Decoding Schemes for Amplify-and-Forward Cooperative OFDM-Based Wireless LAN Systems

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***Abstract*—Cooperative communication can be regarded as virtual multiple-input multiple-output (MIMO) systems where distributed multiple terminals behave as a virtual antenna array. This paper presents two space-time coding (STC) methods, i.e., the Golden code and Alamouti code, for cooperative orthogonal frequency division multiplexing (OFDM)-based wireless local area network (LAN) systems. We propose a decoding scheme for both STC methods. The proposed decoding scheme for the Golden code has approximate maximum likelihood (ML) property, and the one for the Alamouti code has exact ML property. We show the performances of the two proposed schemes for various typical relay locations through simulations. Simulation results reveal that energy gain greater than 2.1 dB can be obtained with the proposed schemes over non-cooperative (direct) communication in each relay location at a packet error rate of 10^{-1}.**

I. Introduction

Recently, there has been a growing interest in the design and analysis of cooperative communication due to improvement of performance [1], [2]. The fundamental principle of cooperative communication is to establish multiple antennas transmission in a distributed manner by allowing neighboring user terminals to share their spatial resources. Consequently, cooperative communication systems can be considered as the virtual multiple-input multiple-output (MIMO) system.

Nabar *et al.* [2] analyze cooperative communication theoretically in a one-relay two-hop (1R2H) topology. Cooperative communication under the 1R2H assumption can be virtual 2×2 MIMO, 1×2 single-input multiple-output (SIMO) or 2×1 multiple-input single-output (MISO) systems depending on the cooperative protocol. Among the cooperative protocols, Protocol I [2], the virtual 2×2 MIMO, has the best capacity and diversity performance due to its highest degree of freedom. In addition to cooperative protocols, a variety of forwarding modes in the relay node have been studied and analyzed extensively, including modes such as amplify-and-forward (AF) and decode-and-forward (DF) [1], [2]. Furthermore, distributed space-time coding (DSTC) can be employed with virtual multiple input systems to obtain diversity gain [2]-[4]. Nabar *et al.* [2] and Shin *et al.* [3] adopt the well-known Alamouti code [5] in a distributed manner since it provides full diversity with a simple decoding algorithm, with complexity of $O(\log_2|\mathcal{A}|)$, where $|\mathcal{A}|$ is the size of the underlying quadrature amplitude modulation (QAM) constellation $\mathcal{A}$. Yang *et al.* [4] adopt the Golden code, one of the STC methods for realizing spatial multiplexing (SM), to improve the spectral efficiency and achieve optimal performance in terms of diversity and multiplexing tradeoff. In [6], it is shown that the Golden code has better error rate performances than the Alamouti code on systems without channel encoding. The Golden code requires a high maximum likelihood (ML) decoding complexity, at least of $O(|\mathcal{A}|^2)$ [7].

As discussed above, cooperative communication techniques have been studied in many papers, but some topics have not been addressed sufficiently. First, many papers have reported theoretical analyses for cooperative communication, but relatively few have investigated the adoption of cooperative communication on practical systems. Second, narrowband transmission in flat fading channel environments has often been assumed for cooperative communication, but frequency selective channel environments should also be investigated for broadband transmission.

In this paper, we evaluate and compare some cooperative schemes on an orthogonal frequency division multiplexing (OFDM) system. The cooperative OFDM system is based on the wireless local area network (LAN) standard IEEE 802.11a [8] and we assume a frequency selective fading environment. Due to its high potential, we use cooperative protocol I in [2] on 1R2H systems. We focus on AF as a forwarding mode in the relay node because of its simple signal processing and short latency compared to DF. We employ two STC methods, the Golden code and the Alamouti code, and propose a decoding scheme for each of them. The scheme we propose for the Golden code utilizes per-antenna-coded soft-decision output maximum likelihood detection (PAC SOMLD) [9], with approximate ML property. For the Alamouti code, we propose a scheme with exact ML property that includes linear space-time decoding based on the maximum ratio combining (MRC) matrix form. We also derive subcarrier weighting factors prior to the soft-decision Viterbi decoding as a channel decoding in order to achieve ML property. The proposed scheme for the Alamouti code has a low complexity of $O(\log_2|\mathcal{A}|)$, which is much smaller than that of decoding for the Golden code, at least $O(|\mathcal{A}|^2)$. Under these assumptions, we evaluated simulated performances of cooperative schemes and non-cooperative (direct) communication in terms of error rate and service range in various typical relay locations. Our simulation results show that our scheme for the Alamouti code gives it performances very close to those of the Golden code.

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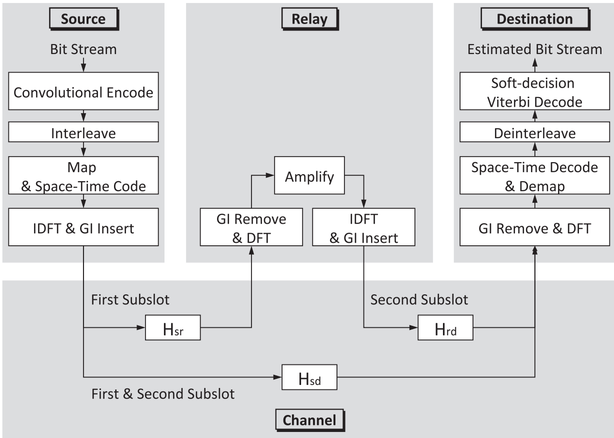

Fig. 1. Block diagram of a cooperative OFDM-based wireless LAN system. The source structure is based on the IEEE 802.11a standard [8].

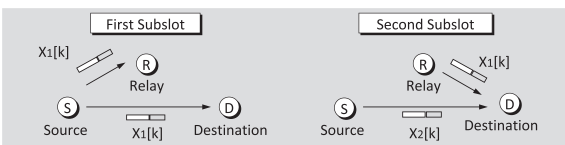

Fig. 2. Schematic of cooperative communication in Protocol I [2], adopted in this paper.

Therefore, we conclude that the scheme for the Alamouti code is appropriate for use in cooperative OFDM-based wireless LAN systems.

The rest of this paper is organized as follows. Section II describes system models, including a transceiver architecture and signal and channel models. Section III introduces two cooperative STC methods and the decoding schemes we propose for them. In Section IV, we use numerical simulation results to analyze the advantages of cooperative communication over non-cooperative communication. Section V concludes the paper with a brief summary of key points.

II. System Models

In this section we describe cooperative communication system models based on the IEEE 802.11a standard [8].

A. Transceiver Architecture

Fig. 1 illustrates block diagrams of the source, relay, and destination. The structure of the source transmitter is similar to that given in the IEEE 802.11a standard [8]. Data bits are first scrambled and then encoded using convolutional encoders with a code rate R_c. Encoded bits are interleaved and then mapped into transmit constellation points from a quadrature amplitude modulation (QAM) constellation $\mathcal{A}$ in a Gray code manner. An inverse discrete Fourier transform (IDFT) of the K subcarriers including data constellations, pilots, and nulls results in K samples of time domain signals. A guard interval (GI), which is a cyclic prefix of the original signal, is appended in front of the K time domain samples to form complete OFDM symbols. Each OFDM symbol is demodulated at the destination by executing the inverse of the source transmitter operation. In particular, the soft-decision Viterbi algorithm is employed to decode the convolutionally encoded data bits. At the relay, the received signals are amplified in frequency domain. The relay has GI remove/insert and DFT/IDFT processes to make them in frequency domain.

B. Signal and Channel Models

Fig. 2 shows the cooperative communication schematic in Protocol I, used in our work. Each channel can be regarded as K flat fading channel systems in the frequency domain with OFDM transmission. In the first subslot, the frequency domain signal received at the destination $Y_1[k]$ is given as

$$Y_1[k] = \sqrt{P_{\mathrm{S1}} L_{\mathrm{SD}}} H_{\mathrm{SD}}[k] X_1[k] + W_1[k] \quad (1)$$

where k is the subcarrier index with $k = 0, 1, \cdots, K-1$, P_{S1} is the transmit power of the source in the first subslot, $H_{\mathrm{SD}}[k]$ is the channel frequency response (CFR) of the k-th subcarrier between the source and the destination (complex Gausssian fading gain with zero mean and unit variance), $X_1[k]$ is the transmit signal of the k-th subcarrier from the source in the first subslot and normalized as $\mathrm{E}[|X_1[k]|^2] = 1$, and $W_1[k]$ is the additive white Gaussian noise (AWGN) of the k-th subcarrier at the destination in the first subslot with zero mean and variance of σ_w^2. Note that the variance σ_w^2 is constant for all the subcarriers because the noise is a white process. We express large-scale propagation loss between the source and destination as L_{SD}. A signal received at the relay of the k-th subcarrier is expressed as

$$Y_{\mathrm{R1}}[k] = \sqrt{P_{\mathrm{S1}} L_{\mathrm{SR}}} H_{\mathrm{SR}}[k] X_1[k] + W_{\mathrm{R1}}[k] \quad (2)$$

where $H_{\mathrm{SR}}[k]$ is the CFR of the k-th subcarrier between the source and the relay (complex Gausssian fading gain with zero mean and unit variance) and $W_{\mathrm{R1}}[k]$ is the AWGN at the relay in the first subslot with zero mean and variance of σ_w^2, respectively. We express large-scale propagation loss between the source and the relay as L_{SR}. When the AF mode is used, the relay uses the amplifying coefficient $\alpha_{\mathrm{R}}[k]$ to amplify the received signal $Y_{\mathrm{R1}}[k]$ and transmits the signal $\alpha_{\mathrm{R}}[k] Y_{\mathrm{R1}}[k]$ to the destination with transmit power P_{R2} in the second subslot. The amplifying coefficient $\alpha_{\mathrm{R}}[k]$ is a normalizing coefficient and various methods are used to calculate it. In this paper, the instantaneous power scaling (IPS) [10] method is employed: $\alpha_{\mathrm{R}}[k] = 1/\sqrt{P_{\mathrm{S1}} L_{\mathrm{SR}} |H_{\mathrm{SR}}[k]|^2 + \sigma_w^2}$. In the second subslot, the source transmits $X_2[k]$ and the relay transmits the amplified signal. Accordingly, we describe the signal received at the destination of the k-th subcarrier as

$$\begin{aligned} Y_2'[k] =& \alpha_{\mathrm{R}}[k] \sqrt{P_{\mathrm{S1}} L_{\mathrm{SR}} P_{\mathrm{R2}} L_{\mathrm{RD}}} H_{\mathrm{SRD}}[k] X_1[k] \\ &+ \sqrt{P_{\mathrm{S2}} L_{\mathrm{SD}}} H_{\mathrm{SD}}[k] X_2[k] + W_{\mathrm{T}}[k] \end{aligned} \quad (3)$$

where P_{S2} is the transmit power of the source in the second subslot, $H_{\mathrm{RD}}[k]$ is the CFR of the k-th subcarrier between the relay and destination (complex Gausssian fading gain with zero mean and unit variance), and $X_2[k]$ is the transmit signal of the k-th subcarrier from the source in the second subslot and normalized as $\mathrm{E}[|X_2[k]|^2] = 1$. We express large-scale propagation loss between the relay and destination as L_{RD}. The equivalent CFR $H_{\mathrm{SRD}}[k]$ satifies $H_{\mathrm{SRD}}[k] = H_{\mathrm{SR}}[k]H_{\mathrm{RD}}[k]$. The equivalent noise term $W_{\mathrm{T}}[k]$ at the destination in the second subslot is the sum of the additive noise at the destination in the second subslot and the propagated noise from the relay, expressed as

$$W_{\mathrm{T}}[k] = H_{\mathrm{RD}}[k]\sqrt{P_{\mathrm{R2}}L_{\mathrm{RD}}}\alpha_{\mathrm{R}}[k]W_{\mathrm{R1}}[k] + W_2'[k] \tag{4}$$

where $W_2'[k]$ is the AWGN of the k-th subcarrier at the destination in the second subslot with zero mean and variance of σ_w^2. If we assume that $W_{\mathrm{R1}}[k]$, $W_1[k]$, and $W_2'[k]$ are independent random processes, then $W_{\mathrm{T}}[k]$ is AWGN with zero mean and variance of

$$\mathrm{E}\left[|W_{\mathrm{T}}[k]|^2 \middle| H_{\mathrm{SR}}[k], H_{\mathrm{RD}}[k]\right] = \frac{\sigma_w^2}{\rho[k]^2} = \sigma_{w_T}[k]^2 \tag{5}$$

where $\mathrm{E}\left[A\middle|B\right]$ means a conditional expectation of A given by B and $\rho[k]$ is $\rho[k] = 1/\sqrt{\alpha_{\mathrm{R}}[k]^2 P_{\mathrm{R2}}L_{\mathrm{RD}}\,|H_{\mathrm{RD}}[k]|^2+1}$. Following [2] or [4], we assume that the receiver uses the factor $\rho[k]$ to normalize $Y_2'[k]$ in order to realize decodings with approximate or exact ML property. Then, from (3) we have

$$Y_2[k] = H_{2,1}[k]X_1[k] + H_{2,2}[k]X_2[k] + W_2[k] \tag{6}$$

where

$$\begin{aligned} Y_2[k] &= \rho[k]Y_2'[k],\\ H_{2,1}[k] &= \rho[k]\alpha_{\mathrm{R}}[k]\sqrt{P_{\mathrm{S1}}L_{\mathrm{SR}}P_{\mathrm{R2}}L_{\mathrm{RD}}}H_{\mathrm{SRD}}[k],\\ H_{2,2}[k] &= \rho[k]\sqrt{P_{\mathrm{S2}}L_{\mathrm{SD}}}H_{\mathrm{SD}}[k],\\ W_2[k] &= \rho[k]W_{\mathrm{T}}[k]. \end{aligned} \tag{7}$$

Here, $\mathrm{E}\left[|W_2[k]|^2 \middle| H_{\mathrm{SR}}[k], H_{\mathrm{RD}}[k]\right] = \sigma_w^2$, and so the variance of $W_2[k]$ becomes σ_w^2.

When DSTC is employed with 1R2H topology, each pair of OFDM symbols in a packet forms a space-time coding block [3]. Specifically, a block of space-time encoded transmit signals for the k-th data subcarrier in the m-th and $(m+1)$-th OFDM symbols can be expressed in matrix form , where m is an odd-number OFDM symbol index with $m = 1, 3, 5, \cdots$. Accordingly, the effective input-output relation between the transmitted and received signals here can be given in matrix form as

$$\boldsymbol{Y}[k] = \boldsymbol{H}[k]\boldsymbol{X}[k] + \boldsymbol{W}[k] \tag{8}$$

where $\boldsymbol{Y}[k]$ and $\boldsymbol{X}[k]$ are the received and transmitted signal matrices given by

$$\boldsymbol{Y}[k] = \begin{bmatrix} Y_{1,m}[k] & Y_{1,m+1}[k] \\ Y_{2,m}[k] & Y_{2,m+1}[k] \end{bmatrix}, \quad \boldsymbol{X}[k] = \begin{bmatrix} X_{1,m}[k] & X_{1,m+1}[k] \\ X_{2,m}[k] & X_{2,m+1}[k] \end{bmatrix} \tag{9}$$

respectively, $\boldsymbol{H}[k]$ and $\boldsymbol{W}[k]$ are the effective 2×2 channel matrix and the AWGN matrix given by

$$\boldsymbol{H}[k] = \begin{bmatrix} H_{1,1}[k] & 0 \\ H_{2,1}[k] & H_{2,2}[k] \end{bmatrix}, \quad \boldsymbol{W}[k] = \begin{bmatrix} W_{1,m}[k] & W_{1,m+1}[k] \\ W_{2,m}[k] & W_{2,m+1}[k] \end{bmatrix} \tag{10}$$

where $Y_{l_x,m_x}[k]$ is the signal received at the destination within the l_x-th subslot of the m_x-th OFDM symbol of the k-th subcarrier, $X_{l_x,m_x}[k]$ is the signal transmitted from the source within the l_x-th subslot of the m_x-th OFDM symbol of the k-th subcarrier and $W_{l_x,m_x}[k]$ is the AWGN generated at the destination within the l_x-th subslot of the m_x-th OFDM symbol of the k-th subcarrier. $H_{1,1}[k]$ is defined as $H_{1,1}[k] = \sqrt{P_{\mathrm{S1}}L_{\mathrm{SD}}}H_{\mathrm{SD}}[k]$.

III. Proposed Decoding Schemes

In this section, we present two STC methods on cooperative OFDM-based wireless LAN systems, and propose a decoding scheme with approximate or exact ML property for each STC method. The subcarrier index k can be omitted in the following equations because we process frequency domain signals in the same fashion for each subcarrier.

A. For the Golden code

We propose a decoding scheme for the Golden code on cooperative OFDM-based wireless LAN systems. Applying the Golden code to cooperative communication has been presented in [4] and other papers, and we employ PAC SOMLD [9] for decoding it. PAC SOMLD involves an exhaustive search of all transmit signals, so with complexity of at least $O(|\mathcal{A}|^2)$ [7], [9], where $|\mathcal{A}|$ is the size of the underlying QAM constellation $\mathcal{A}$. PAC SOMLD inputs approximations of the log likelihood ratio (LLR) of coded bits to the soft-decision Viterbi algorithm as channel decoding, and its ML property is not exact but approximate [9].

B. For the Alamouti code

We propose a decoding scheme with exact ML property for the Alamouti code on cooperative OFDM-based wireless LAN systems. This decoding scheme has four processes, i.e., a conjugate operation for received $(m+1)$-th (even-number) OFDM symbols, a space-time decoding, a demapping operation for soft-decision coded bits and a subcarrier weighting, in order.

A transmission scheme of the Alamouti code for OFDM can be applied to each subcarrier of OFDM symbols [3]. Accordingly, each pair of OFDM symbols in a packet form a space-time coding block. Specifically, a block of space-time encoded transmit symbols for the data subcarrier $k \in \{0, \cdots, K-1\}$ in the m-th and $(m+1)$-th OFDM symbols can be expressed in matrix form. In the STC process for the Alamouti code, $[S_1, S_2]^T$ is set for the m-th OFDM symbol and $[-S_2^*, S_1^*]^T$ is set for the $(m+1)$-th OFDM symbol where $*$ is the complex conjugate operation. Then, (8) can be rewritten as follows

$$\begin{bmatrix} Y_{1,m} & Y_{1,m+1} \\ Y_{2,m} & Y_{2,m+1} \end{bmatrix} = \begin{bmatrix} H_{1,1} & 0 \\ H_{2,1} & H_{2,2} \end{bmatrix}\begin{bmatrix} S_1 & -S_2^* \\ S_2 & S_1^* \end{bmatrix} + \begin{bmatrix} W_{1,m} & W_{1,m+1} \\ W_{2,m} & W_{2,m+1} \end{bmatrix}. \tag{11}$$

Here, we present the proposed decoding scheme. We apply received $(m+1)$-th OFDM symbols, $Y_{1,m+1}$ and $Y_{2,m+1}$, to the conjugate operation in (11), and then we obtain the following equations:

$$\boldsymbol{y}_1 = \boldsymbol{H}_1\boldsymbol{s} + \boldsymbol{w}_1, \quad \boldsymbol{y}_2 = \boldsymbol{H}_2\boldsymbol{s} + \boldsymbol{w}_2 \tag{12}$$

where

$$\boldsymbol{y}_1 = \begin{bmatrix} Y_{1,m} \\ Y^*_{1,m+1} \end{bmatrix},\ \boldsymbol{y}_2 = \begin{bmatrix} Y_{2,m} \\ Y^*_{2,m+1} \end{bmatrix},\ \boldsymbol{s} = \begin{bmatrix} S_1 \\ S_2 \end{bmatrix},\ \boldsymbol{w}_1 = \begin{bmatrix} W_{1,m} \\ W^*_{1,m+1} \end{bmatrix},$$

$$\boldsymbol{w}_2 = \begin{bmatrix} W_{2,m} \\ W^*_{2,m+1} \end{bmatrix},\ \boldsymbol{H}_1 = \begin{bmatrix} H_{1,1} & 0 \\ 0 & -H^*_{1,1} \end{bmatrix},\ \boldsymbol{H}_2 = \begin{bmatrix} H_{2,1} & H_{2,2} \\ H^*_{2,2} & -H^*_{2,1} \end{bmatrix}. \quad (13)$$

The input-output relation between $\boldsymbol{s}$ and $\boldsymbol{y}_E$ by (12) is

$$\boldsymbol{y}_E = \boldsymbol{H}_E \boldsymbol{s} + \boldsymbol{w}_E \quad (14)$$

where

$$\boldsymbol{y}_E = \begin{bmatrix} \boldsymbol{y}_1 \\ \boldsymbol{y}_2 \end{bmatrix},\ \boldsymbol{H}_E = \begin{bmatrix} \boldsymbol{H}_1 \\ \boldsymbol{H}_2 \end{bmatrix},\ \boldsymbol{w}_E = \begin{bmatrix} \boldsymbol{w}_1 \\ \boldsymbol{w}_2 \end{bmatrix}. \quad (15)$$

Space-time decoding based on the MRC matrix form is used as follows:

$$\hat{\boldsymbol{s}} = \begin{bmatrix} \hat{S}_1 \\ \hat{S}_2 \end{bmatrix} = \frac{2(\boldsymbol{H}_1^H \boldsymbol{y}_1 + \boldsymbol{H}_2^H \boldsymbol{y}_2)}{\|\boldsymbol{H}_1\|_F^2 + \|\boldsymbol{H}_2\|_F^2} \quad (16)$$

where

$$\|\boldsymbol{H}_1\|_F^2 + \|\boldsymbol{H}_2\|_F^2 = 2(|H_{1,1}|^2 + |H_{2,1}|^2 + |H_{2,2}|^2) \quad (17)$$

and $\|\cdot\|_F$ is the Frobenius norm of a matrix [11]. This space-time decoding is a linear equalization. A demapping operation for soft-decision coded bits from $\hat{S}_1$ and $\hat{S}_2$ has complexity of $O(\log_2|\mathcal{A}|)$. From (14), the likelihood function of the signal vector corresponding to the received one $\boldsymbol{y}_E$, conditioned on a transmitted constellation point vector $\boldsymbol{s}$, is given by a multi-dimensional complex Gaussian distribution as follows,

$$p(\boldsymbol{y}_E|\boldsymbol{s}) = \frac{1}{(\pi\sigma_w^2)^4} \exp\left(\frac{-\|\boldsymbol{y}_E - \boldsymbol{H}_E \boldsymbol{s}\|^2}{\sigma_w^2} \right). \quad (18)$$

Then, the ML criterion for the receiver is

$$\hat{\boldsymbol{s}}_{\mathrm{ML}} = \arg\min_{\boldsymbol{s}} \{\Lambda\}, \quad \Lambda = \|\boldsymbol{y}_E - \boldsymbol{H}_E \boldsymbol{s}\|^2. \quad (19)$$

In terms of (19), we have

$$\Lambda = \frac{1}{2} \left(\|\boldsymbol{H}_1\|_F{}^2 + \|\boldsymbol{H}_2\|_F{}^2 \right) \left(|\hat{S}_1 - S_1|^2 + |\hat{S}_2 - S_2|^2 \right). \quad (20)$$

From (20), the received signals are space-time decoded as seen in (16) and then the terms $|\hat{S}_1 - S_1|$ and $|\hat{S}_2 - S_2|$, equivalent to the demapping operation, are given. The subcarrier weighting factor is given by the term $\|\boldsymbol{H}_1\|_F{}^2 + \|\boldsymbol{H}_2\|_F{}^2$. Soft-decision coded bits are multiplied by this factor and then input to the soft-decision Viterbi algorithm. Through the above processes, we can realize a decoding scheme with exact ML property [14].

IV. Simulation Results

This section evaluate the performance of two cooperative schemes, the Golden and Alamouti codes, and the non-cooperative scheme by simulations. The main parameters of the systems used in simulations are summarized in Table I. These parameters are based on the IEEE 802.11a standard [8]. For multipath fading models, we use a channel model presented in [12] and the root mean square delay spread is assumed to be 50 ns. Every path gain of the channel is assumed to follow an independent complex Gaussian distribution. The received E_b/N_0 is defined as $E_b/N_0 = (P_0 L_{\mathrm{SD}})/(M R_c \sigma_w^2)$ where P_0 is the reference transmit power, and M is the number of bits per constellation point such that $M = \log_2|\mathcal{A}|$ (e.g. $M = 2$ for QPSK), respectively. Non-cooperative communication has the transmit power of P_0 from the source. The Golden code scheme is set such that $P_{\mathrm{S1}} = P_0$ and $P_{\mathrm{R2}} = P_{\mathrm{S2}} = P_0/2$. The Alamouti code schemes in both Protocol I and III are set such that $P_{\mathrm{S1}} = P_0/2$ and $P_{\mathrm{R2}} = P_{\mathrm{S2}} = P_0/4$. These settings for transmit powers let the transmit energies per data bit for the non-cooperative and cooperative schemes be equal. We evaluate the performance in terms of packet error rate (PER) versus received E_b/N_0 and investigate the service range for three relay location scenarios. Fig. 3 shows the three relay location scenarios, where the relay is close to the source, at the middle point between the source and destination and close to the destination. Path loss and ratio of received energy to noise are specifically set as follows:

TABLE I
Main parameters

Parameter	Value
Bandwidth	20MHz
Number of DFT points (K)	64
Number of data subcarriers	48
Length of GI	16
Reference transmit power (P_0)	15 dBm
Noise power (σ_w^2)	-95 dBm

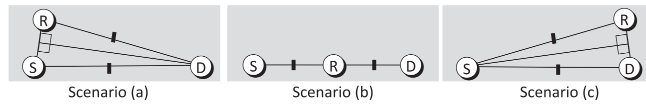

Fig. 3. Three relay location scenarios for performance evaluations.

- Scenario (a): Relay is close to the source

$$L_{\mathrm{SD}} = L_{\mathrm{RD}}, \qquad E_{\mathrm{SR}}/N_0 = \frac{L_{\mathrm{SR}} \cdot P_{\mathrm{S1}}}{\sigma_w^2} = 30\,(\mathrm{dB})$$

- Scenario (b): Relay is at the middle point between the source and destination

$$L_{\mathrm{SD}} = \frac{L_{\mathrm{SR}}}{2^\eta},\ L_{\mathrm{SR}} = L_{\mathrm{RD}}$$

- Scenario (c): Relay is close to the destination

$$L_{\mathrm{SD}} = L_{\mathrm{SR}}, \qquad E_{\mathrm{RD}}/N_0 = \frac{L_{\mathrm{RD}} \cdot P_{\mathrm{R2}}}{\sigma_w^2} = 30\,(\mathrm{dB})$$

where η is a path loss exponent. The path loss exponent η is assumed to be 3.5. The decoding depth of the Viterbi decoder is fixed to 35 bits. Interleavers defined in [8] are used. The payload size of each packet is fixed to $1,000$ bytes.

A. Packet Error Rate Performance Evaluation

We evaluate the performance of cooperative and non-cooperative communication systems in terms of PER. The modulation and coding scheme is QPSK and $Rc = 1/2$. Fig. 4 shows the PER performances obtained in the three scenarios.

From these figures, we can see that Scenario (b) has the best performance of the three scenarios in each cooperative transmission scheme. These results are consistent with those in [6] and [13]. We can also see that the performances of the Golden and Alamouti codes are significantly better than that of non-cooperative. At a PER of 10^{-1}, for example, the energy gains of the Alamouti code over non-cooperative are 2.1, 5.9 and 3.1 dB in Scenarios (a), (b) and (c), respectively. On the other hand, the energy difference between energies of the Golden and Alamouti codes is within 1.0 dB at a PER of 10^{-1} in each scenario. Therefore, we conclude that the Alamouti code with its less complex decoding scheme is better suited than the Golden code on a cooperative OFDM-based wireless LAN system.

B. Service Range Investigation

We also investigate ranges where systems maintain a PER of less than 10^{-1}. The value of 10^{-1} for PER is given in the definition of receiver minimum input level sensitivity of wireless LAN in [8]. We use the channel model D in [14] for path loss models. The modulation and coding scheme is binary phase shift keying (BPSK) and $Rc = 1/2$, which is one of the lowest transmission rates. Under this assumption, the non-cooperative case has a range of 90.0 m and the range of the Alamouti code is 97.1, 117, and 100 m in Scenarios (a), (b) and (c), respectively.

V. Conclusion

We proposed decoding schemes with approximate or exact ML property for STC transmission on cooperative OFDM-based wireless LAN systems. We employed the Golden code as an STC and proposed a decoding scheme utilizing PAC SOMLD for it, with approximate ML property. For the Alamouti code, we proposed a decoding scheme including linear space-time decoding based on MRC matrix form and sub-carrier weighting factors with exact ML property. Simulation results showed that the Golden code and the Alamouti code systems employing the proposed schemes provided energy gains of greater than 2.1, 5.9, and 3.1 dB for non-cooperative (direct) communication in three relay location scenarios at a PER of 10^{-1}.

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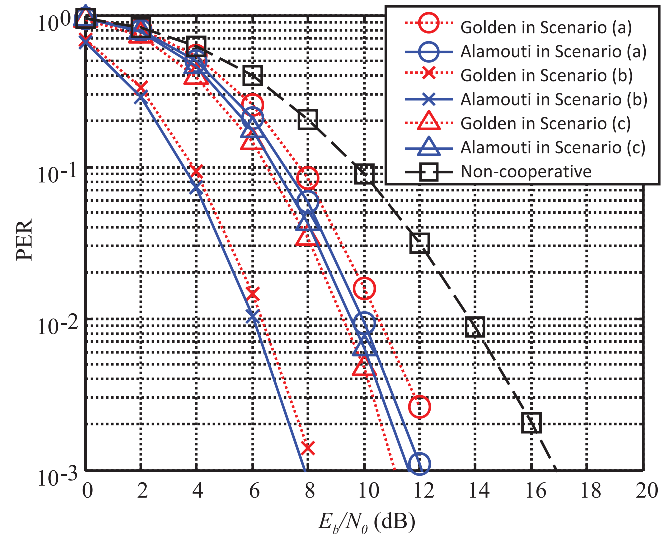

Fig. 4. PER performances of the cooperative Golden and Alamouti schemes in Scenario (a), (b) and (c), and the non-cooperative scheme.

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