Jitter Mitigation in High-Frequency Bandpass-Sampling OFDM Radios

Ville Syrjälä and Mikko Valkama

Tampere University of Technology
Department of Communications Engineering
Korkeakoulunkatu 1, FI-33720 Tampere, Finland
E-mail: {ville.syrjala, mikko.e.valkama}@tut.fi

***Abstract*—This paper presents a new way to address and mitigate sampling jitter in high-frequency bandpass-sampling OFDM radio receivers. Baseband model for mapping the sampling jitter to certain type of phase noise is first presented, and stemming from this model, state-of-the-art phase noise mitigation techniques are then proposed to remove the jitter-induced signal distortion. Performances of the proposed jitter mitigation techniques are analyzed with extensive computer simulations in high-speed bandpass sampling multicarrier system context. In the link performance simulations, both additive white Gaussian noise (AWGN) and extended ITU-R vehicular A multipath (eVehA) radio channel types are used, combined with realistic sampling clock and jitter modelling.**

***Index Terms*—Bandpass sampling, sampling jitter, phase noise, OFDM, intercarrier interference.**

I. Introduction

In modern communications receivers, maximizing flexibility and re-configurability while minimizing power consumption, size and costs are some of the main design criteria. When signals are sampled directly at radio frequencies (RF), very high flexibility and re-configurability can be achieved. Also, the number of analogue components is generally decreased, compared to many other radio architectures, thus resulting in potentially lower power consumption, size and costs.

Still, many problems remain in direct RF sampling receiver concept [4], [7]. With current technologies, maximum sampling frequencies are limited to a few tens or few hundreds of MHz range. The used carrier-frequencies, on the other hand, are typically in the GHz range, which leads to bandpass sub-sampling concepts [4], [14]. This, in turn, sets relatively high demands for RF band-limitation filtering, prior to sampling, to control harmful aliasing effects. In addition to filtering requirements, also the power consumption of the sampling circuitry itself can be somewhat higher, compared to ordinary low-frequency/baseband sampling approaches. This is because the usable sub-sampling frequencies depend essentially on both the signal bandwidth and centre-frequency, and are typically somewhat higher than the corresponding minimum sampling rates in ordinary low-frequency sampling [14]. Furthermore, as the frequency range of the sampled signals increases, the effects of timing inaccuracies or jitter in the sampling process increase as well. In fact, these timing inaccuracies start to noticeably limit the overall receiver and system performance when sampling signals directly at commonly used carrier frequencies [1], [8]. Thus, there is a clear need to study and mitigate the effects of timing jitter in such high-speed bandpass sampling radios.

The impact of sampling jitter on radio system performance has been recently studied in [5], assuming orthogonal frequency-division multiplexing (OFDM) waveforms. Furthermore, the essential signal-to-noise ratio (SNR) reduction due to jitter has been addressed using system calculation principles, e.g., in [8], [11]. This paper, in turn, shows that with fairly reasonable assumptions on received signal and sampling clock characteristics, jitter noise acts similarly as phase noise in bandpass sampling based receivers. Based on this modelling, this paper then also proposes the use of sophisticated phase noise mitigation schemes for jitter mitigation. As a practical example, OFDM waveforms are assumed also in this paper since they are generally found very sensitive to any phase noise –like phenomena [13], [16].

The organization of the rest of the paper is as follows: In Section II, efficient and accurate jitter model is presented for bandpass sampling receivers. In addition, Section II describes how the model essentially maps sampling jitter to phase noise, and thereon to intercarrier interference (ICI) in case of OFDM. Then, Section III describes state-of-the-art techniques for phase noise mitigation to reduce the jitter effects in direct-sampling OFDM receivers. Section IV, in turn, describes the used simulation model and demonstrates the obtained mitigation performance in realistic 3GPP Long Term Evolution (LTE) –type multicarrier radio system context. Finally, Section V concludes the work.

This work was supported by the Finnish Funding Agency for Technology and Innovation (Tekes, under the project "Advanced Techniques for RF Impairment Mitigation in Future Wireless Radio Systems"), the Academy of Finland, the Technology Industries of Finland Centennial Foundation, Finnish Foundation for Technology Promotion, and TUT Graduate School.

II. Jitter Modelling and OFDM Radios

In this Section, a simple yet efficient model for mapping sampling jitter to phase noise in high-frequency bandpass sampling receivers is presented. In addition, using this modelling, jitter impact on direct-sampling OFDM receivers is studied in detail.

A. Jitter Modelling and Connection to Phase Noise

Stemming from the earlier work in [10], we present a simple jitter model for bandpass sampling based radio receivers. The model exploits the general features of bandpass communications waveforms. In general, an arbitrary received bandpass signal is first written as

$$r(t) = x_I(t)\cos(2\pi f_c t) - x_Q(t)\sin(2\pi f_c t)\,, \tag{1}$$

where x_I and x_Q are the I and Q components of the received signal, respectively, and f_c is the corresponding formal centre-frequency. Now, if the jitter is taken into account, we end up with a signal of the form

$$\begin{aligned} r\{t+\xi(t)\} = x_I\{t+\xi(t)\}\cos\{2\pi f_c[t+\xi(t)]\} \\ -x_Q\{t+\xi(t)\}\sin\{2\pi f_c[t+\xi(t)]\}, \end{aligned} \tag{2}$$

where $\xi(t)$ models the uncertainty on the time axis due to jitter. Now, assuming that the centre-frequency f_c of the incoming signal $r(t)$ is large compared to the corresponding bandwidth of the signal, the jitter contribution to the I and Q components is much lower compared to the contribution on the high-frequency carrier components. Furthermore, assuming that the jitter values $\xi(t)$ are relatively small compared to the essential time-dynamics of the I and Q components, it follows that $x_I\{t+\xi(t)\} \approx x_I(t)$ and $x_Q\{t+\xi(t)\} \approx x_Q(t)$, and we can essentially approximate (2) as

$$\begin{aligned} r\{t+\xi(t)\} \approx x_I(t)\cos\{2\pi f_c[t+\xi(t)]\} \\ -x_Q(t)\sin\{2\pi f_c[t+\xi(t)]\}. \end{aligned} \tag{3}$$

Notice that at this stage, no other assumptions have been made yet on the more detailed structure of the modulating I and Q components. Depending on the RF filtering prior to sampling, these can, e.g., contain several radio signals or channels on different carrier-frequencies.

In practice, the above assumption of small jitter values can be seen reasonable with practical processing bandwidths in the order of a few tens of MHz and with jitter RMS values in the tens of ps range. Such relatively small jitter values, in turn, are realistic in many practical sampling circuitries, utilizing some sophisticated (e.g., phase-locked loop, PLL, based) oscillators in generating the sampling instants. For an example spectral characteristics, see Fig. 1. On the other hand, the small jitter approximation does not necessarily hold for free-running type oscillators where the oscillator phase variance, and thereon the variance of the sample instants, increases over time, and thus relatively high jitter values are also possible.

In the actual sampling process, we observe (1) at time instants $t_n = nT_S + \xi_n$, where $\xi_n = \xi(nT_S)$ denotes the time-deviation of the n-th sample instant from the nominal sample grid nT_S, T_S is the time between adjacent samples (in the corresponding ideal sample stream) and sample index $n \in \mathbf{N}$. Thus, based on the approximation in (3), the n-th sample can now be written as

$$\begin{aligned} r_n &= r(nT_S + \xi_n) \\ &\approx x_I(nT_S)\cos[2\pi f_c(nT_S + \xi_n)] \\ &\quad -x_Q(nT_S)\sin[2\pi f_c(nT_S + \xi_n)]. \end{aligned} \tag{4}$$

Stemming from the bandpass sub-sampling principle, the signal is aliased in a controlled manner to an intermediate frequency (IF), i.e., $\cos(2\pi f_c nT_S) = \cos(2\pi f_{\mathrm{IF}} nT_S)$ and $\sin(2\pi f_c nT_S) = \sin(2\pi f_{\mathrm{IF}} nT_S)$ [4], [7], [14]. Now, using simple manipulations, the above signal can also be written as

$$r_n \approx \mathrm{Re}\{[x_I(nT_S) + jx_Q(nT_S)]e^{j2\pi f_{IF} nT_S} e^{j2\pi f_c \xi_n}\}\,. \tag{5}$$

Thus, the final baseband observation after digital IF-to-baseband conversion is given by

$$\begin{aligned} y_n &\approx [x_I(nT_S) + jx_Q(nT_S)]e^{j2\pi f_c \xi_n} \\ &= x_n e^{j2\pi f_c \xi_n} \\ &= x_n e^{j\phi_n}\,, \end{aligned} \tag{6}$$

where $x_n = x_I(nT_s) + jx_Q(nT_s)$ and $\phi_n = 2\pi f_c \xi_n$. Based on the above signal model, sampling jitter results in time-varying excess phase fluctuations in the observed low-frequency signal and can thus be essentially modelled as phase noise in high-frequency bandpass sub-sampling receivers.

B. Jitter Impact on Direct-Sampling OFDM Receiver

Next, we focus on interpreting the above signal model (6) from direct-sampling OFDM receiver point of view. We further assume that the RF filtering stage prior to sampling attenuates the neighbouring channel signals. Then, the signal x_n corresponds to the received OFDM waveform samples, and the jitter contribution is modelled into the system according to (6). Denoting the transmit waveform samples by s_n, the received samples are given by $x_n = h_n \star s_n + z_n$ where h_n represents the radio channel impulse response, $\star$ denotes convolution and z_n represent additive noise. Combining then this signal model and the earlier jitter model in (6), the observed signal after receiver FFT within m-th OFDM symbol can be written as

$$\begin{aligned} Y_m(k) = S_m(k)H_m(k)J_m(0) \\ +\sum_{n=0}^{N-1} S_m(l)H_m(l)J_m(k-l) + Z_m(k). \end{aligned} \tag{7}$$

Here $S_m(k)$ denotes the transmit symbol at k-th subcarrier during m-th OFDM symbol interval, $H_m(k)$ is the corresponding channel transfer function, $J_m(k)$ is the Fourier transform of the complex exponential of the jitter phase noise values during the m-th OFDM symbol and finally $Z_m(k)$ is the FFT of the AWGN term. In deriving (7), it has been further assumed that the transmit signal contains a cyclic prefix longer than the channel delay spread, being then properly discarded in the receiver prior to FFT. For reference, see e.g. [15] where similar analysis is carried out assuming ordinary oscillator phase noise.

The signal model presented in (7) is important as it divides the jitter contribution to two essential parts, i.e., common phase error (CPE) and intercarrier interference (ICI) parts [15]. CPE is common phase rotation within an OFDM symbol for all the subcarriers, where as ICI is the distortion that neighbouring subcarriers cause to each other due to spectral spread caused by the jitter.

III. Jitter Estimation and Mitigation

Stemming from (6) and (7), jitter mitigation in direct-sampling OFDM radios can be accomplished by estimating and cancelling the CPE and ICI caused by jitter-induced phase noise. In this Section, state-of-the-art phase noise estimation techniques used then for jitter mitigation are shortly presented. In short, CPE and ICI estimation corresponds to estimating the essentially non-zero spectral components $J_m(k)$ of the complex exponential of the phase noise. Since the jitter-induced phase noise changes from OFDM symbol to another, this estimation needs to be carried out for each symbol interval. In general, after estimating the essential phase noise spectral components, the actual mitigation is done by circular deconvolution. This results directly from (7) in which phase noise is seen as circular convolution.

A. CPE Estimation

CPE estimation used in this paper is based on the work in [15]. CPE estimation is conceptually easy because CPE causes each subcarrier symbol within an OFDM symbol duration to be multiplied by the same complex multiplier. We can then use e.g. least squares estimation to estimate the CPE contribution of the phase noise with the help of pilot symbols and channel information [15].

B. ICI Estimation Using CPE Interpolation (LI-CPE)

Plain CPE estimation gives us effectively constant estimate for the phase noise within the duration of an individual OFDM symbol. In effect, this is the average value of the true phase noise within one OFDM symbol, so the CPE estimation gives very simplified phase noise estimate. So-called LI-CPE approach [12] is then a very simple method to improve the phase noise estimation performance by linearly interpolating between the CPE estimates of adjacent OFDM symbols. This is done essentially so that, in time domain, a straight line is "drawn" between the CPE estimates at the middle of adjacent OFDM symbols – from the middle of symbol m-1 to the middle of symbol m and from the middle of symbol m to the middle of symbol m+1. In addition, the first spectral component (DC-bin) of this new phase noise estimate inside one OFDM symbol is replaced by the original CPE estimate, because it is likely to be better than the first spectral component of the linear interpolated one. A graphical illustration is given in Fig. 2. For more details, refer to [12].

C. Iterative ICI Estimation Using Tail Interpolation (LI-TE)

Conceptually, the phase noise spectral components can be solved from (7) if both the channel and the transmit symbols are known. In iterative ICI estimation [3], [6], the idea is then to use initial symbol decisions obtained by cancelling first only the CPE, in solving (7) for $J_m(k)$'s (or at least a subset of essentially non-zero ones). Then ICI is cancelled with this estimated ICI profile and the signal is detected again. Altogether this whole procedure can be then repeated to iteratively improve the quality of both detection and ICI estimation. In LI-TE technique [12], this iterative procedure is the starting point. LI-TE method then further improves the quality of the ICI estimates at the borders of each OFDM symbol. This can be done because the original ICI estimation method gives very poor estimates at the border areas of OFDM symbols [3], [6]. LI-TE uses then linear interpolation to obtain better estimates for the phase noise realization over the unreliable region of the original estimate [12]. The interpolation of LI-TE at the OFDM symbol borders is also visible in the example of Fig. 2.

IV. Performance Simulations

In this Section, OFDM link simulation model is first described and then the obtained simulation results are presented and analyzed.

A. Simulation Model

In performance simulations, we use 3GPP LTE downlink – like system [1] operating at so called "2.6 GHz band" (i.e., at 2500-2690 MHz) as a practical example scenario. First we generate OFDM signal with 1024 subcarriers using 15 kHz subcarrier spacing. 600 of these 1024 subcarriers are active, and for these, 16QAM subcarrier modulation is used. This corresponds to 1024×15 kHz = 15.36 MHz transmitter sampling rate for baseband waveform generation, and the RF waveform bandwidth is roughly 10 MHz. Cyclic prefix of 63 samples is then added to the signal, which corresponds to around 4 μs maximum delay spread for the radio channel. This signal is then put through either (*i*) plain AWGN channel or (*ii*) extended ITU-R Vehicular A multipath (eVehA) [9] channel. In the eVehA channel case, the channel impulse response is assumed to be static for a period of 12 OFDM symbols after which a totally new channel realization is drawn (quasi-static simulation approach). In the receiver model, the initial sampling rate for modelling the bandpass sampling stage is assumed to be 138.24 MHz which corresponds to 9 times over-sampling factor compared to basic waveform sample rate of 15.36 MHz. In this sampling process, the jitter contribution is modelled according to the low-frequency model (6), using an assumed example RF carrier frequency of 2661.12 MHz, fitting well in the planned LTE downlink band 7 [1]. This assumed example centre-frequency and the used sampling frequency effectively alias the signal to an IF of 2661.12 – 19×138.24 = 34.56 MHz in a controlled manner.

For sampling clock modelling, we use phase-locked loop (PLL) oscillator model presented in [12]. This models the oscillator phase noise characteristics by taking both white and flicker noise contributions into account. The resulting jitter behaviour is in general normalized to different practical RMS-values in the ps range, as will be illustrated in the following sub-section. An example of the spectral characteristics of non-normalized oscillator is depicted in Fig. 1. After this, the detection of symbols is done with help of state-of-the-art phase noise mitigation techniques described in Section III. We use basic CPE estimation scheme, linear interpolation based ICI estimation (LI-CPE), and basic ICI estimation [5] enhanced with linear interpolation based tail-estimation technique (LI-TE). For more details, refer to [5], [12], [15].

B. *Obtained Results*

In the simulations, the obtained jitter mitigation performances using the above phase noise mitigation techniques are evaluated, in terms of detection error rate, and compared against reference cases with (*i*) no jitter (jitter-free sampling) and (*ii*) no jitter removal. For practicality, both the AWGN and eVehA channels are simulated.

First, we visually demonstrate the jitter estimation performance in time domain. An example result is shown in Fig. 2. There, the true jitter realization is plotted in same figure with the corresponding LI-CPE and LI-TE estimation results. As can be seen in the figure, LI-CPE and LI-TE give very high-quality estimates.

The results of simulated receiver symbol-error rate (SER) in AWGN channel case can be seen in Fig. 3 and Fig. 4. Based on Fig. 3, it can be seen that with 15 ps RMS jitter, one can get very nice performance improvement with phase noise mitigation techniques. The non-mitigated signal, in turn, suffers very heavily from the jitter. Even basic CPE mitigation gives relatively good performance improvements. The LI-CPE method further improves the performance of the system, while the LI-TE method removes almost all the jitter distortion in the interesting SNR region. From Fig. 4, very similar conclusions can be made. When SNR is set to an example value of 18 dB, the LI-TE gives almost perfect jitter estimation up to 15-20 ps RMS jitter values. With higher jitter values, the mitigation performance starts to decrease, but still with greatly increased RMS jitter values even in the order 30-50 ps, relatively good detection performance is obtained.

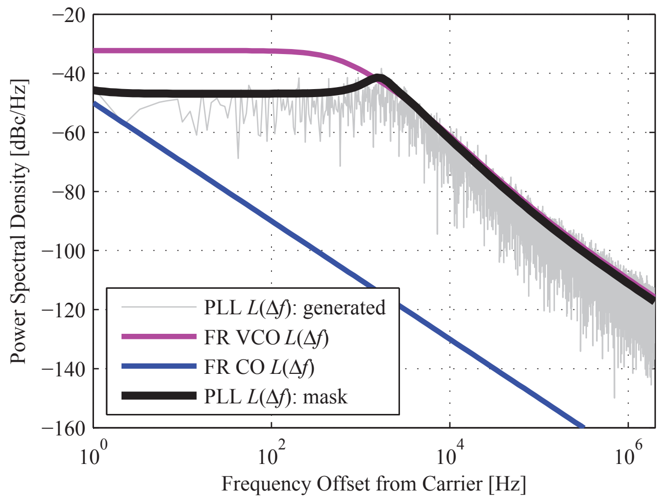

Fig. 1. PLL output phase noise characteristics [12]. Spectra of reference Crystal Oscillator (CO, dominating below 2 kHz) and Voltage Controlled Oscillator (VCO, dominating above 2 kHz) are presented.

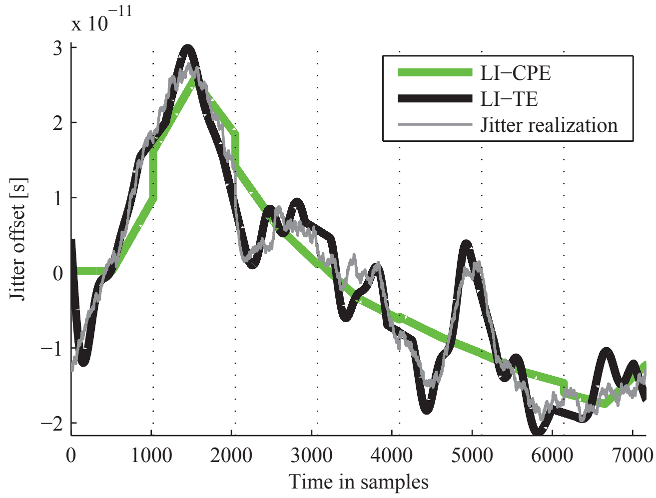

Fig. 2. An example of true jitter realization and the obtained estimates with 20 ps RMS jitter, plotted for a period of 7 OFDM symbols separated by dotted lines. 18 dB received SNR and AWGN channel.

For eVehA channel case, the obtained simulation results can be seen in Fig. 5 and Fig. 6. From these, we can see that when more challenging channel conditions are experienced, the phase noise mitigation techniques still give relatively similar performance as in the AWGN channel case. From Fig. 5, we can see that for an example RMS jitter of 15 ps, LI-TE method mitigates the jitter effects almost perfectly. In addition, LI-CPE method gives also very good performance even in high-SNR region. Furthermore, even the basic CPE mitigation yields considerable performance improvement over the no mitigation case. The Fig. 6 shows further that in the eVehA channel case, the LI-TE method gives almost perfect jitter estimation even up to 20-25 ps RMS jitter range. In addition, the simpler estimation methods also perform relatively well as can be seen in the figure.

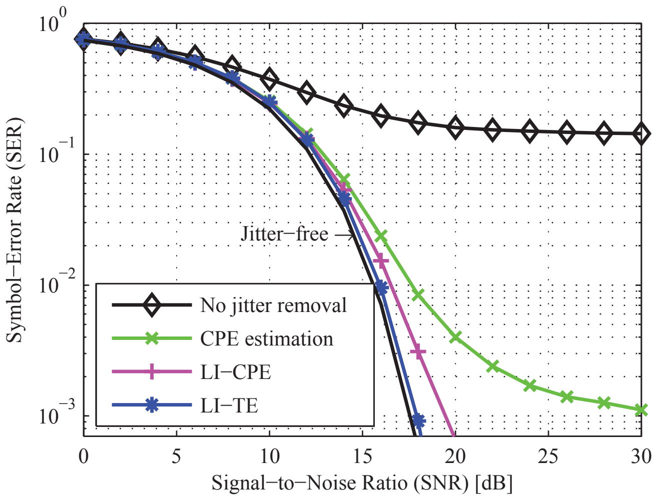

Fig. 3. Simulated SER as a function of SNR in LTE-type OFDM system with direct-sampling receiver. AWGN channel, 15 ps RMS jitter and 16QAM subcarrier modulation are used.

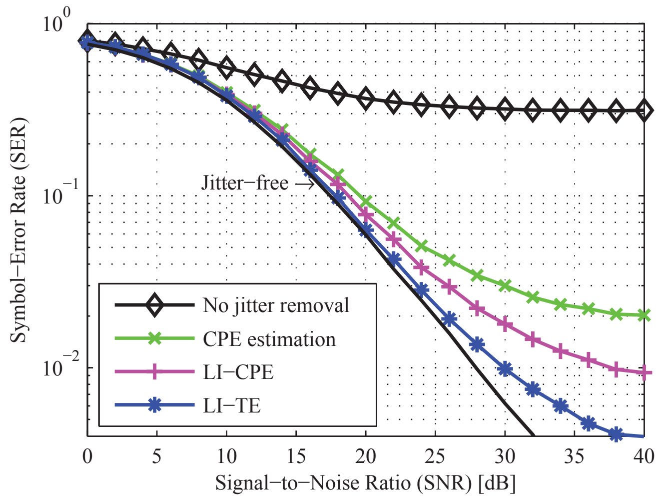

Fig. 5. Simulated SER as a function of SNR in LTE-type OFDM system with direct-sampling receiver. Extended ITU-R Vehicular A multipath channel, 15 ps RMS jitter and 16QAM subcarrier modulation are used.

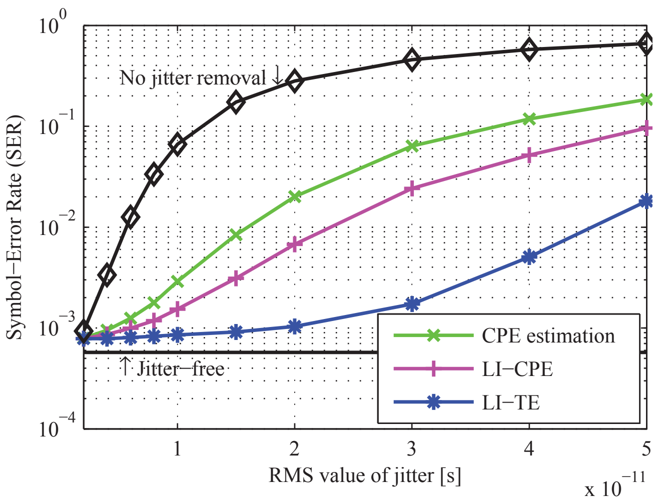

Fig. 4. Simulated SER as a function of RMS jitter with 18 dB received SNR in LTE-type OFDM system with direct-sampling receiver. AWGN channel and 16QAM subcarrier modulation are used.

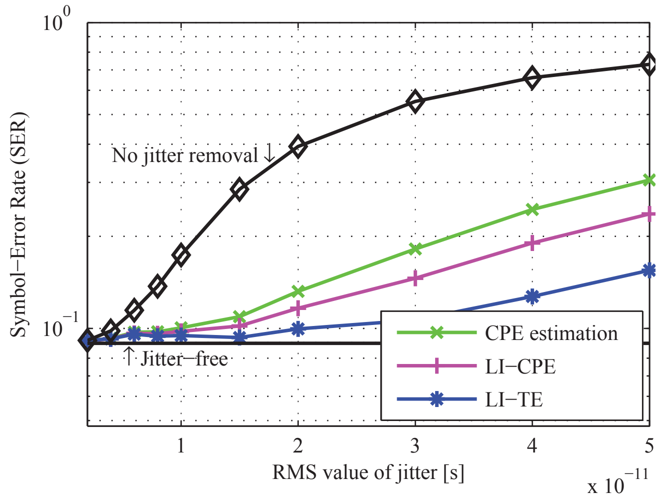

Fig. 6. Simulated SER as a function of RMS jitter with 18 dB received SNR in LTE-type OFDM system with direct-sampling receiver. Extended ITU-R Vehicular A multipath channel and 16QAM subcarrier modulation are used.

V. Conclusions

Sampling jitter causes noticeable performance decrease in high-frequency bandpass-sampling receivers in general. In this paper, we first derived a baseband model, which essentially maps the jitter to certain type phase noise. This allows us to use state-of-the-art phase noise mitigation techniques to reduce the jitter effects in RF sampling receivers. The link simulations showed that, in OFDM systems, using advanced phase noise mitigation techniques, huge performance improvements are generally achieved over the case with no jitter removal, with both AWGN and extended Vehicular A multipath channel types. With the best mitigation method, the so-called LI-TE technique, we got almost perfect jitter mitigation performance with reasonable jitter RMS values in the order of 10 – 20 ps.

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