New Detection Schemes for Transmit Diversity with no Channel Estimation

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Abstract

We present two new detection methods for the simple transmit diversity scheme proposed in [1]. The new detection schemes require no channel state information at either the transmitter or the receiver. Using 2 transmit antennas and 1 receive antenna, the scheme provides a diversity order of two. Simulations have been included to show that the scheme has almost a 3 dB penalty compared to coherent detection.

1. Introduction

A simple and effective transmit diversity scheme was proposed in [1]. This scheme has a very simple implementation and using 2 transmit antennas and 1 receive antenna provides the same diversity order as maximal ratio receiver combining (MRRC) with 1 transmit and 2 receive antennas. One requirement for the simple transmit diversity scheme is that channel state information is needed for decoding at the receiver. An approach to obtain this channel state information is the use of orthogonal pilot sequences as proposed in [2]. It appears from the results of [2] that it is possible to obtain accurate channel state information using pilots in slowly changing channels. It is also beneficial if the statistics of the channels are known. However, measurements show that the statistics of wireless channels are highly variant and finding a general model which holds in all scenarios seems to be a very difficult if not an impossible task. Therefore, there may be scenarios or applications where coherent detection is not plausible.

This motivates us to consider transmit diversity schemes where neither the transmitter nor the receiver require channel state information. Such schemes are well known when there is only 1 transmit antenna. In particular, non-coherent differential detection is a scheme where channel estimation is not needed. In this paper, we use the transmit diversity scheme in [1] and propose new detection schemes that require no channel state information. We also investigate the performance degradation of these schemes compared to coherent detection.

Furthermore, the scheme in [1] has been generalized to higher number of transmit antennas [5]. The new detection schemes may easily be combined with those presented in that paper.

In Section 2, the system model for the transmit diversity scheme in [1] is reviewed, and in Section 3, the new detection method is presented. In Section 4, we introduce a second detection method that derives the channel state from information bearing data. We provide some simulation results in Section 5. In Section 6 we have provided some simple analyses of sensitivity to time variance in the channel, and, finally, the conclusions are made in Section 7.

2. A Review of the Transmit Diversity Scheme

In this section, we present a review of the transmit diversity scheme first published in [1]. For a more detailed review of the scheme within this proceedings you may refer to [3]. Figure 2 shows the baseband representation of the transmit diversity scheme with 2 transmit antennas and 1 receive antenna.

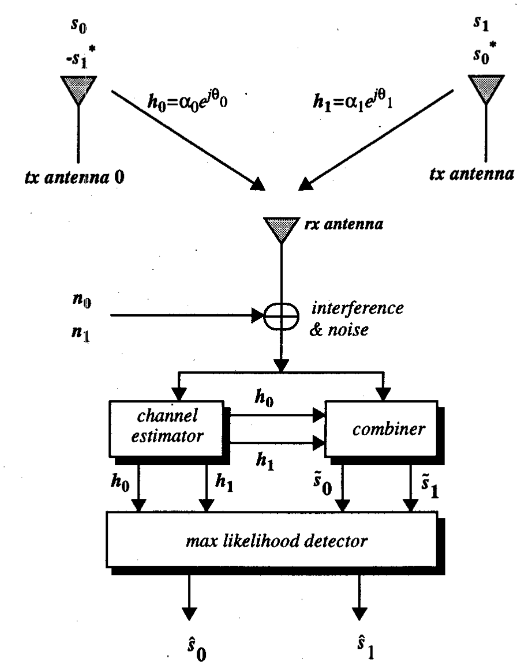

Figure 1 The New Two Branch Transmit Diversity Scheme with 1 Receiver.

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At a given symbol period, 2 signals are simultaneously transmitted from the 2 antennas. The signal transmitted from antenna 0 is denoted by s_0, and from antenna 1 by s_1. During the next symbol period signal ($-s_1*$) is transmitted from antenna 0, and signal s_0* is transmitted from antenna 1, where * is the complex conjugate operation. This sequence is shown in Table 1.

	antenna 0	antenna 1
time t	s_0	s_1
time $t + T$	$-s_1*$	s_0*

Table 1. The Encoding and Transmission Sequence for the New Transmit Diversity Scheme.

The forward transmission channel consisting of the transmit chain, the airlink, and the receive chain, at time t, may be modeled by a complex multiplicative distortion $h_0(t)$ for transmit antenna 0 and $h_1(t)$ for transmit antenna 1. Assuming that fading is constant across 2 consecutive symbols, we can write:

$$\begin{aligned} h_0(t) &= h_0(t+T) = h_o = \alpha_0 e^{j\theta_0} \\ h_1(t) &= h_1(t+T) = h_1 = \alpha_1 e^{j\theta_1} \end{aligned} \tag{1}$$

where T is the symbol duration. The received signals can then be expressed as:

$$\begin{aligned} r_0 &= r(t) = h_0 s_0 + h_1 s_1 + n_0 \\ r_1 &= r(t+T) = -h_0 s_1{}^* + h_1 s_0{}^* + n_1 \end{aligned} \tag{2}$$

where r_0 and r_1 are the received signals at time t and $t+T$, and n_0 and n_1 are complex random variables representing receiver noise and interference.

The combiner shown in Figure 1 builds the following 2 combined signals that are sent to the maximum likelihood detector:

$$\begin{aligned} \tilde{s}_0 &= h_0{}^* r_0 + h_1 r_1{}^* \\ \tilde{s}_1 &= h_1{}^* r_0 - h_0 r_1{}^* \end{aligned} \tag{3}$$

Substituting Equation 1 and Equation 2 into Equation 3, we get:

$$\begin{aligned} \tilde{s}_0 &= (\alpha_0^2 + \alpha_1^2) s_0 + h_0{}^* n_0 + h_1 n_1{}^* \\ \tilde{s}_1 &= (\alpha_0^2 + \alpha_1^2) s_1 - h_0 n_1{}^* + h_1{}^* n_0 \end{aligned} \tag{4}$$

The combined signals in Equation 4 are then sent to the maximum likelihood detector. Assuming n_0 and n_1 are Gaussian distributed, the maximum likelihood decision rule at the receiver for signal s_0 is to choose signal s_i iff [1]:

$$\begin{aligned} &(\alpha_0^2 + \alpha_1^2 - 1)|s_i|^2 + d^2(\tilde{s}_0, s_i) \le \\ &(\alpha_0^2 + \alpha_1^2 - 1)|s_k|^2 + d^2(\tilde{s}_0, s_k) \quad \forall i \ne k \end{aligned} \tag{5}$$

For PSK signals (equal energy constellations):

$$|s_i|^2 = |s_k|^2 = E_s \quad \forall i,k \tag{6}$$

where E_s is the energy of the signal. Therefore, for PSK signals, the decision rule in Equation 5 may be simplified to:

$$\begin{aligned} &\text{Choose } s_i \text{ iff:} \\ &d^2(\tilde{s}_0, s_i) \le d^2(\tilde{s}_0, s_k) \quad \forall i \ne k \end{aligned} \tag{7}$$

Similarly, the decision rule for signal s_1 is to choose signal s_i iff:

$$\begin{aligned} &(\alpha_0^2 + \alpha_1^2 - 1)|s_i|^2 + d^2(\tilde{s}_1, s_i) \le \\ &(\alpha_0^2 + \alpha_1^2 - 1)|s_k|^2 + d^2(\tilde{s}_1, s_k) \quad \forall i \ne k \end{aligned} \tag{8}$$

or for PSK signals:

$$\begin{aligned} &\text{Choose } s_i \text{ iff:} \\ &d^2(\tilde{s}_1, s_i) \le d^2(\tilde{s}_1, s_k) \quad \forall i \ne k \end{aligned} \tag{9}$$

3. Detection Method I

To keep the presentation simple, we assume that the fade coefficients are constant over every 4 consecutive transmissions. This is a reasonable assumption given that the symbol duration T is small when compared to the speed of change in a wireless channel often described by the maximum doppler frequency d_f. We further assume that the constellation points have equal energy which is normalized to $1/2$ here. In the next section, we present another method where the equal energy assumption is not needed.

It is assumed that the receiver knows signals s_0, s_1 and the received words r_0, r_1, r_2, r_3, r_4, r_5 ,... and proceeds to detect the transmitted signals s_2 and s_3. Having s_2 and s_3 computed, the process is repeated to compute s_4 and s_5 and then s_6 and s_7... Therefore the scheme requires the transmission of 2 known symbols at the beginning of an information block. Assuming that the channel for each transmit antenna is constant from t to $t+3T$:

$$\begin{aligned} r_0 &= r(t) = h_0 s_0 + h_1 s_1 + n_0 \\ r_1 &= r(t+T) = h_1 s_0{}^* - h_0 s_1{}^* + n_1 \\ r_2 &= r(t+2T) = h_0 s_2 + h_1 s_3 + n_2 \\ r_3 &= r(t+3T) = h_1 s_2{}^* - h_0 s_3{}^* + n_3 \end{aligned} \tag{10}$$

For decoding, the receiver first builds 2 intermediate values A and B given in the following equations:

$$
\begin{aligned}
A &= r_0 \cdot r_3{}^* - r_2 \cdot r_1{}^* \\
&= (|h_0|^2 + |h_1|^2) \cdot (s_2 \cdot s_1 - s_3 \cdot s_0) + N_1 \\
B &= r_2 \cdot r_0{}^* + r_1 \cdot r_3{}^* \\
&= (|h_0|^2 + |h_1|^2) \cdot (s_2 \cdot s_0{}^* + s_3 \cdot s_1{}^*) + N_2
\end{aligned} \tag{11}
$$

where N_1, N_2 are noise terms. The next step is to build estimates of s_2 and s_3 which are given by:

$$
\begin{aligned}
\tilde{s}_2 &= A s_1{}^* + B s_0 \\
&= (r_0 \cdot r_3{}^* - r_2 \cdot r_1{}^*) s_1{}^* + (r_2 \cdot r_0{}^* + r_1 \cdot r_3{}^*) s_0 \\
\tilde{s}_3 &= -A s_0{}^* + B s_1 \\
&= -(r_0 \cdot r_3{}^* - r_2 \cdot r_1{}^*) s_0{}^* + (r_2 \cdot r_0{}^* + r_1 \cdot r_3{}^*) s_1
\end{aligned} \tag{12}
$$

It may be shown that:

$$
\begin{aligned}
\tilde{s}_2 &= (|h_0|^2 + |h_1|^2) s_2 + N_3 \\
\tilde{s}_3 &= (|h_0|^2 + |h_1|^2) s_3 + N_4
\end{aligned} \tag{13}
$$

where N_3, N_4 are noise terms. It is interesting to note that Equation 13 resembles the equation for 2 branch maximal ratio receiver combining (2 branch MRRC). The decoder now decodes s_2 and s_3 by finding the closest constellation points to $\tilde{s}_2$ and $\tilde{s}_3$. Having s_2 and s_3 computed, s_2, s_3, s_4 and s_5 are replaced for s_0, s_1, s_2 and s_3 respectively in the above equations with r_2, r_3, r_4 and r_5 respectively replaced for r_0, r_1, r_2 and r_3. In this way, s_4 and s_5 are computed. This process is then similarly repeated.

4. Detection Method II

In this section, we present another method of detection where the equal energy assumption is removed. Again, it is assumed that the receiver knows the signals s_0, s_1 and the received words r_0, r_1, r_2, r_3, r_4, r_5,..... and proceeds to detect the transmitted signals s_2 and s_3. Having s_2 and s_3 computed the process is repeated to compute s_4 and s_5 and then s_6 and s_7 and so on. The receiver first forms the estimates

$$
\begin{aligned}
\tilde{h}_0 &= \frac{r_0 \cdot s_0{}^* - r_1 \cdot s_1}{|s_0|^2 + |s_1|^2} = h_0 + \frac{s_0{}^* n_0 + s_1 n_1}{|s_0|^2 + |s_1|^2} \\
\tilde{h}_1 &= \frac{r_0 \cdot s_1{}^* + r_1 \cdot s_0}{|s_0|^2 + |s_1|^2} = h_1 + \frac{s_1{}^* n_0 + s_0 n_1}{|s_0|^2 + |s_1|^2}
\end{aligned} \tag{14}
$$

for h_0 and h_1 and employs these estimates and the receive words r_2 and r_3 to decode for s_2 and s_3 using the detection method given by Equation 5 and Equation 7. Once s_2 and s_3 are computed the receiver uses Equation 14 (with r_0, r_1, s_0 and s_1 replaced by r_2, r_3, s_2 and s_3) to compute new estimates for h_0 and h_1 and decodes for s_4 and s_5. This process is then repeated.

Clearly if the channel is quasi-static all the previous (and present) estimates for the value of h_0 and h_1 can be averaged at each time. This gives less noisy estimates and the performance improves if the rate of change of the channel is small.

5. Error Performance Results

In this section, we provide performance results for the detection methods proposed in this work. First, we have included some simulation results for very slow fading. The simulations were carried out in baseband using SPW (Signal Processing WorkStation). The path gain between each transmit and the receive antenna is given by the Jakes Rayleigh fading model [4]. The path gains between distinct transmit and receive antennas are assumed independent and the average power of the received signal from each transmit antenna is the same. Each block consists of 50 QPSK symbols. At the beginning of each block 2 known symbols are transmitted. This will ensure that the error propagation is confined within a single block. The interference and noise are assumed to have gaussian distribution.

Simulations have shown that the performance of the two detection schemes are identical so we provide a single performance curve for both detection schemes. The simulation results are shown in Figure 2. Three block error rate curves are included: the transmit diversity scheme with perfect channel estimation, the transmit diversity scheme with the new detection schemes, and the performance of coherent QPSK with perfect channel estimation.

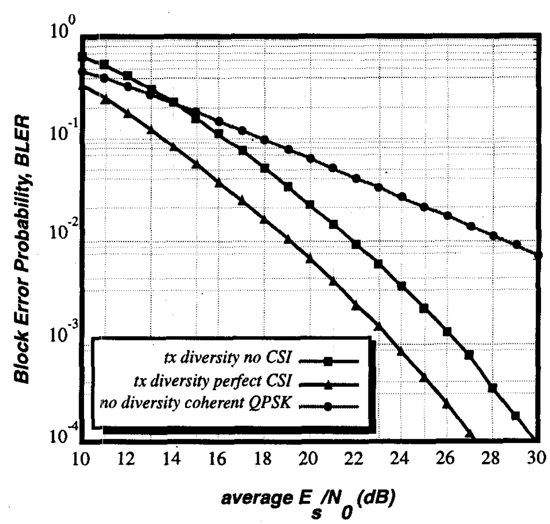

Figure 2 The Error Performance of Transmit Diversity with New Detection Schemes in Rayleigh Fading.

As shown in Figure 2, the performance of the new detection schemes is about 3 dB worse than coherent detection with ideal channel estimation. Nevertheless, the transmit diversity scheme with the new detection techniques provides 6 dB diversity gain at a BLER of 10^{-2}.

6. Sensitivity to Time Variance in the Channel

In this section, we have used simple statistical methods for fade duration statistics to predict the performance degradation as a function of the time variance in the channel. It is clear that the performance is highly dependent on the rate of change in the channel. The rate of change in the channel is often quantified by the normalized doppler frequency $f_d T$. Where f_d is the maximum doppler frequency and T is the symbol duration. The maximum doppler frequency may be described by the following expression[4]:

$$f_d = \frac{s}{c} f_c \tag{15}$$

where s is the speed of the remote unit, c is the speed of light, and f_c is the operating carrier frequency. For fixed applications, where the remote units are stationary, the doppler frequency depends on the scattering environment primarily the combinations of wind and foliage and moving objects. The average duration of a fade event as a function of the fading depth normalized to the symbol duration may then be described by [4]:

$$\tau = \frac{e^{\rho^2} - 1}{\sqrt{2\pi}\rho f_d T} \tag{16}$$

where ρ is fade level normalized to the RMS envelope of the signal. For the new detection schemes to provide their optimum performance, fading events should be constant across 4 symbols. This means that the fade duration should be considerably larger than 4 symbols so that the amplitude of fading remains constant across those symbols. Since the duration of fade is a decreasing function of the fade level, to ensure an acceptable performance from the new detection schemes, very deep fade levels must be considered. For the purpose of illustration, we shall assume a ρ threshold of -40 dB, in which case Equation 16 may be replaced by:

$$\tau = \frac{1}{100\sqrt{2\pi} f_d T} \tag{17}$$

Figure 3 shows the average duration of fade as a function of the normalized doppler frequency for a -40 dB fade level.

Our heuristic rule of thumb is to ensure that a -40 dB fade is constant across 4 symbols, the average duration of fade must be at least an order of magnitude larger. In other words, we expect that for the slow fading assumption to be reasonable, the average fade duration must be 40 symbols or longer. As shown in Figure 3, for the average fade duration to be larger than 40 symbols, $f_d T$ should be smaller than 10^{-4}. We may conclude that the performance of the new detection schemes will be near optimum for $f_d T$ values less than 10^{-4} and will degrade with higher values of $f_d T$.

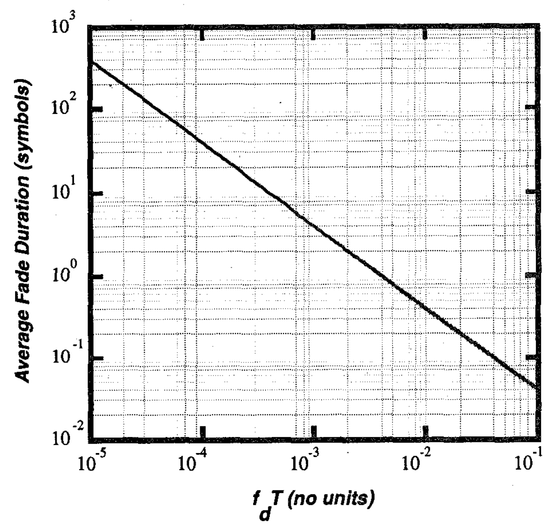

Figure 3 Normalized Average Duration of Fading as a Function of Normalized Doppler Frequency for a -40 dB Fade Level.

7. Conclusions

We presented two detection methods for the transmit diversity scheme in [1]. With 2 transmit antennas and a single receive antenna, a diversity order of 2 is achieved. In general, using these schemes, 2 transmit antennas and M receive antennas provide an order of diversity of $2M$. These detection schemes do not require channel estimation. The schemes require the transmission of 2 known symbols at the start of an information block. The SNR penalty for these detection schemes as compared to maximum likelihood detection with perfect CSI is roughly about 3 dB. The schemes perform optimally when the normalized doppler frequency is less than 10^{-4} and degrade thereafter.

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