

PAPER

Performance Comparison of Delay Transmit Diversity and Frequency-Domain Space-Time Coded Transmit Diversity for Orthogonal Multicode DS-CDMA Signal Reception Using Frequency-Domain Equalization

Takeshi ITAGAKI^{†a)}, Kazuaki TAKEDA[†], *Student Members*, and Fumiyuki ADACHI[†], *Member*

SUMMARY In a severe frequency-selective fading channel, the bit error rate (BER) performance of orthogonal multicode DS-CDMA is severely degraded since the orthogonality property of spreading codes is partially lost. The frequency-selectivity of a fading channel can be exploited by using frequency-domain equalization to improve the BER performance. Further performance improvement can be obtained by using transmit diversity. In this paper, joint transmit diversity and frequency-domain equalization is presented for the reception of orthogonal multicode DS-CDMA signals in a frequency-selective fading channel. As for transmit diversity, delay transmit diversity (DTD) and frequency-domain space-time transmit diversity (STTD) are considered. The achievable BER performance of multicode DS-CDMA in a frequency-selective Rayleigh fading channel is evaluated by computer simulation. It is shown that the frequency-domain STTD significantly improves the BER performance irrespective of the degree of the channel frequency-selectivity while DTD is useful only for a weak frequency-selective channel.

key words: multicode DS-CDMA, frequency-domain equalization, MMSE, transmit diversity

1. Introduction

In mobile radio communications systems, bit error rate (BER) performance of high-speed data transmission may be severely degraded due to frequency-selective multipath fading resulting from the presence of many propagation paths with different time delays [1]. Direct sequence code division multiple access (DS-CDMA) can exploit the channel frequency-selectivity to improve the transmission performance by using rake combining [2]. DS-CDMA provides flexible data transmission in wide range of data rates simply by changing the spreading factor or using orthogonal code-multiplexing technique. For multirate and multiconnection data transmission, orthogonal multicode DS-CDMA is promising [3], [4]. In a severe frequency-selective fading channel, however, its BER performance with coherent rake combining severely degrades since the orthogonality property of spreading codes is partially lost.

Recently, multi-carrier CDMA (MC-CDMA) has been attracting much attention for high-speed data transmissions [5], [6]. In MC-CDMA, frequency-domain spreading is

used unlike DS-CDMA (in which time-domain spreading is used). The orthogonality among spreading codes can be restored while obtaining the frequency diversity effect by simple one-tap frequency-domain equalization [7]–[9]. It has been found [8] that MC-CDMA with frequency-domain equalization can provide a significantly better BER performance than DS-CDMA with coherent rake combining. Recently, application of such frequency-domain equalization to single-carrier (SC) transmission and DS-CDMA has been attracting much attention [10]–[13]. It is shown in [11] that multicode DS-CDMA with frequency-domain minimum mean square error (MMSE) equalization can achieve a BER performance almost identical to MC-CDMA and that the BER performance improves with the increase in the frequency-selectivity of the fading channel (i.e. the number of resolvable paths increases). In MC-CDMA, additional performance improvement can be obtained by the joint use of antenna receive diversity and frequency-domain MMSE equalization [9]. For the downlink case, multiple receive antennas are necessary at a mobile terminal. Recently, transmit diversity is under extensive study for alleviating the complexity problem at the mobile terminals [14].

In this paper, joint transmit diversity and frequency-domain equalization is considered for the orthogonal multicode DS-CDMA signal reception. As for transmit diversity, delay transmit diversity (DTD) and frequency-domain space-time coded transmit diversity (STTD) are considered. Alamouti's STTD [15] for frequency-nonselective fading channel is extended to the case of frequency-domain equalization in a frequency-selective fading channel. The achievable BER performances of multicode DS-CDMA with joint frequency-domain equalization and transmit diversity are evaluated by computer simulation.

According to the best of authors' knowledge, there has been no literature that discusses and compares the BER performances of DS-CDMA with frequency-domain equalization achievable with DTD and STTD. In [16], the BER performance improvement achievable with joint use of frequency-domain equalization and STTD in DS-CDMA is discussed; however, no performance comparison of STTD and DTD is presented. Performance comparison of STTD and DTD can be found in [17], but it is for the case of DS-CDMA with rake combining and furthermore, com-

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[†]The authors are with the Department of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University, Sendai-shi, 980-8579 Japan.

a) E-mail: itagaki@mobile.ecei.tohoku.ac.jp

parison is made only in terms of the achievable signal-to-interference plus noise power ratio (SINR). In our paper, however, performance comparison of DTD and STTD is given for DS-CDMA with frequency-domain equalization. Also shown as a reference in this paper are the BER performance results of DTD and STTD for DS-CDMA with rake combining. Note that it is possible to apply DTD and STTD to MC-CDMA. The simulation result of BER performance of MC-CDMA with joint frequency-domain equalization and STTD is presented in [18], [19], but DTD is not considered.

The remainder of this paper is organized as follows. In Sect. 2, the transmission system model of orthogonal multicode DS-CDMA with frequency-domain equalization is described. Joint frequency-domain equalization and transmit diversity for the reception of orthogonal multicode DS-CDMA signals is presented in Sect.3 and the operation principles of DTD and frequency-domain STTD are given. In Sect. 4, BER performance improvements obtainable by DTD and STTD are evaluated by computer simulation and are compared. Some conclusions and future work are offered in Sect. 5.

2. Orthogonal Multicode DS-CDMA Transmission with Frequency-Domain Equalization

Throughout the paper, the chip-spaced discrete time representation is used. Transmission system model of orthogonal multicode DS-CDMA with frequency-domain equalization is illustrated in Fig. 1. At the transmitter, binary data sequence is transformed into the data-modulated symbol sequence $\{d_i\}$. Transmission of $\{d_i; i = 0 \sim C - 1\}$ is considered here. Each symbol is then spread using orthogonal

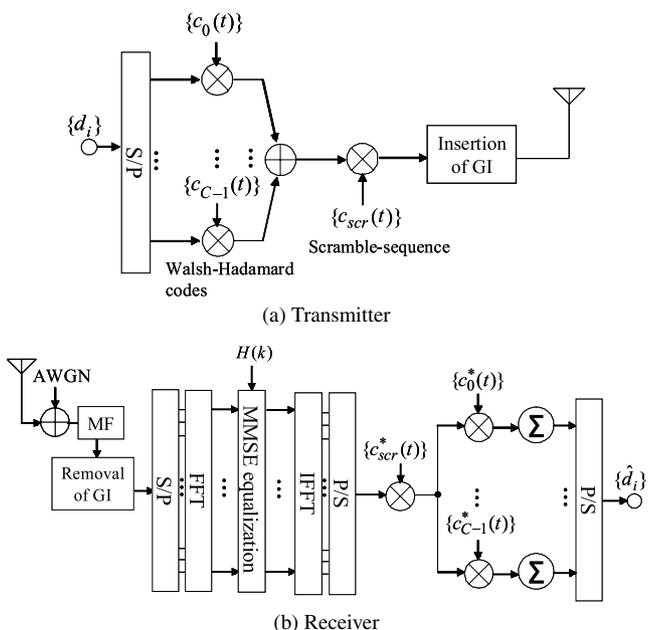


Fig. 1 Transmission system model of orthogonal multicode DS-CDMA with frequency-domain equalization.

spreading codes $\{c_i(t) = \pm 1; i = 0 \sim C - 1, t = 0 \sim SF - 1\}$ of spreading factor SF . The resultant C parallel chip sequences are added (i.e., code-multiplexed) and multiplied by a scramble sequence $\{c_{scr}(t) = \pm 1; t = 0 \sim SF - 1\}$ for making the transmitting signal noise-like. The generated multicode DS-CDMA chip sequence $s(t)$ can be expressed using the equivalent lowpass representation as

$$s(t) = \sqrt{2E_c/T_c} \sum_{i=0}^{C-1} d_i c_{i \bmod C}(t \bmod SF) \times u(i - C \lfloor t/SF \rfloor) c_{scr}(t) \tag{1}$$

for $t = 0 \sim SF - 1$, where E_c and T_c respectively represent the chip energy per code and chip duration, $u(i)=1(0)$ for $i = 0 \sim C - 1$ (otherwise), and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x . Before transmission, the last N_g chips of the multicode DS-CDMA chip sequence is inserted as a cyclic prefix into the guard interval (GI) placed at the beginning of the chip sequence [11], [12].

At the receiver, GI is removed and SF -point fast Fourier transform (FFT) is applied to the received chip sequence to obtain the SF subcarrier components (although DS-CDMA transmission does not use subcarriers for modulation, the terminology “subcarrier” is used for explanation purpose only), and frequency-domain MMSE equalization is performed as in MC-CDMA [7]–[9]. Then, inverse FFT (IFFT) is applied to convert the SF subcarrier components to the time-domain chip sequence, and despreading and coherent demodulation are carried out to obtain the received soft symbol sequence $\{\hat{d}_i; i = 0 \sim SF - 1\}$.

3. Transmit Diversity

3.1 DTD

DTD is to transmit the same chip sequence from N antennas after adding different time delays $\{\tau_{tx,n}; n = 0 \sim N - 1\}$ [14]. A DTD transmitter is illustrated in Fig. 2. The receiver structure is the same as Fig. 1.

The transmitted chip sequences are received via frequency-selective fading channels. Fading channel associated with each transmit antenna is assumed to be composed of L distinct propagation paths with chip-spaced different

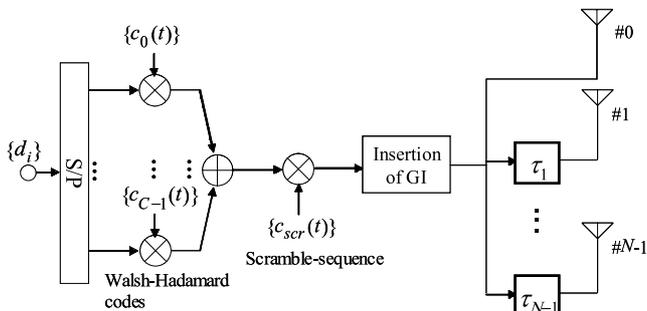


Fig. 2 N-antenna DTD transmitter structure.

time delays $\{\tau_l; l = 0 \sim L - 1\}$. Without loss of generality, time delay of the l th path is assumed to be $\tau_l = lT_c$. In DTD, the time delay of $\tau_{lx,n} = n\Delta\tau$ is added on the n th antenna. The received chip sequence can be expressed, using the equivalent lowpass and chip-spaced time representation, as

$$r(t) = \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} \xi_{n,l} s(t - n\Delta\tau - l) + \eta(t), \quad (2)$$

where $\xi_{n,l}$ is the complex path gain of the l th path corresponding to the n th transmit antenna and $\eta(t)$ is the zero-mean Gaussian noise process with variance $2N_0/T_c$ due to the additive white Gaussian noise (AWGN) having the one-sided power spectrum density N_0 . It is assumed that the path gains $\{\xi_{n,l}\}$ stay constant over one signaling period, i.e., $SF \cdot T_c$.

The received chip sequence $r(t)$, after removal of GI of N_g chips, is decomposed into SF subcarrier components $\{R(k); k = 0 \sim SF - 1\}$ by applying SF -point FFT:

$$R(k) = H(k)S(k) + N(k), \quad (3)$$

where $S(k)$ is the k th subcarrier component of the transmit chip sequence, given by

$$S(k) = \sum_{t=0}^{SF-1} s(t) \exp\left(-j2\pi k \frac{t}{SF}\right), \quad (4)$$

$N(k)$ is the noise component and $H(k)$ is the channel gain at the k th subcarrier frequency. $H(k)$ can be expressed as

$$H(k) = \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} \xi_{n,l} \exp\left(-j2\pi k \frac{n\Delta\tau + l}{SF}\right). \quad (5)$$

Frequency-domain MMSE equalization is carried out as

$$\tilde{R}(k) = w(k)R(k), \quad (6)$$

where $w(k)$ is the MMSE weight given by [7]–[9]

$$w(k) = \frac{H^*(k)}{|H(k)|^2 + \left(C \frac{E_c}{N_0}\right)^{-1}} \quad (7)$$

with $*$ representing the complex conjugate operation and E_c/N_0 being the average received chip energy-to-AWGN power spectrum density ratio. Equation (6) can be rewritten as

$$\begin{aligned} \tilde{R}(k) &= \{w(k)H(k)\}S(k) + w(k)N(k) \\ &= \tilde{H}(k)S(k) + \tilde{N}(k), \end{aligned} \quad (8)$$

where $\tilde{H}(k) = w(k)H(k)$ can be viewed as the equivalent channel gain after frequency-domain equalization and is given by

$$\tilde{H}(k) = \frac{|H(k)|^2}{|H(k)|^2 + \left(C \frac{E_c}{N_0}\right)^{-1}}. \quad (9)$$

3.2 Frequency-Domain STTD

Transmitter and receiver structures for frequency-domain STTD are shown in Fig. 3. The chip sequences of two consecutive signaling periods of multicode DS-CDMA (i.e., even and odd signaling time periods of $\lfloor t/SF \rfloor = 2m$ and $\lfloor t/SF \rfloor = 2m + 1$) are respectively represented as $\{s_e(t); t = 2mSF \sim (2m + 1)SF - 1\}$ and $\{s_o(t); t = (2m + 1)SF \sim 2(m + 1)SF - 1\}$. After STTD encoding, GI is inserted and transmitted from two antennas simultaneously. It should be noted that the transmit power from each antenna is reduced by half so that the total transmit power is kept the same as in the single transmit antenna case. At the receiver, after removal of GI and applying SF -point FFT, STTD decoding and frequency-domain MMSE equalization is carried out simultaneously on each subcarrier component. Then, by applying IFFT, the time-domain multicode chip sequence is obtained from SF subcarrier components, and despreading and coherent demodulation are carried out to obtain the received soft symbol sequence $\{\hat{d}_i; i = 0 \sim C - 1\}$.

(a) Frequency-domain STTD encoding

Alamouti's STTD encoding [15] is applied to each subcarrier component as shown in Table 1. $\{S_e(k); k = 0 \sim SF - 1\}$ and $\{S_o(k); k = 0 \sim SF - 1\}$ represent the subcarrier components of $\{s_e(t); t = 2mSF \sim (2m + 1)SF - 1\}$ and $\{s_o(t); t = (2m + 1)SF \sim 2(m + 1)SF - 1\}$, respectively. The sub-

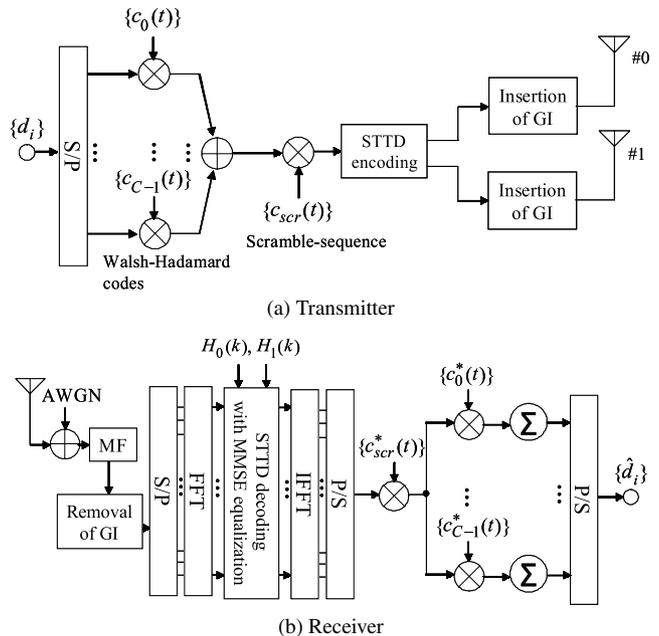


Fig. 3 Transmitter and receiver structures of multicode DS-CDMA using frequency-domain STTD and MMSE equalization.

Table 1 Frequency-domain STTD encoding.

		Antenna index n	
		0	1
Signaling period	$2m$	$\{S_e(k)\}$	$\{S_o(k)\}$
	$2m+1$	$\{-S_o^*(k)\}$	$\{S_e^*(k)\}$

carrier components $\{R_e(k); k = 0 \sim SF - 1\}$ and $\{R_o(k); k = 0 \sim SF - 1\}$ of the multicode chip sequence received during even and odd signaling periods can be expressed as

$$\begin{cases} R_e(k) = H_0(k)S_e(k) + H_1(k)S_o(k) + N_e(k) \\ R_o(k) = -H_0(k)S_o^*(k) + H_1(k)S_e^*(k) + N_o(k) \end{cases}, \quad (10)$$

where $\{H_0(k)\}$ and $\{H_1(k)\}$ are the channel gains corresponding to the two transmit antennas, and are expressed as

$$H_n(k) = \sum_{l=0}^{L-1} \xi_{n,l} \exp\left(-j2\pi k \frac{l}{SF}\right), \quad n = 0 \text{ and } 1 \quad (11)$$

and $N_e(k)$ and $N_o(k)$ are the k th subcarrier components of noise process due to AWGN. STTD decoding suggested in [15] is carried out as

$$\begin{cases} \tilde{R}_e(k) = H_0^*(k)R_e(k) + H_1(k)R_o^*(k) \\ \tilde{R}_o(k) = H_1^*(k)R_e(k) - H_0(k)R_o^*(k) \end{cases}, \quad (12)$$

where channel gains are assumed to stay constant during two signaling periods. Substituting Eq. (10) into Eq. (12) gives

$$\begin{cases} \tilde{R}_e(k) = (|H_0(k)|^2 + |H_1(k)|^2)R_e(k) \\ \quad + H_0^*(k)N_e(k) + H_1(k)N_o^*(k) \\ \tilde{R}_o(k) = (|H_0(k)|^2 + |H_1(k)|^2)R_o(k) \\ \quad + H_1^*(k)N_e(k) - H_0(k)N_o^*(k) \end{cases}. \quad (13)$$

The above STTD decoding of Eq. (12) has turned out to be equivalent to the frequency-domain maximal-ratio combining (MRC) equalization [5], [12], [13].

Frequency-domain STTD encoding shown in Table 1 requires SF -point FFT and IFFT operations in the transmitter. It is possible to perform STTD encoding without FFT and IFFT operations. Applying IFFT to $\{S_e^*(k)\}$, we obtain

$$\frac{1}{SF} \sum_{k=0}^{SF-1} S_e^*(k) \exp\left(j2\pi t \frac{k}{SF}\right) = s_e^*(SF - t). \quad (14)$$

This implies that STTD encoding of Table 1 is equivalent to transmitting $\{s_e^*(t)\}$ and $\{s_o^*(t)\}$ in the reverse order and both FFT and IFFT operations are not required. The equivalent time-domain STTD encoding is shown in Table 2.

(b) Frequency-domain STTD decoding and MMSE equalization

Since STTD decoding using frequency-domain MRC equalization in Eq. (12) emphasizes the frequency-selectivity, orthogonality property among spreading codes is lost and thus, large inter-chip-interference occurs, resulting in a degraded BER performance. We apply frequency-domain MMSE equalization used in MC-CDMA [7]–[9] to STTD

Table 2 Equivalent time-domain STTD encoding.

		Antenna index n	
		0	1
Signaling period	$2m$	$\{s_e(t)\}$	$\{s_o(t)\}$
	$2m+1$	$\{-s_o^*(SF - t)\}$	$\{s_e^*(SF - t)\}$

decoding. STTD decoding operation of Eq. (12) is modified as

$$\begin{cases} \tilde{R}_e(k) = w_0^*(k)R_e(k) + w_1(k)R_o^*(k) \\ \tilde{R}_o(k) = w_1^*(k)R_e(k) - w_0(k)R_o^*(k) \end{cases}, \quad (15)$$

where the weights, $\{w_0(k)\}$ and $\{w_1(k)\}$, are chosen to jointly minimize the mean square error (MSE) between $\tilde{R}_e(k)$ and $S_e(k)$ and that between $\tilde{R}_o(k)$ and $S_o(k)$. After some manipulation following Ref. [9], $w_0(k)$ and $w_1(k)$ are given by

$$\begin{cases} w_0(k) = \frac{H_0(k)}{|H_0(k)|^2 + |H_1(k)|^2 + \left(\frac{C E_c}{2 N_0}\right)^{-1}} \\ w_1(k) = \frac{H_1(k)}{|H_0(k)|^2 + |H_1(k)|^2 + \left(\frac{C E_c}{2 N_0}\right)^{-1}} \end{cases}, \quad (16)$$

which are equivalent to MMSE equalization weights with two-antenna diversity reception of MC-CDMA signals with a 3 dB power penalty [9]. The derivation process of the weights is shown in Appendix. Substituting Eqs. (10) and (16) into Eq. (15), we obtain

$$\begin{cases} \tilde{R}_e(k) = \tilde{H}(k)S_e(k) + \tilde{N}_e(k) \\ \tilde{R}_o(k) = \tilde{H}(k)S_o(k) + \tilde{N}_o(k) \end{cases}, \quad (17)$$

where $\tilde{H}(k)$ is the equivalent channel gain after joint STTD decoding and frequency-domain MMSE equalization and $\tilde{N}_e(k)$ and $\tilde{N}_o(k)$ are the noise components. $\tilde{H}(k)$ is given by

$$\tilde{H}(k) = \frac{|H_0(k)|^2 + |H_1(k)|^2}{|H_0(k)|^2 + |H_1(k)|^2 + \left(\frac{C E_c}{N_0}\right)^{-1}}. \quad (18)$$

Applying SF -point IFFT to $\{\tilde{R}_e(k)\}$ and $\{\tilde{R}_o(k)\}$, the time-domain multicode chip sequences, $\{\tilde{r}_e(t); t = 2mSF \sim (2m+1)SF - 1\}$ and $\{\tilde{r}_o(t); t = (2m+1)SF \sim 2(m+1)SF - 1\}$, are obtained:

$$\tilde{r}_{e(o)}(t) = \frac{1}{SF} \sum_{k=0}^{SF-1} \tilde{R}_{e(o)}(k) \exp\left(j2\pi t \frac{k}{SF}\right). \quad (19)$$

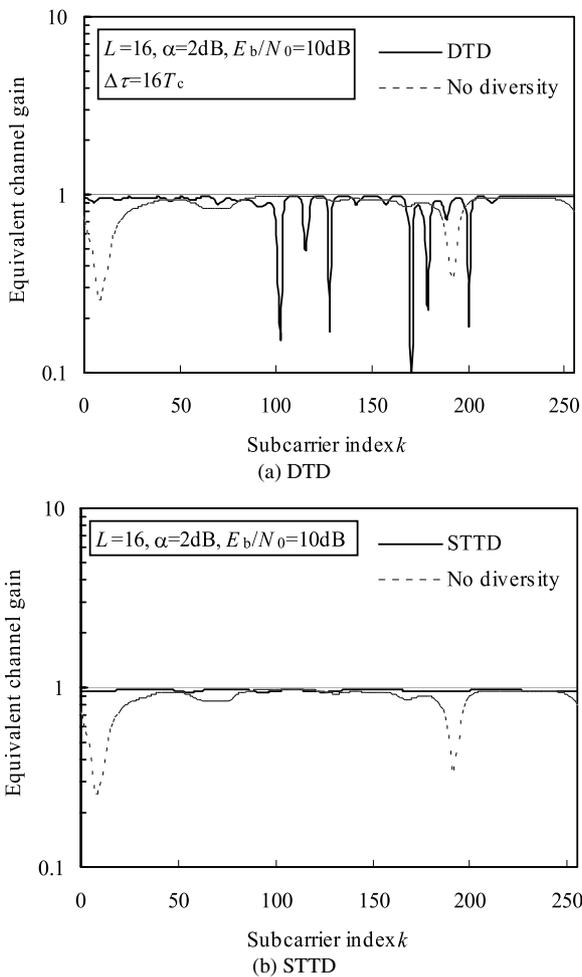
Finally, the soft symbol sequence of $\{\hat{d}_i; i = 0 \sim C - 1\}$ is recovered after despreading and coherent demodulation.

4. Computer Simulation

Simulation condition is shown in Table 3. Quadrature phase shift keying (QPSK) and binary PSK (BPSK) are assumed for data modulation and spreading modulation, respectively. A very slow chip-spaced L -path frequency-selective Rayleigh fading channel having exponential power delay profile with decay factor α dB is assumed. The time delay τ_l of the l th path is assumed to be $\tau_l = lT_c$.

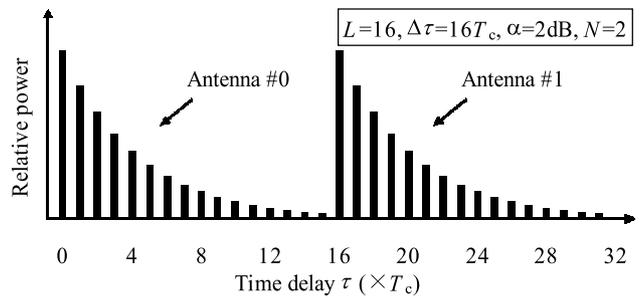
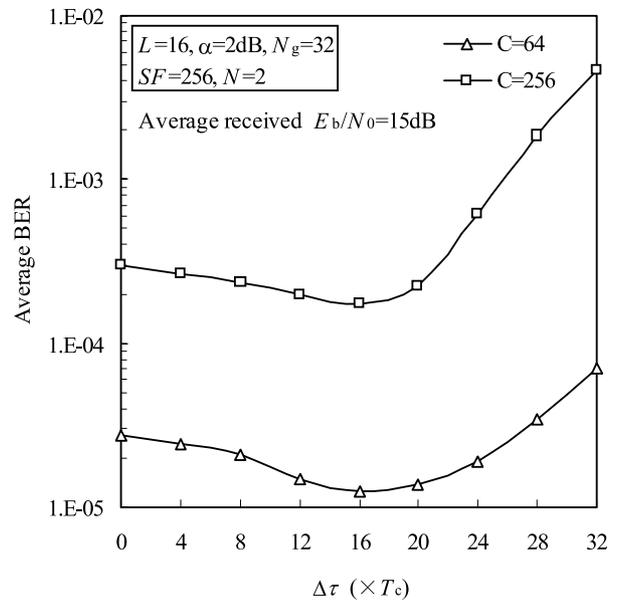
Table 3 Simulation condition.

Data modulation		QPSK
Multicode spreading	Spreading modulation	BPSK
	Spreading codes	Walsh-Hadamard codes
	Spreading factor	$SF=256$
	No. of parallel codes	$C=1\sim 256$
Scramble code		m -sequence with a period of 4095 chips
Guard interval length		$N_g=32$
Transmit diversity		$N=2$ -DTD and STTD
Propagation channel model		$L=16$ -path Rayleigh fading channel
Frequency-domain equalization		MMSE
Channel estimation		Ideal


Fig. 4 Equivalent channel gain for $\alpha = 2$ dB and average received $E_b/N_0 = 10$ dB.

4.1 Equivalent Channel Gain $\tilde{H}(k)$

The equivalent channel gain $\tilde{H}(k)$, defined in Eqs. (9) and (18), after frequency-domain MMSE equalization is shown in Fig. 4 for $\alpha=2$ dB and the average received signal energy per bit-to-AWGN power spectrum density ratio $E_b/N_0=10$ dB. Since DTD can strengthen the fading channel frequency-selectivity by increasing the equivalent num-


Fig. 5 An example of the power delay profile observed at the receiver for $N=2$ and $\Delta\tau = 16T_c$.

Fig. 6 BER dependency on $\Delta\tau$ at the average received $E_b/N_0=15$ dB.

ber of propagation paths, it can increase the frequency diversity gain through the process of frequency-domain MMSE equalization. On the other hand, STTD is equivalent to antenna diversity reception per subcarrier and hence it can reduce the fluctuation of channel gain.

4.2 Effect of DTD

An example of the power delay profile observed at the receiver is illustrated in Fig. 5 for 2-antenna DTD ($N=2$) with time delay difference $\Delta\tau = 16T_c$. Figure 6 plots the dependency of the BER on the added time delay $\Delta\tau$ for the average $E_b/N_0=15$ dB when $N=2$ for $C=64$ and 256. As $\Delta\tau$ increases, the BER reduces due to increasing frequency diversity effect, but starts to increase when the maximum time delay difference resulting from the addition of time delay and the propagation time delay becomes larger than GI. Therefore, the optimum time delay difference that minimizes the BER is given by

$$\Delta\tau = \left\lfloor \frac{N_g - L}{N - 1} \right\rfloor T_c, \quad (20)$$

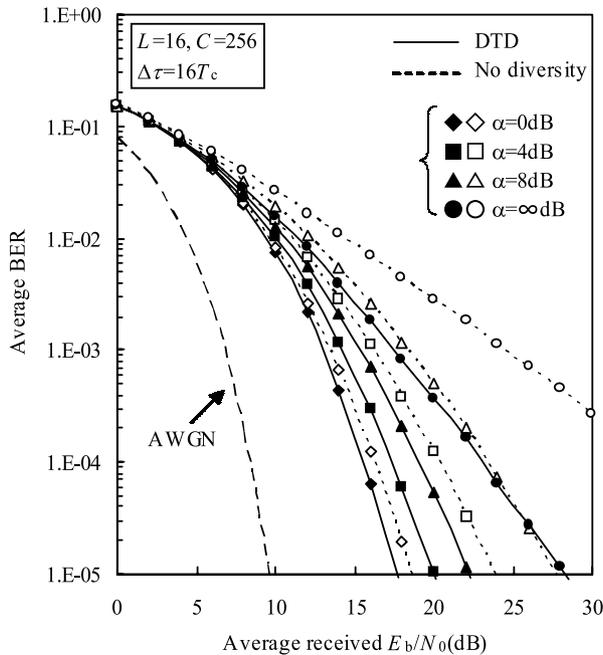


Fig. 7 Average BER performance using joint 2-antenna DTD and frequency-domain MMSE equalization with α as a parameter.

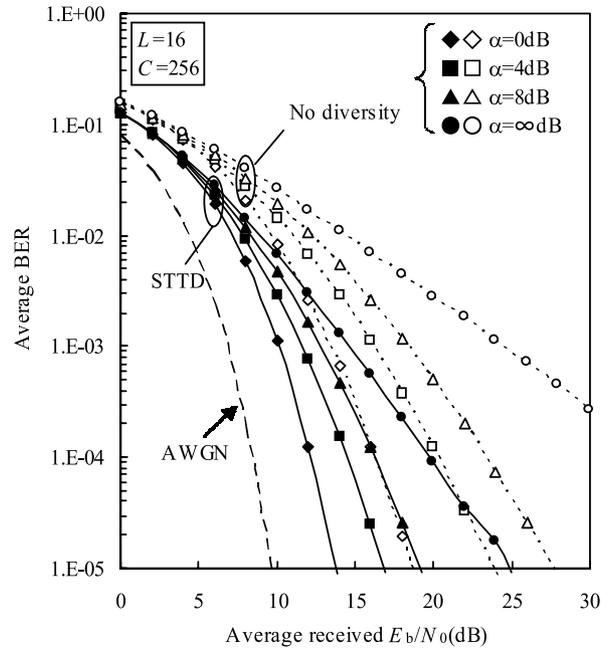


Fig. 9 Average BER performance using joint frequency-domain STTD and MMSE equalization with α as a parameter.

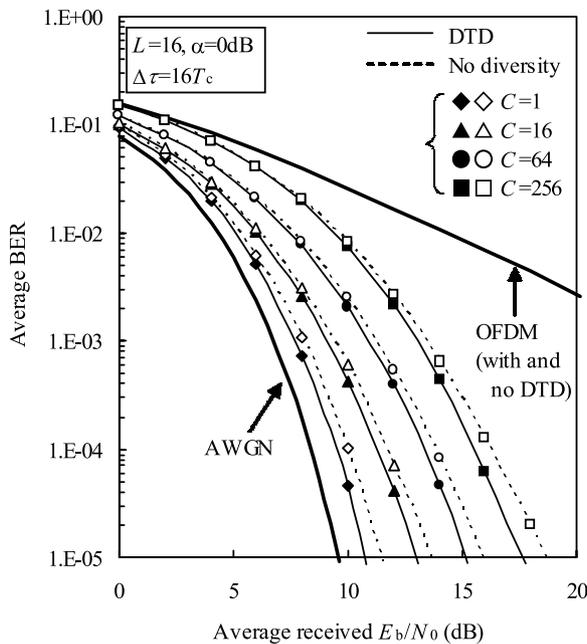


Fig. 8 Average BER performance using joint 2-antenna DTD and frequency-domain MMSE equalization with C as a parameter for $\alpha=0$ dB.

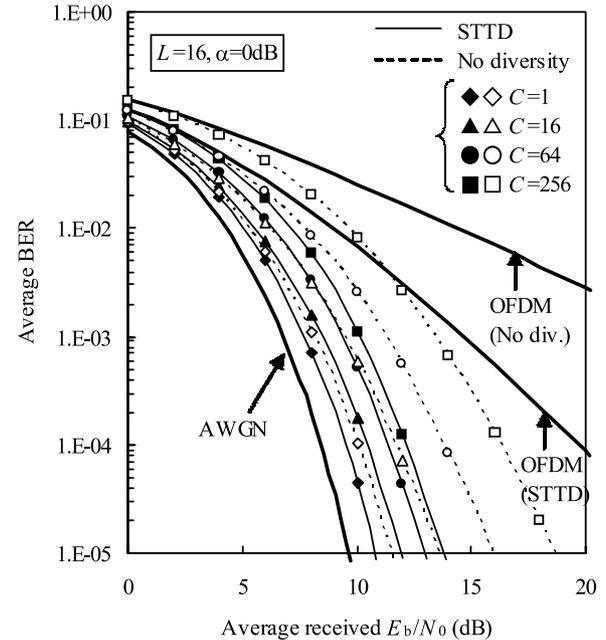


Fig. 10 Average BER performance using joint STTD and frequency-domain MMSE equalization with C as a parameter for $\alpha=0$ dB.

Substituting $N_g=32$, $L=16$, and $N=2$ into Eq. (20) gives $\Delta\tau = 16T_c$. This agrees quite well with the simulation result in Fig. 6. In the following simulation, $\Delta\tau = 16T_c$ is assumed.

The average BER performance of orthogonal multi-code DS-SS using joint 2-antenna DTD and frequency-domain MMSE equalization is plotted with α as a parameter for $C=256$ in Fig. 7. In this case, the transmission

data rate is equal to OFDM using the same bandwidth (256 subcarriers). For comparison, the BER performances for frequency-nonslective fading channel ($\alpha = \infty$ dB) and AWGN channel are also plotted. It is seen that the BER performance improves due to increasing frequency-diversity gain as the value of α decreases. DTD provides better BER performance and is particularly useful when the channel frequency-selectivity is weak, i.e., for large values of α .

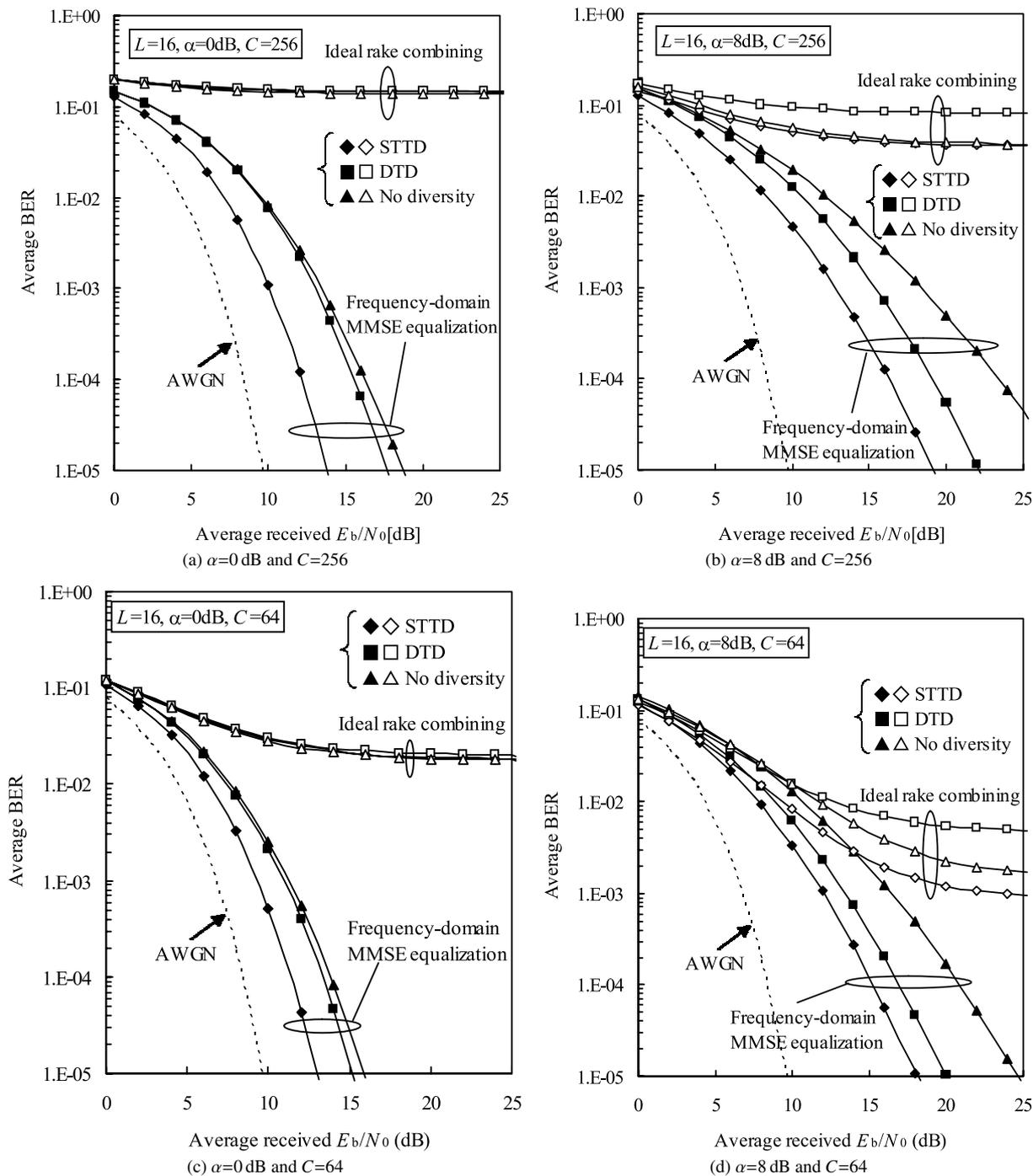


Fig. 11 Performance comparison of DTD and STTD.

For example, a DTD gain of 5.6 (11.8) dB is obtained at $BER=10^{-4}$ when $\alpha = 8$ (∞) dB. It is worthwhile to note that DTD does not help to improve the OFDM transmission performance at all and the BER performance of OFDM is identical to that of multicode DS-CDMA without DTD and with $\alpha = \infty$ dB irrespective of the channel frequency-selectivity.

The average BER performance using joint 2-antenna DTD and frequency-domain MMSE equalization is plotted with C as a parameter in Fig. 8 for $\alpha=0$ dB. When $\alpha=0$ dB, the frequency-selectivity of the fading channel is strong

enough and thus, the performance improvement by the additional use of DTD is small. For the single code case ($C=1$), degradation in the required E_b/N_0 for $BER=10^{-4}$ from the AWGN case is as small as about 1.2 dB even with no DTD. As the code-multiplexing order C increases (in order to increase the transmission data rate), the BER performance degrades, but DTD still provides better BER performance compared with the no DTD case.

4.3 Effect of Frequency-Domain STTD

Figure 9 plots the BER performance of orthogonal multi-code DS-CDMA with joint frequency-domain STTD and MMSE equalization for $C=256$ with α as a parameter. When $\alpha=0$ dB, a STTD gain of 4.1 dB is obtained at $\text{BER}=10^{-4}$ compared with no transmit diversity case. As α increases (i.e., the channel frequency-selectivity becomes weaker), the average BER performance degrades, but the STTD gain becomes larger. When $\alpha=8$ dB, an STTD gain of as much as 7.2 dB is obtained. It is worthwhile to note again that the $C=256$ -multicode DS-CDMA provides the same data rate as OFDM while achieving much improved BER performance. Figure 10 shows the average BER performance with C as a parameter for $\alpha=0$ dB. As C increases, the average BER performance degrades due to the increasing inter-chip-interference, but STTD significantly improves the BER performance compared with no STTD case.

4.4 Comparison of DTD and Frequency-Domain STTD

Performance comparison of DTD and frequency-domain STTD is shown in Fig. 11. DTD is used to get a larger frequency-diversity gain by increasing the equivalent number of paths. Therefore, additional frequency-diversity gain cannot be expected when frequency-selectivity of the channel is already strong (as seen from Fig. 8). On the other hand, STTD can improve the performance even when the frequency-selectivity of the fading channel is strong (as seen from Fig. 9), because it is equivalent to per-subcarrier two-branch MRC antenna diversity reception and the diversity gain independent of the channel frequency-selectivity can be obtained. For the case of strong frequency-selectivity (i.e., $\alpha=0$ dB), STTD can obtain a diversity gain of 4.1 dB at $\text{BER}=10^{-4}$, while DTD can obtain a diversity gain of only 0.7 dB. For a weak frequency-selectivity case of $\alpha=8$ dB, STTD can achieve a diversity gain of 7.2 dB while DTD can obtain a diversity gain of 4.3 dB.

For reference, the BER performance of DS-CDMA using joint rake combining and transmit diversity is also plotted in Fig. 11. When rake combining is used, the code orthogonality property is lost due to large inter-path-interference (IPI) and hence only slight performance improvement can be achieved by STTD when $\alpha=8$ dB and $C=64$ (but, almost no performance improvement can be seen when $\alpha=0$ dB and/or $C=256$). However, DTD enhances the channel frequency-selectivity (i.e., even larger IPI is produced) and therefore, the BER performance is made worse than with no transmit diversity. On the other hand, when frequency-domain MMSE equalization and STTD are jointly used, superior BER performance can be obtained because STTD encoding and decoding are carried out for each subcarrier.

5. Conclusion

In this paper, joint frequency-domain equalization and transmit diversity was presented for the reception of orthogonal multicode DS-CDMA signals in a frequency-selective fading channel. As for transmit diversity, delay transmit diversity (DTD) and frequency-domain space-time coded transmit diversity (STTD) were considered. The BER performance improvement in a frequency-selective Rayleigh fading channel was evaluated by computer simulation. It was found that transmit diversity can improve the average BER performance significantly; however, the DTD gain becomes smaller as the channel frequency-selectivity becomes stronger. STTD was always found to provide a better BER performance than DTD irrespective of the channel frequency-selectivity.

In this paper, ideal channel estimation was assumed. Pilot assisted channel estimation can be applied to estimate the channel gains. Pilot sequence design and the evaluation of the BER performance using a practical channel estimation method are left for future study. In this paper, a very slow frequency-selective fading channel was assumed. If the fading rate becomes faster, the inter-subcarrier interference may be produced, thereby degrading the BER performance. The performance evaluation in a fast fading channel is also left for an interesting future study.

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Appendix

MMSE weights for joint frequency-domain equalization and STTD decoding are derived. Substituting Eq. (15) into Eq. (10), we obtain

$$\begin{cases} \tilde{R}_e(k) = \{w_0^*(k)H_0(k) + w_1(k)H_1^*(k)\} S_e(k) \\ \quad + \{w_0^*(k)H_1(k) - w_1(k)H_0^*(k)\} S_o(k) \\ \quad + w_0^*(k)N_e(k) + w_1(k)N_o^*(k), \\ \tilde{R}_o(k) = \{w_1^*(k)H_0(k) - w_0(k)H_1^*(k)\} S_e(k) \\ \quad + \{w_1^*(k)H_1(k) + w_0(k)H_0^*(k)\} S_o(k) \\ \quad + w_1^*(k)N_e(k) - w_0(k)N_o^*(k). \end{cases} \quad (\text{A} \cdot 1)$$

The equalization error at the k th subcarrier is defined as

$$\begin{cases} \varepsilon_e(k) = \tilde{R}_e(k) - S_e(k) \\ \varepsilon_o(k) = \tilde{R}_o(k) - S_o(k). \end{cases} \quad (\text{A} \cdot 2)$$

The mean square error (MSE) $E[|\varepsilon_{e(o)}(k)|^2]$ is given by

$$\begin{aligned} E[|\varepsilon_{e(o)}(k)|^2] &= \left(\sum_{i=0}^{C-1} 2 \frac{E_c}{T_c} \right) + \left(\sum_{i=0}^{C-1} 2 \frac{E_c}{T_c} \right) \\ &\quad \cdot |w_0^*(k)H_0(k) + w_1(k)H_1^*(k)|^2 \\ &\quad + \left(\sum_{i=0}^{C-1} 2 \frac{E_c}{T_c} \right) |w_0^*(k)H_1(k) - w_1(k)H_0^*(k)|^2 \\ &\quad + \frac{2N_0}{T_c} (|w_0(k)|^2 + |w_1(k)|^2) - 2 \left(\sum_{i=0}^{C-1} 2 \frac{E_c}{T_c} \right) \end{aligned}$$

$$\cdot \text{Re} [w_0^*(k)H_0(k) + w_1(k)H_1^*(k)]. \quad (\text{A} \cdot 3)$$

The weights, $w_0(k)$ and $w_1(k)$, that minimize the MSE, i.e.,

$$\begin{cases} \frac{\partial E[|\varepsilon_{e(o)}(k)|^2]}{\partial w_0(k)} = 0 \\ \frac{\partial E[|\varepsilon_{e(o)}(k)|^2]}{\partial w_1(k)} = 0 \end{cases}, \quad (\text{A} \cdot 4)$$

are called the MMSE weights. After some manipulation following Ref. [9], we can obtain

$$\begin{cases} w_0(k) = \frac{H_0(k)}{|H_0(k)|^2 + |H_1(k)|^2 + \left(\frac{C}{2} \frac{E_c}{N_0}\right)^{-1}} \\ w_1(k) = \frac{H_1(k)}{|H_0(k)|^2 + |H_1(k)|^2 + \left(\frac{C}{2} \frac{E_c}{N_0}\right)^{-1}} \end{cases}. \quad (\text{A} \cdot 5)$$



CDMA.



Takeshi Itagaki received his B.S. and M.S. degrees in communications engineering from Tohoku University, Sendai, Japan, in 2002 and 2004, respectively. Since April 2004, he has been with Sony Ericsson Mobile Communications Japan, Inc., Tokyo, Japan, where he is involved with the research and development of Wireless Access in Communication Core Technology Development Department. His research interests include time and frequency diversity techniques in direct sequence and multicarrier

Kazuaki Takeda received his B.E. degree in communications engineering from Tohoku University, Sendai, Japan, in 2003. Currently he is a graduate student at the Department of Electrical and Communications Engineering, Graduate School of Engineering, Tohoku University. His research interests include frequency-domain equalization for direct sequence CDMA and transmit/receive diversity techniques.



Fumiya Adachi received his B.S. and Dr. Eng. degrees in electrical engineering from Tohoku University, Sendai, Japan, in 1973 and 1984, respectively. In April 1973, he joined the Electrical Communications Laboratories of Nippon Telegraph & Telephone Corporation (now NTT) and conducted various types of research related to digital cellular mobile communications. From July 1992 to December 1999, he was with NTT Mobile Communications Network, Inc. (now NTT DoCoMo, Inc.), where

he led a research group on wideband/broadband CDMA wireless access for IMT-2000 and beyond. Since January 2000, he has been with Tohoku University, Sendai, Japan, where he is a Professor of Electrical and Communication Engineering at Graduate School of Engineering. His research interests are in CDMA and TDMA wireless access techniques, CDMA spreading code design, Rake receiver, transmit/receive antenna diversity, adaptive antenna array, bandwidth-efficient digital modulation, and channel coding, with particular application to broadband wireless communications systems. From October 1984 to September 1985, he was a United Kingdom SERC Visiting Research Fellow in the Department of Electrical Engineering and Electronics at Liverpool University. From April 1997 to March 2000, he was a visiting Professor at Nara Institute of Science and Technology, Japan. He was a recipient of IEICE Achievement Award 2002 and was a co-recipient of the IEICE Transactions best paper of the year award 1996 and again 1998. He is an IEEE Fellow and was a co-recipient of the IEEE Vehicular Technology Transactions best paper of the year award 1980 and again 1990 and also a recipient of Avant Garde award 2000.