# Adaptive Power Allocation for Bi-Directional Single-Carrier Relay Using Analog Network Coding

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***Abstract*— Relay using analog network coding (ANC) can achieve the same maximum throughput as the direct transmission while reducing the required transmit power. In this paper, we propose an adaptive power allocation scheme for bi-directional single-carrier (SC) ANC. The proposed scheme adaptively allocates the transmit power so as to make the instantaneous signal-to-interference plus noise power ratios (SINRs) of both up and down links identical and maximized. Since it is quite difficult to obtain the optimal solution, we derive an approximate solution. We evaluate, by computer simulation, the distributions of bit error rate (BER) and throughput when using the proposed power allocation. It is shown that the proposed scheme can improve the BER and throughput performances compared to equal power allocation.**

***Keywords-component; Relay network, analog network coding, single-carrier transmission***

## I. INTRODUCTION

In the future wireless network, broadband data services of around 1Gbps are demanded. However, for a user near the cell edge, a significantly high transmit power is required due to the path loss and shadowing loss [1]. Cooperative relay is a promising technique to solve this problem [2]. However, its achievable maximum throughput is half the direct transmission since the bi-directional relay needs four time-slots while the direct transmission needs two time-slots.

An application of network coding is an effective way to mitigate the throughput loss for bi-directional relay. There are two type of network coding: digital network coding (DNC) [3] and analog network coding (ANC) [4-6]. DNC is based on decode-and-forward (DF) relaying protocol, while ANC is based on amplify-and-forward (AF) relaying protocol. ANC has been attracting much attention because it requires two time-slots for bi-directional relaying while DNC requires three time-slots and hence, it can achieve the same maximum throughput as the direct transmission. In addition to network coding, transmit power allocation among base station (BS), relay station (RS), and mobile terminal (MT) is an effective method to improve the throughput performance in bi-directional relay network. Several power allocation methods have been studied for orthogonal frequency division multiplexing (OFDM) based two-way AF relay network [7-9]. To the best of our knowledge, the power allocation method has not been studied yet for bi-directional relay using single-carrier (SC) ANC.

In this paper, we propose an adaptive power allocation scheme for bi-directional relay using SC ANC. The proposed scheme adaptively allocates the transmit power (among BS, RS and MT) so as to make the instantaneous signal-to-interference plus noise power ratios (SINRs) after frequency-domain equalization (FDE) of both up and down links identical and maximized. Since it is difficult to obtain the optimal solution, we derive an approximate solution. We evaluate, by computer simulation, the distributions of BER and throughput when using the proposed power allocation scheme.

The remainder of this paper is organized as follows. Section II describes a system model and signal transmission of the bi-directional relay using SC ANC. Section III proposes the adaptive power allocation scheme. Section IV discusses the computer simulation results. Section V offers conclusions.

## II. BI-DIRECTIONAL RELAY USING SC ANC

Figure 1 illustrates the system model of the bi-directional relay using SC ANC considered in this paper. The communication area is assumed to be within the circle of radius $R$. As shown in Fig. 1, there are $K$ RSs in the cell. We consider the single-user environment. The distances between MT and $i$th RS (denoted by $RS_i$) and between BS and $RS_i$ are respectively denoted by $R_{Mi}$ and $R_{Bi}$ ($i$=0,…, $K$–1).

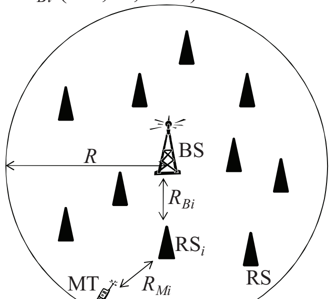


Fig. 1 System model.

ANC requires two time-slots for the bi-directional relay as shown in Figure 2. In the first time-slot, BS and MT transmit simultaneously their signals to RS. RS receives the superposition of the two signals. After amplifying the received signal in the first time-slot, RS broadcasts it to BS and MT in the second time-slot. At BS and MT, the desired signal is detected from the received signal in the second time-slot after the self-interference removal.

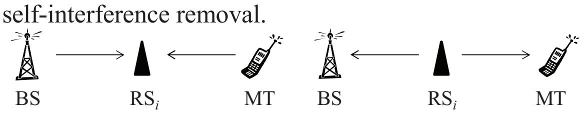


Fig. 2 Bi-directional relay using ANC.

In this paper, symbol-spaced discrete-time signal representation is used. Below, we assume $RS_i$ is selected.

### A. First time-slot

At BS and MT, the data symbol sequence to be transmitted is divided into a sequence of $N_c$-symbol blocks. After insertion of $N_g$-sample cyclic prefix (CP) into the beginning of each block, BS and MT transmit simultaneously their signal blocks to $RS_i$. Denoting the transmit symbol blocks of BS and MT by $\{x_B(t):t=-N_g\sim N_c-1\}$ and $\{x_M(t):t=-N_g\sim N_c-1\}$, respectively, the received signal $\{y_i(t):t=-N_g\sim N_c-1\}$ at $RS_i$ can be expressed as

$$\begin{aligned} y_i(t) &= \sqrt{2\overline{P}_B R_{Bi}^{-\alpha} 10^{-\frac{\eta_{Bi}}{10}}} \sum_{l=0}^{L-1} \overline{h}_{Bi,l} x_B(t-\tau_l) \\ &+ \sqrt{2\overline{P}_M R_{Mi}^{-\alpha} 10^{-\frac{\eta_{Mi}}{10}}} \sum_{l=0}^{L-1} \overline{h}_{Mi,l} x_M(t-\tau_l) + n_i(t) \end{aligned} . \qquad (1)$$

In Eq. (1), $\overline{P}_B$ and $\overline{P}_M$ are the transmit powers at BS and MT, respectively. $\alpha$ denotes the path loss exponent, and $\eta_{Bi}$ and $\eta_{Mi}$ are the shadowing losses (in dB) between BS and $RS_i$ and between MT and $RS_i$, respectively. $\overline{h}_{Bi,l}$ and $\overline{h}_{Mi,l}$ are the complex path gains of the $l$th path of the channels between BS and $RS_i$ and between MT and $RS_i$, respectively. $\tau_l$ denotes the time delay of the $l$th path. $L$ denotes the number of propagation paths. $n_i(t)$ is the independent zero-mean complex-valued Gaussian noise having variance $2N_0/T_s$ with $N_0$ and $T_s$ being the single-sided power spectrum density of the AWGN and the symbol duration, respectively. Eq. (1) can be rewritten as

$$y_i(t) = \sqrt{2P_B} \sum_{l=0}^{L-1} h_{Bi,l} x_B(t-\tau_l) + \sqrt{2P_M} \sum_{l=0}^{L-1} h_{Mi,l} x_M(t-\tau_l) + n_i(t) , \qquad (2)$$

where $P_B = \overline{P}_B R^{-\alpha}$ and $P_M = \overline{P}_M R^{-\alpha}$ are the normalized transmit powers at BS and MT, respectively. $h_{Bi,l}$ and $h_{Mi,l}$ are the $l$th path complex-valued gains, including effects of path loss and shadowing loss of the links between BS and $RS_i$ and between MT and $RS_i$, respectively. They are given as

$$\begin{aligned} h_{Bi,l} &= \overline{h}_{Bi,l} \cdot \sqrt{r_{Bi}^{-\alpha} \cdot 10^{-\eta_{Bi}/10}} \\ h_{Mi,l} &= \overline{h}_{Mi,l} \cdot \sqrt{r_{Mi}^{-\alpha} \cdot 10^{-\eta_{Mi}/10}} \end{aligned} , \qquad (3)$$

where $r_{Bi}=R_{Bi}/R$ and $r_{Mi}=R_{Mi}/R$ are the normalized distances between BS and $RS_i$ and between MT and $RS_i$, respectively.

### B. Second time-slot

In the second time slot, after amplifying the received signal in the first time slot, RS broadcasts it to BS and MT. The received signals $\{y_B(t):t=-N_g\sim N_c-1\}$ and $\{y_M(t):t=-N_g\sim N_c-1\}$, at BS and MT, can be respectively expressed as

$$\begin{cases} y_B(t) = \dfrac{1}{\beta}\sqrt{2P_i} \displaystyle\sum_{l=0}^{L-1} h_{Bi,l} y_i(t-\tau_l) + n_B(t) \\ y_M(t) = \dfrac{1}{\beta}\sqrt{2P_i} \displaystyle\sum_{l=0}^{L-1} h_{Mi,l} y_i(t-\tau_l) + n_M(t) \end{cases} , \qquad (4)$$

where $P_i = \overline{P}_i R^{-\alpha}$ is the normalized transmit power at RS with $\overline{P}_i$ denoting the RS transmit power. $n_B(t)$ and $n_M(t)$ are the independent zero-mean complex-valued Gaussian noises having variance $2N_0/T_s$ at BS and MT, respectively. $\beta$ is the power normalization factor to keep the average transmit power at RS constant and is given as

$$\begin{aligned} \beta &= \sqrt{E\left[|y_i(t)|^2\right]} \\ &= \sqrt{2P_B \sum_{l=0}^{L-1} |h_{Bi,l}|^2 + 2P_M \sum_{l=0}^{L-1} |h_{Mi,l}|^2 + \frac{2N_0}{T_s}} \end{aligned} . \qquad (5)$$

The normalized total transmit power $P_T$ is defined as

$$P_T = P_i + P_M + P_B . \qquad (6)$$

Figure 3 illustrates the receiver structures of BS and MT. After CP removal, the received signal blocks, $\{y_B(t):t=0\sim N_c-1\}$ and $\{y_M(t):t=0\sim N_c-1\}$, are transformed by $N_c$-point fast Fourier transform (FFT) into the frequency-domain signals $\{Y_B(k):k=0\sim N_c-1\}$ and $\{Y_M(k):k=0\sim N_c-1\}$. $Y_B(k)$ and $Y_M(k)$ are given as

$$\begin{cases} Y_B(k) = \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_M} H_{Bi}(k) H_{Mi}(k) X_M(k) \\ \quad + \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_B} H_{Bi}(k) H_{Bi}(k) X_B(k) \\ \quad + \dfrac{1}{\beta}\sqrt{2P_i} H_{Bi}(k) N_i(k) + N_B(k) \\ Y_M(k) = \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_B} H_{Mi}(k) H_{Bi}(k) X_B(k) \\ \quad + \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_M} H_{Mi}(k) H_{Mi}(k) X_M(k) \\ \quad + \dfrac{1}{\beta}\sqrt{2P_i} H_{Mi}(k) N_i(k) + N_M(k) \end{cases} , \qquad (7)$$

where $X_B(k)$ and $X_M(k)$ are respectively the $k$th frequency components of the transmitted signal blocks of BS and MT, and $H_{Bi}(k)$ and $H_{Mi}(k)$ are respectively the $k$th frequency channel gains of the links between BS and $RS_i$ and between MT and $RS_i$. $N_B(k)$, $N_M(k)$, and $N_i(k)$ are the $k$th frequency noise components at BS, MT, and $RS_i$. They are given as

$$\begin{cases} X_B(k) = \sqrt{\dfrac{1}{N_c}} \displaystyle\sum_{t=0}^{N_c-1} x_B(t) \exp(-2\pi kt/N_c) \\ X_M(k) = \sqrt{\dfrac{1}{N_c}} \displaystyle\sum_{t=0}^{N_c-1} x_M(t) \exp(-2\pi kt/N_c) \end{cases} , \qquad (8)$$

$$\begin{cases} H_{Bi}(k) = \displaystyle\sum_{l=0}^{L-1} h_{Bi,l} \exp(-2\pi k\tau_l/N_c) \\ H_{Mi}(k) = \displaystyle\sum_{l=0}^{L-1} h_{Mi,l} \exp(-2\pi k\tau_l/N_c) \end{cases} , \qquad (9)$$

and

$$\begin{cases} N_B(k) = \sqrt{\dfrac{1}{N_c}} \displaystyle\sum_{t=0}^{N_c-1} n_B(t) \exp(-2\pi kt/N_c) \\ N_M(k) = \sqrt{\dfrac{1}{N_c}} \displaystyle\sum_{t=0}^{N_c-1} n_M(t) \exp(-2\pi kt/N_c) \\ N_i(k) = \sqrt{\dfrac{1}{N_c}} \displaystyle\sum_{t=0}^{N_c-1} n_i(t) \exp(-2\pi kt/N_c) \end{cases} . \qquad (10)$$

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Self-interference removal is carried out on each received signal as

$$\begin{cases} \tilde{Y}_B(k) = Y_B(k) - \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_B}\, H_{Bi}(k)H_{Bi}(k)X_B(k) \\ \tilde{Y}_M(k) = Y_M(k) - \dfrac{1}{\beta}\sqrt{2P_i \cdot 2P_M}\, H_{Mi}(k)H_{Mi}(k)X_M(k) \end{cases}, \quad (11)$$

Then, one-tap FDE is carried out as

$$\begin{cases} \hat{Y}_B(k) = \tilde{Y}_B(k)W_B(k) \\ \hat{Y}_M(k) = \tilde{Y}_M(k)W_M(k) \end{cases}, \quad (12)$$

where $W_B(k)$ and $W_M(k)$ are the MMSE-FDE weights given as [5]

$$\begin{cases} W_B(k) = \dfrac{H^*_{Bi}(k)H^*_{Mi}(k)}{|H_{Bi}(k)H_{Mi}(k)|^2 + \left\{|H_{Bi}(k)|^2 + \tilde{\beta}^2\right\}\left(\dfrac{P_M}{N}\right)^{-1}} \\ W_M(k) = \dfrac{H^*_{Bi}(k)H^*_{Mi}(k)}{|H_{Bi}(k)H_{Mi}(k)|^2 + \left\{|H_{Mi}(k)|^2 + \tilde{\beta}^2\right\}\left(\dfrac{P_B}{N}\right)^{-1}} \end{cases} \quad (13)$$

with $\tilde{\beta} = \beta/\sqrt{2P_i}$, $N=N_0/T_s$ being the noise power, and $[.]^*$ denoting the complex conjugate operation.

Finally, the frequency-domain signal after MMSE-FDE is transformed by $N_c$-point inverse FFT (IFFT) back to the time-domain signal for data demodulation.

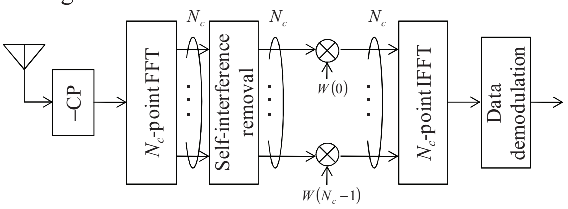



Fig. 3 Receiver structure of BS and MT. ($W(k)=W_B(k)$ for BS and $W_M(k)$ for MT)

## III. Adaptive Power Allocation

### A. Instantaneous SINR

The instantaneous SINRs, $\gamma_B$ and $\gamma_M$, after FDE for the uplink (at BS) and for the downlink (at MT) can be respectively given, from Eqs. (7), (11), and (12), as [5]

$$\begin{cases} \gamma_B = \dfrac{2\dfrac{P_i}{N}\dfrac{P_M}{N}\left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1} H_{Bi}(k)H_{Mi}(k)W_B(k)\right|^2}{\dfrac{P_i}{N}\dfrac{P_M}{N}\hat{H}_B + \dfrac{P_i}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)W_B(k)|^2 + \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|W_B(k)|^2\left\{\dfrac{P_B}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)|^2 + \dfrac{P_M}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Mi}(k)|^2 + 1\right\}} \\ \gamma_M = \dfrac{2\dfrac{P_i}{N}\dfrac{P_B}{N}\left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1} H_{Bi}(k)H_{Mi}(k)W_M(k)\right|^2}{\dfrac{P_i}{N}\dfrac{P_B}{N}\hat{H}_M + \dfrac{P_i}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Mi}(k)W_M(k)|^2 + \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|W_M(k)|^2\left\{\dfrac{P_B}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)|^2 + \dfrac{P_M}{N}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Mi}(k)|^2 + 1\right\}} \end{cases}, \quad (14)$$

where

$$\begin{cases} \hat{H}_B = \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)H_{Mi}(k)W_B(k)|^2 - \left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}H_{Bi}(k)H_{Mi}(k)W_B(k)\right|^2 \\ \hat{H}_M = \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)H_{Mi}(k)W_M(k)|^2 - \left|\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}H_{Bi}(k)H_{Mi}(k)W_M(k)\right|^2 \end{cases}. \quad (15)$$

The first term in the denominator of Eq. (14) is the contribution from the residual ISI. The second, third, and fourth terms are the contributions from noises at RS, BS, and MT, respectively.

### B. Proposed adaptive power allocation

The power allocation problem is expressed as

$$\{P_i, P_M, P_B\} = \arg\max_{P_i, P_M, P_B} \gamma_B, \gamma_M$$
$$\text{s.t.}\begin{cases} P_i + P_M + P_B = P_T \\ P_i > 0,\quad P_M > 0,\quad P_B > 0 \end{cases}. \quad (16)$$

Due to the transmit power constraint, there exists the trade-off relationship between $\gamma_B$ and $\gamma_M$. As the uplink throughput improves, the downlink throughput decreases and vice versa. In this paper, we want to achieve equal throughput for the up and down links (i.e., equal SINR). The optimization problem can be rewritten as

$$\{P_i, P_M, P_B\} = \arg\max_{P_i, P_M, P_B} \gamma_B, \gamma_M$$
$$\text{s.t.}\begin{cases} P_i + P_M + P_B = P_T \\ \gamma_B = \gamma_M \\ P_i > 0,\quad P_M > 0,\quad P_B > 0 \end{cases}. \quad (17)$$

It is quite difficult to solve the optimization problem of Eq. (17). Therefore, we derive an approximate solution by using the FDE weights with equal power allocation ($P_i=P_T/2$, $P_B=P_M=P_T/4$). The proposed scheme allocates the transmit power so as to achieve equal throughput for both up and down links. Therefore, the optimal solution of Eq. (17) is assumed to be close to the equal power allocation. From Eq. (13), the FDE weights when using equal power allocation are given as

$$\begin{cases} W_B(k) = \dfrac{H^*_{Bi}(k)H^*_{Mi}(k)}{|H_{Bi}(k)H_{Mi}(k)|^2 + \left\{|H_{Bi}(k)|^2 + \hat{\beta}^2\right\}\left(\dfrac{P_T}{4N}\right)^{-1}} \\ W_M(k) = \dfrac{H^*_{Bi}(k)H^*_{Mi}(k)}{|H_{Bi}(k)H_{Mi}(k)|^2 + \left\{|H_{Mi}(k)|^2 + \hat{\beta}^2\right\}\left(\dfrac{P_T}{4N}\right)^{-1}} \end{cases}, \quad (18)$$

where

$$\hat{\beta} = \sqrt{\frac{1}{2}\sum_{l=0}^{L-1}|h_{Bi,l}|^2 + \frac{1}{2}\sum_{l=0}^{L-1}|h_{Mi,l}|^2 + 1}\ . \quad (19)$$

When the total transmit signal-to-noise power ratio (SNR) $P_T/N$ is sufficiently high, Eq. (18) can be approximated as

$$W_B(k) \approx W_M(k) = \begin{cases} \dfrac{H_{Bi}^*(k)H_{Mi}^*(k)}{|H_{Bi}(k)H_{Mi}(k)|^2 + \tilde{\beta}^2\left(\dfrac{P_T}{4N}\right)^{-1}} & \text{if } |H_{Bi}(k)|^2, |H_{Mi}(k)|^2 << \tilde{\beta}^2 \\ \dfrac{H_{Bi}^*(k)H_{Mi}^*(k)}{|H_{Bi}(k)H_{Mi}(k)|^2} & \text{otherwise} \end{cases}. \tag{20}$$

The approximate FDE weights for equal power allocation are not a function of the transmit powers $P_i$, $P_M$, and $P_B$. Therefore, the approximate solution to Eq. (17) can be derived by using Karush-Kuhn-Tucher (KKT) conditions [10]. The proposed adaptive power allocation is given as

$$\begin{cases} P_i = P_T \dfrac{\dfrac{(A-B)(C-D)}{(A+B-C-D)}P_T + (CP_T + DP_T + 2N)}{\left[\begin{array}{l} 2\dfrac{(A-B)(C-D)}{(A+B-C-D)}P_T + (CP_T + DP_T + 2N) \\ + \sqrt{2(AP_T+N)(CP_T+N) + 2(BP_T+N)(DP_T+N)} \end{array}\right]}, \\ P_M = \dfrac{(P_T - 2P_i)\{DP_T + N + (A-D)P_i\}}{(C-D)(2P_i - P_T) + (A+B-C-D)P_T} \\ P_B = P_T - P_i - P_M \end{cases} \tag{21}$$

where

$$\begin{cases} A = \dfrac{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)W_B(k)|^2}{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|W_B(k)|^2}, \quad B = \dfrac{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Mi}(k)W_M(k)|^2}{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|W_M(k)|^2} \\ C = \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Bi}(k)|^2, \quad D = \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}|H_{Mi}(k)|^2 \end{cases}. \tag{22}$$

## IV. COMPUTER SIMULATION RESULTS

We evaluate, by computer simulation, the distributions of BER and throughput. Computer simulation condition is shown in Table I. We assume QPSK data modulation. FFT block size $N_c$ and CP size $N_g$ are respectively set to $N_c$=256 and $N_g$=32. One packet consists of 10 blocks. The fading channel is assumed to be a frequency-selective block Rayleigh fading having $L$=16-path uniform power delay profile. The pass loss exponent $\alpha$ and the shadowing loss standard deviation $\sigma$ are respectively assumed to be $\alpha$=3.5 and $\sigma$=7.0(dB). We assume that $K$=6 RSs are located with an equal separation along a circle of a normalized radius of $r_{BR}$=0.5. Since the BS-RS link is a fixed link, the average received SNR $\Gamma$ of the link between BS and RS can be controlled and can be expressed as

$$\Gamma = 10\log_{10}\frac{P_i}{N}r_{BR}^{-\alpha} + \Delta \text{ dB}, \tag{23}$$

where $\Delta$ denotes the SNR deviation from the long-term average received SNR. The location of RS can be determined for the given value of $\Delta$. For simplicity, in this paper, the deviation $\Delta$ is set to $\Delta$=0dB. In the computer simulation, the MT location is randomly generated in a cell. We assume that the nearest RS is selected. Ideal channel estimation is assumed.

Table I Computer simulation condition

| | | |
|---|---|---|
| Transmitter /receiver | Data modulation | QPSK |
| | FFT block size | $N_c$=256 |
| | CP size | $N_g$=32 |
| | Packet size | 10 blocks |
| | Equalization | MMSE-FDE |
| | Channel estimation | Ideal |
| Channel | Fading type | $L$=16-path block Rayleigh fading |
| | Power delay profile | Uniform |
| | Time delay | $\tau_l=l$, $l$=0~$L$−1 |
| | Path loss exponent | $\alpha$=3.5 |
| | Shadowing loss standard deviation | $\sigma$=7.0(dB) |
| | SNR deviation from the long-term average received SNR | $\Delta$=0 |
| Relay | Normalized distance between RS and BS | $r_{BR}$ = 0.5 |

### A. BER distribution

Figure 4 shows the complementary cumulative distribution function (CCDF) of BER when using the proposed power allocation for the normalized total transmit SNR $P_T/N$=20dB. For comparison, the result for equal power allocation is also plotted in Fig. 4(a). When using equal power allocation, different CCDF curves are seen between uplink and downlink. The reason for this is explained below. In ANC-AF relay, the noise which is received and amplified at RS$_i$ is broadcasted to BS and MT. Since the channel gains of BS-RS$_i$ and MT-RS$_i$ links are different, the received noise power forwarded from RS$_i$ is different between BS and MT. Therefore, the instantaneous SINR at BS is different from that at MT when using equal power allocation. On the other hand, when using the proposed power allocation, the CCDF of BER is almost the same for both links. This is because the proposed scheme allocates the transmit power to make the instantaneous SINRs of both links identical.

Fig. 4 (b) compares the proposed power allocation and the optimal solution implemented by greedy algorithm [11]. It can be seen that the proposed power allocation scheme achieves almost identical BER performance to the greedy algorithm.

### B. Outage throughput performance

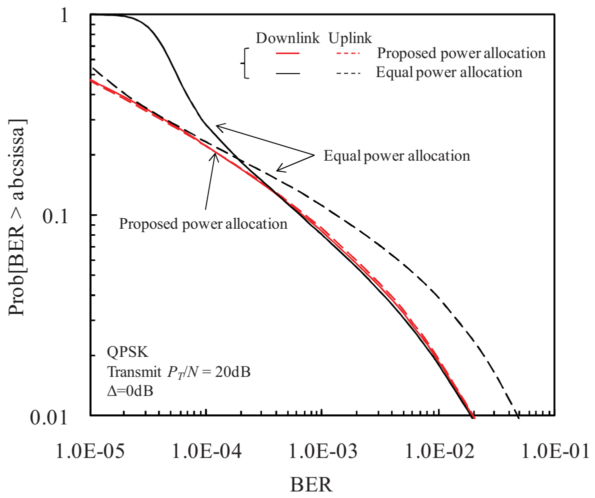
The throughput $S$ is given as

$$S = \frac{M}{2}\cdot(1-PER)\cdot\frac{N_c}{N_c+N_g}, \tag{24}$$

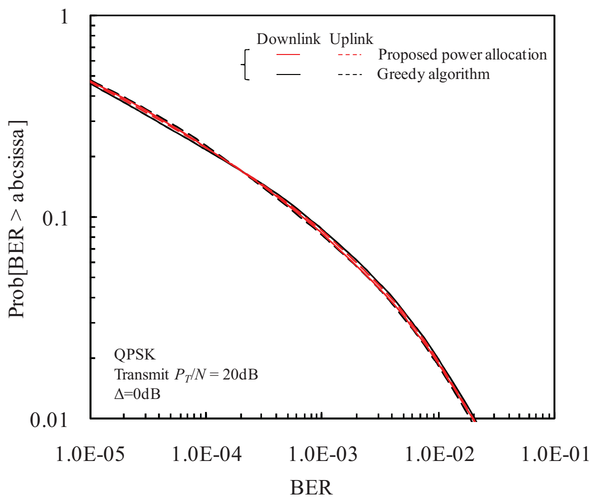
where $M$ denotes the modulation level and $PER$ denotes the packet error rate.

Figure 5 plots the 10%-outage (below which the throughput drops at 10% probability) when using the proposed power allocation with transmit $P_T/N$ as a parameter. For comparison, the performances when using equal power allocation and the direct transmission are also plotted in Fig. 5. When using equal power allocation, the uplink throughput decreases compared to the downlink throughput. The reason for this can be explained below. When MT is located near the cell edge, the channel gains of BS-RS$_i$ link are larger than that of MT-RS$_i$ link due to the shadowing loss. Therefore, the received noise power forwarded from RS$_i$ is larger at BS than at MT. On the other hand, the proposed power allocation achieves the same throughput performance for both up and down links and improves the uplink throughput performance compared to equal power allocation. When transmit $P_T/N$=20dB, the proposed power allocation provides about 1.5 times higher uplink throughput than equal power allocation.

It is seen from Fig. 5 that the bi-directional ANC-AF relay using the proposed power allocation can improve the throughput performances for both up and down links compared to the direct transmission. When transmit $P_T/N$=20dB, the bi-directional ANC-AF relay using the proposed power allocation provides about 10 times higher throughput than the direct transmission. This is because the bi-directional ANC-AF relay can reduce the impact of the path loss and shadowing loss and achieve the same maximum throughput as the direct transmission.




(a) Comparison to equal power allocation




(b) Comparison to the greedy algorithm

Fig. 4 CCDFs of BER

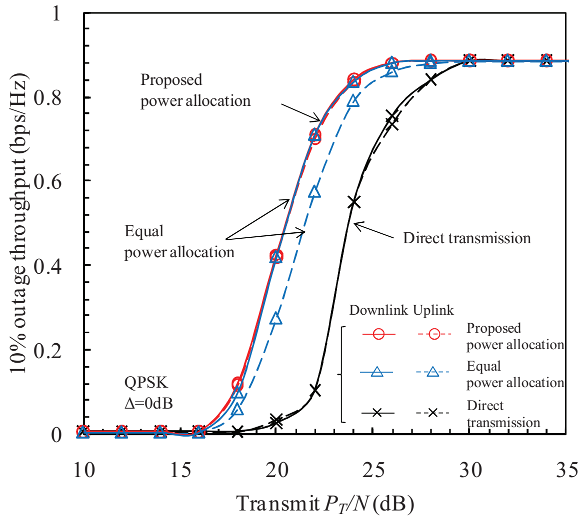



Fig. 5 Throughput performance.

## V. CONCLUSION

In this paper, we proposed an adaptive power allocation for bi-directional relay using SC ANC. The proposed scheme allocates the transmit power so as to make the instantaneous SINR after FDE of up and down links identical and maximized. We derived an approximate power allocation. By computer simulations, it was shown that the proposed scheme can provide about 1.5 times higher uplink throughput than equal power allocation and about 10 times higher throughput than the direct transmission when $P_T/N$=20dB.

The throughputs requirement may be different for up and down links. The power allocation method which can control the up and down link throughputs is left as our future work.

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