Space-Time Signal Design for Time-Correlated Rayleigh Fading Channels

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***Abstract*— The existing constructions of space-time codes or modulation have mainly focused on two ideal situations: either quasi-static or rapid fading channels. In this paper, we consider the design of robust space-time modulation for time-correlated Rayleigh fading channels. We show that the space-time signals of square size achieving full diversity in quasi-static fading channels can also achieve the full diversity in time-correlated fading channels irrespectively of the time correlation matrix. Moreover, we propose a space-time signal construction method by combining orthogonal designs with sphere packings. The simulation results show that our scheme outperforms the previously existing methods. For example, we observe a coding gain of about 1.5 dB over the conventional orthogonal design, about 2 dB over the parametric code, and about 4 dB over the cyclic code under certain fading conditions.**

I. Introduction

Space-time coding for multiple antenna wireless communication systems has attracted considerable attention lately. A large number of space-time codes have been proposed based on two ideal channel conditions: either quasi-static or rapid fading [1], [2], [8]–[13]. These codes do not guarantee robust performance in correlated fading channels. In [3], [4], [5], Fitz *et. al.* presented a general design criterion for Rayleigh fading channels with space-time correlations. However, it is hard to construct space-time codes directly from the general design criterion. In [4], some hand crafted space-time trellis codes that combine the multiple trellis-coded modulation (M-TCM) with the Alamouti scheme show robust performance over space-time correlated fading channels. In [6], the general design criterion [3], [5] was further simplified assuming that the space-time correlation matrix was of full rank. In this case, the resulting design criterion becomes the same as that for rapid fading channels.

In [7], characterizing the performance of space-time codes over space-time-correlated Rayleigh fading channels was also considered. The relationship between the robustness (diversity) and the rank of the space-time correlation matrix was established, and space-time codes designed for the independent fading channel model were proposed for communication over space-time-correlated fading channels.

In this paper, we consider the design of robust space-time block codes, or more precisely *space-time signals*, for time-correlated Rayleigh fading channels. We assume that the wireless channel exhibits temporal correlation, but there is no spatial correlation between the transmit and the receive antennas. A typical physical environment related to this scenario is the downlink from a base station with multiple transmit antennas to a mobile, which is usually modeled by the Jakes fading model [14]. The time correlation is determined by the Doppler frequency shift. Let f_D be the maximum Doppler frequency shift normalized by the sampling period. For stop-and-go mobiles, the maximum Doppler frequency shift f_D changes frequently. Therefore, it is of interest to design *robust* space-time modulation that provide good performance over all time-correlated fading conditions.

First, we show in Section III that for space-time signals of square size, the signal design problem for time-correlated channels can be reduced to the design problem for quasi-static fading channels, independently of the time correlation matrix. Then, in Section IV, we propose a class of space-time signals from orthogonal designs with sphere packings. The simulation results show that our scheme outperforms the previously existing methods.

II. Channel Model and Background

We consider a wireless communication system with M transmit antennas and N receive antennas. We use space-time signals to transmit binary information. Each space-time signal can be expressed as a $T \times M$ matrix

$$C = \begin{bmatrix} c_1^1 & c_1^2 & \cdots & c_1^M \\ c_2^1 & c_2^2 & \cdots & c_2^M \\ \vdots & \vdots & \ddots & \vdots \\ c_T^1 & c_T^2 & \cdots & c_T^M \end{bmatrix}, \qquad (1)$$

with the energy constraint $E||C||_F^2 = MT$ where $||C||_F$ is the Frobenius norm of C, and E stands for the expectation.

The received signal y_t^j at receive antenna j at time t is

$$y_t^j = \sqrt{\frac{\rho}{M}} \sum_{i=1}^{M} c_t^i h_{i,j}(t) + z_t^j, \quad t = 1, 2, \cdots, T, \qquad (2)$$

where z_t^j is the AWGN noise with zero-mean and unit variance, and $h_{i,j}(t)$ is the channel coefficient from transmit antenna i to receive antenna j at time t which is known at the receiver. The channel coefficients are modeled as zero-mean complex Gaussian random variables with variance $1/2$ per dimension. We assume that the channel fading has only temporal correlation, i.e., the channel coefficients $h_{i,j}(t)$ are independent for different index (i, j) and dependent in the time direction. The factor $\sqrt{\rho/M}$ in (2) ensures that ρ is the SNR at each receive antenna, and independent of M.

[1] This work was supported in part by CTA-ARL DAAD 190120011.

0-7803-7802-4/03/$17.00 © 2003 IEEE

The received signal (2) can be rewritten in vector form as

$$\mathbf{Y} = \sqrt{\frac{\rho}{M}}\mathbf{D}\mathbf{H} + \mathbf{Z}, \tag{3}$$

where $\mathbf{D}$ is an $NT \times MNT$ matrix formed from the space-time signal matrix C in (1) as follows [3], [5]:

$$\mathbf{D} = \begin{bmatrix} D_1 & D_2 & \cdots & D_M & 0 & 0 & \cdots & 0 & \cdots & 0 & 0 & \cdots & 0 \\ 0 & 0 & \cdots & 0 & D_1 & D_2 & \cdots & D_M & \cdots & 0 & 0 & \cdots & 0 \\ & \vdots & & & & \vdots & & & \ddots & & \vdots & & \\ 0 & 0 & \cdots & 0 & 0 & 0 & \cdots & 0 & \cdots & D_1 & D_2 & \cdots & D_M \end{bmatrix}, \tag{4}$$

in which $D_i = diag(c_1^i, c_2^i, \cdots, c_T^i), \quad i = 1, 2, \cdots, M$. The channel vector $\mathbf{H}$ of size $MNT \times 1$ is formatted as

$$\mathbf{H} = [\mathbf{h}_{1,1}^{\mathcal{T}} \cdots \mathbf{h}_{M,1}^{\mathcal{T}} \ \mathbf{h}_{1,2}^{\mathcal{T}} \cdots \mathbf{h}_{M,2}^{\mathcal{T}} \cdots \mathbf{h}_{1,N}^{\mathcal{T}} \cdots \mathbf{h}_{M,N}^{\mathcal{T}}]^{\mathcal{T}}, \tag{5}$$

where $\mathbf{h}_{i,j} = [h_{i,j}(1) \ h_{i,j}(2) \cdots h_{i,j}(T)]^{\mathcal{T}}$. The received signal vector $\mathbf{Y}$ of size $NT \times 1$ is $\mathbf{Y} = \left[y_1^1 \cdots y_T^1 \ y_1^2 \cdots y_T^2 \cdots y_1^N \cdots y_T^N\right]^{\mathcal{T}}$, and the noise vector $\mathbf{Z}$ has a same form as $\mathbf{Y}$.

Suppose that $\mathbf{D}$ and $\tilde{\mathbf{D}}$ are two different matrices related to two different space-time signals C and $\tilde{C}$, respectively. The pairwise error probability between $\mathbf{D}$ and $\tilde{\mathbf{D}}$ can be upper bounded as [3], [5]

$$P(\mathbf{D} \to \tilde{\mathbf{D}}) \leq \binom{2K-1}{K} \left(\prod_{i=1}^{K} \gamma_i\right)^{-1} \left(\frac{\rho}{M}\right)^{-K}, \tag{6}$$

where K is the rank of $(\mathbf{D}-\tilde{\mathbf{D}})\mathbf{R}(\mathbf{D}-\tilde{\mathbf{D}})^{\mathcal{H}}$, $\gamma_1, \gamma_2, \cdots, \gamma_K$ are the non-zero eigenvalues of $(\mathbf{D}-\tilde{\mathbf{D}})\mathbf{R}(\mathbf{D}-\tilde{\mathbf{D}})^{\mathcal{H}}$, and $\mathbf{R} = E\{\mathbf{H}\mathbf{H}^{\mathcal{H}}\}$ is the correlation matrix of $\mathbf{H}$. The superscript $\mathcal{H}$ stands for the complex conjugate and transpose.

Based on the upper bound on the pairwise error probability in (6), a general code design criterion has been proposed in [3] and [5]. This criterion is consistent with the well-known criteria [2] for two ideal cases: the quasi-static and the rapid fading channel models, i.e.,

- *For quasi-static fading channels:* The minimum rank of

$$\Delta \triangleq (C - \tilde{C})(C - \tilde{C})^{\mathcal{H}} \tag{7}$$

  over all pairs of distinct signals C and $\tilde{C}$ should be maximized. If Δ is of full rank for distinct signals C and $\tilde{C}$, then the *diversity product* [10], [11] is given by

$$\zeta_{static} = \frac{1}{2\sqrt{M}} \min_{C \neq \tilde{C}} |\det(\Delta)|^{1/(2T)}, \tag{8}$$

  which is related to the coding advantage and should also be maximized.
- *For rapid fading channels:* The minimum number of non-zero rows of $C - \tilde{C}$ should be as large as possible for any pair of distinct signals C and $\tilde{C}$. If for any pair of distinct signals C and $\tilde{C}$, there is no zero row in $C - \tilde{C}$, then the diversity product, given by

$$\zeta_{rapid} = \frac{1}{2\sqrt{M}} \min_{C \neq \tilde{C}} \left(\prod_{t=1}^{T} ||\mathbf{c}_t - \tilde{\mathbf{c}}_t||_F^2\right)^{1/(2T)}, \tag{9}$$

  should be maximized. In (9), $\mathbf{c}_t$ and $\tilde{\mathbf{c}}_t$ are the t-th rows of C and $\tilde{C}$, respectively.

III. Design Criteria for Time-Correlated Fading Channels

In this section, we derive the design criteria assuming only time correlation. In this case, the channel correlation matrix $\mathbf{R}$ becomes

$$\mathbf{R} = diag(R_{1,1}, \cdots, R_{M,1}, \ R_{1,2}, \cdots, R_{M,2}, \cdots \\ \cdots, \ R_{1,N}, \cdots, R_{M,N}),$$

where $R_{i,j} = E\left(\mathbf{h}_{i,j}\mathbf{h}_{i,j}^{\mathcal{H}}\right)$ is the time correlation matrix of the channel coefficients from transmit antenna i to receive antenna j. We may further assume that all of the time correlation matrices $R_{i,j}$ are the same, which is true for the Jakes fading model [14]. Denote $R \triangleq R_{i,j}$, then we have

$$\mathbf{R} = I_{MN} \otimes R, \tag{10}$$

where $\otimes$ denotes the tensor product and I_{MN} is the identity matrix of size $MN \times MN$. Then, we have

$$\begin{aligned}(\mathbf{D}-\tilde{\mathbf{D}})\mathbf{R}(\mathbf{D}-\tilde{\mathbf{D}})^{\mathcal{H}} &= I_N \otimes \left\{\left[(C-\tilde{C})(C-\tilde{C})^{\mathcal{H}}\right] \circ R\right\} \\ &= I_N \otimes \{\Delta \circ R\},\end{aligned} \tag{11}$$

where $\circ$ denotes the Hadamard product[2]. Substituting (11) into (6), the pairwise error probability between C and $\tilde{C}$ can be upper bounded as

$$P(C \to \tilde{C}) \leq \binom{2rN-1}{rN} \left(\prod_{i=1}^{r} \lambda_i\right)^{-N} \left(\frac{\rho}{M}\right)^{-rN}, \tag{12}$$

where r is the rank of $\Delta \circ R$, and $\lambda_1, \lambda_2, \cdots, \lambda_r$ are the non-zero eigenvalues of $\Delta \circ R$. Clearly, the minimum rank of $\Delta \circ R$ over all pairs of distinct signals C and $\tilde{C}$ should be as large as possible.

If the minimum rank of $\Delta \circ R$ is ν for any pair of distinct signals C and $\tilde{C}$, we say that the set of space-time signals achieves a diversity of νN. For fixed time duration T, the number of transmit antennas M, and time correlation matrix R, the *maximum achievable diversity* or *full diversity* is defined as the maximum diversity level that can be achieved by space-time signals of size $T \times M$. For example, for quasi-static fading channels, R is an all one matrix of size $T \times T$. In this case, the maximum achievable diversity is $\min(M, T)N$. For rapid fading channels, $R = I_T$, then the maximum achievable diversity is TN.

Assume that the time correlation matrix R is of rank Γ $(1 \leq \Gamma \leq T)$. According to a rank inequality on Hadamard products ([16] p.307), we have

$$rank(\Delta \circ R) \leq rank(\Delta)rank(R).$$

[2]Assume $A = (a_{i,j})_{1 \leq i \leq m, 1 \leq j \leq n}$ and $B = (b_{i,j})_{1 \leq i \leq m, 1 \leq j \leq n}$, the *Hadamard product* of A and B is $A \circ B = (a_{i,j}b_{i,j})_{1 \leq i \leq m, 1 \leq j \leq n}$.

The rank of Δ cannot be greater than $\min(M,T)$. Therefore, we can state the following theorem.

Theorem 1: *For space-time signals of size $T\times M$ operating in time-correlated fading environment, the maximum achievable diversity is upper bounded by $\min(M\Gamma,T)N$, where Γ is the rank of the time-correlation matrix R.*

For space-time signals of square size, i.e., $T=M$, the maximum achievable diversity cannot be greater than MN. Now we will show that this upper bound can be achieved for any time-correlated fading channel. Note that both Δ and R are non-negative definite, and all of the diagonal entries of R are non-zero. Therefore, we can apply Schur's theorem on Hadamard products ([16], p.309): if Δ is positive definite (i.e. of full rank), then $\Delta\circ R$ is also positive definite (i.e. of full rank). Thus, we arrive at the following theorem.

Theorem 2: *If a set of space-time signals of size $M\times M$ achieves full diversity (MN) for quasi-static fading channels, then it also achieves full diversity (MN) for any time-correlated fading channel, independent of the time correlation matrix R.*

The result in Theorem 2 is very interesting. It is well known that if a set of space-time signals of square size achieves full diversity for quasi-static fading channels, then it also achieves full diversity for rapid fading channels. However, it is not obvious that it can also achieve full diversity for any time-correlated fading channels. It follows from Theorem 2 that all of the space-time signals of square size designed for quasi-static fading channels can also be used to achieve full diversity in any time-correlated fading channel.

Theorem 2 does not hold for non-square signals (i.e. $T\neq M$). However, if the time correlation matrix is of full rank, we can establish another result. We observe that the diagonal entries of Δ are $||\mathbf{c}_t-\tilde{\mathbf{c}}_t||_F^2$, $t=1,2,\cdots,T$, where $\mathbf{c}_t$ and $\tilde{\mathbf{c}}_t$ are the t-th rows of C and $\tilde{C}$, respectively. Thus, we can apply Schur's theorem again: if all of $||\mathbf{c}_t-\tilde{\mathbf{c}}_t||_F^2$'s are non-zero for any pair of distinct signals C and $\tilde{C}$, then $\Delta\circ R$ is of full rank whenever the time correlation matrix R is of full rank. This implies that the maximum achievable diversity is TN for any full rank time correlation matrix R. This result is summarized in the following theorem.

Theorem 3: *If a set of space-time signals of size $T\times M$, $T\geq M$ achieves full diversity (TN) for rapid fading channels, then it also achieves full diversity (TN) for any time-correlated fading channel, provided that the time correlation matrix R is of full rank.*

In the following, we consider the coding advantage of the space-time signals for time-correlated fading channels. Assume that $\Delta\circ R$ is of full rank for any pair of distinct signals C and $\tilde{C}$, then the *diversity product* can be generalized as

$$\zeta_R=\frac{1}{2\sqrt{M}}\min_{C\neq\tilde{C}}|\det(\Delta\circ R)|^{\frac{1}{2T}}. \quad (13)$$

It is easy to see that for quasi-static fading channels, the diversity product (13) reduces to (8); and for rapid fading channels, the diversity product (13) becomes (9). Furthermore, it can be shown that the diversity product ζ_R is upper bounded by ζ_{rapid} and lower bounded by ζ_{static} as follows.

Theorem 4: *If a set of space-time signals has L elements, and $\Delta\circ R$ is of full rank for any pair of distinct signals C and $\tilde{C}$, then ζ_R, the diversity product of these signals for a fading channel with time correlation R, satisfies*

$$\max\left\{\zeta_{static},\,|\det(R)|^{\frac{1}{2T}}\zeta_{rapid}\right\}\leq\zeta_R\leq\zeta_{rapid}\leq\sqrt{\frac{L}{2(L-1)}}. \quad (14)$$

The details of the proof are omitted for brevity. The first inequality in (14) follows from Oppenheim's inequality [16], the second inequality follows from Hadamard's inequality [16], and the third inequality can be established using methods similar to [12]. From Theorem 4, we can see that if $\zeta_{static}=\zeta_{rapid}$, then ζ_R is fixed, no matter what the time correlation matrix R is.

IV. Orthogonal Designs with Sphere Packings

In this section, we consider the construction of space-time signals of square size. In this case, according to Theorem 2 and 4, an efficient way to design robust space-time signals is to make ζ_{static} as large as possible. Thus, the problem of designing robust space-time signals for time-correlated fading channels is reduced to that of designing space-time signals for quasi-static fading channels. It follows that the abundant classes of space-time signals designed for quasi-static fading channels, for example, cyclic codes [10], codes from orthogonal designs [8], [9], parametric codes [12], may also be used for time-correlated fading channels.

We now construct space-time signals from orthogonal designs with sphere packings for $M=2^k$, $k=1,2,3,\cdots$, transmit antennas. In case of the conventional space-time signal design methods, the symbols are chosen independently from PSK or QAM constellations. The basic idea of the new scheme is that we design these symbols jointly with sphere packings to further increase the coding advantage. Notice that, for quasi-static channels, the independent choices of the symbols in orthogonal designs allow for fast maximum likelihood (ML) decoding. However, in time-correlated fading environment, the joint ML decoding is inevitable, so the proposed space-time modulation method does not cause any extra increase in decoding complexity.

Orthogonal designs have a long history in mathematics. Recently, orthogonal designs have attracted considerable attention in space-time coding due to their special structure [8], [9], [13]. A recursive expression of orthogonal designs was given in [13] as follows. Let $G_1(x_1)=x_1I_1$, and

$$G_{2^k}(x_1,\cdots,x_{k+1})=\begin{bmatrix}G_{2^{k-1}}(x_1,\cdots,x_k) & x_{k+1}I_{2^{k-1}}\\ -x_{k+1}^*I_{2^{k-1}} & G_{2^{k-1}}^{\mathcal{H}}(x_1,\cdots,x_k)\end{bmatrix}$$

for $k = 1, 2, 3, \cdots$. Then, $G_{2^k}(x_1, x_2, \cdots, x_{k+1})$ is an orthogonal design with complex variables $x_1, x_2, \cdots, x_{k+1}$ for 2^k transmit antennas. The symbol rate of G_{2^k} is $(k+1)/2^k$, which is the maximum rate for orthogonal designs of square size ([13] and the references therein).

For $M = 2^k, k = 1, 2, 3, \cdots$, transmit antennas, a set of space-time signals can be constructed directly from the orthogonal design G_{2^k} as $C = \sqrt{2^k/(k+1)}\, G_{2^k}(x_1, x_2, \cdots, x_{k+1})$ with some specific choices of $x_1, x_2, \cdots, x_{k+1}$. The factor $\sqrt{2^k/(k+1)}$ is a normalization for the energy constraint.

For two distinct signals C and $\tilde{C}$ with varialbes x_i and $\tilde{x}_i$ respectively, we have

$$(C - \tilde{C})(C - \tilde{C})^{\mathcal{H}} = \frac{2^k}{k+1} \sum_{i=1}^{k+1} |x_i - \tilde{x}_i|^2 I_{2^k}.$$

Since all of the diagonal entries of the auto-correlation matrix R are unity, the diversity product (13) becomes

$$\zeta_R = \frac{1}{2\sqrt{k+1}} \min_{(x_1, \cdots x_{k+1}) \neq (\tilde{x}_1, \cdots \tilde{x}_{k+1})} \left(\sum_{i=1}^{k+1} |x_i - \tilde{x}_i|^2 \right)^{1/2}, \tag{15}$$

which is not affected by a specific time correlation matrix R. In fact, (15) can also be derived from Theorem 4, since $\zeta_{static} = \zeta_{rapid}$ in this case.

Having ensured that the space-time signals achieve full diversity, the next step is to maximize the coding advantage. From (16), we can see that the diversity product is determined by the minimum Euclidean distance of the vectors $\{(x_1, x_2, \cdots, x_{k+1})\}$. Therefore, sphere packings in $\mathrm{R}^{2(k+1)}$ [15] can be used to maximize the coding advantage. In the sequel, we will describe a particular example for 2 transmit antennas, but the proposed method can be easily generalized to other $M = 2^k$ transmit antennas.

For 2 transmit antennas, the 2×2 orthogonal design is

$$G_2(x_1, x_2) = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}, \tag{16}$$

which was first used by Alamouti in space-time coding [8]. Later in [11], a similar structure with constraint $|x_1|^2 + |x_2|^2 = 1$ was used to build 2×2 unitary matrices, a so-called Hamiltonian constellation for differential modulation. This constraint is not necessary for the space-time signals for time-correlated fading channels.

D_4 is a sphere packing with the best known minimum Euclidean distance in R^4 [15]. We combine the orthogonal design G_2 and D_4 to construct space-time signals. Assume that $S = \{[s_{l,1}\, s_{l,2}\, s_{l,3}\, s_{l,4}] \in \mathrm{R}^4 : 0 \le l \le L-1\}$ is a set of L points from D_4. Let $C_l = G_2(s_{l,1} + js_{l,2}, s_{l,3} + js_{l,4})$, $l = 0, 1, \cdots, L-1$, then $\{C_l : 0 \le l \le L-1\}$ is a set of space-time signals whose diversity product is determined by the minimum Euclidean distance of S. We list the diversity products of the orthogonal designs with sphere packings (abbreviated as Orth. S.P.) in Table 1, and compare them with those of cyclic codes [10], parametric codes [12], and orthogonal designs with PSK. Note that for $L = 4$, there are optimal space-time signals [12], in a sense that the diversity product achieves the upper bound $\sqrt{2/3}$, and they actually came out from orthogonal designs with sphere packings.

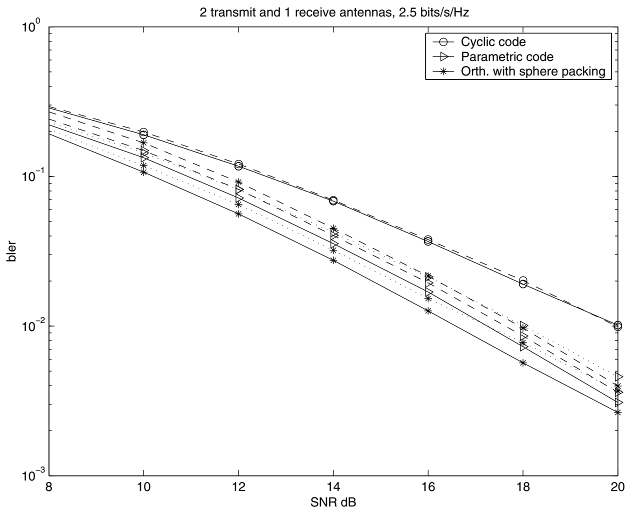
Table 1: Comparison of diversity product for 2 transmit antennas.

Size L	Up. Bound $\sqrt{\frac{L}{2(L-1)}}$	Diversity product ζ_{static}	Diversity product ζ_{rapid}	Comments
4	0.8165	0.8165	0.8165	Orth. S.P.
8	0.7559	0.5946	0.5946	Cyclic code
		0.7071	0.7071	Para. code
		0.7071	0.7071	Orth. S.P.
16	0.7303	0.3827	0.3827	Cyclic code
		0.5946	0.5946	Para. code
		0.5000	0.5000	Orth. QPSK
		0.5535	0.5535	Orth. S.P.
32	0.7184	0.2494	0.2494	Cyclic code
		0.3827	0.3827	Para. code
		0.4658	0.4658	Orth. S.P.
64	0.7127	0.1985	0.1985	Cyclic code
		0.3070	0.3778	Para. code
		0.2706	0.2706	Orth. 8PSK
		0.3860	0.3860	Orth. S.P.
128	0.7099	0.1498	0.1498	Cyclic code
		0.2606	0.3261	Para. code
		0.3226	0.3226	Orth. S.P.

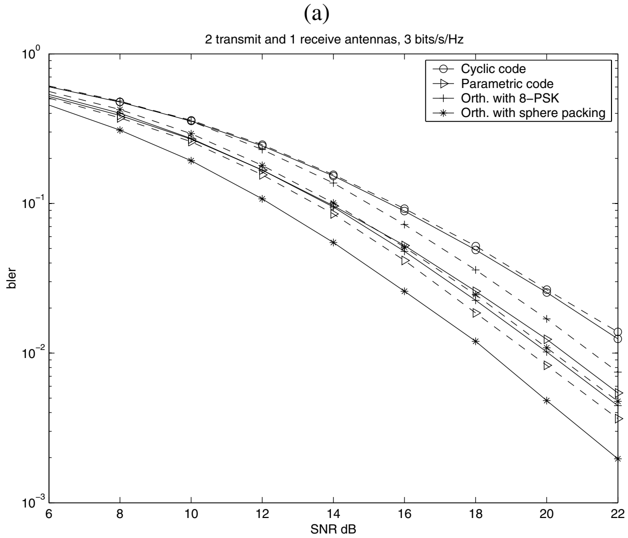
The simulation results are given for two transmit and one receive antennas under three different channel conditions: quasi-static, time-correlated ($f_D = 0.1$) and rapid fading. We used ML decoding and presented block error rate (bler) versus average signal to noise ratio (SNR) curves.

Fig. 1(a) provides the simulation results for the $L = 32$ case, i.e., 2.5 b/s/Hz. The curves demonstrate that the proposed method outperforms the other approaches under almost all channel conditions. For the quasi-static fading channel model, the orthogonal design with sphere packing has an improvement of about 0.5 dB over the parametric code, and about 3.5 dB over the cyclic code at a bler of 10^{-2}. In time-correlated fading environment, the improvement is approximately 0.5 dB and 3 dB over the the parametric code and the cyclic code, respectively. In the rapid fading case, the performance of the proposed scheme is about 2 dB better than that of the cyclic code, but 0.25 dB worse than that of the parametric code at a bler of 10^{-2}. Note that the performance of the cyclic code is almost the same for the three channel conditions.

The block error rate curves for the $L = 64$ scenario, i.e., 3 b/s/Hz, are shown in Fig. 1(b). We compare our method with other three schemes: the conventional orthogonal design with 8PSK, the parametric code, and the cyclic code for two channel models: quasi-static and rapid fading. The curves for the time-correlated channel model are omitted for clarity. It can be observed that the orthogonal de-



(a)



(b)

Fig. 1. Block error rate performance of cyclic codes '○', parametric code '▷', orth. with PSK '+', and orth. with sphere packing '∗' at three channel conditions: quasi-static (solid line), correlated (dotted line), and rapid (dashed line). (a) $L = 32$ and (b) $L = 64$.

sign with sphere packing has superior performance over the other schemes. For the quasi-static fading channel model, the block error rate curves show a performance improvement of about 1.5 dB over the conventional orthogonal design with 8PSK, 2 dB over the parametric code, and 4 dB over the cyclic code at a bler of 10^{-2}. For the rapid fading channel model, the performance of the new scheme is about 1 dB better than that of the conventional orthogonal design with 8PSK, and about 2 dB better than that of the cyclic code, but 0.5 dB worse than that of the parametric code.

The results of Theorem 4 indicate that the performance of the space-time modulation methods is the worst in quasi-static fading environment. Yet, from the simulation results in Fig. 1(a) and (b), it seems that for space-time block codes from orthogonal designs, the worst channel is the rapid fading channel, not the quasi-static fading channel, in contrast to the observations from space-time trellis codes [2], [4], [6].

V. Conclusion

In this paper, we focused on the problem of designing space-time modulation for time-correlated Rayleigh fading channels. We showed that the design problem of square space-time signals can be simplified to the signal design problem for the quasi-static fading channel model, independently of the time correlation matrix. This result implies that various classes of space-time signals of square size designed for quasi-static fading channels may also be used for time-correlated fading channels. We also proposed a class of space-time signals constructed from orthogonal designs with sphere packing. The simulation results show that our scheme outperforms the previously existing methods. For example, we observed a coding gain of about 1.5 dB over the conventional orthogonal design, about 2 dB over the parametric code, and about 4 dB over the cyclic code for two transmit and one receive antennas under certain fading conditions.

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