A Carrier Interferometry Based Channel Estimation Technique for MIMO-OFDM/TDMA Systems

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SUMMARY This paper proposes a channel estimation technique for the dynamic parameter controlled—orthogonal frequency and time division multiple access (DPC-OF/TDMA) systems studied as one of the candidates of the beyond third generation (B3G) systems. In the proposed scheme, the impulse response, which represents the channel state information (CSI) is estimated using carrier interferometry (CI) which is equivalent to impulse signal transmission. Moreover, because the minimum number of subcarriers allocated to terminals is 64, in order to estimate a CSI with its spreading factor of 64, we employ a code-multiplexed CI signal for a cell search process and a time-domain-multiplexed CI signal for transmit antenna identification. Furthermore, we also propose a flexible CSI estimation scheme that supports two cases: multiple subchannel block assignment and MIMO transmission cases. Computer simulation confirms that the proposed scheme can estimate the CSI with high accuracy.

key words: DPC-OF/TDMA systems, MIMO-OFDM/TDMA systems, channel estimation, carrier interferometry, time window

1. Introduction

With extensive research on broadband wireless access techniques, various types of beyond third generation (B3G) systems are under investigation \([1,2]\). One of the most important requirements for the B3G systems is to support data rate of more than 100 Mbit/s with its bandwidth of about 100 MHz using a single-input single-output (SISO) scheme and around 1 Gbit/s by further employing multiple-input multiple-output (MIMO) \([3–6]\) techniques \([7]\).

As one of the possible schemes for B3G systems, we have proposed a dynamic parameter controlled orthogonal frequency and time division multiple access (DPC-OF/TDMA) system \([8]\) based on one-cell reuse OFDM/TDMA systems \([9]\). In this system, all the subcarriers are segmented every 64 subcarriers, each of which will be called a subchannel block in the following, and terminals can use arbitrary number of subchannel blocks depending on functionality and constraint conditions of each terminal, e.g., a small terminal such as a wristwatch-type terminal can use only one subchannel block (64 subcarriers) whereas a large terminal such as a PC might use all the subchannel blocks if available. Moreover, such channel estimation scheme should also be applicable to MIMO systems, because MIMO is one of the most important system level spectral efficiency enhancement technologies for cellular systems while keeping backward compatibility with existing SISO-based systems. Therefore, it is required for its channel estimation scheme to have an accuracy (resolution in frequency domain) depending on the number of subchannel blocks to be used in both SISO and MIMO cases.

When a MIMO scheme is introduced, we have to measure the received power from all the connectable base stations (BSs), as well as to identify all the channel state information (CSI) for all the combination of the transmit and receive antenna elements for each BS. For this purpose, we have to measure an impulse response between any combination of transmit and receive antenna elements for all the connectable BS antennas.

One of the most typical channel estimation technique is to employ a code division multiplexing (CDM) based scheme in which a unique spreading code is assigned to identify both BS and antenna element. When a terminal employs only one subchannel block in which 64 subcarriers are included, we can apply a spreading code length of 64, which is not a sufficient number for identification of both BSs and antenna elements. For example, when the number of detectable surrounding cells is 10 and the number of transmit antenna elements in each BS is 8, the total number of the required codes is 80 (=10 cells \* 8 antenna elements), which is much larger than the number of spreading code to be prepared using a spreading factor of 64. This means that channel estimation for one subchannel block (64 subcarriers) transmission is the severest case than the cases for plural subchannel blocks usage.

Therefore, this paper will propose a channel estimation technique using a carrier interferometry (CI) \([11–13]\) combined with CDM in the frequency domain to estimate CSI with smaller number of spreading factor even in MIMO introduced systems. In the CI technique, when the same value of the amplitude and phase are assigned to all the subcarriers in the OFDM signal, its time-domain signal corresponds to an impulse having a sinc waveform. Moreover, when a linear phase shift is applied to the signal, timing of the impulse can be controlled by applying a linear phase shift in the fre-
quency domain. Based on this concept, we have assigned an antenna element specific time shift to each antenna element in each BS. In this scheme, the amount of time shift is designed to be sufficiently large so that the impulse responses for all the antenna elements do not overlap to each other. Moreover, the CI signal is spread in the frequency domain using a BS-specific spreading code to identify the source BS for each delay profile, which is also effective in suppressing peak level of the CI-based pilot signal.

Computer simulation confirms that the proposed scheme can estimate the CSI with high accuracy.

2. DPC-OF/TDMA Systems

Figure 1 and Fig. 2 show downlink and uplink frame structures for the DPC-OF/TDMA systems. As shown in this figure, three types of slots are multiplexed on the TDMA frame in the time domain [8].

1. FCMS (frame control message slot):
   This is a downlink control slot including system information and current status of user allocation of each slot for the downlink.
2. MDS (message data slot):
   This is mainly used for user data transmission
3. ACTS (activation slot):
   This is an uplink control slot for association request and connection/disconnection request for each terminal.

Moreover, in each time slot, all the subcarriers (1024 subcarriers) are segmented into 16 “subchannel blocks,” each of which consists of 64 subcarriers. Each subchannel block is controlled independently and terminals may use plural number of subchannel blocks according to the functionality and constraint conditions of each terminal. For example, a small terminal such as mobile phones can use only one subchannel block, whereas a large terminal such as PCs can use plural subchannel blocks.

Therefore, in this system, whole the radio resource is segmented in both time and frequency domains, thereby this system can support wide-range of user rate ranging from the minimum rate achieved by using one time slot in one subchannel to 128 (= 16 subchannels * 8 time slots) times higher rate achieved by dominating whole the radio resource, that is, this system can support various types of terminals using the same radio interface.

However, in this case, because the minimum control unit assigned to terminals consists of 64 subcarriers and terminals which can use only one subchannel (64 subcarriers) exist in the same system, it is necessary for channel estimation scheme to be operatable for 64-subcarrier usage cases, that is,

1. impulse response estimation
2. cell search process to identify all the connectable BSs should be operated with 64 subcarriers in both SISO and MIMO cases. Moreover, when the terminal employ plural subchannel blocks,
3. flexible channel estimation scheme regardless of the number of subchannels assigned to the terminals is also required. Especially, when a MIMO technique is introduced, we must identify impulse responses for all the combination of the transmit and receive antenna elements.

3. Proposed Scheme

3.1 Carrier Interferometry

Figure 3 shows a concept of carrier interferometry (CI) technique. In this figure, CI signal is generated by assigning the
same amplitude and phase to all the subcarriers of OFDM signal. CI signal has the following characteristics.

1. Impulse

Because time-domain waveform of CI signal is an inverse Fourier transform of a rectangle spectrum, a CI signal becomes an impulse signal having sinc waveform. The transmitted CI signal \( s(t) \) is represented as

\[
s(t) = \int_{-B}^{B} A \cdot e^{j2\pi ft} df = AB \frac{\sin(\pi B \tau)}{\pi B \tau}
\]

(1)

In this equation, \( B \) is total bandwidth and \( A \) is the amplitude for each subcarrier.

2. Timing control

The timing of impulse can be controlled by assigning a linear phase offset to each subcarrier, which can be represented as the following equation:

\[
\int_{-B}^{B} A e^{-j2\pi ft} \cdot e^{j2\pi ft} = s(t - \tau)
\]

(2)

In the digital signal processing, when the number of subcarriers is \( N_{\text{sub}} \) and the complex amplitude vector for subcarriers is represented as \( S = [1 \ 1 \ldots \ 1]^T \), the CI signal in the time domain \( s \) is obtained as

\[
s = F^H S = \begin{bmatrix} \sqrt{N_{\text{sub}}} & 0 & \ldots & 0 \end{bmatrix}^T
\]

(3)

where \( F \) is a \( N_{\text{sub}} \times N_{\text{sub}} \) discrete Fourier transform (DFT) matrix represented as

\[
F = \frac{1}{\sqrt{N_{\text{sub}}}} \begin{bmatrix} 1 & e^{-j2\pi \frac{1}{N_{\text{sub}}}} & \ldots & e^{-j2\pi \frac{N_{\text{sub}} - 1}{N_{\text{sub}}}} \\
1 & e^{-j2\pi \frac{1}{N_{\text{sub}}}} & \ldots & e^{-j2\pi \frac{N_{\text{sub}} - 1}{N_{\text{sub}}}} \\
\vdots & \vdots & \ddots & \vdots \\
1 & e^{-j2\pi \frac{N_{\text{sub}} - 1}{N_{\text{sub}}}} & \ldots & e^{-j2\pi \frac{N_{\text{sub}} - 1}{N_{\text{sub}}}} 
\end{bmatrix}
\]

(4)

Equation (3) means that the CI signal is equivalent to an impulse signal. Thus, when the CI signal is transmitted in association with the traffic channel signal, we can measure the impulse response for the traffic channel in the receiver. Moreover, when a linear phase offset is imposed to subcarriers, i.e., the complex amplitude vector of \( S_{N_t} = [1 \ e^{-j2\pi \frac{N_{\text{sub}}}{N_t}} \ldots e^{-j2\pi \frac{N_{\text{sub}} - 1}{N_t}}]^T \) is assigned to subcarriers, where \( N_t \) is the slope of phase offset in the frequency domain, the CI signal vector \( s_{N_t} \) is obtained as

\[
s_{N_t} = F^H S_{N_t} = \begin{bmatrix} 0 & \ldots & 0 \sqrt{N_{\text{sub}}} \ldots & 0 \end{bmatrix}^T
\]

(5)

Equation (5) means that the timing of impulse is delayed by \( N_t \) sample period in the time domain. This concept is shown in Fig. 4. As shown in this figure, the CI technique can flexibly control timing of impulse according to the amount of the linear phase offset. Therefore, we can multiplex the impulse responses without overlapping to each other on the time domain, and we can identify the CSI for all the combination of transmit and receive antenna elements.

However, there is one serious disadvantage of the CI technique, i.e., the CI signal has very high peak-to-average power ratio (PAPR) by concentrating energy of all the subcarriers into a certain sampling time. To solve this problem, we will multiply a BS-specific pseudo noise (PN) code in the frequency domain so as to spread energy in the time domain as well as the identification of the source BS [10].

Therefore, in the proposed scheme, by employing a CI technique combined with CDM techniques, we can estimate the CSI for all the combination of transmit and receive antenna elements and operate cell search process with smaller number of spreading factor.

3.2 Channel Estimation Technique

3.2.1 Pilot Signal

Figure 5 shows a pilot signal generator. 64 subcarriers (one
MIMO techniques.

Fig. 6 Transmitter and receiver configuration in the case of employing given a transmit antenna specific linear phase offset to delay impulse response timing for each antenna so as not to overlap the received impulse response timing to each other. In the proposed system, the same pilot signal is applied to all the subchannels so that any terminal, even though it can detect only one subchannel block, can conduct cell search and channel estimation process with the same manner regardless of the assigned subchannel block position. In this case, however, peak power is high because only every 16 sample has a value due to the same spectrum repetition in the frequency domain. To suppress such a high peak power, a part of a known PN sequence with its length of 16 bits is multiplied to each subchannel block, where the first bit is multiplied to all the subcarrier in the first subchannel block of the pilot signal, the second bit is multiplied to all the subcarrier in the second subchannel block, and so on. Moreover, the same pattern is employed for all the BSs. Such a PN pattern multiplication process does not cause any pattern synchronization problem as long as carrier frequency is synchronized.

3.2.2 Impulse Response Estimation

In this section, we will explain the proposed technique under no fractional time delayed paths conditions (We will explain the channel estimation with fractional time delayed paths in the next section).

Figure 6 shows a block diagram to identify the CSI for any combination of the transmit and receive antenna elements in a MIMO system, where two transmit antenna case is shown in the figure. In the proposed scheme, the most important concept of this scheme is that the obtained impulse responses are multiplexed on the time domain without overlapping and each impulse response is extracted by using a time window with its length of the maximum delay time of multipaths.

In the transmitter, when the number of subcarriers is \( N_{sub} \), the longest delay time of multipaths is \( N_g \) sample period, and the number of transmit and receive antenna elements are \( N_{Tx} \) and \( N_{Rx} \), the transmitted signal of \( k \)th subcarrier for \( l \)th (\( l = 1, \ldots, N_{Tx} \)) transmit antenna element is given by

\[
S_l(k) = c(k) e^{-j \frac{2\pi l - 10 N_k}{N_{sub}}} \tag{6}
\]

where \( c(k) \) is the BS-specific complex PN code multiplied to \( k \)th subcarrier in the pilot symbol and \( e^{-j \frac{2\pi l - 10 N_k}{N_{sub}}} \) is a transmit antenna specific linear phase offset.

In the receiver, after the received signal is converted to the frequency-domain signal, the obtained signal is despread in the frequency domain. When the frequency transfer function between \( l \)th transmit antenna element and \( m \)th (\( m = 1, \ldots, N_{Rx} \)) receive antenna element is \( H_{m,l}(k) \), the obtained signal \( R_{m}(k) \) of \( m \)th receive antenna is represented as

\[
R_{m}(k) = \frac{c^*(k)}{|c(k)|^2} \sum_{l=1}^{N_{Tx}} H_{m,l}(k)c(k)e^{-j \frac{2\pi l - 10 N_k}{N_{sub}}} + \eta_m(k) \tag{7}
\]

where \( \eta_m(k) \) is frequency-domain additive white Gaussian noise (AWGN) component. Because transmitted signals from all the transmit antenna elements are superimposed in this signal, when the number of multipaths is \( l_{m,l} \) and complex channel gain of \( l \)th path in the time domain is \( h_{m,l}^d \), the time-domain signal after inverse fast Fourier transform of \( R_{m}(k) \) is given by

\[
\hat{h}_m(n) = \sum_{l=1}^{N_{Tx}} \sum_{j=0}^{l_{m,l}-1} h_{m,l}^d(n-(l-1)N_g - \tau_i) + \nu_m(n) \tag{8}
\]

where \( \nu_m(n) \) is time domain AWGN component which satisfies \( E[|\nu_m(n)|^2] = \sigma^2 \) (noise power) and \( \tau_i \) is the delay time of \( l \)th path. In Eq. (8), if impulse responses are not overlapped (\( \tau_{m,l} \leq N_g \)), impulse response for each transmit antenna can be resolved in the time domain by using a time window \([l-1)N_g \leq N_g \] for \( l \)th transmit antenna element. As the result, in no fractionally time-delayed paths existence case, impulse response between \( l \)th transmit antenna and \( m \)th receive antenna element can be estimated as

\[
\hat{h}_{m,l}(n) = \sum_{j=0}^{l_{m,l}-1} h_{m,l}^d(n-(l-1)N_g - \tau_i) + \nu_{m,l}(n) \tag{9}
\]

where \( \nu_{m,l}(n) \) is time-windowed AWGN signal given by

\[
\nu_{m,l}(n) = \begin{cases} 0 & (0 \leq n < (l-1)N_g) \\ \nu_m(n) & ((l-1)N_g \leq n \leq lN_g) \\ 0 & (lN_g < n \leq N_{sub}) \end{cases} \tag{10}
\]
Finally, the estimated impulse response is converted to the frequency transfer function and we can estimate the CSI from each transmit antenna element by compensating for the transmit antenna specific linear phase offset.

3.2.3 Impulse Response Estimation in Fractional Time Delayed Paths Existence Case

Because the proposed scheme is applied to OFDM-based systems, and waveform of the pilot signal has compatibility with the OFDM signal, we will assume ideal rectangular frequency transfer function for both the transmit and receive filters; i.e., they do not produce any distortion to the signal, although out-of-band noise and interference is eliminated. In the proposed scheme, when we disregard frequency domain spreading process for BS identification, the pulse shape of the transmitted pilot signal has a sinc-like waveform. As a result, delay profile of the received signal is not the impulse response of the channel but the convolution of the pulse waveform and the channel impulse response as shown in Fig. 7. Fortunately, the frequency domain transfer function measured in the receiver is exactly the composite channel characteristics of the pulse waveform and the channel. Therefore, when the measured frequency domain transfer function is converted to the time domain using an IFFT, we can obtain the desired composite impulse response identified by the sampling time spaced discrete impulse response model.

Figure 8 shows an example of the identified impulse response. When a delay component is located at a sampling time spaced positions, the impulse response start from zero timing and it will last until the guard interval position at most. On the other hand, when a delay components is located at a fractional time delayed position, there is also a non-zero value response before the peak position due to the pulse waveform. Even in this case, we can observe a period that impulse response level is very low (shaded area in Fig. 8). In the proposed scheme, impulse response having such low power level is discarded using a time window to suppress noise and interference power included in the measured channel characteristics. Therefore, we preliminary give a cyclic time shift to the to bring $N_t$ tail samples of the impulse response to the head for ease of time division multiplexing of plural impulse responses from different transmit antenna elements, which can be done by giving a linear phase offset to each subcarrier. Figure 9 shows the outline of this process. In the receiver, we extract each impulse response by a time window ($N_g$ sample period) according to the expected longest delay time of identified impulse response. Figure 10 shows transmission and reception process for the channel measurement related sections based on this concept. In this figure, in the 1st transmit antenna, a linear phase offset is given to each subcarrier to equivalently give by $N_t$ sample period delay. On the other hand, $N_g$ sample period is given to the 2nd transmit antenna as the delay time so as to prevent delay profile overlapping to each other in the receiver.

3.2.4 Flexible Channel Estimation for Plural Subchannel Blocks

When a terminal uses plural subchannel blocks, we should flexibly estimate CSI with maximum resolution determined by the allocated number of subchannel blocks. For this pur-
Transmitter

Receiver

Fig. 10 Transmitter and receiver configuration that can cope with fractionally time-delayed paths existence case.

Fig. 11 Phase of each subcarrier in 2 subchannel blocks assignment case.

pose, when plural subchannel blocks are allocated, we will apply another phase shift in the received signal so that phase shift in all the allocated subchannel blocks becomes a desirable linear phase shift. Figure 11 shows its concept in the case of two subchannel block allocation. When the number of subcarriers in one subchannel block is \( N_{\text{sub}} \) and the number of assigned subchannel blocks is \( N_{\text{ch}} \), the allocated phase of \( k \)th subcarrier \( p_l(k) \) (\( k = 0, \ldots, N_{\text{ch}}N_{\text{sub}} - 1 \)) of \( l \)th transmit antenna element is given by

\[
p_l(k) = \begin{cases} 
  e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} & (0 \leq k \leq N_{\text{sub}} - 1) \\
  e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}N_{\text{sub}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} & (N_{\text{sub}} \leq k \leq 2N_{\text{sub}} - 1) \\
  \vdots \\
  e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}N_{\text{sub}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} & ((s-1)N_{\text{sub}} \leq k \leq sN_{\text{sub}} - 1) \\
  \vdots \\
  e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}N_{\text{sub}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} & ((N_{\text{ch}} - 1)N_{\text{sub}} \leq k \leq N_{\text{ch}}N_{\text{sub}} - 1) 
\end{cases}
\] (11)

In the receiver, when the phase offset of \(-(s-1) \cdot 2\pi\) is given to all the subcarriers in the \( s \)th subchannel block, because of \( e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}N_{\text{sub}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} = e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}N_{\text{sub}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} \), the phase shift in all the assigned subchannel blocks becomes the following equation:

\[
p_l(k) = e^{-j2\pi \left( \frac{(l-1)N_{\text{ch}}}{N_{\text{ch}N_{\text{sub}}}} \right) k} & (0 \leq k \leq N_{\text{ch}}N_{\text{sub}} - 1) 
\] (12)

This means that Eq. (11) is equivalent to a desirable linear phase offset. Therefore, even when a terminal uses plural subchannel blocks, we can flexibly estimate the CSI by using the same way as one subchannel block usage case.

4. Computer Simulation

4.1 SISO Transmission Performances

4.1.1 Simulation Parameters

Table 1 shows simulation parameters in the SISO transmission case.

In this simulation, we employ an OFDM based adaptive modulation scheme (OFDM AMS) with its selectable modulation schemes of binary phase shift keying (BPSK), quaternary PSK (QPSK), 16-ary quadrature amplitude modulation (QAM) and 64 QAM as modulation schemes with its coding rate of 3/4. Path model is exponentially decaying 32-path Rayleigh fading with its delay spread of 250 nsec and the maximum Doppler frequency is 20 Hz.

In the OFDM AMS algorithm, a terminal estimates the received SNR for each subcarrier and selects an appropriate modulation scheme for each subcarrier. The selected modulation parameter for each subcarrier is then fed back as a requested modulation and coding schemes (RMCS) to the BS (transmitter) side. In the BS, the data sequence is modulated according to the notified RMCS.

Figure 12 shows a TDMA slot format. Each slot includes 5 preamble symbols, 19 data symbols, and slot guard of one symbol. In the preamble part, the proposed pilot symbol is allocated in one symbol at the head of preamble and four symbols are used for the notification of RMCS and transmission power control (TPC) commands. In the data part, each OFDM symbol is transmitted according to
Table 1  Simulation parameters. (SISO mode)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>OFDM symbol rate</td>
<td>100 ksymbol/s</td>
</tr>
<tr>
<td>Useful symbol duration</td>
<td>8 ( \mu )sec</td>
</tr>
<tr>
<td>GI duration</td>
<td>2 ( \mu )sec</td>
</tr>
<tr>
<td>Modulation (coding rate)</td>
<td>Carrier Hole, BPSK, QPSK, 16QAM, 64QAM (( r = 3/4 ))</td>
</tr>
<tr>
<td>FEC</td>
<td>Convolutional coding (constraint length ( K = 7 ))</td>
</tr>
<tr>
<td>Target QoS</td>
<td>bit error rate = ( 10^{-5} )</td>
</tr>
<tr>
<td>Num. of subcarriers</td>
<td>64–1024</td>
</tr>
<tr>
<td>Path loss model</td>
<td>ITU-R M.1225 indoor to outdoor pedestrian test environment [14]</td>
</tr>
<tr>
<td>Shadowing</td>
<td>log normal distribution (( \sigma = 8 ) dB)</td>
</tr>
<tr>
<td>Path model</td>
<td>exponentially decaying 32-path Rayleigh fading (r.m.s. delay spread = 250 nsec)</td>
</tr>
<tr>
<td>Max. Doppler frequency ( f_D )</td>
<td>20 Hz</td>
</tr>
<tr>
<td>Ave. pilot to data symbol power ratio</td>
<td>0 dB</td>
</tr>
</tbody>
</table>

Fig. 12  TDMA slot format.

the OFDM AMS algorithm.

Because purpose of this simulation is to evaluate accuracy of the estimated CSI, notification of RCS and TPC command is assumed to be perfectly operated. Moreover, the synchronization is operated perfectly because the synchronization technique has been developed in [10] and band limitation is ideally operated.

4.1.2 Simulation Results

In the following, we will evaluate performances of the proposed channel estimation scheme. Although our main target is evaluation of the proposed scheme in multicell and MIMO conditions, this section will conduct evaluation of flexibility to variable number of the assigned subchannel block under SISO and single cell conditions as the basic performance evaluation stage. We will also assume that the delay time of multipath components is a multiple of a sample period. Then, we will evaluate the impact of the proposed channel estimation on throughput performances in MIMO cases including system level throughput in multicell conditions in Sect. 4.2.

(1) Performances in one subchannel block usage case

Figure 13 shows the average received SNR of pilot signal after time windowing process versus the average received SNR of data symbol, and Fig. 14 shows the throughput in data part versus received SNR of data symbol in SISO case when a terminal uses one subchannel block. “Perfect” is the performances when CSI is perfectly estimated and “Estimation” is the performance of the proposed scheme usage case. As shown in this figure, we can obtain 6 dB gain of SNR in channel estimation because time window that corresponds to the expected maximum delay time of the multipath components is set to a quarter of a useful symbol duration in this simulation.

In Fig. 14, “w/o time windowing” means the performances when channel frequency transfer function is directly measured after FFT, i.e., without time windowing process in the time domain. As can be seen from this figure, the proposed CSI estimation scheme has higher estimation accuracy, thereby achieving higher throughput compared to the “w/o time windowing” scheme. When the throughput performance for the proposed scheme is compared to the case of perfect CSI estimation, its degradation in terms of SNR is less than 1 dB. Therefore, we can confirm that the proposed scheme has acceptably high accuracy for CSI estimation.
(2) Performances in plural subchannel blocks usage case

Figure 15 shows the average received SNR of pilot signal versus the number of subchannel blocks assigned to the terminal, and Fig. 16 shows throughput versus the number of assigned subchannel blocks regardless the number of subchannel blocks assigned to the terminal. Actually, as shown in Fig. 16, we confirm that the proposed scheme can estimate CSI with high accuracy regardless of the number of assigned subchannel blocks.

These results show that the proposed scheme can flexibly estimate CSI regardless of the number of assigned subchannel blocks to the terminal.

(3) Performances in fractionally time-delayed paths existence case

In this section, we will evaluate accuracy of channel estimation in fractionally time-delayed paths existence case. In this simulation, the number of subcarriers is 64. Path model is 32-path Rayleigh fading model where each path arrives every 1/2-sample period. The other simulation parameters are the same as Table 1.

((3).1) Optimization of the duration of time window

Figure 17 shows the average normalized power delay profile in 1/2-sample delayed paths existence case. In this figure, we calculate the normalized delay profile by using the following equation:

$$\log_{10} \frac{E[|h_{eq}(l)|^2]}{E[|h_{eq}(0)|^2]} \quad (\text{dB})$$

where $h_{eq}(l)$ is complex gain of the obtained composite impulse response of $l$th discrete time. As can be seen from this figure, delay time of the delayed paths is getting longer due to sinc-like pulse waveform. However, we can observe low level period of the response. For example, $-30$ dB or lower period is located between 16th and 50th samples. Therefore, when we put a time window between 0 and 16th samples, and another time window between 50th and 64th samples, we can suppress noise and interference components located between 16th and 50th samples, and we can improve accuracy of the estimated channel characteristics. With these considerations, we set $N_t = 14$ and $N_g' = 32$ in the following.

Unfortunately, the corresponding time window size for multiplexing of delay profiles is twice more than the guard interval, which reduces the number of multiplexed delay profiles on the time domain. When required SNR for each selectable modulation scheme can be reduced by employment of more powerful channel coding, such as Turbo code, we can increase threshold level for time windowing process, thereby we can reduce $N_t$ and $N_g'$ and increase the number of multiplexed delay profiles. Or, when we focus the line of $-30$ dB, because impulse response components of more than $-30$ dB appear at 50th sample to 64th sample, we can set $N_t = 14$, thereby it is only necessary to set the duration of time window as twice as guard interval duration under this condition ($N_g' = 32$). When the expected maximum delay time is smaller, we can multiplexed more delay profiles, which will be considered as a future system design issue.
Figure 18 shows throughput performances. In this figure, “w/o fractionally time-delayed paths” is the performance in no fractionally time-delayed paths existence case, and “w/ fractionally time-delayed paths” is the performance in fractionally time-delayed paths existence case. As shown in this figure, we can confirm that the performances in fractionally time-delayed paths existence case can be degraded due to the longer duration of time window. However, the proposed scheme can be degraded by less than only 0.7 dB. Therefore, the proposed scheme can estimate channel state with high accuracy.

4.2 MIMO Transmission Performances

4.2.1 Simulation Parameters

Table 2 shows simulation parameters in the MIMO transmission case. In the MIMO simulation, we employ 4-by-4 MIMO technique and maximum likelihood detection (MLD) [15] as signal detection scheme. The number of subcarriers is 64 (one subchannel block). The maximum Doppler frequency is 20 Hz and path model is exponentially decaying 8-path Rayleigh fading model with its r.m.s. delay spread of 200 nsec.

In the MIMO case, we assume the downlink pilot signals are shared by all the terminals, and we improve accuracy of the estimated CSI by averaging it for each slot over \( N_{sym} \) time slots when the accuracy of the estimated CSI is low. Generally, in the TDMA system, the number of time slots in a frame is only one for each user. However, in the DPC-OF/TDMA system, because we apply packet reservation dynamic time-slotted multiple access (PR-DSMA) as a MAC protocol [16], continuous time slots are assigned to one user according to the IP data length. Therefore, in this simulation, each terminal can use time slots continuously.

As for the multi-cell conditions, cell radius is 100 m and each cell is wrapped by 6 cells to equivalently simulate co-channel interference for infinitely and continuously covered service area. We assume that terminals are located randomly in 7-cell area and each terminal searches a BS with the highest received signal power in the downlink. When adaptive modulation and coding scheme is applied, a terminal estimates average received signal to interference plus noise power ratio (SINR) and selects a MCS in each slot. The selected MCS is fed back to the BS as a RMCS, and the data sequence is coded and modulated according to the notified RMCS.

In this simulation, we will assume that TPC and RMCS are perfectly and ideally controlled, and the synchronization and band limitation are operated perfectly.

4.2.2 Simulation Results

(1) Optimization of \( N_{sym} \)

Figure 19 and Fig. 20 show the slot error rate (SER) and the throughput in data part versus the average received SNR per receive antenna in the case of QPSK with its coding rate of 1/2. In these figures, “CDM type pilot” is the performance in CDM based pilot signal usage case, where a conventional direct sequence spread spectrum type pilot signal is employed for delay profile measurements, and the average received SNR is defined as the following equation:

\[
SNR = r \cdot M \cdot N_{Tx} \cdot \frac{E_b}{N_0},
\]

where \( r \) is a coding rate, \( M \) is the number of data bits per symbol, \( N_{Tx} \) is the number of transmit antennas, and \( \frac{E_b}{N_0} \) is the average received energy per information bit to the noise power spectrum density ratio per receive antenna element.

When CDM based pilot signals are used for estimation of the delay profile, because there are four BS antennas for...
each BS and the number of multipath components are eight, we have to resolve 32 paths by despreading the pilot signal. When the number of spreading factor is restricted to 64, it is impossible to accurately measure the impulse response for each BS because the number of total paths, 32 paths, is too large, which means that we cannot accurately discriminate spatially multiplexed signals. That is why the slot error rate for CDM type pilot case shows almost one in Fig. 19 and throughput performance is almost zero in Fig. 20. Next, when $N_{sym}$ is one, the SER and throughput performances are severely deteriorated due to inter-stream interference caused by channel estimation errors. However, when $N_{sym}$ is four, the degradation of SER and throughput performances are reduced to 1 dB. Therefore, we can optimize $N_{sym}$ as four in the case of QPSK with its coding rate of 1/2.

As for the other modulation parameters, we can optimize the value of $N_{sym}$ with the same manner and its result is shown in Table 3.

(2) Performances in fractionally time-delayed paths existence case

In this section, we evaluate the performances in fractionally time-delayed paths existence case. In this simulation, path model is exponentially decaying 16 path Rayleigh fading model where each path arrives every 1/2 sample period, and its r.m.s. delay spread is 200 nsec. Time window size is 32 sample period ($N'_{g} = 32$), and we employ 2-by-2 MIMO techniques. In this case, two delay profiles for each BS antenna element can be detected in one pilot symbol period.

Figure 21 shows bit error rate performances. In this figure, “w/ fractionally time-delayed paths” is performances in fractionally time-delayed paths existence case and “w/o fractionally time-delayed paths” is performances in no fractionally time-delayed paths existence case. We can confirm the performances case are degraded in fractionally time-delayed paths existence case. It's because the accuracy is deteriorated due to longer time window and energy of composite impulse response is discarded a little. However, the degradation is 1 dB approximately. Therefore, we can confirm the proposed scheme can estimate CSI with high accuracy.

(3) Performances under multi-cell conditions

Figure 22 shows a cumulative distribution function (C.D.F.) of throughputs of data part under multi-cell conditions. When a total level TPC is employed in DPC-OF/TDMA systems, the transmit power for the pilot signal and data parts is different because TPC is not applied to the pilot signal part which is shared by all the users. Therefore, we have to estimate subcarrier-level interference plus noise power in the data part for proper operation of the adaptive modulation. Because an subcarrier-level interference power estimation technique is already reported in Ref. [17], we will assume perfect interference level estimation in this paper.

![Fig. 19](image1.png)  
**Fig. 19** SER versus average SNR per receive antenna.

![Fig. 20](image2.png)  
**Fig. 20** Throughput versus SNR per receive antenna.

![Fig. 21](image3.png)  
**Fig. 21** Bit error rate versus average received SNR per antenna.

<table>
<thead>
<tr>
<th>Table 3 Optimum value of $N_{sym}$.</th>
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<tr>
<td>modulation (coding rate)</td>
</tr>
<tr>
<td>BPSK ($r = 1/2$)</td>
</tr>
<tr>
<td>BPSK ($r = 2/3$)</td>
</tr>
<tr>
<td>BPSK ($r = 3/4$)</td>
</tr>
<tr>
<td>QPSK ($r = 1/2$)</td>
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<td>QPSK ($r = 3/4$)</td>
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<td>QPSK ($r = 5/6$)</td>
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<tr>
<td>QPSK ($r = 7/8$)</td>
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As shown in Fig. 22, we can confirm that probability for lower throughput is increasing due to estimation error for the proposed system. We can also find that the throughput performance for the proposed channel estimation employment case is located left to the perfect channel estimation case. This is also due to the channel estimation error for the proposed scheme. However, the difference is considered not to be serious, and we can also confirm more than 10 Mbit/s of the throughput is achievable with more than 99% of probability.

Therefore, we can confirm that the proposed channel estimation scheme has sufficiently high CSI high accuracy even in the MIMO transmission case, thereby we can realize one-cell reuse MIMO-OFDM/TDMA systems by using the CI-based channel estimation scheme.

5. Conclusion

In this paper, we have proposed a CI-based channel estimation technique for the DPC-OF/TDMA systems that can flexibly and accurately estimate CSI regardless of the assigned number of subcarriers in both SISO and MIMO cases.

Computer simulation confirmed that the proposed scheme can estimate CSI with high accuracy even with smaller number of spreading factor and flexibly estimate the CSI in plural subchannel blocks assignment case. Additionally, we confirmed that the proposed scheme can achieve high accuracy of the estimated CSI when it is applied to one-cell reuse MIMO-OFDM/TDMA systems.

As a future work, because the number of multiplexed impulse responses is small due to longer time window in fractional time delay of multipaths, we have to consider some countermeasures to increase the number of multiplexed impulse response with a process for some other system design parameter optimizations.

References

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