ICI Analysis of MIMO-OFDM Systems with Independent Phase Noise at Both Transmit and Receive Antennas

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***Abstract*—This paper analyzes the influence of independent identical distributed phase noise (PN) at different transmit and receive antennas on the performance of multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) based communication systems. Assuming PN is not severe, we can show that the ICI of the phase noises can be separated into two additive terms, which are induced by transmit PN and receive PN, respectively. The ICI caused by transmit PN remains the same in all cases, while the impact of receiver PN depends on specific scenarios: it decreases when the number of receive antennas increases, and is equal to the impact of transmit PN when the number of receive antennas is equal to that of the transmit antennas under optimal LOS condition (unitary channel), or when the number of receive antennas is two times that of transmit antennas under Rayleigh fading.**

***Index Terms*—phase noise, MIMO, OFDM, ICI**

I. Introduction

High throughput and spectral efficiency are important factors for next generation communication systems. Multiple input multiple output (MIMO) techniques provide significant capacity gain and are applicable to both the dense scattering environment, such as the indoor scenarios or outdoor scenarios in urban areas, and LoS conditions such as fixed broadband wireless access (FWA) [1]. In the meantime, orthogonal frequency-division multiplexing (OFDM) has been widely adopted and implemented in wired and wireless communications for its high spectral efficiency and robustness against intersymbol interference (ISI) caused by multipath fading channels. Combining the above two techniques, MIMO-OFDM arises as a promising system design option in more and more applications [2]. However, MIMO-OFDM inherits the disadvantage of MIMO and OFDM as well, namely the increased transceiver complexity and performance degradation due to the implementation impairments, such as I/Q mismatch, clipping, frequency offset, and phase noise (PN) [3].

Phase noise results from non-ideality in the local oscillators (LO) of the system. It causes the power spectral density (PSD) to exhibit skirts around the carrier frequency. Phase noise in single input single output (SISO) OFDM systems has been studied extensively [4]–[6]. The influence of PN is separated into a multiplicative part, common phase error (CPE), and a additive part, inter-carrier interference (ICI). To suppress it, the standard correction method is to compensate CPE using continuous pilots. In order to further improve the performance, the detector-decoder iteration is proposed in [6] [7].

In MIMO-OFDM systems, most analysis on the impact of phase noise assumed that PN is the same for all transmit (TX) or all receive (RX) antennas [6]–[10]. This assumption is valid when TX or RX antennas share a common LO signal. However, many RF transceiver IC's in the market have built-in PLL's or LO's, which produce different phase noises. Besides, transmitters or receivers may not be co-located in such systems as distributed MIMO [11] and FWA MIMO, where antenna spacing may be enlarged to meters or even tens of meters in order to get better channel condition numbers. In these scenarios, sharing same LO signals is too difficult if not impossible, because of the high cost in distributing the LO signal. [9] extended the analysis to the case with independent transmit phase noises, but still assumed the same phase noises at the receive antennas. Moreover, all the previous works analyzed the impact of transmit and receive phase noises separately.

In this paper, we study the influence of phase noise on MIMO-OFDM systems in more realistic scenarios, where the phase noises can be independent at all the transmit and receive antennas. We derive approximate expressions to calculate the total ICI caused by transmit and receive phase noises and show how these two kinds of phase noises contribute to the overall performance degradation. More specifically, when PN is not severe, we can show that the total ICI is the weighted sum of the variances of transmit PN and receive PN, where the coefficients in the expression is determined by the channel statistics. We also provide closed-form expressions for some scenarios with simple channel statistics. Simulation results demonstrate the accuracy of our analysis.

The rest of this paper is organized as follows. Section II introduces the system model. In section III, the influence of phase noise is analyzed. Section IV presents simulation results. Finally, section V concludes the paper.

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II. System Model

Consider a MIMO-OFDM system with N_t TX antennas, N_r RX antennas, and K subcarriers. The sampled symbols on different TX antennas are multiplied by the independent sampled TX PN in time domain, and then pass through the MIMO channel. In the same way, independent sampled RX PN was multiplied to each sampled symbols on different RX antennas. We use $\vartheta_{nt}(t)$, $\varphi_{nr}(t)$ to represent the tth sampled phase noise on the n_tth TX antenna and n_rth RX antenna, respectively. Then $\theta_{nt}(k)$, the discrete frequency transformation (DFT) spectra are defined as (similar for $\phi_{nr}(k)$)

$$\theta_{nt}(k) = \frac{1}{K}\sum_{t=0}^{K-1} e^{-j\frac{2\pi kt}{K}} e^{j\vartheta_{nt}(t)}$$

Since multiplication in the time domain for discrete time systems is mapped to the circular convolution of DFT spectra in the frequency domain, the phase-noised symbol at the k-th subcarrier can be expressed as circular convolution of the original symbols at all subcarriers with the DFT coefficients of phase noise. Combining the effect of TX PN and RX PN, the received signal at the kth subcarrier is

$$\begin{aligned}\mathbf{r}_k &= \sum_{n=0}^{K-1}\sum_{m=0}^{K-1} \mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-m}\mathbf{s}_m + \mathbf{n}_k \\ &= \underbrace{\sum_{n=0}^{K-1} \mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-k}\mathbf{s}_k}_{\mathrm{CPE}_k} \\ &+ \underbrace{\sum_{n=0}^{K-1}\sum_{m=1, m\neq k}^{K} \mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-m}\mathbf{s}_m}_{\mathrm{ICI}_k} + \mathbf{n}_k, \end{aligned} \tag{1}$$

where $\mathbf{s}_k \in \mathbb{C}^{N_t\times 1}$ and $\mathbf{r}_k \in \mathbb{C}^{N_r\times 1}$ are the TX and RX data vectors at subcarrier k, respectively, we assume no precoding at TX, and the TX symbols on different antennas and different subcarriers are identical independent distributed with unit power and zero mean

$$\mathbb{E}(s_{k1}s_{k2}^H) = \delta(k_1 - k_2)\mathbf{I}_{N_t} \tag{2}$$

where $\delta(k_1 - k_2)$ is 1 if $k_1 = k_2$ and 0 otherwise; $\mathbf{\Theta}_k$ is a diagonal matrix whose diagonal elements are $\theta_{nt}(k)$, $n_t = 1...N_t$, and similar for $\mathbf{\Phi}_k$; $\mathbf{H}_k \in \mathbb{C}^{N_r\times N_t}$ is the equivalent channel matrix at subcarrier k, and $\mathbf{n}_k \in \mathbb{C}^{N_r\times 1}$ is the noise vector distributed as $\mathcal{CN}(\mathbf{0}, \sigma_n^2\mathbf{I}_{N_r})$, and are independent w.s.t PN.

As in (1), the influence of PN can be separated into two terms. The first term causes a time-varying rotation and magnitude variation to the desired signals. Since the distortion is equal for all subcarriers and the magnitude variation is very slight, it is referred to as common phase error (CPE). The other term causes ICI.

It is also assumed that LOs at different TX antennas have similar characteristics, so $\vartheta_{nt}(t)$ have zero-mean identical independent distributions, and their one-sided -3dB bandwidth of Lorentzian spectrum is β_t, $\beta_t \ll \frac{1}{KT}$, with T being the sampling interval. So we have

$$\begin{aligned}\mathbb{E}\left(\mathbf{\Theta}_k\right) &= \mathbb{E}(\theta_{nt}(k))\mathbf{I}_{N_t} \\ \mathbb{E}(\theta_{nt}(k)) &= \delta(k) + o(\delta_t) \\ \mathbb{E}\left(|\theta_{nt}(0)|^2\right) &= 1 - \delta_t, \delta_t \ll 1\end{aligned} \tag{3}$$

and since $\sum_{k=0}^{K} |\theta_{nt}(k)|^2 = 1$ [6], we have $\forall k \neq 0$

$$\mathbb{E}\left(|\theta_{nt}(k)|^2\right) = \rho_k\delta_t + o(\delta_t), \rho_k > 0, \sum_{k=1}^{K}\rho_k^2 = 1 \tag{4}$$

where $o(\cdot)$ means infinitesimal of lower order, and $O(\cdot)$ means infinitesimal of the same order.

For example, if PN is modeled as a sampled Brownian motion which is suitable for free-running open-loop LO [10]

$$\begin{aligned}\vartheta(t) &= \sum_{i=0}^{t}\vartheta'(i), \vartheta'(i) \sim i.i.d\ \mathcal{N}(0, \sigma_\vartheta^2) \\ \delta_t &\approx \frac{K}{6}\sigma_\vartheta^2 = \frac{2\pi\beta_t KT}{3}\end{aligned} \tag{5}$$

Similar assumptions and results hold for the RX PN. Besides we assume that the TX and RX PN are independent because TX and RX are always distant apart. And we define $\delta = \max(\delta_t, \delta_r)$.

From (3) and (4), we have

$$\overline{\mathbf{H}}_k = \sum_{n=0}^{K-1}\mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-k} = \mathbf{\Phi}_0\mathbf{H}_k\mathbf{\Theta}_0 + o(\delta) \tag{6}$$

We assume a quasi-static fading channel (channel is constant for several OFDM symbols) and channel state information (CSI) can be obtained by using the preamble. Since CPE varies from symbol to symbol due to the random phase noise, phase noise estimation method has to be implemented for each data symbol using known pilots. Like previous works [6]–[10], we assume that the CPE (6) can be perfectly estimated at RX. For the remainder of the analysis, we only focus our analysis on ICI.

Note that when PN at all RX antennas are assumed to be the same, $\mathbf{\Phi}_k$ is just a scaled identity matrix, since the scaler can be extracted from the matrix multiplication in (1), RX PN produce same influence as TX PN. Therefore replacing multiple LOs at RX by one LO will bring much performance gain, while doing that at TX changes nothing of average ICI power. Note that when PN are independent at different antennas, the rotations taken place at each TX-RX antenna pair are also independent, leaving $N_t + N_r - 1$ parameters to be estimated, which means more pilots or more power allocation are needed to give a good enough estimation.

III. SNR Degradation Induced by Phase Noise

In this section, we will derive the closed-form expression to approximate the SNR degradation caused by phase noise. Assuming perfect channel state information (CSI) and linear equalization at the receiver, the estimated data vector can be written as

$$\begin{aligned}\hat{\mathbf{s}}_k &= \mathbf{G}_k \sum_{n=1}^{K} \mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-k}\mathbf{s}_k + \mathbf{G}_k\mathbf{n}_k \\ &+ \mathbf{G}_k \sum_{n=1}^{K}\sum_{m=1,m\neq k}^{K} \mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-m}\mathbf{s}_m, \end{aligned} \tag{7}$$

where $\mathbf{G}_k$ is the linear equalizer. There are usually two kinds of linear equalizers, namely the zero-forcing (ZF) equalizer and the MMSE equalizer. Since in this paper we are more interested in the high SNR and small PN regime, ZF and MMSE equalizer result in almost the same performance. Due to the simplicity of ZF equalizer, we assume that ZF equalizer is used from now on.

The ZF equalizer $\mathbf{G}_k$ can be written as

$$\mathbf{G}_k = \overline{\mathbf{H}}_k^{\dagger} \tag{8}$$

where $\overline{\mathbf{H}}_k^{\dagger}$ is the pseudo-inverse matrix of $\overline{\mathbf{H}}_k$ defined in (6). Then the mean square error (MSE) of the estimated data vector can be expressed as

$$\begin{aligned}\mathrm{MSE}_k &= \mathbb{E}\Bigg(\|(\overline{\mathbf{H}}_k^{\dagger}\sum_{n=1}^{K}\mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-k}-\mathbf{I})\mathbf{s}_k + \overline{\mathbf{H}}_k^{\dagger}\mathbf{n}_k \\ &+ \overline{\mathbf{H}}_k^{\dagger}\sum_{n=1}^{K}\sum_{m=1,m\neq k}^{K}\mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-m}\mathbf{s}_m\|^2\Bigg) \end{aligned} \tag{9}$$

$$\begin{aligned} &= \mathbb{E}\left(\|\underbrace{(\overline{\mathbf{H}}_k^{\dagger}\sum_{n=1}^{K}\mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-k}-\mathbf{I})\mathbf{s}_k}_{\mathrm{CPE}_k}\|^2\right) \\ &+ \mathbb{E}\left(\|\underbrace{\overline{\mathbf{H}}_k^{\dagger}\sum_{n=1}^{K}\sum_{m=1,m\neq k}^{K}\mathbf{\Phi}_{k-n}\mathbf{H}_n\mathbf{\Theta}_{n-m}\mathbf{s}_m}_{\mathrm{ICI}_k}\|^2\right) \\ &+ \mathbb{E}\left(\|\overline{\mathbf{H}}_k^{\dagger}\mathbf{n}_k\|^2\right). \end{aligned} \tag{10}$$

As mentioned above, the influence of phase noise comes from two aspects, namely the CPE term CPE_k and the ICI term ICI_k. Since we assume that the CPE can be eliminated, we focus on analyzing the average MSE of the estimated data vector caused by ICI, which is given in the following theorem.

Theorem 1: Assuming that the variances of transmit and receive phase noises satisfy $\delta_t \ll 1$ and $\delta_r \ll 1$, and defining $\delta = \max(\delta_t, \delta_r)$, we can calculate the average MSE of each data vector caused by ICI as follows.

$$\begin{aligned}&\mathbb{E}\left(\|\mathrm{ICI}_k\|^2\right) \\ =& \sum_{n\neq k} tr\left(\mathbb{E}\|\phi(k-n)\|^2\overline{\mathbf{H}}_k^{\dagger} diag\left(\overline{\mathbf{H}}_n\overline{\mathbf{H}}_n^{H}\right)\overline{\mathbf{H}}_k^{\dagger H}\right) \\ +& \delta_t N_t + o(\delta)\end{aligned} \tag{11}$$

When the channel is flat fading, we have

$$\begin{aligned}&\mathbb{E}\left(\|\mathrm{ICI}\|^2\right) \\ =& \delta_t N_t + \delta_r \mathbb{E} tr\left(\mathbf{H}^{\dagger} diag(\mathbf{H}\mathbf{H}^{H})\mathbf{H}^{\dagger H}\right) + o(\delta),\end{aligned} \tag{12}$$

where we drop the subscript k, because the frequency-domain channels are the same. We also define $diag(\mathbf{X})$ as a diagonal matrix with the same diagonal elements as $\mathbf{X}$.

When the frequency-domain channels are i.i.d., we have

$$\begin{aligned}&\mathbb{E}\left(\|\mathrm{ICI}_k\|^2\right) \\ =& \delta_t N_t + \delta_r tr\left(\mathbb{E}\left(\mathbf{H}^{\dagger H}\mathbf{H}^{\dagger}\right)\mathbb{E} diag\left(\mathbf{H}\mathbf{H}^{H}\right)\right) + o(\delta),\end{aligned} \tag{13}$$

where we also drop the subscript k, because the frequency-domain channels have the same statistics.

Proof: See Appendix I. ■

From Theorem 1, we can see that the ICI of the phase noises can be separated into two additive components, namely the ICI induced by transmit PN and that induced by receive PN, respectively. Furthermore, if we know enough channel statistics, we can obtain closed-form expression for the MSE induced by ICI.

Corollary 1: If the channel has only one line-of-sight path and is carefully designed to be unitary such that $N_t = N_r$ and $\mathbf{H}_n\mathbf{H}_n^H = \mathbf{I}_{N_r}, \forall n$ [1], the MSE induced by ICI is

$$N_t \cdot \delta_t + N_t \cdot \delta_r. \tag{14}$$

Corollary 2: If the frequency-domain channel matrices are independent and each element of the channel matrices is i.i.d. Rayleigh fading with variance $1/N_t$, the MSE induced by ICI is

$$N_t \cdot \delta_t + \frac{N_t^2}{N_r - N_t} \cdot \delta_r. \tag{15}$$

Corollary 3: If the channel is flat Rayleigh fading, the variance of each element of the channel matrices is $1/N_t$, and the receive phase noises on different antennas are the same, the MSE induced by ICI is

$$N_t \cdot \delta_t + N_t \cdot \delta_r. \tag{16}$$

Note that in this case, our result is the same as that in [10].

To evaluate the total MSE of the kth data vector, we have to calculate the MSE caused by additive noise $\mathbf{n}_k$, which can be easily derived as

$$\mathbb{E}\left(\|\overline{\mathbf{H}}_k^{\dagger}\mathbf{n}_k\|^2\right) = tr\left((\overline{\mathbf{H}}_k^{H}\overline{\mathbf{H}}_k)^{\dagger}\right)\cdot\sigma_n^2. \tag{17}$$

IV. Simulation Result

In this section, we perform Monte-Carlo simulations to demonstrate the accuracy of the ICI analysis in section III. In the simulations, we consider a MIMO-OFDM system with $K = 64$ subcarriers and 64QAM modulation. There are 2 antennas at both the transmitter and the receiver.

The phase noise at transmit and receive antennas is i.i.d. with distribution in (5), where we define $\sigma_\vartheta^2 = \sigma_\varphi^2 = \delta_{\mathrm{samp}}$. We assume that CPE is perfectly corrected.

The channel is flat fading and is fixed at different samples. Since we assume that CPE caused by the transmit and receive

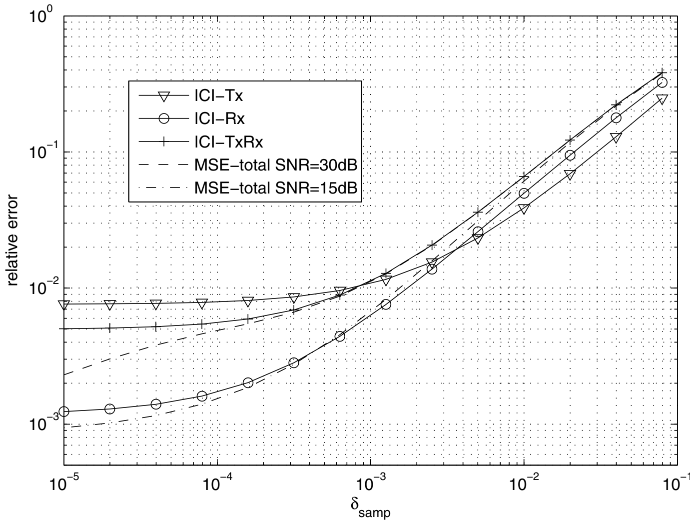

Fig. 1. Comparison of the analytical and simulation results

PN's are perfectly estimated, we can write the channel matrix multiplied by transmit and receive PN matrices as

$$\overline{\mathbf{H}} = \begin{bmatrix} -0.7446 - 0.4168j & -0.3154 - 1.0567j \\ 1.3320 - 0.7176j & 0.0322 + 0.7099j \end{bmatrix}. \quad (18)$$

Under different δ_{samp}, we calculate the relative error between the ICI obtained from (12) and ICI calculated by simulation. In the presence of additive noise, we also calculate the relative error of the total MSE obtained from (12) and (17) and that calculated by simulation. The results are shown in Fig. 1. We can see that the analytical result is relatively accurate when the phase noise is not severe, namely $\delta_{\text{samp}} \leq 10^{-3}$.

V. Conclusion

This paper analyzes the influence of independent identical distributed phase noise at different transmit and receive antennas in MIMO-OFDM based communication systems. We assume that PN is not severe, in other words, the one-sided -3 dB bandwidth of the corresponding Lorentzian spectrum is small compared to the subcarrier spacing. We show that the ICI of the phase noises can be separated into two additive terms, which are induced by transmit PN and receive PN, respectively. The ICI caused by transmit PN remains the same in all cases, while the impact of receiver PN depends on the antenna configuration and channel statistics. In general, ICI induced by receive PN decreases with increasing N_r. When optimal LOS condition is satisfied, which means that the channel matrix is unitary and $N_r = N_t$, the ICI induced by receive PN equals that of transmit PN. When the channel is Rayleigh fading and $N_r = 2N_t$, ICI induced by receive PN equals that of transmit PN.

Appendix A
Proof of Theorem 1

$$
\begin{aligned}
& \mathbb{E}\sum_{n_1,n_2}\sum_{m_1,m_2\neq k} tr\Big(\overline{\mathbf{H}}_k^{\dagger}\mathbf{\Phi}_{k-n_1}\mathbf{H}_{n_1}\mathbf{\Theta}_{n_1-m_1}\mathbf{s}_{m_1} \\
& \qquad \cdot \mathbf{s}_{m_2}^H\mathbf{\Theta}_{n_2-m_2}^H\mathbf{H}_{n_2}^H\mathbf{\Phi}_{k-n_2}^H\overline{\mathbf{H}}_k^{\dagger H}\Big) \\
\overset{(2)}{=}\; & \mathbb{E}\sum_{n_1,n_2}\sum_{m\neq k} tr\Big(\overline{\mathbf{H}}_k^{\dagger}\mathbf{\Phi}_{k-n_1}\mathbf{H}_{n_1}\mathbf{\Theta}_{n_1-m} \qquad (19) \\
& \qquad \cdot \mathbf{\Theta}_{n_2-m}^H\mathbf{H}_{n_2}^H\mathbf{\Phi}_{k-n_2}^H\overline{\mathbf{H}}_k^{\dagger H}\Big) \\
\overset{(3)}{=}\; & \sum_{n_1,n_2}\sum_{m\neq k} tr\Big(\mathbb{E}(\theta(n_1-m)\theta(n_2-m)^*) \\
& \qquad \cdot \mathbb{E}\Big(\overline{\mathbf{H}}_k^{\dagger}\mathbf{\Phi}_{k-n_1}\mathbf{H}_{n_1}\mathbf{H}_{n_2}^H\mathbf{\Phi}_{k-n_2}^H\overline{\mathbf{H}}_k^{\dagger H}\Big)\Big) \\
\overset{(3)}{=}\; & \sum_{n}\sum_{m\neq k} tr\Big(\mathbb{E}\|\theta(n-m)\|^2 \\
& \cdot\mathbb{E}\Big(\overline{\mathbf{H}}_k^{\dagger}\mathbf{\Phi}_{k-n}\mathbf{H}_{n}\mathbf{H}_{n}^H\mathbf{\Phi}_{k-n}^H\overline{\mathbf{H}}_k^{\dagger H}\Big)\Big) + o(\delta) \\
\overset{(4)}{=}\; & \sum_{n\neq k} tr\Big(\mathbb{E}\|\phi(k-n)\|^2\overline{\mathbf{H}}_k^{\dagger} diag\Big(\mathbf{H}_n\mathbf{H}_n^H\Big)\overline{\mathbf{H}}_k^{\dagger H}\Big) + o(\delta) \\
& + \sum_{m\neq k} tr\Big(\mathbb{E}\|\theta(k-m)\|^2\overline{\mathbf{H}}_k^{\dagger}\mathbf{\Phi}_0\mathbf{H}_k\mathbf{H}_k^H\mathbf{\Phi}_0^H\overline{\mathbf{H}}_k^{\dagger H}\Big) \\
\overset{(6)}{=}\; & \sum_{n\neq k} tr\left(\mathbb{E}\|\phi(k-n)\|^2\overline{\mathbf{H}}_k^{\dagger} diag\left(\overline{\mathbf{H}}_n\overline{\mathbf{H}}_n^H\right)\overline{\mathbf{H}}_k^{\dagger H}\right) \\
& + \delta_t N_t + o(\delta)
\end{aligned}
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