Abstract—Millimeter-wave (mmWave) systems must overcome severe phase noise impairment to exploit its full advantage (e.g., high data rate). However, the estimation of phase noise combined with wireless channel requires not only high-computational complexity, but also does not guarantee a stable solution. In this paper, the fact that the coherence bandwidth of a mmWave system is much larger than those of conventional systems is exploited to acquire the knowledge regarding phase noise by linear estimator. The channel coefficients are assumed to be piecewise-constant over  $N_{ch}$ successive subcarriers within coherence bandwidth. Utilizing the coherence structure, we propose a new algorithm for phase noise compensation on orthogonal frequency division multiplexing (OFDM) systems. Numerical results with phase noise generated as a Wiener process demonstrate that the proposed algorithm can provide improved performance even in severe phase noise environments.

Index Terms—Coherence bandwidth, mmWave systems, orthogonal frequency division multiplexing (OFDM), phase noise.

I. INTRODUCTION

As a strong contender for next generation, communication at higher-frequency bands opens a new era of wireless connectivity [1]. The range of the frequency bands, from 30 GHz to 300 GHz, is usually referred to as millimeter-wave (mmWave) bands. Practical systems that operate at high-frequency, however, are generally vulnerable to hardware impairments. One of the predominant source of hardware impairment is the phase noise of the local oscillator (LO), which increases with carrier frequency. For example, multiplying the frequency of a signal by a factor of $M$ increases the phase noise of the multiplied signal by $20 \log(M)$ dB [2]. The non-negligible amount of phase noise inevitably leads to significant performance degradation in coherent systems. The combined effects of phase noise and channel, thus, should be estimated and compensated. Unfortunately, this is not a simple problem to solve. There are two fundamental characteristics of phase noise that make the estimation quite difficult in OFDM systems: 1) spectral spreading 2) fast-varying fluctuation.

The implication of phase noise on orthogonal frequency-division multiplexing (OFDM) has been extensively analyzed in the literature [3], [4]. To compensate the effects of phase noise, early studies used quite strong assumptions such small phase noise [5] and perfect channel state information [6] at receiver. Subsequently, joint channel and phase noise estimation [7], iterative joint phase noise estimation and data detection [8] have been presented. However, unrealistic assumptions or overly complex techniques limit the practical use of phase noise compensation.

The problem of dealing with both spectral spreading and fast-varying fluctuation simultaneously is especially challenging in the presence of severe phase noise (e.g., due to increasing the number of phase noise components that should be necessarily estimated). In OFDM systems, effective channel gain is entangled by two unknown variables of phase noise and wireless channel components. For this reason the required estimation problem of effective channel gain is formulated as an underdetermined system which generally has infinitely many solutions. Obtaining an accurate solution is therefore not only guaranteed, but actually requires high-computational complexity, which makes it more difficult to meet the requirement that instantaneous knowledge of phase noise must be updated every OFDM symbol. In an extreme case where the estimation delay of phase noise is longer than an OFDM symbol duration, it is infeasible to perform coherent detection of all data symbols in the corresponding OFDM symbol due to the fact that phase noise is fast-varying.

In this paper, we consider channel coherence in the frequency domain to manage the underdetermined problem on the basis of the fact that the coherence bandwidth of a mmWave system is inherently much larger than those of conventional systems [9]. It is a promising basis for more suitable phase noise compensation in mmWave systems. We show that larger coherence bandwidth can facilitate the estimation of phase noise components scaled by $\alpha$ in the frequency domain as illustrated in Fig. 1. The deconvolution by the scaled phase noise estimates suppresses the effect of ICI by phase noise. In consequence the estimation problem of effective channel gain is transformed into a fully determined system with the help of pilots as the number of different kinds of channel coefficients in the frequency domain.

II. SYSTEM DESCRIPTION

A. System Model

We consider an OFDM system with $N$ subcarriers, $T_s$ sampling period, a subcarrier spacing of $\Delta f$, and a bandwidth
of \( B = 1/T_s = N\Delta f \). Let \( \{X_k\}_{k=0}^{N-1} \) be the transmitted symbol sequence on \( N \) subcarriers of an OFDM symbol with an average power constraint \( \mathbb{E}[|X_k|^2] = E_s \). An \( N \)-point unitary inverse discrete Fourier transform (IDFT) of sequence \( \{X_k\}_{k=0}^{N-1} \) provides the time-domain representation of the OFDM symbol as

\[
x_n = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j2\pi kn/N}, \quad n \in \{0, \ldots, N-1\},
\]

where time index \( n \in \mathbb{Z}^+ \).

The channel is assumed to be block fading with a coherence time of \( T_c \) and a coherence bandwidth of \( B_c \). For our subsequent analysis we adopt the coherence block model that matches realistic channel. In this model, there are two parameters widely used in the literature. The number of OFDM symbols within coherence time \( T_c \), denoted by \( N_{ct} \), is the number of subcarriers within coherence bandwidth \( B_c \), denoted by \( N_{cb} \).

\[
N_{ct} = \left\lfloor \frac{T_c}{T_{sym}} \right\rfloor,
\]

\[
N_{cb} = \left\lfloor \frac{B_c}{\Delta f} \right\rfloor,
\]

where \( T_{sym} \) is the duration of one OFDM symbol. We assume that coherence block spans \( N_{ct} \) OFDM symbols over which the channel impulse response is constant, and \( N_{cb} \) successive subcarriers over which the channel frequency response is constant. Each channel use (consisting of a block of \( N_{ct} \) OFDM symbols and \( N_{cb} \) subcarriers) is independent of each other, and all channel coefficients are drawn from zero-mean circularly symmetric complex Gaussian distributions with unit variance.

B. Phase Noise Model

We consider the model introduced in [11] to illustrate the phase noise of a free-running oscillator. A random time shift \( \eta(t) \) becomes, asymptotically with time, a Wiener process as

\[
\eta(t) = \sqrt{c} W(t),
\]

denotes the two-sided 3-dB linewidth of the Lorentzian power spectral density \(^1[3]\).

C. OFDM Signal Model with Phase Noise

Phase noise at the receiver influences the channel output as an angular multiplicative distortion in the time domain. In view of the duality of the discrete Fourier transform (DFT), the received signal in the frequency domain \( y_f \in \mathbb{C}^{N \times 1} \) is represented as

\[
y_f = \mathbf{p}_f \odot (\mathbf{x}_f \circ \mathbf{h}_f) + \mathbf{z}_f,
\]

where \( \odot \) and \( \circ \) denote the circular convolution and Hadamard product, respectively; \( \mathbf{p}_f = [P_0, P_1, \ldots, P_{N-1}]^T \in \mathbb{C}^{N \times 1} \) is the DFT coefficient vector of the time-domain phase noise sequence \( \{e^{j\phi_n}\}_{n=0}^{N-1} \); \( \mathbf{x}_f \) and \( \mathbf{h}_f \) are the vectors of transmitted symbols \( \{X_k\}_{k=0}^{N-1} \) and channel gains \( \{H_k\}_{k=0}^{N-1} \) on subcarrier \( k \) in the frequency domain, respectively; the vector \( \mathbf{z}_f \in \mathbb{C}^{N \times 1} \) is the additive white Gaussian noise (AWGN) with zero mean and variance \( \sigma_z^2 \). The phase noise component in the frequency domain \( P_i \) is defined as follows:

\[
P_i = \frac{1}{N} \sum_{n=0}^{N-1} e^{j\phi_n} e^{-j2\pi ni/N}.
\]

We rewrite the received signal for each subcarrier \( k \) in the following sample-wise form:

\[
y_k = P_0 H_k X_k + \sum_{\ell=0,\ell \neq k}^{N-1} P_{k-\ell} H_{k-\ell} X_{k-\ell} + Z_k,
\]

where \( P_0 \) indicates common phase error (CPE) and second term is intercarrier interference (ICI). (5) can be also expressed in the following matrix form:

\[
y_f = \Phi_f \mathbf{H}_f \mathbf{x}_f + \mathbf{z}_f,
\]

\[
\Phi_f = \begin{bmatrix}
P_0 & P_{N-1} & P_{N-2} & \cdots & P_1 \\
P_1 & P_0 & P_{N-1} & \cdots & P_2 \\
P_2 & P_3 & P_0 & P_{N-1} & \cdots \\
\cdots & \cdots & \cdots & \cdots & \cdots \\
P_{N-2} & P_{N-3} & \cdots & P_3 & P_0 & P_{N-1} \\
P_{N-1} & P_{N-2} & \cdots & P_3 & P_2 & P_0 \\
\end{bmatrix},
\]

denoted by \( \Phi_f = \text{circ}(\mathbf{p}_f) \), which is a circulant matrix formed by spectral phase noise components \( \{P_i\}_{i=0}^{N-1} \). \( \mathbf{H}_f = \text{diag}\{H_k\}_{k=0}^{N-1} \) is the diagonal matrix form of vector \( \mathbf{h}_f \).

III. PROPOSED METHOD

In this section, we elaborate on an algorithm for phase noise and channel compensation with an example case. In our algorithm, two kinds of estimations are required for: 1) spectral phase noise components scaled by a channel coefficient and 2) spectral wireless channel components as illustrated in Fig. 1.

\(^1\)In phase noise model by Wiener process, the connection between \( \beta \) in the frequency domain and \( c \) in the time domain is described as \( \beta = 2\pi f_c^2c \).
A. Transmission Structure

Before illustrating our algorithm, we first state the transmission structure for our method as follows. Let us define a coherence block $S_t$ with cardinality $|S_t| = N_{cb}N_{ct}$ and $S = S_0 \cup S_1 \cdots \cup S_{M-1}$ be a set of coherence blocks with the cardinality $|S| = M N_{cb} N_{ct}$, where $M$ is the number of coherence blocks, i.e., $M = N/N_{cb}$. With $S$, we also define a resource element as $r_{k,t}$ where $k$ is the subcarrier index and $t$ is the OFDM symbol index, where $k \in \{0,1,\ldots, N-1\}$ and $t \in \{0,\ldots, N_{ct}-1\}$. To describe resource allocation for pilots and transmitting data, we divide $S$ into two subsets, $F_p$ and $F_c$, in the frequency domain; two subsets, $T_p$ and $T_c$, in the time domain, where $S = F_p \cup F_c = T_p \cup T_c$ and $F_p \cap F_c = T_p \cap T_c = \emptyset$. In the frequency domain, $F_p = S_0 \cup S_{M-1}$ is the set of resource elements that includes the pilots to estimate the scaled spectral components of phase noise, while $F_c$ is the set in which the pilots for the scaled channel estimation are allocated. In the time domain, we consider the fact that phase noise process is still fast-varying within coherence time while wireless channel is invariant. Thus, the first OFDM symbol of coherence block, i.e., $T_p = \{r_{k,0}\}_{k=0}^{N-1}$, is the set that both kinds of pilots are allocated; $T_c$ involves only the pilots for scaled spectral components of phase noise. The remainder of coherence block after the allocation of two kinds of pilots is used for transmitting data.

B. Phase-Noise-Affected Channel Estimation

By the use of coherence bandwidth and $N_p$ dominant spectral phase noise components, we can reduce to the determined system from underdetermined system, enabling the estimation of effective channel coefficients included in received signals. Since the set of effective channel components includes spectral phase noise components scaled by a constant channel coefficient which we call phase-noise (PN)-affected channel, it can be used to suppress the ICI by phase noise. The target of this step is to estimate PN-affected channel by using PN-dedicated pilot in $F_p$. Let us denote the frequency-domain effective channel by $F_{t,k} = P_i H_k$ whose value is drawn from a set of multiplications between possible combinations of $P_i$ and $H_k$ for $i \in \{0,1,\ldots,N-1\}$ and $k \in \{0,1,\ldots,M-1\}$. For the sake of brevity, we adopt $\gamma$ as an approximation order of phase noise spectrum, where $N_p = 2\gamma + 1$ for $\gamma \in \{0,1,\ldots,N/2\}$. The indices set of dominant spectral phase noise is defined as $D \triangleq \{0,1,\ldots,N-1\} \backslash \{\gamma + 1, \gamma + 2, \ldots, N - (\gamma + 1)\}$. With $\gamma$-order approximation of frequency-domain effective channel, the approximated channel matrix $F_\gamma$ and the approximation error matrix $E_\gamma$ are given by

$$F_\gamma = \Phi_{t,\gamma} H_f,$$

$$E_\gamma = F - F_\gamma,$$

where $\Phi_{t,\gamma} = \text{circ}(p_{t,\gamma})$ denoted by $p_{t,\gamma} = [P_0, \ldots, P_{\gamma}, 0, \ldots, 0, P_{N-\gamma}, \ldots]^T \in \mathbb{C}^{N \times 1}$ and $F = \Phi f H_f$ is the effective channel matrix without any approximation.

As an example, suppose $N_{cb} = 6$ and $N_p = 3$. Based on (8) and (10), the received signal is described as

$$y_f = F_1 x_f + E_1 x_f + z_f = F_1 x_f + w_1,$$

where $F_1$ is presented in (13), which denotes the 1-order approximated channel matrix and $E_1$ is its approximation error matrix; $w_1$ is the residual ICI beyond the 1-order approximation plus AWGN. Since PN-affected channel is defined as a set of dominant phase noise components scaled by a coherence channel component in the frequency domain. In this example, PN-affected channel vector $f_p$ is

$$f_p = [F_{N-1,0}, F_{0,0}, F_{1,0}]^T \in \mathbb{C}^{N_p \times 1},$$

where $p_{t,1}$ is non-zero original phase noise vector with 1-order approximation. Based on (14), $\alpha$ in Fig. 1 indicates $H_0$. To estimate $f_p$, we particularly construct a fully determined system by PN-dedicated pilot in $F_p \cap T_p$. As the first criterion of PN-dedicated pilot, zero-pilot symbol vector denoted by $x_{t,1,zero}$ is employed for resource elements of both end subcarriers to avoid the ICI from non-adjacent subcarriers, i.e., $x_{t,1,zero} = [X_{r_{0,0}}, X_{r_{N-1,0}}] = [0, 0]$. After applying $x_{t,1,zero}$, the first three output vector $y_{t,1}^p = [Y_0, Y_1, Y_2]^T$ of $y_f$ is

$$y_{t,1}^p = P_{d} x_{t,1,zero} + w_{t,1}^p,$$

where $w_{t,1}^p$ is the first three output vector of $w_1$. Using the commutative property, the effective output $y_{t,1}^p$ of data vector $x_{t,1,zero}$ mapped to the resource element in $F_p \cap T_p$ can be rewritten by

$$y_{t,1}^p = \begin{bmatrix} F_{N-1,0} & 0 & 0 \\ F_{0,0} & F_{N-1,0} & 0 \\ F_{1,0} & F_{0,0} & F_{N-1,0} \end{bmatrix} \begin{bmatrix} X_{r_{1,0}} \\ X_{r_{2,0}} \\ X_{r_{3,0}} \end{bmatrix} + w_{t,1}^p.$$

By the subset of PN-dedicated pilot $\{X_{r_{k,0}}\}_{k=1}^{3}$ such that rank $(X_{r_{k,0}}^H) = 3$, the PN-affected channel vector can be recovered through three observations $\{Y_k\}_{k=0}^{2}$. We consider the linear minimum mean-square error (LMMSE) estimators to obtain the estimate of PN-affected channel vector $f_p$ from $y_{t,1}^p$, i.e.,

$$Q_{\text{LMMSE}} = R_{f y} R_{yy}^{-1} = R_{ff} (X_{r_{k,0}}^H) (X_{r_{k,0}} R_{ff} (X_{r_{k,0}}^H) + \sigma_{w_{t,1}}^2 I)^{-1},$$

where $R_{f y} = E\{f_p (y_{t,1}^p)^H\}$ is the cross-covariance matrix between $f_p$ and $y_{t,1}^p$; $R_{yy} = E\{y_{t,1}^p (y_{t,1}^p)^H\}$ and $R_{ff} = E\{f_p (f_p)^H\}$ are the autocorrelation matrix of $y_{t,1}^p$ and $f_p$, respectively; $\sigma_{w_{t,1}}^2$ denotes the noise variance of $w_{t,1}^p$. The LMMSE estimates of PN-affected channel vector $f_p$ is given by

$$f_{p,\text{LMMSE}} = Q_{\text{LMMSE}} y_{t,1}^p.$$
C. Inter-carrier-Interference Suppression

In general, the ICI brought on by phase noise can be suppressed by the deconvolution between frequency-domain received signals and spectral phase noise estimates \[11\]. In this subsection, we start with a Lemma that provides the idea behind the ICI suppression.

**Lemma 1.** Let \( z \in \mathbb{C}^{N \times 1} \) be the output vector of circular convolution between \( x \in \mathbb{C}^{N \times 1} \) and vector \( y \in \mathbb{C}^{N \times 1} \). Then the deconvolution of \( cz \) from \( z \), where \( c \in \mathbb{C} \) is a scalar, is given by

\[
\begin{align*}
\tilde{z} &= z \otimes^{-1} c = \frac{1}{c} y,
\end{align*}
\]

where \( \otimes^{-1} \) denotes the deconvolution operator.

**Proof.** By the linear property of circular convolution \[12\],

\[
\begin{align*}
z &= x \otimes y = c \cdot (x \otimes y) = cz \otimes y
\end{align*}
\]

\[
(20)
\]

Based on PN-affected channel vector \( f_n \), the vector \( f_p \) with length \( N \) is defined as

\[
\begin{align*}
f_p &\triangleq H_0 f_n, \gamma \in \mathbb{C}^{N \times 1},
\end{align*}
\]

(21)

where \( i \)th element in \( f_p \) is \( F_{i,0} \) for \( i \in D \), and zero for \( i \in D^c \). Since perfect PN-affected channel vector is the spectral phase noise vector scaled by a channel coefficient \( H_0 \) of the first coherence block, the deconvolution of \( f_p \) from \( y_r \), therefore, yields the effective channel \((1/H_0) \mathbf{h}_f \) from Lemma 1. In other words, the Toeplitz convolution matrix \( \mathbf{F}_1 \) in (13) is converted into diagonal matrix \( \mathbf{H}_f = (1/H_0) \mathbf{h}_f \) named ICI-free channel, which means that the off-diagonal elements causing ICI in \( \mathbf{F}_1 \) can be cancelled. \( \mathbf{H}_f \) is represented as

\[
\begin{align*}
\mathbf{H}_f &= \begin{bmatrix}
I_{N_{ch}} & 0 & \cdots & 0 \\
0 & \hat{\mathbf{H}}_f & \cdots & \cdots \\
0 & \cdots & \cdots & \cdots \\
0 & \cdots & \cdots & H_{M-1}^f
\end{bmatrix},
\end{align*}
\]

(22)

where each diagonal element has the value divided by \( H_0 \) from its own channel coefficient, which is defined as \( H_m^f = H_m^f/H_0 \); \( H_m^f \) is \( H_m^f I_{N_{ch}} \in \mathbb{C}^{N_{ch} \times N_{ch}} \) for \( m \in \{1,2,\ldots,M-1\} \) is the diagonal channel matrix with \( H_m^f \), and \( \hat{\mathbf{H}}_f \) is a \( N_{ch} \times N_{ch} \) matrix.

D. Inter-carrier-Interference-Free Channel Estimation

The objective of this step is to estimate the coefficients of ICI-free channel using channel (CH)-dedicated pilot included in \( F_c \cap T_p \). We consider the least-squares (LS) estimators to obtain the estimate of ICI-free channel. Since the coefficient of ICI-free channel in the first coherence block is 1 as shown in (22), the CH-dedicated pilots are allocated as many as \( M-1 \) to estimate the ICI-free channel vector \( \mathbf{h}_f = [H_1^f,\ldots,H_M^f]^T \in \mathbb{C}^{(M-1)\times 1} \). Let \( \mathbf{X}_c^T = \text{diag} \{ X_{c,0}^T \}_{m=1}^{M-1} \in \mathbb{C}^{(M-1)\times (M-1)} \) for \( k = mN_{ch} \) be the matrix form of CH-dedicated pilot. The ICI-free channel estimate vector \( \hat{\mathbf{h}}_f \) becomes the last estimate for decoding \( x_t \). In the time domain, channel is constant over \( N_{ct} \) OFDM symbols while phase noise is changed every OFDM symbol. In the set of resource elements \( T_c \), therefore, only PN-dedicated pilot is allocated in \( F_c \).

IV. Numerical Results

The benefit of proposed method is illustrated with the performance plots of the compensation by perfect knowledge of \( N_p \) significant phase noise components in the frequency domain. For numerical evaluation, the following settings are assumed. The number of subcarriers \( N = 2048 \), a bandwidth of \( B = 20 \text{ MHz} \) are used. Also, we consider two kinds of 3-dB linewidth \( \beta = \{300,3000\} \) Hz as LO parameters, which have both severe phase noise spectrum compared to the phase noise of LO in conventional transceiver. Fig. 2 and 3 shows the bit error rate (BER) performance for an OFDM system transmitting uncoded quadrature phase shift keying (QPSK) obtained by cancellation of different phase noise components, \( N_p \in \{1,3,5,7\} \). The curve in black displays the performance without any phase noise compensation, but the perfect channel compensation is assumed, while the curve in pink shows the performance with perfect channel and phase noise compensation. As illustrated in Fig. 2, the proposed method has quite good BER performance by using the estimation of even only 3 significant phase noise components. When \( N_p \geq 3 \), it is shown that there is around 3-dB gap between perfect \( N_p \)-phase noise compensation and proposed method with same
N_p at a BER of 10^{-3}. We can find that the performance improvements by phase noise compensation more than three are relatively small. It means that most of phase noise energy is focused in three dominant phase noise components. In Fig. 3, more severe phase noise scenario is considered, which means that the energy of phase noise is significantly spread over more number of subcarriers. Thus, it is shown that use of more number of pilots to estimate more number of PN-affected channel components can be helpful for performance improvement in this case.

V. CONCLUSION

This paper outlines a general framework for phase noise compensation on OFDM systems that exploits coherence bandwidth. Our main conclusion is that we can reformulate the joint estimation problem of phase noise and channel into a determined system by the approximation of phase noise spectrum and the channel coherence structure in the frequency domain. This enables to estimate and compensate the dominant phase noise by using pilot-assisted transmission and linear processing. Compared with perfect phase noise compensation, there is only a small BER loss even in a severe phase noise scenario. Our main result shows that mmWave channel coherence structure can be a key to solve phase noise issue of mmWave systems.

ACKNOWLEDGMENT

This research was partly supported by Intel Corporation (SRC MSR-Intel Research Project P28832, Coordination in Distributed Multi-User High-Performance Dense Networks), and partly supported by Basic Science Research Program through the National Research Foundation of Korea (NRF) funded by the Ministry of Education (2017R1A6A3A03012954).

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