Power Allocation Strategies for Distributed Space-Time Codes in Amplify-and-Forward Mode

Behrouz Maham and Are Hjørungnes
UNIK – University Graduate Center, University of Oslo, Norway
Email: {behrouz, arehj}@unik.no

Abstract—The idea of space-time coding, devised for multiple-antenna systems, can be applied to communication over a wireless relay network. In this paper, we derive an approximate formula for the bit error rate (BER) of distributed space-time codes in amplify-and-forward (AF) mode, in which each relay transmits a scaled version of the linear combination of its received symbols. Assuming $M$-PSK or $M$-QAM modulated, full-rate, and full-diversity space-time codes, we derive two power allocation strategies minimizing the approximate BER expressions, for constrained transmit power. Our analytical results have been verified by simulation results.

Index Terms—Cooperation, distributed space-time codes, power allocation, amplify-and-forward.

I. INTRODUCTION

Space-time coding (STC) has received a lot of attention in the last years as a way to increase capacity and/or reduce the transmitted power necessary to achieve a target bit error rate (BER) using multiple antenna transceivers. In ad-hoc network applications or in distributed large scale wireless networks, the nodes are often constrained in the complexity of their hardware and also in their size. This makes multiple-antenna systems impractical for certain network applications [1]. In an effort to overcome this limitation, cooperative diversity schemes have been introduced [1–4]. Cooperative diversity allows a collection of radios to relay signals for each other and effectively create a virtual antenna array for combating multipath fading in wireless channels. The attractive feature of these techniques is that each node is equipped with only one antenna, creating a virtual antenna array. This property makes them outstanding for deployment in cellular mobile devices as well as in ad-hoc mobile networks, which have problem with exploiting multiple-antenna due to the size limitation of the mobile terminals.

Among the most widely used cooperative strategies are amplify-and-forward (AF) [4], [5] and decode-and-forward (DF) [1], [2], [4]. The authors in [6] applied Hurwitz-Radon space-time codes in wireless relay networks and conjecture a diversity factor around $R/2$ for large $R$ from their simulations, where $R$ is the number of relays. In [7], a cooperative strategy was proposed, which achieves a diversity factor of $R$ in $R$-relay wireless networks, using the so-called distributed space-time codes (DSTC). In this strategy, a two-phase protocol was used. In phase one, the transmitter sends the information signal to the relays and in phase two, the relays send information to the receiver. The signal sent by every relay in the second phase is designed as a linear function of its received signal. It was shown in [7] that the relays can generate a linear space-time codeword at the receiver, as in a multiple antenna system, although they only cooperate distributively. This method does not require decoding at the relays, and for high SNR, it achieves the optimal diversity factor [7]. Recently, the design of practical amplify-and-forward DSTC, that lead to reliable communication in wireless relay networks, has been presented in [8] and [9]. Furthermore, in [9], a formula for the bit error probability of the AF DSTC is derived for Rayleigh fading channels.

In [8] and [9], a network with symmetric channels is assumed, in which all source-to-relay and relay-to-destination links have i.i.d. distributions. In this case, it is obvious that uniform power allocation along relays is optimum. However, this assumption is hardly met in practice and the path lengths among nodes could vary. Therefore, power control among the relays is required for such a cooperation. In [9], a closed-form expression for the moment generating function (MGF) of AF space-time cooperation is derived as a function of Whittaker’s function. However, this function is not well-behaved and cannot be used for finding an analytical solution for power allocation.

In this paper, we derive the average BER of AF DSTC system for Rayleigh fading channels, using two new methods based on moment generating function. Furthermore, we propose two power allocation schemes based on minimizing the target BER. An outstanding feature of the proposed schemes is that they are independent of the instantaneous channel variations, and thus, power control coefficients are varying slowly in time. Simulation results show that using the proposed power allocation strategies, considerable gain will be obtained comparing to the uniform power allocation.

The reminder of this paper is organized as follows: In Section II, the system model is given. The average approximate BER of AF DSTC is derived based on the MGF method in Section III. In Section IV, two power allocation schemes minimizing the BER are proposed. In Section V, the performance of the proposed schemes in terms of average BER is demonstrated. Finally, Section VI summarized the conclusions.

Throughout the article, the following notation is applied: The superscripts $^T$ and $^H$ stand for transposition and conjugate transpose, respectively. $E[\cdot]$ denotes the expectation operation. $\text{Cov}(x_T)$ is the covariance of the $T \times 1$ vector $x_T$ consisting of $T$ random variables. All logarithms are the natural logarithm.
Fig. 1. Wireless relay network consisting of a source $s$, a destination $d$, and $R$ relays.

II. SYSTEM MODEL

Consider the network in Fig. 1 consisting of a source denoted $s$, one or more relays denoted $r = 1, 2, \ldots, R$, and one destination denoted $d$. It is assumed that each node is equipped with a single antenna. We denote the source-to-$r$th relay and $r$th relay-to-destination links by $f_r$ and $g_r$, respectively. Suppose each link has a flat Rayleigh fading, and channels are independent of each others. Therefore, $f_r$ and $g_r$ are i.i.d. complex Gaussian random variables with zero-mean and variances $\sigma_f^2$ and $\sigma_g^2$, respectively. Similar to [10], our scheme requires two phases of transmission. During the first phase, the source node, $s$, transmits a scaled version of the signal $s = [s_1, \ldots, s_T]^T$, consisting of $T$ symbols to all relays. We assume the following normalization, $E\{s^H s\} = 1$. Thus, from time $1$ to $T$, the signals $\sqrt{P_1 T} s_1, \ldots, \sqrt{P_1 T} s_T$ are sent to all relays by the source. The average total transmitted energy in $T$ intervals will be $P_1 T$. Assuming $f_r$ is not varying during $T$ successive intervals, the received $T \times 1$ signal at the $r$th relay can be written as

$$r_r = \sqrt{P_1 T} f_r s + v_r,$$

where $v_r$ is a $T \times 1$ complex zero-mean white Gaussian noise vector with the variance of $N_1$. Using amplify-and-forward, each relay scales its received signal, i.e.,

$$y_r = \rho_r r_r,$$

where $\rho_r$ is the scaling factor at relay $r$. When there is no instantaneous channel state information (ICSI) available at the relays, but statistical channel state information (SCSI) is known, a useful constraint is to ensure that a given average transmitted power is maintained. That is,

$$\rho_r^2 = \frac{P_{2,r}}{\sigma_f^2 P_1 + N_1},$$

where $P_{2,r}$ is the average transmitted power at relay $r$. Similar to [10], we assume that the total transmitted power in the second phase is equal to transmitted power from the source, i.e., $P_1 = \sum_{r=1}^R P_{2,r}$. The total power used in the whole network for one symbol transmission is therefore $P = P_1 + \sum_{r=1}^R P_{2,r}$.

DSTC, proposed in [10], uses the idea of linear dispersion space-time codes of multiple-antenna systems. In this system, the $T \times 1$ received signal at the destination can be written as

$$y = \sum_{r=1}^R g_r A_r y_r + w,$$

where $y_r$ is given by (2), $w$ is a $T \times 1$ complex zero-mean white Gaussian noise vector with the component-wise variance of $N_2$, and the $T \times T$ dimensional matrix $A_r$ is corresponding to the $r$th column of a proper $T \times T$ space-time code. The DSTCs designed in [8] and [9] are such that $A_r, r = 1, \ldots, R$, are unitary. Combining (1)-(4), the total noise vector $w_T$ is given by

$$w_T = \sum_{r=1}^R A_r \sqrt{\frac{P_{2,r}}{\sigma_f^2 P_1 + N_1} f_r g_r + w}.$$

Since $g_r, v_r$, and $w$ are complex Gaussian random variables, which are jointly independent, the covariance matrix of $w_T$ can be shown to be

$$\text{Cov}(w_T) = \left( \sum_{k=1}^R \frac{P_{2,k}}{\sigma_f^2 P_1 + N_1} \sigma_g^2 N_1 + N_2 \right) I_T,$$

where $I_T$ is the $T \times T$ identity matrix. Thus, the noise vector $w_T$ is white.

III. APPROXIMATE BER EXPRESSION

In this section, we derive approximate BER expressions for the AF space-time coded cooperation, using moment generating function method.

The conditional BER of the protocol described in Section II, with $R$ relays, can be written as [11, Eq. (9.17)]

$$P_b \left( R \{f_r\}_{r=1}^R \{g_r\}_{r=1}^R \right) = c Q \left( g \sum_{r=1}^R \mu_r |f_r g_r|^2 \right),$$

where parameters $c$ and $g$ are represented as

$$c_{\text{QAM}} = \frac{\sqrt{M - 1}}{\sqrt{M} \log_2 M}, \quad c_{\text{PSK}} = \frac{2}{\log_2 M},$$

and

$$g_{\text{QAM}} = \frac{3}{M - 1}, \quad g_{\text{PSK}} = 2 \sin^2 \left( \frac{\pi}{M} \right),$$

and $\mu_r$ is the portion of the average symbol energy passing from the $r$th relay to noise power ratio. Using (1)-(7), parameters $\mu_r, r = 1, \ldots, R$ can be written as

$$\mu_r = \frac{\frac{P_{2,r}}{\sigma_f^2 P_1 + N_1}}{\sum_{k=1}^R \frac{P_{2,k}}{\sigma_f^2 P_1 + N_1} \sigma_g^2 N_1 + N_2}.$$

It is important to note that in (8), we approximate the conditional variance of the noise vector $w_T$ in (5) as its expected value, which is obtained in (6). The received SNR in the receiver side is denoted by

$$\gamma = \sum_{r=1}^R \gamma_r.$$
where
\[ \gamma_r = \mu_r |f_r g_r|^2. \] (10)

We can calculate the average BER as
\[
P_e(R) = \int_0^\infty P_e(R | \gamma_r) \sum_{r=1}^R \gamma_r \, d \gamma_r = \int_0^\infty c Q(\sqrt{\gamma}) \gamma \, d \gamma.
\] (11)

Now, we are using the MGF method to calculate the BER expression in (11). We also exploit the property that the \( \gamma_r \)'s are independent of each other, because of the inherit spatial separation of relay nodes in the network. Hence, the average BER in (11) can be rewritten as
\[
P_e(R) = \frac{c}{\pi} \int_0^{2\pi} \prod_{r=1}^R \int_0^{\infty} \left( e^{-\frac{\gamma_r}{2\sigma_r^2}} \right) \, d \phi \, \sum_{r=1}^R \left( p(\gamma_r) \, d \gamma_r \right)
\]
\[
= \frac{c}{\pi} \int_0^{2\pi} \prod_{r=1}^R \left( e^{-\frac{\gamma_r}{2\sigma_r^2}} \right) \, d \phi \, M_r(-s) \, d \phi,
\] (12)

where \( M_r(-s) \) is the MGF of the random variable \( \gamma_r \), and \( s = \frac{\sigma_r}{2\sin^2 \phi} \).

It can be shown that for larger values of average SNR, \( \bar{\gamma} \), the behavior of \( \gamma/\bar{\gamma} \) becomes increasingly irrelevant because the \( Q \) term in (11) goes to zero so fast that almost throughout the whole integration range the integrand is almost zero. However, recalling that \( Q(0) = 1/2 \), regardless of the value of \( \bar{\gamma} \), the behavior of \( p(\gamma) \) around zero never loses importance. On the other hand, it is shown in [9, Eq. (18)] that the probability density function (pdf) of the random variables \( \gamma_r \) is proportional to the modified bessel function of second kind of zeroth order, i.e.,
\[
p(\gamma_r) = \frac{2}{\mu_r \sigma_r^2} K_0 \left( 2 \frac{\gamma_r}{\mu_r \sigma_r^2} \right).
\] (13)

The pdf has a very large value around zero. Thus, the behavior of the integrand in (11) around zero becomes very crucial, and we can approximate \( p(\gamma_r) \) in (13) with a logarithmic function, which is easier to handle. From [12, Eq. (24.40)] the following property can be obtained
\[
\lim_{x \to 0} K_0(x) \to -\log(x).
\] (14)

In Fig. 2, we have shown that \( K_0(x) \) and \( -\log(\frac{1}{x}) \) have the same asymptotic behavior when \( x \to 0 \).

Using the property in (14), we can approximate \( M_r(-s) \) as
\[
M_r(-s) \approx \int_0^\infty e^{-s \gamma_r} \frac{1}{\mu_r \sigma_r^2} \log \left( \frac{4 \gamma_r}{\mu_r \sigma_r^2} \right) \, d \gamma_r
= \frac{1}{s \mu_r \sigma_r^2} \left[ \log \left( \frac{8 \mu_r^2 \sigma_r^2}{4} \right) - \kappa \right],
\] (15)

where \( \kappa \) is Euler’s constant, i.e., \( \kappa \approx 0.5772156 \) [13].

Furthermore, for the case of \( R = 1 \), the closed-form solution for the approximate BER is obtained as
\[
P_e(1) \approx \frac{c}{\pi} \int_0^{\pi} M(-s) \, d \phi
= \frac{2c}{\pi g \mu_r \sigma_r^2} \int_0^{\pi} \sin^2 \phi \left( \log \left( \frac{g \mu_r \sigma_r^2}{8 \sin^2 \phi} \right) - \kappa \right) \, d \phi
= \frac{c}{2 \mu_r \sigma_r^2} \left( \log \left( \frac{g \mu_r \sigma_r^2}{8 \sin^2 \phi} \right) - (\kappa + 1) \right).
\] (16)

IV. POWER CONTROL IN AF SPACE-TIME CODED COOPERATION

In this section, we propose two power allocation schemes for the AF distributed space-time codes introduced in [7]. We are using the approximate value of the MGF, which was derived in Section III for the power control among relays. Furthermore, we are presenting another closed-form solution for the MGF, as a function of the incomplete gamma function, which can be used for a more accurate power control strategy.

The MGF of the random variable \( \gamma_r, M(-s) \), which is the integrand of the integral in (12), is given by the product of MGF of the random variables \( \gamma_r \). Since \( M_r(-s) \) is independent of the other \( \mu_i, \neq r \), we can write
\[
\frac{\partial M(-s)}{\partial \mu_r} = \frac{\partial M_r(-s)}{\partial \mu_r} \prod_{i \neq r} M_i(-s),
\] (17)

which will be used in the next two subsections to find the power control coefficients.

A. Power Allocation Based on Exact MGF

The closed-form solution for MGF of random variable \( \gamma_r \), can be found using [13, Eq. (8.353)] as
\[
M_r(-s) = \frac{2}{s \mu_r \sigma_r^2 g_r} \Gamma \left( 0, \frac{1}{s \mu_r \sigma_r^2 g_r} \right) e^{-\mu_r \sigma_r^2 g_r},
\] (18)
where $\Gamma(\alpha, x)$ is the incomplete gamma function of order $\alpha$ [12, Eq. (6.5)]. Moreover, from [13, Eq. (8.356)], we have

$$-d \frac{\Gamma(\alpha, x)}{dx} = x^{\alpha-1}e^{-x}. \quad (19)$$

Since the MGFs in (15) and (18) is a function of $x_r := \mu_r \sigma_r^2, \sigma_r^2, s$, we can express (17) in terms of $x_r$. Hence, using (19), the partial derivative of $M_r(-s)$ with respect to $x_r$ can be expressed as

$$\frac{\partial M_r(-s)}{\partial x_r} = \frac{\partial}{\partial x_r} \left[ \frac{2}{x_r} \Gamma \left( 0, \frac{1}{x_r} \right) e^{\frac{1}{x_r}} \right]
= \frac{1}{x_r^2} \left[ 1 - \Gamma \left( 0, \frac{1}{x_r} \right) \left( 1 + \frac{1}{x_r} \right) e^{\frac{1}{x_r}} \right]. \quad (20)$$

Furthermore, the power constraint in the second phase, i.e., $\sum_{r=1}^{R} P_{2,r} = P_1$, can be expressed as a function of $x_r$. Thus, using (8) and the definition of $x_r$, under the high SNR assumption, we have the following constraint

$$\sum_{r=1}^{R} \frac{x_r}{\sigma_r^2 s} \leq \frac{P_1}{N_2}. \quad (21)$$

Given the objective function as an integrand of (12) and the power constraint in (21), the classical Karush-Kuhn-Tucker (KKT) conditions for optimality [14] can be shown as

$$\prod_{i=1}^{R} \left[ \frac{2}{x_i} \Gamma \left( 0, \frac{1}{x_i} \right) e^{\frac{1}{x_i}} \right] \frac{1}{x_i^2} \left[ 1 - \Gamma \left( 0, \frac{1}{x_i} \right) \left( 1 + \frac{1}{x_i} \right) e^{\frac{1}{x_i}} \right]
+ \frac{\lambda}{\sigma_r^2 s} = 0, \quad \text{for } r = 1, \ldots, R. \quad (22)$$

By solving (21) and (22), the optimum values of $x_r$, i.e., $x^*_r$, $r = 1, \ldots, R$ can be obtained. Now, we can have the following procedure to find the power control coefficients, $P_{2,r}$. First, $x^*_r$ can be solved by the above optimization problem. Then, recalling the relationship between $x_r$ and $\mu_r$, i.e., $x_r = \mu_r \sigma_r^2, \sigma_r^2, s$, and by taking average $\mu_r$ over different values of $\phi$, since $s$ is a function of $\sin^2 \phi$, the optimum value of $\mu_r$ is obtained. Finally, using (8), we can find the power control coefficients, $P_{2,r}$. If we assume that relays operate in high SNR region, $P_{2,r}$ would be approximately proportional to $\mu_r$.

**B. Power Allocation Based on Approximate MGF**

Since the power allocation proposed in Subsection IV-A needs to solve the set of nonlinear equations presented in (22), we present an alternative scheme in this subsection. For gaining insight into the power allocation based on minimizing the BER, we are going to minimize the approximate MGF of the random variable $\gamma$, obtained in (15). Using (15) and (21), we can formulate the following problem:

$$\min_{\{x_1, x_2, \ldots, x_R\}} \frac{1}{\log \left( \frac{x_r}{4} \right)} \left( x_r \right) \prod_{r=1}^{R} \frac{1}{x_r} \left( \log \left( \frac{x_r}{4} \right) - \kappa \right),$$

subject to

$$\sum_{r=1}^{R} \frac{x_r}{\sigma_r^2 s} \leq \frac{P_1}{N_2},$$

$$x_r \geq 0, \quad \text{for } r = 1, \ldots, R. \quad (23)$$

The objective function in (23) is not a convex function in general. However, it can be shown that for $x_r > e^{1+\log(4)+\kappa}$, which is corresponding to high SNR conditions, this function is convex. Therefore, the problem stated in (23) is a convex problem for high SNR values and has a global optimum point.

In the following, we are going to derive a solution for a problem expressed in (23).

The Lagrangian of the problem stated in (23) is

$$L(x_1, x_2, \ldots, x_R) = \prod_{r=1}^{R} \left[ \frac{\log(x_r) - \kappa'}{x_r} \right] + \lambda \left( \sum_{r=1}^{R} \frac{x_r}{\sigma_r^2 s} - \frac{P_1}{N_2} \right), \quad (24)$$

where $\lambda > 0$ is the Lagrange multiplier, and $\kappa' = \log(4) + \kappa$. For nodes $r = 1, \ldots, R$ with nonzero transmitter powers, the KKT conditions are

$$- \log(x_r) + \frac{1 + \kappa'}{x_r^2} \prod_{r=1}^{R} \left[ \frac{\log(x_r) - \kappa'}{x_r} \right] + \lambda \frac{1}{\sigma_r^2 s} = 0. \quad (25)$$

Using (15) and some manipulations, one can rewrite (25) as

$$\left( \frac{1}{x_r} - \frac{1}{x_r \log(x_r) - \kappa'} \right) M(s) = \frac{\lambda}{\sigma_r^2 s}. \quad (26)$$

Since the strong duality condition [14, Eq. (5.48)] holds for convex optimization problems, we have $\lambda \left( \sum_{r=1}^{R} \frac{x_r}{\sigma_r^2 s} - \frac{P_1}{N_2} \right) = 0$ for the optimum point. If we assume Lagrange multiplier has a positive value, we have $\sum_{r=1}^{R} \frac{x_r}{\sigma_r^2 s} = \frac{P_1}{N_2}$. Thereby, by multiplying two sides of (26) with $x_r$, and applying the summation over $r = 1, \ldots, R$, we have

$$M(s) = \frac{\lambda P_1}{N_2}. \quad (27)$$

Dividing both sides of equalities in (26) and (27), we have

$$\frac{1}{x_r} \left( 1 - \frac{1}{\log(x_r) - \kappa'} \right) = \frac{1}{P_1 \sigma_r^2 s} \left[ \frac{R}{\log(x_r) - \kappa'} - \frac{1}{\log(x_r) - \kappa'} \right], \quad (28)$$

for $r = 1, \ldots, R$. The optimal values of $x_r$ in the problem stated in (23) can be obtained with initializing some positive values for $x_r$, $r = 1, \ldots, R$, and using (28) in an iterative manner. Then, we apply the same procedure stated in Subsection IV-A to find the power control coefficients, $P_{2,r}$.

**V. SIMULATION RESULTS**

In this section, the performances of the AF distributed space-time codes with power allocation are studied through simulations. The error event is the BER. The distributed GABBAb space-time codes [9] is used for signal transmission. The transmitted symbols are modulated as BPSK, and We use the block fading model. We fixed the total power consumed in the whole network as $P$ and assumed equal transmitted power for the two phases, i.e., $P_1 = P_2$.

In Fig. 3, we compare the approximate BER formula based on MGF given in (15) with the full-rate, full-diversity distributed GABBAb space-time codes. For GABBAb codes, we employed $4 \times 4$ GABBAb mother codes, i.e., $T = 4$ [15]. Assume that the relays and the destination have the same noise power, i.e., $N_1 = N_2$, and all the links have unit-variance Rayleigh flat fading. Fig. 3 confirms that the analytical results
attained in Section III for finding the BER approximate well the performance of the practical full-diversity distributed space-time codes for high SNR values.

Fig. 4 presents the BER performance of the AF distributed space-time codes using different power allocation schemes. For transmission power among nodes, we employed the two power control schemes introduced in Section IV, and also uniform power transmission among relays, i.e., $P_1^r = \frac{P}{r}$ and $P_2^r = \frac{\pi}{4\pi}$ [9]. Since the proposed power allocation strategies are designed for high SNR scenarios, we study the performance of system in the high SNR regime. Furthermore, since we supposed that the relays are operating in low noise conditions, here, we assume $N_3 = 2N_1$. Slow Rayleigh flat fading channels are considered, with variance of $\sigma_f^2(r) = \sigma_g^2(r) = \frac{1}{2^{r-1}}$, $r = 1, 2, \ldots, R$. For the power control scheme expressed in Subsection IV-A (based on the exact MGF), we have used MATLAB optimization toolbox command “fmincon” designed to find the minimum of the given constrained nonlinear multivariable function. Fig. 4 demonstrate that using the power control schemes of Section IV, about 1 and 2 dB gain will be obtained for $R = 2$ and $R = 3$ cases, respectively, comparing to uniform power allocation. The power control strategy given in Subsection IV-A (Exact MGF-based power control) has a slightly better performance than the power control strategy presented in Subsection IV-B (Approximate MGF-based power control).

VI. CONCLUSION

In this paper, we derived approximate BER formulas of AF space-time coded cooperation using the moment generating function method, when $M$-PSK and $M$-QAM modulations are employed. Simulations results confirmed that the theoretical expressions have a similar performance as the Monte Carlo simulations at high SNR values. Furthermore, we proposed two power allocation methods based on minimizing the BER. Simulations demonstrated that using these schemes, up to 2 dB can be gained in high SNR region, when using three relays.

REFERENCES

[1] A. Sendonaris, E. Erkip, and B. Aazhang, “User cooperation diversity. Part I. System description,” IEEE Trans. Commun., vol. 51, pp. 1927–1938, Nov. 2003.
[2] A. Sendonaris, E. Erkip, and B. Aazhang, “User cooperation diversity. Part II. Implementation aspects and performance analysis,” IEEE Trans. Commun., vol. 51, pp. 1939–1948, Nov. 2003.
[3] J. N. Laneman and G. Wornell, “Energy-efficient antenna sharing and relaying for wireless networks,” in Proc. Wireless Communications Networking Conf., (Chicago, IL), pp. 7–12, Sep. 2000.
[4] J. N. Laneman and G. Wornell, “Distributed space-time coded protocols for exploiting cooperative diversity in wireless networks,” IEEE GLOBECOM 2002, vol. 1, pp. 77–81, Nov. 2002.
[5] R. U. Nabar, H. Bolcskei, and F. W. Kneubueher, “Fading relay channels: Performance limits and space-time signal design,” IEEE Journal on Selected Areas in Communications, vol. 22, pp. 1099–1109, Aug. 2004.
[6] Y. Hsu, Y. Mei, and Y. Chang, “Wireless antennas-making wireless communications perform like wireline communications,” in IEEE AP-S Topical Conf. on Wireless Commn. Tech., (Honolulu, Hawaii), Oct. 2003.
[7] Y. Jing and B. Hassibi, “Distributed space-time coding in wireless relay networks,” in IEEE Sensor Array and Multichannel Signal Processing Workshop, (Sitges, Spain), 2004.
[8] Y. Jing and H. Jafarkhani, “Using orthogonal and quasi-orthogonal designs in wireless relay networks,” in IEEE GLOBECOM, (Pacific Grove, San Fransisco, CA), Nov. 27–Dec. 1, 2006.
[9] B. Maham and A. Hjørungnes, “Distributed GABBA space-time codes in amplify-and-forward cooperation,” in Proc. IEEE Information Theory Workshop (ITW'07), (Bergen, Norway), pp. 189–193, Jul. 2007.
[10] Y. Jing and B. Hassibi, “Distributed space-time coding in wireless relay networks,” IEEE Trans. Wireless Commun., vol. 5, pp. 3524–3536, Dec. 2006.
[11] M. K. Simon and M.-S. Alouini, Digital Communication over Fading Channels: A Unified Approach to Performance Analysis. New York, NY: Wiley, 2000.
[12] M. Abramowitz and I. A. Stegun, Handbook of Mathematical Functions. New York, USA: Dover Publications, Inc., 1972.
[13] I. S. Gradshteyn and I. M. Ryzhik, Table of Integrals, Series, and Products. San Diego, CA: Academic, 1983.
[14] S. Boyd and L. Vandenberghe, Convex Optimization. Cambridge: Cambridge Univ. Press, 2004.
[15] G. Abreu, “GABBA coders: generalized full-rate orthogonally decodable space-time block codes,” in Thirty-Ninth Asilomar Conference on Signals, Systems and Computers, Nov. 2005.