Double Differential Space-Time Block Coding for Time-Selective Fading Channels

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Abstract— **Most existing space-time coding schemes assume time-invariant fading channels and offer antenna diversity gains relying on accurate channel estimates at the receiver. Based on a diagonal unitary matrix group, a novel double differential space-time block coding approach is derived in this paper for time-selective fading channels. Without estimating the channels at the receiver, information symbols are recovered with antenna diversity gains regardless of frequency offsets. The resulting transceiver has very low complexity and is applicable to an arbitrary number of transmit and receive antennas. Approximately optimal space-time codes are also designed to minimize bit error rate. System performance is evaluated both analytically and with simulations.**

I. Introduction

The rapidly growing demand for reliable high data rate transmissions over fading channels has stimulated much interest in space-time (ST) coding. The effectiveness of most ST coding schemes relies on accurate channel estimates at the receiver, which are either acquired through training sessions [4], or, by employing blind estimation algorithms [6]. Since multiple channels are involved with multiple-antenna links, channel estimation implicitly assumes that the underlying channels remain invariant for sufficiently long time - an assumption that may not be satisfied in rapidly fading mobile environments. Single differential space-time (SDST) block coding schemes were proposed in [2, 10, 1] to achieve diversity gains without channel state information (CSI). By obviating channel estimation, SDST coding schemes allow for slowly changing channels that have to remain invariant within two consecutive blocks. Their performance however, degrades when channel time-selectivity is present.

Generalizing single-antenna double differential coding [8] to the ST context, we develop a novel double differential space-time (DDST) coding scheme along with pertinent modulator design criteria and error bounds. We consider frequency-offset induced time-selective channels which are allowed to vary as fast as one symbol duration. Based on diagonal unitary group code matrices, properly designed encoding at the transmitter enables information symbol recovery at the receiver with diversity gains and without channel knowledge. The resulting algorithm has low complexity and can be applied to an arbitrary number of transmit and receive antennas. We also design an optimal ST code matrix that approximately minimizes bit error rate (BER). In addition to simplicity, our scheme enables 3 dB performance improvement when estimates of frequency offsets are available, while 6 dB performance gain is possible with perfect CSI at the receiver. Thorough simulations illustrate robust performance of our design even in the presence of time-selectivity.

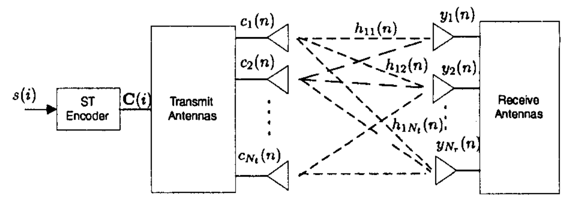

Fig. 1. Baseband system model

The rest of this paper is organized as follows. In Section II we describe the time-selective channel and derive our data model. The double differential ST coding/decoding is developed in Section III. In Section IV, the approximately optimal code matrix designs are derived by minimizing the bit error rate and some examples are given. Supporting simulations are presented in Section V.

II. System Modeling

Fig. 1 depicts a wireless communication system equipped with N_t transmit antennas and N_r receive antennas. At the transmitter, each information symbol $s(i)$ belonging to a finite alphabet set $\mathcal{A}$ is encoded by the space-time encoder which uniquely maps $s(i)$ to the following

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$N \times N_t$ code matrix:

$$\mathbf{C}(i) := \begin{bmatrix} c_1(iN) & \cdots & c_{N_t}(iN) \\ \vdots & \ddots & \vdots \\ c_1(iN+N-1) & \cdots & c_{N_t}(iN+N-1) \end{bmatrix} \quad (1)$$

The rows of $\mathbf{C}(i)$ are generated in N successive time intervals of duration T. Each of the N_t elements in a given row is forwarded to one of the N_t transmit antennas simultaneously. We emphasize that N encoded symbols from each antenna are transmitted for every information symbol. Thus, the overall transmission rate is $R = (1/N)\log_2|\mathcal{A}|$ bits/sec/Hz, where $|\mathcal{A}|$ denotes the cardinality of $\mathcal{A}$.

It is assumed that channel delay spread is small compared to T but channel coherence time is comparable to T. Under these assumptions, channels are frequency-flat but time-selective. In mobile radio channels, time-selectivity is mainly caused by frequency offsets. Two main sources of frequency offsets exist. One is Doppler shift caused by the relative motion between the transmitter and the receiver. The other is the carrier frequency mismatch between transmit- and receive-oscillators that arises due to drifts from their nominal frequencies. Denote by $f_k^{(o)}$ and $f_k^{(d)}$ the carrier frequency offset and Doppler shift at the kth receive-antenna. The channel from the mth transmit-antenna to the kth receive-antenna is modeled as:

$$h_{mk}(n) = \tilde{h}_{mk} e^{j2\pi f_k n}, \quad (2)$$

where: $f_k := (f_k^{(o)} + f_k^{(d)}) \cdot T$ is the normalized frequency offsets; and $\tilde{h}_{mk}$ captures multipath fading effects. To obtain (2), it is also assumed that the Doppler shift $f_k^{(d)}$ (being independent of m) is common to all transmit antennas. This assumption is valid when the receiver is not surrounded by any local scatterers so that multipath components originate far away and arrive at the receiver at a common angle.

Under the channel model (2), the received signal at the kth receive-antenna $y_k(n)$, $k = 1, \ldots, N_r$, is given by:

$$y_k(n) = A \cdot e^{j2\pi f_k n} \sum_{m=1}^{N_t} \tilde{h}_{mk} c_m(n) + v_k(n), \quad (3)$$

where A is the transmit amplitude and $v_k(n)$ denotes AWGN with two-sided power spectral density $N_0/2$. Define the received data block $\mathbf{y}_k(i) := [y_k(iN) \cdots y_k(iN + N-1)]^T$, (3) is cast into a matrix/vector form as:

$$\mathbf{y}_k(i) = A \cdot e^{j2\pi f_k iN} \mathbf{D}_k \mathbf{C}(i) \mathbf{h}_k + \boldsymbol{v}_k(i) \quad (4)$$

where $\mathbf{D}_k := \mathrm{diag}(1, e^{j2\pi f_k}, \cdots, e^{j2\pi f_k(N-1)})$, $\boldsymbol{v}_k(i) := [v_k(iN) \cdots v_k(iN+N-1)]^T$ and $\mathbf{h}_k := [\tilde{h}_{1k} \cdots \tilde{h}_{N_t k}]^T$.

Data model (4) incorporates the Doppler effects, the carrier frequency offsets through f_k's and channel fading effects through $\mathbf{h}_k$'s. If one can estimate f_k's accurately at the receiver, the frequency errors can be compensated for and then the information symbols can be recovered with or without knowledge of $\mathbf{h}_k$'s as discussed in [2]. However, targeting *non-coherent* reception, our goal herein is to design ST encoders and decoders so that the information data can be retrieved with diversity gains at the receiver regardless of the channel variations modeled by (4).

III. Double Differential ST Coding

Given the received data block $\mathbf{y}_k(i)$, we design in this section the code matrix $\mathbf{C}(i)$ to recover the information symbols without requiring knowledge of f_k's and $\mathbf{h}_k$'s. Motivated by the conventional single-antenna double differential coding approaches of [8], where three consecutive samples are considered to remove the unknown frequency and phase errors, our basic idea is to process three consecutive received data blocks, namely, $\mathbf{y}_k(i-2)$, $\mathbf{y}_k(i-1)$ and $\mathbf{y}_k(i)$ from each receive antenna, to recover the data symbol $s(i)$ by exploiting a judiciously designed encoding relationship among $\mathbf{C}(i-2)$, $\mathbf{C}(i-1)$ and $\mathbf{C}(i)$.

We will focus on code matrices $\mathbf{C}(i)$ that satisfy:

$$\mathbf{C}^{\mathcal{H}}(i)\mathbf{C}(i) = N\mathbf{I}_{N_t}, \quad \forall\, i \geq 0, \quad (5)$$

where $\mathcal{H}$ stands for Hermitian transpose and $\mathbf{I}_{N_t}$ denotes the $N_t \times N_t$ identity matrix. Checking the dimensionality of $\mathbf{C}(i)$, we infer that for (5) to hold we have to select $N \geq N_t$. Since $R = (1/N)\log_2|\mathcal{A}|$, we choose $N = N_t$ in order to maximize the transmission rate. For time-invariant channels, two consecutive code matrices in the SDST modulation of [1, 2] are related through a unitary matrix group. The group structure leads to uniform performance over all code matrices and reduces complexity at the receiver, both of which are also desirable in our design. Similar to [1, 2], our approach to double differential space-time (DDST) coding is also based on a unitary matrix group. Unlike [1, 2] however, we will find it necessary to restrict our unitary matrices to be diagonal. Let $\mathcal{G}$ be any finite group of $N \times N$ unitary and *diagonal* matrices ($\mathbf{F}^{\mathcal{H}}\mathbf{F} = \mathbf{F}\mathbf{F}^{\mathcal{H}} = \mathbf{I}_N$, and $\mathbf{F}$ is diagonal, $\forall\, \mathbf{F} \in \mathcal{G}$). We design our ST code matrices to satisfy the following recursion:

$$\mathbf{C}(i) = \mathbf{G}(i)\mathbf{C}(i-1), \quad \forall\, i \geq 1, \quad (6)$$

where the so-termed generating matrix $\mathbf{G}(i)$ obeys also a recursion:

$$\mathbf{G}(i) = \mathbf{F}(i)\mathbf{G}(i-1), \ \forall\, i \geq 2; \tag{7a}$$

$$\mathbf{G}(1) = \mathbf{I}_N. \tag{7b}$$

In the second recursion, the matrix $\mathbf{F}(i) \in \mathcal{G}$ conveys our information symbols and is chosen to correspond one-to-one with $s(i)$. Hence, knowing or detecting $\mathbf{F}(i)$ at the receiver, determines uniquely $s(i)$. This implies that we should design the mapping between $\mathcal{A}$ and $\mathcal{G}$ to be one-to-one and as a first step in this direction, we choose equal cardinalities: $|\mathcal{G}| = |\mathcal{A}| := M$. Next, we define $\mathcal{G} := \{\mathbf{F}_0, \ldots, \mathbf{F}_{M-1}\}$ and $\mathcal{A} := \{s_0, \ldots, s_{M-1}\}$, and establish, without loss of generality, the following one-to-one (ordered) mapping between elements of $\mathcal{A}$ and $\mathcal{G}$:

$$s_i \longleftrightarrow \mathbf{F}_i, \ \ \forall\, i = 0, 1, \ldots, M-1. \tag{8}$$

Due to the group structure of $\mathbf{F}(i)$ and the fact that $\mathbf{G}(1) = \mathbf{I}_N \in \mathcal{G}$, we deduce from (7) that: $\forall\, i \geq 1$, $\mathbf{G}(i) \in \mathcal{G}$. Thus, (6) is the same as the recursion in [2, Eq. (12)], which enables *single* differential decoding of the information data when the underlying channels exhibit flat fading. Note also that choosing $\mathbf{C}(0)$ to satisfy (5), implies (by induction) that $\mathbf{C}(i)$ generated by (6) will satisfy (5) as well.

Using our ST encoder generated by (6), (7) and neglecting the additive noise, three consecutive received blocks of samples at the kth receive antenna can be written as [c.f. (4), (6), (7)]:

$$\mathbf{y}_k(i-2) = P^{i-2} \cdot \mathbf{D}_k \mathbf{C}(i-2)\mathbf{h}_k, \tag{9a}$$

$$\mathbf{y}_k(i-1) = P^{i-1}\mathbf{G}(i-1) \cdot \mathbf{D}_k \mathbf{C}(i-2)\mathbf{h}_k, \tag{9b}$$

$$\mathbf{y}_k(i) = P^{i}\, \mathbf{F}(i)\mathbf{G}^2(i-1) \cdot \mathbf{D}_k \mathbf{C}(i-2)\mathbf{h}_k, \tag{9c}$$

where: $P := A \cdot \exp(j2\pi f_k N)$; and in establishing (9b) and (9c) we exploited the fact that $\mathbf{D}_k$, $\mathbf{G}(i-1)$ and $\mathbf{F}(i)$ are all diagonal matrices so that they commute.

Using (9a), (9b) and (9c), it is shown in [7] that $s(n)$ or (equivalently) $\mathbf{F}(i)$ can be decoded from $\mathbf{y}_k(i)$, $\mathbf{y}_k(i-1)$ and $\mathbf{y}_k(i)$ by selecting:

$$\hat{\mathbf{F}}(i) = \arg\max_{\mathbf{F}\in\mathcal{G}} \sum_{k=1}^{N_r} \mathrm{ReTr}\{\mathrm{diag}[\mathbf{Y}_{1k}(i) \cdot \mathbf{Y}_{1k}^{*}(i-1)] \times \mathrm{diag}^{-1}[\mathbf{Y}_{0k}(i) + 2\,\mathbf{Y}_{0k}(i-1) + \mathbf{Y}_{0k}(i-2)] \cdot \mathbf{F}\}, \tag{10}$$

where: $\mathbf{Y}_{1k}(i) := \mathbf{y}_k(i-1) \cdot \mathbf{y}_k^{\mathcal{H}}(i)$; $\mathbf{Y}_{0k}(i) := \mathbf{y}_k(i) \cdot \mathbf{y}_k^{\mathcal{H}}(i)$; and $\mathrm{diag}(\mathbf{A})$ denotes the diagonal matrix obtained by nulling the off-diagonal elements of the square matrix $\mathbf{A}$. It can be readily verified that the detection rule in (10) recovers uniquely $\hat{\mathbf{F}}(i) = \mathbf{F}(i)$ in the absence of noise. Noise effects will be investigated by simulations in Section V.

The DDST decoder based on (10) processes three consecutive received data blocks and has very low complexity. To gain further insight about our DDST detector in (10), we consider the following special case:

Special Case: In a single transmit- and receive-antenna setting ($N_r = N_t = 1$), the matrix $\mathbf{F}(i)$ reduces to a scalar which we denote by $s(i)$. Thus, the detector in (10) can be rewritten as:

$$\hat{s}(i) = \arg\max_{s}\, [y_k(i-1)y_k^{*}(i)][y_k(i-2)y_k^{*}(i-1)]^{*} s. \tag{11}$$

Note that the unitary group property of $\mathbf{F}(i)$ implies that $s(i)$ must be a PSK symbol. Clearly, (11) reduces to a conventional single-antenna double differential detector [8]. Hence, the conventional single-antenna double differential coding is a special case of our scheme.

We now summarize our DDST coding and decoding stages in the following steps:

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s1) Choose $\mathcal{A}$ and design $\mathcal{G}$ to satisfy $|\mathcal{G}| = |\mathcal{A}|$;
s2) Build a unique mapping between $\mathcal{G}$ and $\mathcal{A}$ as in (8);
s3) Choose $\mathbf{C}(0)$ to satisfy (5) and $\mathbf{G}(1) = \mathbf{I}$;
s4) Transmit code matrices $\mathbf{C}(0)$ and $\mathbf{C}(1) = \mathbf{C}(0)$;
s5) Map $s(i)$ to $\mathbf{F}(i) \in \mathcal{G}$, $\forall\, i \geq 2$, using the mapping in s2);
s6) Obtain $\mathbf{C}(i)$, $\forall\, i \geq 2$ using (6), (7), and transmit $\mathbf{C}(i)$;
s7) Detect $\hat{\mathbf{F}}(i)$ via (10) and recover $s(i)$ by mapping $\hat{\mathbf{F}}(i)$ back to $s(i)$ as in (8).

So far, we required $\mathbf{F}(i) \in \mathcal{G}$ to be unitary and diagonal in order to arrive at the DDST coding/decoding scheme we described in steps s1)-s7). We proceed next to design the group $\mathcal{G}$ aiming at optimal performance in terms of minimizing the BER.

IV. Diagonal Unitary Group Designs

As discussed in Section III, the information symbols are conveyed through the code matrix $\mathbf{F}(i)$ whose group structure leads to the low-complexity DDST decoder in (10). In addition to low-complexity, well designed code matrices help us achieve (or approximate) the optimal performance affordable by the system design. We next derive our group design criteria by analyzing the error performance of our detector in (10).

Dropping the block index i, we define the pairwise code matrix error event $\{\mathbf{F} \to \mathbf{F}'\}$ as the event that the receiver decodes code matrix $\mathbf{F}'$ when $\mathbf{F}$ is actually sent. Let $P\{\mathbf{F} \to \mathbf{F}'\}$ be the pairwise code matrix error probability averaged over the fading channels. As discussed in [9], at high SNR, the BER is mainly determined by $P\{\mathbf{F} \to \mathbf{F}'\}$. Thus, we choose $P\{\mathbf{F} \to \mathbf{F}'\}$ as our figure of merit and proceed to design our diagonal unitary group that minimizes $P\{\mathbf{F} \to \mathbf{F}'\}$.

Assuming $\tilde{h}_{mk}$'s in (2) are i.i.d., zero-mean, unit-variance complex Gaussian variables, it is shown in [7] that $P\{\mathbf{F} \to \mathbf{F}'\}$ can be upper-bounded by

$$P\{\mathbf{F} \to \mathbf{F}\} \le \frac{1}{\left[\det\left(\mathbf{I} + \frac{\log_2 M}{4N_t} \cdot \frac{E_b}{4N_0} \cdot \Delta\mathbf{F}\right)\right]^{N_r}}, \quad (12)$$

where $\det(\cdot)$ stands for matrix determinant; and $\Delta\mathbf{F} := (\mathbf{F} - \mathbf{F}')(\mathbf{F} - \mathbf{F}')^{\mathcal{H}}$. The expression in (12) has important implications for the design of our diagonal unitary group. At high SNR, $P\{\mathbf{F} \to \mathbf{F}'\}$ takes the form $[\Lambda_{dd} \cdot E_b/(4N_0)]^{-(L_{dd} \cdot N_r)}$, where L_{dd} and Λ_{dd} depend on the difference $\Delta\mathbf{F}$. The so-termed transmit diversity advantage L_{dd} is defined as:

$$L_{dd} = \text{rank}(\Delta\mathbf{F}) \le N, \quad (13)$$

and measures the performance gain from multiple transmissions. When $L_{dd} = N = N_t$, the so-termed coding advantage Λ_{dd} is given by

$$\Lambda_{dd} = \frac{\log_2 M}{4N_t}[\det(\Delta\mathbf{F})]^{\frac{1}{N_t}}. \quad (14)$$

In particular, when $N_t = 1$, our DDST decoder reduces to the conventional single antenna double differential coding scheme. In this case, if $M = 2$, then $\Lambda_{dd} = 1/4$. Compared to the corresponding coherent decoding scheme [5, p. 775], our scheme shows a 6 dB performance loss.

In order to optimize our DDST performance, we should maximize both L_{dd} and Λ_{dd}. We summarize the designs of unitary diagonal group in the following Lemma:

Lemma 1 [3, 7]: *For $M = 2^p$, every full-rank group of $N \times N$ diagonal unitary matrices with $|\mathcal{G}| = M$ is equivalent to an $(M; k_1, \ldots, k_N)$ cyclic group, for some odd numbers $0 \le k_1 \le \cdots k_N \le M$. In particular, letting*

$$\mathbf{\Theta} := \begin{pmatrix} \omega_M^{k_1} & 0 & \cdots & 0 \\ 0 & \omega_M^{k_2} & \cdots & 0 \\ & & \ddots & \\ 0 & 0 & 0 & \omega_M^{k_N} \end{pmatrix}, \quad (15)$$

the diagonal unitary group can be represented as:

$$\mathcal{G} = \{\mathbf{I}, \mathbf{\Theta}, \ldots, \mathbf{\Theta}^{M-1}\}. \quad (16)$$

M	R	k_1	k_2	Λ_{dd}
2	0.5	1	1	0.5
4	1	1	1, 3	0.25
8	1.5	1	3, 5	0.179
16	2	1	7,9	0.0732

TABLE I
$(M; k_1, k_2)$ CYCLIC GROUP ($N_t = 2$)

Using Lemma 1, it is easy to verify that $L_{dd} = N_t$ and

$$\Lambda_{dd} = \min_{1 \le l \le M-1} \frac{\log_2 M}{N_t} \left(\prod_{i=1}^{N} \sin\frac{\pi k_i l}{M}\right)^{2/N_t}. \quad (17)$$

Thus, designing our diagonal unitary group codes is equivalent to choosing $k_1, \ldots, k_N$ such that Λ_{dd} is maximized. For example, we collect in Tables I the optimal codes for $N_t = 2$.

Having designed the group $\mathcal{G}$, we next resort to simulations in order to test the performance of the DDST coding/decoding and compare it with existing alternatives.

V. SIMULATIONS

We simulate the performance of our DDST coding scheme using BER as our figure of merit which we average over 100 channel and noise realizations for each E_b/N_0 point. The time-selective fading channels are generated by taking $\tilde{h}_{mk}$ as complex Gaussian variables with variance 0.5 for both real and imaginary parts, and f_k as a random deviate uniformly distributed between 0 and 0.25. Two transmit-antennas and one receive-antenna are used in all examples.

Example 1 (performance of SDST [2] with frequency offsets): To motivate the design of double over *single* differential ST coding, we investigate the performance of SDST coding when the frequency error f_k does exist, but is not compensated for. We consider SNR = 16 dB. The simulation results are shown in Fig. 2 where we observe that the SDST coding is very sensitive to frequency errors. When $f_k = 0.05$, the BER increases from 10^{-3} to 10^{-2}. When $f_k \approx 0.1$, the system is useless because BER ≈ 0.5. For voice transmissions, the data rate is about 10 kbps ($T = 10^{-4}$). Thus, $f_k = 0.05$ is equivalent to $f_k^{(o)} + f_k^{(d)} = 500$ Hz which may happen in practice especially when the carrier frequency is up to GHz.

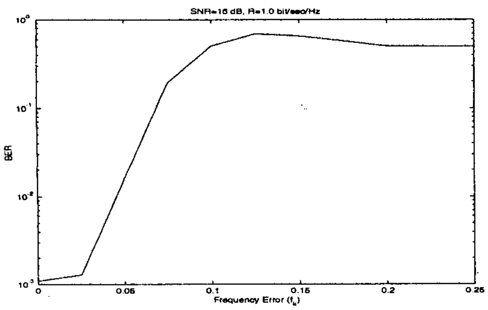

Fig. 2. SDST with frequency offsets

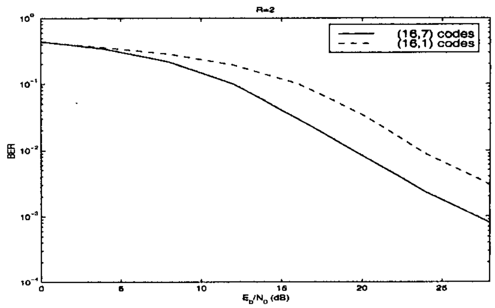

Fig. 3. $(16;1,7)$ vs. $(16;1,1)$ group codes

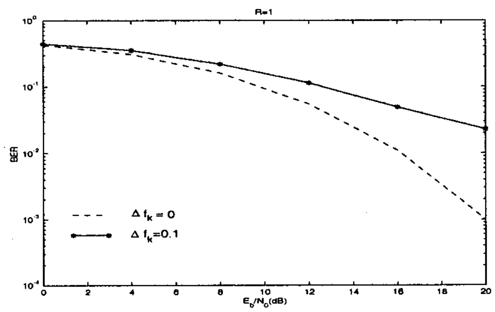

Fig. 4. DDST with $\Delta f_k \neq 0$

Example 2 (performance comparison with different unitary codes): In order to verify our design criteria and investigate the role optimal codes play in our scheme, we compare the DDST schemes with different group codes. The optimal $(16;1,7)$ codes will be compared to $(16;1,1)$ codes. Checking the coding advantage of the (16;1,1) codes, we find $\Lambda_{dd} = 0.019$ which is approximately $1/3.8$ of that for the $(16;1,7)$ optimal codes. This implies that in theory the $(16;1,1)$ codes may induce about 5.8 dB loss compared to the $(16;1,7)$ codes. The simulation results in Fig. 3 show that our optimal group codes outperform the $(16;1,1)$ cyclic group codes by about 4.5 dB. Considering the approximations we have made to arrive at (12), the simulation results are close to our theoretical results.

Example 3 (performance when f_k is different for different transmit antennas): In deriving (4), we assumed that the Doppler frequency shift $f_k^{(d)}$ is common to all transmit antennas. In this example, we test robustness of DDST against cases where $f_k^{(d)}$'s are different for different transmit antennas. The optimal $(4;1,1)$ group codes are used. Let Δf_k denote the difference between f_k's for the two transmit antennas. We emphasize that our system is designed for a common f_k; thus, only the Δf_k should be taken into account. We check performance when $\Delta f_k = 0$ and when $\Delta f_k = 0.1$. Fig. 4 shows that $\Delta f_k = 0.1$ increases the BER from 10^{-2} to 5×10^{-2} at 16 dB. Recall that $f_k = 0.1$ led the single differential space-time coding scheme in Example 1 to an unacceptable performance loss.

REFERENCES

[1] B. Hochwald and W. Sweldens, "Differential unitary space-time modulation," submitted to *IEEE Trans. on Comm.*, July 1999.

[2] B. L. Hughes, "Differential space-time modulation," submitted to *IEEE Trans. on Info. Theory*, 1999; see also *Proc. of 33rd Asilomar Conf.*, pp. 627-631, Pacific Grove, CA, Oct. 1999.

[3] B. L. Hughes, "Optimal space-time constellations from groups," submitted to *IEEE Trans. on Info. Theory*, 2000.

[4] Y. Li, N. Seshadri, S. Ariyavisitakul, "Channel estimation for OFDM systems with transmitter diversity in mobile wireless channels," *IEEE JSAC*, vol.17, no.3, pp. 461-471, March 1999.

[5] J. G. Proakis, *Digital Communications*, Third Edition, McGraw Hill, 1995.

[6] Z. Liu, A. Scaglione, S. Barbarossa, G. B. Giannakis, "Transmit-antennae space-time block coding for generalized OFDM in the presence of unknown multipath," submitted to *IEEE JSAC*, Sept., 1999; see also *Proc. of 33rd Asilomar Conf.*, pp. 1557-1561, Pacific Grove, CA, Oct. 1999

[7] Z. Liu, G. B. Giannakis and B. L. Hughes, "Double differential space-time block coding for time-slective fading channels," submitted to *IEEE Trans. on Comm.*, March, 2000.

[8] M. K. Simon, S. M. Hinedi and W. C. Lindsey, *Digital communication techniques, signal design and detection*, Prentice Hall, 1995.

[9] V. Tarokh, N. Seshadri and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction," *IEEE Trans. on Info. Theory*, pp. 744-765, March 1998.

[10] V. Tarokh and H. Jafarkhani, "A differential detection scheme for transmit diversity," *Proc. of WCNC*, pp. 1043-1047, New Orleans, LA, Sept. 1999.