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OFDM 신호를 이용한 비동기식 증폭 후 전달 중계망에서의 결합 채널 추정

(Joint Channel estimation in Asynchronous Amplify-And-Forward Relay Networks based on OFDM signaling)

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요 약

본 논문에서 증폭 후 전달 전송 기법을 사용하는 중계망의 채널 추정을 하는데 있어서 일어나는 문제점을 해결할 수 있는 방법으로 학습 계열(training sequence)을 이용하는 방법을 제안하였다. 현재의 고속 페이딩 채널 환경에서 기존 파일럿의 추정이 적절하지 않아 송신국(source)과 중계국(relay) 사이의 채널과 중계국(relay)과 수신국(destination) 사이의 채널을 결합하여 추정할 경우 많은 문제점이 초래되기에[1~2] 전송한 신호의 주파수 영역을 선택하여 얻은 정규(Gaussian) 분포에 대하여 최대 우도 함수의 평균을 내어 채널 추정량(estimator)을 유도해 낼 수 있는, 즉, 파일럿 대신에 하나의 OFDM 신호를 사용하여 모든 채널 충격 응답(CIR)을 추정할 수 있는 새로운 방법을 살펴보았다. 컴퓨터 모의실험으로 높은 SNR 영역에서 제안한 채널 추정기(estimator)의 성능이 [1]과 비교하여 약 1dB 정도 높음을 확인할 수 있었다.

Abstract

In this paper, we propose a method on the training sequence based on channel estimation issues for relay networks that employ amplify-and-forward (AF) transmission scheme. In [1] and [2], we have to point out that jointly estimating the channel from source to relay and from relay to destination suffers from many drawbacks in fast fading case because the estimation of previous pilots is not suitable for current channel. In this paper, we consider a new joint estimation of overall channel impulse response (CIR) using one OFDM signal without pilots. Using the maximum likelihood (ML) function, we derive a channel estimator by taking the frequency domain of transmitted signal as Gaussian and averaging the ML function over the resulting Gaussian distribution. Simulation results show that our proposed channel estimator performs a fraction of 1dB compared with [1] in high SNR region.

Keywords : Estimator, detector, OFDM, Gaussian, Relay.

I. INTRODUCTION

Rapid growth in wireless services places the demands on high speed and high throughput requirements. It is well known that the use of multiple input multiple output (MIMO) antenna systems improves the capacity and reliability of wireless communications. However, the use of multiple antennas to achieve transmit diversity in the cellular uplink is impractical due to size constraints at each mobile. In order to overcome this constraint problem, the relay network where the space diversity could be exploited from relay nodes existing in the network[4~5] was recommended to refer. These relay nodes are proposed to cooperate with terminals of other users. The latter scenario is sometimes referred to as "cooperative communication" when each relay

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nodes transmit its own information to destination by processing the information from source.

The relay based transmission is usually divided into two phase. During phase I, the source broadcast its own information bits to all relays. During phase II, the relays would either choose to purely amplify and retransmit the information to the destination, or to decode the information first and then transmit these information bits to the destination. The former process is referred as *amplify-and-forward*(AF) and latter it is referred as *decode-and-forward*(DF). Various cooperative diversity schemes and space time coding(STC) techniques have been developed in [4], [5] for either AF or DF approach.

In this paper, we propose a novel estimator based on an orthogonal frequency division multiplexing (OFDM) signal without pilots in fast fading case, where joint channel estimation from source to destination is estimated at destination. A training sequence will be send to all relay nodes during phase I, and the relay nodes forward the received signals after scaling the power constraint at each relay nodes, to the destination during phase II. Significantly, our estimators need just one OFDM symbol. Assuming that the transmitted frequency domain samples can be modeled as a complex Gaussian process and we derive the joint channel estimator by averaging the maximum likelihood(ML) function over the complex Gaussian distribution. An efficient detector is also proposed by using our currently estimated channel impulse response(CIR).

This paper is organized as follow. Section II describes the system model in consideration. The proposed scheme is described and analyzed in Section III. An efficient detector is also derived in Section IV. The performance is verified by computer simulations in Section V. Finally, Section VI concludes this paper.

Notations: Vectors and matrices are boldface small and capital letters; $\mathbf{A}_{ij}$ is the $\mathbf{A}(i,j)$th entry of $\mathbf{A}$ and $\mathrm{diag}\{\mathbf{a}\}$ denotes a diagonal matrix with the diagonal element constructed form $\mathbf{a}$; $\mathbf{I}$ is the identity matrix; $E\{\bullet\}$ denotes the statistical expectation. The discrete Fourier transform(DFT) matrix of size $N\times N$ is given by

$$\mathbf{F} = 1/\sqrt{N}\exp(-2\pi jkl/N), \quad k,l \in 0,1,\cdots,N-1.$$

$\mathbf{A}$ denotes a diagonal matrix whose diagonal entries are power factor for each relay nodes.

II. SYSTEM MODEL

1. OFDM Baseband Model

In OFDM system, the source data are mapped into the symbols from constellation alphabet $\mathcal{Q}$ (We consider QPSK modulation in this paper), which are modulated by inverse DFT(IDFT) on T parallel subcarriers. Then output samples can be written as $\mathbf{X} = \mathbf{x}\cdot\mathbf{F}$, where $\mathbf{x}$ is an OFDM input vector selected from $\mathcal{Q}^N$. We assume that the transmission delay in the path between source to destination through i^{th} relay node is τ_i. Some subcarriers may be inserted to the left of one OFDM signal $\mathbf{X}$. After the cyclic prefix(CP) insertion, the OFDM symbol is transmitted to the destination node. We also assume that the timing error are known at the receiver, and are within a reasonable range, i.e., all the time delay are less than or equal to a maximum time delay $\tau_{\max}$.

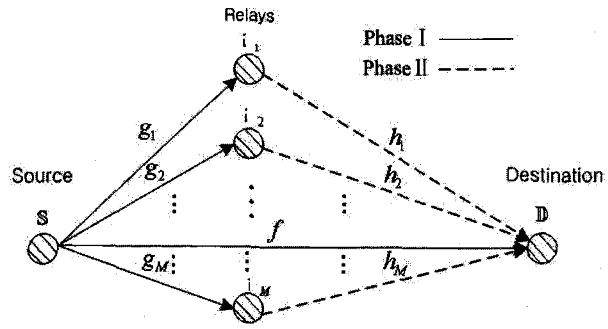

그림 1. 시스템 모델
Fig. 1. system model.

2. Transmission Model

Consider a wireless network with M randomly placed relay nodes $\mathbb{R}_i$, where $i = 1,\cdots,M$, one source node $\mathbb{S}$ and one destination node $\mathbb{D}$ as shown in Fig. 1. Every node has only a single antenna that cannot transmit and receive simultaneously. The channel between each node pair is assumed Rayleigh fast fading which is not constant within one frame and may vary from frame to frame. Denote the channel from $\mathbb{S}$ to $\mathbb{R}_i$ as g_i, from $\mathbb{R}_i$ to $\mathbb{D}$ as h_i, respectively; where $g_i \in \mathcal{CN}(0,\sigma^2_{g_i})$, and $h \in \mathcal{CN}(0,\sigma^2_{h_i})$. Through this paper, we assume non-synchronization at all terminals as in [7]. The source node sends X converted from an information symbol vector $\mathbf{x}$ to the relay nodes in T time slots. The information symbols s_i, where $i \in \{1,2,\cdots,T\}$ are from signal constellation $\mathcal{Q}$. At the i^{th} relay node $\mathbf{r}_i$ is received.

$$\mathbf{r}_i = g_i\mathbf{s} + \mathbf{n}_i, \qquad (1)$$

where $\mathbf{n}_i$ is the independent white complex Gaussian noise at the relays. For convenience all noise variances are assumed as N_0, namely, $\mathbf{n}_i \in \mathcal{CN}(0,N_0\mathbf{I})$. The power constraint of the transmission is $E\{\mathbf{x}^H\mathbf{x}\} = TP_s$, where P_s is the average transmitting power of the source. Then $\mathbf{r}_i$ is scaled by real factor a_i to keep the average power of $\mathbb{R}_i$ as P_{r_i}.

$$\mathbf{t}_i = a_i\mathbf{r}_i, \qquad (2)$$

Basically, there are two different choice of a_i, which are listed as follows:

$$a_i = \sqrt{\frac{P_n}{|g_i|^2 P_s + N_0}}\text{, or} \qquad (3)$$

$$a_i = \sqrt{\frac{P_n}{\sigma^2_{g_i} P_s + N_0}}. \qquad (4)$$

Here, we recommend the second choice of a_i which is not a random value while keeps the power constraint from long term point of view. Otherwise, g_i could be replaced by its estimated value which may not keep the average power exactly as P_{r_i}. In this paper, we always consider (4) for our work. The destination $\mathbb{D}$ in phase II receives,

$$\mathbf{d}_2 = \sum_{i=1}^{M} h_i\mathbf{t}_i + \mathbf{n}_{d_2} = \mathbf{ABW} + \mathbf{n}_d, \qquad (5)$$

where $\mathbf{n}_{d_2} \in \mathcal{CN}(0,N_0\mathbf{I})$ represents the complex white Gaussian noise vector at $\mathbb{D}$ in the second phase, and $\mathbf{X}_i \triangleq \mathbf{X}$ for all $i \in \{1,2,\cdots,M\}$. Other variables are showed as follows:

$$\mathbf{w} = [w_1,\cdots,w_M]^T,\ w_i = g_i h_i^*,\ i \in \{1,\cdots,M\},$$
$$\mathbf{A} = diag\{a_1,\cdots,a_M\}\ \mathbf{B} = [\mathbf{s}_1,\mathbf{s}_i,\cdots,\mathbf{s}_M],$$
$$\mathbf{n}_d = \sum_{i=1}^{M} h_i a_i \mathbf{n}^*_{r_i} + \mathbf{n}^*_{d_2}. \qquad (6)$$

Assume that the statistics of g_i and h_i are known at $\mathbb{D}$ is namely, the statistical channel scenario. Denote the covariance of $\mathbf{h}$ and $\mathbf{g}$ as $\mathbf{R_h}$ and $\mathbf{R_g}$ respectively. The covariance matrix of $\mathbf{W}$ is assuming channels of phase I and phase II are independent, so

$$\mathbf{R_W} = \mathbf{R_h} \odot \mathbf{R_g}, \qquad (7)$$

where $\odot$ denotes the Hadamard product. Furthermore, the covariance of $\mathbf{n}_d$ conditional on specific realization of h_i can be calculated as:

$$Cov(\mathbf{n}_d) = (\sum_{i=1}^{M} |h_i|^2 a_i^2 + 1)N_0\mathbf{I}. \qquad (8)$$

Furthermore, the overall noise under a specific realization of h_i is still white Gaussian with a scaled covariance of $N_0\mathbf{I}$. Assume that the length of OFDM signal is large enough and the number of relay nodes is large, we can simplify (8) into simple form $Cov(\mathbf{n}_d) = (\sum_{i=1}^{M} |h_i|^2 a_i^2 + 1)N_0\mathbf{I}.$

III. JOINT CHANNEL ESTIMATION

At the destination node, after the CP removal and

N-point FFT, the received signal can be written as

$$\mathbf{Y} = \mathbf{AFBW} + \mathbf{n} , \quad (9)$$

where $\mathrm{Y} = \mathrm{Fd_2}$ is the output of DFT transform at the receiver and $\mathrm{n} = \mathrm{Fn}_d$ is also a Gaussian vector in [8] with covariance matrix $Cov(\mathrm{n}) = \left(\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1\right)N_o\mathrm{I}$.

Then, vector Y is a joint Gaussian vector with mean AFBW and covariance matrix $\left(\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1\right)N_0\mathrm{I}$. The likelihood function of vectors s and W is given by

$$P(\mathrm{Y|B,W}) = P_{\mathrm{n}}(\mathrm{Y} - \mathrm{AFBW}) = \frac{1}{\left(\pi\left(\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1\right)N_0\right)^N}\exp\left\{-\frac{\|\mathrm{Y} - \mathrm{AFBW}\|^2}{\left(\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1\right)N_0}\right\} . \quad (10)$$

In [3], the OFDM signal can be modeled as a Gaussian Process while its length is large via CLT theorem. To simplify (10), denote that σ^2 is equivalent to (10). Denote $\mathbf{S} \in \mathcal{CN}(0, \mathbf{\Lambda}_D)$ and we can consider in this paper $E_s\mathrm{I} = \Lambda_D$. In order to propose a joint estimator of W in fast fading case, the transmitted signal S would be used to estimate W depending on its statistical properties. The vector S denotes an unwanted vector parameter in (10) to propose an estimator. To isolate W from (10) as a function with only one unknown parameter W, the averaging $P(\mathrm{Y|S,W})$ over the PDF of S gives the marginal likelihood function $P(\mathrm{Y|W})$

$$P(\mathrm{Y|W}) = \int P(\mathrm{Y|B,W})\mathrm{P}(\mathrm{s})ds , \quad (11)$$

where $P(\mathrm{S})$ is the PDF of S with PDF

$$P(\mathbf{s}) = 1/(2\pi)^{N/2}\det{}^{1/2}(\mathbf{R}_s)\cdot\exp\left(-\frac{1}{2}\mathbf{S}^T\mathbf{R}_s^{-1}\mathbf{S}\right) , \quad (12)$$

is given in [9]. After some manipulation of (12), we can show that the marginal likelihood function (11) is calculated as

$$P(\mathbf{Y}\mid\mathbf{W}) = \frac{1}{\pi^N\det(\sigma^2 + \bar{\mathbf{S}}\mathbf{W}^H\mathbf{W})}\times \exp\Big\{-\Big[\tilde{\mathbf{S}}^H\left(\bar{\mathbf{S}} + (\mathbf{W}^H\mathbf{W})^{-1}\sigma^2\right)^{-1}\tilde{\mathbf{S}} + \mathbf{Y}^H\left(\bar{\mathbf{S}}\mathbf{W}^H\mathbf{W} + \sigma^2\mathbf{I}_N\right)^{-1}\mathbf{Y} - 2\,\mathrm{Re}\left(\mathbf{Y}^H\mathbf{W}(\bar{\mathbf{S}}\mathbf{W}^H\mathbf{W} + \sigma^2\mathbf{I}_N)^{-1}\tilde{\mathbf{S}}\right)\Big]\Big\} . \quad (13)$$

The determinant of $\det(\sigma^2 + \mathrm{W}^H\mathrm{W})$ is $\prod_{k=0}^{N-1}\left(\sigma^2 + |w_k|^2\right)$. After some manipulations of (13), the simple likelihood function can be obtained by using logarithm of (13) and deleting some constant terms. The proposed joint estimator is similar to the log-likelihood function

$$g(\mathbf{W}) \triangleq \ln f(\mathbf{Y}\mid\mathbf{W}) = \sum_{k=N_P}^{N-1}\left(\frac{|Y_k|^2}{E_s|W_k|^2 + \sigma^2} + \ln(E_s|W_k|^2 + \sigma^2)\right) + \sum_{k=0}^{N_P-1}|Y_k - W_k S_k|^2 . \quad (14)$$

This problem can be reformulated as

$$\min g(W) \quad \text{subject to } \mathbf{W} \in C^N . \quad (15)$$

If the channel is fast fading that the length of channel response is larger than N, and when we extend the OFDM signal size or add more OFDM signals to estimate the joint channel coefficients, then the extended channel estimator can expressed as

$$\min \sum_{k=0}^{k=K} g(\mathbf{W}_k) \quad \text{subject to } \mathbf{W}_{KN} \in C^{KN} . \quad (16)$$

The optimum value of $g(\mathrm{W})$ can be evaluated via the gradient descent and related algorithms, i.e., Newton method or Interior point methods. To do these iterative algorithms, we need an initial channel estimate and the initial value of $\widehat{\mathrm{W}}$ is expressed via well-known minimum mean-square error (MMSE) estimator [10] as

$$\hat{\mathbf{W}} = E[\mathbf{W}\mathbf{Y}^H]E^{-1}[\mathbf{Y}\mathbf{Y}^H]\mathbf{Y} = \left(\boldsymbol{\Lambda}_p\mathbf{F}^H\mathbf{F} + \sigma^2\mathbf{R}_{\mathbf{W}}^{-1}\right)^{-1}\mathbf{F}^H\boldsymbol{\Lambda}_D\mathbf{Y}_P, \quad (17)$$

where $\mathbf{F}$ is a $N \times N$ Fourier Matrix and R_W is the covariance matrix of W. Furthermore, R_W and σ^2 are required for this estimator. In [1], assume that two components of $w_i = g_i h_i^*$ are independent and its individual statistics is known in [1] and [2]. We would not discuss these statistics in detail and please refer [1] and [2]. Due to its statistical property, R_W can be calculated by $R_W = R_h \odot R_g$ and the initial estimates for iteratively computing (17).

Ⅳ. AN EFFICIENT DETECTOR FOR OUR SCHEME

Here, (9) can be rewritten explicitly in terms of transmitted symbol in time domain and Gaussian noise

$$\mathbf{Y} = \mathbf{ASFW} + \mathbf{n}, \quad (18)$$

where $\mathbf{s} = [\mathbf{x}_1, \mathbf{x}_i, \ldots, \mathbf{x}_M], \mathbf{x}_i \triangleq \mathbf{x}, i \in \{1, 2, \ldots M\}$ and $\mathbf{x} = diag\{x_i\}$ with entries selected from an alphabet Q with uniform distribution. This form is different from (9) due to that signal in (9) is Gaussian distributed. To derive an efficient detector, it requires knowledge of the autocorrelation matrix R_W of the CIR W and the noise σ^2, where W and noise with zero-mean complex Gaussian are independent. The received vector Y is assumed to be modeled as i.i.d zero mean Gaussian Process conditional on the data vector, data vector and noise are independent random processes. The PDF function of Y can be written as [9]

$$P(\mathbf{Y}) = 1/(2\pi)^{n/2}\det{}^{1/2}(\mathbf{R}_{\mathrm{Y}}) \cdot \exp(-\mathbf{Y}^H\mathbf{R}_{\mathrm{Y}}^{-1}\mathbf{Y}), \quad (19)$$

where Y is Gaussian distributed with n degrees of freedom. The autocorrelation matrix of the received vector is given by

$$\mathbf{R}_{\mathrm{Y}} = E[\mathbf{Y}\mathbf{Y}^H] = \mathbf{S}_d\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{W}}\mathbf{a}^H\mathbf{F}^H\mathbf{S}_d^H + \sigma^2\mathbf{I} = \mathbf{S}_d\left\{\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{w}}\mathbf{a}^H\mathbf{F}^H + (\sum_{i=1}^{M}\sigma_{hi}^2 a_i^2 + 1)N_0\mathbf{I}\right\}\mathbf{S}_d^H \quad (20)$$

From property of matrix computation [11], the determinant of R_Y can be expressed as

$$\det(\mathbf{R}_{\mathrm{Y}}) = \det(\mathbf{S}_d)\det(\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{W}}\mathbf{a}^H\mathbf{F}^H + (\sum_{i=1}^{M}\sigma_{hi}^2 a_i^2 + 1)N_0\mathbf{I})\det(\mathbf{S}_d^H) = \det(\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{W}}\mathbf{a}^H\mathbf{F}^H + (\sum_{i=1}^{M}\sigma_{hi}^2 a_i^2 + 1)N_0\mathbf{I}) \quad (21)$$

In our transmission, PSK signaling is always considered in source node, and the determinant of $\det(\mathbf{S}_d)$ is an identity matrix. The log-likelihood function is given by

$$\Lambda(\mathbf{Y} \mid \mathbf{S}, \mathbf{W}) = -\mathbf{Y}^H\mathbf{R}_{\mathrm{Y}}^{-1}\mathbf{Y}. \quad (22)$$

Due to transmitted symbol S by employing unitary signaling (PSK) can be recognized as a unitary matrix, so we can multiply $\mathbf{S}^H$ and S on the both sides of left side of (20),

$$-\mathbf{Y}^H\mathbf{R}_{\mathrm{Y}}^{-1}\mathbf{Y} = \frac{\mathbf{S}_d^H\mathbf{Y}^H\mathbf{Y}\mathbf{S}_d}{\mathbf{S}_d^H\mathbf{S}_d\left\{\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{w}}\mathbf{a}^H\mathbf{F}^H + (\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1)N_0\mathbf{I}\right\}\mathbf{S}_d} = \frac{\mathbf{S}_d^H\mathbf{Y}^H\mathbf{Y}\mathbf{S}_d}{\left\{\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{w}}\mathbf{a}^H\mathbf{F}^H + (\sum_{i=1}^{M}\sigma_{h_i}^2 a_i^2 + 1)N_0\mathbf{I}\right\}} \quad (23)$$

Then, maximizing the log-likelihood function is equivalent to solving

$$\hat{\mathbf{S}}_d = \arg\min_{\mathbf{S} \in Q^N} \mathbf{S}_d^H\mathbf{Y}^H(\mathbf{F}\mathbf{a}\mathbf{R}_{\mathrm{w}}\mathbf{a}^H\mathbf{F}^H + \sigma^2\mathbf{I})^{-1}\mathbf{Y}\mathbf{S}_d, \quad (24)$$

From our estimation of $\hat{\mathbf{W}}$ in (15), we derive the detector by replacing the term R_W in (23) with the autocorrelation matrix $\hat{\mathbf{R}}_{\mathrm{W}}(m) = 1/N\sum_0^{N-1}\hat{W}(n+m)\hat{W}(n)$ of our estimated $\hat{\mathbf{W}}$ vector. Therefore, our proposed detector in relay networks is given by

$$\hat{\mathbf{S}}_d = \arg\min_{\mathbf{S}\in\mathcal{Q}^N} \mathbf{Y}^H \mathbf{S}_d^H (\mathbf{F}\mathbf{a}\hat{\mathbf{R}}_{\mathbf{w}}\mathbf{a}^H\mathbf{F}^H + \sigma^2\mathbf{I})^{-1}\mathbf{S}_d\mathbf{Y}$$
$$= \arg\min_{\mathbf{S}\in\mathcal{Q}^N} \mathbf{S}^H\mathbf{Y}^H (\mathbf{F}\mathbf{a}\hat{\mathbf{R}}_{\mathbf{w}}\mathbf{a}^H\mathbf{F}^H + \sigma^2\mathbf{I})^{-1}\mathbf{Y}\mathbf{S} \quad (25)$$

Straight forwardly, (25) can be solved by searching all space of $\mathcal{Q}^N$, which costs much complexity as N is large. Here, (25) can be reformulated as a quadratic form and then we can use the fast sphere decoding (SD) algorithm [12] in our proposed detector.

V. NUMERICAL RESULTS

In this section, we numerically show the performance of our proposed channel estimator based on OFDM signal as well as an efficient detector under various scenarios. The effect of individual channels g_i and h_i alone should be considered at the relay and the receiver. The channel covariance $\mathbf{R}_\mathbf{h}$ and $\mathbf{R}_\mathbf{g}$ have the following structures [13]: $[\mathbf{R}_\mathbf{h}]_{a,b} = \varepsilon_1^{|a-b|}$ and $[\mathbf{R}_g]_{a,b} = \varepsilon_2^{|a-b|}$, where ε_1 and ε_2 are two real scalars. The carrier frequency is 5GHz and the data rate is 1000kbps. Eight pilot symbols are inserted into the tail of each OFDM symbol to combat the delay. The delay $\tau_i \le \tau_{\max}$ at each relay is chosen randomly from 0 to 8 with uniform distribution.

Fig. 2 shows the MMSE versus SNR of both the LS estimator [1] and our proposed estimator (15) for M=4 relay nodes with power scalar at each relay as $\{0.8, 0.9, 0.9, 0.8\}$. For convenience, we set $\varepsilon_1 = \varepsilon_2 = \varepsilon = 0.1$ in all simulations. The performance of LS is almost the same as our proposed estimator with SD algorithm at $MSE = 10^{-2}$. At the high SNR region our proposed estimator performs 2-3dB better than the LS. The Cramer-Rao bound (CRB) of

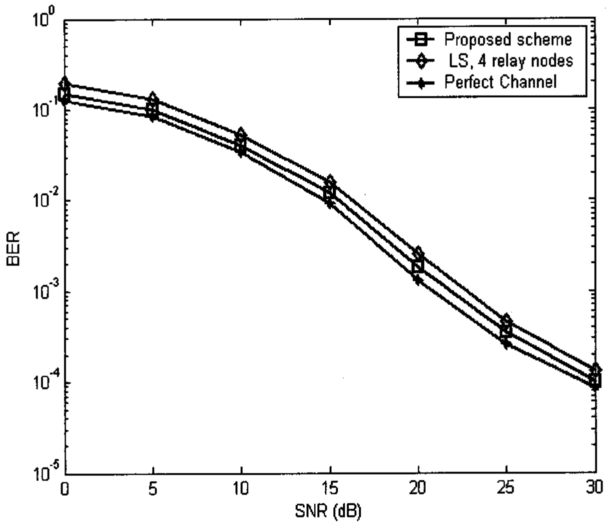

그림 3. 채널 추정 기법에 따른 BER 성능 비교
Fig. 3. Comparisons of BER performance with different estimated channels.

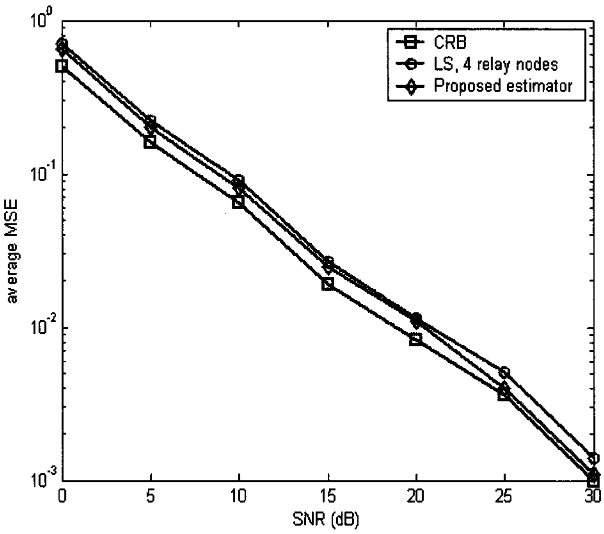

그림 2. Estimator에 따른 MSE의 비교
Fig. 2. Comparison of MSEs.

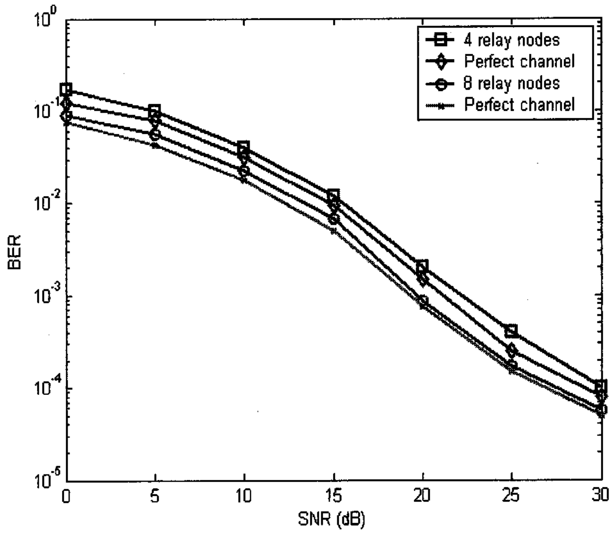

그림 4. 중계국(relay) 수에 따른 BER 성능 비교
Fig. 4. Effects of BER for different relays.

the CIR is also shown in this figure. In the high SNR region (20-30dB), the proposed estimator achieves CRB.

The BER performance is compared in Fig. 3 with different CIRs including CIR estimated by LS, CIR estimated by our proposed estimator, and perfect CIR is known at destination. Substituting the result of (15) into (25) detects the transmitted symbols. Both results are close to the BER performance with perfect CIR. To compare these at different SNR regions, the BER with LS estimator has 1.5dB loss and (25) with proposed estimator and SD algorithm has 0.7dB loss at $BER=10^{-3}$. Even in low SNR region, there exists 1.5dB gap. Our proposed detector and estimator are suitable for low SNR region.

In Fig. 4, we also compare the effects of different relay nodes with 4 relay nodes and 8 relay nodes considering delay case for our proposed estimator and detector. The performance with 4 relay nodes has 3dB loss compared to perfect CIR, and almost 2dB loss compare to 8 relay nodes. The reason to explain this phenomenon is that the relay number affects the channel estimation even when the same power of source and same average power of relays are applied.

Ⅵ. CONCLUSIONS

In this paper, we propose an efficient channel estimator and symbol detector based on an OFDM symbol in asynchronous amplify-and-forward (AF) relay networks. These estimator and detector consider an OFDM symbol in frequency domain (Gaussian distribution) and time domain (uniform distribution), by averaging conditional PDF over its distribution and by maximizing ML function on received symbols using our estimated channel information. To do fast detect transmitted symbols, the SD algorithm is also considered in our work. Simulation results show that our proposed estimator on CRI has better MMSE performance in high SNR region and it is near to the Cramer-Rao bound with a 1.5dB gap. Also the detector shows BER performance similar with perfect channel information. These estimator and detector can be also extended into other complicated communication system based on OFDM signaling.

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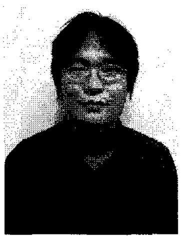

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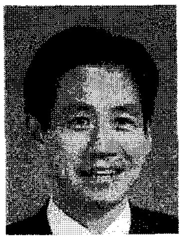

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