Receiver DSP for OFDM Systems Impaired by Transmitter and Receiver Phase Noise

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Abstract—**This paper proposes a time-domain digital signal processing method for estimating and mitigating transmitter and receiver oscillator phase-noise effects in OFDM radio systems on the receiver side of the link. The idea is based on re-constructing time-domain OFDM signal at the receiver from initially detected symbols and using this as a reference in phase noise estimation. The knowledge of heavily low-pass nature of realistic phase noise processes is then utilized in the estimation process to improve the estimation quality. The algorithm can also be used iteratively, inside individual OFDM symbols, to further improve the accuracy of the obtained phase noise estimate. Performance analysis shows that the proposed algorithm outperforms existing state-of-the-art phase noise mitigation techniques, under both additive white Gaussian noise and extended ITU-R Vehicular A multipath channels.**

Index Terms—**OFDM; phase noise; ICI; mitigation; digital signal processing; dirty-RF**

I. Introduction

In recent years, there have been great advances in digital signal processor implementation techniques. This has enabled significant increases in signal processing capabilities of small devices, such as mobile phones, without the increase in size, cost and power consumption. The increased computational power allows moving the complexity of the transceiver units from the analogue component side to the digital signal processing (DSP) side. Actually used analogue components in transceivers can thus be small, cheap and low-power. Even though having more complex signal processing algorithms to handle, overall costs, sizes and power consumptions of transceivers can be significantly lowered [1].

Orthogonal frequency domain multiplexing (OFDM) is a multicarrier modulation technique used in many current and future communications systems (such as DVB-T, WiMAX, 3G LTE, etc.). Even though it has many advantages, OFDM is very prone to transceiver RF-impairments [2]. One of the most harmful of these impairments is phase noise. In addition to well-known rotating effect, called common phase error (CPE), phase noise also causes subcarriers to lose their orthogonality by spectrally spreading subcarriers on top of each other [3]. The spread is called intercarrier interference (ICI) and its effect changes from subcarrier to another making the compensation difficult [3], [4]. It is thus very important to develop new signal processing algorithms to mitigate the phase noise effects in OFDM systems.

By today, the phase noise impaired OFDM systems have already received quite extensive attention in the literature. Phase noise effects on OFDM systems have been studied, e.g., in [3] and [5]. Deeper pen-and-paper analysis of the phase noise effects was carried out, e.g., in [6]. Mitigation of the CPE part of the phase noise has been considered already in [3], and in [7] mitigation performance of such CPE mitigation technique was improved. Furthermore, in [7] they also proposed a simplistic ICI mitigation algorithm. More advanced ICI mitigation technique was then proposed in [4] and some improvements for this method in [8] and [9]. In [10] and [11], the authors of this paper proposed further modifications for the technique proposed in [4] to improve its performance.

In this paper, a novel DSP algorithm is proposed for phase noise mitigation and its performance is compared to the performances of the state-of-the-art techniques from the literature. The technique works on time domain signal in an iterative manner. The technique does not need any prior information about the phase noise process, making it easily implementable and versatile. Only assumption that needs to be made is that the phase noise process is heavily low-pass natured. This is true for all practical oscillators.

The rest of the paper is structured as follows. In Section II, OFDM transmission chain and phase noise are modelled. Section III then gives the proposed phase noise mitigation algorithm and practical considerations related to its implementation. After this, the performance of the proposed algorithm is compared to the state-of-the-art phase noise mitigation techniques in Section IV. At length, the work is concluded in Section V.

II. System Modelling

This Section gives the overall OFDM transmission chain and oscillator phase noise modelling as a basis for the algorithm development.

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A. OFDM Transmission Chain Model

In OFDM symbol generation, first modulated symbols $X_k(m)$ are grouped in blocks of N subcarrier symbols $k = 0,1,\ldots,N-1$, where mth block refers to the mth OFDM symbol. These blocks are then inverse discrete Fourier transformed (IDFT) giving sampled OFDM symbols, where nth sample is

$$x_n(m) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k(m) e^{j2\pi kn/N} . \quad (1)$$

Here, sampling indices are $n = 0,1,\ldots,N-1$. So, the resulting OFDM symbol is N samples long giving it length of NT_s seconds and subcarrier spacing of $f_s = 1/(NT_s)$, where T_s is the sampling interval. Furthermore, to exploit the advantage of having relatively long symbol durations, a cyclic prefix is added to the OFDM symbols before transmission. Namely, we transmit G last samples of each OFDM symbol prior the actual symbol samples. This effectively makes the sent blocks partially circular making the stream immune to inter-symbol interference if the maximum delay spread of the transmission channel is shorter than the duration of the cyclic prefix GT_s. Then, the overall OFDM symbol length is $(G+N)T_s$. [12]

In this work, we assume that the maximum channel delay spread is less than the length of the cyclic prefix. So in the receiver, after the signal has gone through the ideal up-conversion, the transmission channel, the ideal down-conversion and the removal of the cyclic prefix, we can write the received signal as

$$\mathbf{r}^{(m)} = \left(\mathbf{h}^{(m)} * \mathbf{x}^{(m)}\right) + \mathbf{z}^{(m)} . \quad (2)$$

Here, operator $*$ denotes circular convolution between the elements of the operated vectors. $\mathbf{h}^{(m)}$ is a $(D \times 1)$ channel impulse response vector $(D < G)$ and vector $\mathbf{x}^{(m)} = \left[x_0(m), x_1(m), \ldots, x_{N-1}(m)\right]^T$. Vector $\mathbf{z}^{(m)}$ denotes additive white Gaussian noise. Now, when transmitter and receiver oscillator phase noises are assumed present, (2) can be written as

$$\mathbf{r}^{(m)} = \operatorname{diag}\left(e^{j\boldsymbol{\varphi}_R^{(m)}}\right)\left(\mathbf{h}^{(m)} * \left(\operatorname{diag}\left(e^{j\boldsymbol{\varphi}_T^{(m)}}\right)\mathbf{x}^{(m)}\right)\right) + \mathbf{z}^{(m)} , \quad (3)$$

where $\operatorname{diag}(\ )$ is a function which creates a diagonal matrix that has the input vector in its diagonal. The vectors of the sampled transmitter and receiver phase noise realizations during the mth OFDM symbol are $\boldsymbol{\varphi}_T^{(m)}$ and $\boldsymbol{\varphi}_R^{(m)}$, respectively. For nth sample of mth OFDM symbol, they are defined as $\boldsymbol{\varphi}_X^{(m)} = \left[\varphi_{0,X}(m), \varphi_{1,X}(m), \ldots, \varphi_{N-1,X}(m)\right]^T$, $X \in \{T, R\}$, where $\varphi_{n,T}(m)$ and $\varphi_{n,R}(m)$ are the transmitter and receiver phase noise samples, respectively. Since reasonable channel delay spread is assumed, the channel coherence bandwidth is reasonably high. Therefore, since phase noise process is typically a steep low-pass process [2], [4], (3) can be approximated as

$$\begin{aligned}
\mathbf{r}^{(m)} &\approx \mathbf{h}^{(m)} * \left[\operatorname{diag}\left(e^{j\boldsymbol{\varphi}_R^{(m)}}\right)\operatorname{diag}\left(e^{j\boldsymbol{\varphi}_T^{(m)}}\right)\mathbf{x}^{(m)}\right] + \mathbf{z}^{(m)} \\
&= \mathbf{h}^{(m)} * \left[\operatorname{diag}\left(e^{j\left[\boldsymbol{\varphi}_R^{(m)} + \boldsymbol{\varphi}_T^{(m)}\right]}\right)\mathbf{x}^{(m)}\right] + \mathbf{z}^{(m)} \\
&= \mathbf{h}^{(m)} * \left[\operatorname{diag}\left(e^{j\boldsymbol{\varphi}^{(m)}}\right)\mathbf{x}^{(m)}\right] + \mathbf{z}^{(m)}
\end{aligned} . \quad (4)$$

Here, $\boldsymbol{\varphi}^{(m)} = \boldsymbol{\varphi}_T^{(m)} + \boldsymbol{\varphi}_R^{(m)}$ denotes the total effective phase noise. Thus, stemming from the approximation made in (4), we are able to model the receiver phase noise similarly as the transmitter phase noise. The phase noises can be effectively summed together and the sum viewed as transmitter phase noise. Modelling both phase noises as transmitter phase noise is beneficial in algorithm derivation later on. Now, by using a circular convolution matrix [12] for the convolution in (4), we can write the final link model as

$$\mathbf{r}^{(m)} \approx \mathbf{H}^{(m)} \operatorname{diag}\left(e^{j\boldsymbol{\varphi}^{(m)}}\right)\mathbf{x}^{(m)} + \mathbf{z}^{(m)} , \quad (5)$$

where $\mathbf{H}^{(m)}$ is $(N \times N)$ channel circular convolution matrix.

B. Phase Noise Model

In the phase noise modelling, we are interested in how phase noise behaves as a function of sample index. In this Subsection, let us simply denote an arbitrary sampled phase noise process as $\phi_n = \phi(nT_s)$. In the literature, the phase noise is usually assumed to follow the well-known free-running oscillator model [2], [3]. The model assumes that there is no phase lock, so the next phase realization depends only on the previous realization and on the quality of the oscillator. This is perceived as a Brownian motion process, where the variance determines the oscillator quality. The phase can be written as

$$\phi_n = \sqrt{\frac{\beta}{4\pi}} B(nT_s) \ , \quad (6)$$

where $B(\ )$ is the standard Brownian motion function, namely $B(t_1) - B(t_2) \sim \mathcal{N}(0, |t_2 - t_1|)$, and β is one-sided 3-dB bandwidth of the phase noise process. Equation (6) can be also written as

$$\phi_n = \sqrt{\frac{\beta T_s}{4\pi}} B_n , \quad (7)$$

where the B_n is merely a process of cumulatively summed realizations of standard normal distributed $\mathcal{N}(0,1)$ random variable. So eventually, we can characterize the whole phase noise process with just a one parameter β. In general, this kind of free-running oscillator gives very demanding conditions for phase noise mitigation task [10]. For this reason, even though phase locked loop (PLL) type oscillators are more commonly deployed in practice, the free-running oscillator case is assumed in the forthcoming performance simulations.

III. THE PHASE NOISE MITIGATION ALGORITHM

The first part of this Section gives the basic idea and formulation of the proposed algorithm. The second part gives a method to further improve the performance of the algorithm. The last part of the Section gives some practical issues that must be considered when using the proposed algorithm.

A. Proposed Algorithm

The whole algorithm from the sampled down-converted waveform to detected symbols is depicted in Fig. 1.

The derivation of the proposed phase noise mitigation algorithm begins from the received signal waveform (5). After channel equalization and CPE mitigation, (5) reaches a form

$$\begin{aligned}\mathbf{y}^{(m)} &\approx \operatorname{diag}\left(e^{-\hat{\varphi}_{CPE}^{(m)}}\right)\left(\hat{\mathbf{H}}^{(m)}\right)^{-1}\left[\mathbf{H}^{(m)}\operatorname{diag}\left(e^{j\boldsymbol{\varphi}^{(m)}}\right)\mathbf{x}^{(m)}+\mathbf{z}^{(m)}\right]\\ &\approx \operatorname{diag}\left(e^{j\left(\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}\right)}\right)\mathbf{x}^{(m)}+\operatorname{diag}\left(e^{-\hat{\varphi}_{CPE}^{(m)}}\right)\left(\hat{\mathbf{H}}^{(m)}\right)^{-1}\mathbf{z}^{(m)}\\ &= \operatorname{diag}\left(e^{j\left(\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}\right)}\right)\mathbf{x}^{(m)}+\left(\hat{\mathbf{H}}_{CPE}^{(m)}\right)^{-1}\mathbf{z}^{(m)}\end{aligned} \quad (8)$$

Here, $\hat{\varphi}_{CPE}^{(m)}$ is the estimated CPE and $\hat{\mathbf{H}}^{(m)}$ is the estimated channel circular convolution matrix during the m th OFDM symbol. $\hat{\mathbf{H}}_{CPE}^{(m)}$ has also the CPE rotation combined with the channel. Now, after discrete Fourier transform (DFT) and symbol detection, we get the initial symbol estimates. Already at this stage, when only the channel and CPE are compensated, the symbol estimates $\hat{X}_k(m), k=0,\dots N-1$ are relatively reliable with reasonable phase noise levels [3], [10]. Therefore, with the reliability assumption, we can use these symbol decisions in the phase noise estimation process. When we take the symbol estimates, and turn them back into OFDM time-domain waveform with IDFT, we get the estimate of the transmitted waveform $\hat{\mathbf{x}}^{(m)}$. Now, by point-by-point dividing $\mathbf{y}^{(m)}$ in (8) with $\hat{\mathbf{x}}^{(m)}$, we get the estimate for the phase-noise complex exponential

$$\begin{aligned}\boldsymbol{\vartheta}^{(m)} &\approx \left[\operatorname{diag}\left(e^{j\left(\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}\right)}\right)\mathbf{x}^{(m)}+\left(\hat{\mathbf{H}}_{CPE}^{(m)}\right)^{-1}\mathbf{z}^{(m)}\right]\operatorname{diag}^{-1}\left(\hat{\mathbf{x}}^{(m)}\right)\\ &\approx e^{j\left(\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}\right)}+\left(\hat{\mathbf{H}}_{CPE}^{(m)}\right)^{-1}\mathbf{z}^{(m)}\operatorname{diag}^{-1}\left(\hat{\mathbf{x}}^{(m)}\right)\end{aligned} \quad . (9)$$

In some rare realizations $\hat{\mathbf{x}}^{(m)}$ has zero-elements in it. Those elements must be set to non-zero value prior to division to prevent the division by zero. In this study, we use high value $a\ (a\to\infty)$ to replace zeros, because it minimizes the potential error in the resulting phase noise estimate.

Since $\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}$ has heavily decreasing low-pass spectrum, it has almost all of its power at very low frequencies. Thus by taking argument (denoted by arg-function) and filtering the signal with a highly selective low-pass filter (denoted by LPF-function), we obtain phase noise estimate of the form

$$\hat{\boldsymbol{\varphi}}_{est}^{(m)}=\operatorname{LPF}\left\{\arg\left(\boldsymbol{\vartheta}^{(m)}\right)\right\}\approx\arg\left\{\operatorname{LPF}\left(\boldsymbol{\vartheta}^{(m)}\right)\right\}\approx\boldsymbol{\varphi}^{(m)}-\hat{\varphi}_{CPE}^{(m)}. \quad (10)$$

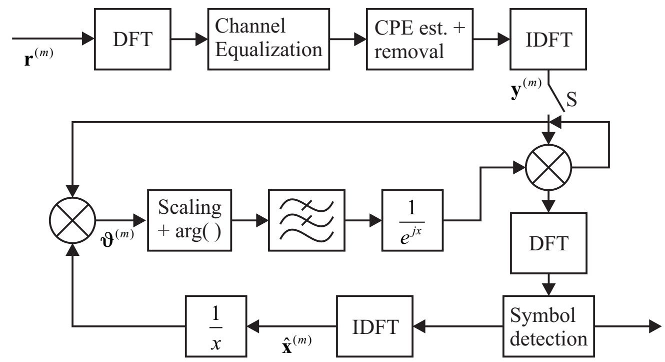

Fig. 1. The overall algorithm described in a block diagram to ease up the implementation. The vectors written in the diagram correspond to vectors with the same names in the equations. Variables x in block markings denote the input signal for the corresponding block. Switch S is connected only for the first iteration.

Due to the fact that phase noise process and its complex exponential are both heavily low-frequency processes, there is actually no correct order to do argument and low-pass filtering tasks in (10). According to simulations, taking the argument before low-pass filtering works better in phase noise dominated systems, whereas the other order of the operations works better in additive noise dominated systems. For simplicity, from now on we consider only the case where argument operator precedes the low-pass filter.

In the first iteration, with above phase noise estimate, the actual mitigation is done by dividing the time-domain waveform $\mathbf{y}^{(m)}$ by the complex exponential of the estimated phase noise. Then, symbols are detected. In later iterations, mitigated signal, namely $\mathbf{y}^{(m)}$ after the phase noise compensation, is fed back to the algorithm input instead of using $\mathbf{y}^{(m)}$ directly, as shown in Fig. 1. The remaining phase noise is then estimated and mitigated, and symbols are detected again.

B. Performance Improvements with Scaling

It is well known that OFDM signal has a very high peak-to-average power ratio (PAPR). Because of this, when we calculate the needed time-domain OFDM signal estimate $\hat{\mathbf{x}}^{(m)}$ for the algorithm, which we use as a point-by-point denominator in (9), the errors in very low amplitude parts of $\hat{\mathbf{x}}^{(m)}$ cause relatively high errors in the corresponding result vector $\boldsymbol{\vartheta}^{(m)}$. The low amplitude samples are also more prone to errors due to AWGN. It is thus beneficial to scale the signal in some part of the estimation process so that the samples more prone to cause errors are given less weight. The scaling can be done, e.g., before the low-pass filtering in (10). This is a good stage of the estimation process for the scaling because the convolving filtering operation would spread the potential error to adjacent samples. Furthermore, even though not marked in (10), the output signal from the low-pass filter is somewhat scaled since a practical highly-selective low-pass filter drains the power of the low-pass natured signal of interest. So at the same time with this scaling, the proposed scaling operation can be also be done.

There are many ways to do the scaling. The optimal scaling depends on the used system parameters. In this paper, we assume simple scaling with squared absolute values of the elements of $\hat{\mathbf{x}}^{(m)}$ divided by the mean power of the time-domain OFDM symbol. The latter can be easily calculated when the used OFDM system configuration is known. In addition, the filter dependent constant scaling is also taken into account at the same time. So the more practical form of (10) is then

$$\mathrm{LPF}\left\{\mathrm{diag}\left(\mathbf{q}\right)\arg\left(\boldsymbol{\vartheta}^{(m)}\right)\right\} \approx \boldsymbol{\varphi}^{(m)} - \hat{\varphi}_{CPE}^{(m)} = \hat{\boldsymbol{\varphi}}_{est,q}^{(m)}, \tag{11}$$

where $\mathbf{q}$ is the scaling vector optimized to correspond to the used filter and the used OFDM configuration. In the simulations later on in the paper, an example for $\mathbf{q}$ is given for the used filter and system parameters considered in this paper.

C. Further Thoughts on the Algorithm

The properties of the used low-pass filter have big impact on the obtained estimation performance. The filter must be optimized while keeping several things in mind. Basically to get the best possible performance, highly selective filter should be used. However, this means that a high-order filter must be used, which increases computational complexity of the overall algorithm and also increases the length of the transients on both ends of the filtered signal. There are many ways to design a highly selective filter with low computational complexity, but the transient problem cannot be averted but by lowering the selectivity of the filter. Anyway, as shown by the simulations later on this paper, very good performance is achievable, even though the transient problem remains.

One of the strengths of this algorithm is that we do not need any prior information about the phase noise in the estimation. Only an assumption of realistic phase noise process has to be made, namely the phase noise process must be heavily low-pass natured. This assumption holds for the free-running oscillator [2] but it also holds for more practical oscillators, such as oscillators using phase-locked loop (PLL) [13].

IV. Performance Simulations and Analysis

In this Section, the performance simulations are carried out and analyzed for OFDM direct-conversion link corrupted by phase noise on the transmitter and the receiver. First, the simulation parameters are given and then the performances of state-of-the-art phase noise mitigation techniques are compared to the performance of the proposed algorithm.

A. Simulation Parameters and Flow

Here, 3GPP Long Term Evolution (LTE) downlink –like OFDM communications link is simulated [14]. We use OFDM system with 1024 subcarriers and 15 kHz subcarrier spacing. Of these 1024 subcarriers, 600 are active 16QAM modulated, 300 on both sides of the centre subcarrier. Other subcarriers are zero-subcarriers. The used cyclic prefix length is 63 samples and 18 of the active subcarriers carry pilot data.

The flow of the simulations is as follows. First, 16QAM modulated symbols are generated. They are then OFDM modulated as described in (1). After this, the cyclic prefix is added, transmitter phase noise is modelled and the resulting signal is send through the communications channel. The channel is either an additive white Gaussian noise (AWGN) channel or an extended ITU-R Vehicular A (VEHA) multipath channel [15] depending on the studied case. In the VEHA case, the channel realizations are independent of each other. After going through the channel, the signal is impaired by the receiver phase noise, cyclic prefix is then removed and the resulting signal is OFDM demodulated. The transmitter and receiver phase noises are independent processes but for easier presentation of the results, they have the same 3 dB bandwidth (β) values. Actual β values reported in the figures below, are the sums of the individual transmitter and receiver β values, namely the total link phase noise.

In the actual impairment mitigation at the receiver, the channel is assumed to be known and it is compensated after the OFDM demodulation. After this, the CPE is estimated and mitigated with very simple least squares and pilot based approach described in [7] and [10]. Now, the phase noise mitigation algorithms to mitigate the ICI part of the phase noise are implemented. We use algorithms proposed in [4] (Petrovic), [9] (Bittner) and [10] (LI-TE) in addition to the proposed technique. Finally, the enhanced signals are detected and symbol-error rates (SER) are computed.

Parameters for the reference phase noise algorithms are the same that were used in [10], namely empirically optimized. For the proposed algorithm, parameters are as follows. For the low-pass filter, we use a selective filter of order 200 for the AWGN case and a filter of order 350 for the VEHA case. Both are designed using the well-known Remez algorithm. This is not the best way to design the filter, but the actual design and optimization are not on the scope of this paper and are left for another study. Prior to low-pass filter, we use a scaling vector

$$\mathbf{q} = \frac{\sqrt{2}\left|\hat{\mathbf{x}}^{(m)}\right|^2}{N_a / N}, \tag{12}$$

where N_a is the number of active subcarriers and N is the total number of subcarriers, namely N_a / N is the expected signal power of $\hat{\mathbf{x}}^{(m)}$. $\left|\hat{\mathbf{x}}^{(m)}\right|^2$ is a vector of squared absolute values of elements of $\hat{\mathbf{x}}^{(m)}$. This scaling assumes that the channel response is normalized to unity, which in practice can be achieved, e.g., with receiver power control. This scaling is good because it gives exponentially less stress on very low-amplitude signal values of the $\hat{\mathbf{x}}^{(m)}$, so the most noisy components of the phase noise estimate are given less weight, as explained earlier.

B. Simulation Results and Performance Analysis

1) Performance of the proposed algorithm

The simulation results for the additive white Gaussian noise channel case for all the studied mitigation techniques are depicted in Fig. 2 and Fig. 3. For Petrovic [4], Bittner [9], LI-TE [10] and the proposed technique, three iterations are used. As depicted in the Fig. 2, the proposed technique clearly outperforms the competition over the whole studied phase noise 3-dB bandwidth region from 0 to 1500 Hz with a fixed SNR of 18 dB. The same conclusions can be made from Fig. 3, in case of fixed β of 350 Hz over the received SNR region from 0 to 30 dB. The proposed technique actually gives almost ideal performance up until 17 dB of received SNR, and then starts to floor. However, the floor is significantly lower than the floor of the reference techniques.

In the more challenging case of extended ITU-R Vehicular A multipath channel, the performance differences change a little as can be seen from Fig. 4 and Fig. 5. As Fig. 4 shows, when system performance is capped by low effective SNR and not by the high phase noise, namely when the phase noise level in the system is very small, the proposed technique cannot reach the performance of the LI-TE. However, the performance differences in such cases are hardly visible, and after β of 200 Hz, the scales are turned for the advantage of the proposed technique. After β of 350 Hz, the proposed technique already outperforms the reference techniques clearly up until the end of the studied region at 1500 Hz. With higher SNR values the proposed technique performs much better also with very low phase noise 3-dB bandwidths as the additive noise is not the dominating noise source in the system. The AWGN case gave a good example of this. From Fig. 5, we can see that when we have fixed β of 350 Hz, the proposed technique outperforms the competitors already after received SNR of 10 dB, and gives very nice performance up until the end of the studied region at 40 dB of received SNR.

Overall, the proposed technique performs very well, and only suffers from very bad noise conditions. In such cases the performance differences are however almost unnoticeable between the studied techniques. Furthermore, in very low-SNR conditions, all the techniques almost achieve the no phase noise performance limit, so the phase noise is indeed dominated by the additive noise.

2) *Iterative performance of the proposed algorithm*

Fig. 6 and Fig. 7 demonstrate how the proposed technique scales up with the increased number of iterations. The scaling of the proposed technique is compared to the best performing reference technique, LI-TE. As depicted in Fig. 6, already two iterations of the proposed technique outperform five iterations of LI-TE technique. Furthermore, the scalability of the proposed technique stays high even with higher number of iterations. There is a clear performance gain in higher β region when comparing the five and eight iterations cases. The same conclusions can be drawn from Fig. 7, where fixed β of 350 Hz is used and simulations are run over the SNR region.

V. Conclusions

OFDM transceivers suffer heavily from phase noise. We proposed a new phase noise mitigation algorithm to address this problem taking into account both the transmitter and receiver phase noise sources. In the proposed algorithm, time-domain estimation of the phase noise is done by exploiting the steep low-pass natured spectrum of the phase noise process in

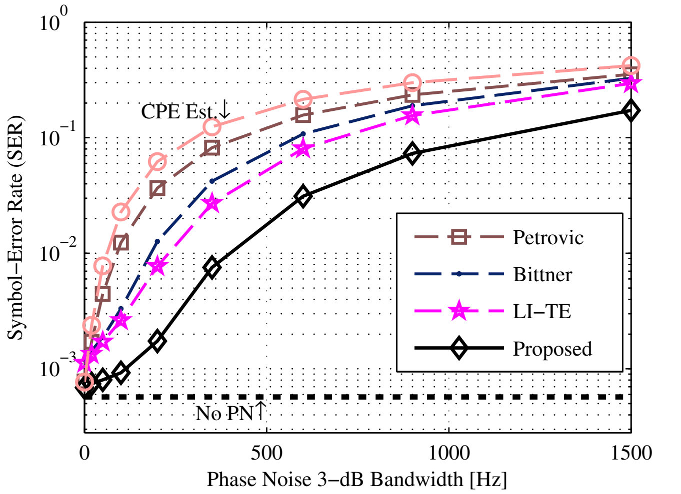

Fig. 2. Simulated SER as a function of β. AWGN channel is used with fixed received SNR of 18 dB.

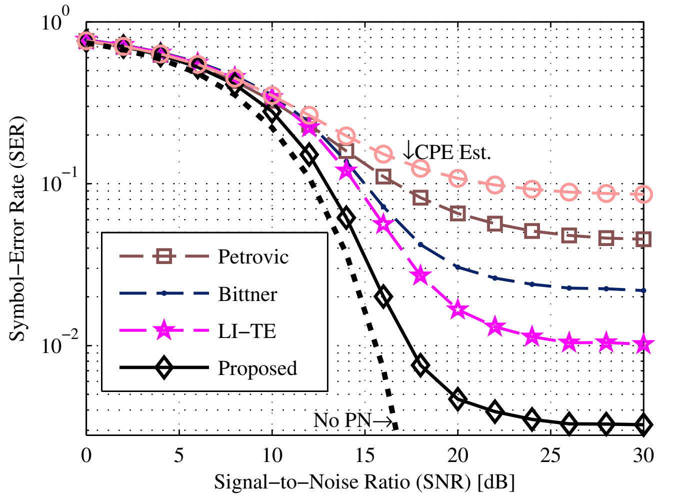

Fig. 3. Simulated SER as a function of received SNR in AWGN channel. Fixed β of 350 Hz is used.

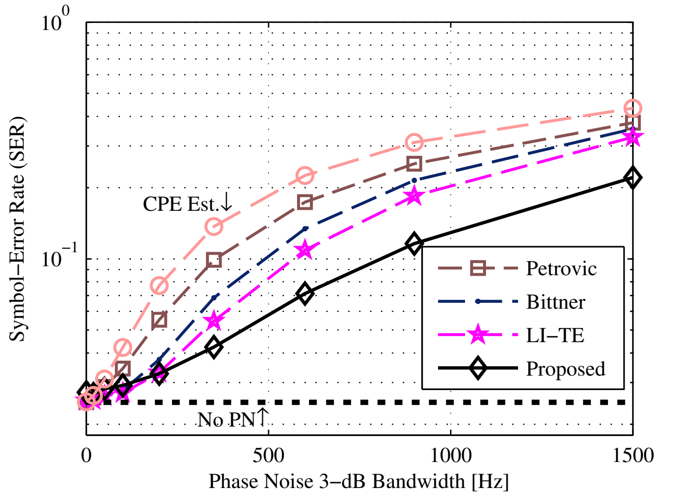

Fig. 4. Simulated SER as a function of β. Extended ITU-R Vehicular A multipath channel is used with fixed received SNR of 24 dB.

iterative manner using also the initial symbol decisions. This allowed accurate estimation of the phase noise without any prior knowledge of the statistics of the phase noise process. The performance of the algorithm was compared to the performances of the state-of-the-art phase noise mitigation techniques from the literature. The performance of the proposed algorithm was very good. It outperformed the reference methods in additive white Gaussian noise channel case as well as in extended ITU-R Vehicular A multipath channel case.

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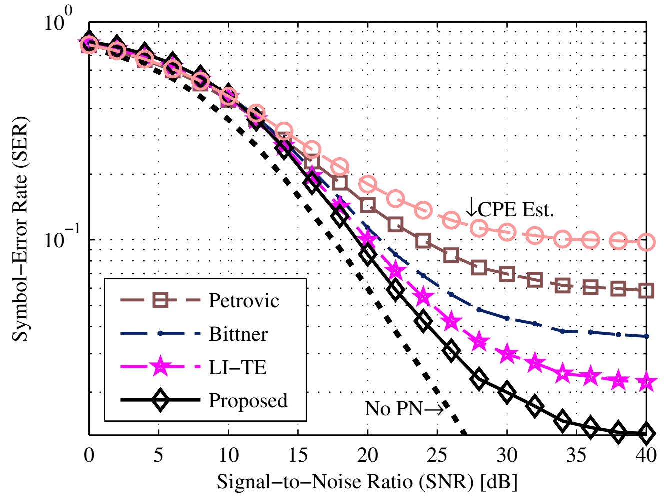

Fig. 5. Simulated SER as a function of SNR in extended ITU-R Vehicular A multipath channel. Fixed β of 350 Hz is used.

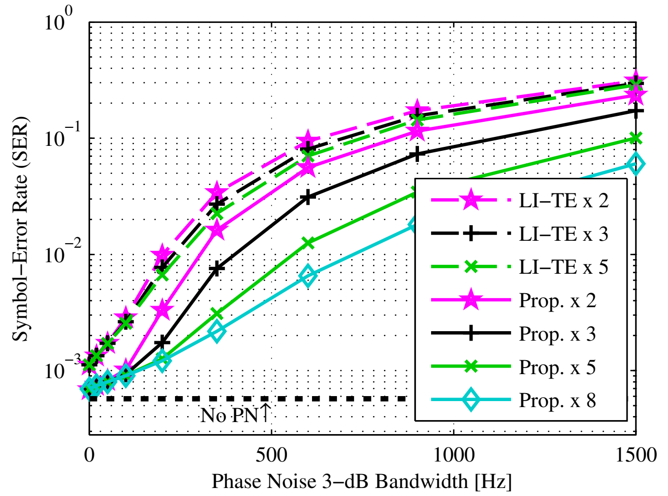

Fig. 6. Simulated SER as a function of β for different levels of iteration for the proposed technique and the best performing reference technique. AWGN channel with fixed received SNR of 18 dB is used.

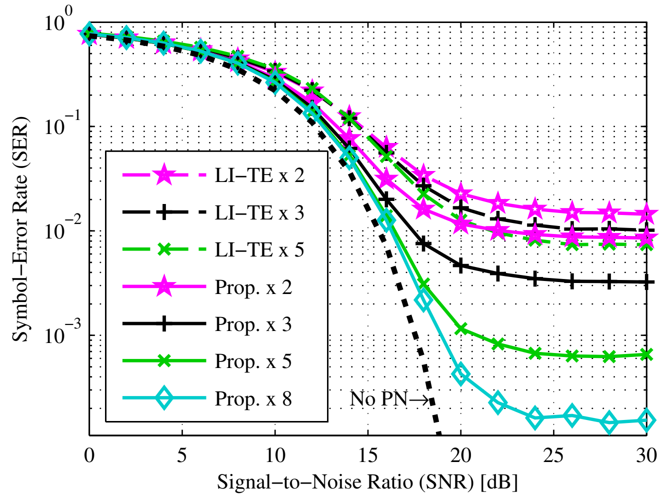

Fig. 7. Simulated SER as a function of received SNR for different levels of iteration for the proposed technique and the best performing reference technique. Fixed β of 350 Hz is used.