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Decision-directed phase noise compensation for millimeter-wave single carrier systems with iterative frequency-domain equalization

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This paper proposes a receiver that repeats iterative frequency-domain equalization (FDE) and decision-directed phase noise compensation (DD-PNC) to alleviate degradation due to the phase noise for millimeter-wave single carrier (SC) systems. High bit-rate SC-FDE transceivers based on the single-chip Si RF-CMOS IC technology in the 60-GHz millimeter-wave band have been extensively studied for wireless personal area network (WPAN) systems, and the relatively large phase noise in a phase-locked loop (PLL) synthesizer severely degrades transmission performance. In an initial processing of the proposed receiver, a cyclic prefix (CP)-based phase noise compensator (CP-PNC) removes the phase noise from a time-domain received signal by using CP, which is known to the receiver, and the channel is equalized by the iterative FDE using the conventional minimum mean-square-error (MMSE) weight. In an iterative processing, DD-PNC estimates the phase noise each symbol by exploiting an output of a channel decoder, and then compensates the time-domain received signal for the phase noise by using the estimate. In order to equalize the compensated received signal, the iterative FDE performs both the MMSE filtering and residual inter-symbol interference cancelation using the decoder output. Computer simulations following the 60-GHz WPAN standard demonstrate that in the 64QAM with the coding rate of 3/4, the proposed receiver with three iterations can drastically remove the phase noise of −85 dBc/Hz at 1 MHz offset, and that it can achieve excellent transmission performance.

Keywords: Millimeter wave, Single carrier, Frequency-domain equalization, Phase noise, Decision-directed estimation

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I. INTRODUCTION

Wireless personal area network (WPAN) systems in 60-GHz millimeter-wave band have been extensively studied, and 60-GHz WPAN standardization in IEEE 802.15.3c investigates both single carrier (SC) transmission employing frequency-domain equalization (FDE), which is called SC-FDE, and OFDM (orthogonal frequency division multiplexing) transmission because of indoor multipath channels [1–4]. Moreover, single-chip RF-CMOS ICs based on the silicon (Si) technology for the 60-GHz WPAN transceiver have been developed to reduce power consumption and cost [5]. However, the relatively large phase noise from the phase-locked loop (PLL) synthesizer occurs. In the 60-GHz SiGe BiCMOS chip, the phase noise of −85 to −90 dBc/Hz at 1 MHz offset was measured when the PLL synthesizers for the transmitter and receiver were cascaded [5]. Therefore, a single-chip Si CMOS transceiver requires a compensation method for the larger phase noise.

It is well known that OFDM is sensitive to the phase noise and frequency offset [6]. This is because the phase noise causes signal distortion in OFDM, which is characterized by common phase error (CPE), which induces a common phase shift at all subcarriers, and inter-carrier interference (ICI). In order to overcome this impairment in OFDM, several phase noise compensation methods have been proposed [7, 8]. Especially, as one of the lowest-complexity phase noise compensation methods with ICI cancelation capability, decision-directed phase noise compensation (DD-PNC) has been proposed for the 60-GHz OFDM transmission [8]. OFDM can suppress CPE by using pilot subcarriers inserted in each OFDM symbol, and the relatively exact tentative decisions can be obtained from the coherent detection. Thus, DD-PNC can compensate a time-domain received signal for the phase noise by exploiting decoded bits that a channel decoder calculates from the tentative decisions.

Similarly to OFDM, SC-FDE suffers from severe performance degradation due to the phase noise, and it has been investigated that SC-FDE is more affected than OFDM [9]. In SC-FDE, it is difficult to compensate for the phase noise before FDE, and thus the time-varying phase noise in the FFT duration drastically causes ICI in the frequency-domain received signals after fast Fourier transform (FFT). Combining decision-feedback equalization (DFE) and the phase noise compensation for the SC transmission without FDE has been investigated [10]. In addition, a phase noise suppression algorithm in single carrier-frequency division multiple access (SC-FDMA), which is also called DFT-spread OFDM, has been proposed [11]. However,

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SC-FDMA can use frequency-domain pilots, and thus it employs the same phase noise correction using the pilots as OFDM. Therefore, these methods cannot be directly applied to the SC-FDE transmission in the 60-GHz WPAN system.

Moreover, a joint method of iterative block decision feedback equalization (IBDFE) and post-equalization carrier synchronization in SC-FDE has been proposed [12]. The method jointly compensates for the frequency offset and equalizes the received signal in an iterative manner, while it does not consider the phase noise and cannot prevent the error propagation in the decision feedback loop because it does not employ the channel coding. On the other hand, a combination method of turbo FDE (TFDE) and the post-equalization carrier synchronization, which copes with the frequency offset and the phase noise, has been proposed [13]. It iterates TFDE with the decision-directed estimation for the unknown parameters by exploiting the log-likelihood ratio (LLR) of coded bits which the maximum *a posteriori* (MAP) decoder provides. These two combination methods can achieve a good performance under the relatively small phase noise condition, while the large phase noise drastically degrades the equalization performance of IBDFE and TFDE. This is because they cannot equalize ICI which the phase noise causes, and the residual inter-symbol interference (ISI) occurs in the FDE outputs. Hence, the post-equalization carrier synchronization, which uses the FDE outputs for the parameter estimation, cannot effectively operate in the 60-GHz SC-FDE transmission with the relatively large phase noise. In addition, TFDE requires the high-complexity processing such as the MAP decoding and calculating the mean and variance of the modulation symbol from LLRs, and the high-complexity method is not suitable for multi-Gbps transmission of the 60-GHz WPAN system because the realistic implementation is not available.

In order to provide an efficient, low-cost and reliable solution to the phase noise problem for the 60-GHz SC system, this paper proposes a SC-FDE receiver employing an iterative FDE and DD-PNC that can cope with multipath delay and phase noise for the SC transmission. In an initial processing of the proposed receiver, a cyclic prefix (CP)-based phase noise compensator (CP-PNC) removes the phase noise from the time-domain received signal before FFT by using CP, which is known to the receiver, and then the channel is equalized by the iterative FDE using the conventional minimum mean-square-error (MMSE) weight [14, 15]. In an iterative processing, DD-PNC estimates the phase noise each symbol by exploiting the decoded bits of the decoder output, and then compensates the time-domain received signal for the phase noise by using the estimate. Since DD-PNC can remove the phase noise from the time-domain received signal before FDE, the iterative FDE does not have to consider the phase noise suppression, and thus the low-complexity iterative FDE that consists of the MMSE filtering and the residual ISI cancelation using the decoded bits can be introduced into the proposed SC-FDE receiver. Similarly to IBDFE and TFDE, the iterative FDE can outperform the conventional MMSE-based FDE (MMSE-FDE) by eliminating the residual ISI. In order to verify the effectiveness of the proposed SC-FDE receiver, computer simulations basically following both the 60-GHz WPAN standard and the one-pole/one-zero phase noise model proposed in IEEE 802.15.3c are conducted. The simulation results demonstrate that the proposed receiver with the iterative FDE and DD-PNC sufficiently removes the phase noise of −85 dBc/Hz at 1 MHz offset, and can realize the multi-Gbps transmission while requiring the low complexity in such a severe phase noise condition.

This paper is organized as follows. In Section II, the proposed SC-FDE receiver is explained briefly, and then transmitted and received signal models are derived along with the clarification of the phase noise problem. Section III describes the detail of the SC-FDE receiver employing CP-PNC and DD-PNC in addition to the iterative FDE. Section IV presents the computer simulation results in terms of packet error rate (PER) performance of the proposed receiver. Finally, some concluding remarks are given in Section V.

II. SC-FDE TRANSMISSION WITH DD-PNC

A) System model

As shown in Fig. 1, a baseband SC-FDE signal is upconverted by a local oscillator, and a received signal through the multipath fading channel is downconverted by another local oscillator. Each local oscillator consists of the PLL synthesizer whose carrier frequency is assumed to be perfectly synchronized. However, the phase noise in the PLL synthesizer severely degrades transmission performance in SC-FDE. Note that this paper handles the phase noise only on the receiver side [6–13], and that the carrier frequency offset is assumed to be negligible.

A block diagram of an SC-FDE transmitter in 60-GHz WPAN is shown in Fig. 2. The transmitter first adds the cyclic redundancy check (CRC) code to detect a decision error of the transmitted packet, and then it encodes information bits by the convolutional code. After interleaving the coded bits, it maps them into a modulation symbol, and CP is inserted. Note that as shown in Fig. 3, Golay code, which is known to both the transmitter and the receiver, is used as CP, and that another complementary Golay code is time multiplexed at the head of the packet as a preamble for the channel estimation [1].

Figure 4 shows a block diagram of the proposed SC-FDE receiver employing the iterative FDE and DD-PNC. The receiver iterates the phase noise compensation and signal detection by exploiting an output of the channel decoder. In an initial processing, a channel estimator first estimates a channel impulse response by using the preamble, and then a CP-PNC removes the phase noise from the time-domain received signal by using CP of the Golay code and the channel estimates. After FFT converts the compensated received signal into the frequency-domain one, the iterative FDE performs the conventional MMSE-FDE [14, 15], and then the output is transformed into the time-domain

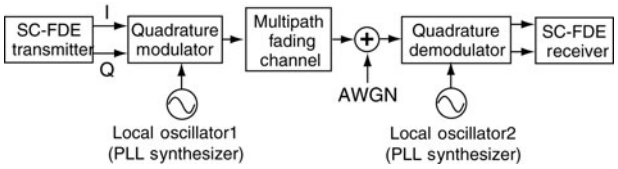


Fig. 1. System model.

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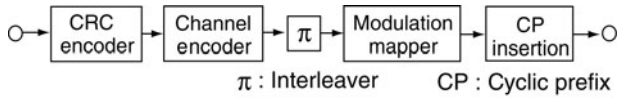


Fig. 2. SC-FDE transmitter.

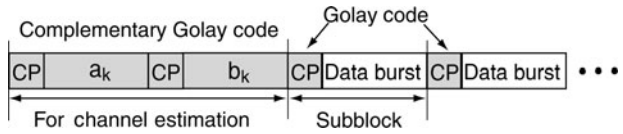


Fig. 3. Packet configuration.

equalized signal by inverse fast Fourier transform (IFFT). Note that the iterative FDE does not employ the residual ISI cancellation in the initial processing because it cannot obtain the decoded bits. Next, the soft-output detector provides soft decision from the equalized signal, and after deinterleaving the soft decisions, a soft-decision Viterbi decoding is carried out. If a CRC decoder detects a decision error from the decoded bits, the receiver shifts the initial processing to an iterative processing. In the iterative processing, the receiver first generates a transmitted signal replica from the decoded bits. Next, DD-PNC recursively estimates the phase noise by the one-tap least-mean-square (LMS) algorithm using the replica, and then it removes the estimate from the time-domain received signal. After FFT, the iterative FDE, which consists of the MMSE filtering and the residual ISI cancellation using the decoded bits [14, 15], is also performed for the compensated signal. The receiver repeats the iterative processing until the number of iterations exceeds a predetermined threshold or no decision error is detected.

B) Signal model

A received signal in the SC-FDE transmission with the phase noise is represented by an equivalent low pass model. Let N and Δ_t denote the number of FFT points and the sampling period which is equal to the symbol duration, respectively, and the number of symbols during the subblock, which consists of CP and data symbols as shown in Fig. 3, is equal to N. A sampled transmitted signal $s_{i,k}$ at the kth symbol of the ith subblock is given by

$$s_{i,k} = \begin{cases} c_{k+N_C} & (-N_c \le k \le -1) \\ z_{i,k} & (0 \le k \le N - N_c - 1) \\ c_{k+N_C-N} = s_{i+1,k-N} & (N - N_c \le k \le N - 1) \end{cases} \quad (1)$$

where c_k is the Golay code with the length of N_c samples, and $z_{i,k}$ is the kth data modulation symbol of the ith subblock. Its mean and variance are 0 and 1, respectively.

Furthermore, let $h(\tau)$ denote an impulse response of a D-path fading channel, and $h(\tau)$ is given using Dirac's delta function $\delta(\tau)$ by

$$h(\tau) = \sum_{d=0}^{D-1} h_d \delta(\tau - d\Delta_t), \quad (2)$$

where h_d and $d\Delta_t$ are a complex amplitude and a delay time of the dth path, respectively, and the maximum delay time $(D-1)\ \Delta_t$ is assumed to be less than $N_c\Delta_t$. Note that the fading fluctuation during one packet is assumed to be negligible.

The phase noise $\phi(t)$ at the time $t = (iN + k)\Delta_t$ is denoted by $\phi(t) = \phi_{i,k}$, and its mean and variance are zero and σ_ϕ^2, respectively. A sampled received signal $r_{i,k}$ is given by

$$r_{i,k} = e^{j\phi_{i,k}} \sum_{d=0}^{D-1} h_d s_{i,k-d} + n_{i,k}, \quad (3)$$

where $n_{i,k}$ is a sampled noise with zero mean and variance σ_n^2. In addition, let us express $e^{j\phi_{i,k}}$ by using a sum of the mean $\bar{\Phi}_i$ of the phase noise during the ith subblock and a remaining component $\Phi_{i,k}$ at the kth symbol from $\bar{\Phi}_i$ as

$$e^{j\phi_{i,k}} = \bar{\Phi}_i + \Phi_{i,k}. \quad (4)$$

Then, $r_{i,k}$ can be written as

$$r_{i,k} = \left(\bar{\Phi}_i + \Phi_{i,k}\right) \sum_{d=0}^{D-1} h_d s_{i,k-d} + n_{i,k}. \quad (5)$$

After FFT converts $r_{i,k}$ except those corresponding to CP, a received signal $R_{i,n}$ at the nth frequency component ($0 \le n \le N - 1$) is expressed as

$$\begin{aligned} R_{i,n} &= \frac{1}{\sqrt{N}} \sum_{k'=0}^{N-1} r_{i,k'} e^{-j2\pi(n/N)k'}, \\ &= \frac{\bar{\Phi}_i}{\sqrt{N}} \sum_{d=0}^{D-1} h_d \sum_{k'=0}^{N-1} s_{i,k'-d} e^{-j2\pi(n/N)k'} \\ &\quad + \Xi_{i,n} + N_{i,n}, \end{aligned} \quad (6)$$

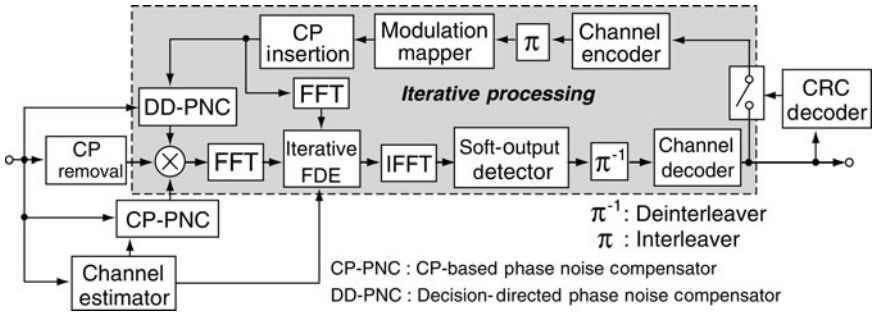


Fig. 4. SC-FDE receiver employing iterative FDE and DD-PNC.

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$$\Xi_{i,n} = \frac{1}{\sqrt{N}} \sum_{d=0}^{D-1} h_d \sum_{k'=0}^{N-1} \Phi_{i,k'} s_{i,k'-d} e^{-j2\pi(n/N)k'}, \quad (7)$$

$$N_{i,n} = \frac{1}{\sqrt{N}} \sum_{k'=0}^{N-1} n_{i,k'} e^{-j2\pi(n/N)k'}. \quad (8)$$

$\Xi_{i,n}$ causes the performance degradation due to the phase noise. If the PLL synthesizer can reduce the phase noise to a sufficiently low level, i.e. $\Phi_{i,k} \simeq 0$, $\Xi_{i,n}$ does not occur.

Let $S_{i,n}$ denote the nth frequency component into which FFT converts $s_{i,k}$, and $S_{i,n}$ is given by

$$S_{i,n} = \frac{1}{\sqrt{N}} \sum_{k'=0}^{N-1} s_{i,k'} e^{-j2\pi(n/N)k'}, \quad (9)$$

and the following term of (6) can be expressed using $S_{i,n}$ as

$$\begin{aligned} &\frac{1}{\sqrt{N}} \sum_{k'=0}^{N-1} s_{i,k'-d} e^{-j2\pi(n/N)k'} \\ &= \frac{e^{-j2\pi(n/N)d}}{\sqrt{N}} \sum_{k'=-d}^{N-d-1} s_{i,k'} e^{-j2\pi(n/N)k'} \\ &= e^{-j2\pi(n/N)d} S_{i,n}. \end{aligned} \quad (10)$$

Hence, from (6) and (10), $R_{i,n}$ is given by

$$R_{i,n} = \bar{\Phi}_i H_n S_{i,n} + \Xi_{i,n} + N_{i,n}, \quad (11)$$

$$H_n = \sum_{d=0}^{D-1} h_d \, e^{-j2\pi \frac{n}{N} d}. \quad (12)$$

III. SC-FDE RECEIVER EMPLOYING ITERATIVE FDE AND DD-PNC

A) Channel estimation

In the initial processing, the channel impulse response, which includes the phase noise, is estimated by using the preamble. The complementary Golay code a_k, b_k (± 1, $-N_c \le k \le N-1$) has the following property:

$$\frac{1}{2N}\left[\rho_a(l) + \rho_b(l)\right] = \begin{cases} 1, & l = 0, \\ 0, & \text{otherwise.} \end{cases} \quad (13)$$

$$\rho_a(l) = \sum_{k'=0}^{N-1} a_{k'} a_{k'-l}, \quad (14)$$

$$\rho_b(l) = \sum_{k'=0}^{N-1} b_{k'} b_{k'-l}, \quad (15)$$

where $0 \le l \le N_c$. By calculating cross correlation between $r_{i,k}$ and the preamble consisting of a_k and b_k in Fig. 3, the channel estimator estimates $g_{i_p,l}$ which is the complex amplitude of the lth path multiplied by the phase noise, and from (5) and (13), $g_{i_p,l}$ is given by

$$\begin{aligned} g_{i_p,l} &= \frac{1}{2N} \sum_{k'=0}^{N-1} (r_{0,k'}\ a_{k'-l} + r_{1,k'}\ b_{k'-l}) \\ &= \frac{1}{2N} \sum_{d=0}^{D-1} h_d \bar{\Phi}_{i_p} \sum_{k'=0}^{N-1} (a_{k'-d} a_{k'-l} + b_{k'-d} b_{k'-l}) \\ &\quad + \nu_{i_p,l} \\ &= h_l \bar{\Phi}_{i_p} + \nu_{i_p,l}, \end{aligned} \quad (16)$$

where $\bar{\Phi}_{i_p}$ is the mean of the phase noise during the preamble. $\nu_{i_p,l}$ is the estimation error that the phase noise and the noise cause, and is given by

$$\begin{aligned} \nu_{i_p,l} &= \frac{1}{2N} \sum_{d=0}^{D-1} h_d \sum_{k'=0}^{N-1} (\Phi_{0,k'}\ a_{k'-d}\ a_{k'-l} + \Phi_{1,k'} b_{k'-d}\ b_{k'-l}) \\ &\quad + \frac{1}{2N} \sum_{k'=0}^{N-1} (n_{0,k'}\ a_{k'-l} + n_{1,k'}\ b_{k'-l}). \end{aligned} \quad (17)$$

In order to improve accuracy of the estimation and estimate a normalized maximum delay L that CP-PNC requires, a path detector adaptively extracts the dominant paths from $g_{i_p,l}$. The path detector sets $g_{i_p,l}$ to zero, if $|g_{i_p,l}|^2 < \alpha(\sigma_n^2 + \sigma_\phi^2)$, where α is a threshold parameter. After that, the maximum path index l that satisfies $g_{i_p,l} > 0$ is set to L. Let Ω denote the total number of extracted paths with $g_{i_p,l} > 0$ in the output of the path detector.

B) CP-PNC and MMSE-FDE

In the initial processing, CP-PNC estimates the phase noise using CP of the Golay code inserted at the head and the end of the subblock, and compensates the time-domain received signals for the phase noise by using the estimates. CP-PNC first generates a received signal replica $\hat{r}_{i,k}$ by using CP c_k and the channel estimate $g_{i_p,l}$ as

$$\hat{r}_{i,k} = \sum_{l=0}^{L} g_{i_p,l}\ c_{k-l}, \quad (18)$$

where $-N_c + L \le k \le -1$, because the multipath delay causes the leakage of the data symbols from the $(i-1)$th subblock as shown in Fig. 5(b). Next, let $\Delta\Phi_i$ denote the relative phase shift of $\bar{\Phi}_i$ to Φ_{i_p}, and the estimate of $\Delta\hat{\Phi}_i$ can be calculated based on the MMSE criterion that should minimize the cost function $J_{\Delta\Phi}$ given by

$$\begin{aligned} J_{\Delta\Phi} &= E\left[\left|\Delta\Phi_i - \Delta\hat{\Phi}_i\right|^2\right] \\ &= E\left[\left|\Delta\Phi_i - \sum_{k'=-N_c+L}^{-1} C_{i,k'}^* r_{i,k'}\right|^2\right], \end{aligned} \quad (19)$$

where $E[\ ]$ denotes an ensemble average, $C_{i,k'}$ is the coefficient of the MMSE estimation, and the superscript * is the complex conjugate. Setting the gradient of $J_{\Delta\Phi}$ with respect to $C_{i,k'}$ to

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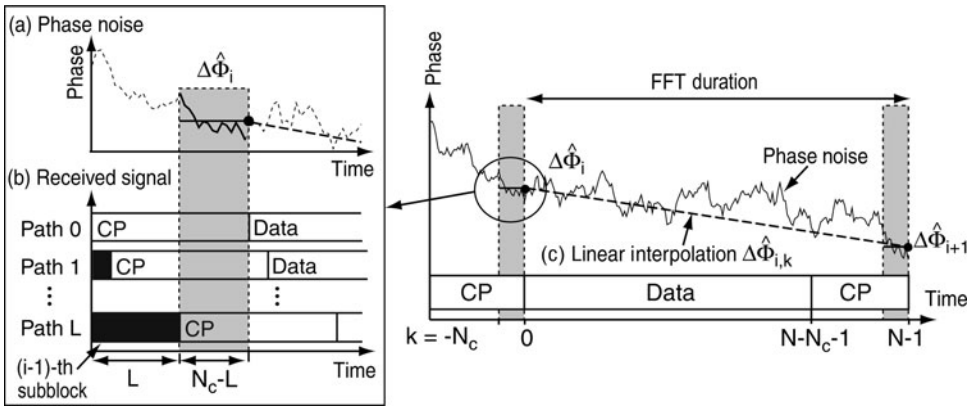


Fig. 5. Phase noise compensation by CP-PNC.

zero, $C_{i,k'}$ is given by

$$C_{i,k'} = \frac{\hat{r}_{i,k'}}{(N_c - L)(|\hat{r}_{i,k'}|^2 + \sigma_n^2)}. \tag{20}$$

The derivation of the above equation is based on the following approximation:

$$E[r_{i,k'}\Delta\Phi_i^*] \simeq \hat{r}_{i,k'}, \quad E[r_{i,k'}r_{i,k'}^*] \simeq |\hat{r}_{i,k'}|^2 + \sigma_n^2. \tag{21}$$

Hence, $\Delta\hat{\Phi}_i$ can be estimated as

$$\Delta\hat{\Phi}_i = \frac{1}{N_c - L}\sum_{k'=-N_c+L}^{-1} \frac{\hat{r}_{i,k'}^*\, r_{i,k'}}{|\hat{r}_{i,k'}|^2 + \sigma_n^2}. \tag{22}$$

CP-PNC also calculates $\Delta\hat{\Phi}_{i+1}$ of the $(i+1)$th subblock, and estimates the relative phase shift $\Delta\Phi_{i,k} = (\bar{\Phi}_i + \Phi_{i,k})/\bar{\Phi}_{i_p}$ at the kth data symbol by linearly interpolating $\Delta\hat{\Phi}_i$ and $\Delta\hat{\Phi}_{i+1}$, and the estimate of $\Delta\Phi_{i,k}$, $\Delta\hat{\Phi}_{i,k}$ is calculated as

$$\Delta\hat{\Phi}_{i,k} = \frac{N-k}{N}\Delta\hat{\Phi}_i + \frac{k}{N}\Delta\hat{\Phi}_{i+1}. \tag{23}$$

Finally, CP-PNC compensates the time-domain received signal for the phase noise by using $\Delta\hat{\Phi}_{i,k}$ as

$$r'_{i,k} = \frac{r_{i,k}}{\Delta\hat{\Phi}_{i,k}} = \bar{\Phi}_{i_p}\sum_{d=0}^{D-1} h_d\, s_{i,k-d} + \xi'_{i,n} + n'_{i,k}, \tag{24}$$

$$\xi'_{i,k} = \left(\frac{\Delta\Phi_{i,k}}{\Delta\hat{\Phi}_{i,k}} - 1\right)\bar{\Phi}_{i_p}\sum_{d=0}^{D-1} h_d s_{i,k-d}, \tag{25}$$

where $n'_{i,k} = n_{i,k}/\Delta\hat{\Phi}_{i,k}$ and ξ'_{i,k_n} is the residual phase noise component of the compensated received signal. FFT converts $r'_{i,k}$ into $R'_{i,n}$ as

$$R'_{i,n} = \bar{\Phi}_{i_p} H_n S_{i,n} + \Xi'_{i,n} + N'_{i,n}, \tag{26}$$

$$\Xi'_{i,n} = \frac{1}{\sqrt{N}}\sum_{k'=0}^{N-1} \xi'_{i,k'}\, e^{-j2\pi(n/N)k'}, \tag{27}$$

$$N'_{i,n} = \frac{1}{\sqrt{N}}\sum_{k'=0}^{N-1} n'_{i,k'}\, e^{-j2\pi(n/N)k'}. \tag{28}$$

It is assumed that the phase noise can be perfectly removed from the received signal, i.e. $\Xi'_{i,n} \simeq 0$, and $R'_{i,n}$ can be approximated as

$$R'_{i,n} \simeq G_n S_{i,n} + N'_{i,n}, \tag{29}$$

$$G_n = \sum_{l=0}^{L} g_{i_p,l}\, e^{-j2\pi(n/N)l}. \tag{30}$$

Since the iterative FDE cannot obtain the decoded bits in the initial processing, and cannot perform the residual ISI cancelation, it operates as the conventional MMSE-FDE. Hence, MMSE-FDE equalizes $R'_{i,n}$ based on the MMSE criterion. Let $\tilde{S}_{i,n}$ denote the estimate of $S_{i,n}$, and the cost function J_S is defined as

$$J_S = E\left[\left|S_{i,n} - \tilde{S}_{i,n}\right|^2\right] = E\left[\left|S_{i,n} - F_n^* R'_{i,n}\right|^2\right], \tag{31}$$

where F_n is a coefficient of the MMSE filter. Setting the gradient of J_S with respect to F_n to zero, F_n is given from (29) by

$$F_n = \frac{G_n}{|G_n|^2 + \sigma_n^2}, \tag{32}$$

and thus the MMSE-FDE calculates $\tilde{S}_{i,n}$ as

$$\tilde{S}_{i,n} = \frac{G_n^* R'_{i,n}}{|G_n|^2 + \sigma_n^2}. \tag{33}$$

Note that MMSE-FDE cannot suppress the residual phase noise component $\Xi'_{i,n}$ because the derivation of F_n ignores $\Xi'_{i,n}$.

C) DD-PNC and iterative FDE

In the iterative processing, DD-PNC estimates the relative phase shift of the phase noise by the one-tap LMS algorithm, which minimizes a mean-square-error between $r_{i,k}$ and its replica [8]. DD-PNC first generates the received signal

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replica $\hat{r}_{i,k}$ excluding the relative phase shift of the phase noise by using $g_{i_p,l}$ as

$$\hat{r}_{i,k} = \sum_{l=0}^{L} g_{i_p,l}\ \hat{s}_{i,k-l}, \tag{34}$$

where $\hat{s}_{i,k}$ is a transmitted signal replica, and is generated by replacing $z_{i,k}$ of (1) with a data symbol replica $\hat{z}_{i,n}$ that the iterative processing regenerates from the decoded bits.

Let $w^*_{i,k}$ denote an estimate of the relative phase shift, and the received signal replica with the phase noise can be expressed as a product of $\hat{r}_{i,k}$ and $w^*_{i,k}$. Thus, an error signal $e_{i,k}$ between $r_{i,k}$ and the replica is given by

$$e_{i,k} = r_{i,k} - w^*_{i,k}\,\hat{r}_{i,k}, \tag{35}$$

and the LMS algorithm recursively updates $w_{i,k}$ as

$$w_{i,k+1} = w_{i,k} + \mu_\phi\,\hat{r}_{i,k}\,e^*_{i,k}, \tag{36}$$

where μ_ϕ is the step-size parameter in DD-PNC and a positive number [16]. Note that an initial value of $w^*_{i,k}$ is set to one. The above recursion is performed until the last sample of the ith subblock. After that, DD-PNC smoothes $w_{i,k}$ by backward exponentially weighting to improve the accuracy of the estimation, because the LMS algorithm is the forward recursion using the post-measured data and cannot exploit the future-measured data [17]. Let $\Delta\tilde{\Phi}_{i,k}$ denote a smoothed estimate, and $\Delta\tilde{\Phi}_{i,k}$ is recursively updated as

$$\Delta\tilde{\Phi}_{i,k} = \mu_\phi w^*_{i,k} + (1-\mu_\phi)\Delta\tilde{\Phi}_{i,k+1}. \tag{37}$$

Note that $\Delta\tilde{\Phi}_{i,N} = w^*_{i,N-1}$. After smoothing $w_{i,k}$, $\Delta\tilde{\Phi}_{i,k}$ is normalized by $\sqrt{|\Delta\tilde{\Phi}_{i,k}|^2}$ because of $|\Delta\Phi_{i,k}| = 1$.

Finally, DD-PNC compensates the time-domain received signals for the phase noise by using $\Delta\tilde{\Phi}_{i,k}$ as

$$r'_{i,k} = \frac{r_{i,k}}{\Delta\tilde{\Phi}_{i,k}}, \tag{38}$$

and FFT converts $r'_{i,k}$ into $R'_{i,k}$.

The iterative FDE performs the MMSE filtering and the residual ISI cancelation using the decoded bits. The MMSE filtering and the residual ISI cancellation consist of the feedforward filter and the feedback filter, respectively [12, 14]. By using the coefficient F_n of the feedforward filter and the coefficient B_n of the feedback filter, $\tilde{S}_{i,n}$ of the iterative FDE output is given by

$$\tilde{S}_{i,n} = F^*_n R'_{i,n} - B^*_n \hat{S}_{i,n}, \tag{39}$$

where $\hat{S}_{i,n}$ is calculated by replacing $s_{i,k}$ of (9) with $\hat{s}_{i,k}$. Unlike IBDFE, the iterative FDE exploits $S_{i,n}$, which is generated from the decoded bits for the residual ISI cancellation. The feedforward and the feedback filters are designed in order to maximize the signal to interference plus noise power ratio (SINR) of the iterative FDE output. Let ρ_m denote a correlation coefficient of $E\left[S^*_{i,n}\hat{S}_{i,n}\right]$ in the mth iteration ($m = 0, 1, \ldots, M$), and from (29) and (39), the feedback coefficient B_n that minimizes the interference power is given by [14]

$$B_n = \rho_m(F_n G^*_n - \mu), \tag{40}$$

$$\mu = \frac{1}{N}\sum_{n=0}^{N-1} F^*_n G_n. \tag{41}$$

In addition, the feedforward coefficient F_n that maximizes SINR is given by

$$F_n = \frac{G_n}{(1-|\rho_m|^2)|G_n|^2 + \sigma^2_n}, \tag{42}$$

and the residual ISI cancelation error depends on ρ_m [14]. The iterative FDE employs the decoded bits instead of LLR, and cannot exactly calculate ρ_m from the variance of the modulation signal [12]. Hence, ρ_m is approximated to a real number that increases linearly with the number of iterations, i.e. $\rho_m = m/M$, which results in the complexity reduction [18]. In the initial processing ($m = 0$), $\rho_m = 0$, and the iterative FDE uses the MMSE weight. In the Mth iteration of the last iterative processing, $\rho_m = 1$, and the maximum ratio combining (MRC) weight is used in addition to the residual ISI cancellation.

Although an inner loop of the iterative FDE that includes IFFT, the decision, the CP insertion, and FFT can repeat the MMSE filtering and the residual ISI cancelation in the decision feedback manner [14], the proposed receiver does not employ the iteration in the inner loop. This is because the phase noise causes the severe performance degradation, and it can hardly be expected that only the inner loop improves the transmission performance. Hence, the proposed receiver repeats the iterative processing, which consists of DD-PNC, the iterative FDE, and the channel decoding.

The iterative FDE requires the scalar multiplication at each frequency component and FFT which generates $\hat{S}_{i,n}$ from $\hat{s}_{i,k}$. The total numerical complexity of the iterative FDE is $O(N)$ per subblock, and the additional complexity for replacing MMSE-FDE with the iterative FDE is small.

D) Numerical complexity of CP-PNC and DD-PNC

Table 1 summarizes the numerical complexity of CP-PNC and DD-PNC, that is, the number of the real multiplications per subblock in each iteration. CP-PNC first generates the received signal replica $\hat{r}_{i,k}$ using (18). Actually, the total number of extracted paths with $g_{i_p,l} > 0$ is Ω, and thus the replica generation in CP-PNC needs only $2\,(N_c - L)\,\Omega$ multiplications. Next, CP-PNC requires $6\,(N_c - L) + 4N$ multiplications to estimate $\Delta\hat{\Phi}_{i,k}$ in (22) and (23). In addition, the number of multiplications of the phase noise compensation of (24) is $4N$. Thus, the total number of real multiplications in CP-PNC becomes $(N_c - L)\,(2\,\Omega + 6) + 8N$.

Moreover, DD-PNC needs $4N\Omega$ multiplications for the replica generation, and the one-tap LMS algorithm in (35) and (36) requires $10N$ multiplications in the subblock duration. In addition, the number of the multiplications in the smoothing of (37) and the phase noise compensation of (24) are $4N$ and $4N$, respectively. Hence, the total number of the real multiplications in DD-PNC becomes $N\,(4\,\Omega + 18)$. For example, when $N = 256$, $N_c = 16$, $L = 12$, and $\Omega = 5$,

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Table 1. Numerical complexity of CP-PNC and DD-PNC.

Operation	CP-PNC	DD-PNC
Replica generation	$2\ (N_c - L)\ \Omega$	$4N\Omega$
Estimation and interpolation	$6\ (N_c - L) + 4N$	–
LMS algorithm	–	$10N$
Smoothing of (37)	–	$4N$
Compensation of (24)	$4N$	$4N$
Total	$(N_c - L)\ (2\ \Omega + 6) + 8N$	$N\ (4\Omega + 18)$

the numerical complexities of CP-PNC and DD-PNC are almost 0.5 and 2.4 times as that of FFT, respectively. Therefore, the proposed receiver can achieve the phase noise compensation with low complexity.

IV. COMPUTER SIMULATIONS

A) Simulation conditions

Computer simulations were conducted to verify the effectiveness of the proposed receiver under the simulation conditions listed in Table 2. Parameters of the SC-FDE transmission basically follow the IEEE 802.15.3c standard [1–4]. In addition to 16QAM, 64QAM was employed as the modulation scheme to achieve the higher bit-rate. The packet configuration consists of the four-subblock long preamble followed by 14 data subblocks. The number of FFT points N was 256, and the number of CP samples N_c was 16. The convolutional code with the constraint length = 7 and the punctured coding rate $R = 3/4$ was employed. The maximum number of iterations was five. In the channel estimation, α for the path detection was set to 0.003. μ_ϕ in DD-PNC was changed from 0.03 to 0.13, which increases linearly with the number of iterations. α and μ_ϕ were optimized by the computer simulation.

The channel model was five-path Nakagami–Rice fading with Ricean factor $K = 10$ dB, where K is defined as the power ratio of the line of sight (LOS) component to the non-line of sight (NLOS) one. The maximum delay in the multipath channel is $12\Delta_t$. Hence, $L = 12$ and $\Omega = 5$, if the path detector can ideally extract the dominant paths.

Table 2. Simulation conditions.

Modulation	16QAM, 64QAM
Sampling rate : $1/\Delta_t$	1.728 GHz
No. of subblocks	Preamble : 4, data : 14
No. of FFT points : N	256
No. of symbols in subblock	CP (N_c) : 16, data : 240
Channel coding	Convolutional code ($R = 3/4$)
Maximum bit rate	7.29 Gbps ($R = 3/4$, 64QAM)
Channel decoding	soft decision Viterbi decoding
Maximum no. of iterations : M	5 (except initial processing)
α for path detection	0.003
Step-size parameter μ_ϕ	0.03–0.13
Channel model	Nakagami–Rice fading
	LOS : $K = 10$ dB
	NLOS : 5-path with equal power
	Maximum delay : $12\ \Delta_t$
Phase noise model	IEEE 802.15.3c [3] by AR model

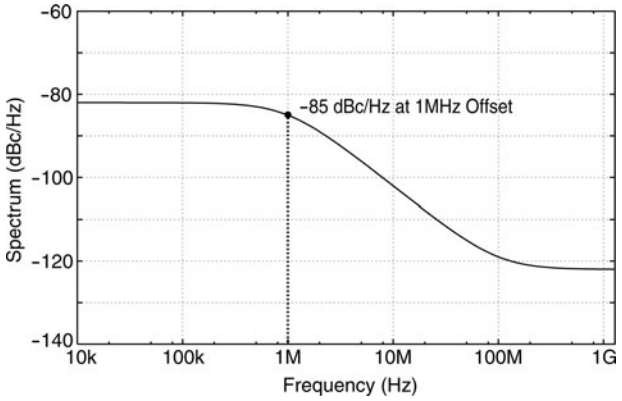


Fig. 6. PSD of phase noise.

Furthermore, the power spectrum density (PSD) of the phase noise was the one-pole/one-zero model proposed in IEEE 802.15.3c as

$$S(f) = K_\phi \frac{1 + (f/f_z)^2}{1 + (f/f_p)^2}, \tag{43}$$

where K_ϕ is the phase noise level that the loop filter determines in the low frequency, and f_p and f_z are pole and zero frequencies, respectively [3]. Figure 6 shows PSD of the phase noise with $K_\phi = -82$ dBc/Hz, $f_p = 1$ MHz, and $f_z = 100$ MHz. The phase noise level was -85 dBc/Hz at 1 MHz offset, and the variance σ_ϕ^2 of the phase noise $\phi_{i,k}$ was 0.02 rad^2. In the computer simulations, K_ϕ was changed as a parameter, and the phase noise level was defined as PSD at 1 MHz offset. In addition, the phase noise was generated by the autoregressive (AR) model, where AR parameters can be calculated from the inverse Fourier transform of (43) [16].

B) PER performance

When the phase noise level is -85 dBc/Hz, average packet error rate (PER) performances of the proposed receiver with 16QAM and 64QAM are shown in Figs 7 and 8. For comparison, PER of the SC-FDE receiver with ideal channel estimation and MMSE-FDE under the phase-noise-free

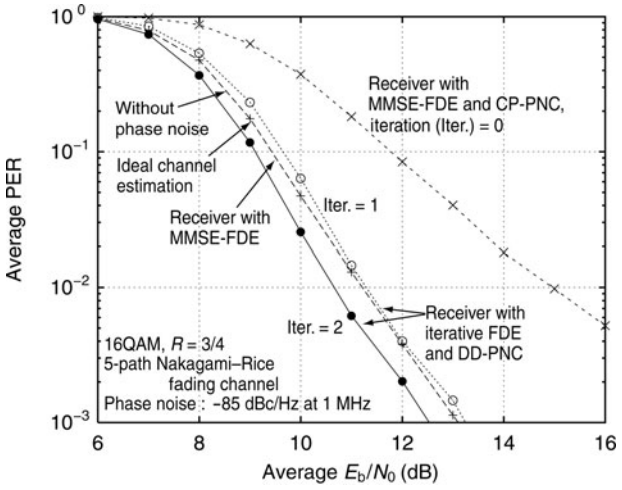


Fig. 7. Average PER performance with 16QAM.

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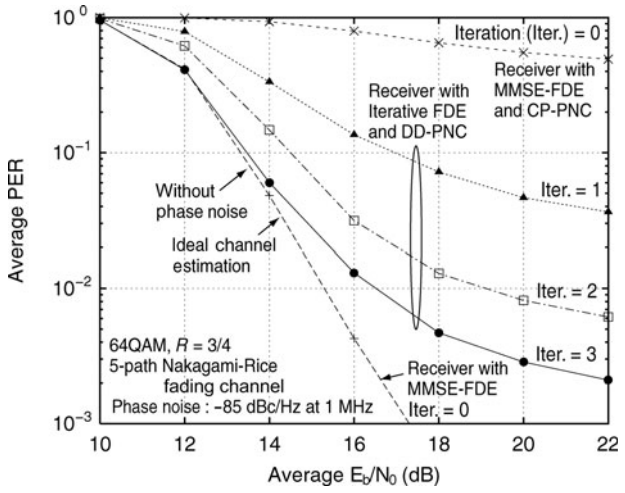


Fig. 8. Average PER performance with 64QAM.

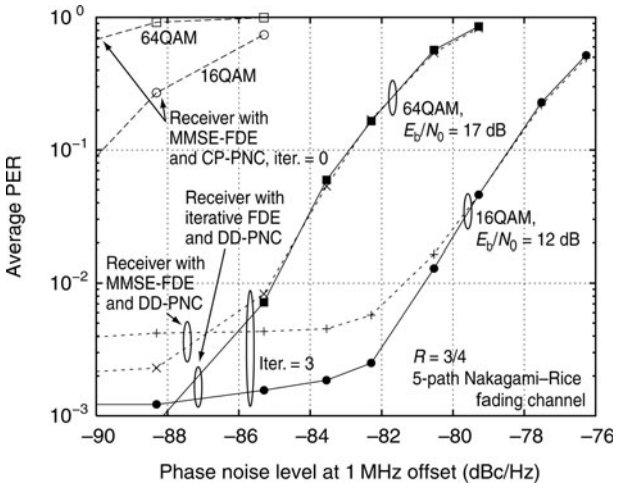


Fig. 10. Robustness of the proposed receiver with iterative FDE and DD-PNC against phase noise.

condition was also plotted (ideal performance). Figure 7 demonstrates that E_b/N_o degradation of the receiver with MMSE-FDE and CP-PNC at PER $= 10^{-2}$ is 4.0 dB from the ideal performance, while one iteration of the proposed receiver with the iterative FDE and DD-PNC can achieve the same performance as the ideal performance, because it can remove the phase noise and obtain the diversity gain by the iterative FDE.

From Fig. 8, it is found that the phase noise severely degrades PER of the receiver with MMSE-FDE and CP-PNC, and that PER cannot reach 10^{-1} even in the high E_b/N_o region. On the other hand, the receiver with the iterative FDE and DD-PNC can achieve PER $= 10^{-2}$ at the $E_b/N_o =$ 16.5 dB by three iterations, and it can reduce the E_b/N_o degradation from the ideal performance to 1.1 dB. With 64QAM, the phase noise prevents the iterative FDE from getting the diversity gain.

C) Iteration performance with different phase noise levels

When the phase noise level is set to −88, −85, and −82 dBc/Hz, average PER performances of the proposed receiver with the iterative FDE and DD-PNC are shown in Fig. 9. The horizontal axis represents the number of iterations, and zero iteration indicates the initial processing. The received E_b/N_o was 17 dB. It can be seen that with −88 and −85 dBc/Hz, the proposed receiver can achieve PER $= 10^{-2}$ by one and three iterations, respectively, while with −82 dBc/Hz, it cannot reach PER $= 10^{-2}$ regardless of the number of iterations. When PER is less than 10^{-2}, the proposed receiver with the iterative FDE and DD-PNC can achieve the better PER performance than the receiver with MMSE-FDE and DD-PNC [19]. It can be also found that with −88 dBc/Hz, two iterations of the proposed receiver with the iterative FDE is superior to the receiver with ideal channel estimation and MMSE-FDE under the phase-noise-free condition.

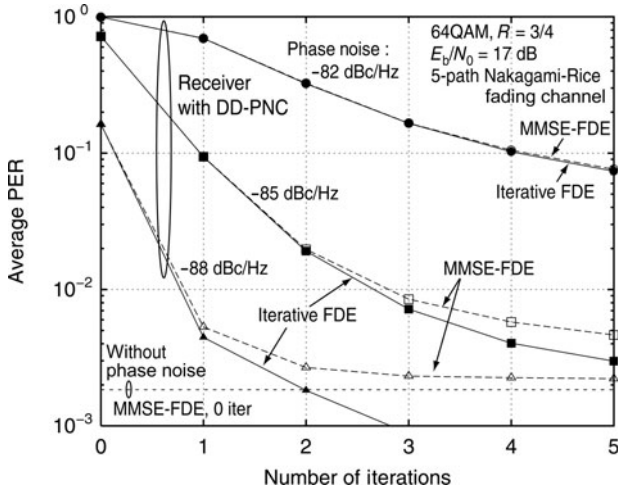


Fig. 9. Iteration performance with different phase noise levels.

D) Effect of DD-PNC on different phase noise levels

When the modulation scheme was 16QAM and 64QAM, average PER performances of the receiver with the iterative FDE or MMSE-FDE and DD-PNC are shown in Fig. 10. The horizontal axis represents the phase noise level at 1 MHz offset. The received E_b/N_o in 16QAM and 64QAM were set to 12 and 17 dB, respectively, and the number of iteration was three. This figure demonstrates that the performance degradation of the receiver with MMSE-FDE and CP-PNC is severe in 16QAM and 64QAM, while the receiver with the iterative FDE and DD-PNC can maintain PER $=$ 10^{-2} with 16QAM and 64QAM until the phase noise level of −80.8 and −85.0 dBc/Hz, respectively.

V. CONCLUSIONS

This paper has proposed the low-complexity SC-FDE receiver employing the iterative FDE and DD-PNC to alleviate the performance degradation due to the phase noise for the millimeter-wave SC systems. The proposed receiver repeats the iterative FDE and DD-PNC by exploiting the output of the channel decoder. In the initial processing,

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CP-PNC estimates the phase noise using the known CP, and it triggers DD-PNC in the iterative processing. DD-PNC estimates the phase noise by the one-tap LMS algorithm with smoothing, and can effectively cope with the phase noise by removing the estimate from the time-domain received signal. In addition, the iterative FDE can obtain the diversity gain by eliminating the residual ISI. The computer simulations have demonstrated that with 64QAM of $R = 3/4$, the phase noise with -85 dBc/Hz at 1 MHz offset drastically degrades PER of the receiver with MMSE-FDE and CP-PNC, while three iterations of the receiver with the iterative FDE and DD-PNC can achieve PER $= 10^{-2}$ at the $E_b/N_o = 16.5$ dB. It has also been shown that with 16QAM, it can realize PER $= 10^{-2}$ at the E_b/N_o of 12 dB until the severe phase noise level of -80.8 dBc/Hz at 1 MHz offset.

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REFERENCES

[1] Merged proposal: new PHY layer and enhancement of MAC for mmWave system proposal, IEEE 802.15-07-0934-01-003c, November 2007.

[2] Proposal for HD AV and data support, IEEE 802.15-07-0942-02-003c, May 2007.

[3] Choi, C.; Shoji, Y.; Harada, H.; Funada, R.: Behavioral model of 60 GHz-band power amplifer for SYS/PHY evaluation, IEEE 802.15-06-0396-01-003c, September 2006.

[4] Kato, S., et al.: Single carrier transmission for multi-gigabit 60-GHz WPAN systems. IEEE J. Sel. Areas Commun., **27** (8) (2009) 1466–1478.

[5] Reynolds, S., et al.: A silicon 60 GHz receiver and transmitter chipset for broadband communications. IEEE J. Solid-State Circuits, **41** (12) (2006), 2820–2830.

[6] Pollet, T.; Bladel, M.V.; Moeneclaey, M.: BER sensitivity of OFDM systems to carrier frequency offset and Wiener phase noise. IEEE Trans. Commun., **43** (2) (1995), 191–193.

[7] Dagres, I.; Polydoros, A.: Decision-directed least-squares phase perturbation compensation in OFDM systems. IEEE Trans. Wireless Commun., **8** (9) (2009), 4784–4796.

[8] Suyama, S.; Suzuki, H.; Fukawa, K.; Izumi, J.: Iterative receiver employing phase noise compensation and channel estimation for millimeter-wave OFDM systems. IEEE J. Sel. Areas Commun., **27** (8) (2009), 1358–1366.

[9] Zamorano, J.L.; Nsenga, J.; Van Thillo, W.; Bourdoux, A.; Horlin, F.: Impact of phase noise on OFDM and SC-CP, in IEEE Global Telecommun. Conf., Washington, DC, November 2007, 3822–3825.

[10] Stark, A.; Raphaeli, D.: Combining decision-feedback equalization and carrier recovery for two-dimensional signal constellations. IEEE Trans. Commun., **55** (10) (2007), 2012.

[11] Ryu, S.B.; Kim, J.; Ryu, H.G.; Li, Y.: PNS algorithm for the SC-FDMA communication system with phase noise, in IEEE and IFIP Inter. Conf. Wireless and Optical Communication Networks (WOCN), Cairo, April 2009, 1–5.

[12] Araujo, T.; Dinis, R.: Iterative equalization and carrier synchronization for single-carrier transmission over severe time-dispersive channels, in IEEE Global Telecommunication. Conf., Dallas, Texas, December 2004, vol. **5**, 3103–3107.

[13] Sabbaghian, M.; Falconer, D.: Joint turbo frequency domain equalization and carrier synchronization. IEEE Trans. Wireless Commun., **7** (1) (2008), 204–212.

[14] Benvenuto, N.; Tomasin, S.: Block iterative DFE for single carrier modulation. IEE Electron Lett., **38** (19) (2002), 1144–1145.

[15] Tuchler, M.; Hagenauer, J.: Turbo equalization using frequency domain equalizers, in Allerton Conf. Communication, Control and Computing, October 2002.

[16] Haykin, S.: Adaptive Filter Theory, 4th ed., Prentice-Hall, 2001.

[17] Kashima, T.; Fukawa, K.; Suzuki, H.: Adaptive MAP receiver via the EM algorithm and message passings for MIMO-OFDM mobile communications. IEEE J. Sel. Areas Commun., **24** (3) (2006), 437–447.

[18] Abe, T.; Fujii, H.; Suda, H.; Tomisato, S.: Performance enhancement of SC/S-MMSE turbo receiver for MIMO-SDM/OFDM transmission, IEICE Trans. Commun., **E88-B** (5) (2005), 2110–2119.

[19] Suyama, S.; Onodera, J.; Suzuki, H.; Fukawa, K.: Decision-directed phase noise compensation for millimeter-wave single carrier transmission systems with frequency-domain equalization, in European Wireless Technology. Conf. (EuWiT), September 2009, 9–11.



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