Broadband Mm-Wave OFDM Communications in Doubly Selective Channel: Performance Evaluation Using Measured Mm-Wave Channel

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Abstract—In this work, we evaluate the performance of the broadband millimeter-wave (mm-wave) OFDM system in the presence of phase noise (PN) of phase-locked loop based oscillator and delay spread of measured mm-wave channel. It is shown, using Akaike’s information criterion, that the channel tap coefficients of the broadband mm-wave channel do not follow Gaussian distribution due to the broad bandwidth. It is also shown that, given a cyclic prefix (CP) length for a certain delay spread, an effective PN mitigation scheme enables a PN corrupted OFDM system to function with small subcarrier spacing and, therefore, small CP overhead, with only slight degradation of the error rate performance.

Index Terms—Mm-wave, propagation, OFDM, phase noise.

I. INTRODUCTION

Millimeter-wave (mm-wave) communication systems are promising techniques to enable multi-gigabit per second data transmission due to the abundant available bandwidth in the mm-wave bands [1], [2]. For example, there are at least 5 GHz available bandwidth in the 60-GHz band globally [3]. As a result, mm-wave communications have been successfully commercialized at 60 GHz, e.g., IEEE 802.11ad [4]. Recently, the 28-GHz band receives a lot of attention for cellular communications, e.g., [5], [6]. (Strictly speaking, the mm-wave frequency ranges from 30 to 300 GHz. Nevertheless, 28 GHz is usually considered to be mm-wave band as well.)

The channel time-dispersiveness is inevitable due to the broadband transmission in multipath environment at mm-wave frequencies. As a result, the orthogonal frequency division multiplexing (OFDM) technique [7] is often used to cope with the multipath effect. In particular, the OFDM has recently been selected as the main waveform for the fifth-generation (5G) communications below 40 GHz according to the 3GPP 5G standardization [8]. A drawback of the mm-wave OFDM system though is the system can be sensitive to oscillator phase noise (PN) [9], which is much severer at mm-wave frequencies as compared to that in the sub-6 GHz band.

The time-dispersiveness (i.e., frequency-selectivity) due to the multipath propagation and the time variation (i.e., time-selectivity) caused by the PN result in a doubly selective channel. For a fixed sampling duration $T_s$ (bandwidth $1/T_s$) and a cyclic prefix (CP) duration $N_gT_s$ (that should be larger than the channel delay spread), increasing the number of subcarriers $N$ (i.e., reducing the subcarrier spacing) increases the OFDM symbol duration $(N_g + N)T_s$, and, therefore, making the OFDM system more robust to the channel time-dispersiveness. However, increasing the OFDM symbol duration makes the OFDM system more sensitive to time-variation caused by the PN. Another way to interpret this dilemma in the frequency domain is that, by reducing the subcarrier spacing, each subchannel tends to see a flatter channel, yet intercarrier interference (ICI) due to the PN also increases. Larger subcarrier spacing can be used to alleviate the ICI effect. However, given the cyclic prefix (CP) length $N_g$, the CP overhead $N_g/(N_g+N)$ increases with increasing subcarrier spacing (reducing $N$), resulting in a reduced spectral efficiency. Hence, effective PN compensation schemes allowing smaller subcarrier spacing (and, therefore, higher spectral efficiency) are highly desirable. Various PN compensation schemes have been proposed, e.g., [10]-[13]. In this work, a simple common phase error (CPE) correction [10] and an advanced PN mitigation scheme [13] are used for the performance evaluation of the mm-wave OFDM system.

While [10]-[13] assume simple multi-tap fading channel models, a more advanced geometry-based stochastic channel model (i.e., QuaDRiGa [14]) was used in [15]. The current version of QuaDRiGa supports mm-wave channel emulation up to 80 GHz. However, the model becomes less accurate for more than 100 MHz bandwidth. Moreover, it is shown that, due to the sparsity and high time resolution of the broadband mm-wave channel, the common assumption that channel coefficient at each (delay) tap follows Gaussian distribution does not hold any more. Therefore, in this work, we choose to use measured mm-wave channel to conduct performance evaluation. It is shown that effective PN mitigation allows the OFDM system to operate at small subcarrier spacing (and, therefore, small CP overhead) with little degradation of the error rate performance of the system.

Notations: Throughout this paper, $\ast$, $^T$, and $^H$ denote complex conjugate, transpose, and Hermitian operators, respectively. Lowercase bold letter ($\mathbf{x}$) and uppercase bold
letter (X) represent column vector and matrix, respectively. I denotes an identity matrix. diag(x) denotes a diagonal matrix whose diagonal elements are given by x.

\[ J(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \exp(j\phi(n)) \exp\left(-j\frac{2\pi kn}{N}\right). \]  

\( J(0) \) in (3) represents the common phase error (CPE) and the 2nd term in (3) represents the ICI. The CPE can be readily estimated as [10]

\[ \hat{j}(0) = \frac{\sum_{k \in S_p} r(k)H^*(k)s^*(k)}{\sum_{k \in S_p} |H(k)s(k)|^2} \]  

where \( S_p \) denotes the set of scattered pilots within each OFDM symbol. When the PN is large and/or the subcarrier spacing is small, it is not sufficient to correct for the CPE alone.

Here, we briefly present an advanced PN mitigation scheme [13] that correct for both CPE and ICI. Let

\[ y = [y(0), y(1), \ldots, y(N-1)]^T \]  

be the received time-domain OFDM symbol (after CP removal). The PN can be mitigated in the time-domain by \( \Phi y \), where

\[ \Phi = \text{diag}\left[\exp(-j\hat{\phi}(0)), \ldots, \exp(-j\hat{\phi}(N-1))\right]^T \]  

with \( \hat{\phi} \) denoting the estimate of \( \phi \). Hence the task of PN mitigation is essentially PN estimation.

Let \( e \) be an \( N_p \times 1 \) vector consisting of the CTFs at the \( N_p \) pilot subcarriers \( (N_p \leq N) \), \( D \) be an \( N_p \times N \) submatrix of the \( N \times N \) unitary DFT matrix \( F \) corresponding to the \( N_p \) pilot subcarriers. Let \( T \) be an \( N \times q \) interpolation matrix (cf. [13]), such that

\[ \Phi = D \text{diag}(Ta) \]  

where \( a \) consists of \( q \) unknowns or anchors \((q \leq N_p)\), and is given as

\[ a = (D \text{diag}(e)T)^\dagger \text{diag}(e)s_p. \]

Both the CPE correction and the PN mitigation scheme are used in performance evaluations in Section IV.

III. MM-WAVE CHANNEL

Measurements of the mm-wave channel were conducted in a basement (of size 7.81m \( \times \) 7.85m) in a line-of-sight (LOS) scenario. Fig. 2 shows the measurement setup. Vertically polarized wideband biconical antennas were used as the transmit and receive antennas, respectively. The transmit antenna is fixed on top of a table in the right side of the Fig. 2. The receive antenna was moved in a positioning turntable. There were 720 positions (that are uniformly distributed in a circle with a radius of 0.5 m). The distance between the circle center and the transmit antenna is 5 m. At each position, the CTF (from 28 to 30 GHz with 750 frequency points) was measured using a vector network analyzer (VNA) based channel sounder. The CIR is obtained by performing a DFT of the Hamming windowed CTF. (Please refer to [18] for a detailed description of the measurement setup.)

The normalized power delay profile (PDP) of the measured channel (averaged over the 720 spatial samples) is shown in Fig. 3. In order to eliminate the inter-block interference due to...

![Fig. 1. Phase noise model of PLL-based oscillator.](image-url)
the delay spread, the CP length is chosen as 350 time samples (with a sampling frequency of 2 GHz), corresponding to a guard time of 175 ns. This CP length is fixed throughout this paper.

For narrowband channel characterizations, it is usually assumed that the (complex-valued) channel at each delay tap follows Gaussian distribution, i.e., the amplitude of the channel tap follows Rayleigh or Rician distribution. This is because that subpath components from a cluster add up in narrowband (flat-fading) channel, whose sum tends towards a Gaussian distribution due to the centra limit theorem. In broadband mm-wave channel, however, different multipath components becomes separable due to the enhanced time resolution of the wide bandwidth and due to the inherent sparsity of the mm-wave propagation. In this case, the assumption that different subpaths add up to each other does not hold any more. In this paper, we use the Akaike’s information criterion (AIC) [16] to study the amplitude distribution of the channel tap.

The AIC is given as [16]

$$AIC_j = -2 \sum_{n=1}^{M} \ln g_{j\theta}(z_n) + 2p$$

where $\ln$ denotes the natural logarithm, $p$ is the number of parameters of the candidate distribution, $g_{j\theta}$ is the probability density functions (PDF) of the $j$th candidate, and $z_m$ denotes the $m$th sample of the measured $M$ sample. The corresponding maximum likelihood (ML) parameter estimator is [19]

$$\hat{\Theta} = \arg \max_{\Theta} \frac{1}{M} \sum_{n=1}^{M} \ln g_{j\theta}(z_n)$$

where $\Theta$ is the $p \times 1$ parameter vector of the candidate distribution.

The AIC weights can be calculated as [20], [21]

$$w_j = \frac{\exp(\beta_j / 2)}{\sum_{j=1}^{J} \exp(\beta_j / 2)}$$

where $\beta_j = AIC_j - \min\{AIC_j\}$. It represents relative feasibilities of different candidates, ranging from 0 (the worst fit) to 1 (the best fit).

In this work, the most relevant distributions, Rayleigh, Rician, Nakagami, Bessel K, and Weibull distributions are chosen as the candidates. Their PDFs and corresponding parameter ML estimators are given in the sequel separately. For the sake of notational convenience, the subscript $j\theta$ is dropped hereafter.

**Rayleigh:** The PDF of the Rayleigh distribution is given as

$$g(x) = \frac{x}{\sigma^2} \exp\left(-\frac{x^2}{2\sigma^2}\right)$$

where the parameter is $\theta = \sigma$, i.e., $p = 1$. The ML estimator of $\sigma$ is

$$\hat{\sigma} = \sqrt{\frac{1}{2M} \sum_{n=1}^{M} x_n^2}.$$  

**Rician:** The PDF of the Rician distribution is

$$g(x) = \frac{x}{\sigma^2} \exp\left(-\frac{x^2 + \nu^2}{2\sigma^2}\right) I_0\left(\frac{x\nu}{\sigma^2}\right)$$

where $I_0$ is the modified Bessel function of the first kind of zero order and the parameter vector is $\Theta = [\nu \sigma]^T$, i.e., $p = 2$. The ML estimator of $\nu$ is
There is no closed-form ML estimator of \( \sigma^2 \). Thus, a numerical ML estimation (which utilizes the \textit{fminsearch} function in MATLAB) is used instead.

\textbf{Nakagami:} The PDF of the Nakagami distribution is
\[
g(x) = \frac{2}{\Gamma(m_0)} \left( \frac{m_0}{\Omega} \right)^{m_0} x^{2m_0-1} \exp \left( -\frac{m_0 x^2}{\Omega} \right)
\]
where \( \Gamma \) is the gamma function and the parameter vector is \( \Theta = \left[ m_0 \quad \Omega \right]^T \), i.e., \( p = 2 \). The ML estimator of \( \Theta \) and \( m_0 \) are
\[
\hat{\Theta} = \frac{1}{M} \sum_{n=1}^{M} x_n^2 \\
\hat{m}_0 = \left( 2 \ln \hat{\Omega} - \frac{2}{M} \sum_{n=1}^{M} \ln x_n^2 \right)^{-1}
\]
respectively.

\textbf{Weibull:} The PDF of the Weibull distribution is
\[
g(x) = ba^{-b}x^{b-1} \exp \left( -\left( \frac{x}{a} \right)^b \right)
\]
where the parameter vector is \( \Theta = [a \quad b]^T \), i.e., \( p = 2 \). There are no closed-formed ML estimators for the Weibull parameters. Thus, the numerical ML estimator \textit{wblfit} in MATLAB is used instead.

\textbf{Bessel K:} The PDF of the Bessel K distribution is
\[
g(x) = \frac{t^{p+1}}{2^{p+1} \Gamma(P)} x^p K_{P,1}(tx)
\]
where the free parameter vector is \( \Theta = [P \quad l]^T \), i.e., \( p = 2 \), and \( K_{P,1} \) denotes the modified Bessel function of the second kind with the order of \( P-1 \). There are no closed-formed ML estimators for the \textit{Bessel K} parameters. Hence, the numerical ML estimator (based on the \textit{fminsearch} function in MATLAB) is used instead.

Fig. 4 shows the comparison of AIC weights of different candidate distributions for the amplitude of the measured channel. Note that, different spatial channel samples do not have the same LOS component (i.e., LOS components at different positions can have very different phases), thus, the comparison between Rayleigh and Rician may not be fair. Nevertheless, as can be seen, the Weibull and Bessel K distributions outperforms other candidates for most of (delay) taps within 70 ns. (Note that the channel tap becomes small beyond 70 ns (cf. Fig. 3), over-fitting problem (due to the noise) may occurs. Therefore, the AIC results become less reliable beyond 70 ns.) The results imply that, for broadband mm-wave channel, the channel coefficient at each tap are probably not Gaussianly distributed (due to the reasons mentioned before). This finding further motivates the use of measured channel (instead of channel models) for mm-wave OFDM simulations.

\section*{IV. SIMULATIONS}

In this section, we evaluate the performance of the mm-wave OFDM system in doubly selective channel due to PN and delay spread.

For PN emulation, we assume a PN model of a phase-locked loop based oscillator [17], whose power spectral density (PSD) is given in Fig. 5. As shown in the previous section, the bandwidth is 2 GHz and the CP length is set to 350 (time samples). Hence, the CP overhead (i.e., throughput loss) depends on the number of subcarriers or the subcarrier spacing. A small subcarrier spacing (large number of subcarriers) ensure a small CP overhead. From throughput point-of-view, small subcarrier spacing is desired. However, the PN impairment becomes prominent at small subcarrier. Hence, from PN impairment point-of-view, the a larger subcarrier spacing is desirable. To solve this dilemma, effective PN compensation schemes are needed. In this work, we use the simple CPE correction method and an advanced PN mitigation scheme that corrects for both CPE and ICI effects. We assume there are scattered pilots (that are evenly distributed among the subcarriers) in each OFDM symbol. The PN mitigation scheme uses the scattered pilots to estimate the time-domain PN at seven anchor points that are evenly distributed among the time-domain OFDM symbol (after CP removal), PN estimates at other time samples are obtained by linear interpolation (cf. Section II).

We choose every four positions from the 720 positions, that is, we have 180 spatial samples with approximately 1.6-wavelength (1.6-\( \lambda \)) inter-sample distance at the lowest frequency (28 GHz). Each spatial sample is regarded as one (hopefully independent) channel realization.

We assume that the channel stay constant over 40 OFDM symbols, after which a new channel realization is drawn. Note that the measured channel has a fix number of frequency points (over the 2-GHz bandwidth) of 750. In order to use the measured channel for simulation, we first interpolate the measured channel transfer function (using the cubic \textit{spline} interpolation function in MATLAB) so that the frequency points match the subcarriers, then transform the interpolated CTF into (time-domain) CIR using the DFT. The CIR is convolved with the transmitted OFDM symbols to obtain the receive signal. The received signal is further corrupted by additive white Gaussian noise (AWGN) and PN. Note that, for simplicity and without loss of generality, only the PN at the receiver side is considered (cf. Appendix).

To see the effect of the subcarrier spacing on the OFDM system, we assume 1024 and 4096 subcarriers, respectively. The former corresponds to a CP overhead of 350/1024 \( \approx 34\% \), whereas the latter corresponds to a CP overhead of 350/4096 \( \approx 8.5\% \). We further assume that the former has 64 scattered pilots and the latter has 256 scattered pilots so that their overheads due to the scattered pilots are the same. Fig. 6 shows the bit-error-rate (BER) performance of OFDM systems with 1024 and 4096 subcarriers as a function of signal-to-noise ratio (SNR), respectively. As can be seen, with CPE correction or no PN correction at all, the OFDM system with 1024 subcarriers clearly outperforms that with 4096 subcarriers in terms of BER performance (at the cost of 4 times CP overhead). Nevertheless, with effective PN mitigation [13], the OFDM system with 4096 subcarriers is
only slightly worse than that with 1024 subcarriers in terms of BER performance, whereas the former has only 1/4 CP overhead as the latter. Therefore, with effective PN mitigation, mm-wave OFDM can achieve higher throughput with little degradation of the BER performance (at the cost of increased computational complexity).

![Fig. 5. PSD of the PN.](image)

![Fig. 6. BER performances of without phase noise (PN) correction, with common phase error (CPE) correction, with PN mitigations and without PN: (a) 1024 subcarrier; (b) 4096 subcarrier.](image)

V. CONCLUSION

In this work, we evaluate the performance of mm-wave OFDM systems in doubly selective channel using emulated PN of a PLL-based oscillator and measured mm-wave channel. It is shown that the channel tap of broadband mm-wave channel does not follow Gaussian distribution due to the high time resolution of the wide bandwidth and the sparsity of the mm-wave propagation. It is also shown that, for an OFDM system with fixed CP length and bandwidth, reducing the subcarrier spacing reduces the CP overhead, yet making the system more vulnerable to PN. Effective PN mitigation scheme can be used to solve this dilemma, allowing the OFDM system to operate at small subcarrier spacing (and, therefore, small CP overhead) with little degradation of the BER performance.

REFERENCES


