Block-Type Pilot Arrangement with Alternating Polarity for ICI Mitigation in Mobile OFDM Systems

Titiek Suryani, and Gamantyo Hendrantoro

***Abstract*—Improvement on Inter Carrier Interference (ICI) mitigation techniques for OFDM caused by Doppler effects through minimizing channel estimation error and decreasing channel time varying rate is investigated. The performance of pilot-aided channel estimation techniques depends on pilot placement and arrangement and also on the channel time varying rate. The block-type and comb-type pilot arrangements are studied through different numbers of guard bands, with or without the involvement of the Doppler shift compensation. The estimation of channel at mid-point of each OFDM symbol is derived from pilot frequencies based on the least square algorithm while the channel interpolation is done using piece-wise linear approximation. For ICI mitigation technique we implement frequency domain zero forcing equalizer. We compare the performance of schemes with different pilot arrangements and Doppler shift compensations by measuring bit error rate with QPSK as sub-channel modulation scheme and with mobile-to-fixed of single ring scattering as channel model. The results are in favour of block-type pilot arrangements with alternating polarity and Doppler compensation of** 0.55 **times the maximum Doppler shift.**

***Index Terms*—pilot tones, OFDM, Doppler, Inter-carrier Interference**

I. INTRODUCTION

MULTIPLEXING techniques that use a large number of orthogonal carriers, i.e., OFDM (Orthogonal Frequency Division Multiplexing), are attractive for high speed data transmission. The orthogonality between sub-carriers allows signal spectra of different sub-carriers to be separated at the receiver without any interference or power loss due to leakage to other sub-carriers. However, time varying channels with large Doppler shift could destroy the orthogonality, resulting in carrier frequency offset and ICI (inter-carrier interference) which degrade the OFDM system bit error rate (BER) performance [1]. The ICI is affected by the maximum Doppler shift, f_d, whose value, normalized to the adjacent sub-carrier spacing, is denoted by $f_{d,norm} = f_d/\Delta f$, where Δf is adjacent sub-carrier spacing. Accordingly, the normalized Doppler shift increases when the mobile terminal moves faster or the frequency spacing between adjacent sub-carriers is decreased. Greater normalized Doppler shift escalates carrier frequency offset and causes spectrum dispersion, resulting in larger ICI [2], [3].

Various methods have been proposed to overcome ICI. Generally, the mitigation techniques can be divided into two types: one that attempts to minimize ICI preventively [4], [5] and the other that attempts to remove it once it occurs, e.g., using frequency domain equalizer (FDE) [3], [6] - [9]. Usually, the former uses coding to minimize ICI by controlling sub-carrier spectrum, and hence, consumes more bandwidth, whereas the latter requires more complex signal processing to remove ICI. In this paper, we focus on the FDE-based ICI mitigation techniques to avoid degradation in bandwidth efficiency.

The capability of FDE to reduce ICI depends on the accuracy of the channel estimation. In time-varying channels, the channel may significantly change even within one OFDM symbol [10]. Hence, a simple and low complexity channel estimation such as a piece-wise linear approximation proposed in [9] for ICI mitigation using FDE is needed. To reduce the complexity of inverse channel matrix computation for finding FDE filter coefficients, the special structure of banded matrix is exploited [11], [12]. For channel estimation using pilot aided technique, conventional methods generally consist of estimating the channel frequency response at pilot frequencies and then interpolating [10]. Most of the channel estimation approaches may be viewed as DFT-based approaches, where LS (least square) channel frequency response estimates are fed into an inverse fast Fourier transform (IFFT) block to obtain the time domain channel impulse response estimate, and then appropriately processed and transformed back to frequency domain by fast Fourier transform (FFT) [13], [14]. In such approaches, the channel estimation using linear approximation studied by Mostofi and Cox [9] works sufficiently well but has limitation when working in channels with large normalized Doppler shift ($f_{d,norm} \geq 0.2$). In [9], channel estimation consists of two step: the first step estimates mid-point channel response at each OFDM symbol using pilots, whereas the second estimates channel between two adjacent mid-points using piece-wise linear interpolation. To reduce the channel estimation error, as in our previous work [15], we propose Doppler shift compensation to reduce channel time variation rate such that the channel remain linear within one OFDM

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Titiek Suryani is with the Department of Electrical Engineering, Institut Teknogi Sepuluh Nopember, Surabaya, 60111, Indonesia e-mail: titiks@ee.its.ac.id.

G. Hendrantoro is with Institut Teknogi Sepuluh Nopember, Surabaya, 60111, Indonesia e-mail: gamantyo@ee.its.ac.id.

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symbol, hence the piece-wise linear interpolation error can be minimized. However, this technique works more effectively if the mid-point channel estimation error can also be reduced. In [15], pilots are inserted in each OFDM symbol and are distributed among sub-carriers in an equally spaced manner (comb-type pilot arrangement). This pilot arrangement makes pilots suffer from large ICI because the power leakage from side-lobe spectrum of symbol at adjacent sub-carrier is high. To overcome this problem, guard bands can be inserted to prevent the impact of ICI on the pilots, but this technique will consume more bandwidth.

In this paper we propose a pilot block-type arrangement with alternating polarity scheme to minimize ICI distortion between pilots or of the data symbols. With alternating polarity, the side-lobe spectra of pilots cancel each other and cause little impact among the pilots themselves; only pilots that are located near the data symbol might suffer from large ICI. Thus, the block-type pilot arrangement with alternating polarity is more robust to ICI distortion than the comb-type. If the ICI distortion to the pilot symbol can be reduced then the mid-point estimation error can be minimized. Moreover, by using Doppler compensation, the overall channel estimation error can be reduced significantly and the ICI mitigation performance using FDE can be improved to operate in large normalized Doppler shift ($f_{d,norm} \geq 0.2$).

To summarize, our main contribution is the adoption of block-type pilot arrangement with alternating polarity, which is combined with the previously proposed Doppler compensation technique, to reduce the effects of ICI on the pilots. As a result, the channel estimation accuracy can be increased and the FDE filter coefficients can be determined more accurately to cancel ICI. This is discussed further in the following sections. Section II presents the channel and system description together with the proposed techniques. Section III reports simulation results where the ICI mitigation performance is investigated for different types of pilot arrangement with and without Doppler compensation at large normalized Doppler shift, $f_{d,norm}$. Conclusions are given in Section IV.

II. SISTEM DESCRIPTION

A. OFDM System with ICI Mitigation

The block diagram of an OFDM system based on pilot-aided channel estimation with Doppler compensation and frequency domain equalizer is given in Fig. 1. The binary information is first mapped into QPSK (quadrature phase shift keying) symbols. The S/P (serial-to-parallel) block will group the symbols and insert pilots with or without guard band into all sub-carriers and the IFFT block will convert them into time-domain OFDM symbols according to the following equation:

$$x_n = \frac{1}{N_{\mathrm{S}}} \sum_{k=0}^{N_{\mathrm{S}}-1} X_k e^{j\frac{2\pi kn}{N_{\mathrm{S}}}}; \quad n = 0, 1, \cdots, N_{\mathrm{S}} - 1 \quad (1)$$

where X_k denotes the k-th frequency-domain sample in the group after pilot insertion and N_{S} is the number of IDFT points. Guard interval or cyclic-prefix is added to prevent the OFDM symbols from inter-symbol-interference due to

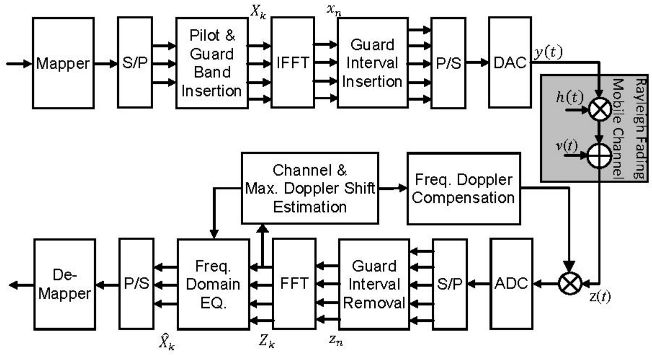

Fig. 1. Baseband Equivalent OFDM System with ICI Mitigation

multipath fading. The guard interval with length G is given as:

$$x_g = x_{N_{\mathrm{S}}-G+g}; \quad 0 \leq g \leq G - 1 \quad (2)$$

Next, output of serial-to-parallel process on input block x_n with guard interval x_g in Fig. 1, i.e., y_n is converted by the DAC (digital-to-analog converter) block into continuous-time signal, $y(t)$. The transmitted signal subsequently passes through the time varying Rayleigh faded channel $h(t)$ and additive white Gaussian noise $v(t)$. The received signal is given by:

$$z(t) = h(t)y(t) + v(t) \quad (3)$$

The mobile-to-fixed (MF) channel is modeled with single-ring scatterers at the mobile transmitter side. It is derived from the mobile-to-mobile (MM) channel model with double-ring scatterers [16]- [18] by assuming that one of the terminals (i.e., the receiver) is a fixed base station on a high tower such that the absence of surrounding scattering objects can be assumed. Hence, the MF channel response can be expressed as,

$$h(t) = \sqrt{\frac{2}{P}} \sum_{p=1}^{P} e^{j(2\pi f_d \cos(\alpha_p)t + \phi_p)} \quad (4)$$

$$\alpha_p = \frac{2p\pi - \pi + \theta_p}{4P}$$

where $f_d = v_d/\lambda_d$, denotes the maximum Doppler shift due to transmitter motion with v_d denoting velocity and λ_d carrier wavelength. Index p refers to the propagation path from the transmitter to the p-th of the scatterers located on the ring surrounding the transmitter (see, e.g., Clarke's model [19]), P is the number of scatterers located on the transmitter ring, $\theta_p \sim U(-\pi, \pi)$ is the random angle of departure (AOD) of the p-th path measured with respect to the transmitter velocity vector, ϕ_p is the phase shift associated with the reflection coefficient of scatterers. The dynamic behavior of the channels can be expressed by the time auto-correlation function, $R(m) = \sigma^2 J_0(2\pi f_d mT)$, where σ^2 denotes the power of the channel, T OFDM symbol period after adding the guard interval and $J_0(\cdot)$ the zeroth-order Bessel function of the first kind. Eq. (4) is assumed that the channel power is normalized to unity.

At the receiver, after conversion to discrete-time domain through ADC (analog-to-digital converter) and low pass filter, the guard interval is removed and the remaining sequence is denoted by z_n. Output of FFT block is denoted by Z_k which, assuming no ISI (inter symbol interference), can be represented as:

$$Z_k = H_{k,0}X_k + \underbrace{\sum_{d=1}^{N_\mathrm{S}-1} H_{k,d}X_{((k-d))_{N_\mathrm{S}}}}_{\mathrm{ICI}} + V_k \tag{5}$$

$$0 \le k \le N_\mathrm{S} - 1$$

where $H_k = FFT\{h_n\}$ with h_n being the sampled version of $h(t)$ at sampling rate T_S, $T_\mathrm{S} = T/(N_\mathrm{S} + G)$, G is guard interval length and N_S denotes the number of sub-carriers, $H_{k,0}$ denotes the channel response associated with data at the k-th sub-carrier, $H_{k,d}$ for $d = 1, \cdots, N_\mathrm{S} - 1$ denotes the channel frequency response associated with data at $((k-d))_{N_\mathrm{S}}$-th sub-carrier that has caused ICI on the data at the k-th sub-carrier, $V_k = FFT\{v_n\}$ and $((\cdot))_{N_\mathrm{S}}$ represents modulo N_S operation. Following the FFT block, the pilot signals are extracted and the estimated channel for the data sub-channels is obtained in the channel estimation block. Then the transmitted QPSK symbols are estimated by:

$$\hat{X}_k = \frac{Z_k}{\hat{H}_k}; \quad k = 0, 1, \cdots, N_\mathrm{S} - 1 \tag{6}$$

where $Z_k = FFT\{z_n\}$. The binary information is obtained back in the signal demapper block. The estimated channel $\hat{H}_k = FFT\{\hat{h}_n\}$ is updated for each symbol. An estimate of $H_{k,0}$ can then be acquired at the pilot tone frequencies based on Least Square estimation as given by:

$$\hat{H}_{l_i,0} = \frac{Z_{l_i}}{P_{l_i}} = H_{l_i,0} + \frac{I_{l_i} + V_{l_i}}{P_{l_i}}; \quad 0 \le i \le N_\mathrm{P} - 1 \tag{7}$$

where P_{l_i} is the complex baseband representation of the i-th pilot that is inserted at the l_i-th sub-channels, N_P is the number of pilots and I_{l_i} denotes ICI at the pilot symbol [marked in (9)]. Through an N_P-point IFFT of $H_{l_i,0}$, the estimate of the average channel over the time duration of $0 \le t \le NT_\mathrm{S}$, $\hat{h}^{ave}$, would be $\hat{h}^{ave} = \hat{h}^{ave}_m|_{m=0} = \hat{h}^{ave}_0$. Where $\hat{h}^{ave}_m$ is defined as,

$$\hat{h}^{ave}_m = \frac{1}{N_\mathrm{P}} \sum_{i=0}^{N_\mathrm{P}-1} \hat{H}_{l_i,0} e^{j\frac{2\pi i m}{N_\mathrm{P}}}; \quad 0 \le m \le N_\mathrm{P} - 1 \tag{8}$$

The estimated channel at mid-point of each symbol is denoted by $\hat{h}^{mid} = \hat{h}^{ave}$. Using $\hat{h}^{mid}$, the channel is estimated through piece-wise linear interpolation as [9],

$$\hat{h}^k = \hat{h}^{mid} + \left(k + 1 - \frac{N_\mathrm{S}}{2}\right) \times \hat{\mu} \times T_\mathrm{S}; \quad 0 \le k \le N_\mathrm{S} - 1 \tag{9}$$

where $\hat{\mu}$ is the line slope from $\hat{h}^{mid}$ to the $\hat{h}^{mid,next}$, whereas $\hat{h}^{mid,next}$ is the estimated channel at mid-point of next symbol. Hence $\hat{\mu}$ is obtained as follows:

$$\hat{\mu} = \frac{\hat{h}^{mid,next} - \hat{h}^{mid}}{T_\mathrm{S}} \tag{10}$$

An accurate estimation of $\hat{h}^{mid}$ will significantly reduce the channel estimation errors. This can be achieved by pilot arrangement techniques to avoid ICI distortion on the pilot itself.

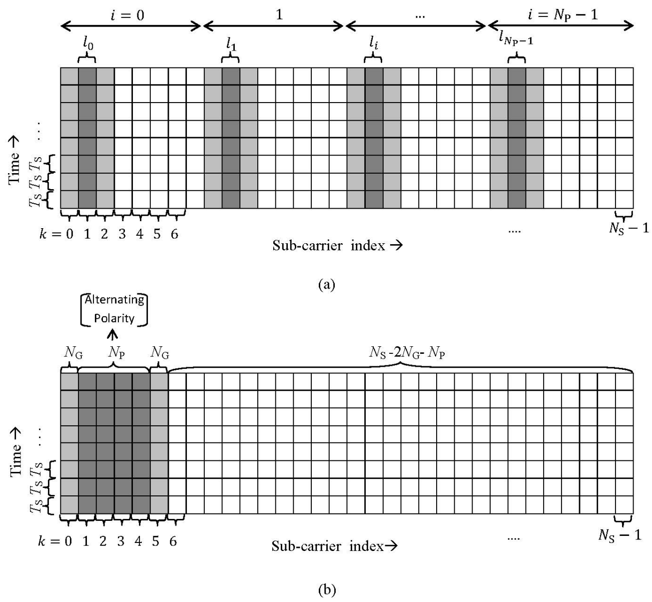

Fig. 2. Pilot Arrangement (a) Comb-type and (b) Block-type

The channel estimation result, $\hat{h}^{(k)}$ is formed into a $N_\mathrm{S} \times N_\mathrm{S}$ matrix form as,

$$\hat{\mathbf{h}} = \mathrm{diag}\left\{\left[\hat{h}^{(0)} \quad \hat{h}^{(1)} \quad \cdots \quad \hat{h}^{(N_\mathrm{S}-1)}\right]\right\} \tag{11}$$

The received signal vector at the output of the FFT block is defined as,

$$\mathbf{Z} = \mathbf{F}\mathbf{z} = \mathbf{F}\mathbf{h}\mathbf{F}^H\mathbf{X} + \mathbf{F}\mathbf{v} = \mathbf{H}\mathbf{X} + \mathbf{F}\mathbf{v} \tag{12}$$

where $\mathbf{H} = \mathbf{F}\mathbf{h}\mathbf{F}^H$ is the frequency-domain channel matrix, $\mathbf{F}$ and $\mathbf{F}^H$ are the N_S-point FFT and IFFT matrices, respectively, $\mathbf{X} = [X_0 \quad X_1 \quad \cdots \quad X_{N_\mathrm{S}-1}]^T$, $\mathbf{v} = [v_0 \quad v_1 \quad \cdots \quad v_{N_\mathrm{S}-1}]^T$, $[\cdot]^T$ denotes the transpose, $[\cdot]^H$ denotes the complex conjugate transpose. Then using frequency domain equalizer (FDE) with filter coefficients that are obtained from the inverse of $\hat{\mathbf{H}}$, where $\hat{\mathbf{H}} = \mathbf{F}\hat{\mathbf{h}}\mathbf{F}^H$, the data vector carried in each OFDM symbol is estimated using the following equation,

$$\hat{\mathbf{X}} = \hat{\mathbf{H}}^{-1}\mathbf{Z} = \hat{\mathbf{H}}^{-1}(\mathbf{H}\mathbf{X} + \mathbf{V}) \tag{13}$$

where $\mathbf{Z} = [Z_0 \quad Z_1 \quad \cdots \quad Z_{N_\mathrm{S}-1}]^T$ and $\mathbf{V} = \mathbf{F}\mathbf{v} = [V_0 \quad V_1 \quad \cdots \quad V_{N_\mathrm{S}-1}]^T$ are vectors of the received signal and noise at the output of the FFT block, respectively and $\hat{\mathbf{H}}^{-1}$ is the inverse of $\hat{\mathbf{H}}$.

B. Pilot Arrangement

We investigate pilot-aided channel estimation with two types of pilots and guard bands arrangements, i.e., comb-type and block-type. Both types are illustrated in Fig. 2. In the comb-type pilot arrangement, the N_P pilot signals are uniformly inserted into X_k according to the following equation:

$$X_k = \begin{cases} P_{l_i}, & k = l_i = 1 + i \times N_\mathrm{S}/N_\mathrm{P} \\ 0, & k = l_i \pm 1 \\ D_q, & k = 3 + 3 \times \lfloor \frac{q}{N_\mathrm{D}} \rfloor + ((q))_{N_\mathrm{D}} \end{cases} \tag{14}$$

$$P_{l_i} = (1+j)/\sqrt{2} \tag{15}$$

$$\begin{gathered} N_S = 2^{n_S}; N_P = 2^{n_P}; n_S - n_P = 3 \\ n_S = 5,6,7,\cdots; j=\sqrt{-1}; N_D = \frac{N_S}{N_P} - 3 \\ 0 \le i \le N_P - 1; 0 \le q \le N_S - 3 \times N_P - 1 \end{gathered}$$

where D_q is the q-th QPSK symbol conveying the information carried in each OFDM symbol, and $\lfloor q \rfloor$ is the largest integer not larger than q . Here, the guard band is required to prevent pilot symbols from experiencing large ICI from adjacent symbols such that the error estimation on $\hat{h}^{mid}$ can be minimized.

As an alternative, we propose the block-type pilot arrangement that is aimed to reduce the guard band requirement. Here, the pilot symbol is arranged in an alternating fashion such that the ICI between pilots can be suppressed as well as their impact on data symbols. The guard band and pilot arrangement for the block-type pilot is given as:

$$X_k = \begin{cases} 0, & k = 0,\cdots,(N_G-1) \quad \text{and} \\ & k = (N_G+N_P),\cdots,(2N_G+N_P-1) \\ P_{l_i}, & k = l_i = N_G + i \\ D_q, & k = 2N_G + N_P + q \end{cases} \tag{16}$$

where N_G is the number of guard band at each pilot side, $q = 0,1,\cdots,N_S - 2N_G - N_P - 1$. For analysis purpose, we define block-type pilots with polarity-alternating pilots as block-type I and non-alternating pilot as block-type II. For the alternating scheme, pilots are defined as,

$$P_{l_i} = \begin{cases} (1+j)/\sqrt{2}, & \text{for} \quad i \quad \text{even} \\ -(1+j)/\sqrt{2}, & \text{for} \quad i \quad \text{odd} \end{cases} \tag{17}$$

$$i = 0,\cdots,N_P - 1$$

whereas in the non-alternating scheme, pilots are defined as in (15).

Fig. 3 (a) presents the frequency spectrum resulting from the pilot tones in comb-type arrangement. The same situation can also happen to the sub-carriers conveying the information symbols. The side-lobe phenomena lead to ICI among the sub-carriers, including the pilot tones themselves. Changing the pilot arrangement into block-type generally does not help reducing the side-lobes, as shown by the dotted-line curve in Fig. 3 (b). However, by employing pilot tones of alternating polarity in block-type arrangement, side-lobe cancellation can be expected.

In Fig. 3 (b), it is depicted that the spectral side-lobe for the block-type arrangement with alternating pilot scheme shown by the solid-line curve is smaller than that for the block-type arrangement with non-alternating polarity scheme as well as the comb-type. Here, we use four pilots and 32 sub-carriers. This pilot sequence is placed on the first four of 32 sub-carriers. In this scheme, the accumulation of pilot side-lobe cancels each other and produces smaller side-lobe compare to the other scheme. Smaller side-lobe will produce smaller ICI distortion not only among pilots but also to the data. If the distortion on pilot can be suppressed, then the channel estimation error will be minimized. Hence, by using the block-type I, the ICI mitigation techniques can be improved through minimizing the channel estimation error without the expense of capacity.

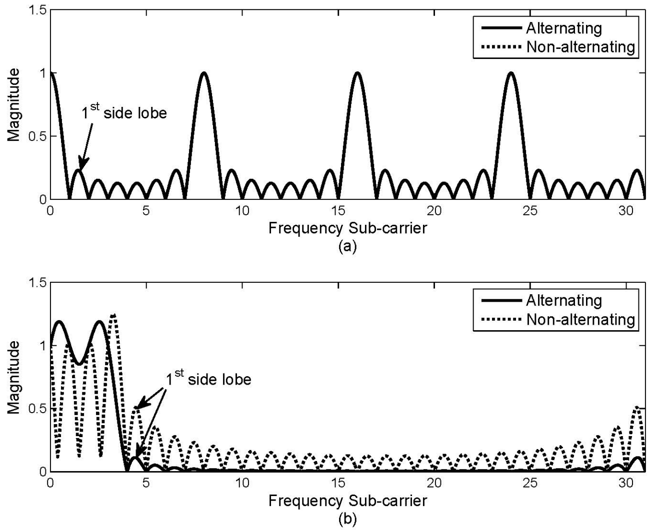

Fig. 3. Spectrum of (a) Comb-type Pilot (The spectra for the alternating and non-alternating type are coincide) and (b) Block-type I Pilot

Furthermore to investigate the best way to use the guard band, we define three variants of block-type I pilot arrangement with different numbers of guard bands placed on one or both sides of the pilot block , i.e., with one guard band only on one side of the pilot block, with one guard band on each of both sides, and with two guard bands only on one side. In the scenario with only one side having a guard band or two, the guard bands are placed at sub-carrier frequency immediately higher than the pilot block and no guard bands are allocated at the other side of the pilot block.

C. Doppler Compensation

The ICI mitigation can be further improved by reducing the channel time variation rate using frequency Doppler compensation technique. The frequency compensation that must be provided depends on the maximum Doppler shift. Therefore, it requires maximum Doppler shift estimation. For this estimation, the Doppler spread estimation algorithm in [20] can be used. Here, we assume that the maximum Doppler shift estimation is perfect. Using Doppler compensation frequency, f_o , the channel impulse response can be written as:

$$\begin{aligned} h'(t) &= e^{-j2\pi f_o t} h(t) \\ &= e^{-j2\pi f_o t} h(t) \sqrt{\tfrac{2}{P}} \textstyle\sum_{p=1}^{P} e^{j(2\pi f_d \cos(\alpha_p)t + \phi_p)} \\ &= \sqrt{\tfrac{2}{P}} \textstyle\sum_{p=1}^{P} e^{j\left(2\pi\left(f_d - \frac{f_o}{\cos(\alpha_p)}\right)\cos(\alpha_p)t + \phi_p\right)} \end{aligned} \tag{18}$$

Analogous to [16], the auto-correlation function of (18) can be derived as:

$$R_{h'h'}(\tau) = \frac{2}{\pi}\int_0^{\frac{\pi}{2}} \cos\left(2\pi\left(f_d - \frac{f_o}{\cos\alpha}\right)\tau\cos\alpha\right) d\alpha \tag{19}$$

In our previous work [15], the optimum Doppler compensation to reduce the Doppler shift was obtained as:

$$f_{o,optimum} = \arg\max_{0 \le f_o \le f_d} \int_{-\infty}^{\infty} \left(Z\left(f_o,\tau\right)\right)^2 d\tau \tag{20}$$

TABLE I
PARAMETERS USED IN THE SIMULATIONS

Parameters	Specifications
FFT size	32
Number of Carriers	32
Pilot Ratio	1/8
Signal Constellation	QPSK
Channel Model	Mobile-to-fixed with single ring scatterer

where $Z(f_o, \tau) = R_{h'h'}(\tau)$. In [15], we studied the CFO compensation taking for generality that all frequency parameters are normalized by the frequency spacing between two adjacent sub-carriers. Using (20), we examine that compensation of carrier frequency offset (CFO) due to Doppler effect with $f_{o,norm} = 0.55 f_{d,norm}$ provides better ICI mitigation performance compared to $f_{o,norm} = f_{d,norm}$, defined as complete Doppler compensation. Furthermore, under perfect channel estimation assumption, the CFO compensation with $f_{o,norm} = 0.55 f_{d,norm}$ is found to be optimum. Therefore, the time-varying channel rate can be reduced significantly if we choose Doppler compensation, $f_o = 0.55 f_d$. We propose and investigate the joint adoption of pilot arrangement and Doppler compensation to improve the ICI mitigation technique capabilities in large maximum normalized Doppler shift $(f_{d,norm} \geq 0.2)$.

III. SIMULATION AND ANALYSIS

A. Description of Simulation

OFDM system parameters used in our simulation are given in Table I. We assume the inter symbol interference is perfectly eliminated by using guard interval greater than maximum delay spread. Simulations are carried out for different values of bit energy-to-noise ratio (E_b/N_0) and pilot types without and with Doppler compensation.

B. Simulation Result and Analysis

To see how ICI mitigation methods can be improved using proper pilots arrangement, Fig. 4 shows the bit error rate (BER) for different pilot schemes. We investigate two block-type pilot arrangements, i.e., block-type I and II. From Fig. 4 block-type I pilot arrangement reduces the error floor considerably even without guard band. Compared to the comb-type pilot arrangement with a guard band on each side of the pilot, the performance of ICI mitigation with block-type is even better. Moreover, if we consider the comb-type pilot arrangement that uses, say, one eighth or 12.5% of all sub-carriers as pilots with guard bands on both sides of each pilot, then the use of block-type I can save $2 \times 12.5\% = 25\%$ the capacity that is lost with the comb-type since with block-type I we do not need guard bands to achieve better performance.

From Fig. 4, if we choose the non-alternating variant of the block type (i.e. block-type II), then the ICI mitigation performance decreases even worse than the comb-type. Hence for the block-type, the alternating pilot scheme is necessary to improve ICI mitigation technique since with this pilot arrangement, the side-lobe will cancel each other significantly, and reduce the ICI distortion not only to the pilot itself but

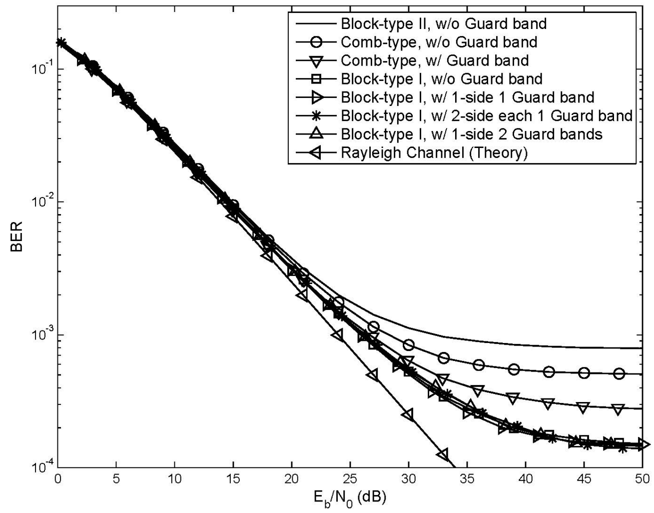

Fig. 4. Performance for Different Pilot Arrangement for $f_{d,norm}$ =0.1

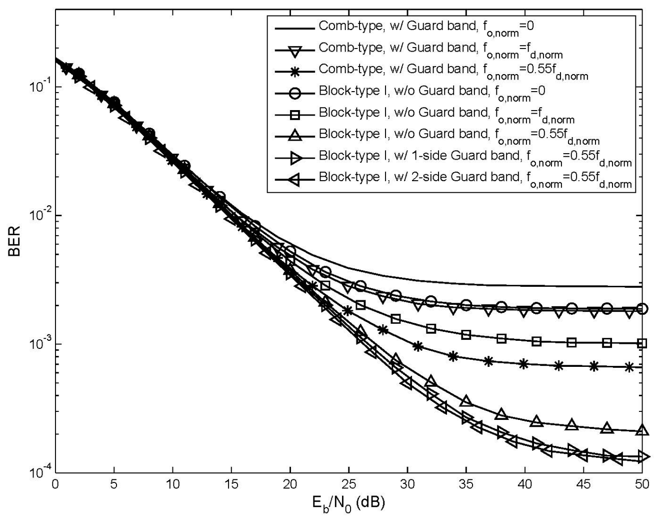

Fig. 5. Performance of ICI Mitigation with Doppler Compensation for $f_{d,norm}$ =0.2

also to the data block. Furthermore, adding guard band on the block-type I pilot will slightly reduce error floor.

Fig. 5. compares ICI mitigation improvement when Doppler compensation is applied. The Doppler compensation reduces the error floor significantly if we use block-type with alternating pilot and Doppler compensation $f_{o,norm}$ equal to $0.55 f_{d,norm}$. As shown in Fig. 5 for $f_{d,norm} = 0.2$, the improvement due to the Doppler compensation of $0.55 f_{d,norm}$ exceeds those obtained with $f_{o,norm} = f_{d,norm}$. Furthermore, using the block type pilot arrangement with alternating polarity, error-floor can be reduced even more from 6.5×10^{-4} to 2×10^{-4} if it is compared to the comb-type pilot with guard band. Hence, the accuracy of channel mid-point estimation is necessary which can be achieve by using the block-type pilot arrangement with alternating polarity, so that the ICI mitigation performance can be improved.

IV. Conclusion

It has been shown that ICI mitigation technique can be improved by proper pilot arrangement and Doppler compensation. Using block-type pilot arrangement with alternating polarity and optimal Doppler compensation, i.e. $f_o = 0.55 \times f_d$, the error floor can be reduced considerably. In addition, 25% sub-carrier loss due to the use of spectrum for guard band requirement can be prevented if block-type pilot arrangement with alternating pilots without guard band is used compare to comb-type pilot arrangement with guard band.

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Titiek Suryani was born on November 30, 1964, in Surabaya, Indonesia. She received the B.Eng. degree in Electrical Engineering from Institut Teknologi Sepuluh Nopember (ITS) Surabaya, Indonesia, in 1988, and M.Eng degree in electrical engineering from Insitut Teknologi Bandung (ITB), Indonesia, in 1999. He is currently working toward the Ph.D. degree in Electrical Engineering from ITS. Since 1989, she was as lecturer with the Department of Electrical Engineering, ITS. Her research interests include digital communications, digital signal processing, mobile wireless communications and cooperative communications.

Gamantyo Hendrantoro received the B.Eng degree in Electrical Engineering from Institut Teknologi Sepuluh Nopember (ITS), Indonesia, in 1992, and the MEng and PhD degrees in Electrical Engineering from Carleton University, Canada, in 1997 and 2001, respectively. He is presently a Professor in the Department of Electrical Engineering of ITS. His research interests include radio propagation channel modelling and wireless communications. He is presently engaged in a number of collaborative studies, including those on milimeter-wave communications, HF communications, cooperative communications and nano-satellites. Dr. Hendrantoro is a Senior Member of the IEEE.