{\sqrt{E_{RD}} r_{R,k}}{\sqrt{E\left\{|r_R|^2\right\}}} h_{RD,k} + \sqrt{E_{SD}} h_{SD,k} x_{2,k} + n_{D,k}. \tag{7}$$

Since this paper does not deal with channel estimation and synchronization, the receiver is assumed to use coherent detection (CD); i.e. perfect channel state information (CSI) is available at the receiver. The receiver minimizes the metric M_{CD}^{AF} for optimum detection.

$$M_{CD}^{AF} = \sum_k \left| h_{RD,k} \exp\{j\theta_{1,k}\} + h_{SD,k} \exp\{j\theta_{2,k}\} - r_{D,k}^{AF} \right|^2 \tag{8}$$

This process can easily be realized via Viterbi decoder.

3.2. Decode-and-forward mode

In DF mode, similar to AF mode, the received signal at the relay terminal is r_R, since both the same cooperation protocol and the channel are considered. Channels $R \rightarrow D$ and $S \rightarrow D$ are also the same. The relay terminal decodes and reencodes the signal r_R in Phase 1 and transmits it during Phase 2. It is assumed that perfect CSI is available at the relay terminal. Therefore, after the filtering and sampling process, CD is also available at the relay terminal. The optimum receiver minimizes the metric

$$M_{CD}^{r} = \sum_{k} |h_{RD,k} \exp\{j\theta_{1,k}\} - r_{r,k}|^2. \tag{9}$$

The minimum metric can also be found via Viterbi decoder. In Phase 2, at the destination terminal, the received signal can be written as

$$r_D^{DF} = \sqrt{E_{RD}}\hat{x}_1 h_{RD} + \sqrt{E_{SD}} h_{SD} x_2 + n_D, \tag{10}$$

where $\hat{x}_1$ denotes the decoded and reencoded signal. From equation (10), it is shown that a normalization process is not necessary at the relay terminal since reencoded signal $\hat{x}_1$ already has the same energy as x_1. The detection process at the destination terminal is the same as in Section 3A.

4. Simulation results

The performance of the proposed system for 2-CPFSK and 4-CPFSK modulations was assessed by Monte Carlo simulations for both AF and DF modes under consideration. All of the channels were subjected to slow, flat Rayleigh fading and the receivers had perfect CSI knowledge, as described in Section 3. E_{SD} and E_{RD} were chosen as equal, which was achieved through power control. Under these conditions, the bit error rate (BER) performances of the proposed system are illustrated in Figures 1-3 for $E_{SR}/N_0 = 10\text{dB},\ 20\text{dB}$ and 30dB, respectively. For simplicity of comparison, the AF and DF modes' performance curves for both 2-CPFSK and 4-CPFSK were plotted together for the full range of E_{SR}/N_0. A reference curve is also plotted in Figure 3. This reference curve is the BER performance of 2-CPFSK with one antenna and therefore no cooperation.

For both AF and DF modes with each modulation, error floors were observed at low E_{SR}/N_0 values (especially $E_{SR}/N_0 < 25$ dB). This is clearly shown in Figures 2 and 3. These error floors were based on the relay terminal where the $S \rightarrow R$ channel conditions were poor and received signals were severely corrupted. These signals were either amplified (noise was also amplified) or decoded and reencoded wrongly by the relay terminal. This made the detection too difficult at the destination terminal. An effective solution to remove the error floor and keep the overall performance at an acceptable level is to stop communicating with this relay once its signal-to-noise ratio drops below a certain threshold. If available, a different relay could assist the source during this time.

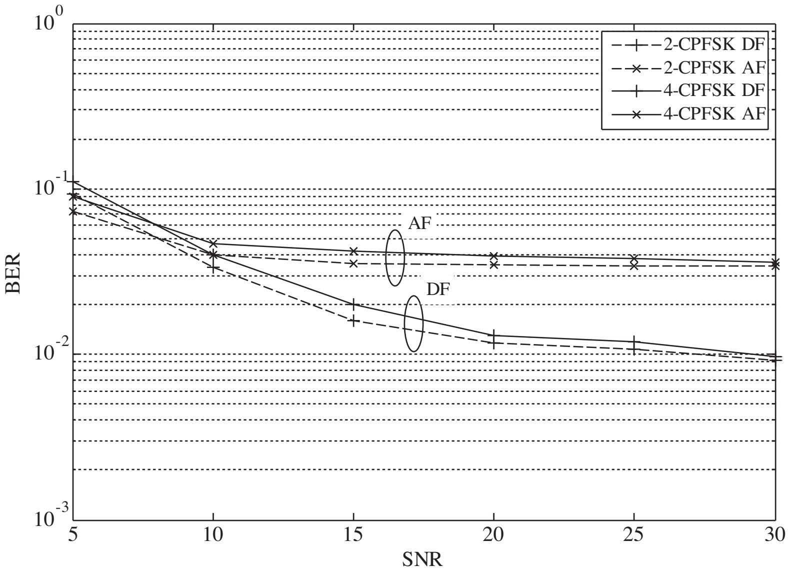

Figure 1. BER performance of proposed system for $E_{SR}/N_0 = 10$dB.

If all 3 Figures are taken into account, it can easily be seen that the AF mode starts to give a better error performance than the DF mode while E_{SD}/N_0 is increasing. On the other hand, the DF mode is seen to outperform the AF mode at high E_{SR}/N_0 regimes. This is a result of using CPM, which has a natural memory and hence allows the use of trellis decoding. The DF mode takes advantage of this natural encoding. It is apparent that if an outer encoder is used, the DF mode provides better error performance.

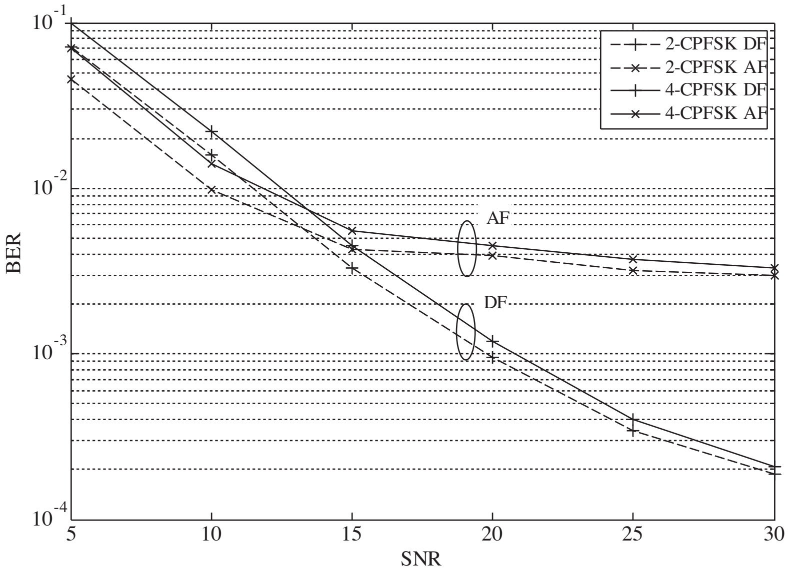

Figure 2. BER performance of proposed system for $E_{SR}/N_0 = 20$dB.

Another important point is that the 2-CPFSK modulation gives a better BER performance than the 4-CPFSK modulation, since the space-time data symbol set for the 4-CPFSK modulation has increased density and achieves a greater data rate than 2-CPFSK. Lastly, Figure 3 illustrates that for cooperative schemes, both AF an DF modes are superior to the "no cooperation" scheme for all E_{SD}/N_0 values, as was expected.

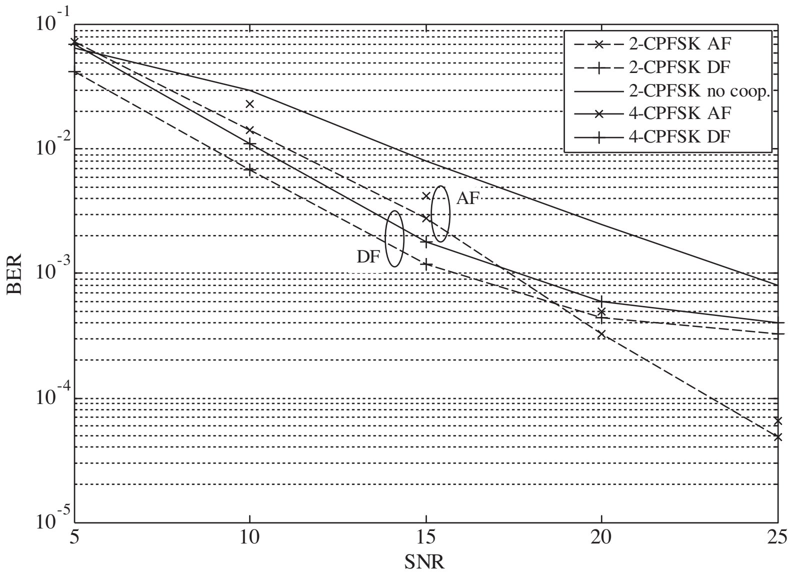

Figure 3. BER performances of proposed system for $E_{SR}/N_0 = 30\text{dB}$ and 2-CPFSK with no cooperation.

5. Conclusion

In this paper, a cooperative diversity system with space-time coded continuous phase modulation is proposed. This system utilizes the advantages of continuous phase modulation as well as diversity gain via virtual antenna array. Thus, the proposed system improves the communication capacity in wireless fading channels even when using multiple antennas is not possible. A simple and efficient differential orthogonal STC technique was used to design the STC-CPM signals. For cooperation mode, both amplify-and-forward and decode-and-forward modes were analyzed. Simulation results were also presented for various scenarios. According to the simulation results, the proposed system is very convenient for wireless fading channels.

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