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Joint Channel and Phase Noise Estimation for Self-Interference Cancelation in Full-Duplex MIMO-OFDM

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ABSTRACT: *Self-interference (SI) and oscillator phase noise (PHN) can adversely impact the performance of the multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) full-duplex wireless system. In this paper, we propose the algorithm which joint channel and phase noise estimation based on Kalman filter (KF) and maximum-likehood (ML) function. Next, we present the self-interference cancelation in digital domain and ML data detection. The analytical and empirical results showed that the proposed algorithm can offer good performance in the presence of phase noise.*

KEYWORDS - *Full-duplex, phase noise estimation, channel estimation, self-interference, Kalman filter, data detection*

I. INTRODUCTION

The prompt increase in demand for data traffic has made system capacity enhancement one of the most important features in next generation wireless communication systems. Key technologies for future wireless communication systems will make it possible to achieve capacity enhancement by increased spectral efficiency, decreased latency. Full-duplex transmission was introduced as a promising candidate that could potentially double the spectral efficiency of wireless systems comparing to the existing half-duplex traditional systems in ideal condition [1]. The main disturbing factor is the strong self-interference signal imposed by the transmit antenna on the receive antenna within the same transceiver. Hence, full-duplex systems must be able to mitigate the self-interference signal to be lower than the receiver noise floor in order to take advantage of simultaneous transmissions and receptions [2], [3].

The other limitation factor in MIMO-OFDM full-duplex system is phase noise (PHN) which arises because of imperfections in the carrier frequency synthesizer. Hence, phase noise tracking and compensate are the main design targets to design the full-duplex system with high synchronism accuracy. Generally, the presence of phase noise in full-duplex systems introduces inter-carrier interference (ICI) and common phase error (CPE) which rotation of the signal constellation. Characteristics of phase noise depend on the type of frequency generator that is being used (phase locked loop, free-running oscillators). The knowledge of the characteristics can play an important role in estimation and compensation in the presence of phase noise [4], [5], [6].

The channel estimation for residual self-interference cancellation at the baseband in a full-duplex transceiver are thoroughly analyzed in [7] and [8]. It's presented a maximum-likelihood (ML) algorithm to jointly estimate both intended channel the residual self-interference channel under block-fading channel and using the self-interference channel to cancel the self-interference signal before decode. In [9], the ML pilot-aided channel estimation algorithm under the fast-fading by using the basis expansion models (BEMs) to reduce the number of fast-fading channel parameters. However, these results are not applicable to full-duplex systems, where a received signal may be affected by phase noise.

In this work, an algorithm which joints phase noise and channel estimation in MIMO-OFDM full-duplex wireless systems equipped with N_t transmit and N_r receive antennas is analyzed. The algorithm which joint Kalman tracking phase noise and ML pilot-aided channel estimation is proposed. Next, self-interference signal is suppressed from received data, detected by log-likelihood function. The free-running oscillators phase noise model and quasi-static and frequency-flat fading channels are assumed in this paper.

The remaining of the paper is organized as follows. In Section 2, we describe the system model for MIMO-OFDM full-duplex wireless communication system and phase noise tracking by Kalman filter. Detail of the ML channel estimation algorithm by using pilot symbols, self-interference cancelation and data detection schemes are presented in Section 3. The simulation results are given in Section 4 and the conclusions are shown in Section 5.

Notations: $(\mathbf{X})^T$ and $(\mathbf{X})^H$ denote the transpose and conjugate transpose (Hermitian operator) of the matrix $\mathbf{X}$, respectively. $[\mathbf{X}]_{i,j}$ is the $(i;j)$ th entry of the matrix **X.** $\mathbf{R_X}$ denotes the autocorrelation $\mathbf{E(xx^H)}$. $|\mathbf{X}|$ and $\|\mathbf{X}\|$

are the determinant and Frobenius norm of the matrix **X**. diag(**x**) is used to denote a diagonal matrix, where the diagonal elements are given by vector **x.** $N(\mu,\sigma^2)$ denotes the Gaussian distributions with mean μ and variance σ^2, respectively. Finally, if x denotes a parameter of an intended channel then $\dot{x}$ stands for the corresponding parameter of a self-interference channel

II. SYSTEM MODEL

We consider the point to point MIMO-OFDM full-duplex system with N_t transmit antennas and N_r receive antennas as shown in Fig. 1. The system allows simultaneous transmission and reception over the same frequency slot. In the considered system, the block-fading channels between two nodes (node A and node B) is assumed. For an OFDM transmitted signal, with $n \in \{0, \dots, N-1\}$, the inverse fast Fourier transform (IFFT) block is used to transform the sequence of symbols $\{X_{k,m}^{(t)}\}$ of length N into the time domain signal:

$$x_{n,m}^{(t)} = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{k,m}^{(t)} e^{\frac{j2\pi kn}{N}}, \tag{1}$$

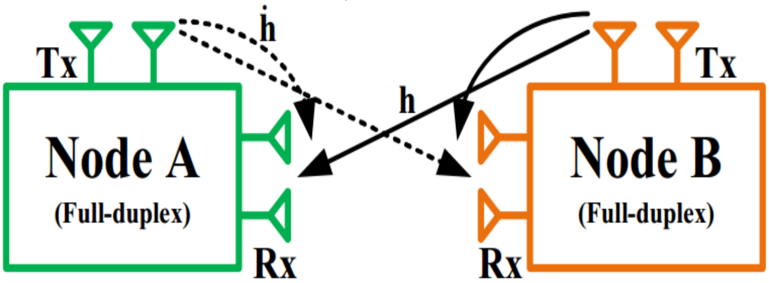

Fig. 1: The considered model of a wireless MIMO-OFDM full-duplex system.

The n-th received in the m-th OFDM symbol at the r-th receive antenna of node without phase noise (after removing cyclic prefix) can be given by:

$$y_{n,m}^{(r)} = \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t)} + \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} \dot{h}_{l,n,m}^{(r,t)} \dot{x}_{n-l,m}^{(t)} + z_{n,m}^{(r)}, \tag{2}$$

where $h_{l,n,m}^{(r,t)}$, $l=0, 1\dots L$ is the L-tap impulse response of the intended channel from the n-th time instance in the m-th OFDM symbol from the t-th transmit antenna of the transmitter (node B) to the r-th receive antenna of the receiver (node A), $\dot{h}_{l,n,m}^{(r,t)}$, $l=0, 1\dots L$ is the L-tap impulse response of the intended channel from the n-th time instance in the m-th OFDM symbol from the t-th antenna to the r-th antenna in same tranceiver (node A). $\dot{x}_{n,m}^{(t)}$, $x_{n,m}^{(t)}$ are the self-interference and intended data sequences, respectively, extended by a cyclic prefix of length N_g and $z_{n,m}^{(r)}$ is the additive white Gaussian noise (AWGN), modeled as a sequence of stationary, complex valued and zero mean Gaussian samples.

Assuming transmitter and receiver share local oscillator (there is not phase noise in self-interference channel) and phase noise only affect in intended signal, the equation (2) can be written as:

$$y_{n,m}^{(r)} = e^{j\theta_n} \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t)} + \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} \dot{h}_{l,n,m}^{(r,t)} \dot{x}_{n-l,m}^{(t)} + z_{n,m}^{(r)}, \tag{3}$$

where θ_n denotes the random phase noise which is modeled as a Brownian process.

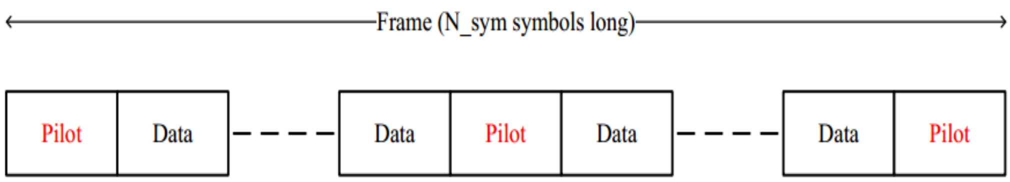

Fig. 2: The timing diagram for transmission of pilot and data symbols within a frame.

As show in the Fig. 2, each frame of length *N_sym* symbols is assumed to consist of pilot of length *N_pilot_sym* symbols which are inserted in some sub-carriers and data symbols. With the optimize position and percentage of the pilot in the frame, the transmitted signal consists of two signals:

$$x_{n,m}^{(t),P} = \frac{1}{\sqrt{N}} \sum_{i=0}^{P} X_{p_i,m}^{(t)} e^{\frac{j2\pi p_i n}{N}}, \tag{4}$$

$$x_{n,m}^{(t),D} = \frac{1}{\sqrt{N}} \sum_{k \neq P} X_{k,m}^{(t)} e^{\frac{j2\pi kn}{N}}, \tag{5}$$

where the sequence $x_{n,m}^{(t),\mathrm{P}}$ and $x_{n,m}^{(t),\mathrm{D}}$ contain the pilot symbols and the data symbol in the transmitted signal from the transmitter to receiver. Insert the equation (4) and (5) into (3), we have:

$$y_{n,m}^{(r)} = e^{j\theta_n} \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} (h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{P}} + h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{D}}) + \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} \dot{h}_{l,n,m}^{(r,t)} \dot{x}_{n-l,m}^{(t)} + z_{n,m}^{(r)}, \tag{6}$$

II.1. Effect of Phase Noise

Phase noise θ_n, denotes rapid, short-term, random fluctuations; generated at both transmitter and receiver imperfect oscillators. Characteristics of phase noise depend on the type of frequency generator that is being used. In this paper, we consider the free-running for phase noise modeling, can be described as a continuous Brownian motion, given by:

$$\theta_n = \theta(nT_s), \tag{7}$$

At the results, the phase noise at the *n*-th sample is related to the previous one as:

$$\theta_n = \theta_{n-1} + \alpha_n, \tag{8}$$

where α is a Gaussian random variable with zero mean and variance $\sigma^2 = 2\pi\beta T/N$. In this notation *T* describes the sample interval and β is an oscillator dependent parameter that determines its quality. The oscillator parameter β is related to the 3dB bandwidth f_{3dB} of the phase noise Lorentzian power density spectrum of the free-running carrier generator.The autocorrelation function of θ_n can be given by:

$$E\left[e^{j\theta_n} e^{-j\theta_{n'}}\right] = e^{-\pi\beta T|n-n'|/N}, \tag{9}$$

II.2. Estimation Phase noise by Kalman Filter

This section presents the Kalman filter to track phase noise parameters θ_n. Using (8), unknown state value is given by:

$$\theta_n = \mathbf{A}\theta_{n-1} + \alpha_n, \tag{10}$$

where θ_n is the phase noise at the *n*-th sample and θ_{n-1} is the previous, **A** are the identity matrix; α_n is the noise process, distributed as $p(\alpha) \sim N(0, Q_\alpha)$ with covariance $Q_\alpha = 2\pi\beta T / N$.

The observation equation for the Kalman filter is given by:

$$y_n = \mathbf{H}\theta_n + w_n, \tag{11}$$

where w_n is the additive white Gaussian noise (AWGN) of observation, where $p(w) \sim N(0, Q_w)$ and **H** are the identity matrix.

Using (10) and (11), the Kalman equations can be written as [10]:

$$\hat{\theta}_{n|n-1} = \mathbf{A}\hat{\theta}_{n-1|n-1}, \tag{13}$$

$$\hat{P}_{n|n-1} = \mathbf{A}P_{n-1|n-1}\mathbf{A}^T + Q_\alpha, \tag{14}$$

$$K_n = \hat{P}_{n|n-1}\mathbf{H}^T(Q_w + \mathbf{H}\hat{P}_{n|n-1}\mathbf{H}^T)^{-1}, \tag{15}$$

$$\hat{\theta}_{n|n} = \hat{\theta}_{n|n-1} + K_n(y_n - \mathbf{H}\hat{\theta}_{n|n-1}), \tag{16}$$

$$P_{n|n} = (1 - K_n)\hat{P}_{n|n-1}, \tag{17}$$

where $\hat{\theta}_{n|n-1}$ and $\hat{\theta}_{n|n}$ are the predicted and update state value at the *n-th* symbol, $\hat{P}_{n|n-1}$ and $P_{n|n}$ are the predicted and update filtering error covariance, K_n is Kalman gain. In order to ensure convergence of the filter, the initial values ($\hat{\theta}_{n-1|n-1}$ and $\hat{P}_{n-1|n-1}$) are chosen by $\hat{\theta}_{n-1|n-1} = \mu_\theta$ and $\hat{P}_{n-1|n-1} = Q_\theta$.

III. CHANNEL ESTIMATION, SELF-INTERFERENCE CANCELATION AND DATA DETECTION

The Fig.3 shows the diagram of joint tracking phase noise and pilot-aided channel estimation, self-interference cancelation and data detection in MIMO-OFDM full-duplex tranceiver.

First, the estimation phase noise $\hat{\theta}_n$, tracked by Kalman filter is used to compensate for affecting received signal. The *n*-th received sample in the *m*-th OFDM symbol at the *r*-th receive antenna of receiver after compensated phase noise can be written as:

$$y_{n,m}^{(r)} = e^{j\theta_n} e^{-j\hat{\theta}_n} \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} (h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{P}} + h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{D}}) + e^{-j\hat{\theta}_n} \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} \dot{h}_{l,n,m}^{(r,t)} \dot{x}_{n-l,m}^{(t)} + e^{-j\hat{\theta}_n} z_{n,m}^{(r)}, \tag{18}$$

The equation (18) can be written as the equivalent equation:

$$\tilde{y}_{n,m}^{(r)} = \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} (h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{P}} + h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),\mathrm{D}}) + \sum_{t=1}^{N_t} \sum_{l=0}^{L-1} \dot{h}'^{(r,t)}_{l,n,m} \dot{x}_{n-l,m}^{(t)} + z'^{(r)}_{n,m}, \tag{19}$$

Then, extracting the pilot in the transmitted signal, the received pilot is given as follows:

$$\mathbf{y}^P = [\dot{\mathbf{S}}\ \mathbf{S}^P]\begin{bmatrix}\dot{\mathbf{h}}' \\ \mathbf{h}\end{bmatrix} + \mathbf{z}' = \mathbf{D}\mathbf{a} + \mathbf{z}', \tag{20}$$

where $\mathbf{y}^P = \left[\mathbf{y}_{m1}^T, ..., \mathbf{y}_{m_P}^T\right]$, $\mathbf{y}_{m_P} = \left[\left[\mathbf{y}_{m_P}^{(1)}\right]^T, ..., \left[\mathbf{y}_{m_P}^{(N_r)}\right]^T\right]^T$, $\mathbf{y}_{m_P}^{(r)} = \left[\mathbf{y}_{0,m_P}^{(r)}, ..., \mathbf{y}_{N-1,m_P}^{(r)}\right]^T$, $\mathbf{D}=[\dot{\mathbf{S}}\ \mathbf{S}^P]$,

$\dot{\mathbf{S}} = \left[\dot{\mathbf{S}}_{m_1}^T, ..., \dot{\mathbf{S}}_{m_P}^T\right]^T$, $\dot{\mathbf{S}}_{m_P} = \left[\dot{\mathbf{S}}_{m_P}^{(1)}, ..., \dot{\mathbf{S}}_{m_P}^{(N_t)}\right]^T$, $\dot{\mathbf{S}}_{m_P}^{(t)} = \left[\dot{\mathbf{s}}_{0,m_P}^{(t)}, ..., \dot{\mathbf{s}}_{L-1,m_P}^{(t)}\right]$,

$\dot{\mathbf{s}}_{l,m_P}^{(t)} = diag(\left[\dot{x}_{0-l,m_P}^{(t)}, ..., \dot{x}_{N-1-l,m_P}^{(t)}\right])$, $\mathbf{S}^P = \left[\mathbf{S}_{m_1}^T, ..., \mathbf{S}_{m_P}^T\right]^T$, $\mathbf{S}_{m_P} = \left[\mathbf{S}_{m_P}^{(1)}, ..., \mathbf{S}_{m_P}^{(N_t)}\right]^T$,

$\mathbf{s}_{l,m_P}^{(t)} = diag(\left[x_{0-l,m_P}^{(t)}, ..., x_{N-1-l,m_P}^{(t)}\right])$, $\mathbf{a} = \left[\dot{\mathbf{h}}'^T, \mathbf{h}^T\right]^T$, $\dot{\mathbf{h}}' = \left[\left[\dot{\mathbf{h}}'^{(1)}\right]^T, ..., \left[\dot{\mathbf{h}}'^{(N_r)}\right]^T\right]^T$,

$\dot{\mathbf{h}}'^{(r)} = \left[\left[\dot{\mathbf{h}}'^{(r,1)}\right]^T, ..., \left[\dot{\mathbf{h}}'^{(r,N_t)}\right]^T\right]^T$ $\mathbf{h} = \left[\left[\mathbf{h}^{(1)}\right]^T, ..., \left[\mathbf{h}^{(N_r)}\right]^T\right]^T$, $\mathbf{h}^{(r)} = \left[\left[\mathbf{h}^{(r,1)}\right]^T, ..., \left[\mathbf{h}^{(r,N_t)}\right]^T\right]^T$.

The ML-based estimates of channel response can be determined as follows:

$$\hat{\mathbf{a}}_{ML} = \arg\max_{a} \ln p(\mathbf{y}^P \mid \mathbf{a}), \tag{21}$$

with $p(\mathbf{y}^P \mid \mathbf{a})$ is the conditional probability, given as:

$$p\left(\mathbf{y}^P \mid \mathbf{a}\right) = \frac{1}{\pi^{NP}\left|\mathbf{R}_{z'}\right|}\exp\left(-\left[\mathbf{y}^P - \mathbf{D}\mathbf{a}\right]^H \mathbf{R}_{z'}^{-1}\left[\mathbf{y}^P - \mathbf{D}\mathbf{a}\right]\right), \tag{22}$$

$$\mathbf{R}_{z'} = \mathrm{E}(\mathbf{z}'\mathbf{z}'^H) = N_0\mathbf{I}, \tag{23}$$

where N_0 is mean noise power AWGN.

After computation, the ML estimates of channel response can be given as:

$$\hat{\mathbf{a}}_{ML} = \arg\max_{a} f_{ML}, \tag{24}$$

where $f_{ML} = \left\|\mathbf{y}^P - \mathbf{D}\mathbf{a}\right\|^2$. Finally, the ML estimates of channel response can be computed as:

$$\hat{\mathbf{a}}_{ML} = \left(\mathbf{D}^H\mathbf{D}\right)^{-1}\mathbf{D}^H\mathbf{y}^P, \tag{25}$$

with the value of the ML-based estimates of the channel response $\hat{\mathbf{a}}_{ML}$, we can determine the intended and self-interference channels response $\mathbf{h}$ and $\dot{\mathbf{h}}'$.

In a MIMO-OFDM full-duplex receiver, the self-interference can be several orders of magnitude higher than the signal of interest because the intended signal crosses longer distance than self-interference signal. Hence, different cancelation stages are needed to gradually suppress the self-interference before restore the data in receiver. Using the estimation self-interference channel response $\dot{\mathbf{h}}$, the self-interference signal can be suppressed as:

$$\begin{aligned} y_{n,m}^{(r),D} &= \sum_{t=1}^{N_t}\sum_{l=0}^{L-1} h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),D} + \sum_{t=1}^{N_t}\sum_{l=0}^{L-1} \dot{h}_{l,n,m}'^{(r,t)} \dot{x}_{n-l,m}^{(t),D} + z_{n,m}'^{(r)} - \sum_{t=1}^{N_t}\sum_{l=0}^{L-1} \hat{\dot{h}}_{l,n,m}'^{(r,t)} \dot{x}_{n-l,m}^{(t),D} \\ &= \sum_{t=1}^{N_t}\sum_{l=0}^{L-1} h_{l,n,m}^{(r,t)} x_{n-l,m}^{(t),D} + \sum_{t=1}^{N_t}\sum_{l=0}^{L-1} (\dot{h}_{l,n,m}'^{(r,t)} - \hat{\dot{h}}_{l,n,m}'^{(r,t)}) \dot{x}_{n-l,m}^{(t),D} + z_{n,m}'^{(r)}, \end{aligned} \tag{26}$$

Transformation the data received into frequency domain by the FFT block:

$$Y_{k,m}^{(r),D} = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} y_{n,m}^{(r),D} \exp\left(\frac{-j2\pi kn}{N}\right), \tag{27}$$

Rewrite the equation (25) as follow:

$$Y_{k,m}^{(r),D} = H_{l,k,m}^{(r,t)} X_{k,m}^{(t),D} + Z_{k,m}'^{(r)}, \tag{28}$$

With the noise is white Gaussian, the ML estimate $\hat{X}_{k,m}$ of the transmit symbols is found as:

$$\hat{X}_{k,m} = \underset{X_{k,m} \in C^t}{\mathrm{argmin}} \left\|Y_{k,m}^{(r),D} - H_{l,k,m}^{(r,t)} X_{k,m}^{(t),D}\right\|^2, \tag{29}$$

where C^t is the set of all possible t dimensional transmit vectors and $\|.\|$ stands for Euclidian norm.

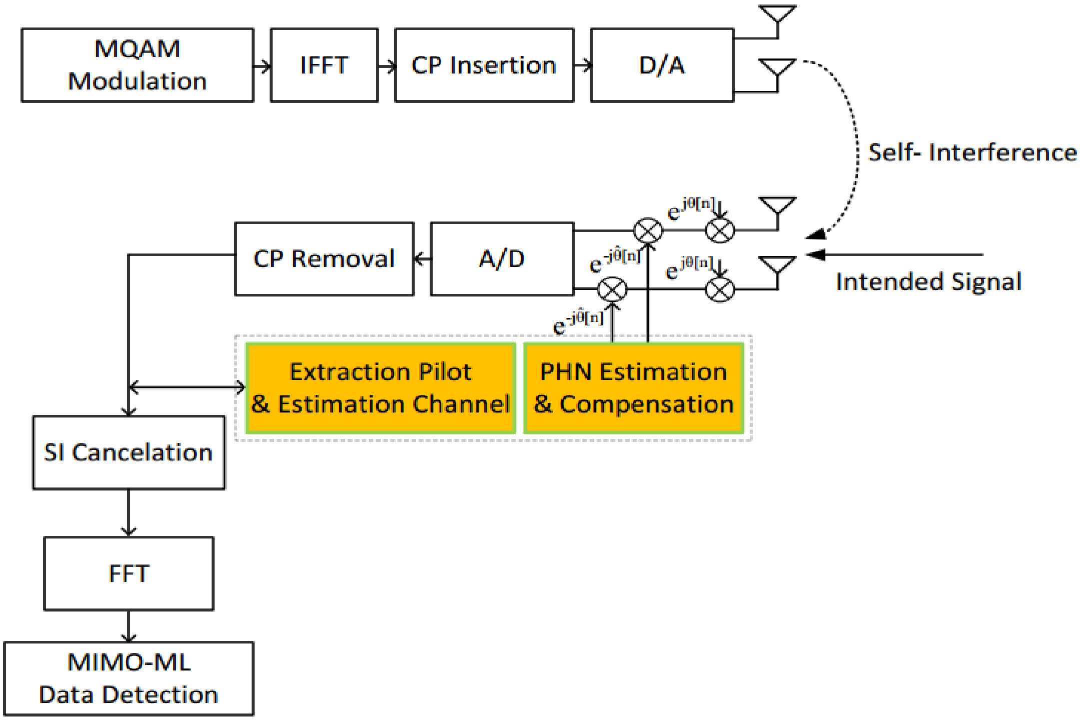

Fig.3: MIMO-OFDM Full-duplex tranceiver using Kalman filter to tracking phase noise and pilot-aided channel estimation.

In summary, the proposed estimation algorithm, self-interference cancellation and data detection works as follows:

1. Initialization:

$$\hat{\theta}_{n-1|n-1} = \mu_\theta = 0,\ \hat{P}_{n-1|n-1} = Q_\theta = 0.$$

// Phase noise Estimation

2. Compute the predicted state value and Kalman gain from (13) and (15).
3. Update the state value from (16) to (17).
4. **for** *i*= length$(\hat{\theta})$ **do**

   Compute from (13) to (17)

   end for

//Channel Estimation

5. Compensate phase noise by (18).
6. Compute the ML estimates CIR with knowledge pilot by (25).
7. Determine the intended and self-interference channels response.

//Self-Interference Cancellation and Data Detection

8. Subtraction the self-interference in received data by (26)
9. Detection the received data by (29)

IV. SIMULATION RESULTS

In this section, we evaluate the performance of the proposed joint channel and phase noise estimation versus the SNR. Subsequently, the BER of a MIMO-OFDM full-duplex system employing the proposed channel and phase noise estimators is investigated in detail under different strategy of operating conditions. Throughout this section, the LTE standard settings are used as a framework for the analysis. The system is assumed to operate over a quasi-static and frequency-flat fading channels, where the channel gains are assumed to remain constant over a OFDM frame. Each tranceiver is equipped with N_t = 2 transmitting antennas and Nr = 2 receiving antennas and the number of channel taps is set to L = 5. The MIMO-OFDM full-duplex system with its center frequency (f_c) at 2 GHz, sampling frequency fs = 1.92MHz and modulator using 64-point FFT. In addition, the transmitted bits are mapped to BPSK and the cyclic prefix of each OFDM symbol is set N_g = 10 sample. Each the burst of transmission with length M = 7 is inserted P = 2 pilot OFDM symbols and the free-running phase noise model is considered.

Fig. 4 shows the phase noise with βT=0.03, βT=0.1 and estimation using Kalman filter. With the bigger βT, the PHN is fluctuated with the larger amplitude and the marked points show that the Kalman filter is efficient in tracking the phase noise.

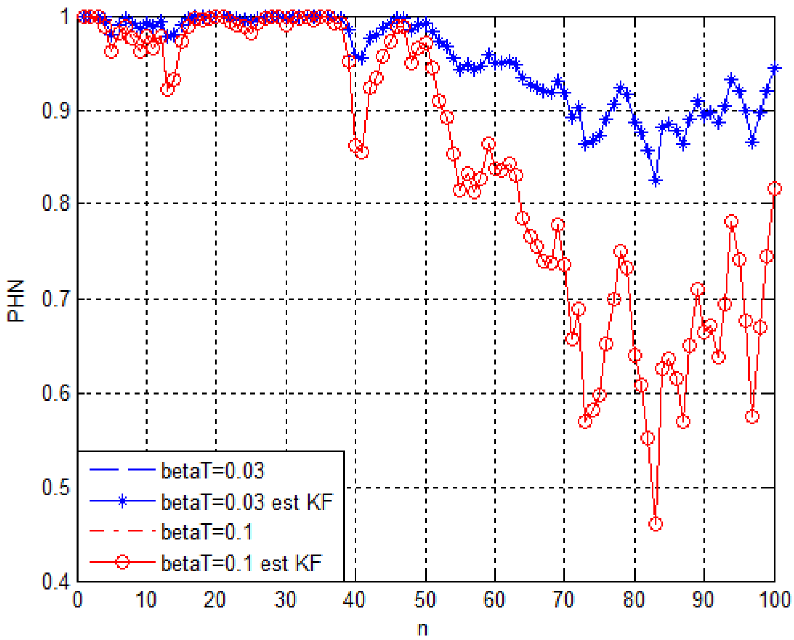

Fig. 4: Phase noise and Phase noise Estimation by Kalman with βT = 0.03, βT =0.1.

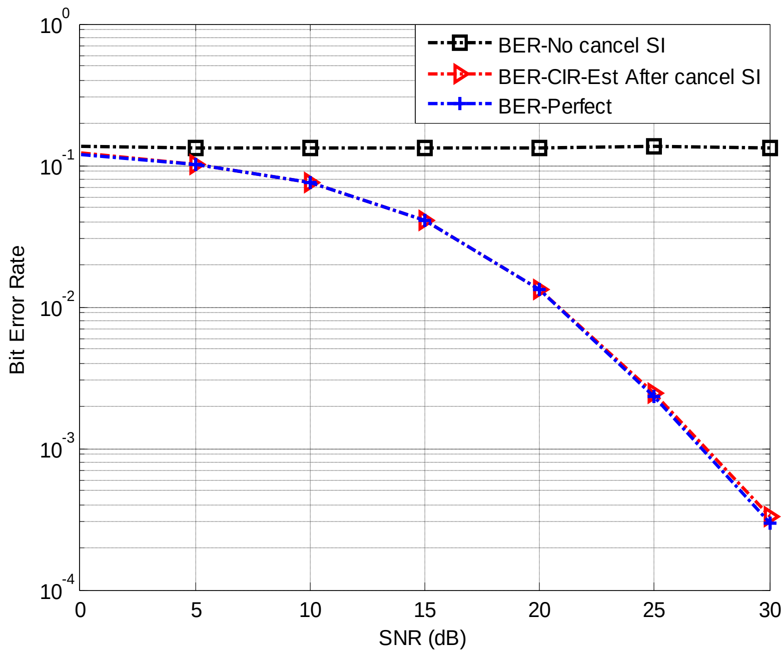

Fig. 5: BER versus SNR with and without Self-interference Cancelation.

To assess the efficient of self-interference cancelation stage in proposed algorithm which joint phase noise and channel estimation, Fig. 5 shows the BER performance using the ML channel estimation under block-fading channel versus SNR in two scenarios: without self-interference cancelation and self-interference cancelation by estimated self-interference channels response. As observed, with the received signal which compressed self-interference before demodulation, the BER is better than the one without compressed. The results in Fig. 5 demonstrate that without self-interference cancelation, the full-duplex system performance deteriorates significantly.

The BER performance versus SNR with and without Kalman phase tracking are showed in Fig. 6. The results show that the BER curve when tracking phase noise and compensation gets relatively near to the ideal case. In addition, the worst performance due to distortion affecting of phase noise when ignoring the presence of phase noise.

To further evaluate the effect of phase noise on BER performance of MIMO-OFDM full-duplex system, Fig. 7 shows the BER results under the use of some various βT of phase noise. In this figure, we can see that the presence of phase noise induces significant BER performance degradation as phase noise level βT increase.

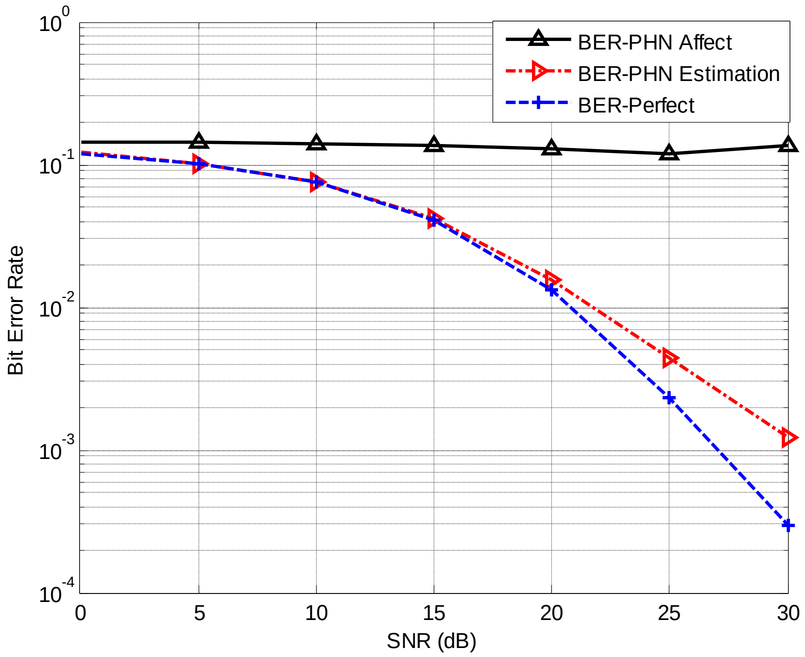

Fig. 6: BER versus SNR with and without Kalman phase tracking (βT=0.05).

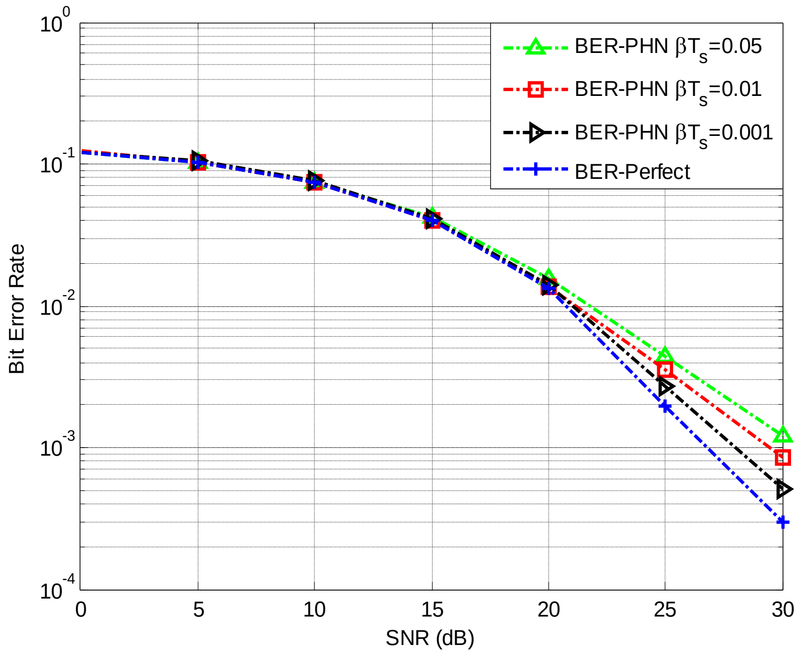

Fig. 7: BER versus SNR with various Phase noise level βT.

The BER performance curves are simulated under the SNR=20dB and 30dB versus phase noise level βT are shown in Fig. 8. It is observed that the BER curves increase insignificant when phase noise increases. The results demonstrate that the proposed algorithms which joint channel and phase noise estimation is efficient.

Fig. 9 compares the BER performance of the system for higher order modulations: BPSK, 4-QAM and 16-QAM. The results show that the system very sensitive to the number of dense constellations in the presence of phase noise.

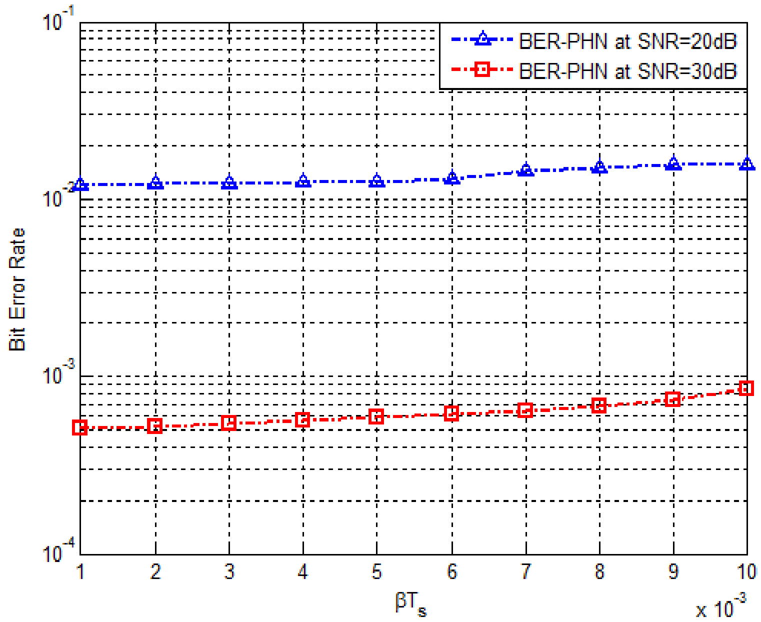

Fig. 8: BER versus Phase noise at SNR=20dB and SNR=30dB.

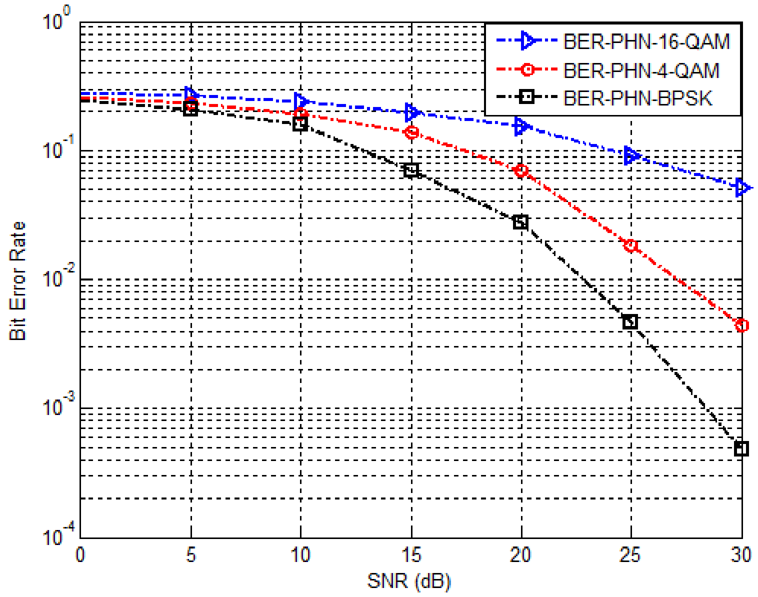

Fig. 9: BER versus SNR of full-duplex system for QPSK, 4-QAM, 16-QAM modulations.

V. CONCLUSION

In this paper, the ML channel estimation algorithm by using both pilot symbols and unknown data symbols from the transmitter and tracking phase noise by KF have been presented for MIMO-OFDM full-duplex systems over a quasi-static and frequency-flat fading channels. The proposed algorithm offers a stable performance with high efficient and robustness against effect of phase noise. Next, simulation results demonstrate that the performance of a system using the proposed ML and KF estimator is close to the idealistic setting of knowledge phase noise. Finally, the modulation with small amount constellations (eg: BPSK, 2-QAM, 4-QAM…) would be a suitable choice in the MIMO-OFDM full-duplex because of the sensitive with phase noise.

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