

Channel Estimation and Equalization Methods for ZP-OFDM under Highly Mobile Environments

Saha TANANGSANAKOOL*, Pongsathorn REANGSUNTEA***, Kazuo MORI** and Pisit BOONSRIMUANG*

*Department of Telecommunication Engineering, Faculty of Engineering, KMITL, Bangkok, 10520 Thailand

**Department of Electrical and Electronic Engineering, Graduate School of Engineering, Mie University, Mie, 514-8507 Japan

***Department of Engineering, Focuz Manufacturing Co., Ltd., Navanakorn Industrial Estate, Patumthanee, 12120 Thailand

Saha_Tan@truecorp.co.th, pongsathorn.air@gmail.com, kmori@elec.mie-u.ac.jp, kbpisit@kmitl.ac.th

Abstract—Zero-Padded OFDM (ZP-OFDM) has been attracted a lot of attention as an efficient technique to improve the performance in severe multipath fading channels. Many works were devoted to the equalization method for ZP-OFDM in quasi-static channel model. However, an inter-carrier interference (ICI) which occurred under highly mobile environments would be destroyed orthogonality among subcarriers and causes the degradation of bit error rate (BER) performance due to the time domain channel impulse responses (CIRs) are assumed to be constant over one OFDM symbol. The time domain equalization (TDE) method in conjunction with a time domain CIR estimation method is proposed to solve this problem for the ZP-OFDM signal in the time domain. The advantage of the proposed method is to apply a training sequence (TS) instead of using ZP which enables for estimation of CIR. And also to reduce computational complexity, the proposed method is employed time domain equalization based on the minimum mean squared error (MMSE-TDE) instead of using the conventional frequency domain equalization based on MMSE (MMSE-FDE) method. This paper presents various simulation results under highly mobile environments to demonstrate the effectiveness of proposed TDE method as compared with conventional demodulation methods.

Keywords— ZP-OFDM, ICI, Training sequence (TS), MMSE-TDE, MMSE-FDE

I. INTRODUCTION

Zero-padded OFDM (ZP-OFDM) systems have recently been received a lot of attention because of its robustness to multipath fading channels of using a zero-padding instead of the more commonly used cyclic-prefix OFDM (CP-OFDM) [1]. ZP-OFDM offers potentially superior performances in the multipath channels only if more complex equalization, such as the least square (LS) or minimum mean square error (MMSE) equalizers is performed at the receiver instead of the more commonly used overlap and add (OLA) equalizer [2-3].

Since the ZP-OFDM is generally known as a robust technique for multi-path fading channels, many works were devoted to the equalization methods for ZP-OFDM in quasi-static channels [4-5] in which the time domain CIRs at every sampling time is assumed to be constant during one OFDM symbol and compensated its ICI precisely by using a simple one-tap OLA-FDE technique. However, the OFDM signal is sensitive to highly mobile environments due to the inter-carrier

interference (ICI) which the CIRs are assumed no more constant during one OFDM symbol. The highly mobile environments depend on the moving velocity of the terminal. The BER performance will be degraded when the moving velocity is increasing. We can remove the ICI for ZP-OFDM signal, the CIRs at every sampling time is required for demodulation of the transmitted data information by using any OLA, LS or MMSE equalizers. since many works [2-5] are assumed that the CIRs at every sampling time are perfectly known for each OFDM symbol at the receiver. Here it should be noted that the practical usage in the communication systems, the time domain CIR estimation technique is required and needed considering with efficient usage of frequency bandwidth especially under highly mobile environments.

Motivated by these facts both for ICI compensation and CIR estimation methods, this paper proposes a demodulation technique in conjunction with a time domain TS for ZP-OFDM. The proposed method enables the CIR estimation which is used in the compensation of both inter-symbol interference (ISI) and ICI. In contrast to existing MMSE-FDE method, our proposed method operates in the time domain and allows a low-complexity implementation. The proposed MMSE-TDE method in conjunction with the CIR estimation method of using a time domain TS can achieve BER performance similarly as compared with conventional demodulation for ZP-OFDM even under highly mobile environments.

The structure of this paper is consist of, Section II shows the proposed transmission system model, which use for the performance evaluation. Section III describes the proposed method of the CIR estimation and MMSE-TDE technique in highly mobile environments which suitable for our transmission systems. Section IV, we show various computer simulation results to verify the effectiveness of the proposed method and conventional methods, and Section V concludes the performances of the proposed method.

II. SYSTEM MODEL

Fig. 1 shows the baseband equivalent system structure model for the proposed OFDM method. At the transmitter side, the input data information is encoded by a forward error correction (FEC) code [6]. The encoded K data are modulated by the quadrature-amplitude modulation (QAM) technique. Then the

modulated signal $X_D(m, k)$ at the k -th subcarrier of m -th symbol is mapped into the certain frequency band within frequency bandwidth with N subcarriers continuously from the subcarrier number N_{Z1} to N_{Z2} ($N_{Z2} - N_{Z1} + 1 = K$) and the null subcarriers (N_Z : zero padding) are added at both ends of K data subcarriers to reject the aliasing occurring at the output of digital to analog (D/A) converter. The modulated signal $X(m, k_1)$ in the frequency domain over N subcarriers is converted to the time domain signal by N -point inverse Discrete Fourier transform (IDFT) which can be shown by the following equation,

$$x_T(m, n_1) = \frac{1}{\sqrt{N}} \sum_{k_1=N_{Z1}}^{N_{Z2}} X(m, k_1) \cdot e^{j \frac{2\pi n_1 k_1}{N}}, 0 \leq n_1 \leq N-1 \quad (1)$$

where $x_T(m, n_1)$ is the transmitted OFDM signal in the time domain. Fig. 2 shows the frame format for the proposed OFDM signal. In the frame format, the training sequence in the time domain [7] is added at the both ends of every data symbol which be used in the estimation of CIR at every symbol and also be used as the role of CP or ZP which is usually employed in the conventional CP or ZP-OFDM signal to avoid the ISI occurred in the multipath fading channels. In the proposed frame format in the time domain, the length of TS1 and TS2 also can be used to increase the system performance because of the observation period for data demodulation is longer than N samples. The transmitted time domain signal $x(m, n_2)$ at the n_2 -th sampling time of m -th symbol including both TS1 and TS2 can be expressed by,

$$x(m, n_2) = \begin{cases} d_1(m, n_2), & 0 \leq n_2 \leq N_{TS} - 1 \\ x_T(m, n_2 - N_{TS}), & N_{TS} \leq n_2 \leq N + N_{TS} - 1 \\ d_2(m, n_2 - N - N_{TS}), & N + N_{TS} \leq n_2 \leq N + 2N_{TS} - 1 \end{cases} \quad (2)$$

where $d_1(m, n_2)$ and $d_2(m, n_2)$ are the lengths of N_{TS} samples of the identical time domain TS1 and TS2, which data patterns are known at the receiver. For simplicity, we assume the data patterns both for TS1 and TS2 are the same as $d(m, n)$. The length of N_{TS} should be taken longer than the length of delay paths (L) which is the same condition of CP or ZP-OFDM signal. The timing synchronization for the received signal is assumed to be perfectly at the receiver, and the received OFDM signal $y(m, n_2)$ in the time domain at n_2 -th sampling time of m -th symbol after passing through the multipath fading channels can be given by,

$$y(m, n_2) = \sum_{l=0}^{L-1} h_l(m, n_2) \cdot x(m, n_2 - l) + z(m, n_2), \quad (3)$$

$$0 \leq n_2 \leq N + 2N_{TS} - 1$$

where $h_l(m, n_2)$ is the time domain CIR at every sampling time occurred in the actual time-varying fading channels, and $z(m, n_2)$ is stationary additive white Gaussian noise (AWGN) of variance σ^2 . Equation (3) and Fig. 2, the received time domain signal $y(m, n_2)$ can be spilled into two parts according to the observation period for the CIR estimation $y_{TS}(m, n_2)$ is considered from 0 to $N_{TS}-1$ period and for the data

demodulation $y_D(m, n_2)$ is considered from 0 to $N+N_{TS}-1$ period as shown in Fig.2.

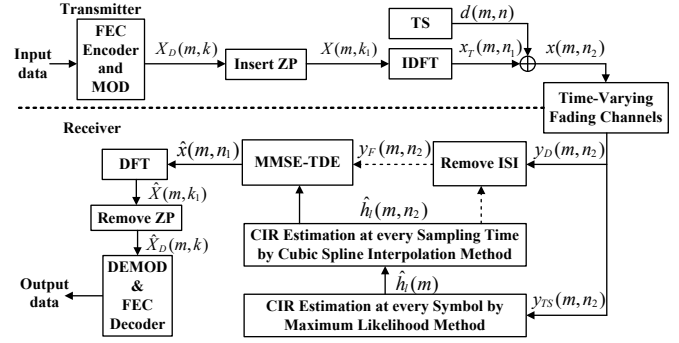


Figure 1. The system model for proposed MMSE-TDE method.

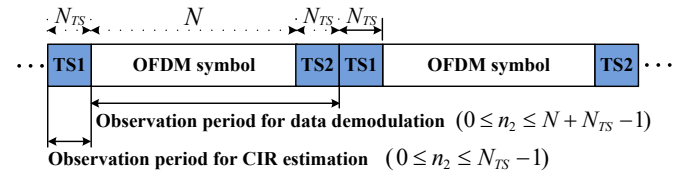


Figure 2. The frame format for the proposed OFDM signal in the time domain.

III. PROPOSED CIR ESTIMATION IN CONJUNCTION WITH MMSE-TDE METHOD

A. Proposed CIR Estimation Method

Equation (3) and the observation period for the CIR estimation, the CIR in highly time-varying fading channels is assumed to be constant during the short time period of TS1 and TS2. The time domain CIR $\hat{h}_l(m)$ is assumed by the unknown parameters, the expected received time domain TS signal $\hat{y}_{TS}(m, n_2)$ can be expressed by,

$$\hat{y}_{TS}(m, n_2) = \sum_{l=0}^{N_{TS}-1} \hat{h}_l(m) \cdot d(m, n_2 - l), 0 \leq n_2 \leq N_{TS} - 1 \quad (4)$$

The unknown parameters for $\hat{h}_l(m)$ at the m -th symbol can be estimated by solving the following maximum likelihood (ML) equation under the constraint condition with minimizing the difference between the actual received TS signal $y_{TS}(m, n_2)$ and the expected received TS signal $\hat{y}_{TS}(m, n_2)$ which is given by,

$$\hat{\gamma} = \arg \min_{\hat{h}_l(m)} \left[\sum_{n_2=0}^{N_{TS}-1} |y_{TS}(m, n_2) - \hat{y}_{TS}(m, n_2)|^2 \right] \quad (5)$$

In this paper, we apply a simple solution for solving ML equation in (5) by the following equations.

$$y_{TS}(m, n_2) = \hat{y}_{TS}(m, n_2) \text{ for } 0 \leq n_2 \leq N_{TS} - 1 \quad (6)$$

By substituting $\hat{y}_{TS}(m, n_2)$ given in (4) into (6). And (6) can be expressed by the following matrix operations.

$$\begin{bmatrix} \hat{h}_l(m) \end{bmatrix}_{N_{TS} \times 1} = \begin{bmatrix} d(m, n_2 - l) \end{bmatrix}_{N_{TS} \times N_{TS}}^{-1} \cdot \begin{bmatrix} y_{TS}(m, n_2) \end{bmatrix}_{N_{TS} \times 1} \quad (7)$$

where $[\cdot]^{-1}$ is inverse matrix operation. By using (7), the every

symbol $\hat{h}_l(m)$ can be estimated. The number of delay paths to be estimated in (7) is assumed by N_{TS} because the actual number of delay paths occurred in the real channel is unknown at the receiver. Here it should be noted that the data pattern of TS $d(m, n_2)$ is known at the receiver, the inverse matrix of $[d(m, n_2 - l)]^{-1}$ in (7) can be calculated in advance and the calculated inverse matrix can be used for every received data symbol which enables the considerable reduction of computation complexity. By using the estimated $\hat{h}_l(m)$ at every symbol, the CIR at every sampling time $\hat{h}_l(m, n_2)$ can be estimated by using the cubic spline interpolation method over several estimated CIRs of $\hat{h}_l(m)$. The estimation accuracy of time domain CIR at every sampling time will be evaluated in section IV.

B. Proposed MMSE-TDE Method

According to the observation period for data demodulation, the received time domain OFDM signal in highly mobile environments can be given by,

$$y_D(m, n_2) = \sum_{l=0}^{L-1} \hat{h}_l(m, n_2) \cdot x(m, n_2 - l) + z(m, n_2), \quad (8)$$

$$0 \leq n_2 \leq N + N_{TS} - 1$$

Equation (8), the received signal $y_D(m, n_2)$ in the time domain includes the ISI of TS which are included at the start and end of the data symbol from the TS1 and TS2, respectively. By using the estimated CIR $\hat{h}_l(m, n_2)$ at every sampling time and the known data pattern of $d(m, n_2)$ both for the time domain TS1 and TS2, the ISI of TS included at both ends of the received signal could be removed by the following equation.

$$y_F(m, n_2) = \begin{cases} y_D(m, n_2) - \sum_{l=n_2+1}^{N_{TS}-1} \hat{h}_l(m, n_2) \cdot d(m, n_2 + N_{TS} - l), & (0 \leq n_2 \leq N_{TS} - 2) \\ y_D(m, n_2), & (N_{TS} - 1 \leq n_2 \leq N - 1) \\ y_D(m, n_2) - \sum_{l=0}^{n_2-N} \hat{h}_l(m, n_2) \cdot d(m, n_2 - N - l), & (N \leq n_2 \leq N + N_{TS} - 2) \end{cases} \quad (9)$$

Since the data demodulation is operated at every symbol, we omitted the parameter m -th symbol and then the received time domain can be given in vector form as

$$y_F = hx + z \quad (10)$$

Here $y_F = [y_F(0), \dots, y_F(n_2), \dots, y_F(N + N_{TS} - 2)]^T$ has received time domain without ISI vector, where $[\cdot]^T$ represents the transposition. $x = [x(0), \dots, x(n_2), \dots, x(N - 1)]^T$ is transmitted time domain signal vector which correspond to $x_T(m, n_1)$ in (1). Furthermore, h is the time domain CIR matrix after removing ISI with the size of $(N + N_{TS} - 1) \times (N)$ which equivalent to CIR matrix of ZP-OFDM signal [3-5]. Equation (10), the conventional OLA, LS and MMSE equalizers both for operation in the frequency domain or time domain can be applied for demodulation of the transmitted data information.

To archive a potentially superior performance with low computational complexity, this paper considers a time domain MMSE ICI equalization amounts for finding the linear equalizer matrix W such that Wy_F is closest to x in the MMSE sense can be expressed by,

$$W_{MMSE} = \arg \min_W \{E|Wy_F - x|^2\} \quad (11)$$

The solution of this minimization problem is straightforwardly obtained as

$$W_{MMSE} = (h^H h + \sigma^2 I)^{-1} h^H \quad (12)$$

where $(\cdot)^H$ is the Hermitian transpose operation, I is an identity matrix, and σ^2 is a variance of AWGN. By using the estimated CIR $\hat{h}_l(m, n_2)$ at every sampling time, the MMSE equalizer output equals

$$\hat{x}_{MMSE} = W_{MMSE} y_F = (h^H h + \sigma^2 I)^{-1} h^H y_F \quad (13)$$

where $\hat{x}_{MMSE}$ is expected transmitted time domain signal. Equation (13), the expected transmitted frequency domain signal $\hat{X}_{MMSE}$ can be obtained by converting the time domain data signal to frequency domain signal with N -point DFT. Equation (13), it also can be seen that the equalization in the time domain (TDE) shows the simplicity structure especially under highly mobile environment because of it easy to construct the CIR matrix. Unlike the equalization in the frequency domain (FDE), the MMSE equalizer output $\hat{x}_{MMSE}$ is mapped to the frequency domain as $F\hat{x}_{MMSE}$, where F is DFT matrix and, it is unitary. $F\hat{x}_{MMSE}$ equals to the solution of MMSE-FDE equalizer and by using $I = F^H F = F F^H$ and $F^{-1} = F^H$, we can obtain as

$$\begin{aligned} F\hat{x}_{MMSE} &= F(h^H h + \sigma^2 I)^{-1} h^H F^H y_F \\ \hat{X}_{MMSE} &= (F(h^H F^H F h + \sigma^2 I)F^H)^{-1} F h^H F^H y_F \\ \hat{X}_{MMSE} &= (H^H H + \sigma^2 I)^{-1} H^H Y_F \end{aligned} \quad (14)$$

where Y_F is received frequency domain signal vector. H is CFR matrix in the frequency domain which obtained from time domain CIR by DFT and IDFT matrixes. Equation (13) and (14), it can be observed that time domain and frequency domain MMSE equalizations are mathematically equivalent and the performances of both methods also have the same results. However, a time domain MMSE equalizer generally has lower computational complexity than frequency domain MMSE equalizer because of the construction for frequency domain matrix operations is lower as shown in (13) and (14).

IV. PERFORMANCE EVALUATIONS

This section presents various simulations results to verify the effectiveness of the proposed MMSE-TDE method as compared with the conventional MMSE-FDE method for ZP-OFDM. For both cases, we consider a rate 1/2 convolutional code is adapted. The data information is encoded by a generator in octal form as (133, 171) while decoding is done by a hard decision Viterbi decoder. 16QAM coded ZP-OFDM with the

length of symbol $M=64$, the subcarrier length $K=96$ within frequency bandwidth 1 MHz operated on radio frequency 5.9 GHz. The number of DFT/IDFT point $N=128$, the length of zero-padded samples and time domain TS are the same as $N_{TS}=16$. The mobile channel in this simulation is modeled by the Rayleigh multipath fading channels having uniform power delay profile with the number of delay paths $L=14$. The normalized Doppler frequency R_D ($R_D=f_{dmax}/\Delta f$ (%)) is applied as the measure of mobile channel conditions where Δf is the subcarrier spacing, and f_{dmax} is the maximum Doppler frequency. As for the fair comparison between proposed OFDM and conventional ZP-OFDM, the CIR estimation accuracy of conventional ZP-OFDM is assumed as the same as the proposed method. And also the all conventional ZP-OFDM systems [2-5] are operated the equalization in the frequency domain. Here, it should be noted that only 1-Tap OLA of the proposed OFDM method is operated the equalization in the frequency domain.

Fig. 3 shows the time domain CIR estimation accuracy at every sampling time which is evaluated by the normalized mean square error (NMSE) for the proposed CIR estimation method of using the time domain TS when changing the operation C/N and normalized Doppler frequency R_D . In the figure (3) can be observed that the estimated CIR for the proposed method can keep the higher accuracy even when R_D is up to 20% (vehicle speed $\approx 381\text{km/hr.}$) at lower operation C/N. From these simulation results, it can be confirmed that the proposed CIR estimation can be used to mitigate the ISI and ICI precisely even under highly mobile environments.

Fig. 4 shows the BER performances for the proposed 1-tap OLA MMSE-FDE, full-tap OLA MMSE-TDE, and full-tap MMSE-TDE methods as compared with conventional ZP-OFDM with MMSE-FDE methods [2-5] when changing normalized Doppler frequency R_D . And Fig. 5 shows the BER performances for the proposed 1-tap OLA MMSE-FDE, full-tap OLA MMSE-TDE and full-tap MMSE-TDE methods as compared with conventional ZP-OFDM with MMSE-FDE methods when changing operation C/N. Here it should be noted that the performances of estimated CIR for conventional ZP-OFDM method and the proposed CIR estimation method are assumed to be as the same accuracy. From both Fig. 4 and 5, it can be observed that the BER performance for the proposed MMSE-TDE method achieves BER performance similar to conventional ZP-OFDM with MMSE-FDE methods even under highly mobile communications. However, the proposed MMSE-TDE method enables for estimation of time domain CIR in practical communication systems. Table I shows the comparison of computation complexity between conventional MMSE-FDE and proposed MMSE-TDE methods. From the table, the proposed method can achieve lower computation complexity than conventional ZP-OFDM with MMSE-FDE methods as mentioned in section III.

V. CONCLUSIONS

This paper proposed low-complexity MMSE-TDE method in conjunction with time domain CIR estimation method for the ZP-OFDM signal under highly mobile environments. The proposed MMSE-TDE method employs the time domain TS signal which is used both for estimation of CIR and also used as the role of CP. From the computer simulation results, it can be concluded that the proposed MMSE-TDE method enables for estimation of the CIR and also can achieve lower computational complexity without degradation of BER performance as compared with the conventional MMSE-FDE for ZP-OFDM.

TABLE I COMPARISON OF COMPUTATION COMPLEXITY

Methods	Computation Complexity
Conv. MMSE-FDE	$4N^3 + 3N^2 + (N + N_{TS})^2N + (N + N_{TS})^2$
Proposed MMSE-TDE	$N^3 + 2N^2 + (N + N_{TS})^2N + (N + N_{TS})^2$

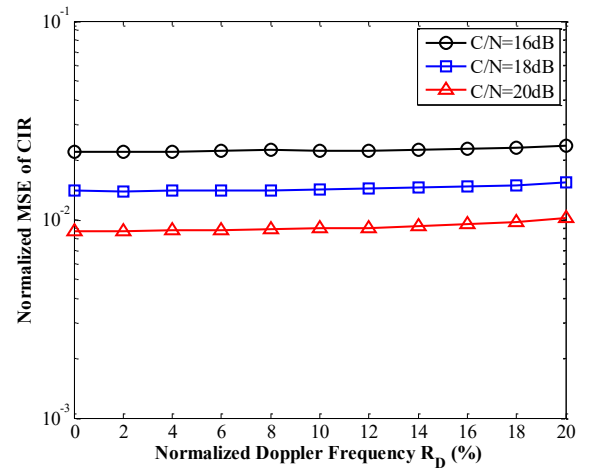


Figure 3. CIR estimation accuracy (NMSE) for the proposed method.

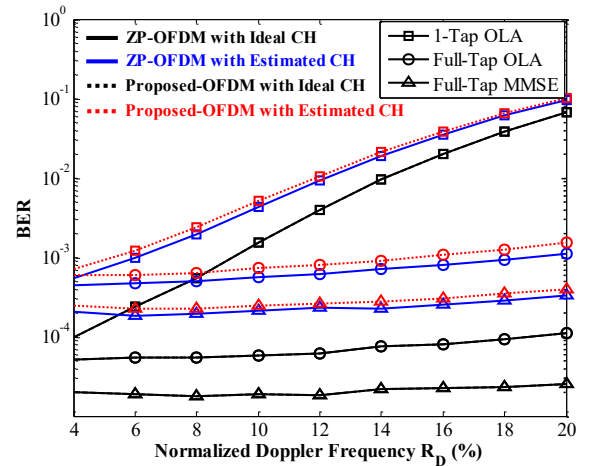


Figure 4. BER performance for proposed MMSE-TDE method when changing R_D at C/N=20dB.

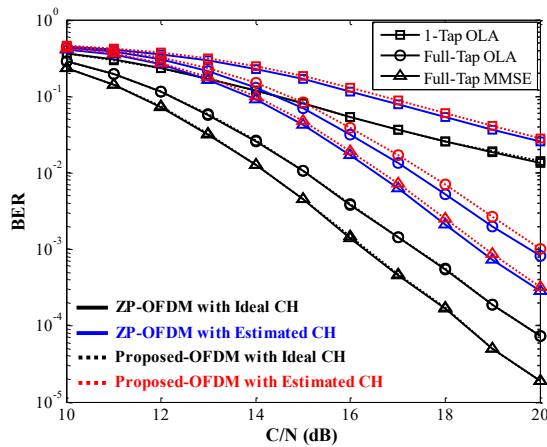


Figure 5. BER performance for proposed MMSE-TDE method when changing C/N at $R_b=15\%$ (vehicle speed $\approx 286\text{km/hr.}$).

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Saha Tanangsanakool received the B.Eng. degree in Electronic Engineering from South-East Asia University and M.Eng. degree in Telecommunication Engineering from King Mongkut's Institute of Technology Ladkrabang (KMUTL), Thailand in 1992, 2001. He is currently a doctoral student at King Mongkut's Institute of Technology Ladkrabang (KMUTL), Thailand. He joined True Corporation PCL in 2005 and engaged in Network Engineering & Quality Assurance. His

research interests include mobile communication system and broadband system.



Pongsathorn Reangsuntea received the B.Eng and M.Eng degrees in Telecommunications Engineering from King Mongkut's Institute of Technology Ladkrabang (KMUTL), Thailand in 2010, and 2012, respectively. He received the Ph.D. degree in systems engineering from Mie University, Japan in 2017. He received the student award of IEEE VTS Japan 2015 Young Researcher's Encouragement Award in VTC2015-Spring, Glasgow, UK, in May 2015. Since 2017, has been a Senior Test Engineer at Focuz Manufacturing Co., Ltd., Thailand. His research

interests include high-speed mobile communications and wireless LAN systems.



Kazuo MORI received the B.Eng. degree in Computer Engineering from Nagoya Institute of Technology, Japan, in 1986 and received the Ph.D. degree in Information Electronics Engineering from Nagoya University, Japan in 2000. In 1986, He joined the Hyper-media Research Center, SANYO Electric Co., Ltd. and was engaged in research and development on Telecommunications Systems. From 1995 to 2000, he was a research engineer at YRP Mobile Telecommunications Key Technology Research Laboratories Co., Ltd., where he was engaged in research on mobile communication systems. Since 2000, he has been an Associate Professor of the Department of Electrical and Electronic Engineering at Mie University, Japan. From 2005 to 2006, he was a Visiting Research Fellow at King's College London, UK. His research interests include mobile communication systems, CDMA schemes, radio packet communications, and tele traffic evaluation. Dr. Mori received the Excellent Paper Award from IEICE, Japan in 2002.



Pisit BOONSIRIMUANG received the B.Eng, M.Eng, and Doctor degrees in Telecommunications Engineering, in 1997, and 2000, and 2007, respectively. He is currently an associate professor at King Mongkut's Institute of Technology Ladkrabang (KMUTL), Thailand. His research interests include transmission techniques for future multimedia wireless LAN systems and next generation of mobile communication systems. He received the student Award of Young Researches Encouragement Award from IEICE Tokai branch and Doctor Thesis Award for information technology from Nation Research council of Thailand (NRCT) in 2005 and 2008, respectively. Since 2012, he has been an Associate Professor of KMUTL. His current research interests include mobile communication and broadband.