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PAPER

A Study on Optimum Weights for Delay Transmit Diversity for DS-CDMA in a Frequency Non-selective Fading Channel

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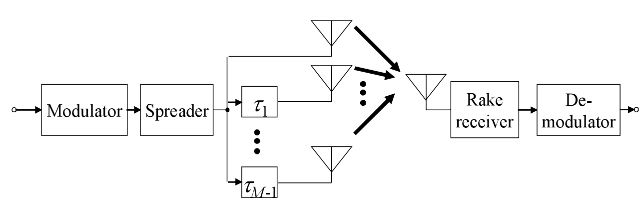
SUMMARY In this paper, we study DS-CDMA delay transmit diversity that transmits the weighted and time-delayed versions of the same signal from multiple antennas in a frequency non-selective fading environment. At a receiver, one receive antenna is used and the received delayed signals are coherently combined by Rake receiver. The set of optimum antenna weights for maximizing the received signal-to-noise power ratio (SNR) is theoretically derived to reveal that the optimum solution is to transmit only from the best antenna that has the maximum channel gain. The bit error rate (BER) performance improvement over conventional delay transmit diversity is theoretically analyzed and confirmed by computer simulations. The combined effect of transmit diversity and transmit power control (TPC) is also evaluated. Furthermore, the impact of fading decorrelation between the transmit and receive channels is also investigated for both the time division duplex (TDD) and frequency division duplex (FDD) schemes.

key words: *transmit diversity, transmit power control, fading, DS-CDMA, Rake combining*

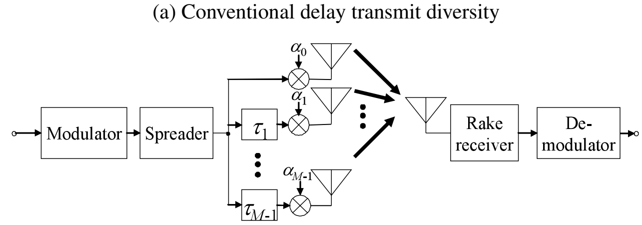
1. Introduction

Recently, direct sequence code division multiple access (DS-CDMA) [1] has been adopted for present cellular mobile communications systems because of its high spectrum efficiency [2]–[6]. In DS-CDMA mobile radio, fast transmit power control (TPC) is an indispensable technique [1], [7], [8] to alleviate the well known near-far problem as well as the adverse effect of multipath fading. Further performance improvement can be achieved by using Rake combining if the channel is frequency selective. However, in a frequency non-selective fading channel (the fading channel becomes frequency non-selective when the multipath spread is much less than the chip duration), the Rake combining effect cannot be obtained. Hence, the frequency non-selective channel can be considered to be the severest channel for DS-CDMA systems. For such a channel, antenna diversity technique is attractive. There are two types of antenna diversity technique: receive diversity and transmit diversity. The former is used in present practical mobile communications systems. However, using more than two receive antennas at a mobile receiver may not be practical. This is the motivation of this paper. In this paper, we consider the use of transmit diversity in a frequency non-selective fading channel, where multiple transmit antennas are used at a base station while a single antenna is used at a mobile station.

Various transmit diversity techniques have been proposed in the open literature [9]–[11]. Delay transmit diversity, in which multiple antennas transmit the delayed versions of the same signal is one of the conventional transmit diversity schemes [12], [13] (see Fig. 1(a)). It is shown [13] that the lower bound of the bit error rate (BER) performance achievable with this delay transmit diversity using channel matched filter is the same as that of maximum ratio combining (MRC) receive diversity, but with reduced received signal power per antenna by a factor of M (M is the number of transmit antennas). In FDMA or TDMA access method, the channel matched filter is realized by maximum likelihood sequence estimation (MLSE). In the DS-CDMA system, Rake receiver can be used as the channel matched filter if the interference among delayed signals is negligibly small (i.e., if very large spreading factor is used). However, the BER performance with conventional delay transmit diversity is still worse than MRC receive diversity by $10 \log M$ dB. In the conventional delay transmit diversity, the delayed versions of the same signal are transmitted with the same power from multiple antennas. To further improve the BER performance, this paper proposes a new delay transmit diversity, in which the delayed versions of the same signal are weighted and then transmitted from different antennas (see Fig. 1(b)). In this paper, the optimum set of antenna weights for maximizing the received signal-to-noise power ratio (SNR), thus minimizing the average BER, is theoretically derived. It will be shown that the optimum weighted delay transmit diver-



(a) Conventional delay transmit diversity



(b) Weighted delay transmit diversity

Fig. 1 Delay transmit diversity model.

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sity becomes selection combining (SC) transmit diversity. Hence, this transmit diversity is called SC transmit diversity in this paper. In addition, the average BER performance achievable by the joint use of the SC transmit diversity and fast TPC is theoretically analyzed.

The SC transmit diversity requires the transmit channel state information to determine the transmitting antenna. If the transmit and receive channels are reciprocal, the transmit channel state information can be estimated from the receive channel. In a practical system, transmit and receive channels are configured using time division duplex (TDD) or frequency division duplex (FDD). Since the time difference (the frequency difference) exists between transmit and receive channels in TDD (in FDD), the transmit channel and receive channel are not perfectly reciprocal and fading decorrelation is present. Fading decorrelation degrades the transmit diversity improvement [11]. The impact of fading decorrelation property between transmit and receive channels on the achievable BER performance is also analyzed in this paper.

The remainder of this paper is organized as follows. In Sect. 2, the proposed weighted delay transmit diversity model is presented. Section 3 derives the set of optimum antenna weights to show that the optimum weighted delay transmit diversity becomes SC transmit diversity and the BER expression is theoretically derived. In Sect. 4, the effect of the joint use of SC transmit diversity and fast TPC is analyzed. The computer simulation results are presented in Sect. 5 to compare with the theoretical analysis. In Sect. 6, the impact of decorrelation property between transmit and receive channels is analyzed and confirmed by computer simulations. Section 7 draws some conclusions.

2. Weighted Delay Transmit Diversity Model

Figure 2 illustrates the block diagram of the proposed weighted delay transmit diversity system using M transmit antennas and a single receive antenna with TPC. The most severe channel for a DS-CDMA mobile radio is a frequency non-selective fading channel since the effect of Rake combining cannot be expected. So, in this paper, frequency non-selective Rayleigh fading is assumed. Quadrature phase shift keying (QPSK) data modulation and binary PSK (BPSK) spreading modulation are assumed. The QPSK modulated signal waveform $d(t)$ is multiplied by the spreading sequence waveform $c(t)$ and then, multiplied by the power coefficient for TPC. M copies of power controlled spread signal are incurred different time delays $\{\tau_m\}$ and then, multiplied by antenna weights $\{\alpha_{m,k}\}$ for transmission from M antennas.

Without loss of generality, $\tau_0(=0) < \tau_1 < \ldots < \tau_{M-1} < T$ is assumed, where T is the QPSK symbol length. The transmit signal from the mth antenna is expressed using the equivalent lowpass representation as

$$s_m(t) = \sqrt{2S}\,d(t-\tau_m)c(t-\tau_m)\alpha_m(t-\tau_m), \tag{1}$$

where S is the transmit signal power and $d(t)$, $c(t)$, and $\alpha_m(t)$ are respectively given by

$$\begin{cases} d(t) = \displaystyle\sum_{k=-\infty}^{\infty} d_k u(t/T - k) \\ c(t) = \displaystyle\sum_{i=-\infty}^{\infty} c_i u(t/T_c - i) \\ \alpha_m(t) = \displaystyle\sum_{k=-\infty}^{\infty} \alpha_{m,k} u(t/T - k) \end{cases}. \tag{2}$$

In Eq. (2), d_k and c_i are the kth transmit QPSK symbol with $|d_k|=1$ and the ith spreading chip with $|c_i|=1$ and a period of T_c, respectively, and $u(x)$ is the unit pulse function with $u(x) = 1(0)$ if $0 \le x \le 1$ (otherwise). We assume

$$\sum_{m=0}^{M-1} \left|\alpha_{m,k}\right|^2 = 1, \tag{3}$$

so that the total power is always kept to be S.

M transmitted signals are received by the single receive antenna via independent and identically distributed (iid) frequency non-selective Rayleigh fading channels. The received signal $r(t)$ is given by

$$\begin{aligned} r(t) &= \sum_{m=0}^{M-1} s_m(t)g_m(t) + n(t) \\ &= \sum_{m=0}^{M-1} \sqrt{2S}\,d(t-\tau_m)c(t-\tau_m)\alpha_m(t-\tau_m)g_m(t) \\ &\quad + n(t), \end{aligned} \tag{4}$$

where $g_m(t)$ is the time varying complex channel gain experienced between the mth transmit antenna and the receive antenna, with $E[|g_m(t)|^2] = 1$ ($E[.]$ denotes ensemble average operation), and $n(t)$ is the additive white Gaussian

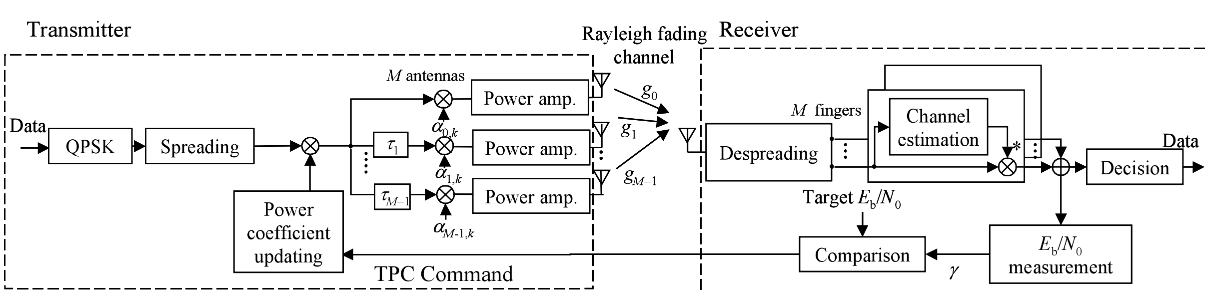


Fig. 2 Block diagram of weighted delay transmit diversity system with TPC.

noise (AWGN) with one-sided power spectrum density N_0. It should be noted that the received signal is equivalent to a signal which has been received via a frequency selective fading channel having M discrete propagation paths, each path following independent frequency non-selective fading.

3. Optimum Antenna Weights and Average BER

The received signal is resolved into M copies of the transmitted QPSK signal and coherently combined by the Rake receiver. Each Rake finger multiplies $r(t)$ with the locally generated spreading sequence waveform $c(t)$, which is time-synchronized to the time delay incurred by each transmit antenna, and integrates over one QPSK symbol duration T. In this paper, it is assumed that the receiver has a perfect knowledge of the time delays and that the receiver sampling timing is ideal. Each Rake finger output is sampled at the symbol rate. The Rake finger output $r_m(k)$ at the kth symbol time epoch, associated with the mth antenna, is given by

$$\begin{aligned} r_{m,k} &= \frac{1}{T}\int_{kT+\tau_m}^{(k+1)T+\tau_m} r(t)c(t-\tau_m)dt \\ &= \sqrt{2S}\, d_k \alpha_{m,k} g_{m,k} + n'_{m,k}, \end{aligned} \tag{5}$$

where $g_{m,k} = g_m(kT)$ and $n'_{m,k}$ is the additive white Gaussian noise (AWGN) with one-sided power spectrum density N_0. We have assumed that the channel gains $\{g_m(t)\}$ remain constant over T. In Eq. (5), we have neglected the interference among M signals with different time delays since very large spreading factor is assumed. Assuming perfect channel estimation, a total of M Rake finger outputs is coherently combined based on MRC [14], resulting in

$$\eta_k = \sum_{m=0}^{M-1} r_{m,k}(\sqrt{2S}\,\alpha_{m,k} g_{m,k})^*, \tag{6}$$

which is the decision variable for QPSK data demodulation, where * denotes complex conjugate operation. The received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0, γ_R, after Rake combining based on MRC is given by

$$\begin{aligned} \gamma_R &= \frac{1}{2}\sum_{m=0}^{M-1}\left|\sqrt{2S}\, d_k \alpha_{m,k} g_{m,k}\right|^2 \Big/ \left(\frac{2N_0}{T}\right) \\ &= \gamma_T \sum_{m=0}^{M-1} |\alpha_{m,k}|^2 |g_{m,k}|^2, \end{aligned} \tag{7}$$

where γ_T is the transmit E_b/N_0. In the following, the subscript k is dropped for the sake of simplicity.

The optimum value of α_m that maximizes γ_R is obtained below. Letting $z = \gamma_R / \gamma_T$, the maximization problem of γ_R is equivalent to maximizing

$$z = \sum_{m=0}^{M-1} c_m x_m, \tag{8}$$

where

$$\begin{cases} x_m = |\alpha_m|^2 \\ c_m = |g_m|^2 \end{cases}, \tag{9}$$

with the constraint:

$$\sum_{m=0}^{M-1} x_m = 1. \tag{10}$$

Using simplex algorithm for linear programming [15], we have (see Appendix A)

$$\begin{cases} x_m = \delta_{ml} \\ z_{\max} = c_l \\ \text{if} \quad c_l = \max\limits_{m=\{0,1,\cdots,M-1\}} (c_m) \end{cases}, \tag{11}$$

where δ_{ml} is the Kronecker's delta function [16]. As a consequence, the optimum weights $\{\hat{\alpha}_m\}$ and the maximum γ_R are given by

$$\begin{cases} \hat{\alpha}_m = \delta_{ml} \\ \gamma_R = \gamma_T |g_l|^2 \\ \text{if} \quad |g_l|^2 = \max\limits_{m=\{0,1,\cdots,M-1\}} (|g_m|^2) \end{cases}. \tag{12}$$

Equation (12) states that the weighted delay transmit diversity is equivalent to the receive diversity using SC, i.e., signal transmission is done only from one antenna that has the maximum channel gain among M antennas. Henceforth, in this paper, this transmit diversity is called SC transmit diversity. Insertion of time delays at transmitter is unnecessary and thus, the Rake combiner can be removed from the receiver. Figure 3 illustrates the SC transmit diversity model for the downlink transmission in a multi-user environment. TPC is performed by multiplying each user's spread signal by the power coefficient. The selection of transmit antenna is carried out by controlling the antenna weights. The mth antenna weight for the uth user at the kth symbol time epoch is denoted by $\alpha_{m,k}^{(u)}$. When the uth user's spread signal is to be transmitted from the m'th antenna, $\alpha_{m,k}^{(u)} = 0\,(1)$ for $m \neq m'$ ($m = m'$). The sum of all users' spread signals after multiplying both power coefficients and antenna weights is power amplified for transmission from each antenna. In the DS-CDMA downlink transmission, the product of different orthogonal spreading code (channelization code) and a common scramble code is used for transmission of different user's signal. In the frequency non-selective channel assumed in this paper, the multi-user interference is not produced when orthogonal spreading codes are used. Hence,

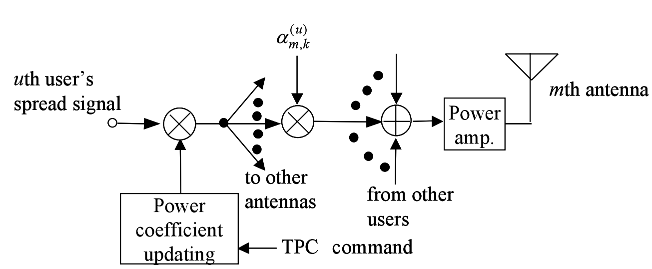


Fig. 3 SC transmit diversity for the downlink transmission in a multi-user environment.

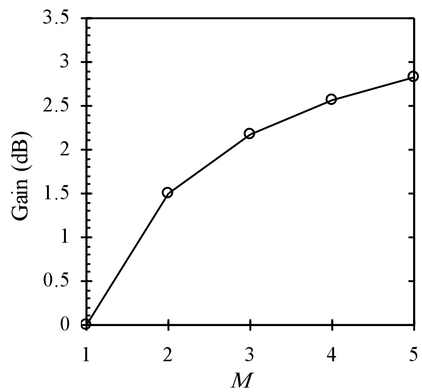


Fig. 4 Gain of SC transmit diversity from delay transmit diversity when no TPC is used.

throughout this paper, the single user case is assumed for simplicity purpose.

In the following BER analysis, it is assumed that the transmitter has a perfect knowledge of transmit channel gains. (In practice, however the transmitter needs to estimate transmit channel gains from the received signals. We will discuss in Sect. 5 the degradation due to the decorelation between transmit and receive channels.) Using Eq. (12), the average BER $P^{(M)}_{\mathrm{b,SC}}(\gamma_{\mathrm{T}})$ is obtained as [14]

$$P^{(M)}_{\mathrm{b,SC}}(\gamma_{\mathrm{T}}) = \frac{1}{2}\sum_{m=0}^{M}\binom{M}{m}\frac{(-1)^m}{\sqrt{1+m/\gamma_{\mathrm{T}}}} \approx \frac{(2M-1)!!}{2^{M+1}}\frac{1}{\gamma_{\mathrm{T}}^M}, \quad \text{for} \quad \gamma_{\mathrm{T}} \gg 1, \tag{13}$$

where

$$\binom{M}{m} = \frac{M!}{m!(M-m)!}$$

and $n!! = n(n-2)\cdots$. The average BER $P^{(M)}_{\mathrm{b,delay}}(\gamma_{\mathrm{T}})$ of the delay transmit diversity, which is equivalent to that of MRC receive diversity with the reduced received signal power per antenna by a factor of M, is given by [14]

$$P^{(M)}_{\mathrm{b,delay}}(\gamma_{\mathrm{T}}) = \frac{1}{2}\left[1 - \frac{1}{\sqrt{1+\frac{M}{\gamma_{\mathrm{T}}}}}\sum_{m=0}^{M-1}\binom{2m}{m}\frac{1}{\left\{4\left(1+\frac{\gamma_{\mathrm{T}}}{M}\right)\right\}^m}\right] \approx \frac{M^M}{2}\frac{(2M-1)!!}{(2M)!!}\frac{1}{\gamma_{\mathrm{T}}^M}, \quad \text{for } \gamma_{\mathrm{T}} \gg 1. \tag{14}$$

It can be shown from Eqs. (13) and (14) that the γ_{T} required for a certain BER with SC transmit diversity can be reduced by about $10\log\{M/(M!)^{1/M}\}$ dB from delay transmit diversity. The gain of SC transmit diversity from delay transmit diversity is plotted in Fig. 4. For example, the gain is 1.5 dB, 2.2 dB and 2.6 dB when M=2, 3 and 4, respectively.

4. Joint Use of Transmit Diversity and Fast TPC

The joint use of transmit diversity and fast TPC is considered. As mentioned earlier, the transmitter is assumed to have a perfect knowledge of transmit channel gains. Also assumed is the perfect measurement of the received E_b/N_0, γ_{R}, at the receiver. As shown in Fig. 2, the receiver measures γ_{R} and compares with the target γ_{tpc}. TPC command is sent to the transmitter to raise (lower) its transmit power S if $\gamma_{\mathrm{R}} < \gamma_{\mathrm{tpc}}$ (otherwise) by the limited amount of Δ_{tpc} dB (step size) [4], [5]. Ideal fast TPC is assumed so that transmit power S is controlled to keep $\gamma_{\mathrm{R}} = \gamma_{\mathrm{tpc}}$. When fast TPC is used, the transmit E_b/N_0, γ_{T}, varies. It can be obtained from Eqs. (7) and (12) for SC transmit diversity. For delay transmit diversity, it can be obtained from the fact that the delay transmit diversity is equivalent to MRC receive diversity with the reduced received signal power per antenna by a factor of M. We have

$$\gamma_{\mathrm{T}} = S\left(\frac{T}{2N_0}\right) = \begin{cases} \gamma_{\mathrm{tpc}}\big/|g_l|^2, & \text{for SC transmit diversity} \\ \gamma_{\mathrm{tpc}}\Big/\left(\frac{1}{M}\sum_{m=0}^{M-1}|g_m|^2\right), & \text{for delay transmit diversity} \end{cases}. \tag{15}$$

The probability density function (pdf) $p(x)$ of $x = \gamma_{\mathrm{T}}/\gamma_{\mathrm{tpc}}$ is expressed as (see Appendix B)

$$p(x) = \begin{cases} \frac{M}{x^2}\exp\left(-\frac{1}{x}\right)\left\{1-\exp\left(-\frac{1}{x}\right)\right\}^{M-1}, \\ \text{for SC transmit diversity} \\ \frac{(M/x)^M}{(M-1)!x}\exp\left(-\frac{M}{x}\right), \\ \text{for delay transmit diversity} \end{cases}. \tag{16}$$

Using Eq. (16), $E[\gamma_{\mathrm{T}}] = \gamma_{\mathrm{tpc}}E[x]$ is given by

$$E[\gamma_{\mathrm{T}}] = \gamma_{\mathrm{tpc}}\int_0^{\infty} xp(x)dx = \begin{cases} \infty, & \text{if } M = 1 \\ \begin{cases} \gamma_{\mathrm{tpc}}A(M), & \text{SC transmit diversity} \\ \gamma_{\mathrm{tpc}}\frac{M}{M-1}, & \text{delay transmit diversity} \end{cases}, & \text{if } M \geq 2 \end{cases} \tag{17}$$

where

$$A(M) = \begin{cases} M\ln\left[\dfrac{2^{\binom{M-1}{1}}4^{\binom{M-1}{3}}\cdots M^{\binom{M-1}{M-1}}}{1^{\binom{M-1}{0}}3^{\binom{M-1}{2}}\cdots(M-1)^{\binom{M-1}{M-2}}}\right], \\ \text{for } M = 2, 4, \cdots \\ M\ln\left[\dfrac{2^{\binom{M-1}{1}}4^{\binom{M-1}{3}}\cdots(M-1)^{\binom{M-1}{M-2}}}{1^{\binom{M-1}{0}}3^{\binom{M-1}{2}}\cdots(M-2)^{\binom{M-1}{M-3}}M^{\binom{M-1}{M-1}}}\right], \\ \text{for } M = 3, 5, \cdots \end{cases}. \tag{18}$$

Since γ_{R} is kept at the target value γ_{tpc}, the BER $P_b(\gamma_{\mathrm{tpc}})$ is given by [14]

$$P_b(\gamma_{tpc}) = \frac{1}{2}\mathrm{erfc}\sqrt{\gamma_{tpc}}, \tag{19}$$

where

$$\mathrm{erfc}(y) = (2/\sqrt{\pi})\int_y^{\infty}\exp(-t^2)dt$$

is the complimentary error function. Using Eqs. (17), (18) and (19), the average BER performance with the joint use of transmit diversity and ideal fast TPC can be obtained as a function of the average γ_T i.e., $E[\gamma_T]$. It can be shown from Eqs. (17) and (18) that the average γ_T required for a certain BER with the joint use of SC transmit diversity and ideal fast TPC can be reduced by $10\log[M/\{(M-1)A(M)\}]$ dB from that of the delay transmit diversity. This is plotted in Fig. 5. It is seen that the gain is 1.6 dB, 2.4 dB and 2.9 dB when M=2, 3 and 4, respectively. Similar to the case of no TPC, the gain of SC transmit diversity over the delay transmit diversity increases as the number of transmit antennas increases.

So far, we have assumed the single user case. When transmit diversity with TPC is applied to the downlink transmission, the TPC operation is performed independently for each user. In delay transmit diversity, transmission of each user's spread signal from M antennas after incurring different time delays produces the multi-user interference (MUI), in addition to the self interference, due to partial destruction of code orthogonality among different users. Additional use of TPC enhances the MUI since large power differences among users may be produced and hence, the transmission performance may degrade. On the other hand, in SC transmit diversity, no MUI is produced even when TPC is used since the best antenna is always selected for transmission. However, this is true only in a frequency non-selective channel. In the case of frequency-selective channel, Rake combining can be applied and combined with SC transmit diversity; the best transmit antenna that has the maximum equivalent channel gain after Rake combining (or the maximum SNR after Rake combining) is selected for transmission [17].

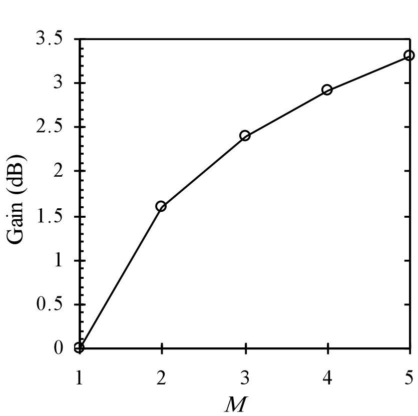


Fig. 5 Gain of SC transmit diversity from delay transmit diversity when fast TPC is used.

5. Computer Simulation

Computer simulation parameters are listed in Table 1. The TPC command period T_{tpc} and step size Δ_{tpc} are T_{tpc}=64 symbols and Δ_{tpc}=1 dB, respectively. Since the transmit power is updated only by the limited amount of Δ_{tpc}=1 dB every T_{tpc}=64 symbols, the fast and deep power drops due to fading cannot be fully regulated and hence the average BER performance may degrade compared to the ideal TPC case. Perfect knowledge of transmit channel gains is assumed in this section. We will investigate how the average BER performance degrades from the ideal one. Since the single user case is assumed in this paper, the orthogonal spreading code (channelization code) is removed and only the scramble code (a long pseudo-noise (PN) sequence of $2^{12}-1$ chips) is used in the computer simulation.

Figure 6 plots the theoretical and simulated average BER performances without TPC as a function of the transmit γ_T. It is seen that the simulated values are the same

Table 1 Simulation parameters.

Propagation channel		Frequency non-selective Rayleigh fading
Data modem.	Modulation	QPSK
	Demodulation	Ideal coherent detection
Spreading modulation	Modulation	BPSK
	Spreading factor	64
	Spreading code	PN sequence of 2^{12}-1 chips
TPC	Period	T_{tpc} =64 symbols
	Step size	Δ_{tpc} =1 dB
	Delay	T_{tpc}
Transmit diversity	Channel estimation	Ideal
	Diversity scheme	Ideal SC

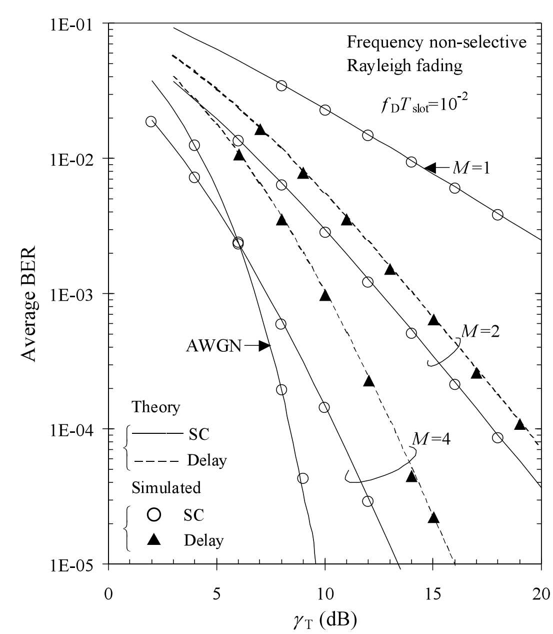
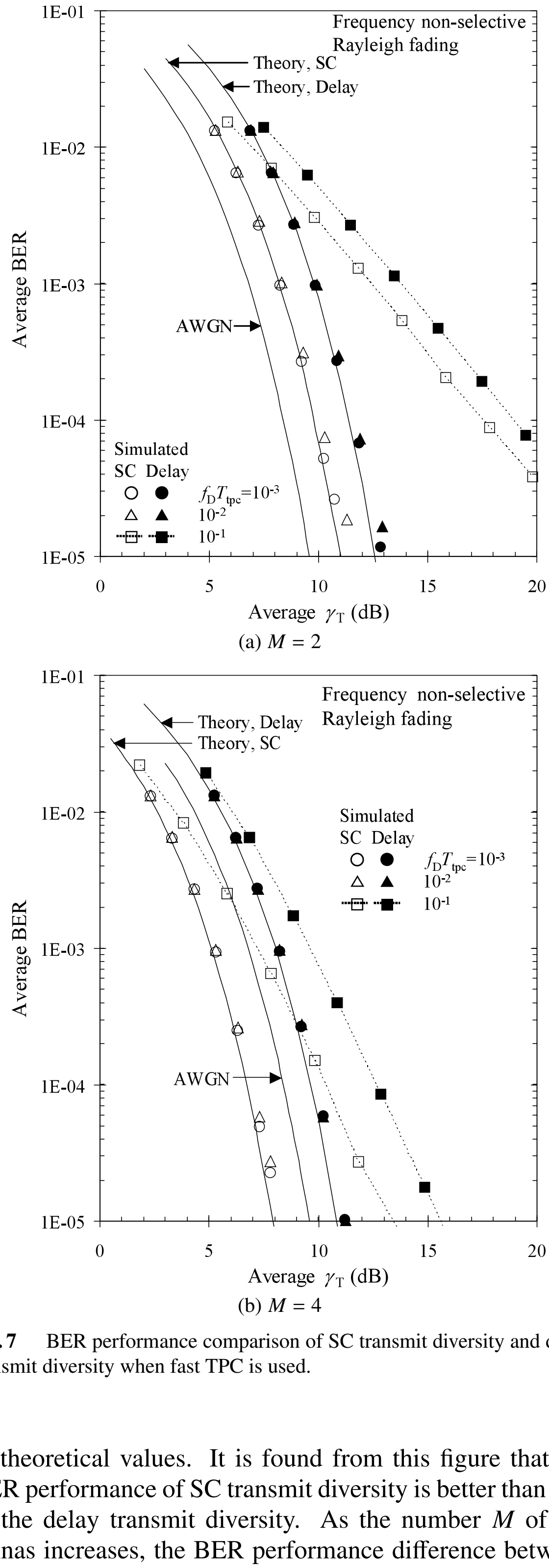


Fig. 6 BER performance comparison of SC transmit diversity and delay transmit diversity when no fast TPC is used.



(a) $M = 2$

(b) $M = 4$

Fig. 7 BER performance comparison of SC transmit diversity and delay transmit diversity when fast TPC is used.

as theoretical values. It is found from this figure that the BER performance of SC transmit diversity is better than that of the delay transmit diversity. As the number M of antennas increases, the BER performance difference between SC transmit diversity and delay transmit diversity becomes larger. The gain of SC transmit diversity in the required average γ_T over delay transmit diversity for BER=10^{-3} is close to the theoretical gain, i.e., 1.5 dB with M=2 and 2.7 dB with M=4.

Figure 7 plots the theoretical and simulated average BER performances with fast TPC as a function of the average transmit γ_T with the maximum Doppler frequency $f_D T_{tpc}$ as a parameter. Similar to the case without TPC, the SC transmit diversity can achieve better performance than the delay transmit diversity. As the number M of antennas increases, the BER performance difference between SC transmit diversity and the delay transmit diversity becomes larger. When $f_D T_{tpc} \leq 10^{-2}$, the gain of SC transmit diversity in the required average γ_T for BER=10^{-3} is close to the theoretical gain, i.e., 1.8 dB with M=2 and 2.9 dB with M=4. It is seen that when $f_D T_{tpc}$=10^{-1}, the simulated BER performances of both diversity schemes are worse than the ideal TPC case. This is because TPC cannot track the fast channel gain variations.

6. Impact of Transmit and Receive Channels Decorrelation

In SC transmit diversity estimation of the transmit channel gain is necessary to determine the transmit antenna, while it is not necessary in the delay transmit diversity. In the previous section, perfect knowledge of transmit channel gains was assumed. However, in practical systems, the channel gains must be estimated. Assuming the reciprocity between transmit and receive channels, the channel gains can be estimated from the receive channels. Transmit and receive channels are configured using TDD or FDD. Since there is time difference between transmit and receive channels in TDD and frequency difference in FDD, transmit and receive channels are not perfectly reciprocal. Fading decorrelation between transmit and receive channels degrades the SC transmit diversity improvement. In this section, this issue is addressed.

6.1 Decorrelation Property

6.1.1 TDD

Figure 8(a) illustrates the TDD frame structure with the frame length T_f and the slot length $T_{slot} = T_f/2$. TPC delay is $T_{tpc} = T_f = 2T_{slot}$. (Note that, for TDD, an open loop TPC can be implemented at the transmitter side and the TPC delay can be $T_f/2$. However, this is not considered.) In [11], pilot symbols are inserted in a frame to estimate the channel gain by linear prediction. However, in this paper, in order to evaluate the degradation caused by the fading decorrelation, simple frame structure is assumed. Pilot symbols are inserted at the beginning of each frame as illustrated in Fig. 8(a). It is assumed that M receive channel gains can be perfectly estimated using known pilot symbols. The estimated receive channel gains $\{g_m(kT_{slot})\}$ are used to determine the transmit channel gains for the next transmit slot. However, the real channel gains become $\{g_m((k + 0.5)T_{slot})\}$ at the beginning of the transmit slot. Assuming Rayleigh fading channel encountered in the mobile radio, the correlation ρ_{TDD} between the channel gains $g_m(kT_{slot})$ and $g_m((k + 0.5)T_{slot})$ is given by [14]

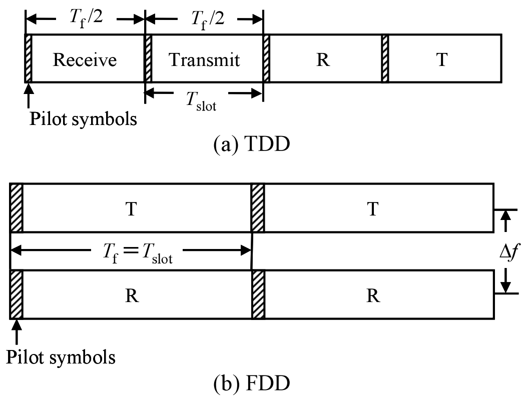


Fig. 8 Frame structure. Pilot symbols are inserted for channel estimation.

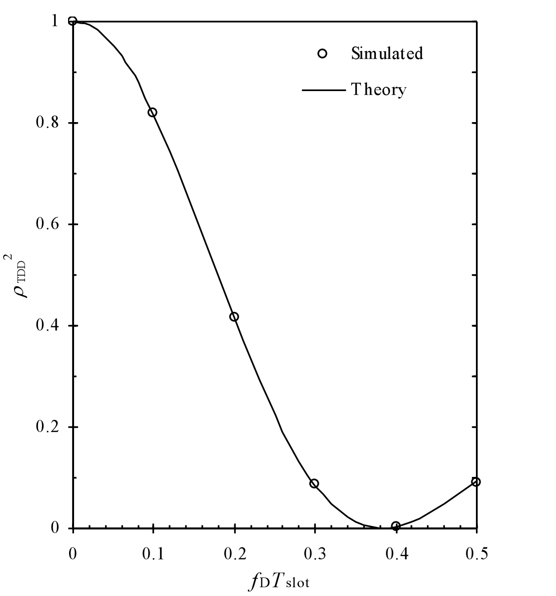


Fig. 9 Decorrelation property between transmit and receive channels for TDD.

$$\rho_{TDD} = J_0(\pi f_D T_{slot}), \tag{20}$$

where $J_0(.)$ is a 0th order Bessel function of the 1st kind. Figure 9 shows the decorrelation property as a function of $f_D T_{slot}$. Assuming that $|\rho_{TDD}| > 0.9$ needs to be satisfied, the value of $f_D T_{slot}$ must be smaller than 0.1 (when T_{slot}=1 ms, f_D must be smaller than 100 Hz). This decorrelation property also affects the fast TPC.

6.1.2 FDD

Figure 8(b) illustrates the FDD frame structure with the frame length T_f and the slot length $T_{slot} = T_f$. TPC delay is $T_{tpc} = T_f = T_{slot}$. The frequency separation between uplink and downlink is denoted as Δf. For the performance analysis, we need to model the channel so that the channel is frequency non-selective within the signal transmit bandwidth of less than $1/T_c$, but is frequency selective between transmit and receive channels. An L-path Rayleigh fading

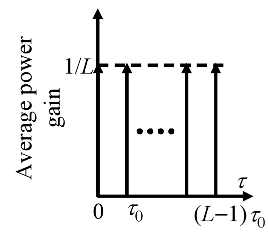


Fig. 10 Power delay profile.

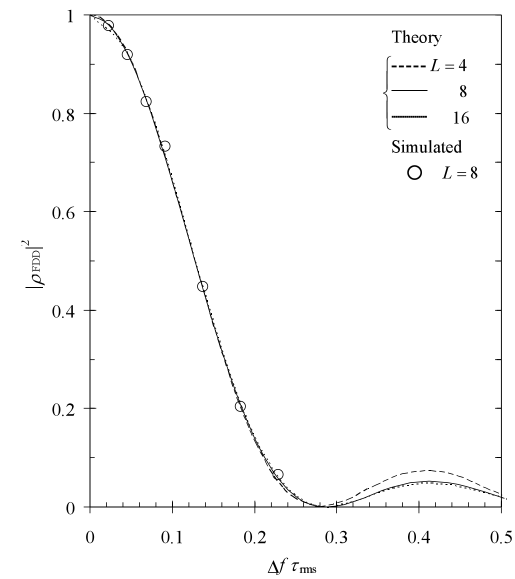


Fig. 11 Decorrelation property between transmit and receive channels for FDD.

with uniform power delay profile is assumed (see Fig. 10). The maximum time delay is assumed to be $(L-1)\tau_0$, where τ_0 much smaller than T_c. The channel impulse response $h(t, \tau)$ is given by

$$h(t,\tau) = \sum_{l=0}^{L-1} g^{(l)}(t)\delta(\tau - l\tau_0), \tag{21}$$

where $g^{(l)}(t)$, $l = 0 \sim L-1$, is the lth path gain. The channel transfer function $H(t, f)$ is given by

$$H(t, f) = \sum_{l=0}^{L-1} g^{(l)}(t)\exp(-j2\pi f l\tau_0), \tag{22}$$

and the rms delay spread τ_{rms} of the channel is given by

$$\tau_{rms} = \frac{\tau_0}{2}\sqrt{\frac{L^2-1}{3}}. \tag{23}$$

The correlation ρ_{FDD} between transmit and receive channels can be found from Eq. (22) and is given by

$$\begin{aligned}\rho_{FDD} &= E[H(t,0)H^*(t,\Delta f)] \\ &= \frac{1}{L}\frac{\sin\left(\frac{2\sqrt{3}\pi L}{\sqrt{L^2-1}}\Delta f\tau_{rms}\right)}{\sin\left(\frac{2\sqrt{3}\pi}{\sqrt{L^2-1}}\Delta f\tau_{rms}\right)}\exp\left(j2\sqrt{3}\pi\sqrt{\frac{L-1}{L+1}}\Delta f\tau_{rms}\right).\end{aligned} \tag{24}$$

Figure 11 plots $|\rho_{FDD}|^2$ as a function of $\Delta f\tau_{rms}$ with L as a parameter.

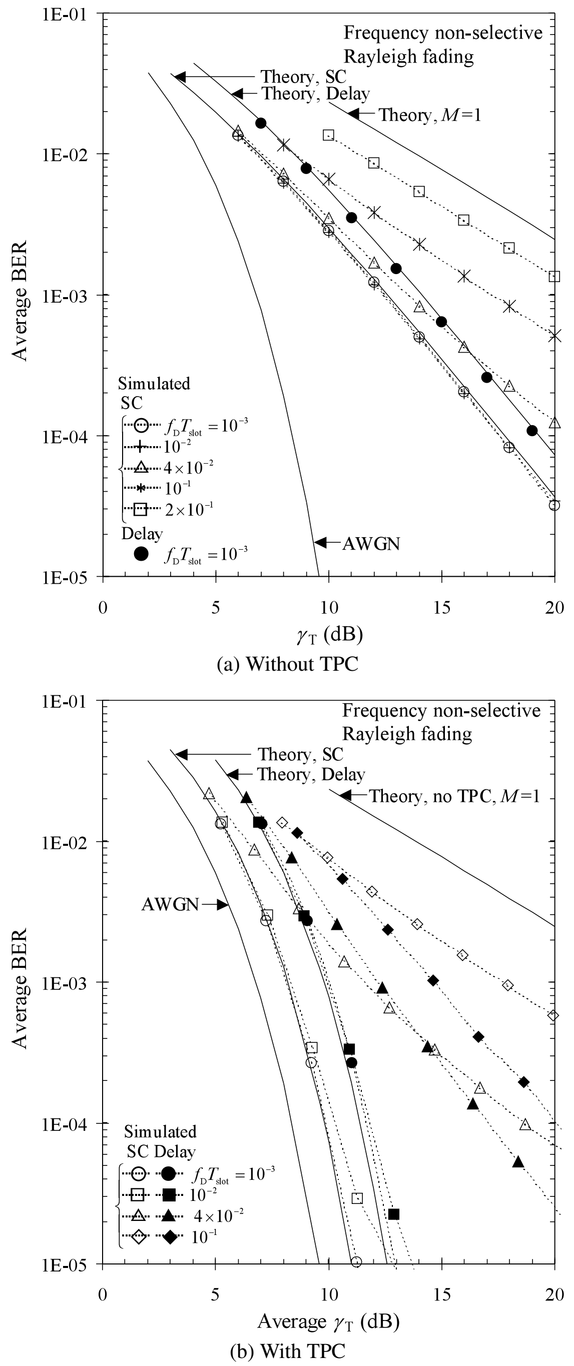


Fig. 12 BER performance with transmit diversity in TDD. M = 2.

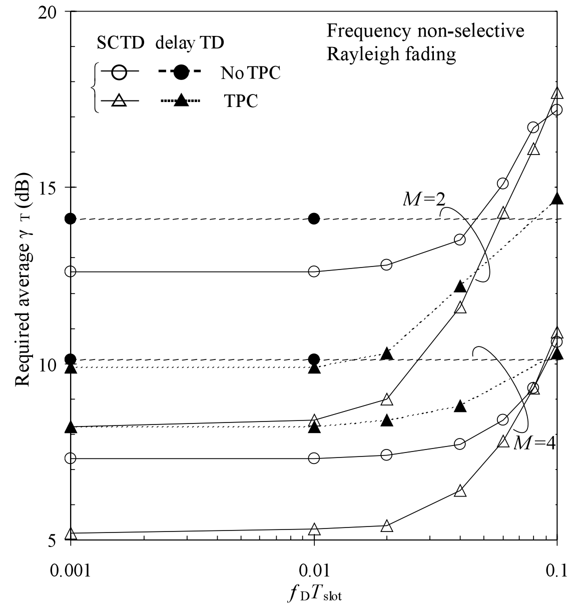


Fig. 13 Required average γ_T for BER=10^{-3} in TDD.

6.2 BER Performance

The impact of the decorrelation between receive and transmit channels on the BER performance is evaluated by computer simulations. Simulation conditions are the same as in Sect. 5.

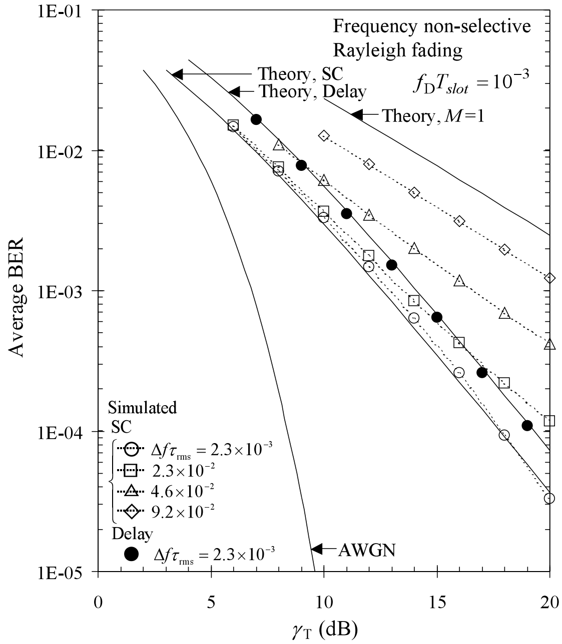
6.2.1 TDD

Figure 12 shows the BER performance of SC transmit diversity with $f_D T_{slot}$ as a parameter for M=2. When $f_D T_{slot}$ is smaller than 0.04, the required γ_T for BER=10^{-3} of SC transmit diversity is smaller than that of delay transmit diversity. When no TPC is used, as the value of $f_D T_{slot}$ increases, the BER performance degrades since the correlation ρ_{TDD} between the transmit and receive channels becomes smaller (see Fig. 9). Note that the delay transmit diversity is not affected by the decorrelation at all since no channel state information is involved in diversity transmission and hence, only the results for $f_D T_{slot}$=10^{-3} are plotted. However, the TPC command is generated at the receiver and sent back to the transmitter and hence, the BER performance of delay transmit diversity with TPC is also affected by the time difference between transmit and receive channels. Therefore, when TPC is used, the BER performances of both SC and delay transmit diversity degrade as $f_D T_{slot}$ increases.

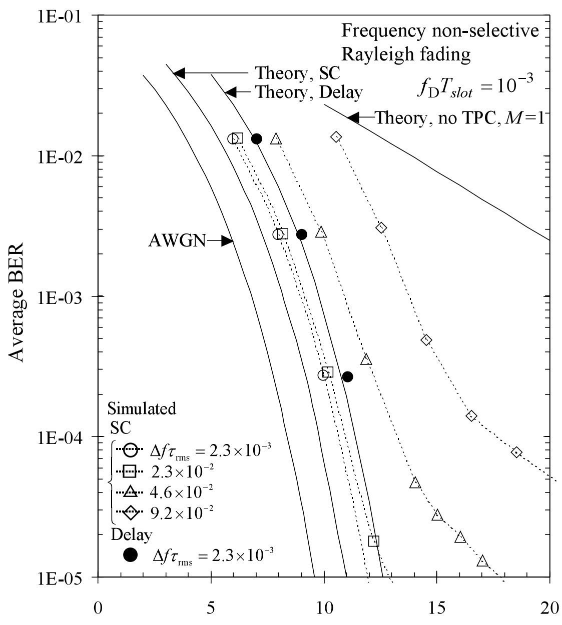
Figure 13 shows the required average γ_T for BER=10^{-3} as a function of $f_D T_{slot}$ With M=2 (4), the required average γ_T of the SC transmit diversity is smaller than that of delay transmit diversity when $f_D T_{slot}$ is less than about 0.04 (0.09). Assuming the carrier frequency of 2 GHz and T_{slot}=1 ms, $f_D T_{slot}$ 0.04 (0.09) corresponds to the mobile speed of 20 km/h (45 km/h).

6.2.2 FDD

Figure 14 plots the BER performance of SC transmit diversity with $\Delta f \tau_{rms}$ as a parameter when M=2. In order to evaluate the impact of $\Delta f \tau_{rms}$, very slow fading ($f_D T_{slot}$ = 10^{-3}) is assumed so that the degradation due to fading variations in time can be neglected. It is found from this figure that as $\Delta f \tau_{rms}$ increases, the BER performance of SC transmit diversity degrades and approaches that with M=1, since $|\rho_{FDD}|$ decreases. Note that the delay transmit diversity is not af-



(a) Without TPC



(b) With TPC

Fig. 14 BER performance with transmit diversity in FDD. $f_D T_{\text{slot}} = 10^{-3}$ and $M = 2$.

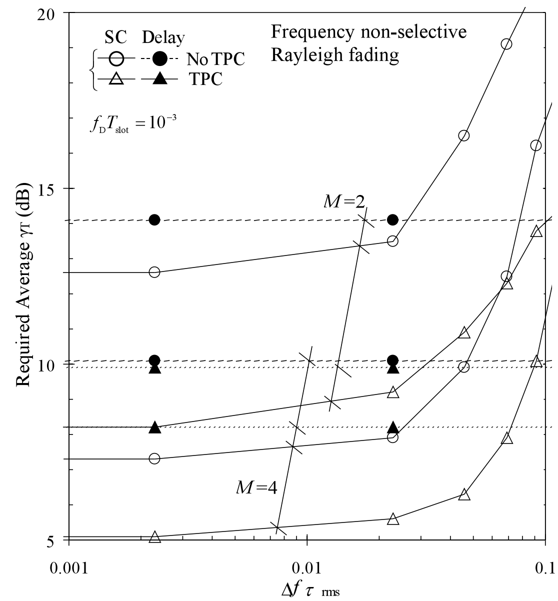


Fig. 15 Required average γ_T for BER=10^{-3} in FDD.

fected by the decorrelation at all since no channel state information is involved in diversity transmission and hence, only the results for $\Delta f\tau_{\text{rms}} = 2.3 \times 10^{-3}$ are plotted. (Of course, as stated in Sect. 6.2.1, the BER performances of both SC transmit diversity and delay transmit diversity are affected by the time difference between the transmit and receive channels when TPC is used. In Figs. 14 and 15, $f_D T_{\text{slot}} = 10^{-3}$ is assumed.)

Figure 15 shows the required average γ_T for BER=10^{-3} as a function of $\Delta f\tau_{\text{rms}}$. When M=2 (4), the required average γ_T of SC transmit diversity is smaller than that of delay diversity when $\Delta f\tau_{\text{rms}}$ is smaller than about 0.03 both for TPC and no TPC (about 0.05 without TPC, while about 0.07 with TPC). If the rms delay spread of the propagation channel is τ_{rms}=1 μs, the minimum required frequency separation Δf between the uplink and downlink becomes 30 kHz when M=2. This implies that the application of SC transmit diversity does not seem to be practical in FDD.

7. Conclusions

In this paper, we proposed weighted delay transmit diversity for DS-CDMA mobile radio. The optimum antenna weights for maximizing the received SNR, thus minimizing the average BER, were theoretically derived to reveal that the optimum solution is to transmit only from the best antenna that has the maximum channel gain (hence, this weighted delay transmit diversity has been called SC transmit diversity). The joint effect of SC transmit diversity and TPC was theoretically analyzed. The BER performance of SC transmit diversity in a frequency non-selective Rayleigh fading environment was compared to that of delay transmit diversity by computer simulations. The transmit antenna needs to be determined from the receive channels. However, since there is time difference between transmit and receive channels in TDD and frequency difference in FDD, transmit and receive channels are not perfectly reciprocal. Therefore, the impact of fading decorrelation between receive and transmit channels was evaluated for TDD and FDD. It was found that when M=2, SC transmit diversity still provides superior BER performance to delay transmit diversity when $f_D T_{\text{slot}}$ is less than about 0.04 in TDD and when $\Delta f\tau_{\text{rms}}$ smaller than about 0.03 in FDD.

References

[1] A.J. Viterbi, CDMA: Principles of spread spectrum communications, Addison-Wesley, 1995.
[2] F. Adachi, M. Sawahashi, and H. Suda, "Wideband DS-CDMA for next generation mobile communications systems," IEEE Commun. Mag., vol.36, no.9, pp.56–69, Sept. 1998.
[3] M. Sawahashi, K. Higuchi, S. Tanaka, and F. Adachi, "Enhanced wireless access technologies and experiments for W-CDMA communications," IEEE Pers. Commun., vol.7, no.6, pp.6–16, Dec. 2000.
[4] R. Kohno, R. Meiden, and L. Milstein, "Spread spectrum access methods for wireless communications," IEEE Commun. Mag., vol.33, no.1, pp.58–67, Jan. 1995.
[5] K. Gilhausen, et al., "On the capacity of a cellular CDMA system," IEEE Trans. Veh. Technol., vol.40, no.2, pp.303–312, May 1991.
[6] C. Kchao and G.L. Stuber, "Analysis of a direct-sequence spread-spectrum cellular radio system," IEEE Trans. Commun., vol.41, no.10, pp.1507–1516, Oct. 1993.
[7] S. Ariyavistakul, "Signal and interference statistics of a CDMA system with feedback power control—PartII," IEEE Trans. Commun., vol.42, no.234, pp.597–605, Feb./March/April 1994.
[8] S. Seo, T. Dohi, and F. Adachi, "SIR-based transmit power control of reverse link for coherent DS-CDMA mobile radio," IEICE Trans. Commun., vol.E81-B, no.7, pp.1508–1516, July 1998.
[9] T. Lo, "Maximum ratio transmission," IEEE Trans. Commun., vol.47, no.10, pp.1458–1461, Oct. 1999.
[10] V. Weerackody, "Diversity for the direct-sequence sqread spectrum system using multiple transmit antennas," Proc. Int. Commun. Conf., pp.1775–1779, May 1993.
[11] J.K. Cavers, "Single-user and multiuser adaptive maximal ratio transmission for Rayleigh channels," IEEE Trans. Veh. Technol., vol.49, no.11, pp.2043–2050, Nov. 2000.
[12] R.T. Derryberry, S.D. Gray, D.M. Ionescu, G. Mandyam, and B. Raghothaman, "Transmit diversity in 3G CDMA systems," IEEE Commun. Mag., vol.40, no.4, pp.68–75, April 2002.
[13] J.H. Winters, "The diversity gain of transmit diversity in wireless systems with Rayleigh fading," IEEE Trans. Veh. Technol., vol.47, no.1, pp.119–123, Feb. 1998.
[14] J.G. Proakis, Digital communications, 3rd ed., McGraw-Hill, 1995.
[15] F.A. Ficken, The simplex method of linear programming, McGraw-Hill, New York, 1995.
[16] G. Stephenson, An introduction to matrices, sets and groups for science students, Dover Publications, New York, 1965.
[17] A. Katoh, E. Kudoh, and F. Adachi, "The optimum weights of multi-antenna transmit and receive diversity in a frequency selective fading environment for DS-CDMA," IEICE Technical Report, RCS 2002-206, Nov. 2002.

Appendix A: Optimum Transmit Weights

From Eqs.(9) and (10), the constraint for $\{x_m\}$ is re-stated as

$$\begin{cases} x_0 + x_1 + \cdots + x_{M-1} = 1 \\ x_m \geq 0, \quad \text{for } m = 0, 1, \cdots, M-1 \end{cases} . \tag{A· 1}$$

The objective function z is given by

$$\begin{cases} z = c_0 x_0 + c_1 x_1 + \cdots + c_{M-1} x_{M-1} \\ c_0 > c_1 > \cdots > c_{M-1} > 0 \end{cases} . \tag{A· 2}$$

Simplex algorithm [15] for linear programming is applied to find the maximum value of z. Firstly, x_0 is determined that satisfies $c_0 = \max_{m=0,1,\cdots,M-1}\{c_m\}$. Replacing the right hand side of Eq. (A· 1) by b'_0 $(= 1 \geq 0)$, Eq. (A· 1) becomes

$$x_0 = b'_0 - (x_1 + x_2 + \cdots + x_{M-1}). \tag{A· 3}$$

Substituting Eqs. (A· 3) into (A· 2), we obtain

$$\begin{aligned} z = {} & c_0(b'_0 - x_1 - x_2 - \cdots - x_{M-1}) \\ & + c_1 x_1 + c_2 x_2 + \cdots + c_{M-1} x_{M-1}, \end{aligned} \tag{A· 4}$$

from which we have

$$z - z'_0 = c'_1 x_1 + c'_2 x_2 + \cdots + c'_{M-1} x_{M-1}, \tag{A· 5}$$

where $z'_0 = c_0 b'_0$ and $c'_n = c_n - c_0$, n=1, 2,..., $M-1$. Supposing that $c_0 = \max_{m=\{0,1,\cdots,M-1\}}(c_m)$, we have $c'_n < 0$. Using the simplex algorithm, z is maximized only when $x_0 = b'_0$=1 and x_n=0 and $z_{max} = z'_0 = b'_0 c_0 = c_0$. Therefore, the maximum z and optimum x_m are given by

$$\begin{cases} x_m = \delta_{ml} \\ z_{\max} = c_l \\ \text{if} \quad c_l = \max_{m=\{0,1,\cdots,M-1\}}(c_m) \end{cases} , \tag{A· 6}$$

where δ_{ml} is Kronecker's delta function [16].

Appendix B: PDF of Transmit E_b/N_0 in SC Transmit Diversity

The equivalent channel gain h when SC transmit diversity is used can be expressed as

$$h = \max_{m=\{0,1,\cdots M-1\}}(|g_m|^2), \tag{A· 7}$$

the pdf of which is given by [14]

$$p(h) = M \exp(-h)\{1 - \exp(-h)\}^{M-1}. \tag{A· 8}$$

When ideal TPC is assumed, i.e., $\gamma_{\text{tpc}} = \gamma_{\text{R}}$, and γ_{R} of Eq. (12) is rewritten as

$$\gamma_{\text{R}} = \gamma_{\text{T}} h. \tag{A· 9}$$

Hence $x = \gamma_{\text{T}}/\gamma_{\text{tpc}}$ is expressed as x=1/h. The Jacobian J for variable transformation to obtain the pdf of $x = \gamma_{\text{T}}/\gamma_{\text{tpc}}$ is $J = -1/x^2$. Therefore, the pdf $p(x)$ of $x = \gamma_{\text{T}}/\gamma_{\text{tpc}}$ becomes

$$p(x) = \frac{M}{x^2}\exp\left(-\frac{1}{x}\right)\left\{1 - \exp\left(-\frac{1}{x}\right)\right\}^{M-1}. \tag{A· 10}$$

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