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Differential Detection of Space-Time Spreading with Two Transmit Antennas

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A differential detection scheme for space-time spreading with two transmit antennas is proposed. The scheme does not require channel state information at either the transmitter or the receiver. With segmentation and preamble symbols padded at the transmitter, the receiver recovers the information using differential detection. Both phase-shift keying (PSK) and quadrature amplitude modulation (QAM) signals are considered. The proposed scheme achieves two-level transmit diversity gain with low complexity and saves the use of channel estimation, while having about 3 dB performance loss as compared to the coherent detection scheme. When multiple receive antennas exist, additional receive diversity gain can be achieved along with the transmit diversity gain. The scheme works fine under block-fading channel as well as slow Rayleigh fading channel, which is a popular scenario for high-rate data communications. The system performance for different segment sizes, channel fading speeds, modulation methods, and numbers of receive antennas is studied through simulations.

1. INTRODUCTION

Receive diversity has been well known as one powerful technique to mitigate the effect of fading and shadowing for high-rate data transmission over wireless hostile channels. The classical approach is to use multiple receive antennas and maximal ratio combining at the receiver. However, implementing receive diversity at the mobile station (MS) is a large cost considering size, power, complexity, and so forth. Therefore, it is intuitive to consider transferring from receive diversity at the MS to transmit diversity at the base station (BS), which can properly balance the problems of electronics, power consumption, size of antenna arrays, and so forth.

Transmit diversity techniques have attracted great enthusiasm in the past few years. Space-time trellis coding (STTC) [1, 2] and space-time block coding (STBC) [3–10] were extensively investigated. More specifically, STBC using coherent detection was introduced in [3, 9, 10]; differential detection for STBC was proposed in [6–8]. Differential STBC under frequency-selective fading channels was considered in [5]. A high-rate differential STBC scheme was discussed in [4].

Inspired by the space-time block codes [3, 9], Hochwald and his colleagues proposed a transmit diversity scheme known as space-time spreading (STS) [11] for the downlink wideband direct-sequence (DS) code-division multiple-access (CDMA) systems, which achieves full transmit diversity with the use of multiple transmit antennas. STS was shown to have more significant advantages than other transmit diversity techniques for CDMA [11, 12]. Buehrer et al. further combine STS with phase-sweep transmit diversity (PSTD) to provide the transmit diversity gain of STS for both 2G and 3G systems [13]. In [14, 15], Yang and Hanzo investigated the performance of wideband CDMA (W-CDMA) and multicarrier (MC) DS-CDMA systems using STS-assisted transmit diversity, respectively.

In all previous STS discussions, coherent detection was conducted with the assumption that perfect knowledge on channel is available at the receiver. However, there exist cases that the estimated channel state information is not reliable and imperfect channel estimation does impact the system performance [12, 16]. In addition, channel estimation requires the transmission of pilots along with the information, which impairs system throughput and consumes the transmission power.

In this paper, we consider the case that no given knowledge on channel is available at either the transmitter or the receiver, and propose a differential detection scheme for STS with two transmit antennas and multiple receive antennas.

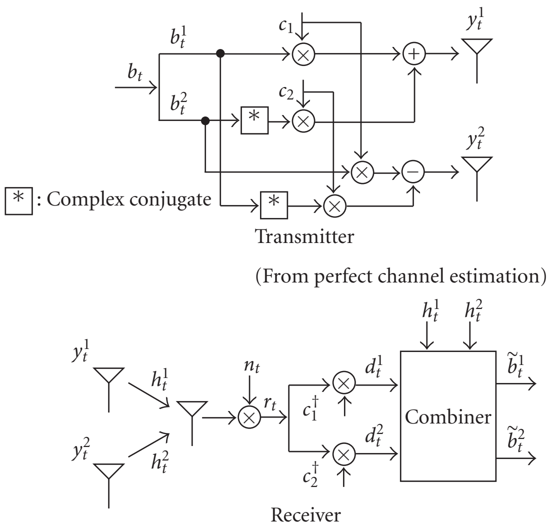

FIGURE 1: Transmitter and receiver of STS with coherent detection.

The two-level transmit diversity gain can still be achieved despite about 3 dB performance loss as compared to the coherent detection scheme. The differential detection scheme is proposed to support phase-shift keying (PSK) as well as quadrature amplitude modulation (QAM) signals. With multiple receive antennas available at the receiver, additional receive diversity gain can also be achieved in addition to the transmit diversity gain. We further study the impact on the performance of the proposed scheme by the segmentation size, fading speed of the channel, modulation methods, and the number of receive antennas.

The remainder of the paper is organized as follows. Section 2 reviews the coherent detection scheme for STS. Section 3 presents the differential detection schemes for STS using PSK and QAM signals. Section 4 discusses the use of multiple receive antennas for additional diversity gains. Section 5 provides simulation results regarding the coherent and differential detection schemes for STS, and investigates the effects of several system parameters. Section 6 concludes the paper and discusses possible future work.

2. COHERENT DETECTION OF STS

To simplify the notation and focus on the multiple-antenna aspects, we consider the downlink transmission of CDMA with orthogonal users experiencing no delay spread [11]. Two transmit antennas and one receive antenna are employed. Figure 1 gives the diagram of the transmitter and receiver of coherent detection of STS (CSTS). The desired user's data sequence $\{b_t\}$ is first split into two substreams $\{b_t^1\}$ and $\{b_t^2\}$. The substreams are then spread and combined in different fashions for transmission on the two transmit antennas:

$$
\begin{aligned}
y_t^1 &= \frac{1}{\sqrt{2}}\left(b_t^1 c_1 + b_t^{2*} c_2\right),\\
y_t^2 &= \frac{1}{\sqrt{2}}\left(b_t^2 c_1 - b_t^{1*} c_2\right),
\end{aligned}
\tag{1}
$$

where "$*$" denotes complex conjugate. The multiplicative coefficient of $1/\sqrt{2}$ is used to make the total transmission power of two transmit antennas the same as that of using one transmit antenna, so that no extra transmission power is required for more transmit antennas at the transmitter. c_1 and c_2 are the orthogonal code sequences for spreading the data, and are constructed by

$$
c_1 = \begin{bmatrix} c \\ 0_{\mathrm{SF}\times 1} \end{bmatrix}, \qquad c_2 = \begin{bmatrix} 0_{\mathrm{SF}\times 1} \\ c \end{bmatrix},
\tag{2}
$$

where c is the primary spreading code sequence for the specified user with spreading factor SF. Hence c_1 and c_2 are made orthogonal to each other and their effective spreading factors are $2\times\mathrm{SF}$. With this construction, no additional resources of spreading codes are required for two transmit antennas.

Assume the channel to be non-frequency selective. The received signal at the receiver can be represented as

$$
\begin{aligned}
r_t &= h_t^1 y_t^1 + h_t^2 y_t^2 + n_t\\
&= \frac{1}{\sqrt{2}}\left[h_t^1\left(b_t^1 c_1 + b_t^{2*} c_2\right) + h_t^2\left(b_t^2 c_1 - b_t^{1*} c_2\right)\right] + n_t,
\end{aligned}
\tag{3}
$$

where h_t^1 and h_t^2 denote the complex channel coefficients for the paths between transmit antennas 1, 2 and the receive antenna, respectively. n_t is complex additive white Gaussian noise, with zero mean and variance σ^2 equal to the double-side noise power spectral density $N_0/2$.

The received signal is then despread by

$$
\begin{aligned}
d_t^1 &= c_1^\dagger r_t = \frac{1}{\sqrt{2}}\left(h_t^1 b_t^1 + h_t^2 b_t^2\right) + c_1^\dagger n_t,\\
d_t^2 &= c_2^\dagger r_t = \frac{1}{\sqrt{2}}\left(h_t^1 b_t^{2*} - h_t^2 b_t^{1*}\right) + c_2^\dagger n_t,
\end{aligned}
\tag{4}
$$

where "$\dagger$" denotes complex conjugate transpose. Perfect knowledge on channel is assumed for coherent detection, that is, accurate estimates of h_t^1 and h_t^2 are assumed to be available at the receiver. Hence, we have

$$
\begin{aligned}
\widetilde{b}_t^1 &= d_t^1 h_t^{1*} - d_t^{2*} h_t^2\\
&= \frac{1}{\sqrt{2}}\left(\left|h_t^1\right|^2 + \left|h_t^2\right|^2\right) b_t^1 + h_t^{1*} c_1^\dagger n_t - h_t^2 n_t^\dagger c_2,\\
\widetilde{b}_t^2 &= d_t^1 h_t^{2*} + d_t^{2*} h_t^1\\
&= \frac{1}{\sqrt{2}}\left(\left|h_t^1\right|^2 + \left|h_t^2\right|^2\right) b_t^2 + h_t^1 n_t^\dagger c_2 + h_t^{2*} c_1^\dagger n_t,
\end{aligned}
\tag{5}
$$

where "$|\cdot|$" denotes the magnitude of a complex number.

The recovered data are then ready for either hard or soft decoding. The expression of the recovered data shows that "a two-level diversity gain" is achieved with this scheme, since the amplitude of the received signal will be very small only when both $|h_t^1|$ and $|h_t^2|$ have small values. It was shown in [11] that when multiple receive antennas are available, additional receive diversity gain can still be achieved with CSTS.

3. DIFFERENTIAL DETECTION OF STS WITH TWO TRANSMIT ANTENNAS AND ONE RECEIVE ANTENNA

Perfect knowledge of the channel is assumed for CSTS in Section 2. Although the receiver generally needs to estimate the channel for synchronization and carrier recovery purposes, it is difficult to guarantee the reliable and accurate channel estimation for all transmitted data symbols. Besides, channel estimation adds much complexity by transmitting pilots along with data.

In this paper, we assume that no channel state information is available at either the transmitter or the receiver. Differential detection of the signal is conducted for STS, where the received data is detected based on the differential relationship between each other. The technical mechanism in [7] is exploited for conducting differential detection.

Define coefficients

$$\begin{aligned} A_c &= b_{t+1}^1 b_t^{1*} + b_{t+1}^2 b_t^{2*}, \\ B_c &= b_{t+1}^2 b_t^1 - b_{t+1}^1 b_t^2, \end{aligned} \tag{6}$$

then given the current data b_t^1, b_t^2, and the coefficients A_c, B_c, the incoming data b_{t+1}^1, b_{t+1}^2 can be recovered by solving the equations in (6), yielding

$$\begin{aligned} b_{t+1}^1 &= \frac{A_c b_t^1 - B_c b_t^{2*}}{|b_t^1|^2 + |b_t^2|^2}, \\ b_{t+1}^2 &= \frac{A_c b_t^2 + B_c b_t^{1*}}{|b_t^1|^2 + |b_t^2|^2}. \end{aligned} \tag{7}$$

Therefore, if we specify the first two symbols $b_1^1 b_1^2$ for transmission to be known at the receiver, and if A_c, B_c can be obtained from the received signal, then we can recover the continued sequence $b_2^1 b_2^2 \cdots$ by recursive calculation based on (7).

The diagram of the proposed differential detection of STS (DSTS) scheme with two transmit antennas and one receive antenna is given in Figure 2. At the transmitter, first we uniformly divide each frame for transmission into segments of equal size. Two preamble symbols that are known to the receiver, b_1^1 and b_1^2, are attached at the beginning of each segment. Then the transmission of STS as that in (1) is followed. It can be noticed that the transmitter of DSTS is very similar to that of CSTS, except that an additional process of segmentation and padding preambles is performed.

The purpose of transmitting preamble symbols is to initialize the recursive calculation in (7) at the receiver. The reason that we pad each segment with preamble symbols rather than to pad each frame itself with preamble symbols is to limit the error propagation caused by differential detection,

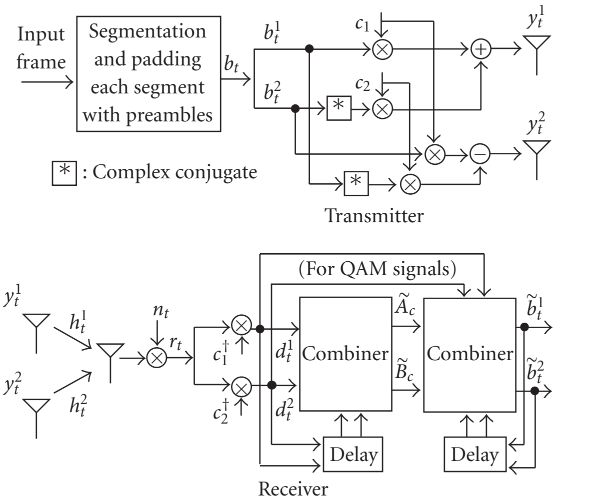

FIGURE 2: Transmitter and receiver of STS with differential detection.

since the detection of the incoming symbol depends on the current symbol, so error on one symbol tends to cause error on later symbols. It can be expected that given a fixed frame size and fixed number of frames transmitted, a segmentation of large size is likely to incur more errors than that of small size.

We assume that a perfect RAKE receiver would take care of the multipath fading. Therefore, the non-frequency-selective fading channel with only one path between each pair of transmit and receive antennas is considered. For each path, the fading channel is assumed to be quasistatic, that is, the path gains for any two neighboring symbols are assumed to have very little difference and can be approximated as identical. So for the paths between transmit antennas 1, 2 and the receive antenna, neighboring fading coefficients satisfy $h_t^1 \approx h_{t+1}^1$ and $h_t^2 \approx h_{t+1}^2$, respectively.

3.1. *Phase-shift-keying constellation symbols*

At the receiver, the received signal can still be expressed as that in (3). The same despreading process is then conducted as that in (4). In order to recover b_{t+1}^1 and b_{t+1}^2, we first recover A_c and B_c by combining the despread neighboring symbols, as shown in (8), where N_A and N_B denote the summation of the corresponding noise terms, respectively. With the assumption of quasistatic fading channel, the fading coefficient h_{t+1}^1 is replaced by h_t^1, and h_{t+1}^2 is replaced by h_t^2.

It can be found that $\widetilde{A}_c$ and $\widetilde{B}_c$ have a multiplicative factor of $1/2(|h_t^1|^2 + |h_t^2|^2)$ as compared to A_c and B_c, regardless of the noise relative terms. For PSK signals, this change in amplitude does not matter since different constellations only differ in phases and we consider only the signs of data. Therefore, $\widetilde{A}_c$ and $\widetilde{B}_c$ can be regarded as recoveries of A_c and B_c. The data symbols b_{t+1}^1 and b_{t+1}^2 are then recovered recursively by (9).

$$
\begin{aligned}
\widetilde{A}_c &= d_{t+1}^1 d_t^{1*} + d_{t+1}^{2*} d_t^2 \\
&\approx \left[\frac{1}{\sqrt{2}}(h_t^1 b_{t+1}^1 + h_t^2 b_{t+1}^2) + c_1^\dagger n_{t+1}\right]\left[\frac{1}{\sqrt{2}}(h_t^{1*} b_t^{1*} + h_t^{2*} b_t^{2*}) + n_t^\dagger c_1\right] \\
&\quad + \left[\frac{1}{\sqrt{2}}(h_t^{1*} b_{t+1}^2 - h_t^{2*} b_{t+1}^1) + n_{t+1}^\dagger c_2\right]\left[\frac{1}{\sqrt{2}}(h_t^1 b_t^{2*} - h_t^2 b_t^{1*}) + c_2^\dagger n_t\right] \\
&= \frac{1}{2}\left(|h_t^1|^2 + |h_t^2|^2\right)(b_{t+1}^1 b_t^{1*} + b_{t+1}^2 b_t^{2*}) \\
&\quad + \frac{1}{\sqrt{2}}[h_t^1 b_{t+1}^1 n_t^\dagger c_1 + h_t^2 b_{t+1}^2 n_t^\dagger c_1 + h_t^{1*} b_t^{1*} c_1^\dagger n_{t+1} + h_t^{2*} b_t^{2*} c_1^\dagger n_{t+1} \\
&\qquad + h_t^{1*} b_{t+1}^2 c_2^\dagger n_t - h_t^{2*} b_{t+1}^1 c_2^\dagger n_t + h_t^1 b_t^{2*} n_{t+1}^\dagger c_2 - h_t^2 b_t^{1*} n_{t+1}^\dagger c_2] \\
&\quad + c_1^\dagger n_{t+1} n_t^\dagger c_1 + n_{t+1}^\dagger c_2 c_2^\dagger n_t \\
&= \frac{1}{2}\left(|h_t^1|^2 + |h_t^2|^2\right) A_c + N_A, \\
\widetilde{B}_c &= d_{t+1}^{2*} d_t^1 - d_{t+1}^1 d_t^{2*} \\
&\approx \left[\frac{1}{\sqrt{2}}(h_t^{1*} b_{t+1}^2 - h_t^{2*} b_{t+1}^1) + n_{t+1}^\dagger c_2\right]\left[\frac{1}{\sqrt{2}}\left(h_t^1 b_t^1 + h_t^2 b_t^2\right) + c_1^\dagger n_t\right] \\
&\quad - \left[\frac{1}{\sqrt{2}}\left(h_t^1 b_{t+1}^1 + h_t^2 b_{t+1}^2\right) + c_1^\dagger n_{t+1}\right]\left[\frac{1}{\sqrt{2}}(h_t^{1*} b_t^2 - h_t^{2*} b_t^1) + n_t^\dagger c_2\right] \\
&= \frac{1}{2}\left(|h_t^1|^2 + |h_t^2|^2\right)(b_{t+1}^2 b_t^1 - b_{t+1}^1 b_t^2) \\
&\quad + \frac{1}{\sqrt{2}}[h_t^{1*} b_{t+1}^2 c_1^\dagger n_t - h_t^{2*} b_{t+1}^1 c_1^\dagger n_t + h_t^1 b_t^1 n_{t+1}^\dagger c_2 + h_t^2 b_t^2 n_{t+1}^\dagger c_2 - h_t^1 b_{t+1}^1 n_t^\dagger c_2 \\
&\qquad - h_t^2 b_{t+1}^2 n_t^\dagger c_2 - h_t^{1*} b_t^2 c_1^\dagger n_{t+1} + h_t^{2*} b_t^1 c_1^\dagger n_{t+1}] + n_{t+1}^\dagger c_2 c_1^\dagger n_t - c_1^\dagger n_{t+1} n_t^\dagger c_2 \\
&= \frac{1}{2}\left(|h_t^1|^2 + |h_t^2|^2\right) B_c + N_B
\end{aligned}
\tag{8}
$$

$$
\widetilde{b}_{t+1}^1 = \frac{\widetilde{A}_c \widetilde{b}_t^1 - \widetilde{B}_c \widetilde{b}_t^{2*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}, \qquad \widetilde{b}_{t+1}^2 = \frac{\widetilde{A}_c \widetilde{b}_t^2 + \widetilde{B}_c \widetilde{b}_t^{1*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}. \tag{9}
$$

The multiplicative coefficient $(|h_t^1|^2 + |h_t^2|^2)$ in (8) will be small only when both $|h_t^1|$ and $|h_t^2|$ have small values, that is, the receiver suffers from the detrimental effect of deep fading only if both subchannels from transmit antennas 1 and 2 to the receive antenna have small path gains. Therefore, a two-level transmit diversity gain is still achieved with this differential detection scheme.

As compared to the differential detection scheme for STBC in [7], one distinction in this paper is that [7] involves signals received during four transmit time slots while our scheme involves signals received in only two transmit time slots. The involved signals are assumed to have the same fading coefficients in the deduction of both schemes. Therefore, our scheme will be more adaptive to changes in channel gains.

3.2. Quadrature-amplitude-modulation constellation symbols

For QAM signals with constellation size larger than 4, different constellation points are identified by both amplitudes and phases. The data recovery process using (8) and (9) produces a multiplicative factor of $1/2(|h_t^1|^2 + |h_t^2|^2)$ as compared to the originally transmitted symbols, regardless of the effect of the noise. So in order to obtain the original amplitudes of QAM signals, the multiplicative factors must be compensated for before making decisions.

We refer to the expressions of d_t^1 and d_t^2 in (4) and find that solving these two equations yields approximations of h_t^1 and h_t^2, which are

$$
\begin{aligned}
\hat{h}_t^1 &= \frac{\sqrt{2}(b_t^{1*} d_t^1 + b_t^2 d_t^2)}{|b_t^1|^2 + |b_t^2|^2} \\
&= h_t^1 + \frac{\sqrt{2}(b_t^{1*} c_1^\dagger n_t + b_t^2 c_2^\dagger n_t)}{|b_t^1|^2 + |b_t^2|^2}, \\
\hat{h}_t^2 &= \frac{\sqrt{2}(b_t^{2*} d_t^1 - b_t^1 d_t^2)}{|b_t^1|^2 + |b_t^2|^2} \\
&= h_t^2 + \frac{\sqrt{2}(b_t^{2*} c_1^\dagger n_t - b_t^1 c_2^\dagger n_t)}{|b_t^1|^2 + |b_t^2|^2}.
\end{aligned}
\tag{10}
$$

Then we get

$$|\hat{h}_t^1|^2 + |\hat{h}_t^2|^2 = \frac{2\left(|d_t^1|^2 + |d_t^2|^2\right)}{|b_t^1|^2 + |b_t^2|^2}, \tag{11}$$

which can be viewed as an approximation of $|h_t^1|^2 + |h_t^2|^2$.

Since the exact values of b_t^1 and b_t^2 are not available at the receiver, we use the recovered data $\widetilde{b}_t^1$ and $\widetilde{b}_t^2$ to replace them in (11), yielding

$$|\widetilde{h}_t^1|^2 + |\widetilde{h}_t^2|^2 = \frac{2\left(|d_t^1|^2 + |d_t^2|^2\right)}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}. \tag{12}$$

Then we use (12) as an approximation of $|h_t^1|^2 + |h_t^2|^2$. Followed from (8), the data symbols can be recovered by

$$\begin{aligned}
\widetilde{b}_{t+1}^1 &= \frac{\widetilde{A}_c\widetilde{b}_t^1 - \widetilde{B}_c\widetilde{b}_t^{2*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2} \cdot \frac{2}{|\widetilde{h}_t^1|^2 + |\widetilde{h}_t^2|^2} \\
&= \frac{\widetilde{A}_c\widetilde{b}_t^1 - \widetilde{B}_c\widetilde{b}_t^{2*}}{|d_t^1|^2 + |d_t^2|^2}, \\
\widetilde{b}_{t+1}^2 &= \frac{\widetilde{A}_c\widetilde{b}_t^2 + \widetilde{B}_c\widetilde{b}_t^{1*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2} \cdot \frac{2}{|\widetilde{h}_t^1|^2 + |\widetilde{h}_t^2|^2} \\
&= \frac{\widetilde{A}_c\widetilde{b}_t^2 + \widetilde{B}_c\widetilde{b}_t^{1*}}{|d_t^1|^2 + |d_t^2|^2}.
\end{aligned} \tag{13}$$

With the above procedures counteracting the effects of the multiplicative factors, the QAM signals can now be recovered by the differential detection method.

It needs to be noted that in the above deduction the channel is assumed to be quasistatic and neighboring symbols do not have significant changes on channel gains. In practice, high-rate data transmission always experiences slow enough quasistatic fading since the data changes much faster than the channel does, which guarantees the effectiveness of the proposed differential detection scheme. In the deduction, we also assume the channel to be non-frequency selective, which includes only one path between each pair of transmit and receive antennas. This is based on the assumption that a perfect RAKE receiver counteracts the multipath.

4. DIFFERENTIAL DETECTION OF STS WITH TWO TRANSMIT ANTENNAS AND MULTIPLE RECEIVE ANTENNAS

Only one receive antenna is assumed in the discussion on differential detection of STS above. We now consider the case where N receive antennas are available at the receiver. Assume that each receive antenna is connected to the transmitter through an independent set of fading coefficients. Then for the signals transmitted in (1), the received signal at the ith receive antenna can be represented as

$$\begin{aligned}
r_t^i &= h_t^{1,i}y_t^1 + h_t^{2,i}y_t^2 + n_t^i \\
&= \frac{1}{\sqrt{2}}\left[h_t^{1,i}(b_t^1c_1 + b_t^{2*}c_2) + h_t^{2,i}(b_t^2c_1 - b_t^{1*}c_2)\right] + n_t^i,
\end{aligned} \tag{14}$$

where the superscript i, $i \in 1, 2, \ldots, N$, denotes the index of the corresponding received signal, fading coefficients, and noise for the ith receive antenna, respectively.

The despread signals on the ith receive antenna can then be represented by

$$\begin{aligned}
d_t^{1,i} &= c_1^\dagger r_t^i = \frac{1}{\sqrt{2}}(h_t^{1,i}b_t^1 + h_t^{2,i}b_t^2) + c_1^\dagger n_t^i, \\
d_t^{2,i} &= c_2^\dagger r_t^i = \frac{1}{\sqrt{2}}(h_t^{1,i}b_t^{2*} - h_t^{2,i}b_t^{1*}) + c_2^\dagger n_t^i.
\end{aligned} \tag{15}$$

A recovery on A_c and B_c can be obtained from the ith receive antenna as

$$\begin{aligned}
\widetilde{A}_c^i &= d_{t+1}^{1,i}d_t^{1,i*} + d_{t+1}^{2,i*}d_t^{2,i} \\
&= \frac{1}{2}\left(|h_t^{1,i}|^2 + |h_t^{2,i}|^2\right)A_c + N_A^i, \\
\widetilde{B}_c^i &= d_{t+1}^{2,i*}d_t^{1,i} - d_{t+1}^{1,i}d_t^{2,i*} \\
&= \frac{1}{2}\left(|h_t^{1,i}|^2 + |h_t^{2,i}|^2\right)B_c + N_B^i,
\end{aligned} \tag{16}$$

where N_A^i and N_B^i denote the sum of the noise relative terms with similar expressions as those for schemes using one receive antenna in (8).

As can be seen from the expression of $(|h_t^{1,i}|^2 + |h_t^{2,i}|^2)$ in (16), the diversity effect is obtained when recovering A_c and B_c. Then, when there are N receive antennas, the N recovered signals in (16), respectively, are combined together, yielding

$$\begin{aligned}
\widetilde{A}_c &= \sum_{i=1}^N \widetilde{A}_c^i = \sum_{i=1}^N (d_{t+1}^{1,i}d_t^{1,i*} + d_{t+1}^{2,i*}d_t^{2,i}) \\
&= \sum_{i=1}^N \frac{1}{2}\left(|h_t^{1,i}|^2 + |h_t^{2,i}|^2\right)A_c + \sum_{i=1}^N N_A^i, \\
\widetilde{B}_c &= \sum_{i=1}^N \widetilde{B}_c^i = \sum_{i=1}^N (d_{t+1}^{2,i*}d_t^{1,i} - d_{t+1}^{1,i}d_t^{2,i*}) \\
&= \sum_{i=1}^N \frac{1}{2}\left(|h_t^{1,i}|^2 + |h_t^{2,i}|^2\right)B_c + \sum_{i=1}^N N_B^i.
\end{aligned} \tag{17}$$

For PSK signals, the transmitted symbols can then be recovered by substituting (17) into (9). It is noticed in (17) that only when all of the $2 \times N$ terms $|h_t^{1,i}|$ and $|h_t^{2,i}|$ have low values the combined signal term will get small amplitude. Thus, a $2 \times N$ diversity gain can be achieved with the PSK-based DSTS scheme using 2 transmit antennas and N receive antennas. That is, additional receive diversity gain can still be achieved with DSTS when multiple receive antennas are available.

For QAM signals, the multiplicative factor

$$\sum_{i=1}^N (1/2)\left(|h_t^{1,i}|^2 + |h_t^{2,i}|^2\right) \tag{18}$$

in (17) still needs to be counteracted when recovering the data. From (12), we can get that for the ith receive antenna,

$$|\widetilde{h}_t^{1,i}|^2 + |\widetilde{h}_t^{2,i}|^2 = \frac{2\left(|d_t^{1,i}|^2 + |d_t^{2,i}|^2\right)}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}, \tag{19}$$

and hence an approximation of $\sum_{i=1}^{N}(1/2)(|h_t^{1,i}|^2 + |h_t^{2,i}|^2)$ is

$$\sum_{i=1}^{N}\left(|\widetilde{h}_t^{1,i}|^2 + |\widetilde{h}_t^{2,i}|^2\right) = \sum_{i=1}^{N}\frac{2\left(|d_t^{1,i}|^2 + |d_t^{2,i}|^2\right)}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}. \tag{20}$$

Followed from (13), the transmitted symbols can then be recovered by

$$\begin{aligned}\widetilde{b}_{t+1}^1 &= \frac{\widetilde{A}_c\widetilde{b}_t^1 - \widetilde{B}_c\widetilde{b}_t^{2*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}\cdot\frac{2}{\sum_{i=1}^{N}\left(|\widetilde{h}_t^{1,i}|^2 + |\widetilde{h}_t^{2,i}|^2\right)}\\ &= \frac{\widetilde{A}_c\widetilde{b}_t^1 - \widetilde{B}_c\widetilde{b}_t^{2*}}{\sum_{i=1}^{N}\left(|d_t^{1,i}|^2 + |d_t^{2,i}|^2\right)},\\ \widetilde{b}_{t+1}^2 &= \frac{\widetilde{A}_c\widetilde{b}_t^2 + \widetilde{B}_c\widetilde{b}_t^{1*}}{|\widetilde{b}_t^1|^2 + |\widetilde{b}_t^2|^2}\cdot\frac{2}{\sum_{i=1}^{N}\left(|\widetilde{h}_t^{1,i}|^2 + |\widetilde{h}_t^{2,i}|^2\right)}\\ &= \frac{\widetilde{A}_c\widetilde{b}_t^2 + \widetilde{B}_c\widetilde{b}_t^{1*}}{\sum_{i=1}^{N}\left(|d_t^{1,i}|^2 + |d_t^{2,i}|^2\right)}.\end{aligned} \tag{21}$$

Similarly, a $2\times N$ diversity gain is still achieved for QAM-based DSTS using 2 transmit antennas and N receive antennas.

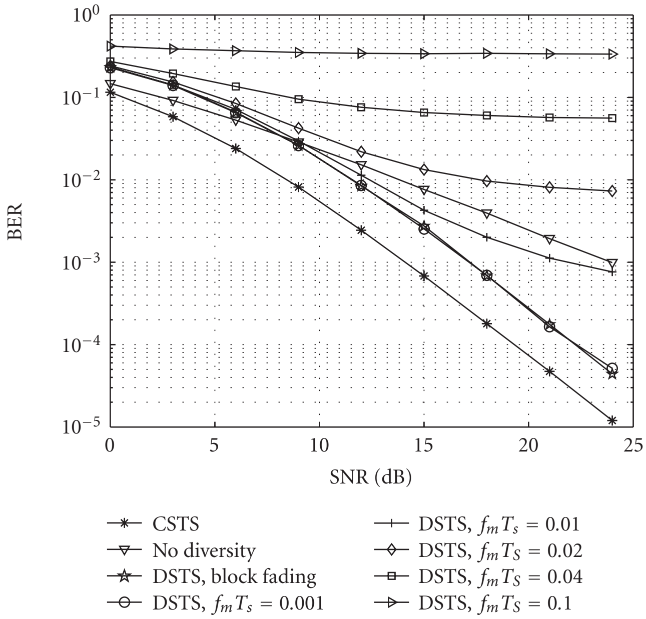

Figure 3: BER of DSTS (2 transmit antennas and 1 receive antenna) under block and Rayleigh fading channels with various f_mT_s, BPSK, DSTS with segment size of 10.

5. SIMULATIONS

In this section, we investigate the performance of STS with coherent and differential detection by simulations. First, we study the performance of the differential detection scheme with respect to the fading speed, where block and Rayleigh fading channels are considered. Then the impact of the segment size on the differential detection scheme is studied. Next, we verify the additional diversity gains for STS when multiple receive antennas are available. Finally, we compare coherent and differential detections of STS using different modulation methods. The spreading factor of the primary spreading code sequence c is set to 8.

5.1. CSTS and DSTS under block and Rayleigh fading channels

Figure 3 gives the BER of coherent and differential detection of STS under block-fading channel and Rayleigh fading channel with different values of normalized Doppler frequency shift f_mT_s, where f_m is the maximum Doppler frequency shift, and T_s is the symbol duration. f_mT_s specifies whether the channel is slow or fast fading as compared to the data rate. Larger value of f_mT_s denotes that the channel is changing faster, while smaller value of f_mT_s means that the channel is changing more slowly than the data does, respectively. In block-fading channel, the path gains are assumed to be constant over each segment and vary from one segment to another. For different segments, the channel gains are modelled as independent complex Gaussian random variables with variance 0.5 per dimension. The segment size of DSTS is set to 10. Two transmit antennas and one receive antenna are employed. Binary PSK (BPSK) modulation is applied. Signal-to-noise ratio (SNR) is defined as E_s/N_0, the average received energy per symbol over noise power spectral density.

For STS using coherent detection, since perfect knowledge on channel is assumed, the performance will not be affected by the fading channel type. Because the fading coefficients for all symbols are known exactly, we do not care whether the channel is block-fading or not, whether the Rayleigh fading is slow or fast. Figure 3 shows that CSTS achieves much lower BER than the scheme without diversity, which employs only one transmit antenna and one receive antenna also using coherent detection.

It is noticed that among all DSTS schemes, DSTS under block-fading channel performs the best, since it perfectly satisfies the condition of quasistatic fading, and it has around 3 dB performance loss as compared to CSTS. This can be explained by the more terms relative to noise in (8) than those in (5). In high SNR range, the noise terms relative to the product of n_t (or $n_t^\dagger$) and n_{t+1} (or $n_{t+1}^\dagger$) in (8) will be much smaller than other terms and hence can be ignored, the rest of the noise terms double the power of noise as compared to the coherent detection scheme, so a performance loss of approximately 3 dB can be expected, similar to that of differential detection of STBC [8]. This difference in performance between coherent and differential detection is also the same as that between coherent and differential modulations of PSK signals [17].

It is found in Figure 3 that for DSTS under Rayleigh fading channels, faster fading leads to much worse performance than slower fading does. This is reasonable since the idea of differential detection is based on the assumption of quasistatic fading. When the fading is fast, neighboring symbols

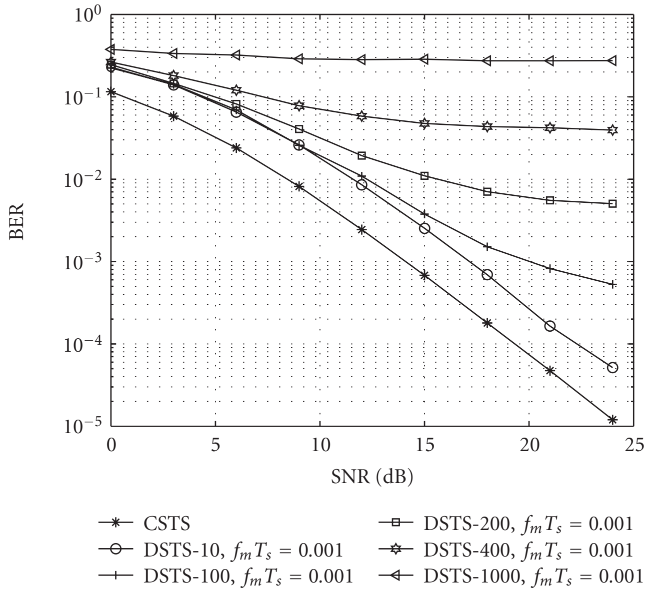

Figure 4: BER of DSTS (2 transmit antennas and 1 receive antenna) under Rayleigh fading channel with $f_mT_s = 0.001$, BPSK, DSTS with various segment sizes.

may experience much different channel gains, so the performance is severely degraded. When the fading is slow, however, there is no significant difference between channel gains of neighboring symbols, so the impact is much smaller. Especially, when f_mT_s is 0.001, the performance of differential detection for STS is very close to that under block-fading channel.

In practice, suppose the carrier frequency f_c is 1850 MHz, the vehicle speed v is 72.5 miles/hour, then $f_m = vf_c/c \doteq 200$ Hz [18], where c is the light speed. So when the data rate is 200 kbps using BPSK, then T_s is 5×10^{-6} s, and f_mT_s will be 0.001, which results in good performance as shown in Figure 3. f_mT_s becomes larger only when f_c or v has even larger values. Hence, for high-rate data transmission in practice, it is not difficult to meet the requirement of having slow enough fading channel, where the differential detection scheme can achieve BER as low as that under block-fading channel.

5.2. DSTS with different segment sizes under Rayleigh fading channel

In Figure 4, we evaluate the impact of the segment size on DSTS under Rayleigh fading channel with $f_mT_s = 0.001$ using BPSK. "DSTS-n" denotes DSTS with segment size of n.

It is shown in Figure 4 that the BER of DSTS increases as the segment size increases. This is the reason why each frame needs to be divided into smaller segments. When the number of symbols per segment reaches 200 and higher, the BER performance will be intolerably poor. This can be explained in that, even though the fading speed is slow, when the segment size is large, the symbols at the beginning and the end of each segment do have very different channel gains. In differential detection, although the detected value of one symbol only directly affects the next symbol, the cumulative effect of this will still lead to the poor performance of those symbols in the rear part of the segment when the segment size is large.

Table 1: Throughput of STS schemes with differential detection.

Segment size	10	100	200	400	1000
Throughput	0.833	0.980	0.990	0.995	0.998

It is also noticed that the curves of DSTS in Figures 3 and 4 are quite similar in shapes. Either a higher fading speed or a larger segment size leads to worse BER. This can be understood by that the system performance is determined by the value of f_mT_{seg}, where T_{seg} denotes the segment period. To make DSTS perform as good as that under block-fading channel, f_mT_{seg} needs to be small enough. When $f_mT_s = 0.001$ and $n = 10$, $T_{seg} = 10\times T_s$, that is, $f_mT_{seg} = 0.01$, DSTS works fine as shown in Figure 4. Increasing either f_m or T_{seg} will increase the BER. Again, the performance of CSTS will not be affected by the segment size, since perfect knowledge on channel is assumed and the detection of one symbol has nothing to do with other symbols in coherent detection.

For DSTS to perform well, we need to have small segment size. However, since each segment is attached with 2 preamble symbols for 2 transmit antennas, the system throughput (the ratio between all accepted symbols and all transmitted symbols) is impacted, and a segmentation of smaller size also leads to lower throughput than that of large size.

In order to evaluate the efficiency of transmission, we include the throughput of the differential detection schemes in Table 1. The redundancy per frame equals to the product of the number of preamble symbols per segment and the number of segments per frame. Table 1 shows that the 10-symbol/segment differential detection scheme pays the cost of the lowest throughput, although achieving the best BER. The throughput increases with larger segment size, but for schemes with segment size larger than 100 bits, the improvement in throughput as size increases is small, and the degradation in BER can be tremendous, as shown in Figure 4. After all, the overload caused by preambles in differential detection schemes is still much less than that of the coherent detection scheme, which needs to transmit pilots and conduct channel estimation depending on the channel variations.

5.3. CSTS and DSTS with multiple receive antennas, and higher spectral efficiency modulation methods

Figure 5 gives the BER of STS schemes with 2 transmit antennas and more than one receive antenna under Rayleigh fading channel with $f_mT_s = 0.001$, still using BPSK, with segment size of 10 for DSTS. The effect of additional diversity gains can be clearly noticed. It is found that for both CSTS and DSTS, the most gain is obtained by going from one receive antenna to two receive antennas. Increasing the number of receive antennas beyond two can still contribute additional gain, but there is a tendency of less gain for additional antennas.

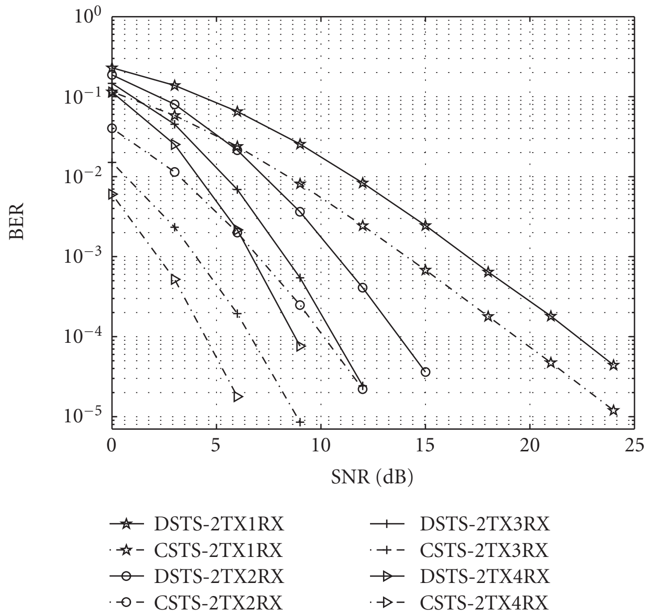

Figure 5: BER of CSTS and DSTS (2 transmit antennas and N receive antennas) under Rayleigh fading channel with $f_m T_s = 0.001$, BPSK, DSTS with segment size of 10.

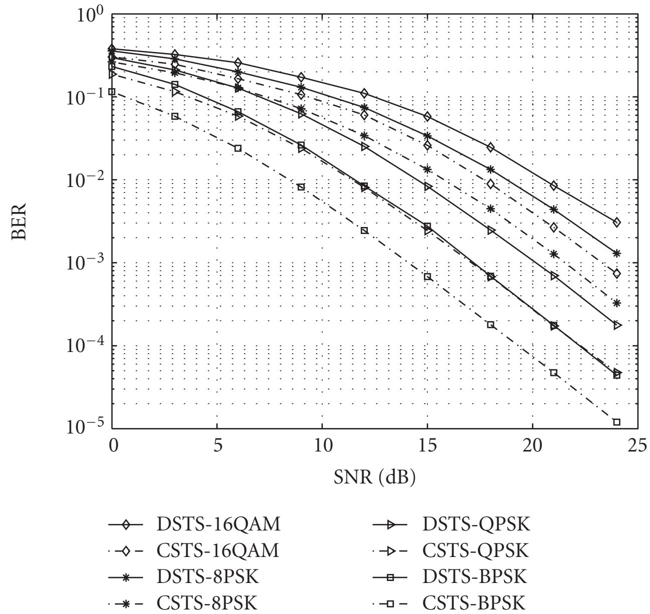

Figure 6: BER of DSTS (2 transmit antennas and 1 receive antenna) under Rayleigh fading channel with $f_m T_s = 0.001$, BPSK, QPSK, 8PSK, and 16QAM, DSTS with segment size of 10.

BPSK has been employed in all above performance results. In the following, we provide some results on higher spectral efficiency modulations. Figure 6 gives the BER of CSTS and DSTS using BPSK, quadrature PSK (QPSK), 8PSK, and 16QAM under Rayleigh fading channel with $f_m T_s = 0.001$, using 2 transmit and 1 receive antennas, DSTS having segment size of 10. The 3 dB loss from CSTS to DSTS can be observed for all modulations. There is also a 3 dB performance loss from BPSK to QPSK, with the SNR defined as the ratio of the average received energy per symbol over noise power spectral density. Actually, since there is no crosstalk or interference between the signals on the two quadrature carriers of QPSK, if we take SNR for energy per bit instead of per symbol, the BER of QPSK will be identical to that of BPSK [17]. Since 8PSK and 16QAM make less constellation space per bit than BPSK and QPSK, larger performance loss of 8PSK and 16QAM can be observed from Figure 6.

Figure 7 provides the performance of DSTS with two transmit antennas and multiple receive antennas under Rayleigh fading channel with $f_m T_s = 0.001$ for QPSK, 8PSK, and 16QAM with segment size of 10. Similar features can be found as those in Figure 5. The differences between CSTS and DSTS are still 3 dB for these cases and so the corresponding CSTS curves are not plotted here.

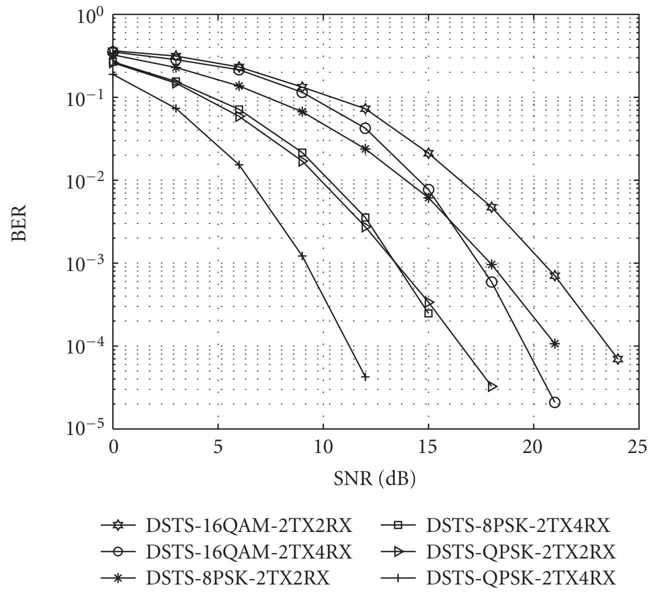

Figure 7: BER of DSTS (2 transmit antennas and N receive antennas) under Rayleigh fading channel with $f_m T_s = 0.001$, QPSK, 8PSK, and 16QAM, DSTS with segment size of 10.

6. CONCLUSIONS AND FUTURE WORK

In this paper, we proposed a differential detection scheme for STS with two transmit antennas, where neither the transmitter nor the receiver requires channel state information. Compared with the coherent detection scheme, the proposed method has about 3 dB performance loss in BER, however, it saves the use of pilots and channel estimation. The scheme was designated to accommodate both PSK and QAM modulated signals.

Experimental results demonstrated that the differential detection scheme works better with smaller segment size, since larger segment size tends to spread the error propagation caused by differential detection. With the assumption of perfect RAKE receiver, the non-frequency-selective fading channel was considered. The scheme was proposed based on quasistatic fading channel, and was shown to work fine under block-fading channel as well as Rayleigh fading channel when the fading speed is slow enough as compared to the data rate, which can generally be satisfied for high-rate data communications in practice. With multiple receive antennas available, additional diversity gain can still be achieved by the proposed scheme.

The complexity of the differential detection scheme for STS is very low. At the transmitter, a process of segmentation and padding preamble symbols with each segment is performed in addition to the space-time spreading. At the receiver, the combinations of neighboring symbols are conducted instead of that with estimated channel gains. Therefore, the proposed differential detection scheme for STS can act as a good alternative of the coherent detection scheme, when either perfect channel estimation is not available or accurate channel estimation is a high cost.

In our analysis and simulation, two transmit antennas are assumed. Future work will focus on generalizing the differential detection scheme for multiple transmit antennas. We assume perfect RAKE receiver counteracts the effect of multipath in the discussion. Future research will also study the performance of DSTS under frequency-selective fading channel. Single-user case is considered in this paper. It can be shown that the differential detection scheme does not impair the orthogonal relationship between data for different users, and so does not incur any additional multiuser interference. The development of two spreading codes in (2) also guarantees that there is no waste of the orthogonal code resources. Therefore, the proposed scheme can be used in multiuser cases as well.

The coherent detection scheme for STS needs to transmit pilots for channel estimation, while the differential detection scheme needs to add a few preamble symbols for each segment. The integrative utilization of these redundancies may be studied in the future to improve the overall performance of BER and throughput under various channel conditions.

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