Preamble-Based Cell Identification for Cellular OFDM Systems

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***Abstract*—Preamble-based cell identification (CID) schemes are derived for orthogonal frequency division multiplexing (OFDM) systems. They include the optimal schemes based on the Bayesian and the maximum likelihood (ML) approaches, a suboptimal scheme that is a simplification of the ML scheme, and a differential decoding-based scheme that does not require any channel information. The complexities and performances of these CID schemes are examined and compared to existing schemes. The differential decoding-based scheme performs like the suboptimal scheme for most practical channels of interest and outperforms existing schemes, yet it is simpler to implement than the others.**

***Index Terms*—Cell identification, orthogonal frequency division multiplexing, preamble, IEEE 802.16.**

I. Introduction

In cellular orthogonal frequency division multiplexing (OFDM) systems, such as IEEE 802.16e [1], cells are distinguished by the cell-specific pattern in the preamble[1]. After the synchronization process in a receiver is completed, the pattern in the received preamble signal can be identified from a set of candidate patterns. This process is called cell identification (CID). Conventional methods for CID are based on some kind of cross-correlations between the received signal and the candidate patterns [4], [5]. They are *ad hoc* techniques which are computationally efficient.

In this paper, we derive optimal and suboptimal schemes for CID and examine their performances and complexities. In addition, we derive an alternative differential decoding-based scheme, which can be thought of as a modification of the CID scheme in [5]. It will be shown that the proposed schemes outperform existing schemes. In particular, we shall show that the proposed differential decoding based CID can perform on a par with the suboptimal schemes and outperform existing schemes, yet its implementation is simpler than the others.

The organization of this paper is as follows. In Section II, we describe the system model of interest. In Section III, we discuss the derivation of several schemes: the optimal CID schemes based on the Bayesian approach and the maximum likelihood (ML) approach, a suboptimal scheme that is a simplified version of the ML approach, and a differential decoding-based scheme. The complexity of the CID schemes is also compared in Section III. In Section IV, we provide a comprehensive performance evaluation and comparison. Section V concludes the paper.

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[1]In some frequency hopping OFDM systems, it is suggested to identify cells from pilot tones [2], [3].

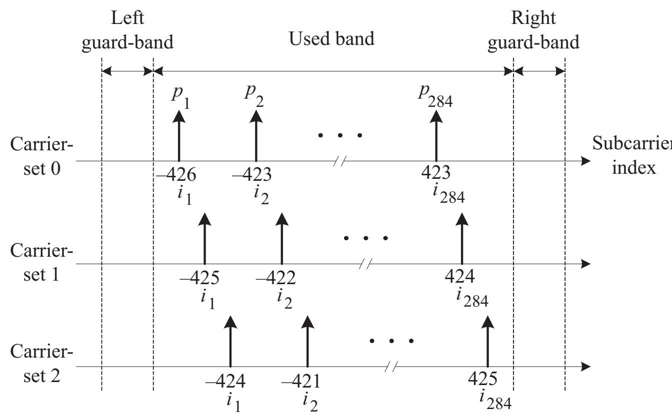

Fig. 1. Preamble structure of IEEE 802.16 system ($N_s = 1024$ and $N = 284$).

II. System model

We consider an OFDM system whose symbols consist of N_s subcarriers. The CID pattern is represented by a complex sequence with a span of $N \leq N_s$, and is denoted by $\mathbf{p} = [p_1, \ldots, p_N]^T$, where p_n is a complex number with $|p_n| = 1$ for all $1 \leq n \leq N$. The CID pattern $\mathbf{p}$ is taken from a finite set $\mathcal{P} = \{\mathbf{p}_1, \ldots, \mathbf{p}_M\}$. In a preamble containing the CID pattern, N subcarriers which are evenly distributed, ignoring the guard bands and DC subcarrier, are selected and modulated by p_n. We denote the index of the subcarrier carrying p_n by i_n, $1 \leq n \leq N$. The remaining subcarriers are left unmodulated and reserved for neighboring cells or sectors. Generally, in a cellular system, more than one neighboring cell or sector is required to be identified at the same time for the purpose of handover. We assume that all cells or sectors are synchronized to simultaneously transmit preambles. Then, if we let adjacent cells or sectors use non-overlapping sets of subcarriers for their preambles, they do not interfere with each other at the receiver, and thus we can identify all of them separately. For example, Fig. 1 shows the preamble structure of the IEEE 802.16e system [1] with $N_s = 1024$ and $N = 284$. There are three preamble carrier-sets. For the k-th carrier set, $k \in \{0, 1, 2\}$, every third subcarrier in the used band is modulated by p_n, starting with $i_1 = -426 + k$. By assigning different carrier-sets to adjacent cells or sectors, co-channel interference can be minimized.

The preamble is transmitted over a frequency selective channel whose impulse response is given by $\mathbf{h} = [h_1, \ldots, h_L]^T$. The channel is assumed to be quasi-static and is invariant during the preamble transmission. At a receiver, the received signal $\mathbf{y} \in \mathbb{C}^N$ denotes the vector containing the N_s-point discrete Fourier transform (DFT) output at the preamble subcarrier locations $\{i_1, i_2, \ldots, i_N\}$. Then we have

$$\mathbf{y} = \mathbf{PBh} + \mathbf{w}, \tag{1}$$

where $\mathbf{P} \in \mathbb{C}^{N \times N}$ is a diagonal matrix given by

$$\mathbf{P} \triangleq \mathrm{diag}(\mathbf{p}) = \mathrm{diag}\{p_1, \ldots, p_N\}, \tag{2}$$

$\mathbf{B} \in \mathbb{C}^{N \times L}$ is a matrix representing the effect of DFT given by

$$[\mathbf{B}]_{n,k} = e^{-j2\pi(k-1)i_n/N_s}, \; n = 1, \ldots, N, \; k = 1, \ldots, L, \tag{3}$$

and $\mathbf{w} \in \mathbb{C}^N$ is a zero-mean circularly symmetric complex Gaussian noise vector with covariance matrix $\sigma^2\mathbf{I}_N$, i.e., $\mathbf{w} \sim \mathcal{CN}(\mathbf{0}, \sigma^2\mathbf{I}_N)$. Since $|p_n| = 1$ for $n \in \{1, \ldots, N\}$, $\mathbf{P}$ is a unitary matrix, i.e., $\mathbf{P}^H\mathbf{P} = \mathbf{P}\mathbf{P}^H = \mathbf{I}_N$. Then the conditional probability density function (pdf) of $\mathbf{y}$ given $\mathbf{p}$ and $\mathbf{h}$ is given by

$$f(\mathbf{y}|\mathbf{p}, \mathbf{h}) = \frac{1}{(\pi\sigma^2)^N} \exp\left\{-\frac{1}{\sigma^2}\|\mathbf{y} - \mathbf{PBh}\|^2\right\}. \tag{4}$$

III. Derivation of CID Schemes

We derive optimal and suboptimal CID schemes on the basis of the Bayesian and the ML approaches as well as a differential decoding-based scheme that requires no channel information. In addition, we examine the computational complexities of the schemes.

A. Bayesian Approach to CID

The Bayesian formulation, or the *maximum a posteriori* (MAP) formulation for identifying $\mathbf{p}$ from $\mathbf{y}$ in (1) is given by

$$\begin{aligned}\hat{\mathbf{p}}_{\mathrm{B}} &= \arg\max_{\mathbf{p} \in \mathcal{P}} f(\mathbf{p}|\mathbf{y}) \\ &= \arg\max_{\mathbf{p} \in \mathcal{P}} f(\mathbf{y}|\mathbf{p}),\end{aligned} \tag{5}$$

where the second equality is from the assumption that $\mathbf{p}$ is uniform over $\mathcal{P}$, i.e., $\Pr\{\mathbf{p} = \mathbf{p}_i \in \mathcal{P}\} = 1/M$. The conditional pdf $f(\mathbf{y}|\mathbf{p})$ is obtained by taking an expectation of $f(\mathbf{y}|\mathbf{p}, \mathbf{h})$ in (4) with respect to the channel $\mathbf{h}$. Thus, we need prior knowledge of the pdf of $\mathbf{h}$. If we assume that $\mathbf{h}$ is distributed as $\mathcal{CN}(\mathbf{0}, \mathbf{C})$, then (5) reduces to

$$\hat{\mathbf{p}}_{\mathrm{B}} = \arg\min_{\mathbf{p} \in \mathcal{P}} \|\mathbf{G}_{\mathrm{B}}\mathbf{Y}\mathbf{p}^*\|^2, \tag{6}$$

where $\mathbf{Y} \triangleq \mathrm{diag}(\mathbf{y})$ and

$$\mathbf{G}_{\mathrm{B}} \triangleq \mathbf{I}_N - \mathbf{BCB}^H(\mathbf{BCB}^H + \sigma^2\mathbf{I}_N)^{-1}. \tag{7}$$

The scheme in (6) will be referred to as the Bayesian CID (B-CID).[2] This method requires knowledge of the channel statistics, especially the covariance matrix $\mathbf{C}$.

[2] This scheme can also be thought of as an ML scheme because, as shown in (5), the MAP criterion reduces to the ML criterion when $\mathbf{p}$'s are equally likely. Even so, we simply call this scheme the Bayesian scheme to avoid confusion with the scheme in Section III-B.

B. Maximum Likelihood Approach to CID

The ML solution for identifying $\mathbf{p}$ from $\mathbf{y}$ in (1) is given by

$$\hat{\mathbf{p}} = \arg\max_{\mathbf{p} \in \mathcal{P}} f(\mathbf{y}|\mathbf{p}, \mathbf{h}) = \arg\min_{\mathbf{p} \in \mathcal{P}} \|\mathbf{y} - \mathbf{PBh}\|^2, \tag{8}$$

where the conditional pdf $f(\mathbf{y}|\mathbf{p}, \mathbf{h})$ is given in (4). Since $\mathbf{h}$ is unknown, $\|\mathbf{y} - \mathbf{PBh}\|^2$ in (8) should be minimized with respect to both $\mathbf{p}$ and $\mathbf{h}$. This minimization can be achieved through the following two step procedure [9]. First, $\|\mathbf{y} - \mathbf{PBh}\|^2$ is minimized with respect to $\mathbf{h}$ for a given $\mathbf{p}$. Assuming $N > L$, this yields the following ML estimate [6] of $\mathbf{h}$:

$$\hat{\mathbf{h}}_{\mathrm{JML}} = (\mathbf{B}^H\mathbf{B})^{-1}\mathbf{B}^H\mathbf{P}^H\mathbf{y}. \tag{9}$$

Second, we substitute $\mathbf{h}$ in (8) with $\hat{\mathbf{h}}_{\mathrm{JML}}$ in (9) to get

$$\left\|\mathbf{y} - \mathbf{PB}\hat{\mathbf{h}}_{\mathrm{JML}}\right\|^2 = \left\|\mathbf{G}_{\mathrm{JML}}\mathbf{P}^H\mathbf{y}\right\|^2, \tag{10}$$

where

$$\mathbf{G}_{\mathrm{JML}} \triangleq \mathbf{I}_N - \mathbf{B}(\mathbf{B}^H\mathbf{B})^{-1}\mathbf{B}^H, \tag{11}$$

and select $\mathbf{p}$ to minimize the right-hand-side (RHS) of (10), i.e.,

$$\begin{aligned}\hat{\mathbf{p}}_{\mathrm{JML}} &= \arg\min_{\mathbf{p} \in \mathcal{P}} \left\|\mathbf{G}_{\mathrm{JML}}\mathbf{P}^H\mathbf{y}\right\|^2 \\ &= \arg\min_{\mathbf{p} \in \mathcal{P}} \|\mathbf{G}_{\mathrm{JML}}\mathbf{Y}\mathbf{p}^*\|^2.\end{aligned} \tag{12}$$

The scheme in (12) is referred to as the joint ML-CID (JML-CID), as this scheme finds the jointly most likely $\mathbf{h}$ and $\mathbf{p}$. The JML-CID requires the knowledge of the channel length L in (9).

It is interesting to note that the B-CID in (6) can also be derived from (8) by using the minimum mean square error (MMSE) estimate [6] of $\mathbf{h}$ given by

$$\hat{\mathbf{h}}_{\mathrm{MMSE}} = \mathbf{CB}^H(\mathbf{BCB}^H + \sigma^2\mathbf{I}_N)^{-1}\mathbf{P}^H\mathbf{y} \tag{13}$$

instead of the ML channel estimate in (9).

If the N subcarriers carrying the CID pattern are evenly spaced over all the N_s subcarriers of the preamble symbol, the columns of $\mathbf{B}$ become orthogonal to each other, and, thus, $\mathbf{B}^H\mathbf{B} = N\mathbf{I}_L$ [7]. Generally, the N preamble subcarriers are not evenly spaced because of the unusable subcarriers such as the guard bands and the DC subcarrier. However, these unusable subcarriers usually take only small portions of the OFDM symbol, and, thus, the columns of $\mathbf{B}$ are almost orthogonal to each other, i.e., $\mathbf{B}^H\mathbf{B} \simeq N\mathbf{I}_L$. Under the assumption that $\mathbf{B}^H\mathbf{B} = N\mathbf{I}_L$, (9) is simplified to

$$\hat{\mathbf{h}}_{\mathrm{SJML}} = \frac{1}{N}\mathbf{B}^H\mathbf{P}^H\mathbf{y}, \tag{14}$$

and by substituting $\mathbf{h}$ in (8) by $\hat{\mathbf{h}}_{\mathrm{SJML}}$ in (14), we obtain

$$\begin{aligned}\hat{\mathbf{p}}_{\mathrm{SJML}} &= \arg\min_{\mathbf{p} \in \mathcal{P}} \left\|\mathbf{y} - \mathbf{PB}\hat{\mathbf{h}}_{\mathrm{SJML}}\right\|^2 \\ &= \arg\min_{\mathbf{p} \in \mathcal{P}} \left\|\mathbf{P}^H\mathbf{y} - \frac{1}{N}\mathbf{BB}^H\mathbf{P}^H\mathbf{y}\right\|^2 \\ &= \arg\min_{\mathbf{p} \in \mathcal{P}} \left(\|\mathbf{y}\|^2 - \frac{1}{N}\left\|\mathbf{B}^H\mathbf{Y}\mathbf{p}^*\right\|^2\right) \\ &= \arg\max_{\mathbf{p} \in \mathcal{P}} \left\|\mathbf{B}^H\mathbf{Y}\mathbf{p}^*\right\|^2,\end{aligned} \tag{15}$$

which is referred to as the simplified JML CID (SJML-CID). On the RHS of (15), multiplication by $\mathbf{B}^H$ represents the inverse DFT operation. This can be implemented with an inverse fast Fourier transform (IFFT) algorithm as follows. The N-dimensional vector $\mathbf{Y}\mathbf{p}^*$ is fed to an N_s-point IFFT unit so that the n-th element of $\mathbf{Y}\mathbf{p}^*$ is applied to the input location i_n of the IFFT unit. The other $N_s - N$ inputs of the IFFT unit are set to zero. Then, the vector containing the first L elements of the IFFT output is identical to $\mathbf{B}^H\mathbf{Y}\mathbf{p}^*$. This process can significantly reduce the complexity of SJML-CID. As with the JML-CID, the SJML-CID needs to know L in (14).

C. Differential Decoding Approach to CID

In [5], a heuristic algorithm for CID was devised on the basis of the cross-correlation between the differentially decoded received signal and the differentially decoded candidate patterns in the frequency domain. This scheme, referred to as the differential CID (D-CID), is described as follows:

$$\hat{\mathbf{p}} = \arg\max_{\mathbf{p}\in\mathcal{P}} \left| \sum_{i=2}^{N} \left(p_{i-1}^* p_i\right)^* \left(y_{i-1}^* y_i\right) \right|, \tag{16}$$

where y_i denotes the i-th element of $\mathbf{y}$. In this subsection, we derive an alternative differential decoding-based CID scheme. The proposed scheme has a similar form to the D-CID, yet it can outperform the existing scheme while requiring less computation.

The i-th element of the received signal $\mathbf{y}$ in (1) can be written as

$$y_i = h_{f,i} p_i + w_i, i = 1, \ldots, N, \tag{17}$$

where w_i is the i-th element of the noise vector $\mathbf{w}$, and $h_{f,i}$ is the i-th element of the frequency-domain channel vector $\mathbf{h}_f \triangleq \mathbf{B}\mathbf{h}$. We now evaluate

$$\begin{aligned} y_{i-1}^* y_i =& h_{f,i-1}^* h_{f,i} p_{i-1}^* p_i + h_{f,i-1}^* p_{i-1}^* w_i \\ &+ h_{f,i} p_i w_{i-1}^* + w_{i-1}^* w_i \\ =& |h_{f,i-1}|^2 x_i + h_{f,i-1}^* w_i' + h_{f,i-1} w_{i-1}'^* + w_{i-1}^* w_i \\ &+ \delta_i h_{f,i-1}^* x_i + \delta_i w_{i-1}'^*, \;\; i = 2, \ldots, N, \end{aligned} \tag{18}$$

where $x_i \triangleq p_{i-1}^* p_i$, $w_i' \triangleq p_{i-1}^* w_i$, $w_{i-1}' \triangleq p_i^* w_{i-1}$, and $\delta_i \triangleq h_{f,i} - h_{f,i-1}$. Since w_i is circularly symmetric and $|p_i| = 1$ for all $i = 1, \ldots, N$, w_{i-1}' and w_i' have the same statistical properties with w_{i-1} and w_i, respectively. We assume that the channel is moderately frequency-selective, that is, the coherence bandwidth is larger than the adjacent preamble subcarrier spacings. Under this assumption, the channel gains of two adjacent preamble subcarriers are almost equal, i.e., $h_{f,i-1} \simeq h_{f,i}$ ($\delta_i \simeq 0$). Thus, we can neglect the last two terms, $\delta_i h_{f,i-1}^* x_i + \delta_i w_{i-1}'^*$ on the RHS of (18). Furthermore, the term $w_{i-1}^* w_i$, which is non-Gaussian, can also be neglected since this term is small relative to the dominant noise term $h_{f,i-1}^* w_i' + h_{f,i-1} w_{i-1}'^*$ at the signal-to-noise ratios (SNRs) of practical interest [8]. Thus, if we ignore the last three terms on the RHS of (18), then $y_{i-1}^* y_i \sim \mathcal{CN}\left(|h_{f,i-1}|^2 x_i, 2|h_{f,i-1}|^2\sigma^2\right)$ and the ML detection of x_i can be written as

$$\begin{aligned} \hat{x}_i &= \arg\max_{x_i\in\mathcal{X}_i} f\left(y_{i-1}^* y_i | x_i\right) \\ &= \arg\min_{x_i\in\mathcal{X}_i} \frac{\left|y_{i-1}^* y_i - |h_{f,i-1}|^2 x_i\right|^2}{2|h_{f,i-1}|^2\sigma^2}, \end{aligned} \tag{19}$$

where $\mathcal{X}_i \triangleq \left\{x_i = p_{i-1}^* p_i | \mathbf{p} \in \mathcal{P}\right\}$. Now, we combine the individual detection rule in (19) to form an identification rule for the sequence $\mathbf{x} \triangleq [x_2, x_3, \ldots, x_N]^T$ as

$$\begin{aligned} \hat{\mathbf{x}} &= \arg\max_{\mathbf{x}\in\mathcal{X}} \prod_{i=2}^{N} f\left(y_{i-1}^* y_i | x_i\right) \\ &= \arg\min_{\mathbf{x}\in\mathcal{X}} \sum_{i=2}^{N} \frac{\left|y_{i-1}^* y_i - |h_{f,i-1}|^2 x_i\right|^2}{2|h_{f,i-1}|^2\sigma^2} \\ &= \arg\max_{\mathbf{x}\in\mathcal{X}} \sum_{i=2}^{N} \Re\left\{x_i^* \left(y_{i-1}^* y_i\right)\right\}, \end{aligned} \tag{20}$$

where $\mathcal{X} \triangleq \left\{\mathbf{x} | x_i = p_{i-1}^* p_i \; (i = 2, \ldots, N), \mathbf{p} \in \mathcal{P}\right\}$ and $\Re\{\cdot\}$ denotes the real part. Note that (20) is not an ML detection rule since $\prod_{i=2}^{N} f\left(y_{i-1}^* y_i | x_i\right) \neq f\left(\left[y_1^* y_2, \ldots, y_{N-1}^* y_N\right]^T | \mathbf{x}\right)$. Though not optimal, (20) is a valid identification rule which corresponds to the *method of least-squares* [9]. Note that the channel gain $h_{f,i-1}$ is not required in (20). Since $x_i = p_{i-1}^* p_i$, $\mathbf{p}$ can be identified if (20) is rewritten as follows:

$$\hat{\mathbf{p}}_{\text{diff}} = \arg\max_{\mathbf{p}\in\mathcal{P}} \sum_{i=2}^{N} \Re\left\{\left(p_{i-1}^* p_i\right)^* \left(y_{i-1}^* y_i\right)\right\}. \tag{21}$$

This scheme is referred to as the modified D-CID (MD-CID). Comparing the D-CID and the MD-CID, we can observe that the absolute value in (16) is replaced by the real part in (21).

Actually, the D-CID (16) can be derived in the same way as the MD-CID (21). Assume that

$$h_{f,i-1} \simeq e^{-j\theta} h_{f,i}, \;\; i = 2, \ldots, N, \tag{22}$$

where $\theta \in [0, 2\pi)$ is an unknown constant. That is, an unknown constant phase rotation between the channel gains of two adjacent preamble subcarriers is assumed to exist, while their magnitudes are almost equal. Following the same procedure as (18)-(20), we obtain

$$\hat{\mathbf{x}} = \arg\max_{\mathbf{x}\in\mathcal{X}} \sum_{i=2}^{N} \Re\left\{e^{j\theta} x_i^* \left(y_{i-1}^* y_i\right)\right\}. \tag{23}$$

We first maximize the RHS of (23) with respect to θ for a given $\mathbf{x}$. Then we can show that the parameter $\hat{\theta}$ that maximizes the RHS of (23) satisfies

$$e^{j\hat{\theta}} = \frac{\left(\sum_{i=2}^{N} x_i^* \left(y_{i-1}^* y_i\right)\right)^*}{\left|\sum_{i=2}^{N} x_i^* \left(y_{i-1}^* y_i\right)\right|}. \tag{24}$$

Substituting $e^{j\theta}$ in (23) with $e^{j\hat{\theta}}$ in (24) and using $x_i = p_{i-1}^* p_i$, we obtain the D-CID (16). To summarize, the D-CID and the MD-CID can be derived in the same procedure, but based on different assumptions for the channel. The D-CID is based on the assumption (22), and the MD-CID on

TABLE I
COMPLEXITY COMPARISON (NUMBER OF FLOATING POINT OPERATIONS).

Complexity	B- and JML-CID	SJML-CID	MD-CID	D-CID [5]
Multiplication	$4N^2M$ (1.1×10^7)	$2MN_s \log N_s$ (6.6×10^5)	$2NM$ (1.8×10^4)	$4NM$ (3.7×10^4)
Addition	$4N^2M$ (1.1×10^7)	$3MN_s \log N_s$ (9.8×10^5)	$2NM$ (1.8×10^4)	$4NM$ (3.7×10^4)

(The values in parentheses represent the complexity evaluated when $N_s = 1024$, $N = 284$, $M = 32$, and $L = 100$.)

the assumption that $h_{f,i-1} \simeq h_{f,i}$. However, (22) is not a usual assumption for practical channels. Particularly for a moderately frequency-selective channel, the assumption that $h_{f,i-1} \simeq h_{f,i}$ is more reasonable than (22). Thus, at least in a moderately frequency-selective channel, the MD-CID would outperform the D-CID. In Section IV, we verify this through computer simulation.

D. Complexity comparison

Table I compares the computational complexities of the CID schemes considered in this section. The computational loads for implementing B-CID and JML-CID are identical, and they are proportional to N^2, where N is the span of CID pattern $\mathbf{p}$. The complexity of SJML-CID is proportional to $N_s \log_2 N_s$ due to the N_s-point IFFT operation, as explained at the end of Section III-B. Although $N_s > N$, in many practical cases of interest, the SJML-CID is simpler to implement than B-CID and JML-CID, as shown in the table. Differential decoding-based schemes are considerably simpler to implement than the optimal and suboptimal schemes. The proposed MD-CID is 50% less complex than the existing D-CID.

IV. Simulation Results

To evaluate the performances of the CID schemes, we consider the IEEE 802.16e system [1] with $N_s = 128$ and $N = 36$. In this system, p_i takes either 1 or -1, and there are 32 different preamble patterns in $\mathcal{P}$. We generate Rayleigh fading channels with three types of power delay profiles, namely, the Pedestrian B profile, the Vehicular A profile, and an exponential profile [10]. The maximum delay spreads of these profiles are $3.70\mu s$, $2.51\mu s$, and $10\mu s$, respectively, corresponding to 5, 4, and 13 channel taps. The decaying factor of the exponential channel, which is defined as the ratio between the powers of two consecutive channel taps, is 2 dB. The probability of an incorrect CID, which will be referred to as a false identification probability (FIP), is evaluated for all the CID schemes presented in the previous section. In addition, we also compare these schemes with the scheme in [4], which is represented as

$$\hat{\mathbf{p}} = \arg\max_{\mathbf{p}\in\mathcal{P}} \left| \sum_{i=2}^{N} \frac{p_{i-1}^{*} y_{i-1}}{p_i^{*} y_i} \right|. \quad (25)$$

The B-CID is implemented for a given channel covariance matrix, and two types of JML-CIDs are realized: one assumes the exact knowledge of the channel length L, and the other sets L at its maximum expected value, which is 13 in this simulation. The SJML-CID also sets L at 13. Of course, the B-CID and the JML-CID that assume the knowledge of channel covariance matrix and length are impractical; these schemes provide performance bounds of the practical methods.

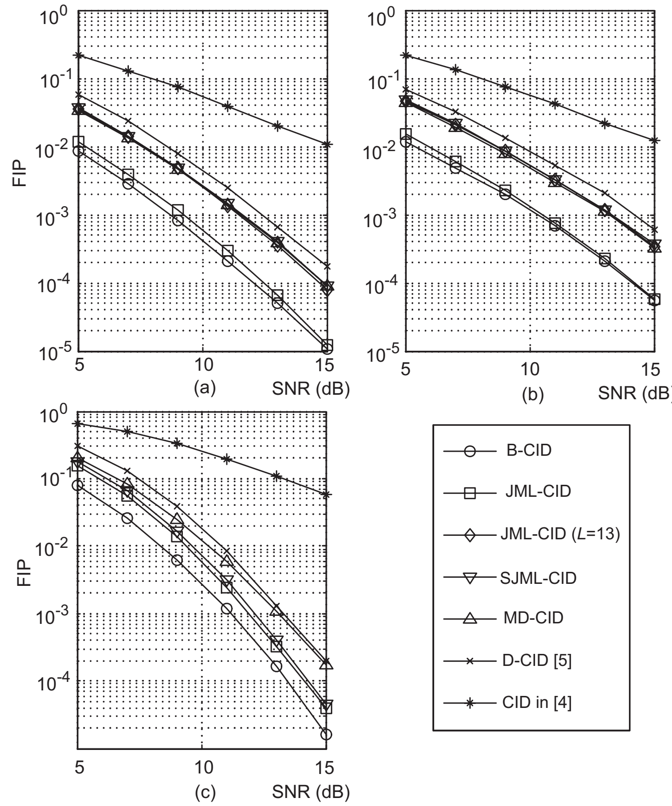

Fig. 2. False identification probability. (a) Ped. B channel. (b) Veh. A channel. (c) Exponentially decaying channel. In (c), JML-CID is identical to JML-CID with $L = 13$.

Fig. 2 shows the FIPs against the SNR. For all channels, the B-CID performs the best and the scheme in [4] performs the worst. In Figs. 2 (a) and (b), the performance of the JML-CID with a known value of L is comparable to that of the B-CID, while the JML-CID with $L = 13$ performs about 2.5 dB worse. This happens because in (10) the CID performance is affected by the accuracy of the channel estimate $\hat{\mathbf{h}}_{\mathrm{JML}}$; setting L at 13, which is considerably larger than the true value, increases the channel estimation error. For the Pedestrian B channel and the Vehicular A channel, the three schemes of JML-CID with $L = 13$, SJML-CID, and MD-CID exhibit almost identical performances and provide a gain of about 1dB over the D-CID in [5]. For the exponential channel, a similar observation can be made for the JML-CID and SJML-CID, but the performance of MD-CID is somewhat degraded and becomes closer to that of the D-CID. This happens because the exponential channel with decaying factor 2 dB is severely frequency selective, and the assumption that neighboring channel gains are identical is less valid.

These results confirm the usefulness of the proposed MD-CID for practical channels such as the Pedestrian B channel and the Vehicular A channel. The MD-CID is the simplest, yet it performs on a par with suboptimal schemes. When the channel is severely frequency-selective, the suboptimal SJML-CID would be recommended because it is simpler to implement than the JML-CID scheme with a maximum L value and exhibits a comparable performance.

V. Conclusion

For preamble-based cell identification in cellular OFDM systems, the optimal schemes were initially derived under the assumption that the fading channel is zero-mean Gaussian with a known covariance matrix (B-CID) and that the channel is deterministic with a known length (JML-CID). A suboptimal scheme (SJML-CID) and a differential decoding based scheme (MD-CID) are then derived. Analysis of their complexities and performance comparison by means of computer simulation indicate that the MD-CID is useful for most practical channels of interest, which are moderately frequency-selective. This is because the MD-CID can perform on a par with the SJML-CID without requiring any channel information and is simpler to implement than the other schemes. If the channel is severely frequency-selective, the SJML-CID is preferred because it is simpler than the JML-CID and exhibits a comparable performance.

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