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EFFICIENT TIME-DOMAIN PHASE NOISE MITIGATION IN CM-WAVE WIRELESS COMMUNICATIONS

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ABSTRACT

The topic of this paper is low-complexity yet high-precision phase-noise mitigation in a realistic cm-wave radio link. The proposed algorithm for phase-noise induced intercarrier interference mitigation works in time domain. We propose a filtering solution that is computationally very simple to implement on silicon. Noise from initial phase-noise estimate is removed with cumulative sum filtering instead of a very high order complex-tap filter. The performance of the proposed algorithm is evaluated in realistic ITU-R Urban Microcell Non-Line-of-Sight channel as well as in white Gaussian noise channel, and with state-of-the-art phase-locked-loop oscillator generated phase noise. The obtained results show that highly-efficient phase-noise mitigation can be obtained using the proposed approach. For the example 5G case, the algorithm suppresses the phase noise so that the bit-error-rate performance is near the case without phase noise.

Index Terms**—**Centimeter-Wave, 5G, Smal Cell, OFDM, Phase Noise Mitigation, Efficient Implementation.

1. INTRODUCTION

Phase noise is one of the most significant performance limiting radio frequency (RF) impairments in multicarrier communications [1]. In orthogonal frequency division multiplexing (OFDM), the phase noise spreads the energy of subcarriers to interfere the other subcarriers [2]. The interference is called intercarrier interference (ICI), and it destroys the orthogonality between the subcarriers. In addition of the ICI, also the rotating effect, called common phase error (CPE), of the phase noise is present. It cases all the subcarrier symbols in an OFDM symbol experience the same rotation.

With current mobile cyclic-prefix (CP) based OFDM (CP-OFDM) transceivers, the phase noise is kept under control by strict oscillator design and CPE mitigation, because the ICI is computationally very complex to estimate. Therefore, especially in mobile transceivers, ICI is generally not mitigated at all. Although the used oscillators and their phase noise profile can be expected to still improve by the launch of 5G products, the shift towards cm-wave communications, and the adoption of larger symbol constellations, e.g. 256-QAM and 1024-QAM, for mobile communications will still require low complexity ICI cancellation algorithms. At the same time, the high-complexity ICI mitigation algorithms proposed in the some of the earlier literature cannot be utilized because small devices are not capable of high-complexity computations.

In this paper, we show that phase noise is a huge problem in cm-wave communications with state-of-the-art oscillator technology. Even though there are also other proposals as waveforms of the 5G systems at cm-wave, the system parameters in this paper are based on CP-OFDM scheme with cm-wave numerology, which is also one of the proposals for 5G communications. We propose computationally efficient yet accurate ICI estimation and mitigation algorithm. The algorithm is building on the algorithm proposed in [3] and also discussed in [4]. The original algorithm provides high quality phase noise estimates, but suffers from relatively high complexity. Especially the used filter in the original algorithm is very complex, and therefore highly impractical for high-data rate communications. In this paper, the filter is substituted by computationally very efficient cumulative sum processing that exploits the spectral shape of the phase noise process better.

In Section II, the state-of-the-art of phase noise mitigation in CP-OFDM is shortly discussed, the proposed phase noise mitigation algorithm is described and its complexity analyzed. In Section III, the simulation environment is described and the results are given with the simulations analysis. Section IV then concludes the work.

2. PHASE-NOISE MITIGATION IN CP-OFDM

2.1 State of the Art

For OFDM communications, usually the CPE is first estimated and mitigated. This is rather simple task with the help of pilot subcarriers, e.g., by averaging [2]. Notice that it is mitigated with channel equalization [5], but if channel is not changing fast and the channel is only estimated, e.g., in a preamble of a frame of multiple CP-OFDM symbols, the CPE mitigation needs to be done separately for each CP-OFDM symbol [4]. The computational load is anyway minor.

ICI estimation on the other hand is much more complex process. Usual way to carry out ICI estimation is to use initial decisions after the CPE mitigation to estimate the ICI [2]. As ICI stems from the change of phase error from sample to sample, it must be estimated separately for each CP-OFDM symbol, and the phase error needs to be estimated accurately for the entire duration of each CP-OFDM symbol. The computational load is therefore very

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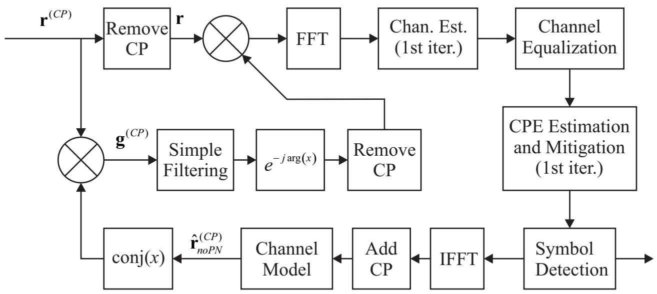

Fig. 1. Illustration of the proposed algorithm.

high. One high-performance phase noise mitigation technique is proposed in [2]. There, initial decisions are used for phase-noise estimation in frequency domain. The method provides high performance, but is at the same time rather computationally complex. The same idea of using the initial decision for phase noise estimation and mitigation is proposed in [3], but in time-domain. The technique provides excellent performance, but is also rather computationally complex.

2.2 Proposed Algorithm

The proposed algorithm is building on the phase-noise mitigation algorithm first proposed in [3], and it is depicted in Fig. 1. Let us start from the received signal of the form

$$\mathbf{r} = \operatorname{diag}\left(e^{j\boldsymbol{\phi}}\right)\mathbf{Hx} + \mathbf{z}\,. \tag{1}$$

Here $\operatorname{diag}\left(\bullet\right)$ makes a diagonal matrix from the input vector, $\boldsymbol{\phi}$ is a $N\times1$ vector of the sampled phase noise, $\mathbf{H}$ is the $N\times N$ circulant convolution matrix of the channel [6], $\mathbf{x}$ is a vector of the samples of the useful signal, and $\mathbf{z}$ is a $N\times1$ vector of additive noise. The $N\times1$ vector $\mathbf{r}$ represents the signal samples during an CP-OFDM symbol without the CP. Notice that we have left out the CP-OFDM symbol index for simpler presentation, as the phase noise needs to be estimated for each CP-OFDM symbol independently anyway [4]. In (1) In (1), we have assumed that the maximum delay spread of the channel is shorter than the length of the CP. We have also assumed that the transmitter and receiver phase noises can be both modelled as receiver phase noise. This is just an approximation, but it holds very well [2], and furthermore no such approximation will be made in the numerical performance evaluations in this paper.

A crude estimate of the phase noise can be obtained relatively easily by using the detected symbols after the channel has been equalized and the CPE has been mitigated. We first make the hard decisions before decoding so that computational complexity can be kept low, pass the decisions through inverse discrete Fourier transform (IDFT) carried out by fast Fourier transform (FFT) algorithm. The resulting signal is denoted by $\hat{\mathbf{x}}$. Then we add CP to the signal, after which the channel is modelled based on the circulant channel estimate matrix $\hat{\mathbf{H}}$. The signal at this point is denoted by $\hat{\mathbf{r}}_{noPN}^{(CP)}$. This signal without CP can be written as

$$\hat{\mathbf{r}}_{noPN} = \hat{\mathbf{H}}\hat{\mathbf{x}}\,. \tag{2}$$

A conjugate, denoted by $\left(\cdot\right)^{*}$, of the signal is then taken, and the signal vector $\mathbf{r}^{(CP)}$ ($\mathbf{r}$ in (1) with CP) is multiplied with this conjugate. The resulting vector of the crude estimates of the phase-noise complex exponential without CP can be written as

$$\begin{aligned}\mathbf{g} &= \operatorname{diag}\left(\hat{\mathbf{x}}^{*}\hat{\mathbf{H}}^{*}\right)\left[\operatorname{diag}\left(e^{j\boldsymbol{\phi}}\right)\mathbf{Hx} + \mathbf{z}\right]\\ &\approx \operatorname{diag}\left(\left|\hat{\mathbf{H}}\hat{\mathbf{x}}\right|^{2}\right)e^{j\boldsymbol{\phi}} + \operatorname{diag}\left(\hat{\mathbf{x}}^{*}\hat{\mathbf{H}}^{*}\right)\mathbf{z} + \mathbf{n}\,.\end{aligned} \tag{3}$$

Here, the $N\times1$ vector $\mathbf{n}$ is a noise vector from channel estimation error and symbol detection errors. In (3), the estimate is still very noisy and contains some scaling. However, as can be seen in (3), the actual phase-noise estimate is scaled so that the vector elements with highest powers in the received signal are given more emphasis than the lower power elements. This is actually desirable in the next step when we filter out some estimation errors. Since phase noise and its complex exponential are steep lowpass processes, we can use a lowpass filter to filter out most of the noise. The actual filter selection is proposed in the next subsection. The resulting signal is

$$\mathbf{b} * \mathbf{g} \approx \operatorname{diag}\left(\mathbf{a}\right)e^{j\hat{\boldsymbol{\phi}}}\,. \tag{4}$$

Here, the filter's impulse response vector is $\mathbf{b}$, the convolution is denoted by $*$, and elements of the vector $\mathbf{a}$ represent the incorrect scaling in the estimate of the phase noise complex exponential. The incorrect scaling can now be removed by using just the phase of the estimate to construct the inverse of the phase-noise complex exponential $\exp\left(-j\hat{\boldsymbol{\phi}}\right)$, that is used for the phase-noise mitigation by multiplying the received time-domain signal with it.

The computational complexity can be lowered by estimating only part of the phase noise samples, and then use interpolation to obtain the rest of the sample estimates.

2.3 Selecting Lowpass Filter for Proposed Algorithm

In [3], the design of the lowpass filter was left as a future work, and order 350 filter designed with Remez-algorithm was used. The used example filter is not realistic for practical utilization as it has 351 complex taps, and imposes huge extra complexity to the overall algorithm. In effect, the filter requires 351 complex multiplications.

We propose to use a filter which is designed based on the physical properties of the phase noise and is very efficiently implementable in silicon. In the literature, a free running oscillator (FRO) is usually modelled by cumulatively summing zero-mean white Gaussian noise. This results in rather realistic –20 dB/dec slope. Our proposed filter is based on this observation. We propose a lowpass filter that cumulatively sums the filtered samples within a summing window. The resulting filter does not require any multiplications and is therefore extremely efficient to implement in silicon. The cumulative summing is done in a window, because the signal to be filtered has huge errors in it, and infinite cumulative summing would result in unstable solution. In the simulations section, the windows size is discussed in more detail. Example responses of the cumulative sums with windows lengths 30 and 10 are depicted in Fig. 2. The filtering vector $\mathbf{b}$ in (4) is therefore a vector of ones, and the length is the length of the window. Notice that this filter amplifies the lower frequencies. However, since after the filtering only the phase of the filtered signal is of any interest, the amplitude does not matter.

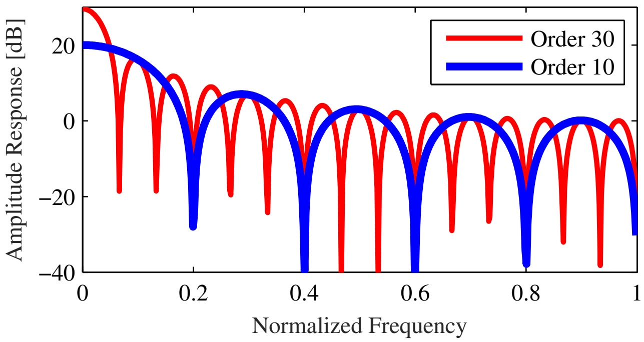

Fig. 2. Amplitude responses of the cumulative sum filters with window lengths (orders) of 30 and 10.

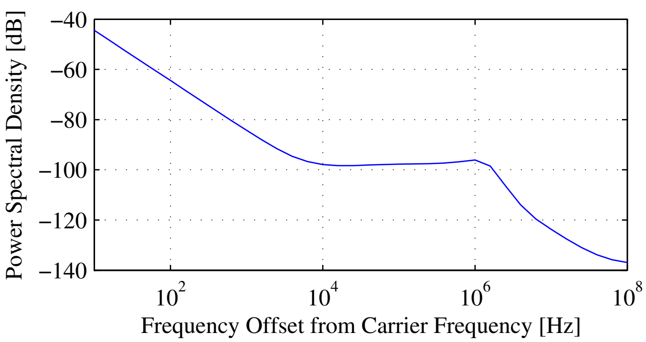

Fig. 3. Spectrum of the phase noise of the used PLL oscillator as a function of the offset from the 6 GHz carrier frequency.

TABLE I. FIXED AND DEFAULT PARAMETERS IN SIMULATIONS

FFT Size	2048
Subcarrier Spacing	60 kHz
Active Subcarriers	1280
Carrier Bandwidth	80 MHz
Subcarrier Modulation	256QAM
Carrier Frequency	28 GHz

2.4 Complexity Analysis

In addition of the regular symbol detection, the proposed algorithm requires about 36 complex multiplications per received sample (subcarrier modulated symbol). Most these multiplications stem from the FFT and IFFT algorithms that both require $N\log_2 N$ complex multiplication each per OFDM symbol, which is $\log_2 N$ complex multiplication per each sample. The actual number depends on the quality of the channel estimate, as the channel estimate is used in time-domain in the algorithm, refer to Fig. 1. In the number 36, it is assumed that the channel estimate has 10 clear taps. The algorithm proposed in [3] requires about 100 to 350 tap complex filter, so compared to the proposed algorithm, it requires 100 to 350 more complex multiplications per sample, resulting in total of around 136 complex multiplications per sample.

In the reference algorithm of [2], if 21 frequency components are estimated, at least 21 *reliably detected* subcarrier symbols are required to construct the minimum mean square estimation (MMSE) matrix [2]. In practice, this minimum amount is far from enough. In this work, we use 40 reliably detected subcarrier symbols. This configuration results in 74 complex multiplications per sample to generate the phase noise estimate. With the minimum amount of 21 reliably detected symbols, this figure would be 23 complex multiplications. In addition of the phase noise estimation, the convolution in the phase noise mitigation requires 21 complex multiplications per sample, resulting in total of 95 complex multiplications per sample. If less frequency components are used in the estimation, the complexity is lowered, but so is the estimate quality. 21 frequency components is a compromise between good performance and reasonable computational complexity.

3. SIMULATIONS AND ANALYSIS

3.1 Simulator and Parameters

The parameterization of the simulated system follows the work done on 5G and cm-wave communications in [7], [8] and [9]. In the simulator, first, bits are generated and transformed to QAM symbols with Gray coding. Then, CP-OFDM signal with 2048 subcarriers is generated so that 640 on the both sides of the zero-subcarrier are carrying 256QAM modulated data, and subcarriers 464, 624, 784, 944, 1104, 1264, 1424 and 1584 are the pilot subcarriers. The subcarrier spacing is 60 kHz. After the CP-OFDM signal is generated, the transmitter phase noise is modelled.

The phase noise is modelled as so-called Ornstein-Steinbeck process [10], [11]. This phase-locked loop oscillator model is constructed by using the physical properties of state of the art oscillators. The used reference oscillator parameters are from Vectron VT-704 [12] and the used voltage controlled oscillator and loop parameters are from [13], scaled to the desired centre frequency. Based on the past development of oscillators used in mobile devices, we expect this to be good baseline for mobile-terminal oscillators in the expected deployment of 5G. The power spectral density (PSD) of the used oscillator is depicted in Fig. 3.

After the transmitter phase noise, the channel is modelled. We use ITU-R Urban Microcell Non-Line-of-Sight (UMi NLoS) channel model [14]. Also, plain additive white Gaussian noise (AWGN) channel (flat channel) is used as another case for reference. After the channel, AWGN is added to achieve desired signal-to-noise ratio (SNR). Then, at the receiver, the receiver phase noise is modelled in the same way as the transmitter phase noise is modelled. Finally, CPE is estimated by averaging over the pilot subcarriers [2] and mitigated, followed by the use of the presented and proposed ICI mitigation techniques. Bit-error rates (BER) are then calculated with Gray coding.

The used ICI mitigation techniques are the proposed technique and the reference techniques from [2] and [3]. There are also many other techniques in the literature, but choice needed to be made which techniques to use as references. The presented techniques give very good performance with relatively low complexity. In the results the technique of [2] is referred as 'ICI est. Freq.' and the technique of [3] is referred as 'ICI est. Time'. As default, all the techniques are iterated 3 times unless otherwise stated. For the technique of [2], 21 frequency bins of the phase noise are estimated. We also use 40 reliable subcarrier symbol decisions for the algorithm (see [2] for details), so that it is able to provide good phase noise estimates. For the algorithm in [3], we used order 200 lowpass filter designed with Remez algorithm. The transition band was set to start from normalized frequency 0 and end at 0.04. Default window size for the proposed algorithm is 20.

The fixed and default parameters for the simulations are summarized in Table I.

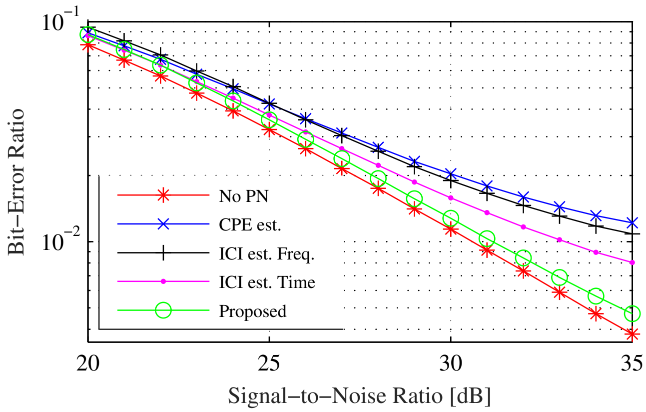

Fig. 4. Bit-error ratio given as a function of signal-to-noise ratio. Default parameters. ITU-R UMi: NLoS channel.

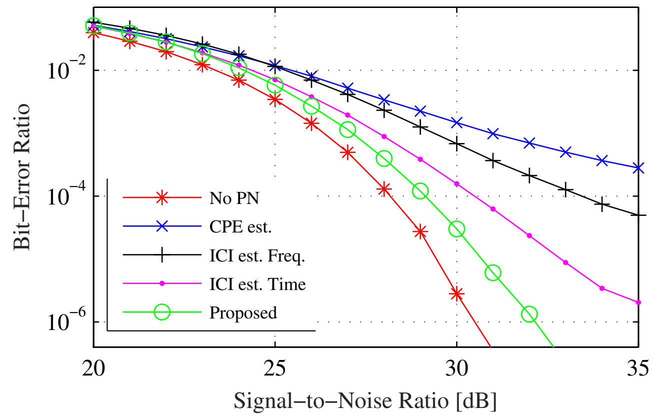

Fig. 6. Bit-error ratio given as a function of signal-to-noise ratio. Default parameters. AWGN channel.

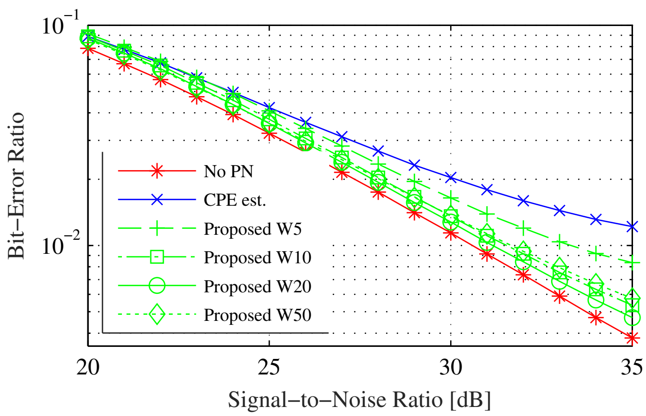

Fig. 5. Bit-error ratio given as a function of signal-to-noise ratio. Default parameters, except for the proposed technique whose windows size is set to 5, 10, 20 or 50 samples. ITU-R UMi: NLoS channel.

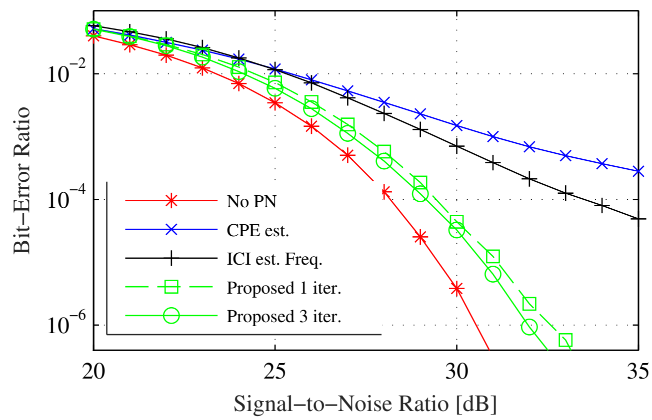

Fig. 7. Bit-error ratio given as a function of signal-to-noise ratio. Default parameters, except for the proposed technique whose amount of iterations is either 1 or 3. AWGN channel.

3.2 Simulation Results and Analysis

The resulting BER results for ITU-R UMi NLoS channel case are given in Fig. 4 and Fig. 5. The default parameters are used in Fig. 4 and the filter size is changed for the proposed algorithm in Fig. 5. First of all, we can see that the impact of the phase noise of a state of the art oscillator on the performance of a realistic CP-OFDM-based 5G radio link is significant, if only CPE is mitigated.

As depicted in Fig. 4, the proposed algorithm gives very near to ideal no phase noise performance in the studied SNR region. Compared to the reference phase-noise mitigation techniques, the performance is very good. For example, at 1 % BER level, the improved performance provided by the proposed algorithm over the reference algorithm of [3] is around 2.5 dB in terms of SNR, and gets higher at lower BER levels. From Fig. 5, we can see that with short windows, the filtering effect is not strong enough to filter out the noise sufficiently. With long filters also the desired frequencies get attenuated for the used phase noise model. Selection of the window size should therefore be done depending on the phase noise model, subcarrier spacing and the sampling frequency.

Notice that already one iteration of the proposed algorithm provided close to identical performance to 3 iterations in the ITU-R UMi NLoS channel, so the number of iterations is only considered for the AWGN channel case.

The results for the AWGN channel case are depicted in Fig. 6 and Fig. 7. The default parameters are used in Fig. 6. The performance differences are similar to the ITU-R UMi NLoS case. In Fig. 7 we see that, we get a small gain with 3 iterations compared to a single iteration. However, the performance with 1 iteration is already very good compared to the reference techniques, and can result in around 10 % of the computational complexity of the reference algorithm with better performance. The practical complexity of the algorithm is therefore much less than that of the reference algorithms.

4. CONCLUSION

Impact of phase noise on the performance of high-frequency communications is significant. The computational resources needed for ICI mitigation in modern CP-OFDM communications systems are generally relatively high, because of large bandwidths and high data rates. The ICI-mitigation algorithm proposed in this paper performs the ICI mitigation task with great accuracy and at the same time with relatively low computational complexity. Compared to the state-of-the-art low-complexity ICI-mitigation algorithms, the performance is significantly better, even with just a single iteration, while the reference techniques use several iterations. Even with the same amount of iterations, the algorithm provides better computational complexity.

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