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Research Article

Transmit Delay Structure Design for Blind Channel Estimation over Multipath Channels

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Wireless communications often exploit guard intervals between data blocks to reduce interblock interference in frequency-selective fading channels. Here we propose a dual-branch transmission scheme that utilizes guard intervals for blind channel estimation and equalization. Unlike existing transmit diversity schemes, in which different antennas transmit delayed, zero-padded, or time-reversed versions of the same signal, in the proposed transmission scheme, each antenna transmits an independent data stream. It is shown that for systems with two transmit antennas and one receive antenna, as in the case of one transmit antenna and two receive antennas, 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 preequalization and has no limitations on channel-zero locations. Moreover, extension of the proposed scheme to systems with multiple receive antennas and/or more than two transmit antennas is discussed. It is also shown that in combination with the threaded layered space-time (TST) architecture and turbo coding, significant improvement can be achieved in the overall system performance.

1. 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 cannot provide enough information for blind channel estimation, it is possible with second-order statistics of fractionally spaced/sampled channel output or baud-rate channel output samples from two or more receive antennas [1–6]. These are, in fact, early examples of 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.

Receive diversity has been widely used in mobile communications (especially in the uplink) to obtain good system performance while minimizing the power consumption at the mobile handset. Exploitation of the transmit diversity, on the other hand, is more challenging, mainly because signals from multiple transmit antennas are mixed before they reach the receiver, and special consideration needs to be taken to separate these signals while allowing low-complexity receiver design. In [7, 8], it is proved that for memoryless channels, increasing the number of receive antennas in a SIMO system only results in a logarithmic increase in the average capacity, but the capacity of a MIMO system roughly grows linearly with the minimum number of antennas placed at both sides of the communication link. With the fundamental works in [9–11], space-time coding and MIMO signal processing have evolved into a promising tool in increasing the spectral efficiency of broadband wireless systems.

In [9], a simple two-branch transmit diversity scheme based on orthogonal design, the Alamouti scheme, is presented for flat fading channels. It is shown that the scheme using two transmit antennas and one receive antenna can achieve the same diversity order as using one transmit antenna and two receive antennas. The Alamouti scheme is directly applicable to systems with multiple receive antennas [9], and can be further extended to systems with any given number of transmit antennas [12], where the latter extension is generally referred to as space-time block coding. The transmit delay diversity scheme, in which copies of the same signal are transmitted from multiple antennas at different times, has been presented in [13, 14]. The transmit delay diversity scheme can also achieve the maximum possible transmit diversity order of the system [15]. Space-time Trellis codes were first developed in [11], and then refined by others, see [16], for example. The layered space-time codes, represented by the BLAST series, have been proposed in [10] and further developed in [17, 18].

Most existing space-time diversity techniques have been developed for flat fading channels. However, due to multipath propagation, wireless channels are generally frequency-selective fading instead of flat fading. Extensions of space-time diversity techniques suited for flat fading channels, especially the Alamouti scheme, to frequency-selective fading channels can be briefly summarized as follows: (i) apply generalized delay diversity (GDD) [19] or the time reversal technique [20]; (ii) convert a frequency-selective fading channel into a flat fading channel using equalization techniques, and then design space-time codes for the resulted flat fading channel(s), see [21] for example; (iii) convert the frequency-selective channels into a number of flat fading channels using OFDM scheme, see [22] and references therein; (iv) reformulate the multipath frequency-selective fading system into an equivalent flat fading system by regarding each single path as a separate channel, see [23] for example.

Space-time coded systems, which generally fall into the MIMO framework, bring significant challenges to channel identification. In fact, in order to fully exploit the space-time diversity, the channel state information generally needs to be estimated for all possible paths between the transmitter and receiver antenna pairs. Training-based channel estimation may result in considerable overhead. To further increase the spectral efficiency of space-time coded system, blind channel identification and signal detection algorithms have been proposed. In [24], blind and semiblind equalizations, 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 [25–28], and a family of convergent kurtosis-based blind space-time equalization techniques is examined in [29]. Blind algorithms based on the MUSIC and Capon techniques can be found in [30, 31], for example. Blind channel estimation for orthogonal space-time block codes (OSTBCs) has also been explored in literature, see [32–34], for example. In [33], based on specific properties of OSTBCs, a closed-form blind MIMO channel estimation method was proposed, together with a simple precoding method to resolve possible ambiguity in channel estimation.

Note that for frequency-selective fading channels, guard intervals are often put between data blocks to prevent interblock-interference, such as in the OFDM system [35], the chip-interleaved block-spread CDMA [36], and the generalized transmit delay diversity scheme [19]. 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 [19] is perhaps the closest to our approach, but unlike [19], and also [24, 27, 28], 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.

Through the proposed approach, we show 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 (SOS) of symbol-rate sampled channel output. This result can be regarded as a counterpart of the blind channel estimation algorithm proposed by Tong et al. [6], which exploits receive diversity. However, unlike [6], the proposed approach has no limitations on channel-zero locations. This is because we have more control over the data structure at the transmitter than at the receiver end, and a properly structured transmitter design can bring more flexibility to the corresponding receiver design.

With the proposed dual-branch transmitter design, when more than one receive antennas are employed, the data rate (in symbols/s/Hz, excluding training symbols or dummy zeros) can be increased by a factor of $2N/(N+L+1)$ (here N is the length of the data block and L is the maximum multipath delay spread, generally, $N \gg L+1$) compared with that of the corresponding SIMO system (under the same modulation scheme and with no training symbols transmitted). A direct corollary of the proposed approach is that for SISO systems, blind channel estimation based only on the second-order statistics of the symbol-rate sampled channel output is possible as long as the actual data rate (in symbols/s/Hz, excluding training symbols or dummy zeros) is not larger than $N/(N+L)$ times of the channel symbol rate. Theoretically, as long as the channel coherence time is long enough, we can choose $N \gg L$ so that $N/(N+L)$ can be arbitrarily close to 1.

The proposed scheme involves no preequalization, and does not rely on the OFDM framework to convert the frequency-selective fading channels to flat fading channels. Furthermore, in this paper, extension of the proposed scheme to systems with more than two transmit antennas is discussed, and it is also shown that in combination with the threaded space-time (TST) architecture [17] and turbo coding, significant improvement can be achieved in the overall system performance.

2. THE PROPOSED STRUCTURED TRANSMIT DELAY SCHEME

The block diagram of the proposed two-branch structured transmit delay scheme with one receive antenna 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. The data block structure depends on the channel model and will be explained subsequently.

The structured data blocks, $\bar{\mathbf{a}}_k$ and $\bar{\mathbf{b}}_k$, are 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,

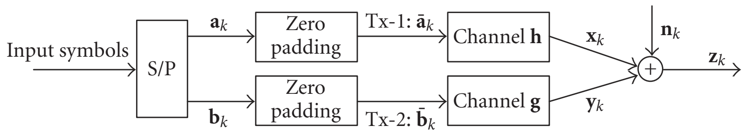


FIGURE 1: Two-branch transmit diversity with one receiver.

$\mathbf{x}_k$ and $\mathbf{y}_k$, from each transmit antenna, and the additive noise $\mathbf{n}_k$.

We assume that the two branches are synchronous and the initial transmit delay is known in this section and in the following two sections. We will discuss the extension of the proposed transmitter design to the synchronous and asynchronous cases with unknown delays, as well as the general MIMO systems in Section 5.

Let L denote the maximum multipath delay spread for both $\mathbf{h}$ and $\mathbf{g}$. When the initial transmission delays are known while the two branches are synchronous, without loss of generality, the channel impulse responses can be represented as

$$\begin{aligned} \mathbf{h} &= [h(0), h(1), \dots, h(L)], \\ \mathbf{g} &= [g(0), g(1), \dots, g(L)], \end{aligned} \tag{1}$$

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 kth block from branch 1 and branch 2 by $\mathbf{a}_k = [a_k(0), a_k(1), \dots, a_k(N-1)]$ and $\mathbf{b}_k = [b_k(0), b_k(1), \dots, b_k(N-1)]$, respectively. Zero-padding is performed for each data block according to the following structure. Define

$$\bar{\mathbf{a}}_k = \left[a_k(0), a_k(1), \dots, a_k(N-1), \underbrace{0, \dots, 0}_{L+1} \right], \tag{2}$$

$$\bar{\mathbf{b}}_k = \left[0, b_k(0), b_k(1), \dots, b_k(N-1), \underbrace{0, \dots, 0}_{L} \right], \tag{3}$$

and assume that there are M blocks in a data frame and the channel is time-invariant within each frame. Transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \dots, \bar{\mathbf{a}}_M]$ from antenna 1 through channel $\mathbf{h}$, and transmit $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \dots, \bar{\mathbf{b}}_M]$ 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

$$\begin{aligned} x_k(n) &= \sum_{l=0}^{L} h(l) a_k(n-l), \\ y_k(n) &= \sum_{l=0}^{L} g(l) b_k(n-l-1). \end{aligned} \tag{4}$$

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

$$\mathbf{x}_k = \underbrace{\begin{bmatrix} h(0) & & & & \\ h(1) & h(0) & & & \\ \vdots & \vdots & & & \\ h(L) & h(L-1) & \cdots & h(0) & \\ & \ddots & \ddots & \ddots & \\ & & & h(L) & h(L-1) \\ & & & & h(L) \\ & & & & 0 \end{bmatrix}}_{\mathbf{H}} \underbrace{\begin{bmatrix} a_k(0) \\ a_k(1) \\ \vdots \\ a_k(N-1) \end{bmatrix}}_{\mathbf{a}_k}, \tag{5}$$

$$\mathbf{y}_k = \underbrace{\begin{bmatrix} 0 & & & & \\ g(0) & & & & \\ g(1) & g(0) & & & \\ \vdots & \vdots & & & \\ g(L) & g(L-1) & \cdots & g(0) & \\ & & \ddots & \ddots & \ddots \\ & & & g(L) & g(L-1) \\ & & & & g(L) \end{bmatrix}}_{\mathbf{G}} \underbrace{\begin{bmatrix} b_k(0) \\ b_k(1) \\ \vdots \\ b_k(N-1) \end{bmatrix}}_{\mathbf{b}_k}, \tag{6}$$

where $\mathbf{H}$ and $\mathbf{G}$ are $(N+L+1) \times N$ matrices. Define $\mathbf{n}_k = [n_k(0), n_k(1), \dots, n_k(N+L)]$ and $\mathbf{z}_k = \mathbf{x}_k + \mathbf{y}_k + \mathbf{n}_k$, it then follows that for $k = 1, 2, \dots, M$,

$$\mathbf{z}_k = \begin{bmatrix} h(0)a_k(0) \\ h(1)a_k(0) + h(0)a_k(1) + g(0)b_k(0) \\ h(2)a_k(0) + h(1)a_k(1) + h(0)a_k(2) \\ +g(1)b_k(0) + g(0)b_k(1) \\ \vdots \\ h(L)a_k(0) + \cdots + h(0)a_k(L) \\ + g(L-1)b_k(0) + \cdots + g(0)b_k(L-1) \\ \vdots \\ h(L)a_k(N-1) + g(L)b_k(N-2) \\ +g(L-1)b_k(N-1) \\ g(L)b_k(N-1) \end{bmatrix} + \mathbf{n}_k. \tag{7}$$

3. 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., which 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 the proposed structured transmit delay and with either one receiver or multiple receivers.

3.1. Systems with single-receive antenna

Consider the autocorrelation matrix of the received signal block $\mathbf{z}_k$, $\mathbf{R_z} = E\{\mathbf{z}_k\mathbf{z}_k^H\}$. It follows from (7) that for $k = 1,\ldots,M$,

$$\mathbf{R_z} = \begin{bmatrix} |X_0|^2+\sigma^2 & X_0X_1^* & \cdots & X_0X_L^* & & \\ X_1X_0^* & a & \cdots & \cdots & X_0X_L^*+Y_0Y_L^* & \\ \ddots & \ddots & \ddots & \ddots & \ddots & \ddots \\ \cdots & \cdots & \cdots & \cdots & b & Y_{L-1}Y_L^* \\ & & Y_LY_0^* & \cdots & Y_LY_{L-1}^* & |Y_L|^2+\sigma^2 \end{bmatrix}, \tag{8}$$

where

$$\begin{aligned} &X_0 = h(0), \quad X_1 = h(1), \quad X_L = h(L) \\ &Y_0 = g(0), \quad Y_L = g(L), \quad Y_{L-1} = g(L-1) \\ &a = \sum_{l=0}^{1} |h(l)|^2 + |g(0)|^2 + \sigma^2, \\ &b = |h(L)|^2 + \sum_{l=L-1}^{L} |g(l)|^2 + \sigma^2. \end{aligned} \tag{9}$$

Based on (5), (6), and assumption (A2), it follows that

$$\mathbf{R_z} = \mathbf{HH}^H + \mathbf{GG}^H + \sigma^2\mathbf{I}_{N+L+1}, \tag{10}$$

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

In the noise-free case,

$$\mathbf{R_z} = \mathbf{HH}^H + \mathbf{GG}^H. \tag{11}$$

Note that when $h(0) \neq 0$, $\mathbf{h} = [h(0), h(1), \ldots, h(L)]$ can be determined up to a phase $e^{j\theta}$ from the first row of $\mathbf{R_z}$. Similarly, when $g(0) \neq 0$, $\mathbf{g} = [g(0), g(1), \ldots, g(L)]$ can be determined up to a phase from the second row of $\mathbf{GG}^H = \mathbf{R_z} - \mathbf{HH}^H$.

Noise variance estimation. In the noisy case, good estimation of 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) Recalling that M is the number of blocks in a frame, without loss of generality, assume that M is even. We transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \ldots, \bar{\mathbf{a}}_{M/2} \mid \bar{\mathbf{b}}_{M/2+1}, \bar{\mathbf{b}}_{M/2+2}, \ldots, \bar{\mathbf{b}}_M]$ from antenna 1 through channel $\mathbf{h}$, and $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \ldots, \bar{\mathbf{b}}_{M/2} \mid \bar{\mathbf{a}}_{M/2+1}, \bar{\mathbf{a}}_{M/2+2}, \ldots, \bar{\mathbf{a}}_M]$ from antenna 2 through channel $\mathbf{g}$. Then for $k = 1,\ldots,M/2$, $\mathbf{R_z}$ is the same as in (8). And for $k = M/2+1,\ldots,M$,

$$\widetilde{\mathbf{R}}_{\mathbf{z}} = \begin{bmatrix} |g(0)|^2+\sigma^2 & g(0)g(1)^* & g(0)g(2)^* & \cdots & g(0)g(L)^* \cdots \\ g(1)g(0)^* & c & d & \cdots & \cdots \\ \ddots & \ddots & \ddots & \ddots & \ddots \end{bmatrix}, \tag{12}$$

where $c = \sum_{l=0}^{1} |g(l)|^2 + |h(0)|^2 + \sigma^2$, $d = \sum_{l=0}^{1} g(l)g(l+1)^* + h(0)h(1)^*$.

Define $r_{01} = g(0)g(1)^*$, $r_{02} = g(0)g(2)^*$, and $r_{12} = g(1)g(2)^*$. Denoting by $A(i,j)$ the (i,j)th entry of a matrix $\mathbf{A}$, it follows from (8) and (12) that

$$g(1)g(2)^* = \widetilde{\mathbf{R}}_{\mathbf{z}}(2,3) - \widetilde{\mathbf{R}}_{\mathbf{z}}(1,2) - \mathbf{R_z}(1,2). \tag{13}$$

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

$$r_{12} = g(1)g(2)^* = \frac{r_{01}^*}{g(0)^*}\frac{r_{02}}{g(0)}. \tag{14}$$

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

$$\sigma^2 = \widetilde{\mathbf{R}}_{\mathbf{z}}(1,1) - |g(0)|^2. \tag{15}$$

When $r_{12} = 0$, then $g(1) = 0$ and/or $g(2) = 0$. If $g(1) = 0$,

$$\widetilde{\mathbf{R}}_{\mathbf{z}}(2,2) = |g(0)|^2 + |h(0)|^2 + \sigma^2. \tag{16}$$

Note that from (8), $\mathbf{R_z}(1,1) = |h(0)|^2 + \sigma^2$, therefore,

$$\sigma^2 = \mathbf{R_z}(1,1) - [\widetilde{\mathbf{R}}_{\mathbf{z}}(2,2) - \widetilde{\mathbf{R}}_{\mathbf{z}}(1,1)]. \tag{17}$$

If $g(1) \neq 0$ but $g(2) = 0$, then

$$\widetilde{\mathbf{R}}_{\mathbf{z}}(3,3) = |g(0)|^2 + |g(1)|^2 + |h(0)|^2 + |h(1)|^2 + \sigma^2, \tag{18}$$

thus

$$|g(1)|^2 = \widetilde{\mathbf{R}}_{\mathbf{z}}(3,3) - \mathbf{R_z}(2,2). \tag{19}$$

Again, we obtain

$$\sigma^2 = \widetilde{\mathbf{R}}_{\mathbf{z}}(1,1) - |g(0)|^2, \quad |g(0)|^2 = \frac{|r_{01}|^2}{|g(1)|^2}. \tag{20}$$

Substituting the estimated noise variance into (8), 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.

Note that in practice, $\mathbf{R_z}$ and $\widetilde{\mathbf{R}}_{\mathbf{z}}$ are generally estimated through time-averaging,

$$\mathbf{R_z} = \frac{2}{M}\sum_{k=1}^{M/2} \mathbf{z}_k\mathbf{z}_k^H, \qquad \widetilde{\mathbf{R}}_{\mathbf{z}} = \frac{2}{M}\sum_{k=M/2+1}^{M} \mathbf{z}_k\mathbf{z}_k^H. \tag{21}$$

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 $[\overline{\mathbf{a}}_1, 0, \overline{\mathbf{a}}_2, 0, \ldots, \overline{\mathbf{a}}_M, 0]$ through $\mathbf{h}$ and $[\overline{\mathbf{b}}_1, 0, \overline{\mathbf{b}}_2, 0, \ldots, \overline{\mathbf{b}}_M, 0]$ through $\mathbf{g}$, then the new correlation matrix $\overline{\mathbf{R}}_\mathbf{z}$ of the channel output is

$$\overline{\mathbf{R}}_\mathbf{z} = \begin{bmatrix} \mathbf{R}_\mathbf{z} & 0 \\ 0 & \sigma^2 \end{bmatrix}. \tag{22}$$

The noise variance σ^2 can then be estimated and used for noise-free channel estimation in combination with $\mathbf{R}_\mathbf{z}$, as discussed above. It should be noted that the transmission scheme in (b) has lower symbol rate compared to that in (a).

Discussion on SISO system. Consider a special case of the two-branch structured transmit delay scheme, in which antenna 2 is shut down, then it reduces to a SISO system. And the related autocorrelation matrix of the channel output is

$$\check{\mathbf{R}}_\mathbf{z} = \mathbf{H}\mathbf{H}^H + \sigma^2 \mathbf{I}_{N+L+1}, \tag{23}$$

and $\mathbf{h}$ can easily be obtained following our discussion above. It should be pointed out that for SISO system, instead of padding $L+1$ zeros to each $\mathbf{a}_k$ as in (2), we can define

$$\overline{\mathbf{a}}_k = \Big[a_k(0), a_k(1), \ldots, a_k(N-1), \underbrace{0, \ldots, 0}_{L} \Big], \tag{24}$$

and still perform blind channel identification with noise variance estimation as discussed above. This implies that as long as the data rate (in symbols/s/Hz, excluding training symbols and the padded zeros) is not larger than $N/(N+L)$ times that of the channel symbol rate, blind channel identification based on SOS of the symbol-rate sampled channel output is possible. Theoretically, as long as the channel coherence time is long enough, we can choose $N \gg L$ so that $N/(N+L)$ can be arbitrarily close to 1.

We observe that in [37], it is shown that with nonconstant modulus precoding, blind channel estimation based only on the SOS of symbol-rate sampled output can be performed for SISO system by exploiting transmission-induced cyclostationarity. Taking into consideration that transmitter-induced cyclostationarity through nonconstant modulus precoding generally implies slight sacrifice on spectral efficiency, as it may reduce the minimum distance of the symbol constellation, our result is consistent with that in [37]. Some related results on transmitter precoder design can be found in [38, 39].

3.2. *Systems with multiple receive antennas*

For systems with two or more receive antennas, channel estimation can be performed at each receiver independently or from more than one receiver 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.

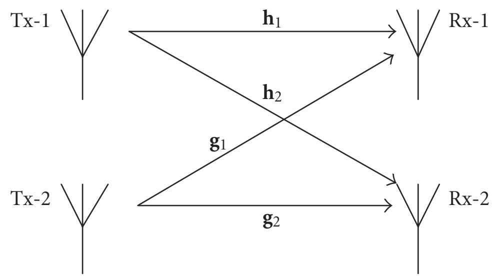


Figure 2: Two-branch transmit diversity with two receive antennas.

Take a synchronous 2×2 system as an example (see Figure 2). Define $\mathbf{H}_1$, $\mathbf{H}_2$ as in (5) and $\mathbf{G}_1$, $\mathbf{G}_2$ as in (6), corresponding to $\mathbf{h}_1$, $\mathbf{h}_2$, $\mathbf{g}_1$, $\mathbf{g}_2$, respectively. If $[\overline{\mathbf{a}}_1, \overline{\mathbf{a}}_2, \ldots, \overline{\mathbf{a}}_M]$ is transmitted through $\mathbf{h}_1$, $\mathbf{h}_2$, and $[\overline{\mathbf{b}}_1, \overline{\mathbf{b}}_2, \ldots, \overline{\mathbf{b}}_M]$ is transmitted through $\mathbf{g}_1$, $\mathbf{g}_2$, the received signal at receivers 1 and 2 can be expressed as

$$\mathbf{z}_k^1 = [\mathbf{H}_1, \mathbf{G}_1] \begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix} + \mathbf{n}_k^1, \qquad \mathbf{z}_k^2 = [\mathbf{H}_2, \mathbf{G}_2] \begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix} + \mathbf{n}_k^2, \tag{25}$$

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 2. Stacking $\mathbf{z}_k^1$, $\mathbf{z}_k^2$ into a $2(N+L+1)$-vector, we obtain

$$\mathbf{z}_k^L = \begin{bmatrix} \mathbf{z}_k^1 \\ \mathbf{z}_k^2 \end{bmatrix} = \underbrace{\begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix}}_{\triangleq \mathbf{F}} \underbrace{\begin{bmatrix} \mathbf{a}_k \\ \mathbf{b}_k \end{bmatrix}}_{\triangleq \mathbf{s}_k} + \begin{bmatrix} \mathbf{n}_k^1 \\ \mathbf{n}_k^2 \end{bmatrix}. \tag{26}$$

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

$$\mathbf{R}_z^L = E\Big\{ \mathbf{z}_k^L \big(\mathbf{z}_k^L\big)^H \Big\} = \mathbf{F}\mathbf{F}^H + \sigma^2 \mathbf{I}_{2(N+L+1)}. \tag{27}$$

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 $\mathbf{R}_z^L$.

4. EQUALIZATION

Once channel estimation has been carried out, equalization can be performed in several ways. Take the 2×2 as an example, define $\mathbf{s}_k = [\mathbf{a}_k^T, \mathbf{b}_k^T]^T$ as before, it follows from (26) that the information blocks $\mathbf{a}_k$ and $\mathbf{b}_k$ can be estimated by

$$\min_{\mathbf{s}_k} \big|\big| \mathbf{z}_k^L - \mathbf{F}\mathbf{s}_k \big|\big|, \tag{28}$$

either using the least-squares (LS) method, the zero-forcing (ZF) equalizer, or through the maximum-likelihood (ML) approach based on the Viterbi algorithm. More specifically, if finite alphabet constraint is put on s_k, then (28) can be solved using the Viterbi algorithm; if this constraint is relaxed, then s_k can be obtained through the LS or ZF equalizer. In the simulations, we choose to use the ZF equalizer.

For systems with two transmit antennas and one receiver, it follows from (5), (6), and (7) that

$$\mathbf{z}_k = [\mathbf{H}, \mathbf{G}]\mathbf{s}_k + \mathbf{n}_k, \tag{29}$$

and $[\mathbf{H}, \mathbf{G}]$ is $(N + L + 1) \times 2N$. The necessary condition for $[\mathbf{H}, \mathbf{G}]$ to be of full-column rank is $N + L + 1 \geq 2N$, that is, $N \leq L + 1$. Here we choose $N = L + 1$ to maximize the spectral efficiency. This implies that the overall data rate (in symbols/s/Hz) of the two-branch transmission system with one receiver will be the same as that of the corresponding single-transmitter and single-receiver system under the same modulation scheme. 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 dual-branch structured transmit delay scheme, blind channel identification and signal detection can be performed with the overall data rate much higher than that of the corresponding SISO system. For a 2×2 system in a slow time-varying environment, for example, blind channel identification and signal detection can be achieved with a data rate (in symbols/s/Hz, excluding training symbols and dummy zeros) of $2N/(N + L + 1)$ times that of the corresponding SIMO system under the same modulation scheme and with no training symbols transmitted.

5. EXTENSION OF THE STRUCTURED TRANSMIT DELAY SCHEME TO GENERAL MIMO SYSTEMS

In this section, extension of the proposed structured transmit delay scheme to general MIMO systems is discussed. We start with the dual-branch transmission systems where the two branches are either synchronous or asynchronous, with unknown transmission delays, and then consider the extension to systems with multiple-(more than two) transmit antennas.

5.1. *Dual-branch transmitter with unknown initial transmission delays*

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 $(L+d+1)\times 1$ vectors,

$$\begin{aligned}\mathbf{h} &= [h(-d_1), h(-d_1+1), \dots, h(0), h(1), \dots, h(L+d-d_1)],\\ \mathbf{g} &= [g(-d_2), g(-d_2+1), \dots, g(0), g(1), \dots, g(L+d-d_2)],\end{aligned} \tag{30}$$

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

(i) *Initial transmission delays are unknown, and the two branches are synchronous* ($0 \leq d_1 = d_2 \leq d$).

Define

$$\begin{aligned}\overline{\mathbf{a}}_k &= \Big[a_k(0), a_k(1), \dots, a_k(N-1), \underbrace{0, \dots, 0}_{L+d+1}\Big],\\ \overline{\mathbf{b}}_k &= \Big[0, b_k(0), b_k(1), \dots, b_k(N-1), \underbrace{0, \dots, 0}_{L+d}\Big].\end{aligned} \tag{31}$$

Suppose that $[\overline{\mathbf{a}}_1, \overline{\mathbf{a}}_2, \dots, \overline{\mathbf{a}}_M]$ are transmitted from antenna 1 through channel $\mathbf{h}$, and $[\overline{\mathbf{b}}_1, \overline{\mathbf{b}}_2, \dots, \overline{\mathbf{b}}_M]$ from antenna 2 through channel $\mathbf{g}$, please refer to Figure 1. For simplicity of the notation, we consider the system with a single-receive antenna. In this case, following our notations in Figure 1, we have

$$\begin{aligned}x_k(n) &= \sum_{l=0}^{L+d} h(l-d_1)a_k(n-l),\\ y_k(n) &= \sum_{l=0}^{L+d} g(l-d_2)b_k(n-l-1).\end{aligned} \tag{32}$$

Define $\mathbf{x}_k = [x_k(0), x_k(1), \dots, x_k(N+L+d)]^T$, $\mathbf{y}_k = [y_k(0), y_k(1), \dots, y_k(N+L+d)]^T$, and $\mathbf{n}_k = [n_k(0), n_k(1), \dots, n_k(N+L+d)]^T$. Define $\mathbf{z}_k = \mathbf{x}_k + \mathbf{y}_k + \mathbf{n}_k$, then it follows that

$$\mathbf{z}_k = \begin{bmatrix} d_1\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \\ h(0)a_k(0) \\ h(1)a_k(0) + h(0)a_k(1) + g(0)b_k(0) \\ h(2)a_k(0) + h(1)a_k(1) + h(0)a_k(2) \\ +g(1)b_k(0) + g(0)b_k(1) \\ \vdots \\ h(L)a_k(0) + \cdots + h(0)a_k(L) \\ +g(L-1)b_k(0) + \cdots + g(0)b_k(L-1) \\ \vdots \\ h(L)a_k(N-1) + g(L)b_k(N-2) \\ +g(L-1)b_k(N-1) \\ g(L)b_k(N-1) \\ d-d_1\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \end{bmatrix} + \mathbf{n}_k. \tag{33}$$

As can be seen from (31), $\overline{\mathbf{a}}_k$ and $\overline{\mathbf{b}}_k$ are defined in the similar manner as that in Section 2. The only difference is that due to the unknown delays, $L + d + 1$ zeros are inserted to each block instead of $L + 1$ zeros as in the delay known case, in order to ensure that there is no interblock interference. At the same time, this design guarantees that in $\mathbf{z}_k$, there is an interference-free item $h(0)a_k(0)$, which plays a critical role in blind channel estimation, please refer to Section 3. Delay estimation will be discussed later on.

(ii) *Initial transmission delays are unknown, and the two branches are asynchronous* ($0 \leq d_1, d_2 \leq d, d_1 \neq d_2$).

Define

$$\begin{aligned}\overline{\mathbf{a}}_k &= \Big[a_k(0), a_k(1), \dots, a_k(N-1), \underbrace{0, \dots, 0}_{L+2d+1}\Big],\\ \overline{\mathbf{b}}_k &= \Big[\underbrace{0, \dots, 0}_{d+1}, b_k(0), b_k(1), \dots, b_k(N-1), \underbrace{0, \dots, 0}_{L+d}\Big].\end{aligned} \tag{34}$$

Again, transmitting $[\overline{\mathbf{a}}_1, \overline{\mathbf{a}}_2, \ldots, \overline{\mathbf{a}}_M]$ from antenna 1 through channel $\mathbf{h}$, and transmitting $[\overline{\mathbf{b}}_1, \overline{\mathbf{b}}_2, \ldots, \overline{\mathbf{b}}_M]$ from antenna 2 through channel $\mathbf{g}$, it turns out that

$$\begin{aligned} x_k(n) &= \sum_{l=0}^{L+d} h(l-d_1)a_k(n-l), \\ y_k(n) &= \sum_{l=0}^{L+d} g(l-d_2)b_k(n-l-d-1). \end{aligned} \tag{35}$$

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

$$\mathbf{z}_k = \mathbf{x}_k + \mathbf{y}_k + \mathbf{n}_k, \tag{36}$$

where

$$\mathbf{x}_k = \begin{bmatrix} d_1\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \\ h(0)a_k(0) \\ h(1)a_k(0)+h(0)a_k(1) \\ h(2)a_k(0)+h(1)a_k(1)+h(0)a_k(2) \\ \vdots \\ h(L)a_k(0)+\cdots+h(0)a_k(L) \\ \vdots \\ h(L)a_k(N-1) \\ 2d-d_1+1\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \end{bmatrix},$$

$$\mathbf{y}_k = \begin{bmatrix} d+d_2+1\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \\ g(0)b_k(0) \\ g(1)b_k(0)+g(0)b_k(1) \\ g(2)b_k(0)+g(1)b_k(1)+g(0)b_k(2) \\ \vdots \\ g(L)b_k(0)+\cdots+g(0)b_k(L) \\ \vdots \\ g(L)b_k(N-1) \\ d-d_2\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \end{bmatrix}. \tag{37}$$

An insight into the design in (34) is provided through two extreme cases. First, consider the case where $d_1 = d$, $d_2 = 0$. It turns out that

$$\mathbf{z}_k = \begin{bmatrix} d\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \\ h(0)a_k(0) \\ h(1)a_k(0)+h(0)a_k(1)+g(0)b_k(0) \\ h(2)a_k(0)+h(1)a_k(1)+h(0)a_k(2) \\ +g(1)b_k(0)+g(0)b_k(1) \\ \vdots \\ h(L)a_k(0)+\cdots+h(0)a_k(L) \\ +g(L-1)b_k(0)+\cdots+g(0)b_k(L-1) \\ \vdots \\ h(L)a_k(N-1)+g(L)b_k(N-2) \\ +g(L-1)b_k(N-1) \\ g(L)b_k(N-1) \\ d\left\{\begin{matrix}0\\ \vdots\\ 0\end{matrix}\right. \end{bmatrix} + \mathbf{n}_k. \tag{38}$$

Second, exchange the role in the previous example and consider the case where $d_1 = 0$ and $d_2 = d$, then we have

$$\mathbf{z}_k = \begin{bmatrix} h(0)a_k(0) \\ \vdots \\ g(L)b_k(N-1) \end{bmatrix} + \mathbf{n}_k. \tag{39}$$

From the above two extreme cases, we can see that structured transmit delay is elaborately designed in (34) to ensure that there is no interblock interference and that there is an interference-free term to allow simple blind channel estimation. At the receiver end, to retrieve the channel status information, we still rely on the covariance matrix $\mathbf{R}_z$. In the absence of noise, the first d_1 rows of $\mathbf{R}_z$ are all zeros. The (d_1+1)th row contains all the information needed to estimate $\mathbf{h}$, that is, $[\ldots, |h(0)|^2, h(0)h(1)^*, \ldots, h(0)h(L)^*, \ldots]$. In the presence of noise, the first (d_1+1) rows of $\mathbf{R}_z$ become

$$\mathbf{R}_z[1:d_1+1,:] = \begin{bmatrix} \sigma_2 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & \cdots & 0 \\ 0 & \sigma_2 & 0 & \cdots & 0 & 0 & 0 & 0 & 0 & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \ddots & \sigma_2 & 0 & 0 & 0 & 0 & 0 & \cdots & 0 \\ 0 & \cdots & 0 & 0 & |X_0|^2+\sigma 2 & X_0X_1^* & \cdots & X_0X_L^* & 0 & \cdots & 0 \end{bmatrix}, \tag{40}$$

where

$$X_0 = h(0), \quad X_1 = h(1), \quad X_L = h(L). \tag{41}$$

Clearly, delay d_1 can be estimated from the first (d_1+1) rows of $\mathbf{R}_z$ since $|h(0)|^2 + \sigma_2 > \sigma_2$. Similarly, delay d_2 can be estimated by exchanging the role of (34). Recall that M

is the number of blocks in a frame, without loss of generality, assume that M is even. We transmit $[\bar{\mathbf{a}}_1, \bar{\mathbf{a}}_2, \ldots, \bar{\mathbf{a}}_{M/2}, | \bar{\mathbf{b}}_{M/2+1}, \bar{\mathbf{b}}_{M/2+2}, \ldots, \bar{\mathbf{b}}_M]$ from antenna 1 through channel $\mathbf{h}$, and $[\bar{\mathbf{b}}_1, \bar{\mathbf{b}}_2, \ldots, \bar{\mathbf{b}}_{M/2}, | \ \bar{\mathbf{a}}_{M/2+1}, \bar{\mathbf{a}}_{M/2+2}, \ldots, \bar{\mathbf{a}}_M]$ from antenna 2 through channel $\mathbf{g}$. After the transmission delays are estimated, noise variance estimation and channel estimation for systems with either one or more receive antenna(s) follow directly from our discussion in Section 3.

5.2. Extension to systems with more than two transmit antennas

Extension to systems with multiple-(more than two) transmit antennas is not unique. Here we illustrate one possibility by taking a three-branch transmitter as an example. First, we convert the input sequence into three parallel data streams and partition each data stream into N-symbol blocks $\mathbf{a}_k$, $\mathbf{b}_k$, $\mathbf{c}_k$. In the case when the system is synchronous and the transmission delay is known (other cases can be extended similarly), define

$$
\begin{aligned}
\bar{\mathbf{a}}_k &= \Big[a_k(0), a_k(1), \ldots, a_k(N-1), \underbrace{0, \ldots, 0}_{L+1}\Big], \\
\bar{\mathbf{b}}_k &= \Big[0, b_k(0), b_k(1), \ldots, b_k(N-1), \underbrace{0, \ldots, 0}_{L}\Big], \\
\bar{\mathbf{c}}_k &= \Big[0, c_k(0), c_k(1), \ldots, c_k(N-1), \underbrace{0, \ldots, 0}_{L}\Big].
\end{aligned}
\tag{42}
$$

Assume that there are $2M$ blocks in a frame, transmit

$$
\begin{aligned}
&[\bar{\mathbf{a}}_1, \ldots, \bar{\mathbf{a}}_M, | \ \bar{\mathbf{b}}_{M+1}, , \ldots, \bar{\mathbf{b}}_{2M}], \\
&[\bar{\mathbf{b}}_1, \ldots, \bar{\mathbf{b}}_M, | \ \bar{\mathbf{a}}_{M+1}, , \ldots, \bar{\mathbf{a}}_{2M}], \\
&[\bar{\mathbf{c}}_1, \ldots, \bar{\mathbf{c}}_M, | \ \bar{\mathbf{c}}_{M+1}, , \ldots, \bar{\mathbf{c}}_{2M}],
\end{aligned}
\tag{43}
$$

from three antennas, through channels $\mathbf{h}_1$, $\mathbf{h}_2$, $\mathbf{h}_3$, respectively. Again, our transmit delay structure here is designed to avoid interblock interference and to ensure that there is an interference-free item for simple blind channel estimation. Channel $\mathbf{h}_1$ can be estimated through the covariance matrix of the received signal obtained from the first M blocks in the frame, and channels $\mathbf{h}_2$ and $\mathbf{h}_3$ can be estimated from the second half of the frame. Compared with the two-branch case, the block size N needs to be kept small enough such that a $2M$-block interval is less than the channel coherence time. In view of the computational complexity and the essential improvement on spectral efficiency, the number of transmit antennas to be used in the system would be channel-and application-dependent.

6. SIMULATION RESULTS

In this section, simulation results are provided to illustrate the performance of the proposed approach. In the simulation examples, each antenna transmits BPSK signals and 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. Systems with different block sizes are tested. It will be seen that as the block size gets larger, we get more accurate approximation of desired statistics, and hence obtain better results.

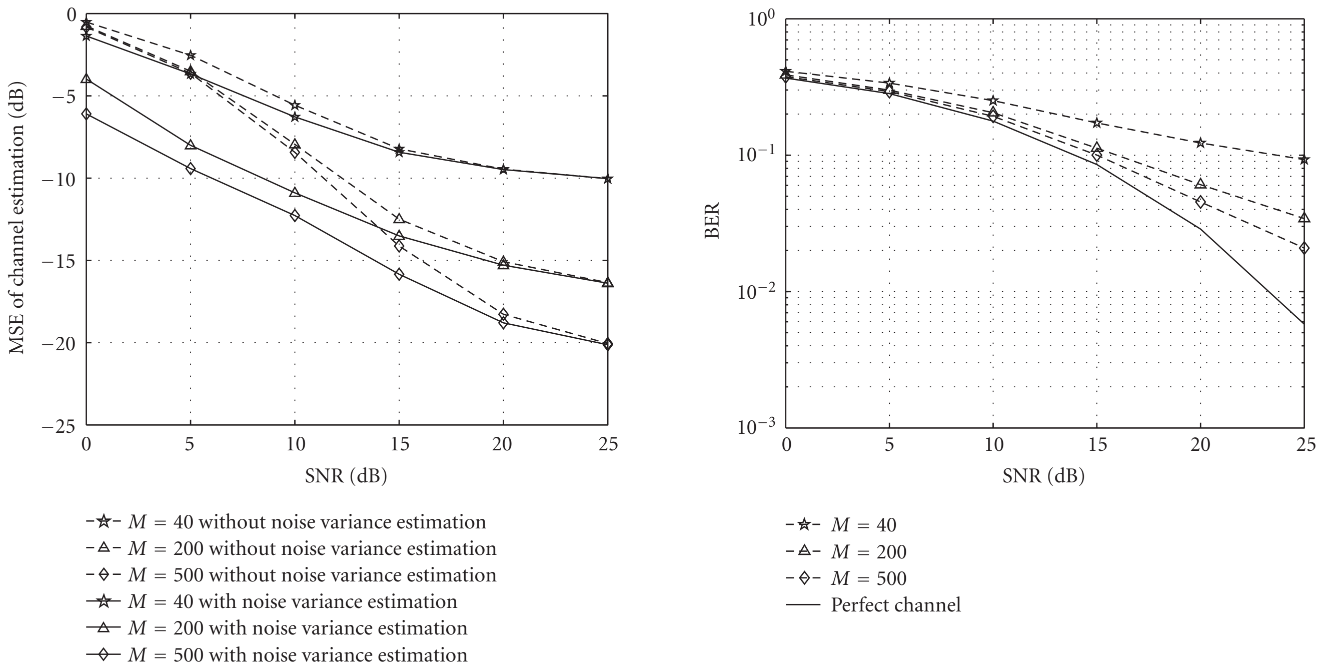
In the simulation,

$$
\text{normalized channel estimation MSE} \triangleq \frac{1}{I}\sum_{i=1}^{I} ||\hat{\mathbf{h}}_i - \mathbf{h}_i||^2 / ||\mathbf{h}_i||^2,
\tag{44}
$$

where $\mathbf{h}_i$ and $\hat{\mathbf{h}}_i$ denote the estimated channel and the true channel in the ith run, respectively. I is the total number of Monte Carlo runs. At each receive antenna, SNR is defined as the ratio between the total received signal power and the noise power, and it is assumed that the receive antennas have the same SNR level. For systems with a single-receive antenna, we choose $N = L + 1$, resulting in an overall data rate of 1. For systems with two receive antennas, 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. In all examples, the zero-forcing equalizer is used for signal detection. As it is well known, blind equalization can only be achieved up to an unknown constant phase and delay. In the simulation, the phase ambiguity is resolved by adding one pilot symbol for every M block at each transmit antenna. All the simulation results are averaged over $I = 500$ Monte Carlo runs. In the following, three examples are considered.

Example 1 (synchronous two-branch transmission with a single receiver). The multipath channels are assumed to have 6 rays, the amplitude of each ray is zero-mean complex Gaussian with unit variance, the first ray has no initial delay, and the delays for the remaining 5 rays are uniformly distributed over $[1, 5]$ symbol periods. Figure 3 shows the MSE of blind channel estimation both with and without noise variance estimation, and resulted BER (with noise variance estimation only) for various block sizes. It can be seen that (i) good noise variance estimation results in significantly more accurate channel estimation, (ii) when the number of blocks, M, increases, better results are achieved as the time-averaged statistics approach their ensemble values, (iii) BER is not satisfying even if the channel estimation is good, as there is only one-receive antenna.

The performance of the proposed channel estimation algorithm versus the relative power of the first tap has been provided in Figure 4. In the simulation, the multipath channel is assumed to have 6 rays, each is complex Gaussian with zero mean and variance 1/6. The results are averaged over 500 Monte Carlo runs where the channel is randomly generated for each run. These 500 channels are grouped into two classes: (a) power of the first tap lower than the average tap power, and (b) power of the first tap larger than the average tap power. As expected, class (b) delivers better result.



(a) (b)

Figure 3: Example 1, synchronous two-branch transmission with a single receiver, $N = L+1$, data rate = 1, (a) normalized channel estimation MSE versus SNR, (b) BER versus SNR (with noise variance estimation).

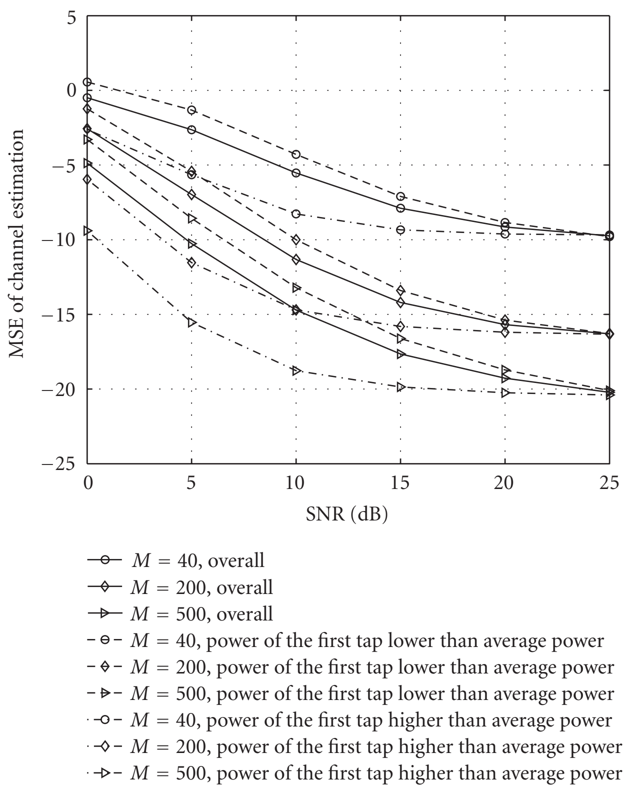


Figure 4: Performance sensitivity of the proposed channel estimation method to the relative power of the first tap.

Example 2 (*synchronous two-branch transmission with two receivers*). In this example, the channels have the same characteristics as in Example 1, channel estimation is carried out at each receive antenna independently with and without noise variance estimation, and noise variance estimation is performed as described in Section 3.2. It should be noted that although it is possible to obtain good channel estimation results with one-receive antenna, two-receive antennas are necessary to recover the original inputs when we are transmitting at a higher data rate (1.5 times that of the corresponding SISO system).

We also consider to improve the system performance by combining the threaded layered space-time (TST) architecture [17] with the proposed transmission scheme, as shown in Figure 5. Here “SI” stands for *spatial interleaver*, and “Int” for *interleaver*. Turbo encoder is chosen for forward error control. At the receiver, hard decisions are made on the equalizer outputs, and there is no iteration between the receiver front end and the turbo decoder. The decoding algorithm is chosen to be log-MAP [40]. The number of decoding iterations is set to be 5, and no early termination scheme is applied. The rate of the turbo code is 1/2. The block length is 6000. The generation matrix of the constituent code is given by $[1, (7)_{\text{octal}}/(5)_{\text{octal}}]$, where $(7)_{\text{octal}}$ and $(5)_{\text{octal}}$ are the feedback and feedforward polynomials with memory length 2, respectively. After space-domain interleaving and time-domain interleaving, the symbols transmitted from each antenna will be independent with each other and with the symbols transmitted from other antennas.

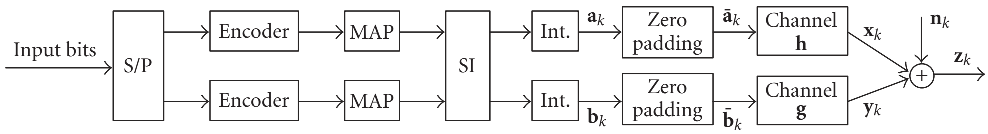


Figure 5: Space-time diversity with the threaded layered space-time (TST) architecture.

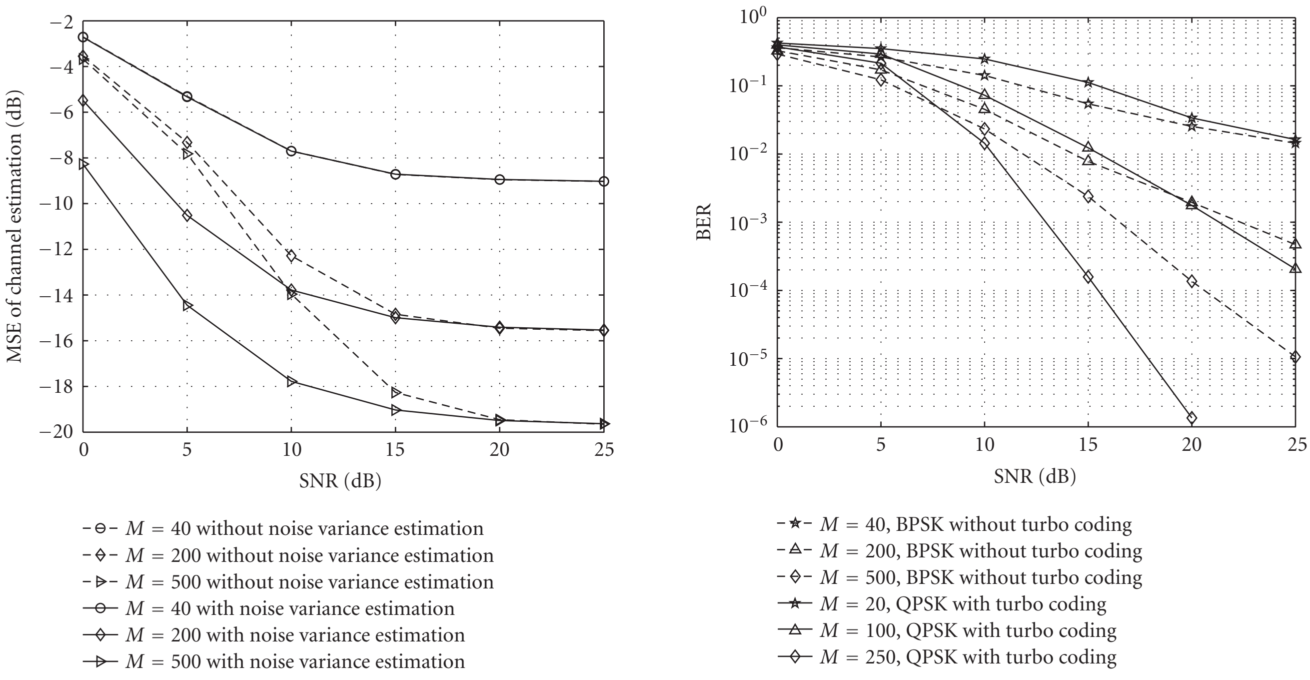


Figure 6: Example 2, synchronous two-branch transmission with two receivers, $N = 3(L+1)$, data rate = 1.5, (a) normalized channel estimation MSE versus SNR, (b) BER versus SNR (with noise variance estimation).

The proposed transmission scheme, therefore, can be directly concatenated with the TST structure. From Figure 6, it can be seen that the 2-by-2 systems can achieve much better BER at a higher data rate. Significant improvement can be observed when turbo coding and TST structure are employed. For fair comparison, when turbo encoder is added, the system transmits QPSK signals instead of BPSK signals, which are used for the systems without turbo coding.

Example 3 (*asynchronous two-branch transmission with two receivers*). In this example, system performance is tested in two cases: (i) transmission delays are known, (ii) transmission delays are unknown. 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. That is, the maximum initial delay is $d = 2$, and the maximum multipath delay spread is $L = 5$. We use the same turbo encoder as in Example 2. As can be seen from Figure 7, when the delays are unknown, for the first arrival to be detected, the signal power of the first path should be sufficiently large in comparison with the noise power. And when SNR $\geq$ 10 dB, channel estimation with unknown delays is as good as that with known delays.

7. CONCLUSIONS

In this paper, a dual-branch transmission scheme that utilizes guard intervals for blind channel estimation and equalization is proposed. Unlike existing transmit diversity schemes, in which each antenna transmits different versions of the same signal, the proposed transmission scheme promises higher data rate since each antenna transmits an independent data stream. 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. This can be

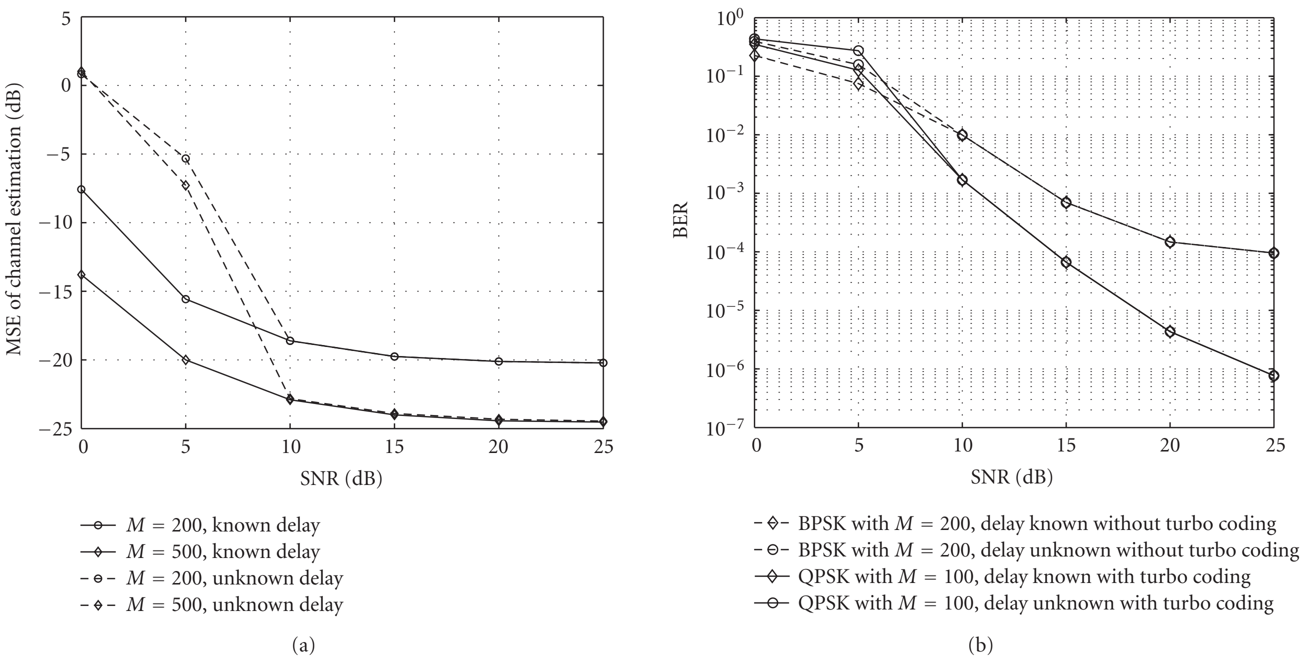


FIGURE 7: Example 3, asynchronous two-branch transmission with two receivers, $N = 3(L+1)$, data rate = 1.5, with noise variance estimation, (a) normalized channel estimation MSE versus SNR, (b) BER versus SNR when $M = 200$.

regarded as a counterpart of [6] which exploits *receive diversity*. The proposed approach involves no preequalization and has no limitations on channel-zero locations. Higher system capacity can be achieved with better performance when more than one-receive antennas are available. It is also shown that in combination with the TST structure and turbo coding, significant improvement can be observed in the overall system performance.

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