Space-Time Diversity Design for Blind Estimation and Equalization over Frequency Selective Channels

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***Abstract*— Wireless communications often exploit guard intervals between data blocks to reduce inter-block-interference in frequency selective channels. Here we propose a dual-branch transmission scheme that utilizes guard intervals for blind channel estimation and equalization. Unlike existing diversity schemes, in which different antennas transmit delayed, zero-padded, or time reversed versions of the same signal, we use two antennas to transmit independent data streams. It is shown that for systems with two transmit antennas and one receive antenna, blind channel estimation can be carried out based only on the second order statistics of symbol rate sampled channel output. The proposed approach involves no pre-equalization and has no requirement on channel coprimeness. It is also shown that in combination with the T-BLAST structure [5] and Turbo coding, significant improvement can be achieved in the overall system performance.**

I. Introduction

Aiming for high spectral efficiency, recent years have witnessed broad research activities on blind channel estimation and signal detection. Although second order statistics of symbol rate sampled channel output alone can not provide enough information for blind channel estimation, it is possible with second order statistics of fractionally spaced/sampled channel output [7], [12] or baud-rate channel output samples from two or more receive antennas. This is, in fact, an early example on blind channel identification by exploiting space-time diversity techniques, the fractionally spaced sampling takes advantage of time diversity, while multiple receive antennas indicate spatial diversity at the receiver end.

Space-time coded systems, which generally fall into the MIMO framework, bring significant challenge to channel identification. In fact, in order to fully exploit the space-time diversity, the channel state information generally needs to be estimated at the receiver for all possible paths between Tx and Rx antenna pairs. Training based channel estimation may require considerable overhead. To further increase the spectral efficiency of space-time coded system, blind channel identification and signal detection algorithms have been proposed. In [10], blind and semiblind equalization, which exploit the structure of space-time coded signals, are presented for generalized space-time block codes which employ redundant precoders. Subspace based blind and semiblind approaches have been presented in [1]–[3], [13], and a family of convergent kurtosis based blind space-time equalization techniques are examined in [9].

Note that for frequency selective channels, guard intervals are often inserted between data blocks to prevent inter-block-interference, such as in the OFDM system [8], the chip-interleaved block-spread CDMA [14] and the generalized transmit delay diversity scheme [6]. In this paper, a simple two-branch transmission scheme, which is independent of modulation (OFDM or CDMA) format, is proposed to exploit the guard intervals for blind channel estimation and equalization. The generalized delay diversity proposed in [6] is perhaps the closest to our approach, but unlike [6], and also [3], [10], [13], in which different antennas transmit the delayed, zero-padded, or time reversed versions of the same signal, the proposed transmission scheme promises higher data rate since each antenna transmits an independent data stream.

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It is shown that with two transmit antennas and one receive antenna, blind channel estimation can be carried out based only on the second order statistics of symbol rate sampled channel output. This can be regarded as a counterpart of [12] which exploits *receive diversity*, and the proposed approach requires no channel coprimeness and involves no pre-equalization. For an overall data rate higher than that of the SISO system, two or more receivers are necessary for accurate equalization. Furthermore, it is also shown that in combination with the T-BLAST structure [5] and Turbo coding, significant gain can be achieved in the overall system performance.

II. The Proposed Transmit Diversity Scheme

The block diagram of the proposed two-branch transmit diversity scheme is shown in Figure 1. The input symbols are first split by a serial-to-parallel converter (S/P) into two parallel data streams; Each data stream then forms blocks with specific zero-padding structure. Data block structure may depends on the channel model and will be explained subsequently. The structured data blocks, $\bar{\mathbf{a}}_k$ and $\bar{\mathbf{b}}_k$, are

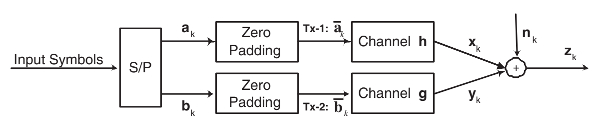


Fig. 1. Two-Branch Transmit Diversity

transmitted through two transmit antennas over frequency selective fading wireless channels, with channel impulse response vectors denoted by $\mathbf{h}$ and $\mathbf{g}$, respectively. The received signal is therefore the superposition of distorted information signals, $\mathbf{x}_k$ and $\mathbf{y}_k$, from each transmit antenna, and the additive noise $\mathbf{n}_k$.

Let L denote the maximum multipath delay spread for both $\mathbf{h}$ and $\mathbf{g}$. We consider the following two cases:

1) *Initial transmission delays are known while the two branches are synchronous*. In this case, without loss of generality, the channel impulse responses can be represented as:

$$\mathbf{h} = [h(0), h(1), \cdots, h(L)], \quad (1)$$
$$\mathbf{g} = [g(0), g(1), \cdots, g(L)], \quad (2)$$

with $h(0) \neq 0, g(0) \neq 0$.
Partition the data stream from each branch into N-symbol blocks ($N \geq L+1$), denote the k-th block from branch 1 and branch 2 by $\mathbf{a}_k = [a_k(0), a_k(1), \cdots, a_k(N-1)]$ and $\mathbf{b}_k = [b_k(0), b_k(1), \cdots, b_k(N-1)]$, respectively. Zero-padding is performed for each data block according to the

following structure. Define

$$\bar{\mathbf{a}}_k = [a_k(0), a_k(1), \cdots, a_k(N-1), \underbrace{0, \cdots, 0}_{L+1}], \quad (3)$$

$$\bar{\mathbf{b}}_k = [0, b_k(0), b_k(1), \cdots, b_k(N-1), \underbrace{0, \cdots, 0}_{L}], \quad (4)$$

and assume that there are M blocks in a data frame and the channel is time-invariant within each frame. Transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \cdots, \bar{\mathbf{a}}_M]$ from antenna 1 through channel $\mathbf{h}$, and transmit $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \cdots, \bar{\mathbf{b}}_M]$ from antenna 2 through channel $\mathbf{g}$. With the notation that $a_k(n) = b_k(n) = 0$ for $n < 0$ and $n > N-1$, we have

$$x_k(n) = \sum_{l=0}^{L} h(l) a_k(n-l), \quad (5)$$

$$y_k(n) = \sum_{l=0}^{L} g(l) b_k(n-l-1). \quad (6)$$

Define $\mathbf{x}_k = [x_k(0), x_k(1), \cdots, x_k(N+L)]^T$ and $\mathbf{y}_k = [y_k(0), y_k(1), \cdots, y_k(N+L)]^T$. For $k = 1, 2, \cdots, M$, it follows that

$$\mathbf{x}_k = \underbrace{\begin{bmatrix} h(0) & & & & \\ h(1) & h(0) & & & \\ \vdots & \vdots & & & \\ h(L) & h(L-1) & \cdots & h(0) & \\ & \ddots & \ddots & \ddots & \\ & & & h(L) & h(L-1) \\ & & & & h(L) \\ & & & & 0 \end{bmatrix}}_{\mathbf{H}} \underbrace{\begin{bmatrix} a_k(0) \\ a_k(1) \\ \vdots \\ a_k(N-1) \end{bmatrix}}_{\mathbf{a}_k}, \quad (7)$$

$$\mathbf{y}_k = \underbrace{\begin{bmatrix} 0 & & & & \\ g(0) & & & & \\ g(1) & g(0) & & & \\ \vdots & \vdots & & & \\ g(L) & g(L-1) & \cdots & g(0) & \\ & & \ddots & \ddots & \ddots \\ & & & g(L) & g(L-1) \\ & & & & g(L) \end{bmatrix}}_{\mathbf{G}} \underbrace{\begin{bmatrix} b_k(0) \\ b_k(1) \\ \vdots \\ b_k(N-1) \end{bmatrix}}_{\mathbf{b}_k}, \quad (8)$$

where $\mathbf{H}$ and $\mathbf{G}$ are $(N+L+1) \times N$ matrices. Define $\mathbf{n}_k = [n_k(0), n_k(1), \cdots, n_k(N+L)]$ and overall received signal is

$$\mathbf{z}_k = \mathbf{x}_k + \mathbf{y}_k + \mathbf{n}_k \quad (9)$$

2) *Initial transmission delays are unknown, and the two branches are either synchronous or asynchronous.*
Assume that the maximum transmission delay is d symbol intervals and the maximum multipath delay spread is L symbol intervals, the channel impulse responses corresponding to the two air links can be represented with two $(L+d+1)$ vectors,

$$\mathbf{h} = [h(-d_1), h(-d_1+1), \cdots, h(L+d-d_1)], \quad (10)$$

$$\mathbf{g} = [g(-d_2), g(-d_2+1), \cdots, g(L+d-d_2)], \quad (11)$$

where $0 \leq d_1, d_2 \leq d$. Define

$$\bar{\mathbf{a}}_k = [a_k(0), \cdots, a_k(N-1), \underbrace{0, \cdots, 0}_{L+2d+1}], \quad (12)$$

$$\bar{\mathbf{b}}_k = [\underbrace{0, \cdots, 0}_{d+1}, b_k(0), \cdots, b_k(N-1), \underbrace{0, \cdots, 0}_{L+d}]. \quad (13)$$

Again, transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \cdots, \bar{\mathbf{a}}_M]$ from antenna 1 through channel $\mathbf{h}$, and transmit $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \cdots, \bar{\mathbf{b}}_M]$ from antenna 2 through channel $\mathbf{g}$. It turns out that

$$x_k(n) = \sum_{l=0}^{L+d} h(l-d_1) a_k(n-l), \quad (14)$$

$$y_k(n) = \sum_{l=0}^{L+d} g(l-d_2) b_k(n-l-d-1). \quad (15)$$

Define $\mathbf{x}_k = [x_k(0), x_k(1), \cdots, x_k(N+L+2d)]^T$, $\mathbf{y}_k = [y_k(0), y_k(1), \cdots, y_k(N+L+2d)]^T$, $\mathbf{n}_k = [n_k(0), n_k(1), \cdots, n_k(N+L+2d)]^T$, and again define

$$\mathbf{z}_k = \mathbf{x}_k + \mathbf{y}_k + \mathbf{n}_k. \quad (16)$$

When multiple receive antennas are available, the received signal at each antenna can be characterized similarly using the same model as described above. It will be shown in later sections that increasing the number of receive antennas can improve the spectral efficiency, which is consistent with the results in [4], [11].

III. Blind Channel Identification

Our discussion in this section is based on the following assumptions:

(A1) The input information sequence is zero mean, mutually independent and *i.i.d.*. Absorbing any non-identity variance of the input symbols into the channel, this implies that $E\{a_k(m)a_l(n)\} = \delta_{k-l}\delta_{m-n}$, $E\{b_k(m)b_l(n)\} = \delta_{k-l}\delta_{m-n}$, and $E\{a_k(m)b_l(n)\} = 0$.

(A2) The noise is additive white Gaussian, independent of the information sequences, with variance σ^2.

Note that we impose *no limitation on channel zeros*. In what follows, blind channel identification is addressed for systems with proposed transmit diversity and with either one receiver or multiple receivers.

A. Two-branch transmit diversity with one receiver

We look at the synchronous two-branch diversity case first. Consider the auto-correlation matrix of the received signal block $\mathbf{z}_k$, $\mathbf{R}_\mathbf{z} = E\{\mathbf{z}_k \mathbf{z}_k^H\}$. It follows from (9) that for $k = 1, \cdots, M$,

$$\mathbf{R}_\mathbf{z} = \begin{bmatrix} |h(0)|^2 + \sigma^2 & h(0)h(1)^* & \cdots & h(0)h(L)^* & \\ h(1)h(0)^* & \sum_{l=0}^{1} |h(l)|^2 + |g(0)|^2 + \sigma^2 & \cdots & \cdots & \\ \ddots & \ddots & \ddots & \ddots & \ddots \end{bmatrix} \quad (17)$$

Based on (7), (8) and assumption (A2), it follows that

$$\mathbf{R}_\mathbf{z} = \mathbf{H}\mathbf{H}^H + \mathbf{G}\mathbf{G}^H + \sigma^2 \mathbf{I}_{N+L+1}, \quad (18)$$

where $\mathbf{I}_{N+L+1}$ denote the $(N+L+1) \times (N+L+1)$ identity matrix.

In the noise-free case,

$$\mathbf{R}_\mathbf{z} = \mathbf{H}\mathbf{H}^H + \mathbf{G}\mathbf{G}^H.$$

Note that $h(0) \neq 0$, $\mathbf{h} = [h(0), h(1), \cdots, h(L)]$ can be determined up to a phase $e^{j\theta}$ from the first row of $\mathbf{R}_\mathbf{z}$, and similarly, $\mathbf{g} = [g(0), g(1), \cdots, g(L)]$ can be determined up to a phase from the second row of $\mathbf{G}\mathbf{G}^H = \mathbf{R}_\mathbf{z} - \mathbf{H}\mathbf{H}^H$.

Noise variance estimation: In the noisy case, good estimation on the noise variance can improve the accuracy of channel estimation significantly, especially when the SNR is low. Here we provide two methods for noise variance estimation.

a) Recall that M is the number of blocks in a frame, without loss of generality, assume that M is even. We transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \cdots, \bar{\mathbf{a}}_{\frac{M}{2}}, |\bar{\mathbf{b}}_{\frac{M}{2}+1}, \bar{\mathbf{b}}_{\frac{M}{2}+2}, \cdots, \bar{\mathbf{b}}_M]$

from antenna 1 through channel $\mathbf{h}$, and $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \cdots, \bar{\mathbf{b}}_{\frac{M}{2}}, | \ \bar{\mathbf{a}}_{\frac{M}{2}+1}, \bar{\mathbf{a}}_{\frac{M}{2}+2}, \cdots, \bar{\mathbf{a}}_M]$ from antenna 2 through channel $\mathbf{g}$. Then for $k = 1, \cdots, \frac{M}{2}$, $\mathbf{R_z}$ is the same as in (17). And for $k = \frac{M}{2}+1, \cdots, M$,

$$\tilde{\mathbf{R}}_{\mathbf{z}} = \begin{bmatrix} |g(0)|^2+\sigma^2 & g(0)g(1)^* & \cdots \\ g(1)g(0)^* & \sum\limits_{l=0}^{1}|g(l)|^2+|h(0)|^2+\sigma^2 & \cdots \\ \ddots & \ddots & \ddots \end{bmatrix} \quad (19)$$

Define $r_{01} = g(0)g(1)^*, r_{02} = g(0)g(2)^*$ and $r_{12} = g(1)g(2)^*$. Let $A(i,j)$ denote the (i,j)-th entry of a matrix A, it follows from (17) and (19) that

$$g(1)g(2)^* = \tilde{\mathbf{R}}_{\mathbf{z}}(2,3) - \tilde{\mathbf{R}}_{\mathbf{z}}(1,2) - \mathbf{R}_{\mathbf{z}}(1,2). \quad (20)$$

Therefore r_{01}, r_{02}, r_{12} are all available, and

$$r_{12} = g(1)g(2)^* = \frac{r_{01}^*}{g(0)^*}\frac{r_{02}}{g(0)}, \quad (21)$$

When $r_{12} \neq 0$, we obtain the noise-free estimation $|g(0)|^2 = \frac{r_{01}^* r_{02}}{r_{12}}$ and the noise variance can be calculated from

$$\sigma^2 = \tilde{\mathbf{R}}_{\mathbf{z}}(1,1) - |g(0)|^2. \quad (22)$$

When $r_{12} = 0$, σ^2 can be estimated through similar discussion by considering $g(1) = 0$ and/or $g(2) = 0$. Substitute the estimated noise variance into (17), the noise-free estimation of $|h(0)|^2$ is obtained. It then follows directly that $\mathbf{h}$ and $\mathbf{g}$ can be estimated up to a phase difference. This method requires that M be large enough to obtain an accurate estimation of the correlation matrices. As an alternative, we may insert zeros and obtain noise variance estimate from a frame with almost half the length.

b) If we insert a zero after each block, that is, we transmit $[\bar{\mathbf{a}}_1, 0, \bar{\mathbf{a}}_2, 0, \cdots, \bar{\mathbf{a}}_M, 0]$ through $\mathbf{h}$ and $[\bar{\mathbf{b}}_1, 0, \bar{\mathbf{a}}_2, 0, \cdots, \bar{\mathbf{a}}_M, 0]$ through $\mathbf{g}$, then the new correlation matrix $\bar{\mathbf{R}}_{\mathbf{z}}$ of the channel output is

$$\bar{\mathbf{R}}_{\mathbf{z}} = \begin{bmatrix} \mathbf{R_z} & 0 \\ 0 & \sigma^2 \end{bmatrix} \quad (23)$$

The noise variance σ^2 can then be estimated and used for noise-free channel estimation in combination with $\mathbf{R_z}$, as discussed above.

Channel estimation in the asynchronous case follows directly from (12),(13) and our previous discussions. It can be seen that when the initial delays are unknown, extra overhead zeros are needed.

B. Systems with multiple receive antennas

For systems with two or more receivers, channel estimation can be performed at each receiver independently or from more than one receivers jointly. The major advantage of joint channel estimation is that accurate noise variance estimation becomes possible without inserting extra zeros or extending the frame length.

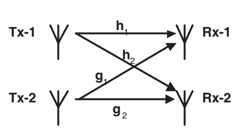


Fig. 2. Two-Branch Transmit Diversity with Two Receivers

Take a synchronous 2×2 system as an example (see Figure 2). Define $\mathbf{H}_1$,$\mathbf{H}_2$ as in (7) and $\mathbf{G}_1$,$\mathbf{G}_2$ as in (8), corresponding to $\mathbf{h}_1, \mathbf{h}_2, \mathbf{g}_1, \mathbf{g}_2$, respectively. If $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \cdots, \bar{\mathbf{a}}_M]$ is transmitted through $\mathbf{h}_1, \mathbf{h}_2$, and $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \cdots, \bar{\mathbf{b}}_M]$ is transmitted through $\mathbf{g}_1, \mathbf{g}_2$, the received signal at receiver 1 and 2 can be expressed as:

$$\mathbf{z}_k^1 = [\mathbf{H}_1, \mathbf{G}_1]\begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix} + \mathbf{n}_k^1, \quad \mathbf{z}_k^2 = [\mathbf{H}_2, \mathbf{G}_2]\begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix} + \mathbf{n}_k^2. \quad (24)$$

where $\mathbf{z}_k^1, \mathbf{z}_k^2$, $\mathbf{n}_k^1, \mathbf{n}_k^2$ are defined in the same manner as in Section II. Stacking $\mathbf{z}_k^1, \mathbf{z}_k^2$ into a $2(N+L+1)$-vector, we obtain

$$\mathbf{z}_k^L = \begin{bmatrix} \mathbf{z}_k^1 \\ \mathbf{z}_k^2 \end{bmatrix} = \underbrace{\begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix}}_{\triangleq \mathbf{F}} \underbrace{\begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix}}_{\triangleq \mathbf{s}_k} + \begin{bmatrix} \mathbf{n}_k^1 \\ \mathbf{n}_k^2 \end{bmatrix}. \quad (25)$$

Consider the correlation matrix of $\mathbf{z}_k^L$, it follows that

$$\mathbf{R}_z^L = E\{\mathbf{z}_k^L(\mathbf{z}_k^L)^H\} = \mathbf{F}\mathbf{F}^H + \sigma^2\mathbf{I}_{2(N+L+1)}. \quad (26)$$

Note that $\mathbf{F}$ is a $2(N+L+1)\times 2N$ tall matrix, the noise variance σ^2 can be estimated through the SVD of $\mathbf{R}_z^L$, by averaging the least $2(L+1)$ eigenvalues of $\mathbf{R}_z^L$. Extension to asynchronous systems and systems with more than two transmit antennas is straightforward and is omitted here.

IV. Equalization

Once channel estimation is carried out, equalization can be performed in several ways. Take the two-branch transmit diversity with two receivers as an example, define $\mathbf{s}_k = [\mathbf{a}_k^T, \mathbf{b}_k^T]^T$ as before, it follows from (25) that the information blocks $\mathbf{a}_k$ and $\mathbf{b}_k$ can be estimated by

$$\min_{\mathbf{s}_k} \| \mathbf{z}_k^L - \mathbf{F}\mathbf{s}_k \|, \quad (27)$$

either via least square or through the maximum likelihood (ML) approach based on the Viterbi algorithm.

For systems with two transmit antennas and one receiver, it follows from (7), (8) and (9) that

$$\mathbf{z}_k = [\mathbf{H}, \mathbf{G}]\mathbf{s}_k + \mathbf{n}_k, \quad (28)$$

and $[\mathbf{H}, \mathbf{G}]$ is $(N+L+1)\times 2N$. This implies that generally, we should choose $N = L+1$ such that $[\mathbf{H}, \mathbf{G}]$ is not severely rank deficient. In other words, the overall data rate of the two-branch transmit system with one receiver will be the same as the symbol rate of the corresponding single transmitter and single receiver system. While in the 2×2 system, $\mathbf{F}$ is $2(N+L+1)\times 2N$. Obviously, N is no longer constrained by L, and can be chosen as large as possible, as long as the frame length is within the channel coherence time range and the computational complexity is acceptable.

With the proposed transmit diversity scheme, blind channel identification and signal detection can be performed with the overall data rate much higher than the symbol rate of the SISO system. For a 2×2 system in a slow time-varying environment, for example, blind channel identification and signal detection can be achieved at a data rate $\frac{2N}{N+L+1}$ times that of the SISO system.

V. Simulation results

In this section, simulation results are provided to illustrate the performance of the proposed approach. In the example, each antenna transmits BPSK signals as the channel impulse response between each transmitter-receiver pair is generated randomly and independently. The channel is assumed to be static within each frame consisting of M blocks. In the simulation,

$$\text{Channel Estimation MSE} \triangleq \frac{1}{I}\sum_{i=1}^{I}\|\hat{\mathbf{h}}_i - \mathbf{h}_i\|^2/\|\mathbf{h}_i\|^2, \quad (29)$$

where $\mathbf{h}_i$ and $\hat{\mathbf{h}}_i$ denote the true channel and the estimated channel in the i-th run, respectively. I is the total number of Monto Carlo runs. At each receive antenna, the SNR is defined as ratio between the total received signal power and the noise power, and it is assumed that the receive antennas have the same SNR level. In the simulations, we choose $N = 3(L+1)$ so that the overall data rate is 1.5 times that of the corresponding SISO system over the same bandwidth, and all the simulation results are averaged over $I = 500$ Monte Carlo runs.

Consider *asynchronous two-branch transmission with two receivers*, either with or without the knowledge of the transmission delays. The channels are assumed to have three rays, the initial delays are uniformly distributed over [0,2] symbols, the direct path amplitude is normalized to 1, and the two successive paths have relative delays (with respect to the first arrival) uniformly distributed over [1,5] symbols with complex Gaussian amplitudes of zero mean and standard deviation 0.3. As shown in Figure 4.(a), when the delays are unknown, to detect the first arrival, the signal power of the first path should be sufficiently large in comparison with the noise power. *In order to illustrate the proposed approach in this example, channel estimation is obtained from one receiver, and equalization part is based on two receivers*.

We also consider to improve the system performance by combining the threaded layered space-time (TLST) coding [5] with the proposed transmission scheme, as shown in Figure 3. Here "SI" stands for *spatial interleaver*, and "Int" for *interleaver*. As can be seen in Figure 4.(b), significant performance improvement can be achieved with Turbo encoding (here we use a rate 1/2 Turbo encoder) and the TLST structure.

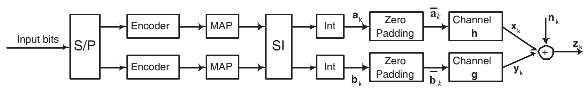
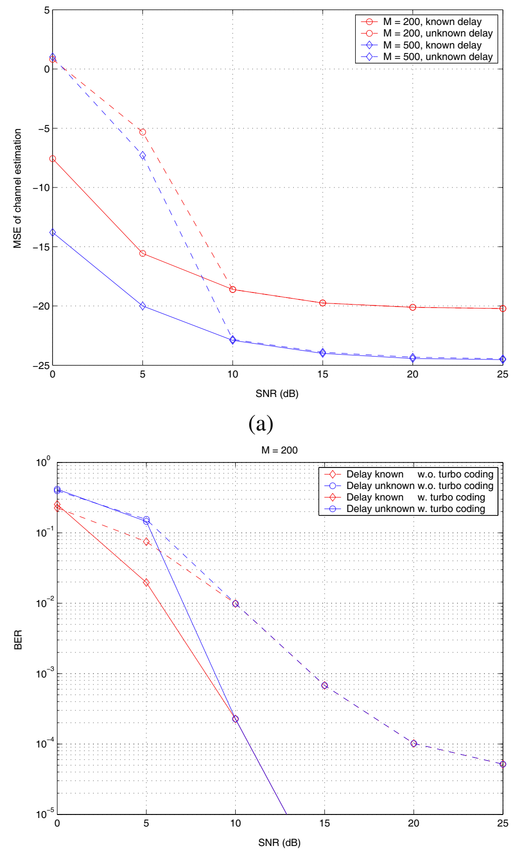


Fig. 3. ST-diversity-with-TBLAST



(a)

(b)

Fig. 4. (a) Channel estimation MSE versus SNR, (b) BER versus SNR.

VI. Conclusions

In this paper, a dual-branch transmission scheme that utilizes guard intervals for blind channel estimation and equalization is proposed. It is shown that with two transmit antennas and one receive antenna, blind channel estimation and equalization can be carried out based only on the second order statistics of symbol rate sampled channel output. The proposed approach involves no pre-equalization and has no requirement on channel zero locations. It is also shown that in combination with the T-BLAST structure and Turbo coding, significant improvement can be achieved in the overall system performance.

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