Robust Progressive Image Transmission Over OFDM Systems Using Space-Time Block Code

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***Abstract*—A joint source-channel coding (JSCC) scheme for robust progressive image transmission over broadband wireless channels using orthogonal frequency division multiplexing (OFDM) systems with spatial diversity is proposed for the application environments where no feedback channel is available such as broadcasting services. Most of current research about JSCC focus on either binary symmetric channel (BSC) or additive white Gaussian noise (AWGN) channels. To deal with fading channels in most of previous methods, the fading channel is modeled as two state Gilbert-Elliott channel model and the JSCC is normally aimed at the BER of *bad* channel status, which is not optimal when channel is at *good* status. By using diversity techniques and OFDM, the frequency selective fading effects in broadband wireless channels can be significantly decreased and we show that subchannels in OFDM systems approach Gaussian noisy channels when the diversity gain gets large, as a result, the system performance can be improved in terms of throughput and channel coding efficiency. After analyzing the channel property of OFDM system with spatial diversity, a practical JSCC scheme for OFDM system is proposed. The simulation results are presented for transmit diversity with different number of antennas and different multipath delay and Doppler spread. It is observed from simulations that the performance can be improved more than 4 dB in terms of peak signal-to-noise ratio (PSNR) of received image *Lena* and the performance is not very sensitive to different multipath spread and Doppler frequency.**

***Index Terms*—Image transmission, multimedia communications, OFDM.**

I. INTRODUCTION

THE public's desire for multimedia communications over broadband wireless channels combined with the increasing demands for Intranet access suggests a very promising future for wireless data services. For multimedia communications, the delay constraint limits the use of Automatic Repeat reQuest (ARQ), and in some cases the feedback channel is not available such as broadcasting service. The unequal importance property and delay constraint of multimedia data suggests the use of joint source-channel coding (JSCC) scheme which has been studied in the last decade [1], [2], [4], [8]. The existing techniques for JSCC can be divided roughly, into two main categories: 1) the source and channel coding is combined together to minimize the total distortion incurred by source coding and noisy channel (e.g., [4]). These can be regarded as *true* joint source-channel coding and 2) the source code is optimized for a noiseless channel, and is then made robust toward channel errors by a optimal channel coding and/or modulation policy to minimize the channel errors while optimizing the throughput [2], [8]–[10], [12]. This kind of JSCC is of practical interests because most of current image and video coding standards are based on noiseless channel assumption. We denote this kind of method as joint source-channel matching (JSCM) as mentioned in [9]. Most of the JSCM schemes focus on the binary symmetric channel (BSC) or additive white Gaussian noise (AWGN) channel where the bit error rate or signal-to-noise ratio (SNR) is constant. To deal with fading channels in wireless communications, the fading channel is modeled as two-state Gilbert-Elliot model and the JSCM normally aims at the BER of *bad* channel status [10], [14]. However, this scheme is not optimal when channel is in *good* condition. Another way is adaptive JSCM, where the channel state information (CSI) is available at the transmitter through a feedback channel. But in wireless communications it is normally impractical to get the CSI through feedback channel because of limited bandwidth of feedback channel, or there is no feedback channel available such as in broadcasting services.

The transmission of the images coded using set partitioning in hierarchical trees (SPIHT) [26] across noisy channels is an active research topic recently. For example, in [8], a concatenated channel coding scheme was applied to SPIHT coded image to achieve performance gains over previous coding systems for memoryless channels with known statistics. In [11] and [12], a optimal source and channel rate allocation scheme is proposed for channels with and without feedback channel. A packetization scheme in [13] was proposed that the output bitstream of zerotrees encoder was reordered and packetized in such a way that complete trees of wavelet coefficients were contained within packets. This allows graceful degradation of an image in the presence of packet erasures, instead of loss of synchronization typically experienced with the error-sensitive zerotree encoder. In [7], a product channel code structure is used to make the system in [8] robust to fading channels. A combination of channel coding and packetization is proposed in [6], which can perform well on channels that can suffer packet losses as well as statistically varying bit errors.

In this paper, a joint source-channel matching scheme for progressive image transmission over broadband wireless channels using OFDM systems with spatial diversity is proposed, where neither feedback channel nor CSI is available at transmitter. Multimedia transmission over asymmetric digital subscriber lines (ADSL) have been studied recently [5], [3], where

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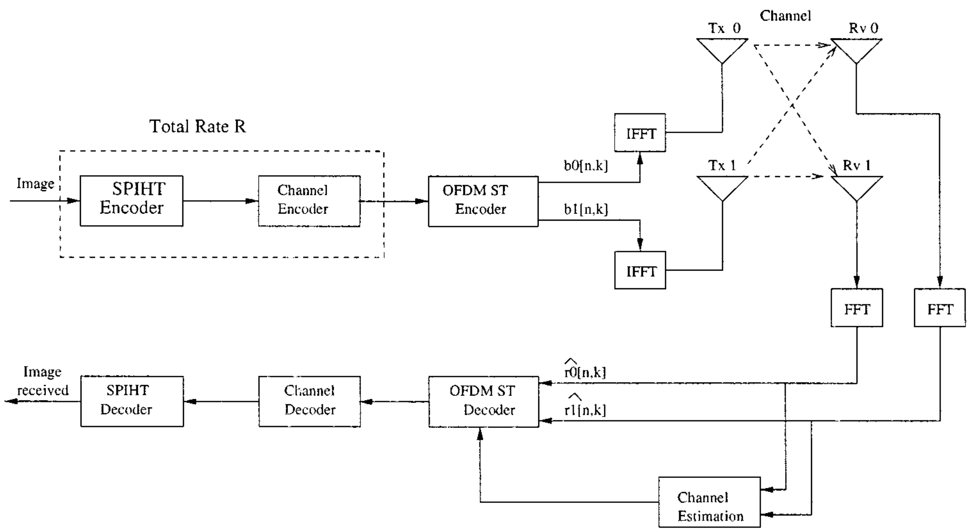

Fig. 1. System structure for joint source–channel image transmission over OFDM channels using multiple antennas.

the channel is assumed to be time-invariant. For high bit-rate wireless communications, OFDM [16] is an attractive technique to be used because of its simplicity in dealing with frequency-selective, time-dispersive wireless fading channels [16], [19]. Diversity techniques, including spatial, frequency, and time domain diversity, have been suggested to decrease the fading effect. Sufficiently spaced antennas are an attractive source of diversity since they do not typically incur bandwidth expansion as in frequency division diversity, and does not incur delay as in time diversity. Though spatial diversity can be available at both transmitter and receiver, most of past work has focused on exploiting receiver diversity since simple combining techniques, such as maximum ratio combining (MRC) or antenna selection, are available to obtain sufficient diversity gain [15]. However, it may not be possible to obtain much diversity gain at mobile terminal because of the space and power limitations at mobile terminals. A common technique to solve this problem is to use multiple transmit antennas at base station to provide transmit diversity. Some transmit diversity schemes have been proposed recently such as delay diversity, space-time trellis coding [17], and space-time block coding [18], [20]. It should be noted that our scheme focuses on broadband wireless communications, and, using spatial diversity to deal with fading effects which are different from other proposed image transmission schemes, the proposed techniques can definitely be used in conjunction with all of these techniques as previously mentioned.

We consider using space–time block codes (STBC) as a spatial diversity technique to decrease the fading effects in both time an frequency domain for OFDM systems, and show that the subchannels in OFDM can be modeled as parallel flat Rayleigh fading channels and *independent Gaussian noisy channels when the diversity gain goes to infinity*. Then, a JSCM scheme for OFDM system with STBC is presented for SPIHT encoded image transmission, where the channel state information is not available at the transmitter but known at the receiver.

The remainder of the paper is organized as follows. In Section II, the system structure is described, then, in Section III, the channel property of OFDM system with STBC is studied and we show that the subchannels of OFDM system can be modeled as Gaussian noisy channels when the diversity gain is large enough. In Section IV, a JSCM scheme is proposed for image transmission over OFDM system using STBC. In Section V, simulation results are presented under different diversity gain and fading parameters. Finally, conclusions are reached.

II. System Description

The system structure is shown in Fig. 1. The image coding algorithm used is SPIHT [26], which is a refined version of the Embedded Zerotree algorithm [25]. The output of SPIHT encoder is progressive, such that wherever the transmission is stopped, the received data can be used to reconstruct an image with corresponding quality. The SPIHT bitstream is packetized into source-packets of size B_i bits for packet $i, i = 1, \ldots, L$. A linear block code (N, B_i) (e.g., RS code or convolutional code with known tail state) is used to protect source-packet i. The resulted codewords are of equal size and treated as OFDM blocks to the transmitter. There are total of L OFDM blocks, where L is decided by the transmission rate. The source-packet size $B_i, i = 1, 2, \ldots, L$ is obtained by a JSCM algorithm to minimize the average distortion of the reconstructed image at the receiver. The STBC with N transmit antennas and M receiver antennas is denoted as (N, M) multiantenna system. The STBC for two transmit antennas is proposed by Alamouti [20]. Tarokh *et al.* extended it to a general STBC by orthogonal design [17], [18]. The maximum-likelihood decoding can be achieved for STBC based only on linear processing at the receiver. Alamouti and Tarokh *et al.* have shown in [20] and [17] that (N, M) STBC is equivalent to $(1, MN)$ receiver diversity with maximal ratio combining (MRC). This is an important result because the performance analysis of STBC can be transfered to receiver diversity with MRC which has been well studied. The multipath Rayleigh fading channels between transmit antenna $i, i = 1, \ldots, N$ and receive antenna $j, j = 1, \ldots, M$ is denoted as $h_{ij,p}, p = 1, \ldots, P$ where P is the total number of paths between i and j. Each path $h_{ij,p}$ is zero mean complex Gaussian process with variance $\sigma^2_{i,jp}$, which are normalized such that

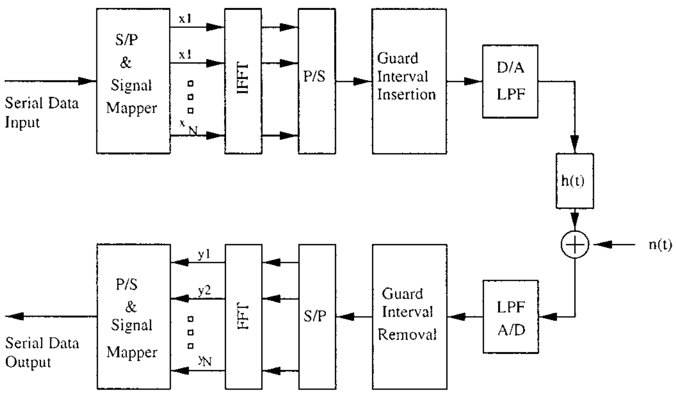

Fig. 2. Based-band OFDM system.

$\sum_1^P \sigma_{ij,p}^2 = 1$. All the $h_{ij,p}$'s are statistically independent from each other for different i, j and p.

A. Orthogonal Frequency Division Multiplexing

The base-band OFDM system is shown in Fig. 2, where x_k are the transmitted symbols, $h(t)$ is the channel impulse response, $n(t)$ is the white complex Gaussian channel noise and y_k are the received symbols. The x_k's are taken from a M-ary signal constellation. The D/A and A/D converters contain ideal low-pass filters with bandwidth $1/T_s$, where T_s is the sampling interval. A cyclic extension of time length T_g is used to eliminate inter-block interference and preserve the orthogonality of tones. The channel impulse response $h(t)$ is a time-limited pulse of the form

$$h(t, \tau) = \sum_{p=1}^{P} \alpha_p(t)\delta(\tau - \tau_p) \quad (1)$$

where P is the number of paths between transmitter and receiver; each path is complex Gaussian process with zero-mean and variance σ_p^2. The path delay satisfies $\tau_p \leq T_g$, i.e., the entire impulse response lies inside the guard space. The channel response is normalized such that $\sum_{p=1}^{P} \sigma_p^2 = 1$.

Using (1), the frequency response of the time-varying radio channel at time t is

$$H(t, f) = \int_{-\infty}^{\infty} h(t, \tau)e^{-j2\pi f\tau\, d\tau} = \sum_{p=1}^{P} \alpha_p(t)e^{-j2\pi f\tau_p}. \quad (2)$$

For an OFDM system with block length T_f and tone spacing (subchannel spacing) Δf, the output signal after FFT at the kth tone of the nth OFDM block can be expressed as

$$y[n, k] = H[n, k]x[n, k] + w[n, k] \quad (3)$$

where $w[n, k]$ is additive Gaussian noise at the kth tone and the nth block, with zero-mean and variance σ^2. We also assume $w[n, k]$ is independent for different n's, k's. $H[n, k] = H(nT_f, k\Delta f)$ is the frequency response at the kth tone of the nth block.

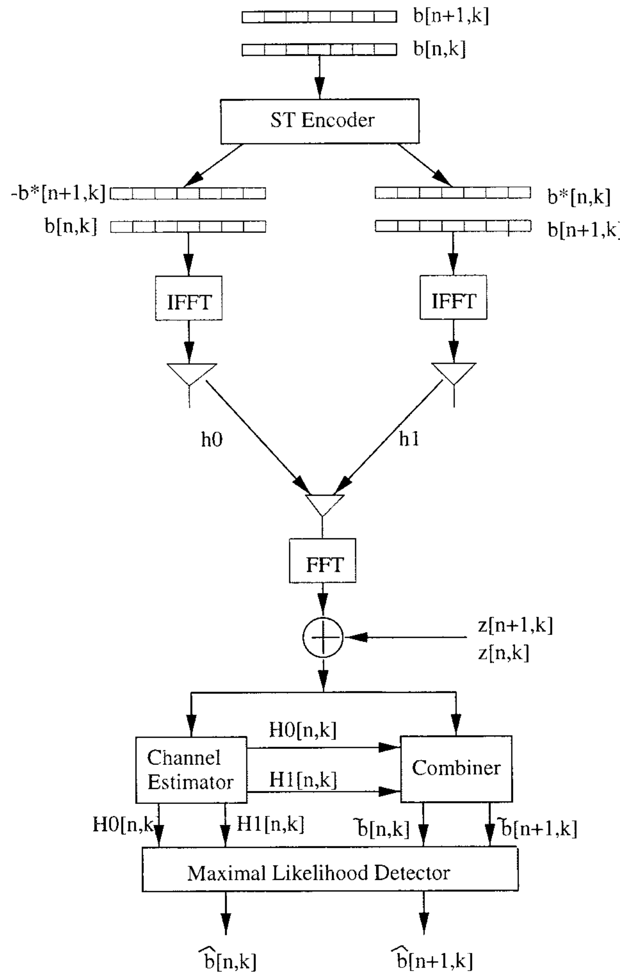

Fig. 3. Space-time block code for OFDM systems.

B. Space-Time Block Coding Scheme for OFDM

A simple transmit diversity for OFDM systems with two transmit antennas is proposed in Fig. 3. In this paper, we use $(2, M)$ multiple antenna system for simplicity. An OFDM block is denoted as a complex vector $b(n)$, where n is the discrete time index for $t = nT_f$, with T_f being the OFDM symbol interval. There are N_s complex subsymbols in $b(n)$ denoted as $b(n, k), k = 1, 2, \ldots, N_s$, where N_s is the total number of subchannels in OFDM systems.

Assume there are two transmit antennas and two receive antennas. Two OFDM blocks $b(n)$ and $b(n+1)$ are collected from the output of signal mapper where each subsymbol is mapped to a point in a signal constellation set. For $b(n)$ and $b(n+1)$, the OFDM blocks for transmit antenna 0 and 1 after STBC encoding are denoted as $b_0(n), b_0(n+1)$ and $b_1(n), b_1(n+1)$, respectively. For subchannel $k, k = 1, \ldots, N_s$, the STBC encoder works as follows:

$$\begin{aligned} b_0(n, k) &= b(n, k) \\ b_0(n+1, k) &= -b^*(n+1, k) \\ b_1(n, k) &= b(n+1, k) \\ b_1(n+1, k) &= b^*(n, k) \end{aligned} \quad (4)$$

where $*$ denotes complex conjugate.

The channel impulse response between transmit antenna i and receive antenna j is

$$h_{ij}[n,\tau]=\sum_{p=1}^{P}\alpha_{ij,p}(nT_f)\delta(\tau-\tau_{ij,p}), \tag{5}$$

where $\tau_{ij,p}$ is the delay of the pth path between i and j. The channel frequency response at time n and subchannel k is

$$H_{ij}[n,k]=\sum_{p=1}^{P}\alpha_{ij,p}(nT_f)e^{-j2\pi k\Delta f\tau_{ij,p}}. \tag{6}$$

Assume the channel state is constant over two OFDM symbol intervals, we have

$$\begin{aligned}H_{ij}[n,k]&=H_{ij}[n+1,k]=H_{ij}(k)\\&=\sum_{p=1}^{P}\alpha_{ij,p}e^{-j2\pi k\Delta f\tau_{ij,p}}=C_{ij}(k)e^{-j\theta_{ij}}\end{aligned} \tag{7}$$

where $i,j=0,1, C_{ij}$ is the amplitude of $H_{ij}(k)$, which is a Rayleigh fading channel.

As a result, the received subsymbol at subchannel k from receiver antenna $j=0,1$ after FFT demodulation are

$$\begin{aligned}r_j(n,k)&=H_{0j}(k)b_0(n,k)+H_{1j}(k)b_1(n,k)+w_j(n,k)\\&\overset{(4)}{=}H_{0j}(k)b(n,k)+H_{1j}(k)b(n+1,k)+w_j(n,k)\\r_j(n+1,k)&=H_{0j}(k)b_0(n+1,k)+H_{1j}(k)b_1(n+1,k)\\&\quad+w_j(n+1,k)\\&\overset{(4)}{=}-H_{0j}(k)b^*(n+1,k)+H_{1j}(k)b^*(n,k)\\&\quad+w_j(n+1,k)\end{aligned} \tag{8}$$

and the combiner for receive antenna $j, j=0,1$ is [20]

$$\begin{aligned}\hat{b}_j(n,k)&=H_{0j}^*(k)r_j(n,k)+H_{1j}(k)r_j^*(n+1,k)\\\hat{b}_j(n+1,k)&=H_{1j}^*(k)r_j(n,k)-H_{0j}(k)r_j^*(n+1,k).\end{aligned} \tag{9}$$

After combining the outputs of the two receive antenna 0 and 1, we have

$$\begin{aligned}\tilde{b}(n,k)&=\hat{b}_0(n,k)+\hat{b}_1(n,k)\\\tilde{b}(n+1,k)&=\hat{b}_0(n+1,k)+\hat{b}_1(n+1,k).\end{aligned} \tag{10}$$

Substituting (7)–(9) into (10), we get

$$\begin{aligned}\tilde{b}(n,k)&=\left[C_{00}^2(k)+C_{01}^2(k)+C_{10}^2(k)+C_{11}^2(k)\right]b(n,k)\\&\quad+H_{00}(k)^*w_0(n,k)+H_{10}(k)w_0^*(n+1,k)\\&\quad+H_{01}^*(k)w_1(n,k)+H_{11}(k)w_1^*(n+1,k)\\\tilde{b}(n+1,k)&=\left[C_{00}^2(k)+C_{01}^2(k)+C_{10}^2(k)+C_{11}^2(k)\right]\\&\quad\times b(n+1,k)-H_{00}(k)w_0^*(n+1,k)\\&\quad+H_{10}^*(k)w_0(n,k)-H_{01}(k)w_1^*(n+1,k)\\&\quad+H_{11}^*(k)w_1(n,k).\end{aligned} \tag{11}$$

The combined signals in (11) are then sent to the maximum likelihood detector where for $b(n,k)$ choose $b(n,k)=s_i$ if and only if

$$\begin{aligned}&\left[C_{00}^2(k)+C_{01}^2(k)+C_{10}^2(k)+C_{11}^2(k)-1\right]|s_i|^2\\&\quad+d^2(\tilde{b}(n,k),s_i)\\&\qquad\le\left[C_{00}^2(k)+C_{01}^2(k)+C_{10}^2(k)\right.\\&\qquad\quad\left.+C_{11}^2(k)-1\right]|s_l|^2+d^2(\tilde{b}(n,k),s_l)\end{aligned} \tag{12}$$

or for MPSK signals choose $b(n,k)=s_i$ if and only if

$$d^2(\tilde{b}(n,k),s_i)\le d^2(\tilde{b}(n,k),s_l),\quad\forall i\neq l. \tag{13}$$

where s_i, s_l are points in a signal constellation set. Similarly, the decision rule can be used for subsymbol $b(n+1,k)$.

It is straightforward to implement $M\ge 2$ receiver antennas to get more diversity gains [20]. It is shown in [20], [18] that the (M,N) STBC scheme is equivalent to $(1,MN)$ receiver diversity using MRC.

III. Channel Property of OFDM System Using Space-Time Block Code

By using OFDM, the whole bandwidth is divided into a large number of subchannels, where each subchannel is nonfrequency selective and the intersymbol interference (ISI) can be neglected when guard interval is inserted in each OFDM symbol. It is already known that the subchannels in OFDM are flat Rayleigh fading channels. Next, we study the property of OFDM system with (N,M) STBC.

We know that the channel capacity of a bandwidth limited Gaussian noisy channel can be expressed as [21]

$$C_g=\log_2(1+\mathrm{SNR})b/s/Hz \tag{14}$$

where SNR is the signal-to-noise ratio.

The channel capacity for Rayleigh fading has to be calculated in an average sense because the SNR varies in time. In flat Rayleigh fading environment, the signal model for M-branch receiver diversity is

$$r_i(t)=h_i(t)s(t)+n_i(t),\quad i=1,\ldots,M \tag{15}$$

where $h_i(t)$ is zero-mean complex Gaussian variable with variance 0.5 in each dimension, $s(t)$ is the baseband transmitted signal with signal energy $E, n_i(t)$ is the additive white Gaussian noise with variance N_0 at receiver antenna i. Note that $h_i(t)$ can also be expressed as $h_i(t)=\alpha(t)e^{j\phi_i(t)}$, then the MRC combines the M received signal as

$$\tilde{s}=\sum_{i=1}^{M}r_i(t)h^*(t)=s(t)\sum_{i=1}^{M}\alpha_i(t)^2+\sum_{i=1}^{M}h^*(t)n(t) \tag{16}$$

the instantaneous SNR at receiver at time t is [24]

$$\gamma(t)=\frac{E}{N_0}\sum_{i=1}^{M}\alpha_i^2(t) \tag{17}$$

and the average receiver SNR Γ is

$$\Gamma=E(\gamma(t))=\frac{EM}{N_0}. \tag{18}$$

Let $p_{\Gamma,M}$ denote the probability density of γ for M-fold receiver diversity with average receiver SNR Γ, then the average channel capacity is [24]

$$C = \int_0^\infty \log(1+\gamma) p_{\Gamma,M}(\gamma)\, d\gamma, \tag{19}$$

and

$$p_{\Gamma,M} = \frac{M}{\Gamma} \frac{1}{(M-1)!} \left(\frac{M\gamma}{\Gamma}\right)^{M-1} \exp\left(\frac{-M\gamma}{\Gamma}\right). \tag{20}$$

The average channel capacity for AWGN and flat Rayleigh fading channel with different M is plotted in Fig. 7. It can be noted that the channel capacity in the Rayleigh fading for $M = 8$ and greater is very close to that in the Gaussian noise environment.

Several points can be summarized as follows [22].

1) The channel capacity in a Rayleigh fading is in an average sense.
2) The channel capacity in a Rayleigh fading is always lower than that of a Gaussian noise channel.
3) The diversity scheme can increase the channel capacity in a Rayleigh fading environment.

When the diversity gain increases, we have the following proposition.

Proposition 1: When the number of diversity branches $M \rightarrow \infty$, the channel capacity of a M-branch signal in a Rayleigh fading environment approaches the channel capacity in a Gaussian noise environment.

Proof: From (18), the transmitted signal energy E for a given average receiver SNR Γ is

$$E = \frac{\Gamma N_0}{M}. \tag{21}$$

From (17), the instantaneous SNR $\gamma(t)$ at receiver with M antennas is

$$\gamma(t) = \frac{E}{N_0} \sum_{i=1}^{M} \alpha_i^2(t) = \Gamma \frac{\sum_{i=1}^{M} \alpha_i^2(t)}{M}. \tag{22}$$

From the Law of Large Numbers [23], we have

$$P[\lim_{M\rightarrow\infty} \gamma(t) = \Gamma] = 1 \tag{23}$$

then the channel capacity at time t is

$$C(t) = \log_2(1+\gamma(t)) \stackrel{M\rightarrow\infty}{\longrightarrow} \log_2(1+\Gamma) = C_g. \tag{24}$$

Now we have the following conclusion. ■

Proposition 2: When OFDM system with (N, M) space-time block code is used in a frequency-selective, time-dispersive Rayleigh fading environment, the frequency response of OFDM subchannels can be modeled as parallel, independent Gaussian noisy channels when $MN \rightarrow \infty$.

Proof: We know that the subchannels in OFDM are parallel flat Rayleigh fading channels. It is also shown that a subchannel in OFDM system with (N, M) space-time block code is equivalent to $(1, MN)$ receiver diversity system with MRC. Furthermore, it is proved in Proposition 1 that a flat Rayleigh fading channel with M-branch receiver diversity approaches Gaussian noisy channel when $M \rightarrow \infty$. Then when $MN \rightarrow \infty$, the conclusion is reached. ■

IV. Joint Source–Channel Matching for OFDM Using STBC

Now we consider the joint source-channel matching scheme for progressive image transmission over OFDM systems using (N, M) STBC. First, we derive a JSCM method for OFDM systems with STBC with the assumption that the OFDM subchannels are parallel, independent Gaussian noisy channels with the same SNR. For OFDM systems with finite antenna (N, M) STBC, we propose to use *target* SNR to satisfy the assumption based on which the JSCM is proposed. We assume the average SNR is known at both transmitter and receiver.

A. Joint Source-Channel Matching Scheme for OFDM Systems

The problem considered is as follows. The SPIHT encoded bitstream is to be transmitted over OFDM systems using (N, M) STBC. The transmission rate is R_t bit-per-pixel (bpp). There are N_s subchannels in the OFDM and each subchannel is modulated by a complex symbol from a M-ary modulation set, e.g., QPSK. The SPIHT stream is packetized into L source-packets with B_i bits for packet $i, i = 1, 2, \ldots, L$, where L is the total number of OFDM blocks decided by R_t. For channel coding, we assume that we are provided with a finite number of block code. These channel codes operate on source packets which are CRC-16 outer coded first, here we use the terms that are used in the concatenated code, CRC is outer code and RS is inner code. The resulted blocks of codeword are of equal size and are subsequently transmitted over the channels as OFDM blocks. The total number of OFDM blocks L used for transmission should satisfy

$$L \leq \frac{R_t HW}{N_s \log_2 M_{\text{ary}}} \tag{25}$$

where H and W is the dimension of the image, M_{ary} is the constellation size for subchannels.

The source packets are channel-coded by the assigned channels codes and transmitted over the noisy channel. The receiver tries to recover the source-packets from the noisy received OFDM blocks. The channel decoder either correctly decodes a source-packet or detects an error. We assume that the probability of undetected error is zero because of CRC-16 error detection. For SPIHT bitstream, if a packet cannot be decoded correctly then the subsequent packets cannot be used to improved the quality of the source. Hence, at the receiver, the source is reconstructed only from the decoded bit-stream up to the first packet that is error-detected.

RS channel code is used here because of its bursty error correction capability. It should be pointed out that other channel codes combined with interleaver could also be used to have better performance. Our objective is to show that JSCM based on spatial diversity is an alternative solution for wireless multimedia communications where delay constraint and performance could be satisfied without interleaver and ARQ. Each OFDM block b_i is a RS codeword $\mathrm{RS}(N, K_i)$ over $\mathrm{GF}(2^m)$, where

$N = ((N_s log_2 M_{\rm ary})/m)$ is the total number of RS code symbols in each OFDM block, and $K_i = (B_i/m)$ is the number of RS information symbols in OFDM block $b_i, i = 1, 2, \ldots, L$. We denote the error-correction failure probability of OFDM block b_i as $P_e(b_i)$.

If the first i source-packets are correctly received, the image can be reconstructed to a rate $R = ((\sum_{j=1}^{i}(K_j m - 16))/HW)$ bpp. If the peak-signal-to-noise-ratio (PSNR) function of the source coder for the image is given by $\mathrm{PSNR}(R)$, then the JSCM algorithm is to look for a source packetization scheme $\mathcal{A}$, which specifies the sequence of source-packet size as $\{K_1 m, K_2 m, \ldots, K_L m\}$ to maximize the expected PSNR of the reconstructed image [10], [9]

$$\overline{\mathrm{PSNR}}_{\mathcal{A}} = \mathrm{PSNR}(0)P_e(b_1) + \sum_{i=1}^{L-1}\mathrm{PSNR}\left(\sum_{j=1}^{i}K_j m\right) \times P_e(b_{i+1})\prod_{j=1}^{i}[1 - P_e(b_j)] + \mathrm{PSNR}\left(\sum_{j=1}^{L}K_j m\right)\prod_{j=1}^{L}[1 - P_e(b_j)]. \quad (26)$$

It is difficult to use (26) in practice because the $\mathrm{PSNR}(R)$ function is not available at both transmitter and receiver, or a high quality channel is required to transmit the $\mathrm{PSNR}(R)$ side information from transmitter to receiver. An alternative criteria is to use the expected number of source bits correctly received as suggested in [12] where dynamic programming is used to solve the optimization problem similar to (26). Even though the use of the number of received bits is not as precise as using $\mathrm{PSNR}(R)$ or rate-distortion function $D(R)$, it was shown in [28] that the performance of using expected received bits is almost as good as using PSNR or rate-distortion function. The major difference between the JSCM we proposed here and the others is that, in our scheme, the channel codewords are of equal length, and we look for the length of source packet in each codeword such that the average distortion at receiver is minimized. We call this scheme variable source-packet length (VSL) JSCM to distinguish the equal source-packet length (ESL) algorithm used in most of current literatures where the source-packets are of equal length. The VSL method is more suitable than ESL here because the length of the codewords is limited in block data transmission scheme for OFDM and RS block codes.

If the average number of source bits received before errors occur is used to find the source-packet packetization scheme $\mathcal{A}$ for a given L, it can be written as

$$\max_{K_1, K_2, \ldots, K_L}\mathrm{BITS}_{\mathcal{A}} = \sum_{i=1}^{L-1}\left(\sum_{j=1}^{i}K_j m\right)P_e(b_{i+1}) \times \prod_{j=1}^{i}(1 - P_e(b_j)) + \left(\sum_{j=1}^{L}K_j m\right)\prod_{j=1}^{L}(1 - P_e(b_j)). \quad (27)$$

This optimization problem can be solved by dynamic programming as proposed in [12]. For integers $l, L, 1 \le l \le L$ and a packetization scheme $\mathcal{A}$, define

$$\Delta(l, L, \mathcal{A}) = \sum_{i=l}^{L-1}\left(\sum_{j=l}^{i}K_j m\right)\prod_{j=l}^{i}(1 - P_e(b_j))P_e(b_{i+1}) + \prod_{j=l}^{L}(1 - P_e(b_j))\left(\sum_{j=l}^{L}K_j m\right) \quad (28)$$

then we have the recursive relation for dynamic programming

$$\Delta(L, L, \mathcal{A}) = (1 - P_e(b_L))K_L m$$
$$\Delta(l, L, \mathcal{A}) = (1 - P_e(b_l))(K_l m + \Delta(l+1, L, \mathcal{A})). \quad (29)$$

Now the optimization problem in (27) becomes

$$\mathcal{A} = \max_{\{K_1, \ldots, K_L\}}\Delta(1, L, \mathcal{A}). \quad (30)$$

The optimization solution $\mathcal{A}^* = \{K_1^* m, \ldots, K_L^* m\}$ is used to packetize the SPIHT bitstream and protected by $\mathrm{RS}(N, K_i^*)$ over $\mathrm{GF}(2^m)$.

B. *Compute the Error Probability of OFDM Block*

The decoding failure probability of a OFDM block $\mathrm{RS}(N, K)$ over $\mathrm{GF}(2^m)$ under the assumption of mutual independence of OFDM subchannels is [27]

$$P_e(b) = 1 - \sum_{v=0}^{(N-K)/2}\binom{N}{v}P_s^v(1 - P_s)^{N-v} \quad (31)$$

where P_s is the error probability of RS code symbol (m bits per RS code symbol), which is

$$P_s = 1 - (1 - P_{ce})^{m/b} \quad (32)$$

where b is the number of bits for each subchannel symbol, and m/b is the number of subchannel symbols in a RS code symbol. e.g., for QPSK and $\mathrm{GF}(2^6), m = 6, b = 2$.

All the computations in (31) and (32) are based on the assumption that the subchannels in OFDM with (N, M) STBC are mutually independent Gaussian noisy channels with the same SNR, which is true when NM goes to infinity. For finite NM of STBC (N, M), the assumption doesn't hold. Also, the instantaneous SNR at receiver, denoted as γ_c, is time-varying. Because the transmitter cannot know the γ_c, the JSCM scheme has to be designed for a single SNR, which we call *target SNR* and denote as γ_T. The *target SNR* γ_T is computed for a given probability P_B such that $P(\gamma_c \le \gamma_T) = P_B$. By specifying a relatively small P_B, the γ_c is larger than γ_T with high probability and the system performance can be guaranteed. We call this scheme *worst-case JSCM design*. The advantage of using STBC or spatial diversity is that the *target SNR* γ_T approaches the average SNR as the NM increases. As a result, the system performance can be improved.

For a OFDM system using (N, M) STBC, the instant SNR γ_c at subchannels with a known average receiver SNR $\bar{\gamma}_c$ is

$$\gamma_c = \frac{\bar{\gamma}_c}{NM}\sum_{i=1}^{NM}C_i^2 \quad (33)$$

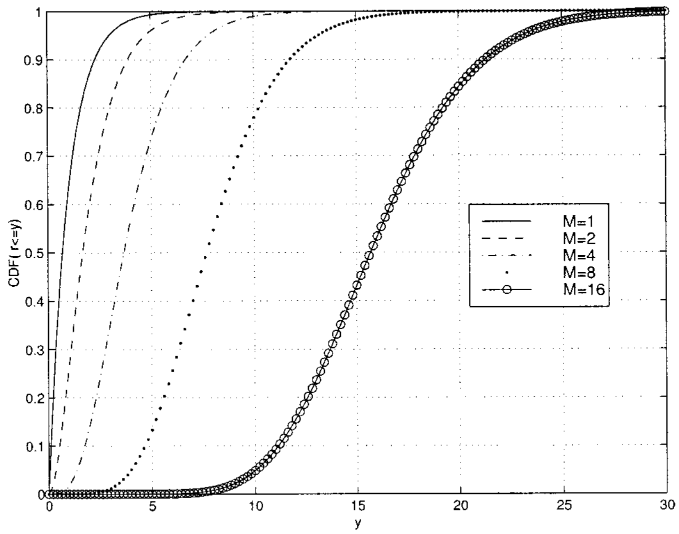

Fig. 4. The CDF of diversity gains for receiver diversity using M antennas.

TABLE I
THE *TARGET SNR* FOR AVERAGE RECEIVER $\mathrm{SNR} = 1$ AT DIFFERENT P_B UNDER DIFFERENT (N, M) STBC CASES

	$(1,1)$	$(2,1)$	$(2,2)$	$(2,4)$	$(2,8)$
$P_B = 0.01$	0.0101	0.0743	0.2058	0.3633	0.5113
$P_B = 0.05$	0.0513	0.1777	0.3416	0.4976	0.6272
$P_B = 0.10$	0.1054	0.2659	0.4362	0.5820	0.6960

where C_i is the amplitude of complex zero-mean Gaussian random variable with variance 0.5 at each dimension. Let $x = \sum_{i=1}^{NM} C_i^2$, then x is a χ^2 random variable with $2NM$-degree freedom. To find the *target SNR* γ_T such that $p(\gamma_c \le \gamma_T) = P_B$ for a given P_B, we only need to find a real value y such that $p(x \le y) = P_B$, then using (33) we have

$$\gamma_T = \frac{\bar{\gamma}_c y}{NM}. \tag{34}$$

The cumulative distribution function (CDF) of $x = \sum_{i=1}^{NM} C_i^2$ is [24]

$$p(x \le y) = 1 - e^{-y/2N_0} \sum_{k=0}^{NM-1} \frac{1}{k!} \left(\frac{y}{2N_0}\right)^k, \quad y \ge 0 \tag{35}$$

where the CDF of x is shown in Fig. 4 for different NMs. It shows that the fading effect can be significantly decreased as NM increases. To show the advantage of using diversity, the γ_T for different (N, M) STBC at same average receiver SNR $\bar{\gamma}_c = 1$ is listed in Table I for several P_B's. Indeed, as $NM \to \infty, p(\gamma_c = \bar{\gamma}_c) \to 1$, which follows the Proposition 1.

After the *target SNR* γ_T is obtained, the *worst-case* subchannel symbol error probability of Gray-mapped QPSK is [24]

$$P_{\mathrm{ce}} = bP_b = bQ(\sqrt{\gamma_T}). \tag{36}$$

For other MPSK symbols except BPSK and QPSK, P_{ce} has to be obtained numerically for a given γ_T.

In general, the scheme for robust progressive image transmission over OFDM systems with (N, M) STBC is summarized as follows.

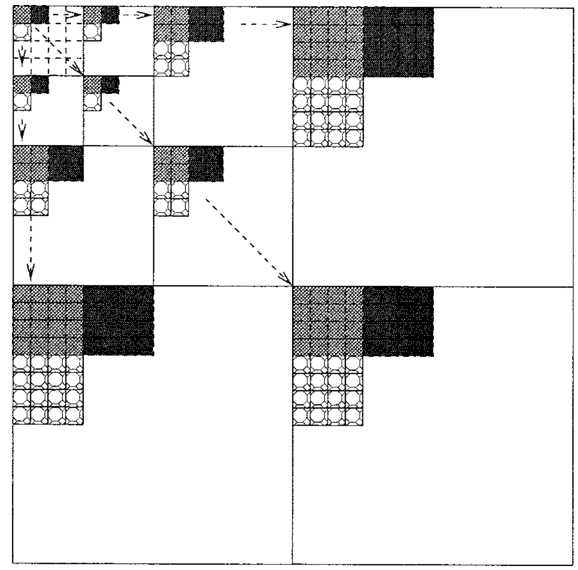

Fig. 5. Partition and group method for substreams of wavelet coefficients.

1) Given the system parameters:

N_s	number of subchannels in OFDM;
L	number of OFDM blocks will be transmitted;
$\bar{\gamma}_c$	average SNR at receiver;
(N, M)	number of transmit antennas and receive antennas;
m	RS channel code is based on $\mathrm{GF}(2^m)$, or m bits per channel code symbol;
b	number of bits for each subchannel symbol.

2) The *target SNR* γ_T is computed for a specified P_B using (34).
3) Compute the BER and subchannel symbol error rate (SER) using (36).
4) Compute the RS code symbol SER and OFDM block decoding failure probability using (32) and (31).
5) Find the source-packet packetization scheme $\mathcal{A}^* = \{K_1^*, \ldots, K_L^*\}$ by solving (27). The obtained variable source-packet lengths are known by both transmitter and receiver through a reliable control channel.
6) The SPIHT bitstream is packetized into L blocks with $K_i^* m - 16$ bits for block $i, i = 1, \ldots, L$ which is CRC-16 outer coded, then RS inner coded using $\mathrm{RS}(N, K_i^*)$ over $\mathrm{GF}(2^m)$ to form a OFDM block.
7) The OFDM blocks is space-time coded and transmitted as described in Section II-B.

The receiver performs the same computation as given above to get the size of source-packet $K_i^*, i = 1, 2, \ldots, L$ for the given parameters. Then, after demodulation and RS decoding, the receiver uses the CRC-16 to detect the first error source-packet if any, and reconstructs the image using SPIHT decoder.

V. SIMULATION RESULTS

In our simulation, we use a two-ray channel model with delay spread from 0 to 40 μs and Doppler frequency from 10 Hz to 200 Hz. Two transmit antennas and $M = 2, 4$ receive antennas are used for space-time block code. The entire channel bandwidth, 800 kHz, is divided into 128 subchannels. Four subchannels on each end are used as guard tones and the rest 120 tones are used

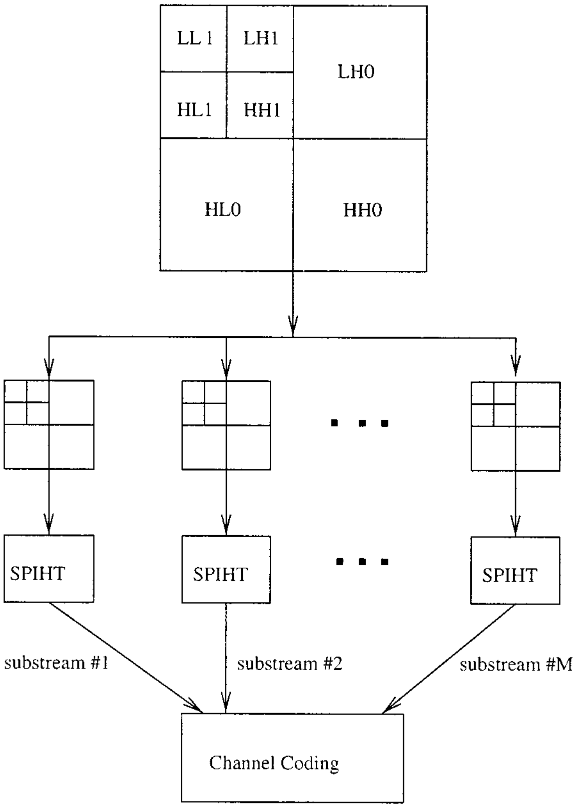

Fig. 6. Structure of SPIHT for substreams.

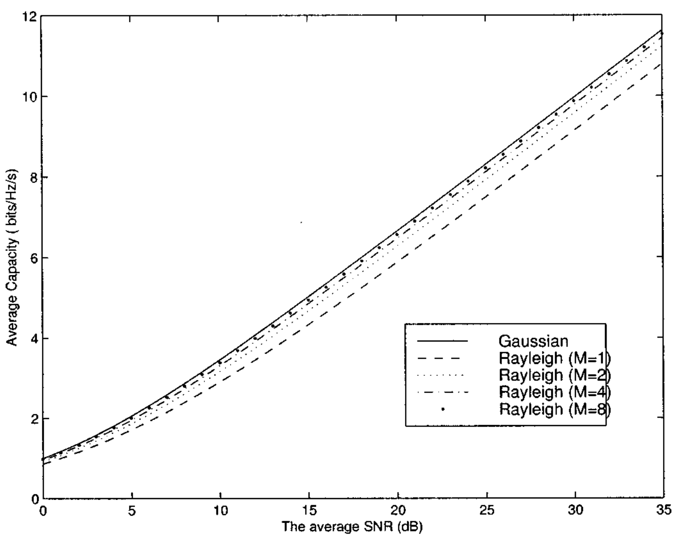

Fig. 7. The spectral capacity density per unit time on the Rayleigh fading channel, M-fold diversity channels $(M = 2, 4, 8)$ and the AWGN channel.

to transmit data. To make the tones orthogonal to each other, the symbol duration is 160 μs. An additional 40 μs guard interval is used to provide protection from ISI due to channel multipath delay spread. This results in a total block length $T_f = 200$ μs and a subchannel symbol rate $r_b = 5$ kBd. QPSK modulation and coherent estimation is used with the assumption of perfect channel information at receivers. The punctured and/or shortened RS code over $\mathrm{GF}(2^8)$ is used such that each OFDM block has $120 \times 2/8 = 30$ RS code symbols, and the number of information symbols in each OFDM block is in the range

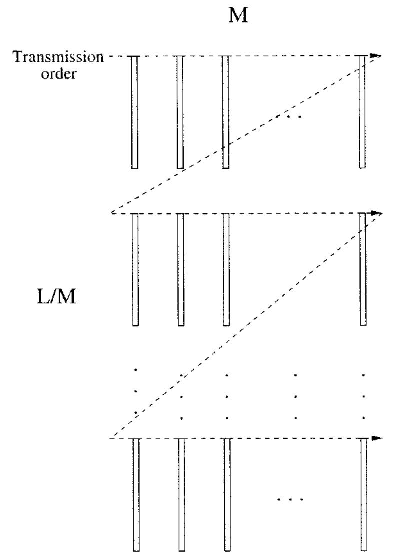

Fig. 8. The transmission order for MDC SPIHT substreams.

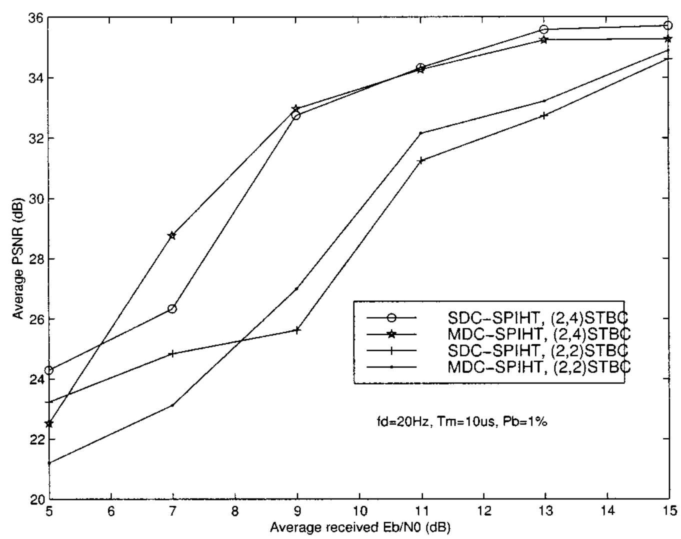

Fig. 9. PSNR comparison between MDC-SPIHT and SDC-SPIHT at (2, 4) and (2, 2) STBC, $f_d = 20$ Hz, $\tau_m = 10$ μs, $P_B = 0.01$.

$\{3, 4, \ldots, 26\}$. In all the simulations, 300 times image transmission are simulated under each case. The transmit rate used for all the simulations is 0.5 bpp, which is approximately equivalent to $L = 576$ OFDM blocks.

A. Performance Comparisons

The discrete wavelet transform of 512×512 gray image Lena is obtained first, then two schemes are compared for given system parameters.

1) *Single stream transmission (SDC-SPIHT):* The DWT coefficients is encoded into a SPIHT bitstream, and is packetized into L packets with $K_i^* m - 16, i = 1, 2 \ldots, L$ bits for packet i. The packetization scheme is obtained through the JSCM scheme described above. Then the source-packets are CRC-16 outer coded and RS inner coded to form OFDM blocks and sent out through STBC encoder.

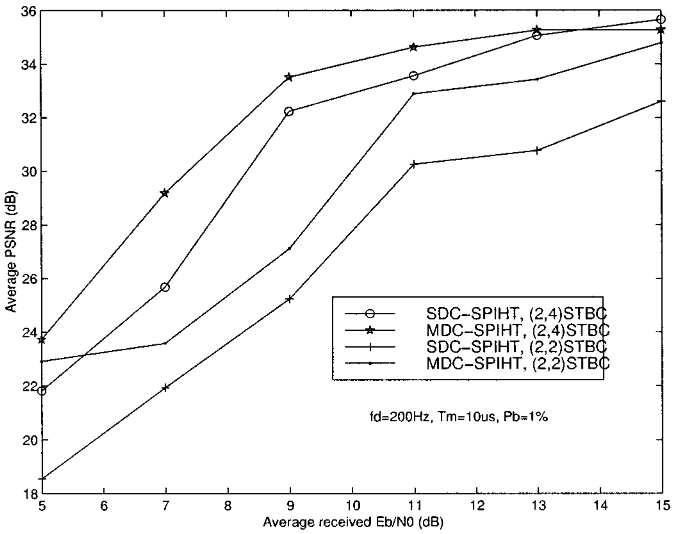

Fig. 10. PSNR comparison between MDC-SPIHT and SDC-SPIHT at (2, 4) and (2, 2) STBC, $f_d = 200$ Hz, $\tau_m = 10\ \mu$s, $P_B = 0.01$.

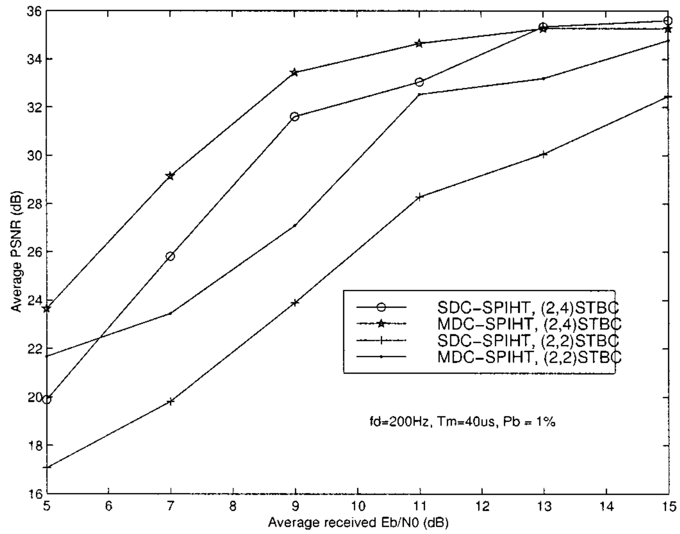

Fig. 12. PSNR comparison between MDC-SPIHT and SDC-SPIHT at (2, 4) and (2, 2) STBC, $f_d = 200$ Hz, $\tau_m = 40\ \mu$s, $P_B = 0.01$.

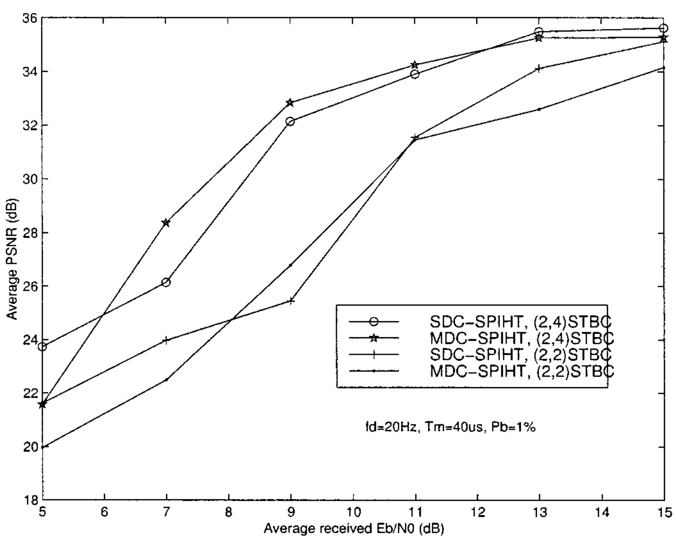

Fig. 11. PSNR comparison between MDC-SPIHT and SDC-SPIHT at (2, 4) and (2, 2) STBC, $f_d = 20$ Hz, $\tau_m = 40\ \mu$s, $P_B = 0.01$.

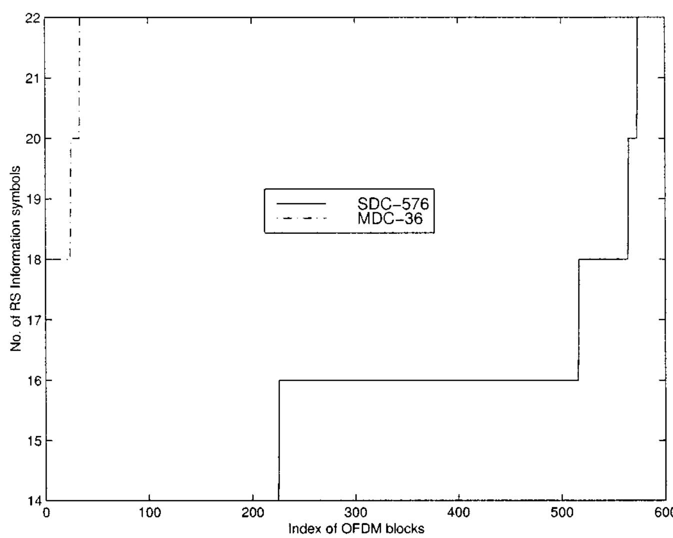

Fig. 13. Packetization scheme comparison between MDC-SPIHT and SDC-SPIHT at (2, 4) STBC, where packetization for MDC-SPIHT is based on 36 OFDM blocks, SDC-SPIHT is based on 576 OFDM blocks.

2) *Multiple substream transmission (MDC-SPIHT):* The DWT coefficients are partitioned into M_s subimages, and each subimage keeps the tree structure across scales as shown in Fig. 5, then each subimage is SPIHT coded as shown in Fig. 6. It was shown in [29], [30] that the partition of the coefficients into subimage is robust to bits and packet error in noisy channels. Each SPIHT substream is packetized into $K_i^* m - 16, i = 1, 2 \ldots, L/M_s$, which is obtained through the JSCM scheme for L/M_s blocks. Then the source packet i is CRC-16 outer coded and $\mathrm{RS}(30, K_i)$ inner coded to form a OFDM block. The OFDM blocks is transmitted across the substreams as shown in Fig. 8 such that the OFDM blocks within one substream are interleaved while the progressive transmission property is still maintained with a delay of M_s OFDM blocks. In the simulation the number of subimage used is 16.

In Figs. 9–12, the average PSNR of the received images is compared with $P_B = 0.01$ using STBC (2, 4) and (2, 2), under different delay spread and Doppler spread parameters. It can be observed that at the same average received SNR, the image quality using STBC (2, 4) is always better than using STBC (2, 2) because the *target SNR* of STBC (2, 4) is larger than that of STBC (2, 2), as a result, the JSCM scheme has larger throughput for STBC (2, 4) than that of STBC (2, 2). Also, when the diversity gain NM is increased, the subchannels in OFDM system approaches Gaussian noisy channel, which is the assumption on which the JSCM algorithm is based.

We can also note that the performance of MDC-SPIHT is better than that of SDC-SPIHT at moderate average SNR, this is because of the following two reasons.

- In SDC-SPIHT, when one source-packet is error-detected, then all the source-packets followed can not be used; In

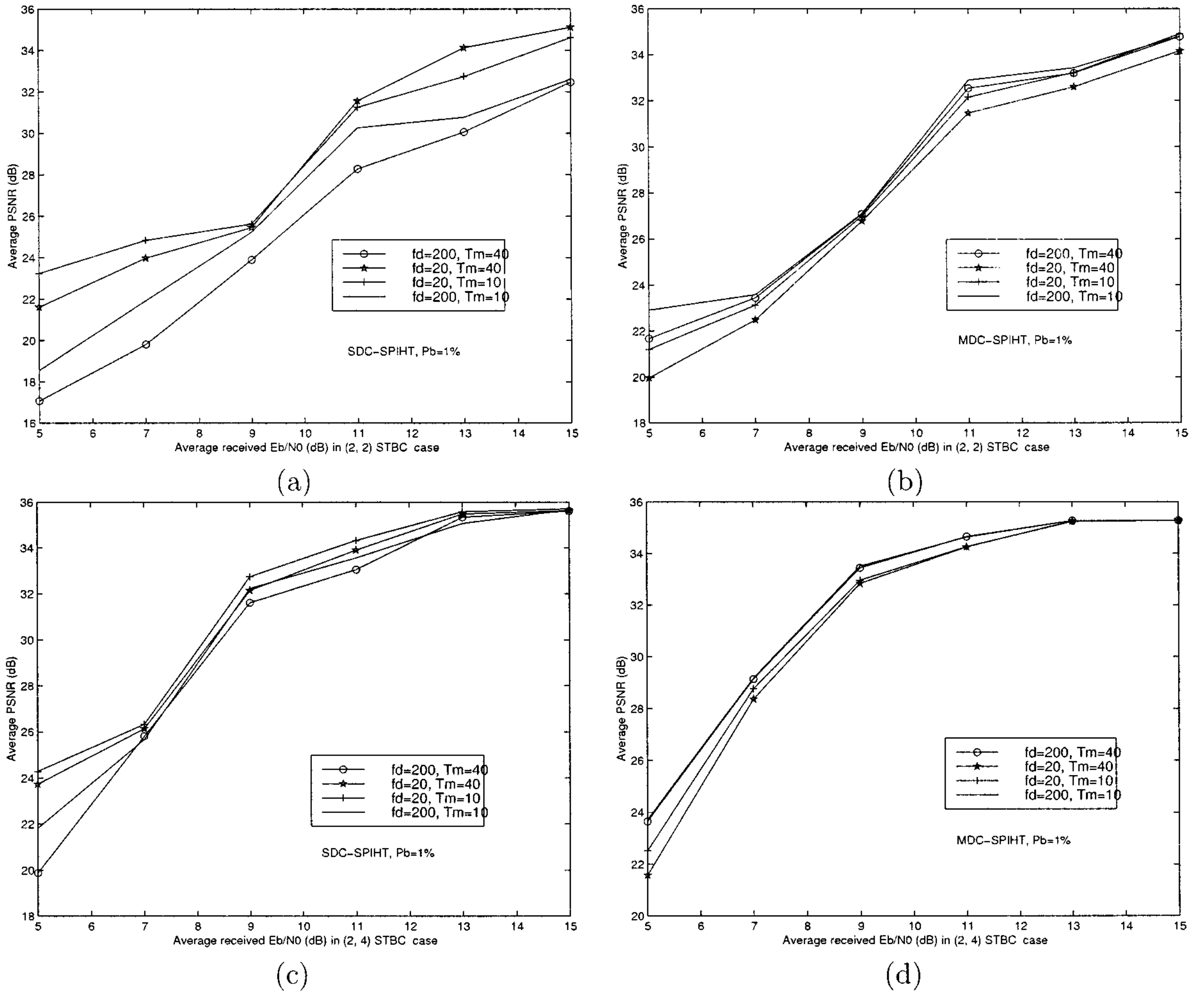

Fig. 14. Performance comparison at different multipath delay spread and Doppler frequency: (a) SDC-SPIHT with STBC(2, 2), (b) MDC-SPIHT with STBC(2, 2), (c) SDC-SPIHT with STBC(2, 4), and (d) MDC-SPIHT with STBC(2,4).

MDC-SPIHT, however, the error of one source-packet only affect the source-packets followed in that single substream. The interpolation from the neighbor coefficients at LL Subband from other substreams can be used to conceal the error effects.

- The effective source-rate of MDC-SPIHT is larger than the source-rate of SDC-SPIHT in most cases. For a given transmission rate R_t, the corresponding source rate for MDC-SPIHT is

$$R_{\mathrm{mdc}} = R_t \frac{M_s \sum_{i=1}^{L/M_s}(K_i - 2)}{LN} \tag{37}$$

and for SDC-SPIHT is

$$R_{\mathrm{sdc}} = R_t \frac{\sum_{i=1}^{L}(K_i - 2)}{LN}. \tag{38}$$

For a given average SNR, the packetization scheme for one MDC-SPIHT substream (36 OFDM blocks) is the same as the last 36 OFDM blocks in SDC-SPIHT. For example, the source-packetization scheme obtained when $E_b/N_0 = 9$ dB for both MDC-SPIHT and SDC-SPIHT is shown in Fig. 13. Because of the progressive property of SPIHT bitstream, we know that $K_j \leq K_i$, for $j \leq i$, then we have $M_s \sum_{i=1}^{L/M_s}(K_i - 2) > \sum_{i=1}^{L}(K_i - 2)$, as a result, $R_{\mathrm{sdc}} \leq R_{\mathrm{mdc}}$.

On the other hand, when the SNR is large enough, the throughput limit is reached such that $R_{\mathrm{mdc}} = R_{\mathrm{sdc}}$. Then SDC-SPIHT has better performance than MDC-SPIHT because the source coding efficiency of SDC-SPIHT is better than MDC-SPIHT.

The performance under different multipath delay spread and Doppler frequency for same STBC configuration are also compared. The average PSNR of SDC-SPIHT and MDC-SPIHT at different delay spread and Doppler spread is shown in Fig. 14(a)–(b) using STBC (2, 2). The same comparison for STBC (2, 4) is also shown in Fig. 14(c)–(d). We can see that the multipath delay spread and Doppler spread have less effects to the performance when the diversity gain NM increases. Actually, when $NM \geq 8$, the system performances are almost the same under different multipath delay and Doppler spread because the OFDM subchannels approach Gaussian noisy channels.

The received images of “Lena” is also shown in Fig. 15, when $E_b/N_0 = 9$ dB, where the delay spread $\tau_m = 40$ μs, Doppler frequency $f_d = 200$ Hz. The probability for *target SNR*

(a) PSNR = 25.72 dB

(b) PSNR = 27.12 dB

(c) PSNR = 31.24 dB

(d) PSNR = 33.59dB

Fig. 15. Received images of "Lena" when $E_b/N_0 = 9$ dB at (a) SDC-SPIHT with STBC(2, 2), (b) MDC-SPIHT with STBC(2, 2), (c) SDC-SPIHT with STBC(2, 4), and (d) MDC-SPIHT with STBC(2, 4). Doppler frequency $f_d = 200$ Hz, multipath delay $\tau_m = 40\ \mu$s, $P_B = 1\%$ for JSCM.

$P_b = 1\%$ is used in JSCM algorithm. At the same STBC configurations [(2, 4) or (2, 2)], MDC-SPIHT has better image quality (Fig. 15(b) and (d)) than that of SDC-SPIHT [Fig. 15(a) and (d)], respectively. The image quality of STBC (2, 4) as shown in Fig. 15(c) and (d) is much better than the image quality of STBC (2, 2) as shown in Fig. 15(a) and (b).

B. Error Propagation and Concealment

Even though the fading effects can be decreased significantly, the errors may still occur and propagate. Fig. 16(a) shows a sample of channel gain of OFDM channels with one transmitter and one receiver over two-ray fading channels. Fig. 16(b) shows a sample of channel gain with two transmitter and four receivers with the same average SNR. It can be noted that the dynamic range of Fig. 16(b) is much smaller than Fig. 16(a) because of using multiple antennas. To illustrate the error propagation and concealment effects, we assume that one of the 16 substreams is lost. Fig. 17(a) shows the decoded image when one of the substream is lost and the corresponding coefficients are replaced by zero. The decoded image has dark "holes" spread out to an extent determined by the number of decomposition levels and the

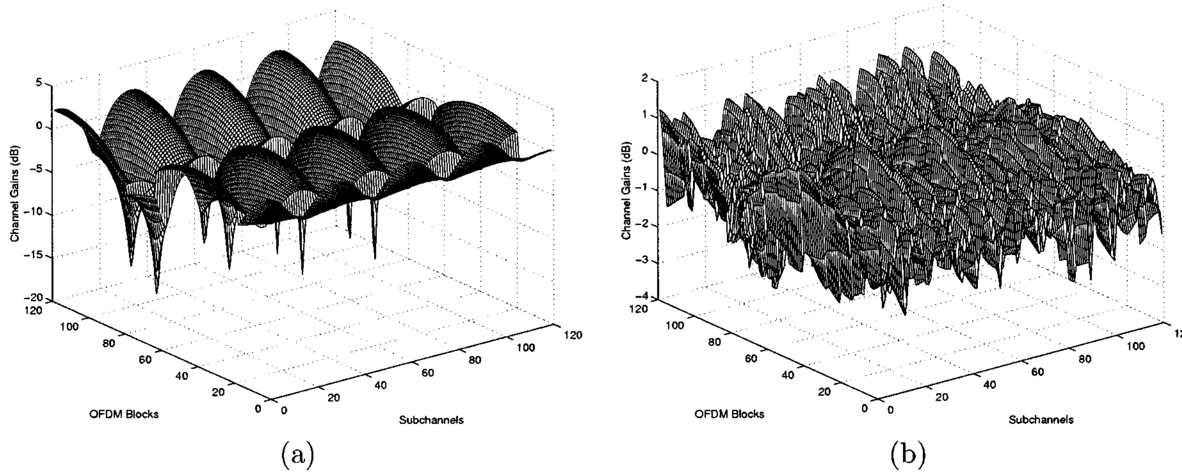

Fig. 16. Channel gains using two-ray Rayleigh fading channel model: (a) one transmitter and one receiver and (b) two transmitters and four receivers.

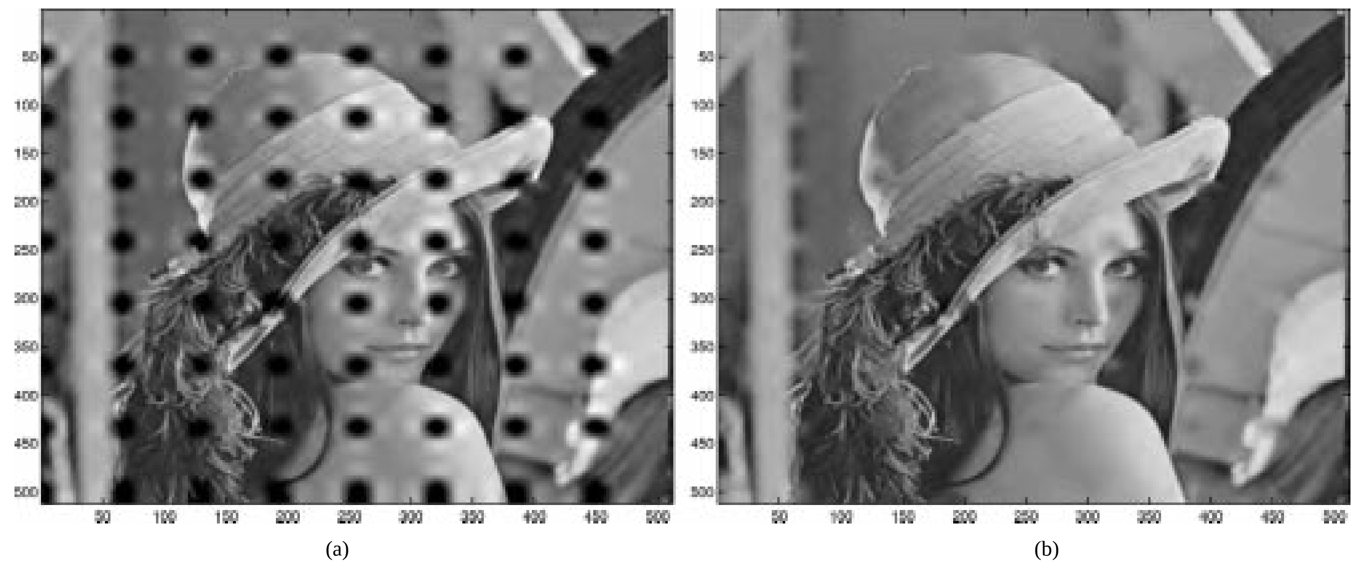

Fig. 17. The reconstructed images using MDC when one substream is lost: (a) without error concealment (the lost coefficients are replaced by zero) and (b) With simple error concealment. (The lost low frequency coefficients are replaced by interpolation from neighboring coefficients.)

filter length. To conceal the lost coefficients, we simply substitute low-frequency coefficients of the lost coefficients by the average of the neighboring correctly received coefficients. The corresponding high-frequency coefficients are still replaced by zero. Fig. 17(b) shows the recovered image using the simple concealment method. It can be noticed that the image quality is improved significantly at smooth area. However, there is blurring in synthesized edges since high frequency coefficients are not recovered. Advanced image reconstruction techniques, such as in [31], can be employed to improve the quality by recovering both low frequency and high frequency coefficients using different schemes.

VI. Conclusion

A joint source–channel matching framework is proposed for SPIHT coded image transmission over OFDM systems with multiple antennas using space–time block code. By using spatial diversity, the fading effect of wireless channels can be significantly decreased. After showing that the OFDM subchannels are indeed parallel independent Gaussian noisy channel in a frequency-selective, time-dispersive Rayleigh fading environment when the number of antennas goes to infinity, a joint source–channel matching scheme is proposed for OFDM systems using multiple antennas. The simulation results show that the performance is improved significantly in terms of image rate throughput and channel coding efficiency. To further decrease the channel impairment, we use an error resilient SPIHT image coding algorithm to divide the image into several subimages, where the proposed joint source–channel matching scheme can be applied directly to each SPIHT encoded subimage. By doing so, the error propagation is only constrained within one substream and error concealment techniques can be employed to achieve better reconstruction at the receiver. The simulation results show that this scheme is robust to different Doppler and multipath delay spread in wireless communications.

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