2017 International Symposium on Nonlinear Theory and Its Applications,
NOLTA2017, Cancun, Mexico, December 4-7, 2017

Compensation of Deteriorated CSI Allocation Based on Decision Direct and Turbo Equalization for MUDiv/OFDMA

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Abstract—In a wireless propagation channel, the channel for each user is independent. By using this characteristic, orthogonal frequency division multiple access (OFDMA) with the multiuser diversity (MUDiv) effect is achieved. However, in the fast fading channel, since channel state information (CSI) is changed between the head and latter of packets, the system performance is degraded. Moreover, in this case, CSI is also changed between the cases with and without the feedback delay. To solve these problems, in this paper, we propose the compensation of the deteriorated CSI allocation based on the decision direct and the turbo equalization for MUDiv/OFDMA systems.

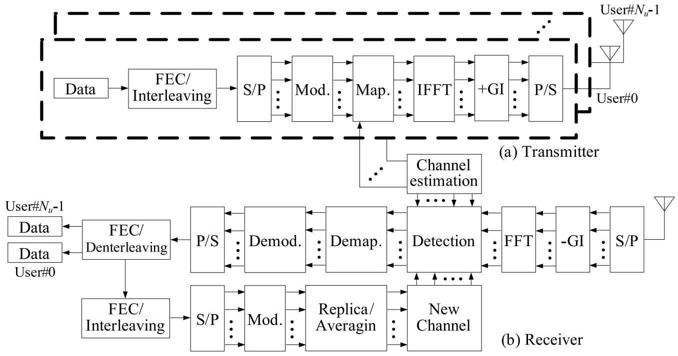

Figure 1: Block diagram of the proposed system.

1. Introduction

Orthogonal frequency division multiplexing (OFDM) and OFDM access (OFDMA) are standardized in mobile, wireless, and broadband systems [1]. OFDM keeps the orthogonality between the adjacent subcarriers and achieves high capacity transmissions. OFDMA divides the subcarrier for several users and allocates it in the frequency band. As a result, OFDMA deals with the OFDM signal for each user in the same time and achieves orthogonal multiple access (OMA) systems [2].

In a wireless propagation channel, channel variance is occurred due to a multipath Rayleigh fading. In this case, since the channel for each user is independent, the channel variance for each user is also independent. By using this characteristic, OFDMA with the multiuser diversity (MUDiv) effect has been proposed [3]. In this paper, we assume that the subcarrier for each user is allocated based on the channel state information (CSI) in the head of packet. Moreover, we assume also that the subcarrier for each user is allocated in the good subcarrier channel and the number of subcarriers for each user is equal. In a MUDiv/OFDMA, the block allocation has been also proposed to mitigate the complexity. On the other hand, in OFDM systems, many error is occurred at a high Doppler frequency [4]. This is because CSI is changed between the head and latter of packets due to the fast fading channel. MUDiv/OFDMA has also the same problem and the MUDiv effect is destroyed due to the fast fading channel. Moreover, in this case, if the feedback delay is occurred, the MUDiv effect is more destroyed. This is because CSI is also changed between the cases with and without the feedback delay [5]. As a result, large deterioration of the system performance is occurred due to the fast fading channel and the feedback delay. Before now, we have proposed the decision direct method to compensate the fast fading channel [6]. Moreover, turbo equalization is also effective in the fast fading channel [7]. Therefore, in this paper, we propose the compensation of the deteriorated CSI allocation based on the decision direct and the turbo equalization for MUDiv/OFDMA systems.

2. MUDiv/OFDMA

2.1. Channel Model

For the channel model, the channel impulse response for the user n consists of L discrete paths with the different delay time as

$$h_n(\tau) = \sum_{l=0}^{L-1} h_{n,l}\delta(\tau - \tau_{n,l}), \tag{1}$$

where $h_{n,l}$ and $\tau_{n,l}$ are the complex channel gain and the delay time of the lth propagation path with $\sum_{l=0}^{L-1} E|h_{n,l}^2| = 1$, and $E|\cdot|$ is the ensemble average operation. After the fast Fourier transform (FFT) operation, the channel response for the user n is given by

$$H_n(k) \quad = \quad \int_0^{\infty} h_n(\tau)\exp(-j2\pi k\tau)d\tau$$

$$= \sum_{l=0}^{L-1} h_{n,l} \exp(-j2\pi k \tau_{n,l}). \quad (2)$$

2.2. Subcarrier Allocation with MUDiv Effect

The channel variance for each user base on Eq. (2) is independent since the channel for each user is independent. By using this characteristic, the subcarrier allocation based on the MUDiv effect is defined by

$$J = \max \left(\sum_{n=0}^{N_u-1} \sum_{k=0}^{\lfloor N_c/B \rfloor -1} \sum_{b=0}^{B-1} \frac{|H_n(Bk+b,0)|}{B} \alpha_n(k) \right), \quad (3)$$

where $\alpha_n(k)$ is the selection parameter as

$$\alpha_n(k) = \begin{cases} 1 & \text{for } \textit{allocation} \\ 0 & \text{for } \textit{no allocation}, \end{cases} \quad (4)$$

N_c and N_u are the number of subcarriers and users, respectively, $H_n(k,i)$ is the channel response for the kth subcarrier and the ith symbol, B is the block length as $\lfloor N_c/N_u \rfloor$, $\lfloor \beta \rfloor$ stands for the integer lower and closer to β, and $\max(\gamma)$ is the maximum value of γ. In this paper, since we assume that the channel response $H_n(k,i)$ is only known for $i = 0$, the channel response $H_n(k,0)$ is utilized in Eq. (3).

2.3. MUDiv/OFDMA

At the transmitter, as shown in Figure 1(a), the binary data signal is coded by the convolutional code with the interleaving. After the serial-to-parallel (S/P) conversion, the coded signal is modulated as $d_n(k,i)$ and is mapped in the frequency band as

$$x_n(k,i) = \alpha_n(k) d_n(k,i). \quad (5)$$

The mapped signal $x_n(k,i)$ is converted to the time domain signal by the inverse FFT (IFFT) operation and guard interval (GI) is inserted. After the parallel-to-serial (P/S) conversion, the time domain signal is given by

$$x_n(t) = \sum_{i=0}^{N_d-1} g(t - iN_{cg}) \cdot \left\{ \sqrt{\frac{2S}{N_c}} \sum_{k=0}^{N_c-1} x_n(k,i) \exp[j2\pi(t - iN_{cg})k]/N_c \right\}, \quad (6)$$

where $g(t)$ is the transmission pulse as

$$g(t) = \begin{cases} 1 & \text{for } -N_g \le t \le N_c \\ 0 & \text{otherwise}, \end{cases} \quad (7)$$

S is the average transmission power, N_d is the number of data symbols, N_g is the GI length, and N_{cg} is the symbol length as $N_c + N_g$.

At the receiver, as shown in Fig. 1(b), the time domain received signal is given by

$$y(t) = \sum_{n=0}^{N_u-1} \int_{-\infty}^{\infty} h_n(\tau) x_n(t-\tau) d\tau + z(t), \quad (8)$$

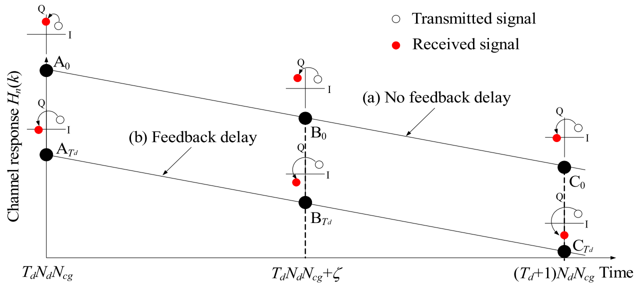

Figure 2: Channel response with and without the feedback delay at a high Doppler frequency.

where $z(t)$ is an additive white Gaussian noise (AWGN) with a single side power spectral density of N_0. After the S/P conversion and GI elimination, the time domain signal is converted to the frequency domain signal by the FFT operation as

$$\begin{aligned} y(k,i) &= \frac{1}{N_c} \int_{iN_{cg}}^{iN_{cg}+N_c} y(t) \exp\left[\frac{-j2\pi(t - iN_{cg})k}{N_c}\right] dt \\ &= \sqrt{\frac{2S}{N_c}} \sum_{n=0}^{N_u-1} H_n(k,i) x_n(k,i) + z(k,i), \end{aligned} \quad (9)$$

where $z(k,i)$ is an AWGN noise with zero-mean and variance $2N_0/N_c$. The frequency domain received signal $y(k,i)$ is detected and is demapped as

$$\begin{aligned} \tilde{d}_n(k,i) &= \sum_{n=0}^{N_u-1} \lambda_n^{-1}(k,i) \alpha_n(k) y(k,i) \\ &= \sum_{n=0}^{N_u-1} \alpha_n(k) \tilde{x}_n(k,i) \ \text{for} \ \alpha_n(k) = 1, \end{aligned} \quad (10)$$

where $\tilde{x}_n(k,i)$ is the detected signal before the demapping and $(\cdot)^{-1}$ is the inverse operation. $\lambda_n(k,i)$ is the channel weight and will be shown in the next section. After the P/S conversion, the demapped signal is returned to the binary data signal.

3. Proposed Compensation of Deteriorated CSI Allocation

Figure 2 shows the example of the channel response with and without the feedback delay at a high Doppler frequency. In a small Doppler frequency and no feedback delay, the MUDiv effect is kept as $H_n(k,0) \approx H_n(k,N_d)$. However, in a high Doppler frequency, since CSI between the head and latter of packets is changed, the system performance is degraded due to the fast fading as shown in Fig. 2(a). Moreover, in this case, if the feedback delay is occurred, CSI is more changed as shown in Fig. 2(b). Therefore, the proposed method achieves the compensation for the fast fading and the feedback delay.

For $i < \zeta$, since the changing for the CSI is a small, the channel weight $\lambda_n(k,i)$ between $0 \le i \le \zeta - 1$ adapts the known channel responses $H_n(k,i)$ for $i = 0$ as

$$\lambda_n(k,i) = H_n(k,0), \tag{11}$$

where ζ is the compensated interval. On the other hand, when $i \ge \zeta$, the changing of the CSI is a large. To solve this problem, the proposed method adapts the decision direct and the channel weight $\lambda_n(k,i)$ between $\zeta \le i \le 2\zeta - 1$ is given by

$$\begin{aligned}\lambda_n(k,i) &= \sum_{n=0}^{N_u-1} \frac{|y(k,\zeta)|^2}{y^*(k,\zeta)\tilde{x}_n(k,\zeta)} \\ &= \sum_{n=0}^{N_u-1} (H_n(k, i + T_d N_d) + \omega_n(k, i + T_d N_d)) \\ &\qquad \text{for } \alpha_n(k) = 1, \end{aligned} \tag{12}$$

where T_d is the feedback delay, $\omega_n(k,i)$ is the detection error, and $(\cdot)^*$ is a conjugate operation. Observing Eq. (12), the proposed method obtains the channel weight between $\zeta \le i \le 2\zeta - 1$. However, since the proposed method adapts the detected signal $\tilde{x}_n(k,\zeta)$, the detection error $\omega_n(k,i)$ is included. To mitigate this error, the proposed method is averaged as

$$\lambda_n(k,i) = \sum_{n=0}^{N_u-1} \frac{\sum_{i_2=0}^{N_{ave}-1} |y(k,\zeta - i_2)|^2}{\sum_{i_1=0}^{N_{ave}-1} y^*(k,\zeta - i_1)\tilde{x}_n(k,\zeta - i_1)} \quad \text{for } \alpha_n(k) = 1, \tag{13}$$

where N_{ave} is the number of symbols for the averaging. And then, the new detected signals $\hat{x}_m(k,i)$ and $\hat{d}_m(k,i)$ are obtained by using Eqs. (10) and (13). Moreover, to more reduce the detection error $\omega_n(k,i)$, the proposed method obtains the channel weight $\lambda_n(k,i)$ again between $\zeta \le i \le 2\zeta - 1$ by using the detected signal $\hat{x}_m(k,i)$. This operation iterates N_t times by using the turbo equalization. For $i \ge 2\zeta$, the proposed method performs the same processing in the interval ζ and compensates the fast fading and the feedback delay.

4. Computer Simulation Results

In this section, we evaluate the system performance of the proposed method by using the computer simulation. Table 1 shows the simulation parameters. In this simulation, we assume 4 users OFDMA systems. Moreover, the propagation channel is that the symbol duration is 10 μs, the GI length is 2 μs, the path model is 15 paths Rayleigh fading at a Doppler frequency of 5 and 200 Hz, and the feedback delay is 0, 3, and 6 packet time. As shown in Figure 1, at the transmitter, the original data signal is coded by the convolutional code for the rate 1/2 and the constraint length 7 with the interleaving. After the S/P conversion, the parallel signal is modulated by a quadrature phase shift keying (QPSK) and is allocated based on Eq. (3) in the frequency band. The time domain signal is generated by the IFFT operation and GI is inserted. After the P/S conversion, the time domain signal is transmitted to the receiver. At the receiver, GI is eliminated and is converted to the frequency domain signal by the FFT operation after the S/P conversion. The proposed method estimates the channel weight by using the decision direct and the turbo equalization based on Eqs. (10) and (13), and the frequency domain signal is detected by the zero-forcing (ZF) equalization. After the demapping and the demodulation, the detected signal is decoded by the Viterbi soft decoding algorithm with the deinterleaving.

Fig. 3 shows the bit error rate (BER) performance for the conventional and proposed methods at $(f_d, T_d) = (5,0)$, $(200,0)$, and $(200,3)$, where f_d is a Doppler frequency. The conventional method for $(f_d, T_d) = (200,0)$ shows about 9 dB penalty compared with $(f_d, T_d) = (5,0)$ due to the fast fading channel. The conventional method for $(f_d, T_d) = (200,3)$ shows about 11.5 and 8 dB penalties compared with $(f_d, T_d) = (5,0)$ and $(200,0)$ due to the fast fading channel and the feedback delay. The proposed method for $(f_d, T_d, \zeta, N_t) = (200,3,5,1)$ shows about 10 dB gain compared with the conventional method for $(f_d, T_d) = (200,3)$. This is because the proposed method approximates the CSI in the head of packet without the feedback delay by using the decision direct. The proposed method for $(f_d, T_d, \zeta, N_t) = (200,3,5,4)$ shows about 4 dB gain compared with $(f_d, T_d, \zeta, N_t) = (200,3,5,1)$. This means that the proposed method achieves more approximated CSI by using the turbo equalization.

Fig. 4 shows the BER performance for the conventional and proposed methods at $(f_d, T_d) = (5,0)$, $(200,0)$, and $(200,6)$. The conventional method for $(f_d, T_d) = (200,6)$ shows about 13 and 11 dB penalties compared with $(f_d, T_d) = (5,0)$ and $(200,0)$. This is because the conventional method for $(f_d, T_d) = (200,6)$ has a large feedback delay compared with $(f_d, T_d) = (200,3)$ for Fig. 3. The

Table 1: Simulation parameters.

Data modulation	QPSK
Data detection	Coherent
Symbol duration	10 μs
Number of data symbols	20
Number of subcarriers	64
Number of users	4
FFT size	64
Block length	16
Guard interval	16 sample time
Path model	15 paths Rayleigh fading
Doppler frequency	5, 200 Hz
Feedback delay	0, 3, 6 packet time
Number of symbols for averaging	$N_{ave} = 5$
FEC	Convolutional code (R= 1/2, $\mathcal{K} = 7$)

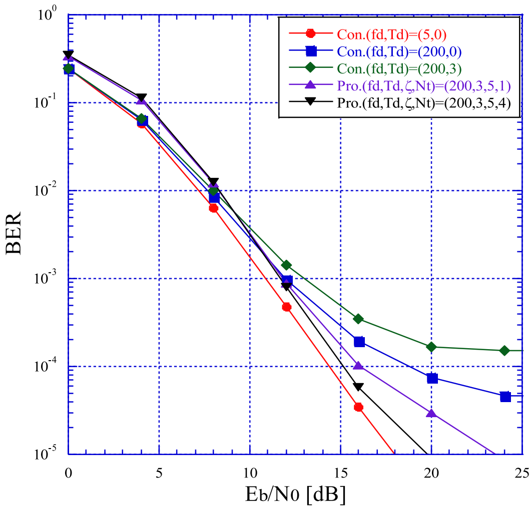

Figure 3: BER performance for the conventional and proposed methods at $(f_d, T_d) = (5, 0)$, $(200, 0)$, and $(200, 3)$.

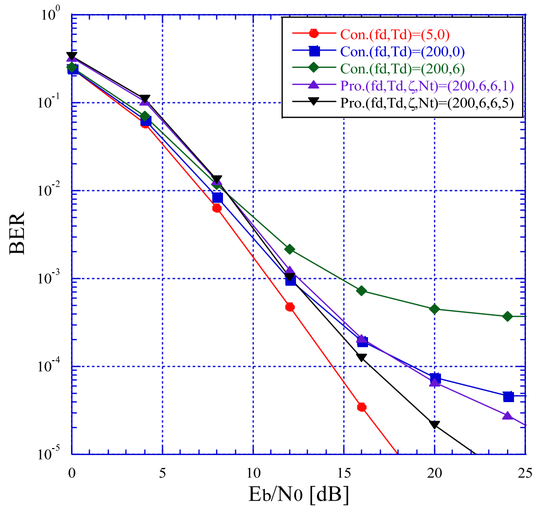

Figure 4: BER performance for the conventional and proposed methods at $(f_d, T_d) = (5, 0)$, $(200, 0)$, and $(200, 6)$.

proposed method for $(f_d, T_d, \zeta, N_t) = (200, 6, 6, 1)$ shows about 11 dB gain compared with the conventional method for $(f_d, T_d) = (200, 6)$. Moreover, the proposed method for $(f_d, T_d, \zeta, N_t) = (200, 6, 6, 5)$ shows about 5 dB gain compared with $(f_d, T_d, \zeta, N_t) = (200, 6, 6, 1)$. These mean that the proposed method is also effective in $T_d = 6$.

5. Conclusion

In this paper, we have proposed the compensation of the deteriorated CSI allocation based on the decision direct and the turbo equalization for MUDiv/OFDMA systems. The proposed method compensates the deteriorated CSI due to the fast fading channel and the feedback delay by using the decision direct and the turbo equalization. From the computer simulation results, the proposed method has shown the good BER performance in $(f_d, T_d) = (200, 3)$ and $(200, 6)$. In the future plan, we reduce the complexity due to the repeat processing.

Acknowledgments

This work was partly supported by JSPS KAKENHI Grant-in-Aid for Young Scientists (B) Number 15K18065.

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