Non-orthogonal Access Scheme over Multiple Channels with Iterative Interference Cancellation and Fractional Sampling in OFDM Receiver

Hiroyuki Osada, Mamiko Inamori, Yukitoshi Sanada
Department of Electronics and Electrical Engineering, Keio University
Yokohama, 223-8522 Japan
Email:hosada@snd.elec.keio.ac.jp, {inamori, sanada}@elec.keio.ac.jp

***Abstract*—A diversity scheme with Fractional Sampling (FS) in OFDM receivers has been investigated recently. FS path diversity makes use of the imaging components of the desired signal transmitted on the adjacent channel. In this paper non-orthogonal access over multiple channels with iterative interference cancellation (IIC) and FS is proposed. The proposed scheme transmits the imaging component non-orthogonally on the adjacent channel. In order to accommodate the imaging component, it is underlaid on the other desired signal. Through diversity with FS and IIC, non-orthogonal access on multiple channels is realized. Our propose scheme can accommodate non-orthogonal signals with limited diversity gains.**

***Index Terms*—Fractional Sampling, OFDM, Path Diversity, Interference Cancellation.**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) has been used as a modulation scheme in various wireless communication schemes such as terrestrial digital broadcasting, wireless broadband communications, or wireless local area networks. This is because a OFDM symbol has long time duration, and the relative amount of time dispersion caused by multipath delay spread is decreased [1], [2].

On the other hand, diversity is necessary to improve the performance of the OFDM system on the multipath channel. One of the typical diversity techniques is antenna diversity [3]. Multiple antenna elements must be spatially separated in order to reduce the correlation among the received signals [4], [5]. However, it is difficult for small terminals to implement multiple antenna elements. Therefore, a FS scheme that realizes diversity with a single antenna has also been proposed [6]. The FS scheme achieves path diversity by sampling a received signal at a rate higher than the symbol rate and by demodulating them in parallel.

FS path diversity makes use of the imaging components of the desired signal transmitted on the adjacent channel. In the previous paper, a non-orthogonal access scheme with the use of FS and decision feedback cancellation has been evaluated [7]. In this paper, with the use of FS and iterative interference cancellation (IIC), and a non-orthogonal access scheme on multiple channels is investigated. It is demonstrated that our propose scheme can accommodate non-orthogonal signals with limited diversity gains.

This paper is organized as follows. In Section 2, the system model with the proposed cancellation scheme is described. Section 3 shows numerical results through computer simulation. The conclusions of this paper are presented in Section 4.

II. SYSTEM MODEL

A. Channel Allocation and Cancellation Process

Figure 1 shows the channel allocation model. It is for the downlinks of two different terminals. The OFDM signals for the terminals 1 and 2 are transmitted on the adjacent

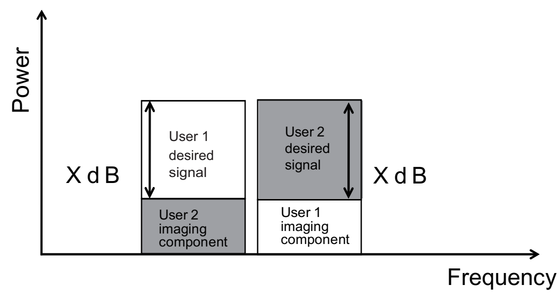

Fig. 1. Channel allocation model.

two channels. Their imaging components are underlaid non-orthogonally on the other desired signals, respectively. The powers of the imaging components are XdB smaller than the desired signals on both of the channels. Therefore, the non-orthogonal signal has to be cancelled in order to demodulate the desired signal. A block diagram of the proposed FS-OFDM receiver is shown in Fig. 2. The signals of User 1 ($S_1[k]$) and User 2 ($S_2[k]$) pass through the analog filters whose impulse responses are given as $p'_{rx}(t)$ and $p_{rx}(t)$. First, the User 1 signal through $p'_{rx}(t)$ is demodulated, its replica is regenerated, and the replica signal is subtracted from the received signal through $p_{rx}(t)$. After the subtraction, the User 2 signal is demodulated with FS, and it's replica is regenerated, and the replica signal is subtracted from the received signal through $p_{rx}(t)$. The cancellation of the signal from User 2 is then realized.

The OFDM signals of User 1 and User 2 are given as

$$u_1[n] = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} S_1[k] \exp\left(j\frac{2\pi nk}{N}\right), \quad (1)$$

$$u_2[n] = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} S_2[k] \exp\left(j\frac{2\pi nk}{N}\right), \quad (2)$$

where $S_1[k]$ and $S_2[k]$ are the information symbols on the kth subcarrier for the User 1 and User 2, respectively, $n(n = 0, 1, ..., 2N-1)$ is the time index, and N is the size of the inverse discrete Fourier transform (IDFT). Then, $u_1[n]$ and $u_2[n]$ pass through different transmit filters. These signals can be generated by IDFT with the size of $2N$ in the base station at a time. However, they are treated separately in this paper for the convenience of the explanation. Suppose the last part of the OFDM signal is copied and appended at the beginning as a guard interval (GI), the transmit signals in the baseband

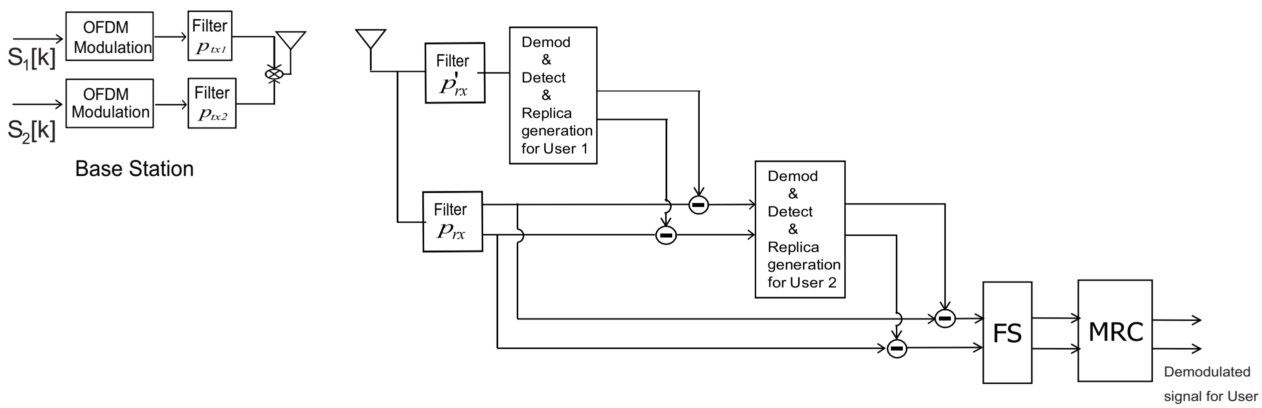

Fig. 2. Block diagram of system model.

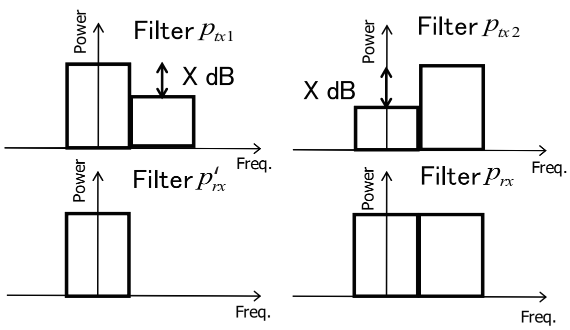

Fig. 3. Frequency response of filters.

form are given by

$$x_1(t) = \sum_{n=-N_{GI}}^{N-1} u_1[n]p_{tx1}(t-nT_s), \quad (3)$$

$$x_2(t) = \sum_{n=-N_{GI}}^{N-1} u_2[n]p_{tx2}(t-nT_s), \quad (4)$$

where $p_{tx1}(t)$ and $p_{tx2}(t)$ are the impulse responses of the respective transmit filters that include the responses of band-pass filters at D/A converters, T_s is the sampling interval for the OFDM signal, and N_{GI} is the GI length. The transmit filters for User 1 and User 2 signals have the symmetric frequency responses as shown Fig. 3. At the receiver two different baseband filters are used as shown Fig. 3. The filter $p'_{rx}(t)$ has the passband only on the channel for the desired component of User 1. The signal that passes through $p'_{rx}(t)$ is used for the tentative decision of the User 1 signal. The replica signal is regenerated, subtracted from the received signal that passes through the filter $p_{rx}(t)$. The signal for the terminal 2 is then demodulated with FS, it's replica is regenerated, and subtracted from the signal output from the filter $p_{rx}(t)$.

B. First Cancellation

The received signal through $p'_{rx}(t)$ is given as

$$y'(t) = y'_1(t) + y'_2(t) + v'(t) \quad (5)$$

where $y'_1(t)$ and $y'_2(t)$ represent the signals of User 1 and User 2 that are

$$y'_1(t) = \sum_{n=-N_{GI}}^{N-1} u_1[n]h'_1(t-nT_s), \quad (6)$$

$$y'_2(t) = \sum_{n=-N_{GI}}^{N-1} u_2[n]h'_2(t-nT_s), \quad (7)$$

and $v'(t)$ is the noise through the filter $p'_{rx}(t)$. On the other hand, the received signal through the filter $p_{rx}(t)$ is given as

$$y(t) = y_1(t) + y_2(t) + v(t) \quad (8)$$

where $y_1(t)$ and $y_2(t)$ represent the signals of User 1 and User 2 that are

$$y_1(t) = \sum_{n=-N_{GI}}^{N-1} u_1[n]h_1(t-nT_s), \quad (9)$$

$$y_2(t) = \sum_{n=-N_{GI}}^{N-1} u_2[n]h_2(t-nT_s), \quad (10)$$

and $v(t)$ is the noise through $p_{rx}(t)$. Since these signals are transmitted through a multipath channel with the impulse response of $c(t)$, $h'_1(t)$, $h'_2(t)$, $h_1(t)$ and $h_2(t)$ are the impulse responses of the composite channels that are given as

$$h'_1(t) = p_{tx1}(t) \star c(t) \star p'_{rx}(t), \quad (11)$$
$$h'_2(t) = p_{tx2}(t) \star c(t) \star p'_{rx}(t), \quad (12)$$
$$h_1(t) = p_{tx1}(t) \star c(t) \star p_{rx}(t), \quad (13)$$
$$h_2(t) = p_{tx2}(t) \star c(t) \star p_{rx}(t), \quad (14)$$

where $\star$ denotes convolution. The signal after $p_{rx}(t)$ is converted to a digital form by the A/D converter at the rate of T_s for the first cancellation. Therefore, the received digital signal is expressed as

$$y'_1[n] = y'_1(nT_s) \quad (15)$$

where T_s is the sampling rate. After removing the GI and taking the DFT to N samples, the signal on the kth subcarrier is

$$\begin{aligned} Z'[k] &= \sum_{n=-N_{GI}}^{N-1} y'_1[n] \exp\left(-\frac{j2\pi nk}{N}\right) \\ &= H'_1[k]S_1[k] + H'_2[k]S_2[k] + W'[k] \end{aligned} \quad (16)$$

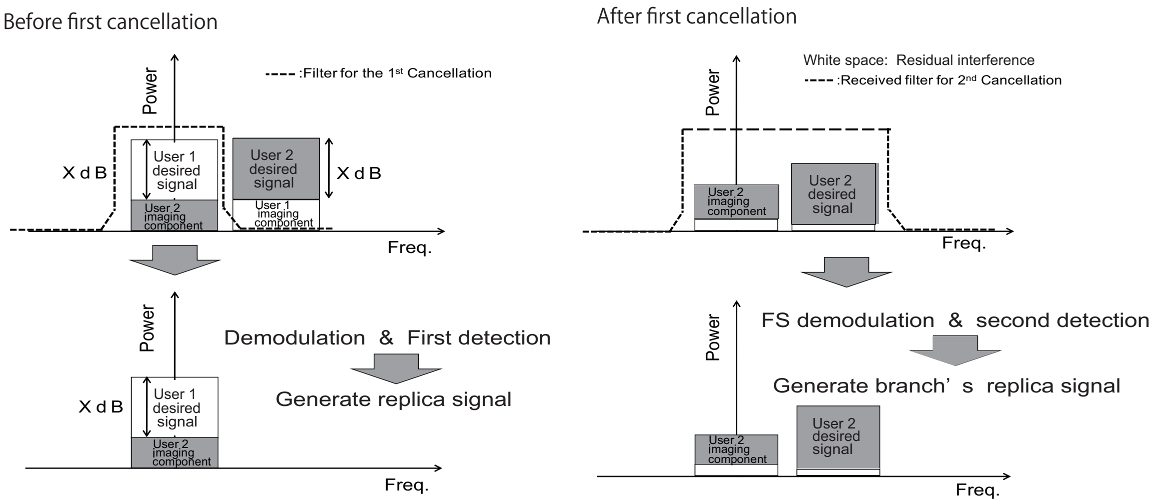

Fig. 4. Proposed cancellation scheme.

where $H_1'[k]$ and $H_2'[k]$ are the frequency responses of the User 1 and 2 signals and $W'[k]$ is the noise through $p_{rx}'(t)$ on the kth subcarrier. These are calculated as

$$H_1'[k] = \sum_{n=0}^{N-1} h_1'[n] \exp\left(-\frac{j2\pi nk}{N}\right), \quad (17)$$

$$H_2'[k] = \sum_{n=0}^{N-1} h_2'[n] \exp\left(-\frac{j2\pi nk}{N}\right), \quad (18)$$

$$W'[k] = \sum_{n=0}^{N-1} v'[n] \exp\left(-\frac{j2\pi nk}{N}\right), \quad (19)$$

where $h_1'[n]$, $h_2'[n]$, and $v'[n]$ are given by

$$\begin{aligned} h_1'[n] &= h_1'(nT_s), && (20)\\ h_2'[n] &= h_2'(nT_s), && (21)\\ v'[n] &= v'(nT_s). && (22) \end{aligned}$$

Suppose that the estimated frequency response on the kth subcarrier of the User 1 signal is expressed as $\hat{H}_1'[k]$, the demodulated signal on the adjacent channel is

$$\tilde{S}_1'[k] = \frac{\hat{H}_1'^*[k] Z'[k]}{\hat{H}_1'^*[k]\hat{H}_1'[k]}. \quad (23)$$

The symbol of the User 1 signal, $\hat{S}_1'[k]$, is detected with $\tilde{S}_1'[k]$. Therefore, the replica of the User 1 signal on all the branches of FS is given in the $G \times 1$ vector form as

$$\hat{\mathbf{Y}}_1[k] = \mathbf{H}_1[k]\hat{S}'_1[k]. \quad (24)$$

In the meantime, the signal after $p_{rx}(t)$ is converted to a digital form by the A/D converter at the rate of T_s/G with FS.

$$y_g[n,g] = \sum_{l=0}^{P-1} u_1[l]h_1[n-l,g] + \sum_{l=0}^{P-1} u_2[l]h_2[n-l,g] + v[n,g], \quad g=0,...,G-1 \quad (25)$$

where $y_g[n]$, $h_{1,g}[n]$, $h_{2,g}[n]$ and $v_g[n]$ are the polynomials of sampled $y(t)$, $h_1(t)$, $h_2(t)$ and $v(t)$, respectively, and are given as

$$\begin{aligned} y_g[n,g] &:= y(nT_s + gT_s/G), && (26)\\ h_1[n,g] &:= h_1(nT_s + gT_s/G), && (27)\\ h_2[n,g] &:= h_2(nT_s + gT_s/G), && (28)\\ v[n,g] &:= v(nT_s + gT_s/G). && (29) \end{aligned}$$

These samples are put into G DFT demodulators in parallel and the sampling rate on each branch reduces to $1/T_s$. Therefore, the images are fold down to the main channel as alias components. After removing the GI and taking DFT on each subcarrier, the received symbol is given by

$$\mathbf{Z}[k] = \mathbf{H}_1[k]S_1[k] + \mathbf{H}_2[k]S_2[k] + \mathbf{W}[k] \quad (30)$$

where $\mathbf{Z}[k] = [Z[k,0]...Z[k,G-1]]^T$, $\mathbf{W}[k] = [W[k,0]...W_[k,G-1]]^T$, $H_1[k] = [H_1[k,0]...H_1,[k,G-1]]^T$, and $\mathbf{H}_2[k] = [H_2[k,0]...H_2[k,G-1]]^T$ are $G \times 1$ column vectors, each gth component representing

$$[\mathbf{Z}[k]]_g := Z[k,g] = \sum_{n=0}^{N-1} y[n,g]e^{-j\frac{2\pi kn}{N}}, \quad (31)$$

$$[\mathbf{W}[k]]_g := W[k,g] = \sum_{n=0}^{N-1} v[n,g]e^{-j\frac{2\pi kn}{N}}, \quad (32)$$

$$[\mathbf{H}_1[k]]_g := H_1[k,g] = \sum_{n=0}^{N-1} h_1[n,g]e^{-j\frac{2\pi kn}{N}}, \quad (33)$$

$$[\mathbf{H}_2[k]]_g := H_2[k,g] = \sum_{n=0}^{N-1} h_2[n,g]e^{-j\frac{2\pi kn}{N}}, \quad (34)$$

respectively. The replica signal is subtracted from the received signal through $p_{rx}(t)$ on each FS branch and the remaining User 2 signal is expressed as

$$\begin{aligned} \bar{\mathbf{Z}}_2[k] &= \mathbf{Z}[k] - \hat{\mathbf{Y}}_1[k] \\ &= \mathbf{H}_2[k]S_2[k] + \mathbf{W}[k] + \mathbf{R}_1[k], && (35) \end{aligned}$$

where $\bar{\mathbf{Z}}_2[k] = [\bar{Z}_2[k,g]...\bar{Z}_2[k,G-1]]^T$ is the signal after the first cancellation for demodulating the User 2 signal, and $\mathbf{R}_1[k] = [R_1[k,0]...R_1[k,G-1]]^T$ is the residual interference of the User 1 signal.

C. Second Cancellation

After the first cancellation, the User 2 signal is demodulated. As already stated, $v(t)$ is the filtered noise. When $v(t)$ is sampled at the Nyquist rate of $1/T_s$, the samples of $v(t)$ are independent one another. However, when the sampling rate is a multiple of the baud rate, the noise samples are correlated. Consequently, it is necessary to whiten the colored noise samples. In order to perform noise-whitening, it is required to calculate the noise covariance matrix on each subcarrier whose (g_1, g_2)th element is given as

$$\begin{aligned} R_{W_{g1,g2}}[k] &= E[W[k,g_1]W^*[k,g_2]] \\ &= \sigma_v^2 \frac{1}{N} \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} p_{rx}((n_2 - n_1 + (g_2 - g_1)/G)T_s) \\ &\quad \times \exp\left(j\frac{2\pi k(n_2-n_1)}{N}\right) \end{aligned} \quad (36)$$

where σ_v^2 is the noise variance. Multiplying both sides of Eq. (35) by $\mathbf{R}_W^{-\frac{1}{2}}[k]$,

$$\begin{aligned} \mathbf{R}_W^{-\frac{1}{2}}[k]\bar{\mathbf{Z}}_2[k] &= \mathbf{R}_W^{-\frac{1}{2}}[k](\mathbf{H}_2[k]S_2[k] + \mathbf{W}[k] + \mathbf{R}_1[k]), \\ \tilde{\mathbf{Z}}_2[k] &= \tilde{\mathbf{H}}_2[k]S_2[k] + \tilde{\mathbf{W}}[k] + \tilde{\mathbf{R}}_1[k]. \end{aligned} \quad (37)$$

The output of the subcarrier-based maximum ratio combining (MRC) among the whitened samples is given as

$$\begin{aligned} \tilde{S}_2[k] &= \frac{\tilde{\mathbf{H}}_2^H[k]\bar{\mathbf{Z}}_2[k]}{\tilde{\mathbf{H}}_2^H[k]\tilde{\mathbf{H}}_2[k]} \\ &= \frac{(\mathbf{R}_W^{-\frac{1}{2}}[k]\hat{\mathbf{H}}_2[k])^{H_2}(\mathbf{R}_W^{-\frac{1}{2}}[k]\bar{\mathbf{Z}}[k])}{(\mathbf{R}_W^{-\frac{1}{2}}[k]\hat{\mathbf{H}}_2[k])^{\mathbf{H}_2}(\mathbf{R}_W^{-\frac{1}{2}}[k]\hat{\mathbf{H}}_2[k])} \end{aligned} \quad (38)$$

where $\hat{\mathbf{H}}_2[k] = [\hat{H}_2[k,0]...\hat{H}_2[k,G-1]]^T$ is the estimated frequency response on the kth subcarrier for the User 2 signal. The symbol of the User 2 signal, $\hat{S}_2[k]$ is detected based on $\tilde{S}_2[k]$. From $\hat{S}_2[k]$, the replica signal on each branch for cancelling the User 1 signal is generated in the $G \times 1$ vector form as

$$\hat{\mathbf{Y}}_2[k] = \mathbf{H}_2[k]\hat{S}_2[k], \quad (39)$$

The replica signal are subtracted from the received signal as

$$\begin{aligned} \mathbf{Z}_1[k] &= \mathbf{Z}[k] - \hat{\mathbf{Y}}_2[k] \\ &= \mathbf{H}_1[k]S_1[k] + \mathbf{W}_1[k] + \mathbf{R}_2[k], \end{aligned} \quad (40)$$

where $\mathbf{Z}_1[k] = [Z_1[k,0]...Z_1[k,G-1]]^T$ is the signal after the second cancellation for demodulating the User 1 signal, and $\mathbf{R}_2[k] = [R_2[k,0]...R_2[k,G-1]]^T$ is the residual interference of the User 2 signal. Finally, the User 1 signal is demodulated with the same manner as Eqs. (38)-(40).

III. Numerical Results

A. Simulation Conditions

Simulation conditions are presented in TABLE I. The symbols are modulated with QPSK for both signals of the User 1 and User 2 , and multiplexed with OFDM. The numbers of data subcarriers and pilot subcarriers are 48 and 4 while the DFT size is 64. The bandwidth of the subcarrier is 312.5 kHz. The oversampling rate, G, is 1 or 2, the number of packets per trial is 100000, and the number of OFDM symbols per packet is 1. The channel responses of the subcarriers are estimated with the preamble symbols at the beginning of the packet. The orthogonal symbols for channel estimation may be transmitted like a MIMO system and the channel responses for the signal of User 1 and User 2 are estimated, respectively. For simulation ideal channel estimation is assumed. As a channel model, 16 path Rayleigh fading with uniform delay profile is employed [9].

TABLE I
SIMULATION CONDITIONS

Modulation scheme	1st:QPSK 2nd:OFDM
Number of subcarriers	64
Number of data subcarriers	48
Bandwidth of subcarriers	312.5[kHz]
Preamble length (GI+Preamble)	$1.6 + 6.4[\mu s]$
OFDM symbol length (GI+Data)	$0.8 + 3.2[\mu s]$
Number of OFDM symbols per packet	1
Number of OFDM packets per trial	100000
Oversampling rate	$G = 1, 2$
Channel model	Rayleigh (16 path)
Channel estimation	Ideal

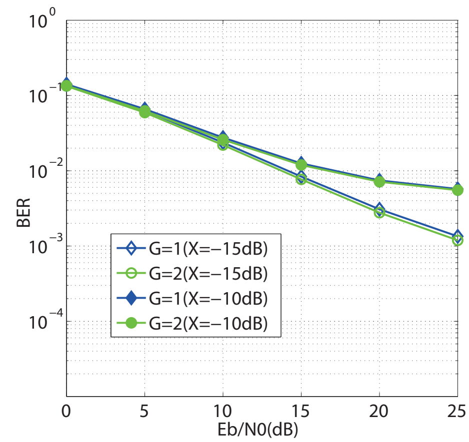

Fig. 5. BER vs. E_b/N_0(Before first cancellation).

B. Before First Cancellation

Figure 5 shows the relationship between the BER and E_b/N_0 on the 16 path Rayleigh fading channel before the first cancellation. The power difference between the desire and imaging components is set to X=15dB, or X=10dB. Obviously, the BER with X=15dB is smaller than that with X=10dB.

C. After Second Cancellation (X=15dB)

Figure 6 shows the relationship between the BER and E_b/N_0 for X=15dB on the 16 path Rayleigh fading channel after the second cancellation. If the power difference is 15dB, the replica of the User 2 signal is regenerated accurately and the interference to the User 1 signal is eliminated with the proposed scheme.

D. After Second Cancellation (X=10dB)

Figure 7 shows the relationship between the BER and E_b/N_0 for X=10dB on the 16 path Rayleigh fading channel

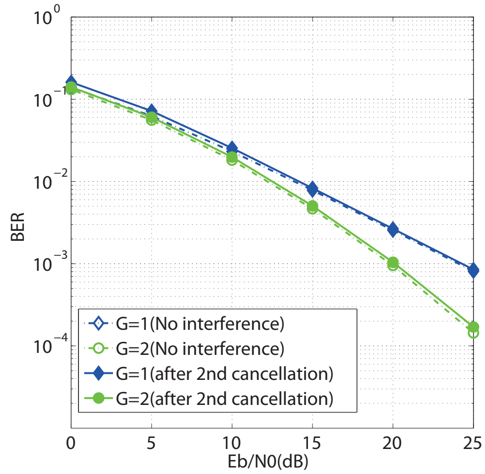

Fig. 6. BER vs. E_b/N_0(after second cancellation (X=15dB)).

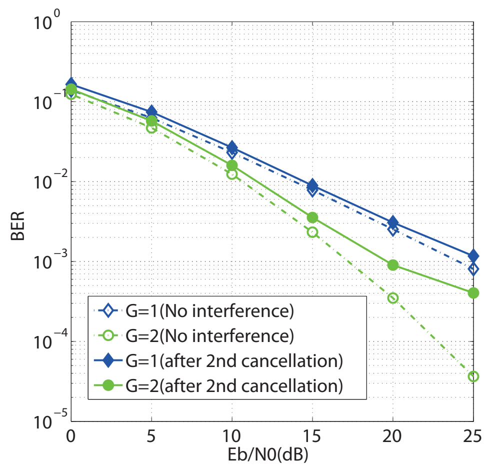

Fig. 7. BER vs. E_b/N_0(after second cancellation (X=10dB)).

after the second cancellation. In this case, the signal-to-interference and noise ratio (SINR) is lower before the first cancellation. Lower SINR leads to higher error rates on the tentative detection of the User 1 signal even though more diversity gain can be expected with larger imaging components. The interference due to the detection errors then increases and, the BER curves exhibit error floors in the proposed scheme.

E. BER versus X (after second cancellation)

Figure 8 shows the relationship between the BER and the power difference X on the 16 path Rayleigh fading channel after the second cancellation. E_b/N_0 is set to 20dB. When X is close to 12dB, the BER is the smallest. If X is larger than 12dB, the BER increase because the SINR is low before the first cancellation, and the more tentative detection errors occur. If X is smaller than 12dB, the BER also grows. In this case the proposed scheme can not improve the BER, this is because the power of the imaging components for diversity is very small.

IV. Conclusions

In this paper, non-orthogonal access over multiple channels with IIC and FS is proposed. In the proposed scheme, the replica signal is regenerated and subtracted from the received signal for each desired signals. The proposed scheme improves the BER performance when the power difference between the desire and imaging components is larger. Numerical results through computer simulation have shown that the power of the imaging components should be large enough for diversity at the same time as keeping errors on the tentative decisions small in order to avoid the error floors.

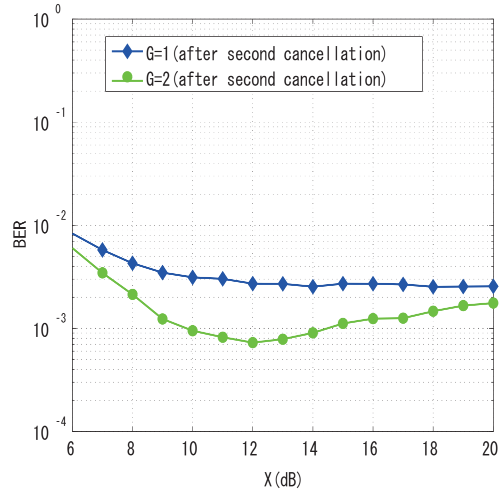

Fig. 8. BER vs. X (after second cancellation (E_b/N_0=20dB))

Acknowledgement

This work is supported in part by a Grant-in-Aid for the Global Center of Excellence for high-Level Global Cooperation for Leading-Edge Platform on Access Spaces.

References

[1] R. Nee, G Awater, M. Morikura, H. Takanashi, M. Webster and K. Halford, "New High-Rate Wireless LAN Standards," IEEE Communications Magazine, Dec. 1999.

[2] S. Weinstein, "The History of Orthogonal Frequency Division Multiplexing," IEEE Trans. on Broadcasting, IEEE Communications Magazine, pp.26-35, Nov. 2009

[3] A. N. Barreto, "Antenna Transmit Diversity for Wireless OFDM Systems," IEEE Vehicular Technology Conference, 2002 Spring, vol.2, pp.757-761, Nov. 2002.

[4] J. Salz, J. H. Winters, "Effect of Fading Correlation on Adaptive Arrays in Digital Mobile Radio," IEEE Trans. Vehicular Technology, vol.43, no.4, pp.1049-1057, Nov. 1994.

[5] R. C. Kizilirmak and Y. Sanada, "Multipath Diversity through Time Shifted Sampling for Spatially Correlated OFDM-Antenna Array Systems," IEICE Trans. on Fundamentals, vol.E91-A, no.11, pp.3104-3111, Nov. 2008.

[6] C. Tepedelenlioĝlu and R. Challagulla, "Low-Complexity Multipath Diversity Through Fractional Sampling in OFDM," IEEE Trans. on Signal Processing, vol.52, no.11, pp.3104-3116, Nov. 2004.

[7] H. Osada, H. Nishimura, M. inamori and Y. Sanada, "Adjacent channel interference cancellation in Fractional Sampling OFDM Receiver," IEEE Vehicular Technology Conference Fall, Sept. 2011.

[8] T. Shinkai, H. Nishimura, and Y. Sanada, "Improvement on Diversity Gain with Filter Bandwidth Enlargement in Fractional Sampling OFDM Receiver," IEICE Trans. on Communication, vol.E93-B, no.6, pp.1526-1533, June 2010.

[9] Joint Technical Committee of Committee T1 R1P1.4 and TIA TR46.3.3/TR45.4.4 on Wireless Access, "Draft Final Report on RF Channel Characterization," Paper No.JTC(AIR)/94.01.17-238R4, Jan. 1994.