Abstract—The paper considers a time domain channel estimation approach for uplink OFDMA (Orthogonal Frequency Division Multiple Access) systems. Although frequency domain channel estimation schemes are widely used for those systems, we propose time domain channel estimation schemes by taking advantage of the sparsity of channel impulse response with compressed sensing. Numerical simulations show the merit of the proposed schemes, which demonstrates the validity of the time domain channel estimation approach for OFDMA systems.

I. INTRODUCTION

It has been known that, in OFDM (Orthogonal Frequency Division Multiplexing) based systems, the channel estimation accuracy can be improved by performing the estimation in time domain instead of frequency domain, because the number of parameters to be estimated is at most the length of the cyclic prefix plus one, which is much less than the number of subcarriers in general [1]–[3]. On the other hand, the frequency domain approach has been commonly employed for the channel estimation of uplink OFDMA (Orthogonal Frequency Division Multiple Access) systems [4]. This is mainly because the information of channel frequency responses is required only at the allocated subcarriers for the demodulation of the received signal at the base station (BS), hence the whole frequency responses (or equivalently, the impulse response) of each user’s channel are not necessary. Moreover, another reason might be that the signals coming from different users are distorted by different frequency selective channels, while the available pilot signals for the estimation of each user’s channel impulse response are limited within subcarriers assigned to the corresponding user.

In [5], a time domain channel estimation scheme for uplink OFDMA is proposed by considering the estimate of a virtual channel impulse response, which is defined as the inverse discrete Fourier transform (IDFT) of a frequency response vector, whose response at each subcarrier is given by the corresponding user’s channel frequency response. Assuming the localized subcarrier allocation, where a consecutive set of subcarriers are assigned to each user, it has been shown that the number of nonzero elements of the virtual channel impulse response is much less than the number of subcarriers as far as the number of simultaneous users is small. However, the method cannot be applied for the distributed subcarrier allocation scheme, and the estimation performance rapidly degrades as the number of users increases.

In this paper, we consider a time domain channel estimation scheme for uplink OFDMA systems taking advantage of the novel sparse signal processing framework of compressed sensing [6]. So far, a considerable number of studies have been made on sparse channel estimation for OFDM systems using compressed sensing [7]–[12], however, to the best of our knowledge, there is little work on the application of compressed sensing to channel estimation for uplink OFDMA systems. We propose two different approaches for channel estimation, namely, a method making use of compressed sensing for the estimation of the virtual channel impulse response considered in [5], and a method estimating each user’s channel impulse response using compressed sensing. The performance of the proposed schemes is evaluated by computer simulations assuming both the localized and the distributed subcarrier allocations, and is compared with that of conventional channel estimation schemes to show the merit of the proposed approaches.

In the rest of the paper, we use the following notations for describing the proposed schemes. An $M \times M$ identity matrix will be denoted as $1_M$, a zero matrix of size $A \times B$ as $0_{A \times B}$, an all one matrix of size $A \times B$ as $1_{A \times B}$, a diagonal matrix, whose $(i, j)$ element is $a_{ij}$, as diag$[a_{0}, a_{1}, \ldots, a_{M-1}]$, and the discrete Fourier transform (DFT) matrix of size $M \times M$, whose $(i, j)$ element is $\frac{1}{\sqrt{M}} e^{-j \frac{2\pi (i-1)(j-1)}{M}}$, as $D$. We use $E[\cdot]$ to denote ensemble average, $(\cdot)^T$ for transpose, and $(\cdot)^{H}$ for Hermitian transpose. The cardinality of a set $A$ is denoted by $|A|$.

II. SYSTEM MODEL

Figure 1 shows the system model of the uplink OFDMA system considered in this paper. The number of users is assumed to be $N$, and the $n$-th ($n = 0, \ldots, N-1$) user transmits the information bearing symbol vector $s_n = [s_{n0}^0, s_{n1}^1, \ldots, s_{nM-1}^0] \in \mathbb{C}^M$, where $M$ is the number of subcarriers. Note that we assume that only one user is assigned to each subcarrier. Denoting by $S_n$ the set of indices of the
subcarriers allocated to the $n$-th user, we have

$$s^n_m = \begin{cases} s_m & (m \in S_n) \\ 0 & (m \notin S_n) \end{cases}, \quad (1)$$

$$\bigcup_{n=0}^{N-1} S_n = \{0, 1, \ldots, M-1\}, \quad (2)$$

$$S_n \cap S_{n'} = \emptyset \quad (n \neq n'). \quad (3)$$

Typical subcarrier assignment strategies are local and distributed subcarrier allocations. In the localized subcarrier allocation, a consecutive set of subcarriers is assigned to a user. If we assume that the subcarriers are assigned in ascending order of the subcarrier index to the users in ascending order of the user index, $s_n$ is given by

$$s_n = \begin{bmatrix} 0_{1 \times (\sum_{k=0}^{n-1} N_k)} s^n_0 \ldots s^n_{N_n-1} 0_{1 \times (M-\sum_{k=0}^{n} N_k)} \end{bmatrix}^T, \quad (4)$$

where $N_n$ is the number of subcarriers assigned to the $n$-th user, which satisfies $\sum_{n=0}^{N-1} N_n = M$. On the other hand, the distributed subcarrier allocation assigns randomly selected subcarriers to users in a non-overlapping manner. In this paper, we consider both subcarrier assignment strategies in numerical evaluations, since they have a large impact on the performance of the proposed schemes. Since we consider pilot-assisted channel estimation schemes in this paper, we assume that some of the allocated subcarriers are used to send pilot signals, and denote the set of indices of pilot subcarriers for the $n$-th user as $L_n(\subset S_n)$.

Each user transmits $s_n$ after performing IDFT and insertion of the cyclic prefix (CP). The received signal vector at the BS after CP removal is given by

$$r = \sum_{n=0}^{N-1} C_n D^H s_n + \mathbf{w}, \quad (5)$$

where $\mathbf{w}$ is a zero-mean additive white noise vector of correlation matrix $E[\mathbf{w}\mathbf{w}^H] = \sigma^2 \mathbf{I}$ and $C_n \in \mathbb{C}^{M \times M}$ is the channel matrix between the $n$-th user and the BS. Assuming the order of each channel impulse response to be at most the length of the CP $L$ and defining $h_n = [h^n_0, h^n_1, \ldots, h^n_L]^T$ as the impulse response vector of the channel between the $n$-th user and the BS, $C_n$ becomes a circulant matrix having the first column $[h^n_0, 0_{1 \times (M-L-1)}]^T$ and the first row $[h^n_0, h^n_1, \ldots, h^n_L]^T$. Thus, after performing DFT on $r$, we have

$$\rho = [\rho_0, \rho_1, \ldots, \rho_{M-1}]^T = Dr = \sum_{n=0}^{N-1} DC_n D^H s_n + Dw \quad (6)$$

$$= \sum_{n=0}^{N-1} \Lambda_n s_n + \nu, \quad (7)$$

where $\nu = [\nu_0, \nu_1, \ldots, \nu_{M-1}]^T \in \mathbb{C}^M$ and

$$\Lambda_n = DC_n D^H = \text{diag}[\lambda^n_0, \lambda^n_1, \ldots, \lambda^n_{M-1}], \quad (8)$$

where $\lambda^n_m$ is the frequency response of the channel between the $n$-th user to the BS at the $m$-th ($m = 0, 1, \ldots, M-1$) subcarrier, which is calculated as $[\lambda^n_0, \lambda^n_1, \ldots, \lambda^n_{M-1}]^T = D_h^T n_0, 0_{1 \times (M-L-1)}]$. If we define a vector $s = [s_0, s_1, \ldots, s_{M-1}]^T \in \mathbb{C}^M$ as

$$s = \sum_{n=0}^{N-1} s_n, \quad (10)$$

$s_n$ can be written as

$$s_n = P_n s, \quad (11)$$

where $P_n \in \mathbb{R}^{M \times M}$ is defined as

$$P_n = \text{diag}[p^n_0, p^n_1, \ldots, p^n_{M-1}] \quad (12)$$

$$p^n_m = \begin{cases} 1 & (m \in S_n) \\ 0 & (m \notin S_n) \end{cases} (m = 0, 1, \ldots, M-1), \quad (13)$$

since subcarriers are assigned to users in a non-overlapping manner. Thus, by using (11), (7) can be rewritten as

$$\rho = \sum_{n=0}^{N-1} \Lambda_n P_n s + \nu \quad (14)$$

$$= \Lambda s + \nu, \quad (15)$$

where

$$\Lambda = \sum_{n=0}^{N-1} \Lambda_n P_n \quad (16)$$

$$= \text{diag}[\lambda_0, \lambda_1, \ldots, \lambda_{M-1}]. \quad (17)$$

Here, we define a vector composed of the diagonal elements of $\Lambda$ as

$$\lambda = [\lambda_0, \lambda_1, \ldots, \lambda_{M-1}]^T. \quad (18)$$

The channel estimation problem considered in this paper is to estimate $\lambda$ from the received signal vector $r$. Note that $\lambda$ is composed of subsets of channel frequency responses of each user corresponding to the allocated subcarriers, and that the estimation of all frequency responses (or impulse responses) of all users’ channels is not required.
III. CONVENTIONAL CHANNEL ESTIMATION SCHEMES

A. Frequency Domain Approach

The estimates of the channel frequency responses at the pilot subcarriers can be obtained as

$$\hat{\lambda}_m = \frac{\rho_m}{s_m} = \lambda_m + \frac{\nu_m}{s_m} \quad (m \in \mathcal{I}),$$

(19)

where $\mathcal{I} = \cup_{n=1}^N I_n$, while the frequency responses at the data subcarriers are not directly obtained since $s_m$ is unknown for $m \notin \mathcal{I}$. In the frequency domain channel estimation approach, the estimates of $\lambda_m$ for $m \notin \mathcal{I}$ are obtained via some kind of interpolation in frequency domain [4].

For example, if we employ the 0-th order interpolation scheme, the estimate of the frequency response of the channel between the $n$-th user and the BS at the data subcarrier $\hat{\lambda}_m^{n'}$ ($m' \in S_n \setminus I_n$) is given by the estimate of the channel frequency response at the pilot subcarrier whose subcarrier index $m \in I_n$ is the closest to $m'$ as $\hat{\lambda}_m^{n'} = \hat{\lambda}_m$.

The interpolation based on minimum mean-square-error (MMSE) criterion has been also proposed in the literature [13]. Defining a vector of the frequency responses at the pilot subcarriers as $\lambda_p \in \mathbb{C}^{\mathcal{I}}$, it can be written as

$$\lambda_p = E \lambda,$$

(20)

where $E \in \mathbb{R}^{\mathcal{I} \times M}$ is a matrix constructed by omitting the $m$-th ($m = 0, 1, \cdots, M-1$) row vector of the identity matrix $I_M$ if $m \notin \mathcal{I}$. Thus, the estimate of $\lambda_p$ via (19) is given by

$$\hat{\lambda}_p = E \lambda + \nu_p,$$

(21)

where $\nu_p \in \mathbb{C}^{\mathcal{I}}$ is a noise vector, whose elements are given by $\nu_m/s_m$ ($m \in \mathcal{I}$). By assuming $E[|s_m|^2] = 1$, the MMSE estimate of the frequency response vector is obtained as

$$\hat{\lambda}_{\text{mmse}} = \text{RE}^T (E \text{RE}^T + \sigma^2 I_{\mathcal{I}})^{-1} \lambda_p,$$

(22)

where $\text{R} = E[\lambda H^H]$. Since $\text{R}$ is unknown a priori in general, uniform delay power spectrum with uncorrelated channel gains is often assumed for the calculation of $\text{R}$.

B. Time Domain Approach

In [5], a time domain channel estimation scheme is proposed based on the property of the IDFT of $\lambda$ defined as

$$h = [h_0, h_1, \ldots, h_{M-1}]^T = D^H \lambda,$$

(23)

which is called the overall impulse response of the virtual channel since it is obtained as the IDFT of the frequency response vector $\lambda$, although $\lambda$ is not the frequency response of any physical channel since it is composed of subsets of the frequency responses of allocated users. Assuming localized subcarrier allocation as in (4) and uncorrelated path gains as

$$E[h_i^* h_{i'}^*] = \begin{cases} \sigma^2_{l,n}, & l = l', \ n = n' \\ 0, & \text{otherwise} \end{cases},$$

(24)

it has been shown in [5] that the delay power spectrum of $h$ can be obtained as

$$E[|h_i|^2] = \frac{1}{M^2} \sum_{n=0}^{N-1} \sum_{l=0}^{L} \sigma^2_{l,n} \frac{1 - \cos \frac{2 \pi}{M} (i - l) N} {1 - \cos \frac{2 \pi}{M} (i - l)}.$$  (25)

Moreover, if we assume the same exponentially decaying delay power profile for all users as $\sigma^2_{l,n} = \sigma^2_l$, $(n = 0, 1, \cdots, N-1)$ and

$$\sigma^2_l = \frac{1}{C} a^l, \quad C = \sum_{l=0}^{L} a^l,$$

(26)

we have

$$E[|h_i|^2] = \frac{1}{CM^2} \sum_{n=0}^{N-1} \sum_{l=0}^{L} a^l \frac{1 - \cos \frac{2 \pi}{M} (i - l) N} {1 - \cos \frac{2 \pi}{M} (i - l)}.$$  (27)

Figure 2 shows an example of the delay power spectrum of the virtual overall impulse response $h$ obtained from (27), where parameters are set to be $M = 128$, $N = 4$, $N_n = 32$, $L = 16$, and $a = 1.0$.

From Fig.2, we can see that the nonzero elements of $h$ are located only for some first and last some elements. Based on this observation, the method in [5] assumes $h_i = 0$ for $i = L + \alpha + 1, \cdots, M - \alpha - 1$, $(\alpha > 0)$, estimates $h' \in \mathcal{C}^{L+2\alpha+1}$ defined as

$$h' = [h_0, \cdots, h_{L+\alpha}, h_{M-\alpha}, \cdots, h_{M-1}]^T,$$

(28)

via the method of least squares (LS), and finally obtains the estimate of $\lambda$ by performing the DFT on $h'$ after insertion of the zeros for $h_i, \ i = L + \alpha + 1, \cdots, M - \alpha - 1$. Specifically, defining a diagonal matrix $Q = \text{diag}[q_0, q_1, \cdots, q_{M-1}] \in \mathbb{R}^{M \times M}$, whose diagonal elements are given by

$$q_m = \begin{cases} 1, & (m \in \mathcal{I}) \\ 0, & (m \notin \mathcal{I}) \end{cases},$$

(29)

the left multiplication of $D^H Q$ for $\rho$ gives

$$D^H Q \rho = D^H Q s + D^H Q \nu,$$

(30)

$$= C_h s' + D^H Q \nu,$$

(31)
where \( C_h = D^H A D \) and \( s' = D^H Q s \). Since \( C_h \) is a circulant matrix, (31) can be rewritten as
\[
D^H Q \rho = C_h' h + D^H Q \nu,
\]
(32)
where \( C_h' \in \mathbb{C}^{M \times M} \) is a circulant matrix composed of \( s' \) defined as \( C_h' = D^H \text{diag}[s'_D] \). Moreover, using the assumption on the elements of \( h \), we have
\[
D^H Q \rho = C_h' h + D^H Q \nu,
\]
(33)
where \( C_h' \in \mathbb{C}^{M \times (L + 2a + 1)} \) is a matrix composed of the first \( L + 1 + \alpha \) and the last \( \alpha \) column vectors of \( C_h' \). Thus, if \( |Z| \geq L + 2a + 1 \) holds, the LS estimate of \( h' \) is given by
\[
\hat{h}' = \left( C_h'^H C_h' \right)^{-1} C_h'^H D^H Q \rho.
\]
(34)

IV. PROPOSED CHANNEL ESTIMATION SCHEMES VIA COMPRESSED SENSING

In this section, we propose two different channel estimation schemes in time domain using the idea of compressed sensing [6].

A. Proposed Scheme 1: estimation of the virtual overall channel impulse response

In the first method, we utilize compressed sensing to obtain the estimate of \( h' \) in (28). To be precise, (33) can be regarded as linear simultaneous equations obtained from linear measurements, where \( h' \) is the unknown vector, \( D^H Q \rho \) is the known measurement vector, \( C_h' \) the known sensing matrix, and \( D^H Q \nu \) an additive noise vector. Therefore, if \( h' \) can be assumed to be sparse, the estimate of \( h' \) might be obtained even when \( |Z| < L + 2a + 1 \) by using compressed sensing. If we employ the \( l_1-l_2 \) reconstruction approach for compressed sensing, the estimate of \( h' \) is given by
\[
\hat{h}' = \arg \min_{h'} \left( \frac{1}{2} \| D^H Q \rho - C_h' h' \|_2^2 + \mu \| h' \|_1 \right),
\]
(35)
where \( \mu \) is a regularization parameter. The estimate of \( \lambda \) is obtained by taking the DFT of \( \hat{h}' \) after insertion of the zero elements.

Note that, while the proposed scheme 1 is available only for the localized subcarrier allocation as in the case of the conventional method in Sect.III-B, it has the advantage that the channel estimation can be achieved by solving a single compressed sensing problem.

B. Proposed Scheme 2: estimation of all users’ channel impulse responses

The second scheme firstly estimates the channel impulse responses \( h_n \) for all users \( (n = 0, \ldots, N - 1) \) and then uses them to obtain the estimate of \( \lambda \).

By the left multiplication of \( D^H Q P_n \) on both sides of (14), we have
\[
D^H Q P_n \rho = \sum_{k=0}^{N-1} D^H Q P_n \lambda_k P_k s + D^H Q P_n \nu
\]
(36)
\[
= D^H \Lambda_n s_n + D^H Q P_n \nu
\]
(37)
\[
= C_h s_n + D^H Q P_n \nu,
\]
(38)
where \( s_n = D^H Q P_n s \). Moreover, using the property of the circulant matrix, we have
\[
D^H Q P_n \rho = C_{h_n} [0_{(M - L \cdot n - 1) \times 1}] + D^H Q P_n \nu
\]
(39)
\[
= C_h s_n + D^H Q P_n \nu
\]
(40)
where \( C_{h_n} \in \mathbb{C}^{M \times M} \) is a circulant matrix defined as \( C_{h_n} = D^H \text{diag}[s_n] D \), and \( C_{h_n} \in \mathbb{C}^{M \times (L + 1)} \) is a matrix composed of the first \( L + 1 \) column vectors of \( C_{h_n} \). Thus, if we use the \( l_1-l_2 \) reconstruction as in the proposed scheme 1, the estimate of \( h_n \) is obtained as
\[
\hat{h}_n = \arg \min_{h_n} \left( \frac{1}{2} \| D^H Q P_n h_n - C_{h_n} h_n \|_2^2 + \mu \| h_n \|_1 \right).
\]
(41)

After performing the above channel estimation procedure for all users, the estimate of \( \lambda \) is given by
\[
\lambda = \sum_{n=0}^{N-1} P_n D \hat{h}_n.
\]
(42)

Note that, although the number of available pilot signals for each impulse response estimation decreases as the number of users increases in the proposed scheme 2, the scheme can be used regardless of the subcarrier allocation strategy.

V. NUMERICAL EXPERIMENTS

In order to evaluate the performance of the proposed channel estimation schemes, computer simulations are conducted in comparison with conventional channel estimation schemes. The evaluated channel estimation schemes are as follows:

- Proposed 1 ··· proposed scheme 1 in Sect. IV-A
- Proposed 2 ··· proposed scheme 2 in Sect. IV-B
- Perfect CSI ··· known channel state information (CSI)
- LS (overall) ··· time domain approach in [5]
- 0th IPO ··· the 0-th order interpolation in freq. domain
- 1st IPO ··· the 1-st order interpolation in freq. domain
- MMSE IPO ··· MMSE interpolation in freq. domain [13]

A. System Parameters

The number of subcarriers and the total number of pilot subcarriers are set to be \( M = 128 \) and \( |Z| = 32 \), respectively. For the subcarrier allocation, we assume both the localized and the distributed allocations, and the numbers of subcarriers \( |S_n| \) and pilot subcarriers \( |Z_n| \) allocated to the \( n \)-th user are set to be the same for all \( n = 0, 1, \ldots, N - 1 \) for both allocation strategies. The pilot subcarriers are located with equal spacing within the allocated subcarriers for the case of the localized subcarrier allocation, while their positions are randomly selected from the allocated subcarriers for the distributed subcarrier allocation. QPSK modulation with coherent detection is employed as the baseband modulation/demodulation schemes. Frequency selective Rayleigh fading channel with the order of up to \( L = 16 \) is assumed to generate \( h_n \), where the number of paths (i.e., the number of nonzero elements in \( h_n \)) is set to 2 or 5, the temporal positions of the paths are randomly selected for each
simulation trial within the channel order, and the powers of all paths are uniform. As the solver for the $l_1$-$l_2$ optimization problem in the proposed schemes, we have employed FISTA (Fast Iterative Shrinkage-Thresholding Algorithm) [14]. For Proposed 1 and LS (overall), the number of nonzero elements in $h$ is assumed to be $L + 2\alpha + 1 = 29$ by setting $\alpha = 6$, and the performance of these methods is evaluated only for the localized subcarrier allocation. The number of trials for all simulations are set to $N_{\text{trial}} = 4,000$ times.

B. BER Performance

Figures 3 and 4 show the average bit error rate (BER) performance for the localized subcarrier allocation with the different number of paths and users when the demodulation using frequency domain equalization is performed based on the estimated channel frequency responses using the proposed and the conventional channel estimation schemes. From the figures, we can see that the performance of Proposed 1 is almost the same as that of LS (overall), which is worse than the performance of MMSE IPO. This is because the overall channel impulse response $h'$ is not sufficiently sparse even when $h_n$ is sparse due to the discontinuous change of frequency response at the allocation boundaries. On the other hand, Proposed 2 achieves the best performance among the feasible estimation schemes. Although the performance of Proposed 2 is degraded in Fig. 4 compared to Fig. 3 because of the reduction of the number of pilot signals per user as well as the reduction of the sparsity of $h_n$, it still outperforms MMSE IPO, which is widely used in existing systems. This is because MMSE IPO does not take advantage of the sparsity of the channel impulse response.

The average BER performance for the distributed subcarrier allocation is depicted in Figs. 5 and 6 for different number of paths and users. We can see that Proposed 2 achieves the best BER performance among the feasible channel estimation schemes for the case of the distributed allocation as well. However, by comparing the BER performance of Proposed 2 in Figs. 4 and 6, it can be observed that the BER performance for the case with the distributed allocation is worse than the case with the localized allocation. This is somewhat strange from the viewpoint of compressed sensing, because it is well known that the compressed sensing using the sub-DFT matrix obtained by the random selection of the rows as the sensing matrix can achieve the exact reconstruction under certain conditions [15]. In the next section, we clarify the reason by evaluating the channel estimation error.

C. Channel Estimation Error

Figures 7 and 8 show the normalized root mean-square-error (RMSE) of the channel impulse response, which is calculated as

$$\sqrt{\frac{\sum_{k=1}^{N_{\text{trial}}} \sum_{n=1}^{N} ||\hat{h}_n(k) - h_n(k)||^2}{\sum_{k=1}^{N_{\text{trial}}} \sum_{n=1}^{N} ||h_n(k)||^2}}. \quad (43)$$

where $h_n(k)$ denotes the impulse response of the $n$-th user’s channel in the $k$-th simulation trial and $\hat{h}_n(k)$ is the corresponding estimated channel response with Proposed 2. From the figures, we can see that the estimation accuracy of $h_n$ with Proposed 2 is better for the case with the distributed subcarrier allocation than for the case with the localized allocation. These results suggest that the average channel estimation error for all the subcarriers for a user with Proposed 2 is smaller for the case with the distributed subcarrier allocation, but the estimation error evaluated only on the allocated subcarriers is smaller for the case with the localized allocation. In order to confirm the validity of the discussion, the channel estimation error for each subcarrier is shown in Figs. 9 and 10 for the localized and the distributed subcarrier allocations, respectively. From the figures, we can confirm indeed that the channel estimation error within the allocated subcarriers is smaller for the case with the localized subcarrier allocation.
VI. CONCLUSION

This paper has proposed channel estimation schemes in time domain for uplink OFDMA systems. The key idea in the proposed schemes is the utilization of compressed sensing assuming the sparsity of the channel impulse response. The performance of the proposed schemes is evaluated via computer simulations in comparison with that of the conventional schemes, and it has been demonstrated that, even for the uplink OFDMA channel estimation, the time domain approach is beneficial if the sparsity of the channel impulse response can assumed.

Future works include the investigation of subcarrier and/or pilot subcarrier allocation schemes suited for the proposed channel estimation schemes, and the analysis of the performance of Proposed 2 with the localized subcarrier allocation.

REFERENCES

Fig. 9. Per-subcarrier normalized RMSE of the overall frequency response with proposed scheme 2 (localized subcarrier allocation).

Fig. 10. Per-subcarrier normalized RMSE of the overall frequency response with proposed scheme 2 (distributed subcarrier allocation).


