SDM-COFDM Technologies for Broadband Wireless Access over 100 Mbit/s

Takatoshi Sugiyama†*, Satoshi Kurosaki, Daisei Uchida, and Yusuke Asai*

Abstract

This paper describes a simple SDM-COFDM (space division multiplexed, coded orthogonal frequency division multiplexing) scheme for MIMO (multiple input multiple output)-based broadband wireless access. The proposed scheme reduces inter-channel interference with a simple feed-forward canceller by using a new space-time coding preamble. Moreover, the proposed likelihood weighting based on the signal-to-noise power ratio improves the soft decision error correction performance. The proposed SDM-COFDM scheme with 2×2 antennas can double the transmission bit rate to more than 100 Mbit/s per 18 MHz by using 64QAM with a coding rate of 3/4 in frequency selective fading environments.

1. Introduction

The demand for broadband wireless access continues to increase rapidly to provide seamless connection to optical fiber services such as FTTH (fiber to the home), providing a transmission bit rate of more than 100 Mbit/s on the PHY (physical) layer. The third-generation mobile access, IMT-2000, provides up to 2 Mbit/s [1], and wireless LAN standards such as IEEE802.11a/g in the U.S.A., HiperLAN2 in Europe, and HiSWANa in Japan support a maximum transmission bit rate of 54 Mbit/s [2]-[5]. To achieve a breakthrough, IEEE802.11n [6] has just started to be standardized and its target has been set at more than 100-Mbit/s throughput even in the MAC (media access control) layer in 2006. Moreover, it is desirable to achieve a higher transmission bit rate without any occupied frequency bandwidth expansion because of the current frequency spectrum deficiency, especially in the relatively lower bands useful for mobile systems.

Space division multiplexing (SDM) [7] is one of the most promising techniques to satisfy the above-mentioned requirements. It can achieve N times the transmission bit rate of the single transmission via a MIMO (multiple input multiple output) channel by using N sets of antennas at both the transmitter and receiver without any bandwidth expansion, as shown in **Fig. 1**. However, in the SDM schemes, inter-channel interference (ICI) significantly degrades the transmission quality.

To reduce the degradation due to ICI, V-BLAST [8] serially repeats detecting the transmitted signal of one SDM channel by hard decision, regenerating the replica from the detected signal, and subtracting it from the received signal until all signals have been detected. Therefore, both the hardware size and signal processing delay would become very large, even when more antenna elements or subcarriers are used. Another conventional scheme uses maximum likelihood decoding (MLD) [9]. Unfortunately, it is necessary to calculate the likelihood values of K (K=M^N) possible sets of one symbol transmitted from each of N antennas for each OFDM (orthogonal frequency division multiplexing)subcarrier in order to presume one set with M representing the number of constellation points according to the modulation scheme, so the signal processing complexity is very high, especially in higher bit rate modulation schemes such as 64QAM (quadrature amplitude modulation).

To solve these problems, this paper proposes a sim-

† NTT Network Innovation Laboratories
Yokosuka-shi, 239-0847 Japan
E-mail: sugiyama.takatoshi@lab.ntt.co.jp

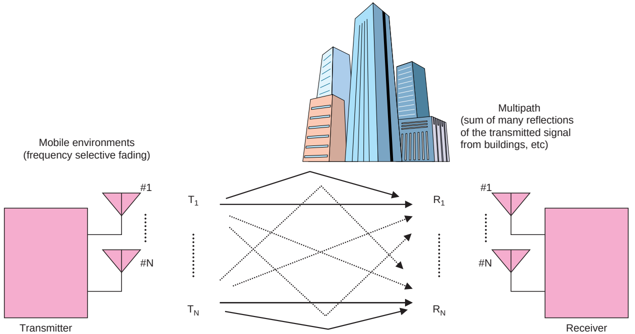
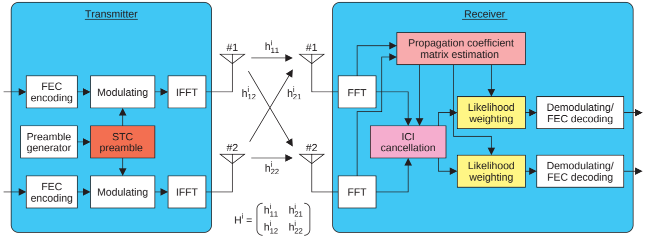


Fig. 1. Structure of SDM scheme.



H^i: propagation coefficient matrix, i: subcarrier number

Fig. 2. Block diagram of the proposed SDM-COFDM scheme.

ple SDM-COFDM (SDM-coded orthogonal frequency division multiplexing) scheme that uses a simple feed-forward ICI canceller [10]. The canceller only multiplies the received symbols by the estimated inverse propagation coefficient matrix for each OFDM subcarrier by using a new preamble based on space-time coding (STC) to improve power efficiency. Moreover, to improve the required CNR (carrier-to-noise power ratio) in frequency selective fading environments, this paper also proposes the SNR (signal-to-noise power ratio)-based likelihood weighting for the soft decision error correction. It is clarified that the proposed SDM-COFDM scheme with two sets of transmission and reception antennas can double the transmission bit rate without any bandwidth expansion and achieves similar PER (packet error rate) performances to the conventional SISO (single-input single-output) transmission in frequency selective fading environments. In particular, it provides more than 100 Mbit/s per 18 MHz by using 64QAM with a coding rate of 3/4.

2. Proposed SDM-COFDM scheme

2.1 Configuration of the SDM-COFDM scheme

A block diagram of the proposed SDM-COFDM scheme is shown in **Fig. 2**. It has two sets of antennas on both the transmitting and receiving sides to construct a 2×2 MIMO channel. Each data stream has its own convolutional encoder. The encoded data is interleaved, distributed to OFDM subcarriers, and mapped to data symbols according to the modulation scheme. Then, the preamble symbols, which are used for the propagation coefficient matrix estimation, are combined at each subcarrier. The combined symbols are transmitted through the transmission antennas after inverse fast Fourier transformation (IFFT). The symbols received via the 2×2 MIMO channel are received by the reception antennas and fed to the FFT blocks, and their ICIs are cancelled. After that, the ICI-cancelled symbols are likelihood-weighted, demodulated, and FEC (forward error correction) decoded.

The relationship between the transmitted data symbols $\mathbf{t}^i$ and the received data symbols $\mathbf{r}^i$ may be expressed as follows.

$$\mathbf{r}^i = \mathbf{H}^i \cdot \mathbf{t}^i + \mathbf{n}^i \quad , \quad \mathbf{H}^i = \begin{pmatrix} h^i_{1,1} & h^i_{2,1} \\ h^i_{1,2} & h^i_{2,2} \end{pmatrix} \qquad (1)$$

Here, $\mathbf{H}^i$ is the 2×2 propagation coefficient matrix of the i-th subcarrier, $h^i_{n,m}$ is the propagation coefficient between the n-th transmission antenna and the m-th reception antenna (Fig. 2), and $\mathbf{n}^i$ represents the zero mean, complex additive white Gaussian noise (AWGN) contained in the received data symbol $\mathbf{r}^i$. Therefore, $\mathbf{H}^i$ must be estimated precisely in order to presume the transmitted data symbol $\mathbf{t}^i$. Considering that each subcarrier transmits narrow-band modulated symbols in an OFDM system, $\mathbf{H}^i$ can be estimated by calculating each propagation coefficient directly from the relationship between the transmitted and received symbols.

2.2 STC preamble

From Eq. (1), in the preamble transmission, the relationship between the transmitted preamble symbols $\mathbf{C}^i$ and the received preamble symbols $\mathbf{B}^i$ can also be described by $\mathbf{B}^i = \mathbf{H}^i \cdot \mathbf{C}^i$, if we assume noise-free conditions. Therefore, the determinant of $\mathbf{C}^i$ must be non-zero to obtain $\mathbf{H}^i$. This paper introduces two preambles for the SDM-COFDM scheme to meet this requirement: the conventional scattered preamble [11] and the newly proposed STC preamble. The frame formats of the i-th subcarrier with the scattered and STC preambles are shown in **Fig. 3**. Here, $c^i_{n,m}$ represents the m-th time slot preamble symbol of the n-th transmission antenna, and $t^i_{n,m}$ represents the m-th time slot data symbol of the n-th transmission antenna.

In the scattered preamble, each transmission antenna is used exclusively to transmit its preamble symbol, and the other preamble symbol must be null. The propagation coefficient $h^i_{n,m}$ can be calculated by $b^i_{n,m}/\ c^i_{n,n}$. Since $c^i_{n,m}$ equals zero when n does not equal m, the time slot cannot be efficiently used for $\mathbf{H}^i$ estimation. Equation (2) gives examples of the scattered preambles.

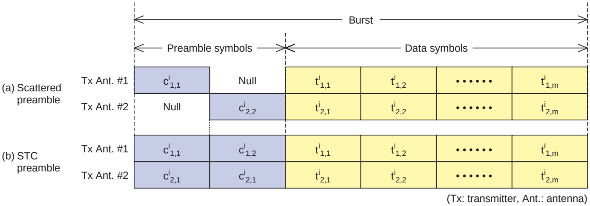


Fig. 3. Frame formats for i-th subcarrier.

$$\mathbf{C}^i{}_{scat1} = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \cdot c^i \;\; , \;\; \mathbf{C}^i{}_{scat2} = \sqrt{2} \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \cdot c^i, \qquad (2)$$

where c^i is a preamble symbol having the same amplitude as the data symbols. $\mathbf{C}^i{}_{scat1}$ has the same peak power as the data symbols, so the total transmission power of the preambles in two time slots is half that of the data symbols. On the other hand, $\mathbf{C}^i{}_{scat2}$ has the same total transmission power as the data symbols, but double the peak power of the data symbols is needed.

To solve these problems, $\mathbf{C}^i$ must satisfy the condition that both $c^i{}_{n,m}$ and the determinant of $\mathbf{C}^i$ must be non-zero because the preamble symbols are simultaneously transmitted in parallel over all the channels. As one of the preambles to satisfy this condition, we propose STC [12] -based preamble $\mathbf{C}^i{}_{STC}$, as shown in Fig. 3.

$$\mathbf{C}^i{}_{STC} = \begin{pmatrix} c^i & -c^i \\ c^i & c^i \end{pmatrix} = X \cdot c^i = \begin{pmatrix} 1 & -1 \\ 1 & 1 \end{pmatrix} \cdot c^i, \qquad (3)$$

where X is one of the simplest space-time encoders for 2×2 antennas. The advantage of the STC preamble compared with the conventional scattered one is the next point. STC preamble enables us to use twice the total preamble transmission power as $\mathbf{C}^i{}_{scat1}$ and needs only half the peak power of the transmission amplifier compared with $\mathbf{C}^i{}_{scat2}$.

2.3 ICI cancellation

We can estimate $\mathbf{H}^i$ for each OFDM subcarrier by using the above-mentioned STC preamble. Consequently, the inverse propagation coefficient matrix $\mathbf{G}^i$ is defined as

$$\mathbf{G}^i \equiv (\mathbf{H}^i)^{-1} = \mathbf{C}^i \cdot (\mathbf{B}^i)^{-1}, \;\; \mathbf{G}^i = \begin{pmatrix} g^i_{1,1} & g^i_{2,1} \\ g^i_{1,2} & g^i_{2,2} \end{pmatrix}, \qquad (4)$$

where, $g^i{}_{n,m}$ is a component of $\mathbf{G}^i$. Thus, the transmitted symbol $\mathbf{t}^i$ can be obtained by simply multiplying the received symbol $\mathbf{r}^i$ by $\mathbf{G}^i$ in Eq. (1).

2.4 SNR-based likelihood weighting

After the cancellation in the proposed SDM-COFDM scheme, the ICI-cancelled symbol loses channel attenuation information because the ideal canceller operates to make the symbol amplitude constant as shown in **Fig. 4**. Therefore, the error correction performance of soft decision Viterbi decoding degrades. Although SISO channel transmission can use the likelihood weighting based on the measured amplitude of each subcarrier in the receiver [13], it is

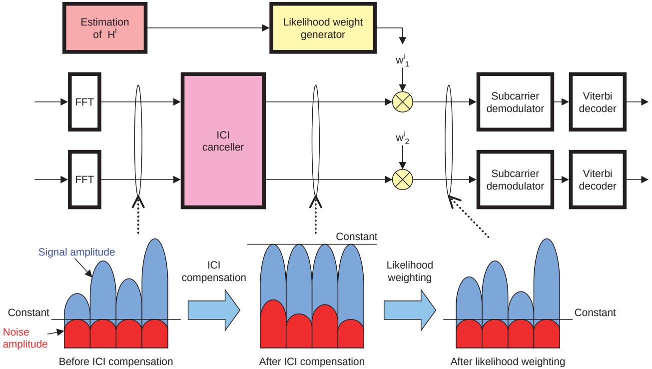


Fig. 4. Concept of the proposed likelihood weighting.

impossible for the SDM-COFDM scheme to measure subcarrier amplitude independently because of the ICI.

In order to solve this problem, we propose likelihood weighting based on the SNR because the ICI-cancelled symbol amplitude is proportional to the SNR in each subcarrier if the noise power of all subcarriers is constant. Since the AWGN $\mathbf{n}^i$ takes an independent zero-mean Gaussian distribution and the phase distribution of $n^i_{l,m}$ is uniform, the square amplitude values of $g^i_{n,m}$ are approximately proportional to the SNR. Therefore, the amplitude weighting is given by Eq. (5) [10].

$$w_n^i = \frac{1}{\sqrt{\sum_{l=1}^{2} \left| g_{l,n}^i \right|^2}} \tag{5}$$

Finally, the likelihood values are weighted by Eq. (5) and fed to the soft decision Viterbi decoders to obtain larger coding gain.

3. Performance of the proposed SDM-COFDM scheme

3.1 Main parameters

We evaluated the performance of the proposed SDM-COFDM scheme through both computer simulation and experiments using a 1/8-scale prototype. The main parameters of the proposed SDM-COFDM scheme, which are based on HIPERLAN/2 and HiSWANa, are listed in **Table 1**. In evaluating the proposed SDM-COFDM scheme with two sets of antennas on both the transmitting and receiving sides, we employ the STC preamble given by Eq. (3). The RF band is the 5-GHz band. In this paper, the total transmission power of the proposed SDM-COFDM scheme is assumed to be equivalent to the SISO transmission. There are 48 data subcarriers and 4 pilot subcarriers in each OFDM symbol. The duration of each symbol T is 4.0 μs including the guard interval length of 0.8 μs. The OFDM subcarriers are modulated using PHY modes that are BPSK (binary phase shift keying) with a coding rate R=1/2, QPSK (quadrature phase shift keying) with R=1/2, 16QAM with R=9/16, 16QAM with R=3/4, and 64QAM with R=3/4. The transmission bit rate values are 12, 24, 54, 72 and 108 Mbit/s, respectively. The occupied frequency bandwidth is 18 MHz. The interference-cancelled data symbols of each subcarrier are demodulated with coherent detection. All encoded data bits of every OFDM symbol are interleaved with a depth of 16 bits. Convolutional coding and 6-bit soft decision Viterbi decoding are used for FEC.

For mobile communication environments, the 18-ray Rayleigh fading HiperLAN 2 model [14] is deployed. The 2×2 MIMO channel is assumed to have four independent fading characteristics between the transmission and the reception antennas. Normalized delay spread by the symbol duration D/T is set to 1.25×10^{-2} and normalized maximum Doppler frequency f_dT is 2×10^{-4}. These values correspond to HiperLAN 2 model A with delay spread of 50 ns and maximum Doppler frequency of 50 Hz for 18 MHz occupied bandwidth in the 5-GHz band.

The prototype is shown in **Fig. 5**. It is 155 mm high,

Table 1. Main parameters of the proposed SDM-COFDM scheme.

RF band	5-GHz band
Number of antennas	2×2
Number of FFT points	64
Number of subcarriers	52 (including 4 pilots)
Effective symbol length	3.2 µs
Guard interval length	0.8 µs
Subcarrier modulation	BPSK, QPSK, 16QAM, 64QAM
Interleaving depth	16 bits
FEC	Convolutional coding (constraint length of 7) - 6-bit soft decision Viterbi decoding
Occupied bandwidth	18 MHz
Transmission bit rate	12 Mbit/s (BPSK, R=1/2) 24 Mbit/s (QPSK, R=1/2) 54 Mbit/s (16QAM, R=9/16) 72 Mbit/s (16QAM, R=3/4) 108 Mbit/s (64QAM, R=3/4)

Main changes are highlighted in red. Prototype: 1/8 scale model

Fig. 5. Appearance of the SDM-COFDM prototype.

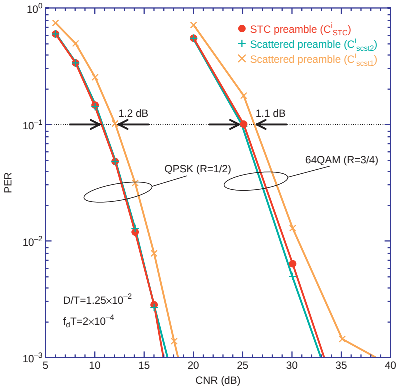


Fig. 6. PER performances of the proposed STC preamble.

430 mm wide, and 450 mm deep.

3.2 Performance of the STC preamble

The PER performance of the proposed STC preamble measured by computer simulation is shown in **Fig. 6**. This figure also shows the error performances of the conventional scattered preamble $\mathbf{C}^i_{scat1}$ and $\mathbf{C}^i_{scat2}$. The CNRs required by $\mathbf{C}^i_{STC}$ are 1.2 and 1.1 dB less than $\mathbf{C}^i_{scat1}$ at a PER of 0.1 on the condition of QPSK with R=1/2 and 64QAM with R=3/4, respectively. Although the CNR required for $\mathbf{C}^i_{STC}$ is almost equal to $\mathbf{C}^i_{scat2}$ because of the same SNR, the STC preamble needs only half the peak power of the transmission amplifier compared with $\mathbf{C}^i_{scat2}$.

3.3 Performance of the SNR-based likelihood weighting

The PER performances of the SNR-based likelihood weighting measured by computer simulation are shown in **Fig. 7**. The required CNR improvements at a PER of 0.1 are 7.1 and 5.6 dB for QPSK with R=1/2 and 64QAM with R=3/4, respectively, compared with those without the weighting. The results show that the proposed likelihood weighting significantly improves error correction performance for soft decision Viterbi decoding.

3.4 PER performance of the prototype

We performed an experiment to confirm the overall performances of the SDM-COFDM prototype using the proposed ICI cancellation with the STC preamble and SNR-based likelihood weighting.

The PER performances of the prototype are shown in **Fig. 8**. It is an example for QPSK with R=1/2 and 64QAM with R=3/4. It shows that, when QPSK with R=1/2 is used, the required CNR at a PER of 0.1 of the proposed SDM-COFDM degrades by less than 4 dB compared with that of the SISO-COFDM. However, assuming double the total transmission for the SDM-COFDM, this corresponds to less than 1 dB degradation of Eb/No (the ratio of energy per bit to noise density), even though the transmission bit rate of SDM-COFDM is roughly double that of SISO-COFDM without any expansion of the occupied frequency bandwidth. Furthermore, even in 64QAM with R=3/4 which is significantly sensitive to the degradation due to the hardware imperfection and the residual estimation error of $\mathbf{H}^i$, the prototype keeps the required CNR degradation below 8 dB (equivalent to 5 dB in Eb/No) at a PER of 0.1 compared with that of the SISO-COFDM. Therefore, it achieves more than 100 Mbit/s per 18 MHz in a frequency selective fading environment with a delay spread of 50 ns and maximum Doppler frequency of 50 Hz in the 5-GHz band without any fatal performance degradation.

4. Conclusion

This paper introduced a simple SDM-COFDM scheme for MIMO-based broadband wireless access to achieve a higher transmission bit rate without any occupied frequency bandwidth expansion, with a simple feed-forward canceller by using a new STC preamble. Moreover, the proposed likelihood weighting based on the SNR improves the error correction performance. The proposed SDM-COFDM scheme with 2×2 antennas can double the transmission bit rate to more than 100 Mbit/s per 18 MHz by using 64QAM with a coding rate of 3/4 in frequency selective fading environments. The feasibility of implementing the proposed approach has been confirmed by the experimental results for the developed prototype. In the near future, a real speed experiment using large-scale integrated circuits will be carried out.

References

[1] ITU-R Rec. M.1455, "Key characteristics for international mobile telecommunications-2000 (IMT-2000) radio interfaces," May 2000.

[2] IEEE802.11a, "High speed physical layer (PHY) in 5GHz band," 1999.

[3] IEEE802.11g, "Higher speed physical layer (PHY) extension to IEEE 802.11b," Feb. 2003.

[4] "Broadband radio access networks (BRAN); HIPERLAN Type 2; Physical (PHY) layer," ETSI TS 101 475 V1.1.1, Apr. 2000.

[5] "Low Power Data Communication Systems Broadband Mobile Access Communication System (HiSWANa)," ARIB STD-T70, Association of Radio Industries and Businesses, 2000.

[6] J. Rosdahl, "HT-SG Draft-PAR," IEEE Doc:802.11-02-798r7, Mar. 2003.

[7] G. J. Foschini and M. J. Gans, "On Limits of Wireless Communication in a Fading Environment

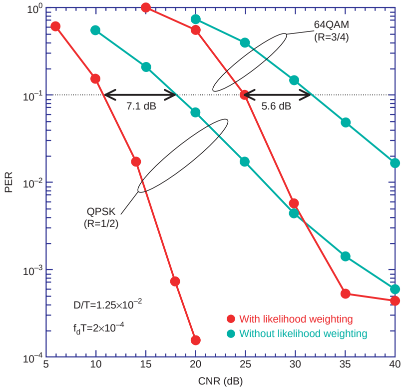


Fig. 7. PER performances of the proposed likelihood weighting.

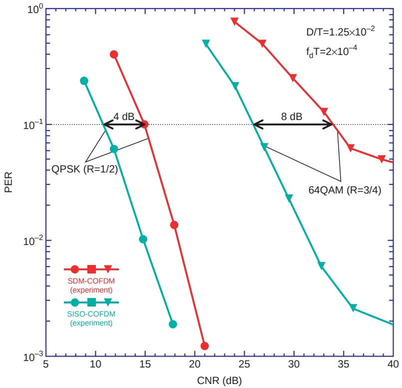


Fig. 8. PER performances of the SDM-COFDM prototype.

When Using Multiple Antennas," Wireless Pers. Commun., Vol. 6, No. 3, pp. 311-335, 1998.
[8] D. K. C. So and R. S. Cheng, "Layered Maximum Likelihood Detection for V-BLAST in Frequency Selective Fading Channels," IEEE VTC 2002 Spring, May 2002.
[9] A. van Zelst, R van Nee, and G. A. Awater, "Space Division Multiplexing (SDM) for OFDM systems," IEEE VTC 2000 Spring, Sep. 2000.
[10] S. Kurosaki, Y. Asai, T. Sugiyama, and M. Umehira, "An SDM-COFDM Scheme Employing a Simple Feed-Forward Inter-Channel Interference Canceller for MIMO Based Broadband Wireless LANs," IEICE Trans. Commun., Vol. E86-B, No. 1, pp. 283-290, Jan. 2003.
[11] F. Tufvesson and T. Maseng, "Pilot assisted channel estimation for OFDM in mobile cellular systems," IEEE VTC 1997, Vol. 3, pp. 1639-1643, May 1997.
[12] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction," IEEE Trans. on IT, Vol. 44, pp. 744-765, Mar. 1998.
[13] H. Sari, G. Karam, and I. Jeanclaude, "Transmission techniques for digital terrestrial TV broadcasting," IEEE Communications Magazine, Vol. 33, No. 2, pp. 100-109, Feb. 1995.
[14] J. Medbo and P. Schramm, "Channel Models for HIPERLAN 2," ETSI/BRAN document No. 3ERI085B, 1998.

Takatoshi Sugiyama
Senior Research Engineer, Wireless Systems Innovation Laboratory, NTT Network Innovation Laboratories.

He received the B.E., M.E. and Ph.D. degrees in electrical engineering from Keio University, Yokohama in 1987, 1989, and 1998, respectively. Since joining NTT in 1989, he has worked on forward error correction, interference compensation, CDMA, modulation-demodulation schemes for wireless communication systems such as satellite, personal, and wireless ATM communication systems. From 1998 to 2001, he was engaged in business planning of international satellite communication services in NTT Communications Corporation, Tokyo, Japan. He is currently responsible for the developing the next-generation broadband wireless access systems using MIMO-OFDM technology. He is a member of IEEE and the Institute of Electronics, Information and Communication Engineers (IEICE). He received the Young Engineer Award from IEICE in 1996.

Satoshi Kurosaki
Research Engineer, Wireless Systems Research Group, Wireless Systems Innovation Laboratory, NTT Network Innovation Laboratories.

He received the B.E. and M.E. degrees in electronic engineering from Kyoto University, Kyoto in 1992 and 1994, respectively. Since joining NTT in 1994, he has worked on antenna switching control and radio link control for wireless communication systems such as personal and wireless ATM communication systems. His current work is the development of broadband SDM-Coded OFDM techniques for the next-generation wireless LANs and cellular systems. He is a member of IEEE.

Daisei Uchida
Member, Wireless Systems Research Group, Wireless Systems Innovation Laboratory, NTT Network Innovation Laboratories.

He received the B.S. and M.S. degrees in applied physics from Tokyo Institute of Technology, Tokyo, in 1994 and 1997, respectively. Since joining NTT in 1997, he has worked on the access scheme for satellite systems, the radio channel assignment and tree topology of PHS-based local positioning and information systems to form wireless multi-hop networks autonomously, and the radio channel assignment and transmission power control for mesh-type broadband FWA systems. He is currently engaged in R&D of modulation-demodulation schemes for SDM-COFDM systems. He is a member of the Institute of Electronics, Information and Communication Engineers (IEICE). He received the Young Engineer Award from IEICE in 2001.

Yusuke Asai
Member, Wireless Systems Research Group, Wireless Systems Innovation Laboratory, NTT Network Innovation Laboratories.

He received the B.E. and M.E. degrees in information and communication engineering from Nagoya University, Nagoya in 1997 and 1999, respectively. Since joining NTT in 1999, he has worked on modulation-demodulation schemes of software defined radio system and OFDM system. He is currently developing the next-generation broadband wireless access systems using MIMO-OFDM technology. He is a member of IEICE.