Interleaved block subcarrier allocation and power combination for frequency symbol spreading multiuser diversity OFDMA

Yuta Ida¹* and Takahiro Matsumoto²

Abstract

In a FSS-OFDMA, the OFDM symbol is spread, and the subcarriers are assigned for each user in the frequency band. As the principal subcarrier allocation for each user, localized and interleaved allocations have been proposed, and the interleaved allocation is improved for the BER performance to avoid continuous deep faded subcarrier channels. Also, in the independent channel fluctuation for each user, the MUDiv gain is obtained by the individual and block subcarrier allocations. However, they have a trade-off problem between the BER performance and the computational complexity. To solve this problem, the interleaved block subcarrier allocation has been proposed. On the other hand, the interleaved block is only evaluated for the individual subcarrier and is not evaluated for the several subcarriers. Moreover, the power combination between the subcarrier and the channel is also improved for the BER performance, but the individual and block subcarrier allocations are only evaluated. Therefore, in this paper, we propose the interleaved block subcarrier allocation and the power combination for a FSS-MUDiv/OFDMA.

Keywords: OFDMA, FSS, MUDiv, Interleaved block subcarrier allocation, Power combination

1 Introduction

1.1 Introduction

OFDM is very important technique to achieve high capacity transmissions by multicarrier systems [1]. Therefore, OFDM has been adopted in many standards such as WLAN, digital broadcasting, and mobile communication systems [2–4]. Also, as the access technique with an OFDM, OFDMA has been proposed and has been also adopted in these techniques. In recent access technique, OMA and NOMA are discussed [5]. In a NOMA, the capacity is expanded by assigning the signal for each user non-orthogonally, but the elaborate signal separation and detection such as the SIC are required. On the other hand, OFDMA is classified as an OMA, and the orthogonally between subcarriers and users is kept in the frequency band. Therefore, the signal separation and detection are achieved linearly such as the ZF and the MMSE. Moreover, in an OFDM, FSS-OFDM has been proposed and spreads the OFDM symbol by the spreading technique such as
the WHT and the DFT in the frequency band [6–9]. And then, in a FSS-OFDMA, the subcarrier allocation for each user is very important [10]. As the principal subcarrier allocation for each user, localized and interleaved allocations have been proposed. In the localized allocation, the subcarriers for each user are assigned continuously in the frequency band. This is a simple allocation. However, when continuous deep faded subcarrier channels occurred, the BER performance is deteriorated due to a burst error. On the other hand, in the interleaved allocation, the subcarriers for each user are assigned alternately in the frequency band [11–13]. In this case, the BER performance is improved compared with the localize allocation since a burst error is prevented by avoiding continuous deep faded subcarrier channels.

1.2 Related research status, problems, and this paper’s contribution

In Table 1, the related research states, problems, and this paper’s contribution are summarized. In a wireless propagation channel, the channel fluctuation occurred due to a multipath fading at the frequency band [14]. And then, since the propagation channel for each user is independent, the channel fluctuation for each user is also independent. By using this characteristic, the MUDiv gain is obtained by assigning the subcarrier based on the CSI for each user [15]. In this paper, the subcarriers for each user are assigned based on the following condition that the number of subcarriers for each user is equal, and the BER performance is improved in all users. In this case, the individual and block subcarrier allocations have been proposed [16, 17]. The individual subcarrier allocation obtains the good BER performance compared with the block subcarrier allocation. However, large computational complexity is required. On the other hand, in the block subcarrier allocation, the block consists of the continuous several subcarriers, and the computational complexity is reduced. However, its BER performance is deteriorated compared with the individual subcarrier allocation. As a result, since they have a trade-off problem, we have proposed the following method to solve this problem [18–20]. As mentioned in Sect. 1.1, a burst error occurred due to a deep faded subcarrier channel. In [18], deep faded subcarrier channels are avoided by using cooperative communications, but several relay nodes are required. On the other hand, in [19, 20], the subcarriers for each user are assigned by the interleaved allocation as the same [11–13], and the MUDiv gain is obtained. In this case, deep faded subcarrier channels are avoided without cooperative communications. However, in [19], large computational complexity is required. Also, in [20], the interleaved block is

| Method | Interleaved allocation | Flexibility | MUDiv gain | Computational complexity |
|--------|------------------------|-------------|------------|-------------------------|
| Ref. [15] | X | X | O | Large |
| Refs. [16, 18, 21, 22] | X | O | O | Small to large |
| Refs. [11–13] | O | O | X | Small |
| Refs. [17, 19] | O | O | O | Large |
| Ref. [20] | O | X | O | Small |
| Refs. [23, 24] | X | O | X | Small |
| Proposed method | O | O | O | Small |
only evaluated for the individual subcarrier, and its flexibility is not evaluated for the several subcarriers. Moreover, the power combination between the subcarrier and the channel has been proposed [21, 22]. It improves also the BER performance, but the individual and block subcarrier allocations are only evaluated. As the other method to solve a deep faded subcarrier channel, WF is also effective, but the adaptive threshold is required [23, 24]. Therefore, in this paper, we propose the interleaved block subcarrier allocation and the power combination for a FSS-MUDiv/OFDMA. The proposed method is the new FSS-OFDMA with the interleaved subcarrier allocation, and the utilization for the conventional FSS-OFDMA with the interleaved subcarrier allocation is introduced in the existing standards [9, 12, 13]. And then, since the proposed method is achieved by changing the subcarrier allocation method from the conventional method, authors think that the proposed method is compatible with existing standards.

1.3 Notation and this paper’s organization
In Table 2, we define the notation for the symbols. And then, the rest of this paper is organized as follows. In Sect. 2, we present the system model for a FSS-MUDiv/OFDMA. Then, in Sect. 3, we indicate the methods for the conventional individual and block subcarrier allocations, and the proposed interleaved block subcarrier allocation.
allocation and power combination. And then, we discuss the system performance for the conventional and proposed methods by using the computer simulation in Sect. 4. Finally, we give the conclusion in Sect. 5.

2 System model

In this section, we indicate the system model for a FSS-MUDiv/OFDMA. Here, in Fig. 1, the block diagram of the proposed system is shown.

2.1 WHT and DFT

In a FSS-OFDM, the OFDM symbol is spread. In this paper, WHT and DFT are utilized as the spreading technique. WHT and DFT are expressed as the \((\hat{N}_c \times \hat{N}_c)\) matrix. Firstly, the WHT matrix is defined by

\[
W_{\hat{N}_c} = \begin{bmatrix}
W_{\hat{N}_c/2} & W_{\hat{N}_c/2} \\
W_{\hat{N}_c/2} & -W_{\hat{N}_c/2}
\end{bmatrix},
\]

where \(W_1 = 1\). Next, the DFT matrix is defined by

\[
F_{\hat{N}_c} = \begin{bmatrix}
\omega^0_{\hat{N}_c} & \omega^0_{\hat{N}_c} & \cdots & \omega^0_{\hat{N}_c} \\
\omega^0_{\hat{N}_c} & \omega^1_{\hat{N}_c} & \cdots & \omega_{\hat{N}_c-1} \\
\vdots & \vdots & \ddots & \vdots \\
\omega^0_{\hat{N}_c} & \omega_{\hat{N}_c-1} & \cdots & \omega^1_{\hat{N}_c}
\end{bmatrix},
\]

where \(\omega^k_{\hat{N}_c} = \exp(-j2\pi k/\hat{N}_c)\), and \(j = \sqrt{-1}\).

2.2 Channel model

Firstly, the CIR for the \(u\)th user is constituted by

\[
h_{u}(\tau) = \sum_{l=0}^{L-1} h_{u,l} \delta(\tau - \tau_{u,l}),
\]

where \(\sum_{l=0}^{L-1} E[|h_{u,l}|^2] = 1\). And then, the CIR is converted to the CR by the FFT operation as

![Block diagram of the proposed system](image-url)
2.3 Transmitter and receiver

Figure 1a shows the block diagram of the proposed system for the transmitter. Firstly, the original binary data signals are coded by the convolutional code with an interleaving. Next, after the S/P conversion, the parallel signals are modulated to $x_u(k, i)$ between the $k$th subcarrier and the $i$th symbol for the $u$th user. Here, its matrix form for the $i$th symbol is defined by

$$X_u(i) = [x_u(0, i), \ldots, x_u(\tilde{N}_c - 1, i)]^T. \quad (5)$$

In a FSS-OFDM, the $i$th OFDM symbol for Eq. (5) is spread by using $s_u(k, i)$ between the $k$th subcarrier and the $i$th symbol for the $u$th user as

$$S_u(i) = [s_u(0, i), \ldots, s_u(\tilde{N}_c - 1, i)]^T = \frac{1}{\sqrt{\tilde{N}_c}} \Phi_{\tilde{N}_c} X_u(i). \quad (6)$$

Here, if $\Phi_{\tilde{N}_c} = W_{\tilde{N}_c}$, the OFDM symbols are spread by the WHT. Also, if $\Phi_{\tilde{N}_c} = F_{\tilde{N}_c}$, the OFDM symbols are spread by the DFT. After the symbol spreading, the spreading signals are mapped to the frequency band as

$$m_u(k, i) = \begin{cases} s_u(k, i) & \text{for } \gamma_u(k) = 1 \\ 0 & \text{for } \gamma_u(k) = 0, \end{cases} \quad (7)$$

where $\gamma_u(k)$ is the selection parameter as shown in Eq. (15). The mapping operation is indicated in Sect. 3. In the mapped signal, GI is inserted after the $N_c$ points IFFT operation. Finally, after the P/S conversion, the time domain signal for the $u$th user is sent to the receiver as

$$m_u(t) = \sqrt{2S} \sum_{i=0}^{N_u-1} \sum_{k=0}^{N_c-1} m_u(k, i) \exp \left( j2\pi \frac{kt}{N_c} \right). \quad (8)$$

Figure 1b shows the block diagram of the proposed system for the receiver. Firstly, the time domain signal is given by

$$n(t) = \sum_{u=0}^{N_u-1} \int_{-\infty}^{\infty} h_u(\tau)m_u(t - \tau)d\tau + w(t), \quad (9)$$

where $w(t)$ is AWGN with a single side power spectral density of $N_0$. Next, after the S/P conversion and the GI elimination, the time domain signal is converted to the frequency domain signal by the $N_c$ points FFT operation as

$$H_u(k) = \int_{0}^{\infty} h_u(\tau) \exp(-j2\pi k \tau)d\tau = \sum_{l=0}^{L-1} h_{u,l} \exp(-j2\pi k \tau_{u,l}). \quad (4)$$
\[
n(k, i) = \sqrt{\frac{2S}{N_c}} \sum_{t=N/2}^{(i+1)N_c-1} n(t) \exp \left( -j2\pi \frac{kt}{N_c} \right)
\]
(10)

where \( w(k, i) \) is AWGN with zero-mean and variance \( 2N_0/N_c \). Here, we assume \( H_u(k) = H_u(k, i) \) since \( H_u(k, 0) \approx H_u(k, 1) \approx \cdots \approx H_u(k, N_u - 1) \). And then, the frequency domain signals are detected by using the weight \( \omega(k) \) for the ZF or the MMSE and are demapped as

\[
\tilde{s}_u(k, i) = \omega(k)\gamma_u(k)n(k, i) = \begin{cases} 
\frac{\gamma_u(k)n(k, i)}{|H_u(k)|^2 + N_0} & \text{for ZF} \\
\frac{\gamma_u(k)n(k, i)}{|H_u(k)|^2 + N_0} & \text{for MMSE.}
\end{cases}
\]
(11)

Observing Eq. (11), the demapped signals for the \( u \)th user are obtained in \( \gamma_u(k) = 1 \), and its matrix form is expressed as

\[
\tilde{\mathbf{s}}_u(i) = \left[ \tilde{s}_u(0, i), \ldots, \tilde{s}(N_c - 1, i) \right]^T.
\]
(12)

And then, the demapped signals are despread as

\[
\tilde{\mathbf{x}}_u(i) = \left[ \tilde{x}_u(0, i), \ldots, \tilde{x}(N_c - 1, i) \right]^T = \frac{1}{\sqrt{N_c}} \Phi_{(N/2)}^{-1} \tilde{\mathbf{s}}_u(i),
\]
(13)

where \( \tilde{x}_u(k, i) \) is the demapped signal for the \( u \)th user. Here, if \( \Phi_{(N/2)}^{-1} = \mathbf{W}_{(N/2)}^{-1} = \mathbf{W}_{N_c} \), the OFDM symbols are despread by the WHT. Also, if \( \Phi_{(N/2)}^{-1} = \mathbf{F}_{(N/2)}^{-1} \), the OFDM symbols are despread by the IDFT. Finally, after the demodulation and the S/P conversion, the demodulated signals are decoded with a deinterleaving and are returned to the bit signals.

### 3 Methods

In this section, we indicate the methods for the conventional individual and block subcarrier allocations, and the proposed interleaved block subcarrier allocation and power combination. In this paper, as shown in Sect. 2, since \( H_u(k, 0) \approx H_u(k, 1) \approx \cdots \approx H_u(k, N_u - 1) \), we assume \( H_u(k) = H_u(k, i) \). Moreover, we assume also that the subcarriers for each user are not assigned in the same frequency band, and the number of subcarriers for each user is equal as \( N_c \). Additionally, Figs. 2, 4, and 5 show the example of the conventional and proposed subcarrier allocations for \( N_c = 8 \) and \( N_u = 2 \), and Fig. 3 shows the flowchart of the proposed interleaved subcarrier allocation and power combination.

#### 3.1 Individual and block subcarrier allocations

As shown in Figs. 2, 4, and 5, in wireless communications, since the propagation channel for each user is independent, the channel fluctuation for each user is also independent at the frequency domain. By using this characteristic, the subcarriers for each user are assigned to obtain the MUDiv gain in the frequency band as
Fig. 2 Example of the conventional individual and block subcarrier allocations for $N_c = 8$, $N_u = 2$, and $B = 1$, 2, and 4

Fig. 3 Flowchart of the proposed interleaved block subcarrier allocation and power combination

Fig. 4 Example of the proposed interleaved block subcarrier allocation for $N_c = 8$, $N_u = 2$, and $I = 1$ and 2
where $\gamma_u(k)$ is the selection parameter as $H_{\max}B, u(k)$ is the subcarrier block channel which sorts the power of $H_{B, u}(k)$ as $|H_{\max}B, u(0)| \geq \cdots \geq |H_{\max}B, u(\lfloor N_c/B \rfloor - 1)|$. $H_B, u(k)$ is defined by Eq. (16). For Eq. (14), if the individual subcarrier allocation for $B = 1$, $H_1, u(k) = H_u(k)$. In this case, at the example of Fig. 2a, $J_B$ for $B = 1$ is given by

$$J_B = \max \left( \sum_{\alpha=0}^{(N_c/B)^{N_u-1}} J_B, \alpha \right) = \max \left( \sum_{u=0}^{N_u-1} \sum_{k=0}^{(N_c/B)-1} |H_{B, u}(k)| \gamma_u(k) \right) \leq \sum_{u=0}^{N_u-1} \sum_{k=0}^{(N_c/B)/N_u-1} |H_{B, u}(k)|,$$

(14)

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

$$\gamma_u(k) = \begin{cases} 1 & \text{for allocation} \\ 0 & \text{for no allocation} \end{cases}$$

(15)

$H_{B, u}(k)$ is the subcarrier block channel which sorts the power of $H_{B, u}(k)$ as $|H_{\max}B, u(0)| \geq \cdots \geq |H_{\max}B, u(\lfloor N_c/B \rfloor - 1)|$. $H_B, u(k)$ is defined by Eq. (16). For Eq. (14), if the individual subcarrier allocation for $B = 1$, $H_1, u(k) = H_u(k)$. In this case, at the example of Fig. 2a, $J_B$ for $B = 1$ is given by

$$J_1 = |H_1(0)| + |H_0(1)| + |H_1(2)| + |H_1(3)| + |H_1(4)| + |H_0(5)| + |H_0(6)| + |H_0(7)|.$$

This means that the subcarriers for the 0th user are assigned in $k = 1, 5, 6,$ and 7, and the subcarriers for the 1st user are assigned in $k = 0, 2, 3,$ and 4. As a result, the subcarriers for each user are assigned fairly. And then, the MUDiv gain is obtained since the subcarriers for each use are assigned in the good channel condition. The individual subcarrier allocation for $B = 1$ gives the good BER performance, but large computational complexity is required compared with $B > 1$. To solve this problem, the block subcarrier allocation for $B > 1$ is also considered. In the block subcarrier allocation, the block consists of several adjacent subcarriers as

![Fig. 5 Example of the proposed interleaved block subcarrier allocation and power combination for $N_c = 8$, $N_u = 2$, and $I = 1$](image-url)
$$\mathcal{H}_{B,u}(k) = \sum_{b=0}^{B-1} H_{u}(Bk + b). \quad (16)$$

In this case, at the example of Fig. 2b, c, $J_B$ for $B = 2$ and 4 is given by

$$J_2 = |\mathcal{H}_{2,1}(0)| + |\mathcal{H}_{2,1}(1)| + |\mathcal{H}_{2,0}(3)| + |\mathcal{H}_{2,0}(4)|$$
$$= |H_1(0)| + |H_2(1,0) + |H_2(2)| + |H_1(3) + |H_0(4)| + |H_0(5)| + |H_0(6)| + |H_0(7)|,$$

$$J_4 = |\mathcal{H}_{4,1}(0)| + |\mathcal{H}_{4,1}(1)|$$
$$= |H_1(0)| + |H_1(1)| + |H_1(2)| + |H_1(3)| + |H_0(4)| + |H_0(5)| + |H_0(6)| + |H_0(7)|.$$

These mean that the subcarriers for the 0th user are assigned in $k = 0$ to 7, and the subcarriers for the 1st user are assigned in $k = 0$ to 3. From Eqs. (14) and (16), the block subcarrier allocation for $B > 1$ is assigned for the subcarriers of each user, and the MUDiv gain is also obtained. Moreover, the computational complexity for $B > 1$ is reduced compared with $B = 1$.

### 3.2 Interleaved block subcarrier allocation

As shown in Sect. 3.1, the MUDiv gain is obtained by using the individual and block subcarrier allocations. However, the MUDiv gain for $B > 1$ is to be small since the possibility for the selection of the deep faded subcarrier channel is to be high compared with $B = 1$. Therefore, in the conventional individual and block subcarrier allocations, a trade-off problem is occurred between the MUDiv gain and the computational complexity. Therefore, we propose the interleaved block subcarrier allocation, and its flowchart is indicated in Fig. 3. In the interleaved block subcarrier allocation, the subcarriers for each user are assigned alternately to avoid a deep faded channel, and the block consists of the interleaved subcarrier channel to obtain the MUDiv gain as

$$K_I = \max \left( \sum_{u=0}^{N_c/B_{\text{max}}} \sum_{i=0}^{N_u-1} K_{I,u} \right) = \max \left( \sum_{u=0}^{N_c/B_{\text{max}}} \sum_{k=0}^{N_u-1} |\tilde{\mathcal{H}}_{I,u}(k)| \gamma_u(k) \right)$$

$$\leq \sum_{u=0}^{N_u-1} \sum_{k=0}^{N_c/B_{\text{max}}/N_u-1} |\tilde{\mathcal{H}}_{I,u}^{\text{max}}(k)|,$$ \quad (17)

where

$$\tilde{\mathcal{H}}_{I,u}^{\text{max}}(k) = \sum_{i=0}^{I} H_u(I[N_c/B_{\text{max}}]k + i'), \quad (18)$$

$\tilde{\mathcal{H}}_{I,u}^{\text{max}}(k)$ is the subcarrier block channel which sorts the power of $\tilde{\mathcal{H}}_{I,u}(k)$ as $|\tilde{\mathcal{H}}_{I,u}^{\text{max}}(0)| \geq \cdots \geq |\tilde{\mathcal{H}}_{I,u}^{\text{max}}(N_c/B_{\text{max}})|$. Observing Eq. (17), the interleaved block subcarrier allocation is only achieved in $B = B_{\text{max}}$. And then, in this paper, $B_{\text{max}}$ equals to $N_c$. In this case, at the example of Fig. 4a, $K_I$ for $I = 1$ is given by

$$K_1 = |\tilde{\mathcal{H}}_{1,1}(0)| + |\tilde{\mathcal{H}}_{1,0}(1)|$$
$$= |H_1(0)| + |H_0(1)| + |H_1(2)| + |H_0(3)| + |H_1(4)| + |H_0(5)| + |H_1(6)| + |H_0(7)|.$$
This means that the subcarriers for the 0th user are assigned in $k = 1, 3, 5,$ and 7, and the subcarriers for the 1st user are assigned in $k = 0, 2, 4,$ and 6. Therefore, in the proposed method, the subcarriers for each user are assigned alternately with the MUDiv gain in the frequency band. Moreover, in this paper, we consider also the case of $I > 1$. This is achieved by using Eq. (18), and the interleaved block consists of several subcarriers. And then, the subcarriers for each user are assigned as shown in Fig. 4b. In this case, $K_I$ for $I = 2$ is given by

$$K_2 = |\hat{H}_{2,1}(0)| + |\hat{H}_{2,0}(1)| = |H_1(0)| + |H_1(1)| + |H_0(2)| + |H_0(3)| + |H_1(5)| + |H_0(6)| + |H_0(7)|.$$  

This means that the subcarriers for the 0th user are assigned in $k = 2, 3, 6,$ and 7, and the subcarriers for the 1st user are assigned in $k = 0, 1, 4,$ and 5. Therefore, in the proposed method, the subcarriers of each user for $I > 1$ are also assigned.

### 3.3 Interleaved block subcarrier allocation and power combination

In a FSS-OFDM, the OFDM symbols are spread as shown in Eq. (6). In this case, the power of each subcarrier is different in the frequency band as shown in Fig. 5. Moreover, as the above mentioned, the channel is also fluctuated in the frequency band. By using these characteristics, the power combination between the subcarrier and the channel has been proposed. In this paper, we propose the interleaved block subcarrier allocation and the power combination, and its flowchart is indicated in Fig. 3.

As shown in Sects. 3.1 and 3.2, when the subcarriers for each user are assigned, the utilized frequency is also determined by Eq. (17). Here, the assigned channel responses are defined by the matrix form as

$$H_u^\text{sel} = \begin{bmatrix} H_u^\text{sel}(0), \ldots, H_u^\text{sel}(\tilde{N_c} - 1) \end{bmatrix}^\top,$$  \hspace{1cm} (19)

where $H_u^\text{sel}(k) = H_u(k)\gamma_u(k)$ for $\gamma_u(k) = 1$. Next, the selected channel response matrix for Eq. (19) and the spread signal matrix of the $i$th symbol for Eq. (6) are sorted based on these powers as

$$H_u^\text{pow} = \begin{bmatrix} H_u^\text{pow}(0), \ldots, H_u^\text{pow}(\tilde{N_c} - 1) \end{bmatrix}^\top,$$  \hspace{1cm} (20)

$$S_u^\text{pow}(i) = \begin{bmatrix} s_u^\text{pow}(0, i), \ldots, s_u^\text{pow}(\tilde{N_c} - 1, i) \end{bmatrix}^\top,$$  \hspace{1cm} (21)

where $H_u^\text{pow}(k)$ is the channel response for $|H_u^\text{pow}(0)| \leq |H_u^\text{pow}(1)| \leq \cdots \leq |H_u^\text{pow}(\tilde{N_c} - 1)|$, and $s_u^\text{pow}(k, i)$ is the spread signal for $|s_u^\text{pow}(0, i)| \leq |s_u^\text{pow}(1, i)| \leq \cdots \leq |s_u^\text{pow}(\tilde{N_c} - 1, i)|$. In the proposed method, since the $k$th largest power for the channel and the subcarrier is combined, the spread signals for Eq. (21) are mapped in the frequency band based on Eq. (20) as

$$m_u^\text{pow}(k, i) = \begin{cases} s_u^\text{pow}(k, i) & \text{for } H_u^\text{pow}(k) = \sum_{k'}^{N_c - 1} H_u(k')\gamma_u(k) \text{ and } \gamma_u(k) = 1, \\ 0 & \text{for } \gamma_u(k) = 0. \end{cases}$$  \hspace{1cm} (22)

In this case, at the example of Fig. 5, $H_u^\text{pow}(k)$ is given by
\[ H_0^{\text{pow}}(0) = H_0(3), \quad H_0^{\text{pow}}(1) = H_0(5), \quad H_0^{\text{pow}}(2) = H_0(1), \quad H_0^{\text{pow}}(3) = H_0(7), \]
\[ H_1^{\text{pow}}(0) = H_1(2), \quad H_1^{\text{pow}}(1) = H_1(6), \quad H_1^{\text{pow}}(2) = H_1(0), \quad H_1^{\text{pow}}(3) = H_1(4). \]

Moreover, \( s_k^{\text{pow}}(k, i) \) is given by
\[ s_0^{\text{pow}}(0, i) = s_0(1, i), \quad s_0^{\text{pow}}(1, i) = s_0(3, i), \quad s_0^{\text{pow}}(2, i) = s_0(2, i), \quad s_0^{\text{pow}}(3, i) = s_0(0, i), \]
\[ s_1^{\text{pow}}(0, i) = s_1(3, i), \quad s_1^{\text{pow}}(1, i) = s_1(0, i), \quad s_1^{\text{pow}}(2, i) = s_1(2, i), \quad s_1^{\text{pow}}(3, i) = s_1(1, i). \]

In this case, if \( k = 0, 2, 4, \) and \( 6, m_0^{\text{pow}}(k, i) = 0. \) Otherwise, \( m_0^{\text{pow}}(k, i) \) is assigned as
\[ m_0^{\text{pow}}(1, i) = s_0^{\text{pow}}(2, i) = s_0(2, i) \quad \text{for} \quad H_0^{\text{pow}}(2) = H_0(1), \]
\[ m_0^{\text{pow}}(3, i) = s_0^{\text{pow}}(0, i) = s_0(1, i) \quad \text{for} \quad H_0^{\text{pow}}(0) = H_0(3), \]
\[ m_0^{\text{pow}}(5, i) = s_0^{\text{pow}}(1, i) = s_0(3, i) \quad \text{for} \quad H_0^{\text{pow}}(1) = H_0(5), \]
\[ m_0^{\text{pow}}(7, i) = s_0^{\text{pow}}(3, i) = s_0(0, i) \quad \text{for} \quad H_0^{\text{pow}}(3) = H_0(7). \]

Similarly, if \( k = 1, 3, 5, \) and \( 7, m_1^{\text{pow}}(k, i) = 0. \) Otherwise, \( m_1^{\text{pow}}(k, i) \) is assigned as
\[ m_1^{\text{pow}}(0, i) = s_1^{\text{pow}}(2, i) = s_1(2, i) \quad \text{for} \quad H_1^{\text{pow}}(2) = H_1(0), \]
\[ m_1^{\text{pow}}(2, i) = s_1^{\text{pow}}(0, i) = s_1(3, i) \quad \text{for} \quad H_1^{\text{pow}}(0) = H_1(2), \]
\[ m_1^{\text{pow}}(4, i) = s_1^{\text{pow}}(3, i) = s_1(1, i) \quad \text{for} \quad H_1^{\text{pow}}(3) = H_1(4), \]
\[ m_1^{\text{pow}}(6, i) = s_1^{\text{pow}}(1, i) = s_1(0, i) \quad \text{for} \quad H_1^{\text{pow}}(1) = H_1(6). \]

Therefore, the proposed method achieves the interleaved block subcarrier allocation and the power combination based on Eq. (22).

### 3.4 Computational complexity for conventional and proposed subcarrier block allocations

In this Section, we indicate the computational complexity for the conventional and proposed block subcarrier allocations. Firstly, the computational complexity for the individual and block subcarrier allocations for Eq. (14) is given by

\[
C_{\text{blo}} = \left\lceil \frac{N_c}{B} \right\rceil^{N_u}.
\]  

Next, the interleaved block subcarrier allocation for Eq. (17) is only achieved in \( B = B_{\text{max}} \), and \( \lfloor N_c / B_{\text{max}} \rfloor \) equals to \( N_u \). Therefore, the computational complexity for the interleaved block subcarrier allocation is given by

\[
C_{\text{int}} = \left\lceil \frac{N_c}{B_{\text{max}}} \right\rceil^{N_u} = N_u^{N_u}.
\]  

Finally, the power combination for Eq. (22) is achieved after the utilized frequency selection for each user based on Eq. (17). Therefore, the computational complexity for the power combination is given by
In Sect. 4, we will compare these computational complexities by using the computer simulation parameter of Table 3.

### 4 Results and discussion

In this section, we discuss the system performance for the conventional and proposed methods by using the computer simulation. Table 3 shows the computer simulation parameters. In this simulation, we assume FSS-MUDiv/OFDMA systems for $N_u = 4$. And then, in the propagation channel, the symbol duration is $10 \mu s$, the GI length is $2 \mu s$, and the path model is 15 path Rayleigh fading at Doppler frequency of 5 Hz. Moreover,
Table 5  Computational complexity for the individual and block subcarrier allocations, the interleaved block subcarrier allocation, and the power combination

| B  | 1             | 2             | 4             | 8             | 16            | 32            |
|----|---------------|---------------|---------------|---------------|---------------|---------------|
| C<sub>Blo</sub> | 268,435,456   | 16,777,216    | 1,048,576     | 65,536        | 4096          | 256           |
| C<sub>int</sub>  | –             | –             | –             | –             | –             | 256           |
| C<sub>pow</sub>  | –             | –             | –             | –             | –             | 11,776        |

**Fig. 6** BER of the various B for the individual and block subcarrier allocations, the interleaved block subcarrier allocation, and the power combination with the DFT, the ZF, and the MMSE, without the coding at \( E_b/N_0 = 25 \) dB

the path model is indicated in Table 4. Firstly, at the transmitter as shown in Fig. 1a, the original data signals are coded by the convolutional code for \((R, K) = (1/2, 7)\) with the bit interleaving. After the S/P conversion, the parallel signals are modulated by 16QAM and are spread per the OFDM symbol by the WHT or the DFT. The spread signals for each user are assigned in the frequency band. Here, in the proposed method, the subcarriers for each user are assigned by the interleaved block subcarrier allocation and the power combination. And then, the mapped signals are converted to the time domain signal by the IFFT operation. Finally, after the GI insertion and the P/S conversion, the time domain signal is sent to the receiver via the propagation channel. Next, at the receiver as shown in Fig. 1b, after the S/P conversion and the GI elimination, the time domain signal is converted to the frequency domain signal by the FFT operation. The frequency domain signals are detected by the ZF or the MMSE. Here, we assume the perfect channel estimation. Moreover, the detected signals are demapped and are despread by the WHT or the IDFT. Finally, after the 16QAM demodulation, the demodulated signals are decoded by the Viterbi soft decoding algorithm with the deinterleaving and are returned to the binary signals.

---

The diagram in Fig. 6 illustrates the BER (Bit Error Rate) for various block sizes (B) when using different subcarrier allocation methods: the individual allocation, the block allocation, the interleaved block allocation, and the power combination. The diagram compares performance with the DFT, ZF, and MMSE detection schemes without coding, at an SNR of 25 dB. Each line represents a different allocation method, with distinct markers for the DFT, ZF, and MMSE techniques. The x-axis represents the block size (B), while the y-axis shows the BER on a logarithmic scale.
4.1 BER and computational complexity for various $B$

Figure 6 shows the BER of the various $B$ for the individual and block subcarrier allocations, the interleaved block subcarrier allocation, and the power combination with the DFT, the ZF, and the MMSE, without the coding at $E_b/N_0 = 25$ dB. Here, we define that “Blo” is the individual and block subcarrier allocations, “Int” is the interleaved block subcarrier allocation, and “Pow” is the power combination. And then, Table 5 shows the computational complexity for the individual and block subcarrier allocations, the interleaved block subcarrier allocation, and the power combination. In Table 5, they are calculated from Eqs. (23) to (25) by using the parameter of Table 3.

4.1.1 Results

In the ZF case, the block subcarrier allocation for $B = 4$ and the interleaved block subcarrier allocation for $I = 1$ show the same BER performance. Next, in the MMSE case, the block subcarrier allocation for $B = 4$ and the power combination for $I = 1$ show the same BER performance. Moreover, the block subcarrier allocation for $B = 8$ and the interleaved block subcarrier allocation for $I = 1$ show the same BER performance. Also, in all methods, the MMSE shows the good BER performance compared with the ZF.

4.1.2 Discussion

In the ZF case, the computational complexity for the interleaved block subcarrier allocation is reduced to $256/1,048,576 = 0.0002$ times compared with the block subcarrier allocation for $B = 4$, and they show the same BER performance. Next, in the MMSE case, the computational complexity for the power combination is reduced to $11,776/1,048,576 = 0.0112$ times compared with the block subcarrier allocation for $B = 4$, and they also show the same BER performance. Moreover, the computational complexity for the interleaved block subcarrier allocation is reduced to $256/65,536 = 0.0039$ times compared with the block subcarrier allocation for $B = 8$, and they also show the same BER performance. Additionally, the method for the MMSE obtains the good BER performance compared with the ZF since the detected error is reduced by using the noise power and the spreading technique. Therefore, since the proposed method reduces the computational complexity for the subcarrier allocation dramatically and obtains the good BER performance, it solves the trade-off problem for the conventional method.

4.2 BER for various $I$

Figure 7 shows the BER of the various $I$ for the interleaved block subcarrier allocation and the power combination with the WHT, the DFT, the ZF, and the MMSE, without the coding at $E_b/N_0 = 25$ dB. Here, the interleaved block subcarrier allocation for $I = 32$ equals to the block subcarrier allocation for $B = 32$. 
4.2.1 Results
When \( I \) is small, the interleaved block subcarrier allocation for the ZF and the MMSE, and the power combination for the ZF show the good BER performance. On the other hand, they show the floor in \( B \geq 4 \). Next, the power combination for the MMSE shows the good BER performance in \( I = 1 \) and \( I = 32 \), and they show the same BER performance. Also, in all methods, the MMSE shows the good BER performance compared with the ZF same as the case of Fig. 6. Finally, since the interleaved block subcarrier allocation and the power combination for the WHT and the DFT show the same BER performance.

4.2.2 Discussion
The interleaved block subcarrier allocation and the power combination for \( I = 1 \) achieve the good BER performance in the ZF and MMSE cases. This means that they are effect when the one chunk length for interleaved block subcarrier allocation (\( I \)) is small. And then, they have large effect for the avoiding of continuous deep faded subcarrier channels. Moreover, since their methods with the WHT for the biphase code and the DFT for the polyphase code show the same BER performance, they give the same effect for the BER performance. Additionally, the method for the MMSE obtains the good BER performance compared with the ZF as shown in Section 4.1.2. The advantage of the WHT is the simple processing by the biphase code, and the advantage of the DFT is the low PAPR by the polyphase code [7]. As a result, if the fast processing is required, WHT is better. Also, if the low PAPR is required, DFT is better. Therefore, the proposed method for the WHT and the DFT is effective in a small \( I \).

4.3 BER various \( E_b/N_0 \) without coding
Figures 8 and 9 show the BER of the various \( E_b/N_0 \) for the interleaved block subcarrier allocation, the power combination, and the WF, with the DFT and the ZF, without the
coding. For the WF method, we assume only the simple method in a MUDiv-OFDMA without the spreading technique. In this case, the method for the ZF and the MMSE shows the same BER performance [25]. Therefore, we show only the ZF case in Fig. 9.

The modulated signal between the $k$th subcarrier and the $i$th symbol for the $u$th user after the WF is defined by

$$m_{u}^{WF}(k,i) = \begin{cases} W_{u}^{W}m_{u}(k,i) & \text{for } T_{w} > |H_{u}(k)| \\ m_{u}(k,i) & \text{for otherwise.} \end{cases}$$ (26)
And then, Figs. 10 and 11 show the BER of the various $E_b/N_0$ for the interleaved block subcarrier allocation and the power combination with the DFT and the MMSE, without the coding. In Fig. 11, the interleaved block subcarrier allocation for $I = 1$ is also shown to compare the power combination.

**4.3.1 Results**

From the ZF case of Fig. 8, the interleaved block subcarrier allocation for $I = 1$ shows about 4.5 dB gain compared with $I = 2$ in the BER of $1.5 \times 10^{-5}$. And then, the interleaved block subcarrier allocation for $I = 2$ shows about 5.5 dB gain compared with $I = 32$ in the BER of $8 \times 10^{-5}$. The power combination shows also the same trend,
but the interleaved block allocation and the power combination show the same BER performance in $I = 1, 2, 32$ as shown in Fig. 7. Next, from Fig. 9, the WF for $T_{wf} = -20$ dB and $W_{wf} = 5$ shows about 2 dB gain performance compared with $T_{wf} = -10$ dB and $W_{wf} = 5$ in the BER of $1.5 \times 10^{-5}$ and shows about 3 and 5 dB gain compared with $T_{wf} = -20$ dB and $W_{wf} = 2$, and $T_{wf} = -30$ dB and $W_{wf} = 5$ in the BER of $4.5 \times 10^{-5}$. Moreover, it shows about 6 dB gain compared with the interleaved block subcarrier allocation for $I = 32$. On the other hand, it shows about 3 dB penalty compared with the interleaved block subcarrier allocation for $I = 1$.

From the MMSE case of Fig. 10, the interleaved block subcarrier allocation for $I = 1$ shows about 3.5 dB gain compared with $I = 2$ in the BER of $1 \times 10^{-6}$ and shows about 6 dB gain compared with $I \geq 4$ in the BER of $3 \times 10^{-6}$. Moreover, in the BER of $3 \times 10^{-6}$, the interleaved block subcarrier allocation for $I = 2$ shows about 3 dB gain compared with $I \geq 4$, and the interleaved block subcarrier allocation for $I \geq 4$ shows the same BER performance. Moreover, from Fig. 11, the power combination for $I = 1$ shows about 3 dB gain compared with $I = 2$ in the BER of $3 \times 10^{-7}$ and shows about 4 dB gain compared with $I = 4, 8$ and $16$ in the BER of $4 \times 10^{-7}$. On the other hand, the power combination for $I = 1$ and 32 shows the same performance in the BER of $1 \times 10^{-7}$. Also, the power combination for $I = 2$ and the interleaved block subcarrier allocation for $I = 1$ also show the same performance in the BER of $4 \times 10^{-7}$. Additionally, from Figs. 8, 10, and 11, the MMSE shows the good BER performance compared with the WF in all methods same as the case of Figs. 6 and 7.

### 4.3.2 Discussion

From the ZF case of Fig. 8, the interleaved block allocation and the power combination show the good BER performance in a small $I$ at various $E_b/N_0$. This reason is discussed in Sect. 4.2.2. However, they show the same BER performance. This means that the influence of the detected error due to the ZF is strong compared with the diversity gain for the power combination. Next, from Fig. 9, the WF for $T_{wf} = -20$ dB and $W_{wf} = 5$ shows the good BER performance compared with the interleaved block subcarrier allocation for $I = 32$ by assigning the adaptive threshold and power. This is because WF is achieved based on the interleaved block subcarrier allocation for $I = 32$ without the spreading. On the other hand, it shows the bad BER performance compared with the interleaved block subcarrier allocation for $I = 1$. This means that the proposed method is mitigated for a deep faded subcarrier channel compared with the WF method. Therefore, the proposed method is effective compared with the WF method.

From the MMSE case of Figs. 10 and 11, the interleaved block allocation and the power combination also show the good BER performance in a small $I$ at various $E_b/N_0$ same as the ZF case. And then, from Fig. 11, since the power combination shows the good BER performance compared with the interleaved block subcarrier allocation, the diversity gain for the power combination is obtained in the MMSE case. Therefore, in the proposed method, MMSE is effective to obtain the diversity gain for the power combination. Moreover, the power combination for $I = 2$ and the interleaved block subcarrier allocation for $I = 1$ show the same BER performance. In this case, the computational complexity for the interleaved block subcarrier allocation is reduced to $256/11776 = 0.0217$ times compared with the power combination from Table 5. Additionally, the method
for the MMSE obtains the good BER performance compared with the ZF as shown in Sects. 4.1.2 and 4.2.2. Therefore, the interleaved block subcarrier allocation for $I = 1$ is effective for the BER and the computational complexity in the MMSE case without the coding.

4.4 BER various $E_b/N_0$ with coding
Figures 12 and 13 show the BER of the various $E_b/N_0$ for the interleaved block subcarrier allocation and the power combination with the DFT, the MMSE, and the coding.
4.4.1 Results

From Fig. 12, the interleaved block subcarrier allocation for $I = 1$ shows about 1 dB gain compared with $I = 2$ and shows about 4 dB gain compared with $I \geq 4$ in the BER of $1 \times 10^{-7}$. Moreover, since the interleaved block subcarrier allocation for $I = 2$ shows about 3 dB gain compared with $I \geq 4$. Next, from Fig. 13, the power combination for $I = 1, 2,$ and 32 shows the same performance in the BER of $1 \times 10^{-7}$. Moreover, they show about 3 to 4 dB gain compared with $I = 4, 8,$ and 16. Also, the interleaved block subcarrier allocation for $I = 1$ also shows the good BER performance same as the power combination for $I = 4$. 

Fig. 14 BER of the various $E_b/N_0$ for the interleaved block subcarrier allocation with the DFT and the MMSE, without the coding at $E_b/N_0 = 25$ dB

Fig. 15 BER of the various $E_b/N_0$ for the power combination with the DFT and the MMSE, without the coding at $E_b/N_0 = 25$ dB
4.4.2 Discussion
From Figs. 12 and 13, in the coding case, since the interleaved block subcarrier allocation and power combination for $I = 2$ show the good BER performance, they are effective as $I = 1$. Therefore, the proposed method for $I = 2$ is also effective by obtaining the coding gain. Moreover, since the interleaved block subcarrier allocation for $I = 1$ shows also the good BER performance, it is effective as the uncoding case.

4.5 BER various $E_b/N_{icl}$
Figures 14 and 15 show the BER of the various $E_b/N_{icl}$ for the interleaved block subcarrier allocation and the power combination with the DFT and the MMSE, without the coding at $E_b/N_0 = 25$ dB.

4.5.1 Results
From Fig. 14, the interleaved block subcarrier allocation for $I = 1$ shows about 4 dB gain compared with $I = 2$ in the BER of $1.5 \times 10^{-5}$. And then, it shows about 8 dB gain compared with $I = 4, 8,$ and 16 in the BER of $5 \times 10^{-5}$. Also, the interleaved block subcarrier allocation for $I = 2$ shows also the good BER performance compared with the $I = 32$ (i.e., the conventional block subcarrier allocation for $B = 32$), but $I = 4, 8,$ and 16 show the bad BER performance. Next, from Fig. 15, the power combination for $I = 1$ shows about 1 to 6 dB gain compared with $I = 2$ to 32 in BER of $1 \times 10^{-5}$. Moreover, the interleaved block subcarrier allocation for $I = 1$ shows about 1 to 4 dB gain compared with $I = 2$ to 16.

4.5.2 Discussion
From Figs. 14 and 15, the interleaved block subcarrier allocation and the power combination for $I = 1$ and 2 show the good BER performance. Therefore, the proposed method is effective to reduce the influence of the ICI by using the interleaved allocation and the MUDiv gain same as the $E_b/N_0$ case.

5 Conclusion
In this paper, we have proposed the interleaved block subcarrier allocation and the power combination for a FSS-MUDiv/OFDMA. In the individual and block subcarrier allocations, the MUDiv gain is obtained by using the independent channel fluctuation for each user, but they have a trade-off problem between the BER performance and the computational complexity. To solve this problem, in the proposed method, the interleaved block subcarrier allocation which consists of the interleaved block for the individual and several subcarriers is achieved. Moreover, the power combination between the selected interleaved subcarrier and channel is also achieved. From the computer simulation results, the proposed method reduces the computational complexity and shows the good BER performance. Especially, in the proposed method, the interleaved block subcarrier allocation for $I = 1$ and 2 is effective for the ZF and the MMSE cases with and without the coding.
Abbreviations

AWGN Additive white Gaussian noise
BER Bit error rate
CIR Channel impulse response
CR Channel response
CSI Channel state information
DFT Discrete Fourier transform
FEC Forward error correction
FFT Fast Fourier transform
FSS Frequency symbol spreading
GI Guard interval
ICI Inter-cell interference
IDFT Inverse DFT
IFFT Inverse FFT
MUDiv Multiuser diversity
MMSE Minimum mean square error
NOMA Non-orthogonal multiple access
OFDM Orthogonal frequency division multiplexing
OFDMA OFDM access
OMA Orthogonal multiple access
PAPR Peak average power ratio
P/S Parallel-to-serial
QAM Quadrature amplitude modulation
SIC Successive interference cancellation
S/P Serial-to-parallel
WF Water filling
WHT Walsh-Hadamard transform
WLAN Wireless local area network
ZF Zero-forcing

Author contributions

Yi proposed the idea and approach and involved in writing. TM contributed the confirmation for the approach and draft. Both authors read and approved the current manuscript.

Funding

This work was supported by Japan Society for the Promotion of Science (JSPS) KAKENHI Numbers 18K04145 and 21K04062, and Strategic Information and Communications R&D Promotion Programme (SCOPE) Number JP225008001.

Declarations

Competing interests

The authors declare that they have no competing interests.

Received: 15 August 2021   Accepted: 29 May 2022

Published online: 07 June 2022

References

1. L.J. Cimini, Analysis and simulation of a digital mobile channel using orthogonal frequency division multiplexing. IEEE Trans. Commun. 33(7), 666–675 (1985)
2. J.J. Gimenez, D.G. Barquero, J. Morgade, E. Stare, Wideband broadcasting: a power-efficient approach to 5G broadcasting. IEEE Commun. Mag. 56(3), 119–125 (2018)
3. M. Nurchis, B. Bellalta, Target wake time: scheduled access in IEEE 802.11ax WLANs. IEEE Wirel. Commun. 26(2), 142–150 (2019)
4. L. Zhu, Z. Xiao, X.-G. Xia, D.O. Wu, Millimeter-wave communications with non-orthogonal multiple access for BS/GiG. IEEE Access 7, 116123–116132 (2019)
5. Z. Wei, J. Guo, D.W.K. Ng, J. Yuan, Fairness comparison of uplink NOMA and OMA, in IEEE 85th Vehicular Technology Conference (2017). https://doi.org/10.1109/VTC2017.8108680
6. M. Munster, L. Hanzo, Performance of SDMA multiuser detection techniques for walsh-hadamard-spread OFDM schemes, in IEEE 54th Vehicular Technology Conference (2001). https://doi.org/10.1109/VTC.2001.957162
7. Z. Dong, S. Matsufuji, Y. Ida, T. Matsumoto, A study on PAPR reduction in OFDM using complex hadamard matrices, in The 8th International Workshop on Signal Design and its Applications in Communications (2017), pp. 154–158
8. J. Kim, Y.H. Yun, C. Kim, J.H. Cho, Minimization of PAPR for DFT-spread OFDM with BPSK Symbols. IEEE Trans. Veh. Technol. 67(12), 11746–11758 (2018)
9. A. Sahin, N. Hosseini, H. Jamal, S.M. Hoque, D.W. Matolak, DFT-spread-OFDM-based chip transmission. IEEE Commun. Lett. 25(3), 902–906 (2021)
10. F. Adachi, H. Tombeba, K. Takeda, Frequency-domain equalization for broadband single-carrier multiple access. IEICE Trans. Commun. E92–B(5), 1441–1456 (2009)
11. T. Svensson, T. Frank, T. Eriksson, D. Aronsson, M. Sternad, A. Klein, Block interleaved frequency division multiple access for power efficiency, robustness, flexibility, and scalability. EURASIP J. Wirel. Commun. Netw. (2009). https://doi.org/10.1155/2009/720973
12. Y. Shao, S.C. Liew, Flexible subcarrier allocation for interleaved frequency division multiple access. IEEE Trans. Wirel. Commun. 19(11), 7139–7152 (2020)
13. S.C. Liew, Y. Shao, New transceiver designs for interleaved frequency-division multiple access. IEEE Trans. Wirel. Commun. 19(12), 7765–7778 (2020)
14. Y. Karasawa, H. Iwai, Modeling of signal envelope correlation of line-of-sight fading with applications to frequency correlation analysis. IEEE Trans. Commun. 42(6), 2201–2203 (1994)
15. C.Y. Wong, R.S. Cheng, K.B. Letaief, R.D. Murch, Multituser OFDM with adaptive subcarrier, bit, and power allocation. IEEE J. Sel. Areas Commun. 17(10), 1747–1758 (1999)
16. C. Ahn, D. Har, T. Omori, K. Hashimoto, Frequency symbol spreading based adaptive subcarrier block selection for OFDMA system. Digit. Signal Process. 22(3), 518–525 (2012)
17. T. Mert, O. Kaya, H. A. Cirpan, Jointly optimal chunk and power allocation in uplink SC-FDMA, in IEEE International Conference on Communications (2013), pp. 3393–3397
18. Y. Ida, C. Ahn, T. Matsumoto, S. Matsufuji, Removing deep faded subcarrier channel for cooperative multituser diversity OFDMA based on low granularity block. IEICE Trans. Fundam. E97–A(12), 2586–2594 (2014)
19. C.-H. Choi, H.-J. Kim, T.-K. Kim, G.H. Im, Spectral efficient multituser technique with channel-dependent resource allocation schemes. IEEE Trans. Wirel. Commun. 11(3), 990–999 (2012)
20. Y. Ida, T. Matsumoto, S. Matsufuji, Interleaved MUDiv block allocation and optimum power control for SC-FDMA, in RISP International Workshop on Nonlinear Circuits. Communications and Signal Processing, vol. 2017 (2017), pp. 517–520
21. A. Kuroha, C. Ahn, T. Omori, K. Hashimoto, Multituser diversity OFDMA using power priority selection and adaptive clipping. Int. J. Dist. Syst. Technol. 5(4), 18–30 (2014)
22. T. Iwata, H. Miyazaki, F. Adachi, Capacity-fairness controllable scheduling algorithms for single-carrier FDMA. IEICE Trans. Commun. E97–B(7), 1474–1482 (2014)
23. J. Mao, G. Xie, J. Gao, Y. Liu, Energy efficiency optimization for OFDM-based cognitive radio systems: a water-filling factor aided search method. IEEE Trans. Wirel. Commun. 12(5), 2366–2375 (2013)
24. S. Khakurel, L. Musavian, H.V. Vu, T.L. Ngoc, QoS-aware utility-based resource allocation in mixed-traffic multi-user OFDM systems. IEEE Access 6, 21646–21657 (2018)
25. Y. Ida, T. Matsumoto, S. Matsufuji, Different antenna interleaved allocation with full and divided DFT/IDDT spreading for HT-RCI-MIMO/OFDM. IEICE Trans. Commun. E103–B(12), 1438–1446 (2020)

Publisher’s Note
Springer Nature remains neutral with regard to jurisdictional claims in published maps and institutional affiliations.