An Embedded Space-Time Coding (STC) Scheme for Broadcasting

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***Abstract*—An embedded space-time coding method is proposed for wireless multimedia broadcast with heterogeneous receivers. In the proposed system, a transmitter sends out multi-layer source signals by encoding different layers with different space-time codes. Then, a receiver can retrieve a different amount of information depending on the number of its antennas. The receiver with only one antenna can decode only the base layer information with a low complexity while the receiver with more antennas can retrieve more layers of information. We also investigate a differential design of the embedded space-time system, which is used when the channel state information (CSI) is unknown to the receiver. Both analytic and experimental results are provided to demonstrate the performance of the proposed system. Differential detection with Kalman filtering is also investigated to help improve performance.**

***Index Terms*—Broadcast channel, differential design, space-time codes (STC).**

I. Introduction

THE PROGRESS in signal processing and communication technologies has enabled digital wireless broadcast applications, leading to the commercialization of high-definition television (HDTV), digital video broadcast (DVB), and digital audio broadcast (DAB). In this paper, we investigate a wireless broadcast system with multiple transmit antennas that supports the transmission of a scalable media source.

Wireless channels are susceptible to environmental noise, and can be degraded due to the motion of the mobile station. The concept of diversity has been developed to overcome these channel impairments, which is a technique to provide the receiver several replicas of signals over independent fading channels [1]. The antenna diversity, which uses more than one antenna in the transmitter and/or the receiver, is also capable of providing the diversity for wireless transmission. Recently, space-time coding (STC) has been extensively studied to exploit the transmission antenna diversity, which uses more than one antenna in the transmitter. STC integrates the antenna diversity with coding techniques to achieve a higher capacity and reduce co-channel interference in multiple access. Tarokh *et al.* [2] derived an analytical bound for the symbol error rate and presented the criteria for STC that can achieve the maximum diversity.

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A broadcast system differs from the point-to-point transmission in that different receivers can have different receiving capabilities. In [3], the performance of a broadcast channel was analyzed in an information theoretic framework. In [4], a multi-resolution modulation technique was proposed to provide different receiving quality according to the distance between the transmitter and the receiver. The source is divided into several layers so that the receiver with a lower signal power can still reconstruct the base-layer information while the receiver with a higher signal power can get all the information.

Designers of conventional communication systems consider each information bit and each user equally important. Thus, they have focused on maximizing the channel capacity or minimizing the bit error rate. However, in the context of multimedia communication, the importance of each bit may vary according to its carried content. For example, in the video bitstream, the header is more important than the content part, since a little amount of corruption in the header may lead to totally wrong results. In the layered coding technique, some layers are more important than others. The basic layer is usually critical to the visual quality while enhancement layers can be discarded with much less impact on the perception of end users. Also, different users may have different requirements. Some users may wish to gain higher quality at a higher cost while others may only demand the minimum level of quality of service (QoS). These differences among users and QoS requirements are important in a commercial system. That is our motivation to find a new structure of space-time codes that is applicable to systems with different QoS requirements.

The layered or scalable source coding technique has become mature in the last few years. Two well-known scalable standards are the JPEG 2000 [5] image compression standard and the MPEG-4 fine granularity scalability (FGS) [6], [7] format. The encoded bitstream has the property that the receiver can reconstruct the source with a rate-distortion tradeoff, even if only parts of the data are received. The scalable coding technique has inspired the exploration of many Internet applications, since the server can perform multicast and/or broadcast with a less storage space. Some researchers have also attempted to apply STC to multimedia transmission. For example, Zheng and Liu [8] proposed a layered multimedia transmission scheme over a STC system. Their focus was on the power allocation over antennas, loaded with different layers of multimedia data contents, instead of exploiting the structure of STC.

In this research, we study a special form of STC for multimedia transmission that utilizes the property of space-time codes and layered source coding. Specifically, we consider the case that a receiving terminal may have a different number of antennas. Some receivers may have more antennas to receive higher quality signals while some may not have an enough

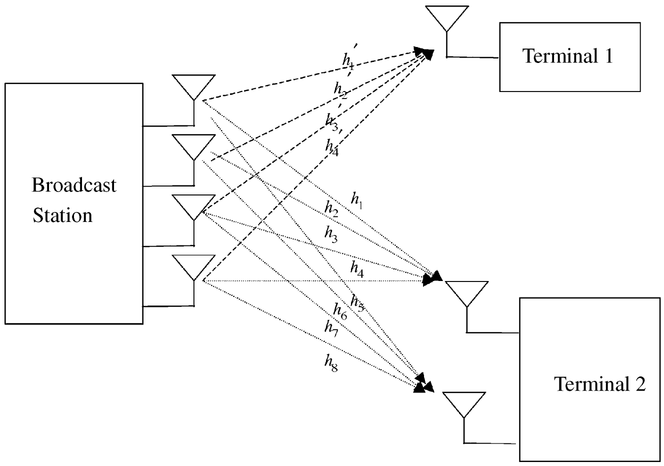


Fig. 1. The channel model of a broadcast system with multiple antennas.

number of antennas due to the cost consideration or the physical size constraint (for example, the mobile handset or wireless PDA) so that they may communicate at the lowest quality level.

The proposed embedded STC scheme may carry several layers of information. In addition to transmission of the layered source bitstream, it can be used in heterogeneous broadcast services. For example, if the content of a DVB-Handheld (DVB-H) service is a subset of DVB-Terrestial (DVB-T), we can transmit the DVB-H bitstream in the base layer that can be reconstructed by handheld devices with less complexity and lower power consumption. The other part of the DVB-T content can be broadcast in the same bandwidth to the base station using enhanced layers, which are designed to be received by higher end receivers.

The rest of this paper is organized as follows. The channel model is described in Section II. In Section III, embedded space-time codes are proposed and discussed. Differential design of embedded STC is investigated in Section IV. Finally, the paper is summarized, and the future direction is discussed in Section V.

II. Channel Model

The channel model for a wireless broadcast system with two different receivers is depicted in Fig. 1. Let the broadcast station encode the media source by space-time block codes (STBC) and send them out with multiple antennas. Receivers are allowed to have a different number of antennas. Each receive antenna receives signals from transmit antennas through different paths. In the next section, it is assumed that these paths are statistically independent, and signals experience slow fading such that channel coefficients are unchanged during the transmission of a single space-time block code. The fading coefficients are assumed to be perfectly estimated in Section III. In Section IV it is assumed that the fading coefficients are not known.

Suppose that the broadcast system has N transmit antennas and M receive antennas. Let T denote the block length. Then, each code word C is a $N \times T$ matrix. The received signal matrix is of size $M \times T$ and represented by $R = \sqrt{\gamma_t} HC + N_o$, where $\gamma_t = \gamma/N$ with γ being the Signal to Noise Ratio (SNR) level. The element $h_{m,n}$ of $M \times N$ matrix H represents channel coefficients, and N_o is the additive white-Gaussian noise matrix.

The purpose of this work is to design a space-time block code that can carry a multi-layered source bitstream so that the receiver with a different number of antennas can recover a different amount of information. This means that the receiving diversity is used to differentiate the service type. For example, if the source is encoded with two layers, then terminal no. 1 with one antenna can only decode the first layer while terminal no. 2 with two antennas can decode both layers successfully since more diversity can be exploited with more receive antennas.

III. Embedded Space-Time Codes

A. Transmitter Design

Suppose that the bitstream is modulated by the phase shift keying. For the ith layer, the bitstream is mapped to a sequence of symbols by M-ary constellation, and then encoded into matrix C_i. We transmit the sum

$$C = C_1 + C_2 + \cdots + C_L \quad (1)$$

with multiple antennas. The receiver is expected to retrieve a different amount of data, depending on the number of antennas. More specifically, for the receiver with a single antenna, only C_1, hence the first layer, can be successfully decoded. For the n-antenna case $(1 \le n \le L)$, the first n layers are decoded. Let $P_i = \mathrm{Tr}(C_i C_i^*)$ denote the power of each layer code, where $\mathrm{Tr}(\cdot)$ is the trace operator. It is often that $P_1 > P_2 > \cdots > P_L$, since the first layer is the most important one and requires the highest transmission power.

Take a two-layer embedded STC system with four transmit antennas as an example. We use a repeated Alamouti code as the first layer code. This code satisfies the orthogonal design condition, thus reducing the complexity of the maximum likelihood decoder. Naguib *et al.* [9] proposed a method that transmits K Alamouti codes simultaneously. In this case, $2K$ interfering signals arrive at the receiver. They showed that K receive antennas are required to perfectly suppress the interference effect, and developed a minimum mean-square error (MMSE) interference suppression technique. We adopt this method as the second layer space-time code, which needs at least two antennas to perform the decoding.

Therefore, in the embedded STC system with four transmit antennas and two receive antennas, we transmit the signal in the following way. Suppose the first layer modulates x_1 and x_2 in the duration of one block code while the second layer transmits y_1, y_2, y_3 and y_4. Then, the first layer space-time code C_1 is

$$C_1 = \begin{bmatrix} x_1 & x_2 \\ x_1 & x_2 \\ x_2^* & -x_1^* \\ x_2^* & -x_1^* \end{bmatrix},$$

and the second layer code C_2 is

$$C_2 = \begin{bmatrix} y_1 & y_2 \\ y_2^* & -y_1^* \\ y_3 & y_4 \\ y_4^* & -y_3^* \end{bmatrix}.$$

The transmitter then sends $C = C_1 + C_2$ through the air interface with four antennas. Since the first layer is more important, we assign more power to the first layer symbols. Because of the

PSK modulation, elements in C_1 and C_2 have fixed amplitudes. The amplitude ratio ρ is defined as the ratio of the second layer amplitude to the first layer amplitude, *i.e.* $\rho = |y_j|/|x_i|$ for all possible i's and j's.

Two points are worthy of mentioning here. First, this coding scheme is designed for layered multimedia transmission, wherein layers may be of different importance and hence require different transmission power. This is different from other schemes that are designed to increase the data rate, though the concepts of 'layers' are adopted. For example, a special category of multi-stratum space-time coding was proposed in [10], [11]. In their construction, the input data bit-stream is partitioned into separate strata, which are then encoded using block STC and orthogonally superposed to enable a higher data rate. The sub-bitstreams of each stratum are treated with equal importance, and hence cannot be directly integrated into layered multimedia source coding. Second, this proposed embedded STC does not increase the transmission capacity. As a matter of fact, the signal of the first layer may be interferred by those of other layers and lead to a higher receiving bit error rate. The performance depends on the power allocation among layers whose power levels can be determined by the requirement of source bit streams for different layers.

B. Receiver Design

In the decoding of two layers, joint detection can provide optimal solutions for the embedded STC. This can be performed using the maximum-likelihood criterion with

$$\begin{aligned}\{\hat{C}_1, \hat{C}_2, \ldots, \hat{C}_L\} \\ = \arg \min_{\{C_1, C_2, \ldots, C_L\}} \operatorname{Tr}\left((R - HC)(R - HC)^*\right), \quad (2)\end{aligned}$$

where C is composed of layer codes from (1).

Although optimal detection can be achieved by jointly decoding two layers, it requires a high computational complexity. We develop a fast sub-optimal decoding algorithm for embedded STC with two layers based on the interference cancellation technique.

Now consider the receiver with one antenna. Let h_1, h_2, h_3, h_4 be channel coefficients of paths from four transmit antennas. In the first layer, each symbol is transmitted by two antennas. Thus, we can combine the corresponding two coefficients together and reconstruct the symbols with Alamouti's decoder [12], where the second layer symbols are treated as noise. More specifically, symbols x_1 and x_2 can be estimated as

$$\hat{x_1} = \kappa_2 \left[(h_1 + h_2)^* r_1 + (h_3 + h_4) r_2^*\right], \quad (3)$$
$$\hat{x_2} = \kappa_2 \left[(h_3 + h_4)^* r_1 - (h_1 + h_2) r_2^*\right], \quad (4)$$

where $\kappa_2 = 1/(|h_1 + h_2|^2 + |h_3 + h_4|^2)$ is the normalization factor. The estimated symbols are then detected with the maximum likelihood (ML) method to find the most suitable decisions.

For the case of a two-antenna receiver, there are eight paths in total from the transmitter to the receiver. Let h_1, h_2, h_3, h_4 denote the channel coefficients at the first receive antenna, and h_5, h_6, h_7, h_8 denote those at the second receive antenna. We first estimate the first layer symbols by the following combination:

$$\begin{aligned}\hat{x_1} &= \kappa_4 \times \left[(h_1 + h_2)^* r_1 + (h_3 + h_4) r_2^*\right. \\ &\quad \left. + (h_5 + h_6)^* r_3 + (h_7 + h_8) r_4^*\right], \\ \hat{x_2} &= \kappa_4 \times \left[(h_3 + h_4)^* r_1 - (h_1 + h_2) r_2^*\right. \\ &\quad \left. + (h_7 + h_8)^* r_3 - (h_5 + h_6) r_4^*\right]. \quad (5)\end{aligned}$$

where $\kappa_4 = 1/(|h_1 + h_2|^2 + |h_3 + h_4|^2 + |h_5 + h_6|^2 + |h_7 + h_8|^2)$. These signals are reconstructed with fading coefficients, and subtracted from the original input signals by

$$\begin{aligned}\Delta r_1 &= r_1 - (h_1 + h_2)\hat{x_1} - (h_3 + h_4)\hat{x_2}, \\ \Delta r_2 &= r_2 + (h_1 + h_2)\hat{x_2}^* - (h_3 + h_4)\hat{x_1}^*, \\ \Delta r_3 &= r_3 - (h_5 + h_6)\hat{x_1} - (h_7 + h_8)\hat{x_2}, \\ \Delta r_4 &= r_4 + (h_5 + h_6)\hat{x_2}^* - (h_7 + h_8)\hat{x_1}^*. \quad (6)\end{aligned}$$

Denote the residual signal vector by $\Delta \mathbf{r} = [\Delta r_1, \Delta r_2^*, \Delta r_3, \Delta r_4^*]^T$. In this way, the interference from the first layer can be eliminated if its estimation is correct. To estimate the second layer signals y's, the MMSE method is adopted to find a linear combination $\hat{y_i} = \mathbf{w}_i^* \Delta \mathbf{r}$, $i = 1, 2, 3, 4$, to minimize the mean square error of the estimation value. The weighting vector $\mathbf{w}_i$ can be obtained by $\mathbf{w}_i = [\mathbf{H}\mathbf{H}^* + 1/\gamma \mathbf{I}_4]^{-1} \mathbf{h}_i$, where H is the channel coefficient matrix

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_5 & h_6 & h_7 & h_8 \\ h_6^* & -h_5^* & h_8^* & -h_7^* \end{bmatrix},$$

and $\mathbf{h}_i$ is the ith column of $\mathbf{H}$, $\gamma = E_s/N_0$ is the signal-to-noise ratio. Then, we can detect y_i by finding the codeword closest to $\mathbf{w}_i^* \Delta \mathbf{r}$.

C. Analysis of Error Probability

In the proposed embedded STC system, the received signal for the one-antenna receiver can be written as

$$\mathbf{r} = \mathbf{H_x x} + \mathbf{H_y y} + \mathbf{n}, \quad (7)$$

where $\mathbf{r} = [r_1, r_2^*]^T$, $\mathbf{x} = [x_1, x_2]^T$, $\mathbf{y} = [y_1, y_2, y_3, y_4]^T$, $\mathbf{n} = [n_1, n_2^*]^T$, and

$$\mathbf{H_x} = \begin{bmatrix} h_1 + h_2 & h_3 + h_4 \\ h_3^* + h_4^* & -h_1^* - h_2^* \end{bmatrix},$$
$$\mathbf{H_y} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \end{bmatrix}.$$

Hence, estimation rules in (3) and (4) can be rewritten as $[\hat{x}_1, \hat{x}_2]^T = \kappa_2 \mathbf{H_x^* r} = \mathbf{x} + \eta + \mathbf{n}'$, where the term $\eta = \kappa_2 \mathbf{H_x^* H_y y}$ is the interference from the second layer, and $\mathbf{n}' = \kappa_2 \mathbf{H_x^* n}$ is an AWGN vector, whose element has the same variance $\sigma_{n'}^2 = \sigma_n^2/(|h_1 + h_2|^2 + |h_3 + h_4|^2)$.

To investigate the interference effect on the error probability, we model the interference η as a Gaussian random number.

Without loss of generality, we consider only the pairwise error probability of x_1, whose interference term η_1 has mean 0 and variance $\sigma_{\eta_1}^2 = E[\eta_1\eta_1^*] = \kappa_2\rho^2(|h_1|^2 + |h_2|^2 + |h_3|^2 + |h_4|^2)$. The total symbol error probability for M-ary modulation is upper-bounded by the following union bound of average pairwise error probability

$$P_e \leq (M-1)E \times \left[Q\left(\sqrt{\frac{|x-x'|^2\,(|h_1+h_2|^2+|h_3+h_4|^2)}{4\,[\rho^2\,(|h_1|^2+|h_2|^2+|h_3|^2+|h_4|^2)+\sigma_n^2]}} \right) \right], \quad (8)$$

where $E[\cdot]$ denotes the expectation over all channel coefficients h_i's, which are complex Gaussian random variables in the Rayleigh fading model. Note that the integration over eight dimensions is required in (8). It is difficult to find a closed form for this integration. Thus, we may find its approximate value by a computer program.

Now, for the receiver with two antennas, (7) still holds with some proper replacement of variables. The error probability of the second layer can be derived in a similar way as the first layer. However, the interference term is more complex and demands a more sophisticated approximation technique.

D. Experimental Results

We conducted a set of simulations for the proposed embedded STC system. The channel models and the transceiver mentioned above are simulated by a computer program written in C. In the transmitter, two layers of data are transmitted with four antennas. The input binary data of both layers are randomly generated and then modulated using the QPSK constellation. In two consecutive symbol durations, two symbols are transmitted at the first layer and four symbols at the second layer. All paths experience Rayleigh fadings with the Doppler coefficient 2.5×10^{-4}. The receiver requires two antennas to retrieve both layers while the one-antenna receiver can reconstruct only the first layer. All the SNR values reported in the following experiments represent the signal-to-noise ratios of the first layer signals.

Fig. 2(a) shows the bit error rate for the one-antenna receiver. Each curve represents the case of a fixed ρ, which is the amplitude ratio of the second-layer signal to the first layer signal. The curve labeled ‘Single layer’ represents the case that only the first layer is transmitted; namely, $\rho = 0$. Other curves show the BER performances of embedded STC systems by superimposing second layer signals with different magnitudes. As expected, the performance degrades as ρ increases. Notice that the bit-error rate in each curve with $\rho > 0$ does not decrease without limitation as SNR increases. That is, there is a saturation point. This indicates that the interference from the second layer limits the performance of the system.

The performance of the two-antenna receiver is shown in Figs. (2b) and (c) for the first and second layers, respectively . The two-antenna receiver achieves a lower bit error rate for the first layer data than the one-antenna receiver, since the two-antenna receiver yields a higher diversity gain. The bit error rate for the second layer decreases as ρ increases from 0.1 to 0.3. This can be understood as the increase of the second layer amplitude is equivalent to the increase of SNR. However, when ρ continues to increase to be above 0.4, the higher signal power of the second layer interferes the decoding of the first layer. Thus, the falsely detected first-layer data result in severe degradation in the decoding performance of the second layer.

We see from simulation results that the range of the amplitude ratio, ρ, is limited to provide a reasonable performance in the second layer. The range from 0.2 to 0.4 is acceptable, since the corresponding bit error rates fall in the range of 10^{-3} to 10^{-2}, when the SNR of the first layer is above 15 dB. Notice that the equivalent SNR of the second layer is lower than that of the first layer by $|20 \log \rho|$ dB (for $\rho < 1$). For example, if $\rho = 0.4$, the SNR of the second layer is 8 dB lower than that of the first layer. The high BER of the second layer is due to its low SNR value as well as the interference between layers. This could be tolerable since the information of the second layer is less important and error correction techniques can be used to enhance the quality of the received signal in the second layer at the receiver.

IV. Differential Design for Embedded Space-Time Coding

In this section, we consider the case when the channel state information is unknown. Differential encoding is adopted so that the receiver can recover the information even without explicit knowledge of the channel status.

A. Differential Design

Tarokh and Jafarkhani [13] proposed a differential detection scheme for a space-time system with two transmit antennas, which imposes a lower complexity on the receiver. Later, they generalized the construction scheme to systems with more transmit antennas in [14]. We will extend their work to the context of embedded space-time coding.

The main idea is that we apply the conventional differential coding scheme to the base layer encoding. Since it has higher power than signals in other layers, the receiver can detect the signal by ignoring those from other layers and treat them as interference. For other layers, the information is encoded with some relation to the first layer signal. This encoding method will allow a receiver with a sufficient number of antennas to decode higher layer signals in accordance with the decoded information obtained from the base layer.

The differential coding for embedded STC is briefly described as follows. At time t, the base-layer information is first mapped to a N-component vector $\mathbf{d}(t)$. Here, the transmission duration of the code block is denoted by T, which is set to N symbols, and transmit matrix $X(t)$ of size $N \times T$ has all elements coming from components in the N-component vector $\mathbf{x}(t)$. The relationship between $\mathbf{x}(t)$ and $\mathbf{d}(t)$ can be expressed as

$$\mathbf{x}(t) = X(t-1)\mathbf{d}(t). \quad (9)$$

Matrix $X(t)$ has to satisfy the unitary condition $X(t)X(t)^H = NI_N$, while the vector length of $\mathbf{x}(t)$ is fixed, *i.e.* $\mathbf{x}(t)^H\mathbf{x}(t) = N$. Under these conditions, the receiver can detect the signal by combining previously received signals. This can be seen more clearly from an example given in Section IV-B.

As to the ith $(2 \leq i \leq L)$ layer information, there is more freedom in choosing the encoding method. We choose the encoding method that depends on the first layer. The ith layer information is first encoded into matrix $E_i(t)$, which has the form

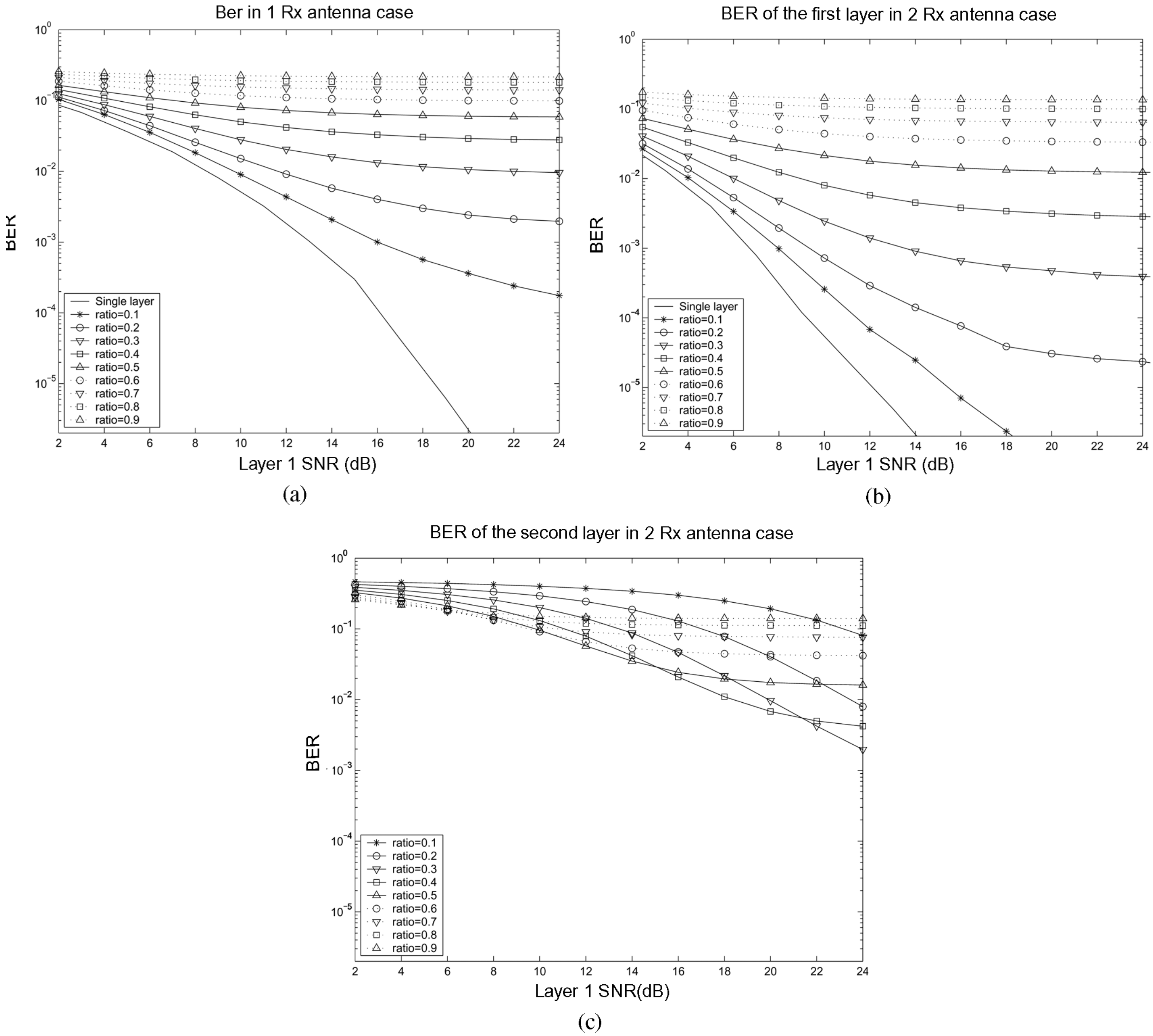


Fig. 2. The performance of bit-error rates in an embedded STC system: (a) the first layer of the one-antenna receiver, (b) the first layer for the two-antenna receiver, and (c) the second-layer for the two-antenna receiver.

that only those receivers with a sufficient number of antennas can decode successfully, as described in the previous section. The code $Y_i(t)$ representing the ith layer information is computed from $Y_i(t) = X(t)E_i(t)$. Finally, the resulting code to be sent through antennas is $Z(t) = X(t) + Y_2(t) + Y_3(t) + \cdots + Y_L(t)$. Since the importance decreases as i increases, the power constraint should be satisfied as in non-differential case, *i.e.* $P_2 > P_3 > \cdots > P_L$.

At the receiver, the first layer $X(t)$ can be easily retrieved, even signals in other layers are not detected. For those with more antennas, higher layer information $Y_i(t)$ can be decoded from signals received by different antennas.

B. An Example: Broadcast With Four Antennas

Consider a broadcast system using 4 transmit antennas. As described above, the encoding process can be viewed as a mapping to a vector space spanned by previous code matrix $X(t-1)$. Then, the receiver can retrieve the information via proper matrix computation, and this code can be decoded with only one receive antenna. The matrix of rank two can serve as the non-differential STC for the base layer. However, while we attempt to construct the second layer signal in the differential STC that has to be decoded by at least two receive antennas, matrix $X(t)$ should be invertible and hence the full rank of four is needed for this code matrix. To achieve this, the code block has to be transmitted over 4 symbol durations in the broadcast station with four transmit antennas.

To encode the first layer signal with $N = 4$ transmit antennas, we first map the input 4 bits to a signal vector denoted by $\mathbf{d}(t) = [d_1(t), d_2(t), d_3(t), d_4(t)]^T$. The codeword $\mathbf{d}(t)$ is taken from the set $\mathcal{S} = \{\mathbf{s}_i | 0 \le i \le 15\}$, which contains $2^4 = 16$ elements. The length of every element $\mathbf{s}_i \in \mathcal{S}$ should be unity, *i.e.* $\mathbf{s}_i^H \mathbf{s}_i = 1$, so that the length of $\mathbf{x}(t)$ is fixed. There are several possible mappings to construct set $\mathcal{S}$. We choose the set to be an obvious mapping from $[1,1,1,1]^T$ to the column space of any $X(t)$. After mapping a codeword, the differential modulation is obtained via (9). Since the length of vector $\mathbf{d}(t) \in \mathcal{S}$ is unity, the length of vector $\mathbf{x}(t)$ is fixed.

The construction of columns of code matrix $X(t)$ needs some permutation and re-arrangement of elements in vector $\mathbf{x}(t)$. Let us explicitly denote the code vector by

$$\mathbf{x}(t) = [x_1(t), x_2(t), x_3(t), x_4(t)]^T .$$

Furthermore, let us define four permutation operations below:

$$\begin{aligned}
\pi_1(\mathbf{x}(t)) &= [x_1(t), x_2(t), x_3(t), x_4(t)]^T, \\
\pi_2(\mathbf{x}(t)) &= [-x_2(t), x_1(t), -x_4(t), x_3(t)]^T, \\
\pi_3(\mathbf{x}(t)) &= [-x_3(t), x_4(t), x_1(t), -x_2(t)]^T, \\
\pi_4(\mathbf{x}(t)) &= [-x_4(t), -x_3(t), x_2(t), x_1(t)]^T .
\end{aligned}$$

Then, code matrix $X(t)$ is composed by column vectors as

$$X(t) = [\pi_1(\mathbf{x}(t)) | \pi_2(\mathbf{x}(t)) | \pi_3(\mathbf{x}(t)) | \pi_4(\mathbf{x}(t))] . \quad (10)$$

Note that elements in the above matrix are restricted to be real since it was proved in [15] that code matrices of a size greater than 2×2 cannot be complex in a differential STC system. This means that only PAM is needed. Since we have chosen $\mathbf{d}(t)$ to be real, matrix $X(t)$ would also be real with real initialization values.

As to the second layer, we map the codeword into 8 symbols $e_1(t), e_2(t), e_3(t), \ldots, e_8(t)$. Unlike the first layer, these symbols can be complex numbers. The information matrix for the second layer is encoded using the following equation

$$E(t) = \begin{bmatrix} e_1(t) & e_2(t)^* & e_5(t) & e_6(t)^* \\ e_2(t) & -e_1(t)^* & e_6(t) & -e_5(t)^* \\ e_3(t) & e_4(t)^* & e_7(t) & e_8(t)^* \\ e_4(t) & -e_3(t)^* & e_8(t) & -e_7(t)^* \end{bmatrix} .$$

One can easily find that the above construction is a concatenation of two codewords for the second layer signals as described in the non-differential scheme. This construction can be detected by the receiver with at least two antennas. In this construction, we also fix the amplitude of each element in $E(t)$ to be ρ. The actual codeword to be transmitted for the second layer is $Y(t) = X(t)E(t)$. The sum of two matrices $X(t) + Y(t) = X(t) + X(t)E(t) = Z(t)$ are then transmitted through 4 antennas.

Now, let us examine the decoding of the first layer information. Consider that four consecutive symbols $r_1(t)$, $r_2(t)$, $r_3(t)$ and $r_4(t)$ are received by a terminal with one antenna. For any receive antenna, suppose that channel coefficients are $h_i(t)$ from the ith transmit terminal $i = 1, 2, 3, 4$, and form the vector $\mathbf{h}(t) = [h_1(t), h_2(t), h_3(t), h_4(t)]^T$. The received signal at the kth time slot can be expressed as $r_k(t) = \mathbf{h}(t)^T \Pi_k(\mathbf{x}(t))$. Let $\mathbf{r}(t) = [r_1(t), r_2(t), r_3(t), r_4(t)]^T$. From this received signal $\mathbf{r}(t) = Z(t)^T \mathbf{h}(t) + \mathbf{n}(t)$, it can be shown that

$$\mathbf{r}(t) = R(t-1)\mathbf{d}(t) + \mathbf{r}_Y(t) + \mathbf{n}(t), \quad (11)$$

where $R(t) = [\pi_1(\mathbf{r}(t))| - \pi_2(\mathbf{r}(t))| - \pi_3(\mathbf{r}(t))| - \pi_4(\mathbf{r}(t))]^T$ and the term $\mathbf{r}_Y(t)$ denotes the signal generated from the second layer code matrix Y.

A simple combining rule can be derived by multiplying the received signal vector with the inverse of the matrix $R(t)$ to find the estimation of $\mathbf{d}(t)$. That is, $\hat{\mathbf{d}}(t) = R(t)^{-1}\mathbf{r}(t)$. However, the performance of this approach is not acceptable due to the interference from the second layer signal $r_Y(t)$. Instead, we adopt the MMSE method to suppress the interference. By approximating the second layer interference as additive Gaussian noise, the estimate of the information signal $\mathbf{d}$ can be computed as

$$\hat{\mathbf{d}}(t) = W^H(t-1)\mathbf{r}(t), \quad (12)$$

where $W(t) = \left(R(t)R(t)^H + \rho^2 I_4\right)^{-1} R(t)$, and where ρ is the amplitude of the components in the second layer code matrix.

For a terminal with two receive antennas, the above procedure will be conducted twice for two antennas, separately. Then, their average is used as an estimate for the base layer signal. Let us use $\mathbf{r}^{(i)}(t)$, $H^{(i)}(t)$ and $\hat{\mathbf{d}}^{(i)}(t)$ to denote the received signal, the channel coefficients matrix, and estimated symbols computed from the ith receive antenna, respectively. The first layer symbol is simply estimated as $\hat{\mathbf{d}}(t) = (\hat{\mathbf{d}}^{(1)}(t) + \hat{\mathbf{d}}^{(2)}(t))/2$. To decode the first layer, we can find the nearest codeword $\tilde{\mathbf{d}}(t) = \max_{s \in \mathcal{S}} |s - \hat{\mathbf{d}}(t)|^2$.

To further decode the second layer, we first reconstruct the base layer signal and then subtract it from the received signal

$$\Delta\mathbf{r}^{(1)}(t) = \mathbf{r}^{(1)}(t) - R^{(1)}(t-1)\tilde{\mathbf{d}}(t), \quad (13)$$

$$\Delta\mathbf{r}^{(2)}(t) = \mathbf{r}^{(2)}(t) - R^{(2)}(t-1)\tilde{\mathbf{d}}(t), \quad (14)$$

where $R^{(1)}(t)$ and $R^{(2)}(t)$ are signal matrices constructed from the two receive antennas, respectively. Let $\hat{r}_j^{(i)}(t)$ denote the jth reconstructed signal of the first layer at the ith antenna. Similarly, let $\Delta r_j^{(i)}(t)$ denote the jth residual signal at the ith antenna. If the first layer is correctly decoded, it is shown in the Appendix that

$$\Delta R(t) \equiv \begin{bmatrix} \Delta r_1^{(1)}(t) & \Delta r_3^{(1)}(t) \\ \Delta r_2^{(1)}(t) & \Delta r_4^{(1)}(t) \\ \Delta r_1^{(2)}(t) & \Delta r_3^{(2)}(t) \\ \Delta r_2^{(2)}(t) & \Delta r_4^{(2)}(t) \end{bmatrix} \approx \Theta(t) \times E_d(t) + N(t), \quad (15)$$

where

$$\Theta(t) = \begin{bmatrix} \hat{r}_1^{(1)}(t) & \hat{r}_2^{(1)}(t) & \hat{r}_3^{(1)}(t) & \hat{r}_4^{(1)}(t) \\ -\hat{r}_2^{(1)}(t)^* & \hat{r}_1^{(1)}(t)^* & -\hat{r}_4^{(1)}(t)^* & \hat{r}_3^{(1)}(t)^* \\ \hat{r}_1^{(2)}(t) & \hat{r}_2^{(2)}(t) & \hat{r}_3^{(2)}(t) & \hat{r}_4^{(2)}(t) \\ -\hat{r}_2^{(2)}(t)^* & \hat{r}_1^{(2)}(t)^* & -\hat{r}_4^{(2)}(t)^* & \hat{r}_3^{(2)}(t)^* \end{bmatrix}$$

$$E_d(t) = \begin{bmatrix} e_1(t) & e_5(t) \\ e_2(t) & e_6(t) \\ e_3(t) & e_7(t) \\ e_4(t) & e_8(t) \end{bmatrix},$$

and $N(t)$ is a 4 by 2 matrix whose components are AWGN signals. By applying MMSE again, we can get estimates of $e_i(t)$ so that the nearest signal can be detected via

$$\hat{E}_d(t) = W_E^H(t-1)\Delta R(t), \tag{16}$$

where $W_E(t) = \left(\Theta(t)\Theta(t)^H + (\rho^2/\Gamma)I_4\right)^{-1}\Theta(t)$ and Γ is the SNR value of the first layer signal.

C. Differential Scheme With Kalman Filtering

The performance of differential STC could degrade a lot in a time-selective fading environment. This is especially true for the second layer signals, since they have less power and need to be estimated more accurately. In this section, we examine a channel tracking mechanism designed for a terminal with two receive antennas while the one-antenna terminal can still decode base-layer signals differentially as described before.

Liu *et al.* [16] proposed a Kalman filtering scheme for the STC system with 2 transmit antennas. With the assistance from Kalman filtering, the time-varying channel status can be tracked with decision feedback. Here, we propose a new scheme that combines differential STC with channel tracking. At the encoder, the first layer still adopts differential encoding, which allows the low-end terminal with a single antenna to receive the base layer signal with a low computational complexity. The Kalman filtering mechanism is adopted by the high-end terminal with two receive antennas.

Let us consider a system consisting of a 4-antenna transmitter to send out Space-time codewords as before. Assume that the initial channel coefficients can be trained by the pilot signal of the Kalman filter [17] in the start-up phase. The first-order autoregressive (AR) model for the time-selective fading channel is adopted to describe the channel variation. That is, for the i-th path channel coefficients received by the k-th antenna, denoted by $h_i^{(k)}(t)$, varies according to the following form $h_i^{(k)}(t) = \alpha h_i^{(k)}(t-1) + v_i^{(k)}(t)$, where $v_i^{(k)}(t)$ is a zero-mean complex Gaussian variable with variance σ_v^2, and α is the correlation factor that can be derived easily from the expected value of channel realizations $\alpha = E[h_i(t)h_i^*(t-1)]$.

At the encoder, the base-layer code $X(t)$ is generated using the differential code as described before. The second layer code $Y(t)$ is then simply obtained using the following formula:

$$Y(t) = \begin{bmatrix} e_1(t) & e_2(t)^* & e_5(t) & e_6(t)^* \\ e_2(t) & -e_1(t)^* & e_6(t) & -e_5(t)^* \\ e_3(t) & e_4(t)^* & e_7(t) & e_8(t)^* \\ e_4(t) & -e_3(t)^* & e_8(t) & -e_7(t)^* \end{bmatrix}.$$

The sum of two layer codes $Z(t) = X(t) + Y(t)$ is transmitted through antennas.

For the receiver with one antenna, only the first layer can be decoded with conventional differential STC. For the receiver with two antennas, the base layer can also be estimated using the conventional differential coding system. To decode the second layer more accurately, we get the help from Kalman filtering. Let the state vector of Kalman filtering be a collection of channel coefficients, *i.e.* $\chi(t) = [h_1^{(1)}(t), h_2^{(1)}(t), h_3^{(1)}(t), h_4^{(1)}(t), h_1^{(2)}(t), h_2^{(2)}(t), h_3^{(2)}(t), h_4^{(2)}(t)]^T$. The observation vector can be naturally chosen to be the received signals from antennas $\zeta(t) = [r_1^{(1)}(t), r_2^{(1)}(t), r_3^{(1)}(t), r_4^{(1)}(t), r_1^{(2)}(t), r_2^{(2)}(t), r_3^{(2)}(t), r_4^{(2)}(t)]^T$. We obtain the state equation

$$\chi(t) = A\chi(t-1) + \mathbf{v}(t), \tag{17}$$

where $A = \alpha I_8$ is an 8×8 diagonal matrix and the measurement equation is

$$\zeta(t) = S(t)\chi(t) + n(t), \tag{18}$$

where

$$S(t) = \begin{bmatrix} Z(t)^T & O_4 \\ O_4 & Z(t)^T \end{bmatrix},$$

and where $Z(t) = X(t) + Y(t)$ is a 4×4 transmitted code matrix and O_4 is a zero matrix of size 4 by 4. The value of $X(t)$ is from the differentially decoded signal of the first layer, which does not demand the knowledge of channel coefficients. The second layer code matrix $Y(t)$ can be estimated using the MMSE algorithm from coarsely predicted channel coefficients

$$\hat{\chi}(t) = A\chi(t-1). \tag{19}$$

Once observation matrix $Z(t)$ is available, $\chi(t)$ can be obtained using a standard Kalman filtering procedure [18].

The updated value $\chi(t)$ can be further utilized to refine the decoded result. At this pass, since we have an estimation for the second layer information $\hat{Y}(t)$, it can be reconstructed and subtracted from the received signal to help decode the first layer more accurately. This residual signal can be expressed as

$$\Delta\zeta^x(t) = \zeta(t) - \begin{bmatrix} \hat{Y}(t)^T & O_4 \\ O_4 & \hat{Y}(t)^T \end{bmatrix} \times \chi(t), \tag{20}$$

which can be used to decode the first layer to get new $\hat{X}(t)$ by differential decoding. Again, to update the second layer signal, the first layer has to be removed. That is, the signal

$$\Delta\zeta^y(t) = \zeta(t) - \begin{bmatrix} \hat{X}(t)^T & O_4 \\ O_4 & \hat{X}(t)^T \end{bmatrix} \times \chi(t) \tag{21}$$

has to be used as the input for the second layer decoding. The filtered channel states $\chi(t)$ along with $\Delta\zeta^x(t)$ and $\Delta\zeta^y(t)$ can be used to decode in this phase. More iterations of these three vectors can be performed to achieve better estimation whenever it is necessary.

The decoding procedure of the receiver with 2 antennas is summarized below.

Step 0: For $t = 0$, initialize the Kalman filter by the training sequence.

Step 1: Estimate the first layer signal $\hat{X}(t)$ by differential decoding.

Step 2: Obtain prediction $\chi_p(t)$ from (19).

Step 3: Compute the residual signal $\Delta\mathbf{r}_y$.

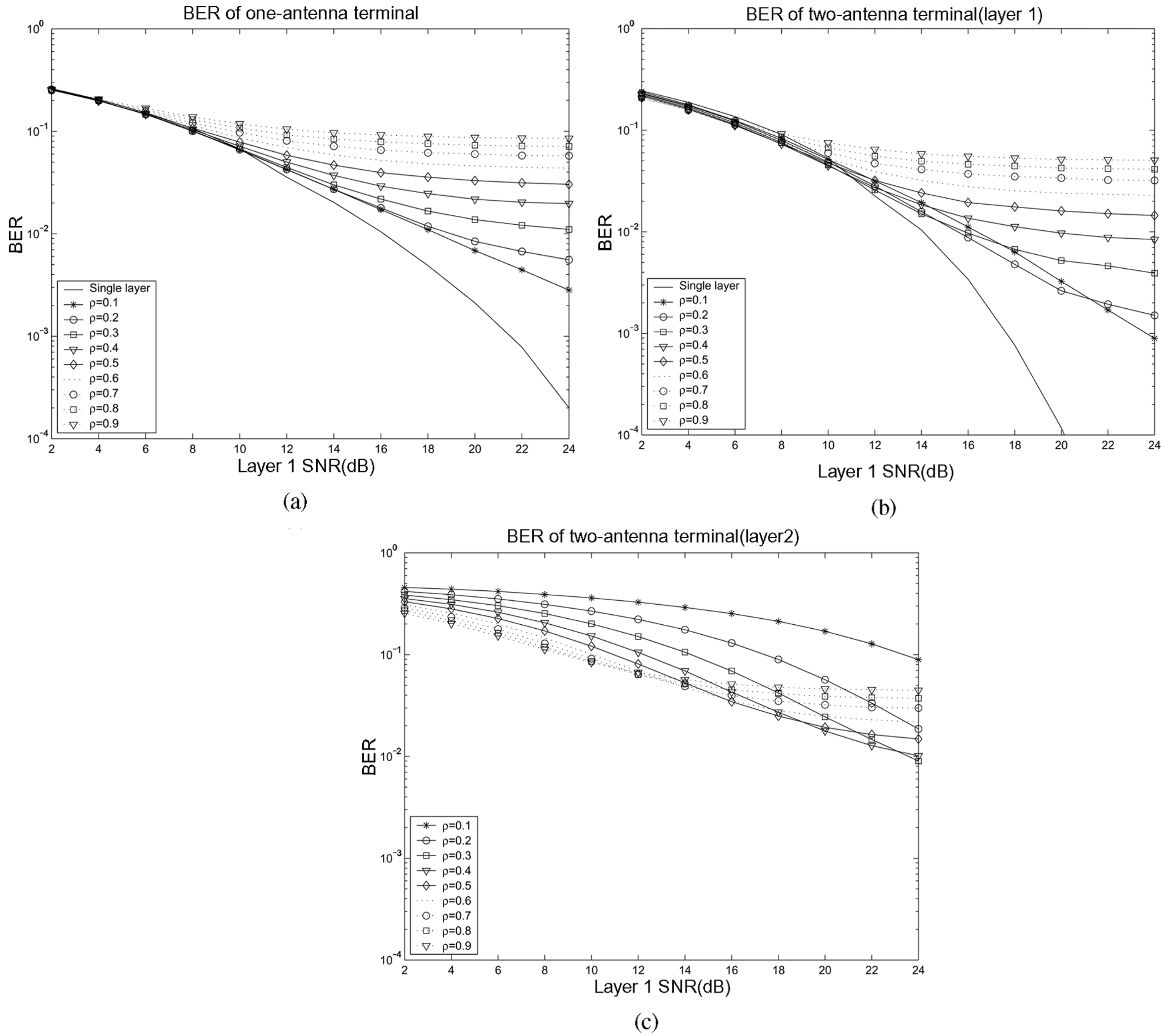


Fig. 3. The bit-error rates of the receiver with differential detection: (a) the first layer for the one-antenna system, (b) the first layer for the two-antenna system, and (c) the second layer for the two-antenna.

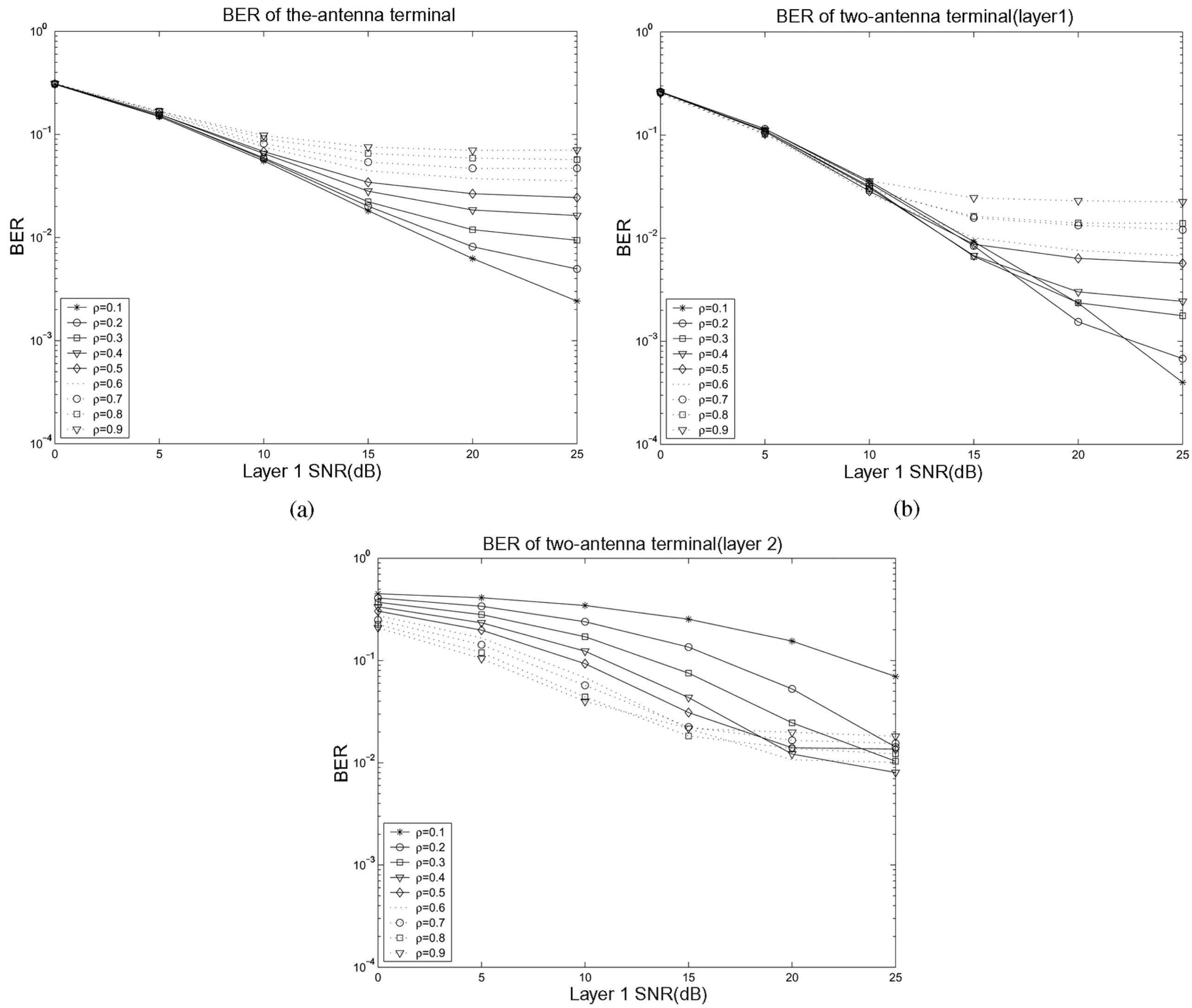
Step 4: Detect the second layer signal $\hat{Y}(t)$ from $\Delta\mathbf{r}_y$.
Step 5: Reconstruct matrix $Z(t) = X(t) + Y(t)$.
Step 6: Perform Kalman filtering to update $\mathbf{h}(t)$.
Step 7: Decode X from $\Delta\zeta^x(t)$ in (20).
Step 8: Decode Y from $\Delta\zeta^y(t)$ in (21).
Step 9: Iterate Steps 6–8 several times to improve the tracking performance.
Step 10: Update time t by $t+1$ and return to Step 1.

D. Experimental Results

1) Differential Detection: In this subsection, we report computer simulation results based on the proposed detection scheme for embedded differential STC as described in Section IV-A. In the transmitter, two layers of data are transmitted with four antennas. For a four-consecutive-symbol duration, four symbols are sent at the first layer, and eight symbols at the second layer. The mapping of the first layer is from 4 bits onto the space $\mathcal{S}$. The second layer modulation is simply BPSK in the quadrature phase with value $\{+i, -i\}$. All paths experience the Rayleigh fading effect. The receiver requires two antennas to retrieve both layers while the one-antenna receiver can reconstruct only the first layer. Suppose that channel coefficients do not change during one block duration. The second layer has only the imagery part while the first layer has only the real part.

The channel model has a slow fading coefficient rate set to 2.5×10^{-4}. The simulation shows that the uncertainty of channel status and noise can deteriorate the performance severely, since errors tend to accumulate as differential detection is performed. Hence, we insert a re-synchronization symbol for every duration of 80 symbols so that the receiver can adjust decoded signals to prevent error propagation.

Fig. 3(a) shows simulation results of decoded bit error rates for the base layer in a terminal with only one receive antenna at different ρ and SNR values. Compared with those in Fig. 2(a) of



(c)

Fig. 4. The bit-error rates of (a) the first layer for one antenna system, (b) the first layer for two-antenna system, and (c) the second layer for the two-antenna receiver with differential detection and Kalman filtering.

the non-differential scheme, we see that the performance curves of the differential scheme have a higher ‘floor’ when the ρ value is the same, and the saturation SNR point is lower. This means that the interference effect is more severe in the differential case. Figs. 3(b) and (c) show the corresponding results of layers 1 and 2 in the 2-antenna terminal, respectively, of the differential scheme. As expected, the performance of the first layer in this case is better than that of the one-antenna system since it exploits more receiving diversity. The overall performance is slightly worse than that of the non-differential scheme.

2) Kalman Filtering Tracking: In this subsection, we report the simulation result of the differential detection scheme for embedded differential STC with channel tracking via Kalman filtering as described in Section IV-C. The setting of the simulation is the same as that in IV-D-1. Again, the channel is modeled as a slow fading one with the coefficient rate set to 2.5×10^{-4}. Resync words are inserted as done before. They also serve as the pilot signal for Kalman filter tracking. Parameter α is computed off-line based on channel realizations. It is observed in our experiments that more iterations in updating coefficients and decoding do not improve the performance much. Hence, we show results with no extra iterations.

Fig. 4(a) shows simulation results of decoded bit error rates for the base layer in a terminal with only one receive antenna at different ρ and SNR values. Comparing them with results of the pure differential scheme, we see that the first layer has almost the same performance for these two one-antenna receive systems. This is easily understood since the interference level from the second layer is almost identical.

The results for the two-antenna receiver are shown in Figs. (4b) and (c). We see that the BER floor for each ρ value has been reduced to about one half. For the second layer information, there is a similar performance. This is due to the tracking effect of the Kalman filter, which reduces the interference of two layers. These results indicate that the proposed system with Kalman filtering can improve the differential scheme at the cost of the complexity of the receiver with two antennas. However, the receiver with only one antenna can still get the same quality without increasing its complexity. This is suitable for low-end receivers such as handsets and wireless PDA.

V. Conclusion and Future Work

In this research, we proposed a special class of space-time codes to enable wireless broadcast for heterogeneous receivers. The layered multimedia source can be nicely integrated with this kind of systems for scalable content delivery. The terminal with more antennas can retrieve contents with higher quality. It is worthwhile to point out that conventional space-time codes utilize the transmitter diversity while this work exploits the receiver diversity. Our basic idea is that some space-time codes can only be detected with more than one antenna. Thus, one can trade more receive antennas with a higher computational complexity for higher quality. Nevertheless, receivers with less diversity can still get the minimum level of guaranteed QoS. Embedded STC can integrate the receiver diversity with the layered source coding technique to provide an embedded wireless broadcast service.

However, the proposed system is far from completion, and quite a few research problems remain in order to make the system practical. For example, it is an interesting yet challenging task to study better space-time codes to enhance the proposed system. Please note that the proposed space-time codes are not orthogonal so that there is an interference effect between layers. It is desirable to remove or reduce this interference effect by code design. It is also worthwhile to integrate multimedia source coding and/or the forward error correction (FEC) technique with the proposed embedded STC system. The integration with FEC can reduce the bit error probability to combat wireless channel fading and noise so that the receiver can decode more accurately. In addition, FEC can be integrated with interference cancellation and iterative channel tracking to further improve the overall performance. Unequal error protection can be used along with better resource allocation for video transmission to achieve better quality.

Appendix

In this Appendix, we briefly summarize the derivation of (15). Let us first focus on antenna 1. From (13) and by expanding (11), we obtain

$$\begin{bmatrix} \Delta r_1^{(1)}(t) \\ \Delta r_2^{(1)}(t) \\ \Delta r_3^{(1)}(t) \\ \Delta r_4^{(1)}(t) \end{bmatrix} \equiv \Delta \mathbf{r}^{(1)}(t) = \mathbf{r}_{1,Y}(t) + \mathbf{n}^{(1)}(t)$$
$$= (X(t)E(t))^T\, \mathbf{h}^{(1)}(t) + \mathbf{n}^{(1)}(t)$$
$$= E(t)^T X(t)^T \mathbf{h}^{(1)}(t) + \mathbf{n}^{(1)}(t)$$
$$\approx E(t)^T \hat{\mathbf{r}}^{(1)}(t) + \mathbf{n}^{(1)}(t)$$
$$= \begin{bmatrix} e_1(t) & e_2(t) & e_3(t) & e_4(t) \\ e_2(t)^* & -e_1(t)^* & e_4(t)^* & -e_3(t)^* \\ e_5(t) & e_6(t) & e_7(t) & e_8(t) \\ e_6(t)^* & -e_5(t)^* & e_8(t)^* & -e_7(t)^* \end{bmatrix} \times \begin{bmatrix} \hat{r}_1^{(1)}(t) \\ \hat{r}_2^{(1)}(t) \\ \hat{r}_3^{(1)}(t) \\ \hat{r}_4^{(1)}(t) \end{bmatrix} + \mathbf{n}^{(1)}(t).$$

The element $\Delta r_1^{(1)}(t)$ can be represented as

$$\Delta r_1^{(1)}(t) = e_1(t)\hat{r}_1^{(1)}(t) + e_2(t)\hat{r}_2^{(1)}(t) + e_3(t)\hat{r}_3^{(1)}(t) + e_4(t)\hat{r}_4^{(1)}(t) + n_1^{(1)}(t)$$
$$= \begin{bmatrix} \hat{r}_1^{(1)}(t) & \hat{r}_2^{(1)}(t) & \hat{r}_3^{(1)}(t) & \hat{r}_4^{(1)}(t) \end{bmatrix} \begin{bmatrix} e_1(t) \\ e_2(t) \\ e_3(t) \\ e_4(t) \end{bmatrix} + n_1^{(1)}(t),$$

and the conjugate of $\Delta r_2^{(1)}(t)$ becomes

$$\Delta r_2^{(1)}(t)^* = \begin{bmatrix} -\hat{r}_2^{(1)}(t)^* & \hat{r}_1^{(1)}(t)^* & -\hat{r}_4^{(1)}(t)^* & \hat{r}_3^{(1)}(t)^* \end{bmatrix} \times \begin{bmatrix} e_1(t) \\ e_2(t) \\ e_3(t) \\ e_4(t) \end{bmatrix} + n_2^{(1)}(t)^*.$$

We can perform similar manipulations on $\Delta r_3^{(1)}(t)$ and $\Delta r_4^{(1)}(t)^*$, and put them together with $\Delta r_1^{(1)}(t)$ and $\Delta r_2^{(1)}(t)^*$ to yield

$$\begin{bmatrix} \Delta r_1^{(1)}(t) & \Delta r_3^{(1)}(t) \\ \Delta r_2^{(1)}(t)^* & \Delta r_4^{(1)}(t)^* \end{bmatrix} = \begin{bmatrix} \hat{r}_1^{(1)}(t) & \hat{r}_2^{(1)}(t) & \hat{r}_3^{(1)}(t) & \hat{r}_4^{(1)}(t) \\ -\hat{r}_2^{(1)}(t)^* & \hat{r}_1^{(1)}(t)^* & -\hat{r}_4^{(1)}(t)^* & \hat{r}_3^{(1)}(t)^* \end{bmatrix} \times \begin{bmatrix} e_1(t) & e_5(t) \\ e_2(t) & e_6(t) \\ e_3(t) & e_7(t) \\ e_4(t) & e_8(t) \end{bmatrix} + N^{(1)}(t),$$

where $N^{(1)}(t)$ is a 2×2 matrix with the correponding elements taking from the vector $\mathbf{n}^{(1)}(t)$. Similarly, for antenna 2, we can obtain

$$\begin{bmatrix} \Delta r_1^{(2)}(t) & \Delta r_3^{(2)}(t) \\ \Delta r_2^{(2)}(t)^* & \Delta r_4^{(2)}(t)^* \end{bmatrix} = \begin{bmatrix} \hat{r}_1^{(2)}(t) & \hat{r}_2^{(2)}(t) & \hat{r}_3^{(2)}(t) & \hat{r}_4^{(2)}(t) \\ -\hat{r}_2^{(2)}(t)^* & \hat{r}_1^{(2)}(t)^* & -\hat{r}_4^{(2)}(t)^* & \hat{r}_3^{(2)}(t)^* \end{bmatrix} \times \begin{bmatrix} e_1(t) & e_5(t) \\ e_2(t) & e_6(t) \\ e_3(t) & e_7(t) \\ e_4(t) & e_8(t) \end{bmatrix} + N^{(2)}(t).$$

Consequently, we get (15) by combining the above two equations.

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References

[1] J. G. Proakis, *Digital Communications*. New York: McGraw-Hill, 1989.

[2] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction," *IEEE Tr. Information Theory*, vol. 44, pp. 744–764, 1998.

[3] T. M. Cover, "Broadcast channels," *IEEE Trans. on Information Theory*, vol. IT-18, pp. 2–14, 1972.

[4] K. Ramchandran, A. Ortega, K. M. Uz, and M. Vetterli, "Multiresolution broadcast for digital HDTV using joint source/channel coding," *IEEE Journal on Selected Areas in Communications*, vol. 11, pp. 6–23, 1993.

[5] *Information Technology—JPEG 2000 Image Coding System—Part 1: Core Coding System*, ISO/IEC 15444-1:2000, Dec. 2000.

[6] *Coding of Audio-Visual Objects, Part-2 Visual, Amendment 4: Streaming Video Profile*, ISO/IEC 14496-2/FPDAM4, July 2000.

[7] W. Li, "Overview of fine granularity scalability in MPEG-4 video standard," *IEEE Journal on Circuits and Systems for Video Technology*, vol. 11, pp. 301–317, 2001.

[8] H. Zheng and K. J. R. Liu, "Space-time diversity for multimedia delivery over wireless channels," in *Proc. ISCAS 2000*, 2000, vol. IV, pp. 285–288.

[9] A. F. Naguib, N. Seshadri, and A. R. Calderbank, "Applications of space-time block codes and interference suppression for high capacity and high data rate wireless systems," in *Proc. 32nd Asilomar Conference on Signals, Systems and Computers*, 1998, pp. 1803–1810.

[10] U. Wachsmann, J. Thielecke, and H. Schotten, "Exploiting the datarate potential of MIMO channels: multi-stratum space-time coding," in *Proc. Vehicular Technology Conference*, May 6–9, 2001, vol. 1, pp. 199–203.

[11] M. Stege and G. Fettweis, "Multistratum-permutation codes for MIMO communication," *IEEE Trans. Commun.*, vol. 21, no. 5, pp. 774–782, June 2003.

[12] S. M. Alamouti, "A simple transmit diversity technique for wireless communication," *IEEE Journal on Selected Areas in Communications*, vol. 16, no. 8, pp. 1451–1458, 1998.

[13] V. Tarokh and H. Jafarkhani, "A differential detection scheme for trasmit diversity," *IEEE J. Selected Areas Commun.*, vol. 18, pp. 1169–1174, 2000.

[14] H. Jafarkhani and V. Tarokh, "Multiple transmit antenna differential detection from generalized orghogonal designs," *IEEE Trans.. Information Theory*, vol. 47, pp. 2626–2631, 2001.

[15] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Trans. Information Theory*, vol. 45, pp. 1456–1467, 1999.

[16] Z. Liu, X. Ma, and G. B. Giannakis, "Space-time coding and Kalman filtering for time-selective fading channels," *IEEE Trans. Communications*, vol. 50, no. 2, pp. 183–186, February 2002.

[17] R. Emil Kalman, "A new approach to linear filtering and prediction problems," *Transactions of the ASME-Journal of Basic Engineering*, vol. 82, no. Series D, pp. 35–45, 1960.

[18] J. M. Mendel, *Lessons in Estimation Theory for Signal Processing, Communications, and Control*. Upper Saddle River, New Jersey: Prentice Hall, 1995.

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