Single-Carrier Cooperative DF Relay Using Adaptive Modulation

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***Abstract*— A throughput performance significantly degrades near the cell edge due to the path loss, shadowing loss, and multipath-fading. A cooperative relay is a well-known technique to increase the cell-edge throughput. In the cooperative relay, it is expected that assigning different levels of data modulation for mobile terminal (MT) and relay station (RS) may improve the throughput performance. In this paper, we propose a 2-time slot single-carrier (SC) cooperative decode-and-forward (DF) relay using adaptive modulation and evaluate, by computer simulation, its throughput performance. It is shown that the cooperative DF relay using adaptive modulation can reduce the normalized total transmit signal-to-noise power ratio (SNR) by about 1.5 dB compared to the cooperative DF relay using fixed modulation for a required throughput of 0.5 bps/Hz and can reduce the normalized total transmit SNR by about 9 dB compared to the direct transmission for a required throughput of 0.8 bps/Hz.**

Keywords; cooperative decode-and-forward (DF) relay, single-carrier transmission, adaptive modulation

I. INTRODUCTION

In the next generation mobile communication systems, broadband data services of around 1 Gbps are demanded. However, the throughput performance significantly degrades near the cell edge due to the propagation path loss, shadowing loss, and multipath-fading [1]. Therefore, the transmit power must be significantly increased to achieve the required throughput. The cooperative relay [2-5] is a well-known technique to increase the cell-edge throughput. The throughput achievable by the cooperative relay is determined by the worth link between the mobile terminal (MT) - relay station (RS) link and the RS - base station (BS) link [4]. Using different levels of data modulation at MT and RS may improve the throughput performance; lower level of modulation is used for a link with worse condition.

In this paper, we propose a single-carrier (SC) 2-time slot cooperative decode-and-forward (DF) relay using adaptive modulation. The modulation level is determined based on the signal-to-noise power ratio (SNR). According to the modulation levels of the first time-slot and the second time-slot, the ratio of time-slot lengths is also changed. We evaluate, by computer simulation, the throughput of 2-time slot cooperative DF relay using adaptive modulation.

The rest of this paper is organized as follows. Section II presents the 2-time slot cooperative DF relay network model. Sect. III presents the transmit/receive signal representation and signal combining. Computer simulation results are discussed in Sect. IV. Section VI concludes this paper.

II. COOPERATIVE DF RELAY NETWORK USING ADAPTIVE MODULATION

Figure 1 illustrates the 2-time slot cooperative DF relay network model considered in this paper. The single-cell and single-user environment is assumed. The communication area is within the radius R from the BS. It is assumed that 6 RSs are located equal distantly to each other at the distance $R/2$ from BS. In this paper, RS which has the best link is chosen for the cooperative relay and is denoted by RS_i (i=0~5). The distances between MT and BS, between MT and RS_i, and between RS_i and BS are denoted by R_{MB}, R_{Mi}, and R_{iB}, respectively. The total transmit power P is assumed to be equally allocated to MT and RS_i as $P_M = P_i = P/2$, where P_M and P_i are the transmit power at MT and RS_i, respectively.

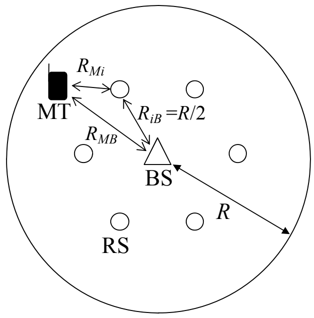

Fig. 1 System model.

Figure 2 shows the uplink transmission in 2-time slot cooperative DF relay. In this paper, it is assumed that the data symbol duration T_s is the same for each time slot.

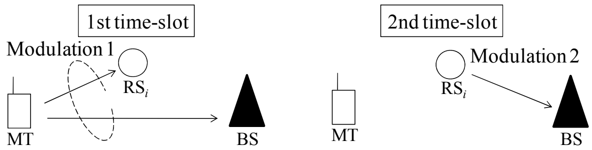

Fig. 2 Uplink transmission in 2-time slot cooperative DF relay.

During the first time slot, at the MT, the data-modulated symbol sequence is divided into a sequence of N_c-symbol blocks. The last N_g symbols of each N_c-symbol block are inserted into the guard interval (GI) as a cyclic prefix (CP). Then, the MT broadcasts the sequence of CP-inserted symbol blocks to both RS and BS. RS applies the minimum mean square error frequency-domain equalization (MMSE-FDE) [6] to each CP-removed N_c-symbol block and carries out the block data detection.

During the second time slot, RS remodulates the recovered data block and transmits to BS after CP insertion.

BS combines the received signal blocks from MT in the first time slot and RS in the second time slot. The signal combining scheme depends on whether the data modulation levels in the first and second time-slots are the same or not.

In this paper, it is assumed that BS has the perfect knowledge of channel state information (CSI) of the MT-BS link, MT-RS$_i$ link and RS$_i$-BS link. BS selects the modulation combination which maximizes the throughput and informs the selected modulation combination to MT and RS.

The throughput S is defined as

$$S=\frac{M}{T_s N(M/m_1+M/m_2)\left(1+N_g/N_c\right)}\quad \text{[bps]}\ , \qquad (1)$$

where the denominator of Eq. (1) represents the required time for M bits transmission. M is the packet size. m_1 and m_2 are the modulation levels for the first and second time-slots, respectively. N is the average number of the packet retransmissions. Since N=1–PER, where PER denotes the packet error rate achieved by the cooperative DF relay, Eq. (1) can be rewritten as

$$S=\frac{m_1 m_2}{(m_1+m_2)\left(1+N_g/N_c\right)}\left(1-PER\right)\quad \text{[bps/Hz]}\ . \qquad (2)$$

The optimum modulation combination which maximizes S is chosen. The set of possible modulation combinations is shown in Table 1. An example of modulation combination is illustrated by Fig. 3. The frame length is always kept constant. The amount of transmitted data bits per frame changes according to the modulation combination.

Table 1 Set of possible modulation combinations.

BPSK-BPSK, BPSK-QPSK, BPSK-16QAM, QPSK-BPSK, QPSK-QPSK, QPSK-16QAM, 16QAM-BPSK, 16QAM-QPSK, 16QAM-16QAM

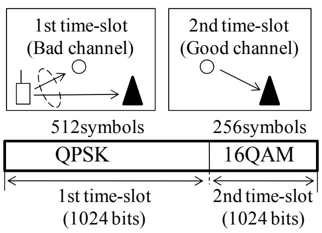

Fig.3 An example of modulation combination.

III. SIGNAL COMBINING

The received powers, P_{MB}, P_{Mi}, and P_{iB}, of MT-BS, MT-RS$_i$, and RS$_i$-BS links can be expressed as

$$\begin{cases} P_{MB}=P_M\cdot R_{MB}^{-\alpha}\cdot 10^{-\eta_{MB}/10}=\left(P_M\cdot R^{-\alpha}\right)\cdot\left(R_{MB}^{-\alpha}/R^{-\alpha}\right)\cdot 10^{-\eta_{MB}/10} \\ P_{Mi}=P_M\cdot R_{Mi}^{-\alpha}\cdot 10^{-\eta_{Mi}/10}=\left(P_M\cdot R^{-\alpha}\right)\cdot\left(R_{Mi}^{-\alpha}/R^{-\alpha}\right)\cdot 10^{-\eta_{Mi}/10} \\ P_{iB}=P_i\cdot R_{iB}^{-\alpha}\cdot 10^{-\eta_{RB}/10}=\left(P_i\cdot R^{-\alpha}\right)\cdot\left(R_{iB}^{-\alpha}/R^{-\alpha}\right)\cdot 10^{-\eta_{iB}/10} \end{cases}, \qquad (3)$$

where α is the path loss exponent. η_{MB}, η_{Mi}, and η_{iB} are the log-normally distributed shadowing loss in dB of MT-BS link, MT-RS$_i$ link, and RS$_i$-BS link, respectively. They are independent zero mean Gaussian random variables with standard deviation σ. By introducing the normalized distances r_{MB}=R_{MB}/R, r_{Mi}=R_{Mi}/R and r_{iB}=R_{iB}/R of MT-BS, MT-RS$_i$, RS$_i$-BS links, Eq. (3) can be rewritten as

$$\begin{cases} P_{MB}=\overline{P}_M\cdot r_{MB}^{-\alpha}\cdot 10^{-\eta_{MB}/10} \\ P_{Mi}=\overline{P}_M\cdot r_{Mi}^{-\alpha}\cdot 10^{-\eta_{Mi}/10} \\ P_{iB}=\overline{P}_i\cdot r_{iB}^{-\alpha}\cdot 10^{-\eta_{iB}/10} \end{cases}, \qquad (4)$$

where $\overline{P}_M=P_M\cdot R^{-\alpha}$ and $\overline{P}_i=P_i\cdot R^{-\alpha}$ are the normalized transmit powers at MT and RS$_i$, respectively.

A. Signal processing at relay

In this paper, the equivalent baseband T_s-spaced discrete time signal representation is used.

The CP-removed received signal blocks at RS$_i$ and BS in the first time slot are respectively denoted by $\{y_{Mi}(t);t=0,\ldots,N_c-1\}$ and $\{y_{MB}(t);t=0,\ldots,N_c-1\}$. $y_{Mi}(t)$ and $y_{MB}(t)$ are given as

$$\begin{cases} y_{Mi}(t)=\sqrt{2P_{Mi}}\sum_{l=0}^{L-1}h_{l,Mi}s\left((t-\tau_{l,Mi})\bmod N_c\right)+n_{Mi}(t) \\ y_{MB}(t)=\sqrt{2P_{MB}}\sum_{l=0}^{L-1}h_{l,MB}s\left((t-\tau_{l,MB})\bmod N_c\right)+n_{MB}(t) \end{cases}, \qquad (5)$$

where $\{s(t);t=0,\ldots,N_c-1\}$ is the transmitted symbol block from MT. $h_{l,Mi}$ and $h_{l,MB}$ are complex path gains and $\tau_{l,Mi}$ and $\tau_{l,MB}$ are the time delays of the l-th path of the links between MT and RS$_i$ and between MT and BS, respectively. $n_{Mi}(t)$ and $n_{MB}(t)$ are zero-mean complex-valued noise samples with variance $2N_0/T_s$, where N_0 is the single-sided power spectrum density of the additive white Gaussian noises (AWGNs).

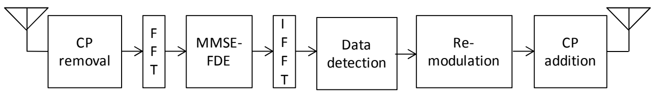

Fig. 4 RS structure.

Figure 4 illustrates the RS structure. At the RS, N_c-point fast Fourier transform (FFT) is applied to the CP-removed received signal block $\{y_{Mi}(t);t=0,\ldots,N_c-1\}$ to transform it into the frequency-domain received signal $\{Y_{Mi}(k);k=0,\ldots,N_c-1\}$. $Y_{Mi}(k)$ is given as

$$\begin{aligned} Y_{Mi}(k)&=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}y_{Mi}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &=H_{Mi}(k)S(k)+\Pi_{Mi}(k) \end{aligned}, \qquad (6)$$

where

$$\begin{cases} S(k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}s(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ H_{Mi}(k)=\sqrt{2P_{Mi}}\sum_{l=0}^{L-1}h_{l,Mi}\exp\left(-j2\pi k\frac{\tau_{l,Mi}}{N_c}\right) \\ \Pi_{Mi}(k)=\frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}n_{Mi}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \end{cases}. \qquad (7)$$

One-tap MMSE-FDE is applied as

$$\hat{Y}_{Mi}(k)=Y_{Mi}(k)W_{Mi}(k)\ , \qquad (8)$$

where $W_{Mi}(k)$ is the MMSE-FDE weight (which minimizes the mean square error (MSE) between $\hat{Y}_{Mi}(k)$ and $S(k)$) given as [7]

$$W_{Mi}(k) = \frac{H^*_{Mi}(k)}{\left|H_{Mi}(k)\right|^2 + 2N_0/T_s} \ , \tag{9}$$

where $(.)^*$ denotes the complex conjugate operation.

The frequency-domain signal $\{\hat{Y}_{Mi}(k); k = 0,\ldots,N_c-1\}$ after MMSE-FDE is transformed by N_c-point inverse FFT (IFFT) back to the time-domain signal block $\{\hat{d}_{Mi}(t); t = 0,\ldots,N_c-1\}$. $\hat{d}_{Mi}(t)$ is given as

$$\begin{aligned}\hat{d}_{Mi}(t) &= \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\hat{Y}_{Mi}(k)\exp\left(j2\pi t\frac{k}{N_c}\right) \\ &= \left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{Mi}(k)W_{Mi}(k)\right)s(t) \\ &\quad + \frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{Mi}(k)W_{Mi}(k)\sum_{\substack{\tau=0\\ \neq t}}^{N_c-1}s(\tau)\exp\left(j2\pi k\frac{t-\tau}{N_c}\right) \\ &\quad + \frac{1}{N_c}\sum_{k=0}^{N_c-1}W_{Mi}(k)\Pi_{Mi}(k)\exp\left(j2\pi k\frac{t}{N_c}\right)\end{aligned} \ , \tag{10}$$

where the first term is the desired signal, the second and third terms are the residual inter-symbol interference (ISI) and the noise, respectively.

Finally, the data decision and remodulation is done by the RS to form the symbol block $\{\hat{s}(t); t = 0 \sim N_c-1\}$, to be transmitted to BS in the second time-slot, as

$$\hat{s}(t) = \underset{\hat{x}(t)\in\chi}{\arg\min}\left|\hat{d}_{Mi}(t) - \left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{Mi}(k)W_{Mi}(k)\right)\hat{x}(t)\right|^2 , \tag{11}$$

where χ is the set of the data-modulated symbol candidate.

B. Signal processing at BS

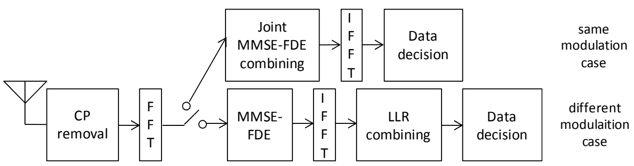

Fig. 5 BS structure.

BS receives the signal block directly from MT in the first time-slot and the remodulated signal block from RS_i in the second time-slot. Figure 5 shows the BS structure. The received signal $\{y_{iB}(t); t=0,\ldots,N_c-1\}$ at BS in the second time slot is given as

$$y_{iB}(t) = \sqrt{2P_{iB}}\sum_{l=0}^{L-1}h_{l,iB}\hat{s}\big((t-\tau_{l,iB}) \bmod N_c\big) + n_{iB}(t) \ , \tag{12}$$

where $h_{l,iB}$ and $\tau_{l,iB}$ are the complex path gain and the time delay of the RS_i-BS link. $n_{iB}(t)$ is the Gaussian noise with zero mean and variance $2N_0/T_s$.

N_c-point FFT is applied to $\{y_{MB}(t); t=0,\ldots,N_c-1\}$ and $\{y_{iB}(t); t=0,\ldots,N_c-1\}$ to transform them into the frequency-domain received signals $\{Y_{MB}(k); k=0,\ldots,N_c-1\}$ and $\{Y_{iB}(k); k=0,\ldots,N_c-1\}$, respectively. $Y_{MB}(k)$ and $Y_{iB}(k)$ are given as

$$\begin{cases} Y_{MB}(k) = \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}y_{MB}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \qquad\quad = H_{MB}(k)S(k) + \Pi_{MB}(k) \\ Y_{iB}(k) = \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}y_{iB}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \qquad\quad = H_{iB}(k)\hat{S}(k) + \Pi_{iB}(k) \end{cases} , \tag{13}$$

where

$$\begin{cases} \hat{S}(k) = \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}\hat{s}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ H_{MB}(k) = \sqrt{2P_{MB}}\sum_{l=0}^{L-1}h_{l,MB}\exp\left(-j2\pi k\frac{\tau_{l,MB}}{N_c}\right) \\ H_{iB}(k) = \sqrt{2P_{iB}}\sum_{l=0}^{L-1}h_{l,iB}\exp\left(-j2\pi k\frac{\tau_{l,iB}}{N_c}\right) \\ \Pi_{MB}(k) = \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}n_{MB}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \Pi_{iB}(k) = \frac{1}{\sqrt{N_c}}\sum_{t=0}^{N_c-1}n_{iB}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \end{cases} . \tag{14}$$

1) *Combining in case of same modulation level*

When the same data modulation level is used in the first and second time-slots, joint MMSE-FDE combining [7] is applied. The frequency-domain signal $\{\hat{Y}(k); k = 0,\ldots,N_c-1\}$ after joint MMSE-FDE combining is expressed as

$$\hat{Y}(k) = Y_{MB}(k)W_{MB}(k) + Y_{iB}(k)W_{iB}(k) \ , \tag{15}$$

where $W_{MB}(k)$ and $W_{iB}(k)$ are the MMSE weights which jointly minimizes the MSE between $\hat{Y}(k)$ and $S(k)$ and are given as [7]

$$\begin{cases} W_{MB}(k) = \frac{H^*_{MB}(k)}{\left|H_{MB}(k)\right|^2 + \left|H_{iB}(k)\right|^2 + 2N_0/T_s} \\ W_{iB}(k) = \frac{H^*_{iB}(k)}{\left|H_{MB}(k)\right|^2 + \left|H_{iB}(k)\right|^2 + 2N_0/T_s} \end{cases} . \tag{16}$$

The frequency-domain signal $\{\hat{Y}(k); k = 0,\ldots,N_c-1\}$ is transformed by N_c-point IFFT back to the time-domain signal $\{\hat{d}(t); t = 0,\ldots,N_c-1\}$ as

$$\hat{d}(t) = \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\hat{Y}(k)\exp\left(j2\pi t\frac{k}{N_c}\right) . \tag{17}$$

2) *Combining in case of different modulation level*

When the data-modulation level is different in the first and second time-slots, the received signals from MT and RS are combined using log-likelihood ratio (LLR) combining [8] after MMSE-FDE.

One-tap MMSE-FDE for $\{Y_{MB}(k);k=0,\ldots,N_c-1\}$ and $\{Y_{iB}(k);k=0,\ldots,N_c-1\}$ are expressed as

$$\begin{cases}\hat{Y}_{MB}(k)=Y_{MB}(k)W_{MB}(k)\\ \hat{Y}_{iB}(k)=Y_{iB}(k)W_{iB}(k)\end{cases}, \tag{18}$$

where $W_{MB}(k)$ and $W_{iB}(k)$ are the MMSE weights given as [7]

$$\begin{cases}W_{MB}(k)=\dfrac{H^*_{MB}(k)}{\left|H_{MB}(k)\right|^2+2N_0/T_s}\\ W_{iB}(k)=\dfrac{H^*_{iB}(k)}{\left|H_{iB}(k)\right|^2+2N_0/T_s}\end{cases}. \tag{19}$$

The equalized frequency-domain signals $\{\hat{Y}_{MB}(k);k=0,\ldots,N_c-1\}$ and $\{\hat{Y}_{iB}(k);k=0,\ldots,N_c-1\}$ are transformed by N_c-point IFFT back to the time-domain signals $\{\hat{d}_{MB}(t);t=0,\ldots,N_c-1\}$ and $\{\hat{d}_{iB}(t);t=0,\ldots,N_c-1\}$ as

$$\begin{cases}\hat{d}_{MB}(t)=\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{k=0}^{N_c-1}\hat{Y}_{MB}(k)\exp\left(j2\pi t\dfrac{k}{N_c}\right)\\ \hat{d}_{iB}(t)=\dfrac{1}{\sqrt{N_c}}\displaystyle\sum_{k=0}^{N_c-1}\hat{Y}_{iB}(k)\exp\left(j2\pi t\dfrac{k}{N_c}\right)\end{cases}, \tag{20}$$

from which the bit-LLRs, $\lambda_{MB,x}(t)$ and $\lambda_{iB,x}(t)$ of x-th bit within t-th symbol, are computed using [8]

$$\begin{cases}\lambda_{MB,x}(t)=\dfrac{1}{2\sigma^2_{MB}}\left(\begin{aligned}&\left|\hat{d}_{MB}(t)-\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{MB}(k)W_{MB}(k)d^{\min}_{MB,b(x)=0}\right|^2\\&-\left|\hat{d}_{MB}(t)-\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{MB}(k)W_{MB}(k)d^{\min}_{MB,b(x)=1}\right|^2\end{aligned}\right)\\ \lambda_{iB,x}(t)=\dfrac{1}{2\sigma^2_{iB}}\left(\begin{aligned}&\left|\hat{d}_{iB}(t)-\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{iB}(k)W_{iB}(k)d^{\min}_{iB,b(x)=0}\right|^2\\&-\left|\hat{d}_{iB}(t)-\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_{iB}(k)W_{iB}(k)d^{\min}_{iB,b(x)=1}\right|^2\end{aligned}\right)\end{cases}. \tag{21}$$

$d^{\min}_{MB,b(x)=0(1)}$ and $d^{\min}_{iB,b(x)=0(1)}$ are the symbol candidates which have the maximum LLR within the set of the data symbols. $2\sigma^2_{MB}$ and $2\sigma^2_{iB}$ are the sums of variances of the residual ISI and the noise given as

$$\begin{cases}2\sigma^2_{MB}=\dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left|H_{MB}(k)W_{MB}(k)\right|^2-\left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}H_{MB}(k)W_{MB}(k)\right|^2\\ \qquad+\dfrac{1}{N_c}\dfrac{2N_0}{T_s}\displaystyle\sum_{k=0}^{N_c-1}\left|W_{MB}(k)\right|^2\\ 2\sigma^2_{iB}=\dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left|H_{iB}(k)W_{iB}(k)\right|^2-\left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}H_{iB}(k)W_{iB}(k)\right|^2\\ \qquad+\dfrac{1}{N_c}\dfrac{2N_0}{T_s}\displaystyle\sum_{k=0}^{N_c-1}\left|W_{iB}(k)\right|^2\end{cases}. \tag{22}$$

The combined bit-LLR $\lambda_x(t)$ is given as

$$\lambda_x(t)=\lambda_{MB,x}(t)+\lambda_{iB,x}(t). \tag{23}$$

If $\lambda_x(t)$ is positive (negative), the estimated transmit bit is 1 (0).

IV. COMPUTER SIMULATION

The simulation condition is shown in Table 1. A frequency-selective block Rayleigh fading channel having a symbol-spaced L=16-path uniform power delay profile is assumed. The frame length (sum of the first and second time-slots in time) is kept constant and is always 512 symbols. MT is assumed to randomly move around in the cell.

The RS selection is based on the throughput maximization of the worst link between MT-RS and RS-BS. The RS selection is expressed as

$$RS_i=\arg\max_{i'}\left\{\min\left(\sum_{l=0}^{L-1}\left|h_{Mi',l}\right|^2\cdot r_{Mi'}^{-\alpha}\cdot 10^{-\eta_{Mi'}/10},\sum_{l=0}^{L-1}\left|h_{i'B,l}\right|^2\cdot r_{i'B}^{-\alpha}\cdot 10^{-\eta_{i'B}/10}\right)\right\}. \tag{24}$$

Table 1 Simulation condition.

Transmission data	Block size	N_c=512
	GI length	N_g=16
	Packet size	60 (blocks)
Channel	Fading type	Block Rayleigh fading
	Power delay profile	Uniform
	No. of paths	L=16
	Path loss exponent	α=3.5
	Shadowing loss standard deviation	σ=7.0 (dB)
RS receiver	Equalization	MMSE-FDE
	Channel estimation	Ideal
BS receiver	Equalization	MMSE-FDE
	Combining scheme	MMSE-FDE combining, LLR combining
	Channel estimation	Ideal

A. Optimal modulation combination

First, we investigate the optimal modulation combination, which maximizes the throughput, of the MT in the first time slot and the RS in the second time slot. Figure 6 shows the simulation model. It is assumed that the average received SNRs of MT-RS and RS-BS links are Δ_{Mi} and Δ_{iB} dB higher than that of MT-BS link, respectively.

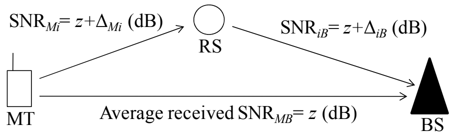

Fig. 6 Simulation model.

Figure 7 shows the optimal modulation combination which maximizes the throughput when the average received SNR of MT-BS link is z=5dB. Using higher level modulation in the better time-slot, more transmission time is given to the worse time-slot and hence lower level modulation can be used to transmit the same amount of data bits in the first and second time-slots. It should be noted that the same amount of data bits must be transmitted in the first and second time-slots. It can be seen from Fig. 7 that using higher level modulation in the first time-slot maximizes the throughput when Δ_{Mi} is high and using higher level modulation in the second time-slot maximizes the throughput when Δ_{iB} is high. For example, when $(\Delta_{Mi},\Delta_{iB})$ = (15,25), using a modulation combination of QPSK-16QAM achieves higher throughput than using the same modulation at MT and RS.

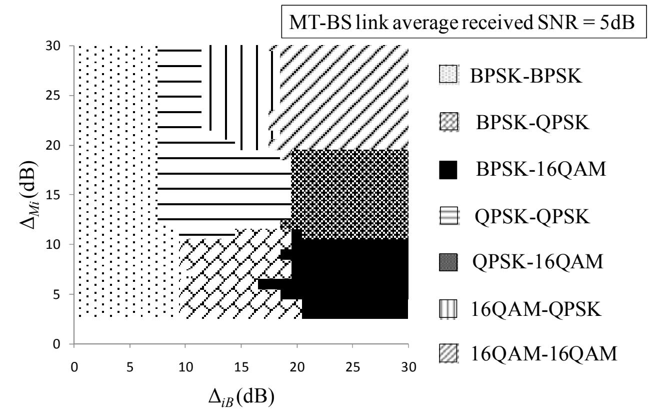

Fig. 7 Optimal modulation combination.

B. *Throughput performance*

In this subsection, we evaluate the average throughput performance of cooperative DF relay with adaptive modulation based on the result obtained by the computer simulation. In the adaptive modulation, the different modulation combination is chosen each time a symbol block is transmitted. Therefore, the average throughput of cooperative DF relay using adaptive modulation is given as

$$S = \frac{1}{\left(1+N_g/N_c\right)}\sum_{a=0}^{A-1} P_a\left\{\frac{m_{a,1}m_{a,2}}{m_{a,1}+m_{a,2}}\left(1-PER_a\right)\right\} \quad \text{[bps/Hz]} \quad ,(25)$$

where P_a denotes the probability that the a-th(a=0,…,A−1) modulation combination is selected. PER_a, $m_{a,1}$ and $m_{a,2}$ are the PER and the modulation levels in the first and second time-slots when the a-th modulation combination is used, respectively. A is the number of possible modulation combination.

Figure 8 plots the 10%-outage throughput, below which the throughput falls with the probability of 10%. For comparison, the 10%-outage throughputs of cooperative DF relay using fixed modulation combination are also plotted as the dashed line and that of direct communication using adaptive modulation as the dashed-dotted line.

It can be seen from Fig. 8 that the cooperative DF relay using proposed adaptive modulation can achieve the best throughput performance. When the required throughput is 0.5 bps/Hz, the proposed adaptive modulation can reduce the normalized total transmit SNR by about 1.5 dB compared to the modulation combination of QPSK-QPSK. Also as seen from Fig. 8 is that, except for the very high SNR region, cooperative DF relay using adaptive modulation provides higher throughput than the direct communication using adaptive modulation since the spatial diversity gain is obtained by cooperative DF relay. When the required throughput is 0.8 bps/Hz, the proposed adaptive modulation can reduce the normalized total transmit SNR by about 9.0 dB. However, for a very high transmit SNR region (e.g; >26dB), the direct communication provides higher throughput than the cooperative DF relay because of the use of 2-time slot relay.

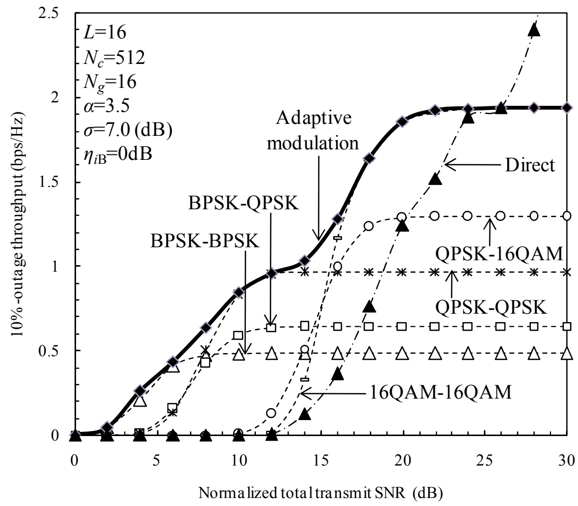

Fig. 8 10%-outage throughput performance.

V. CONCLUSION

In this paper, we proposed 2-time slot cooperative DF relay using adaptive modulation. We first discussed the optimal modulation combination. Then, we evaluated the throughput performance of proposed cooperative SC-DF relay. It was confirmed by the computer simulation that the cooperative DF relay using adaptive modulation can achieve higher throughput than both the DF relay using fixed modulation and the direct transmission.

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