Sampling Jitter Cancellation in Direct-Sampling Radio

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Abstract—**This paper addresses the sampling jitter estimation and cancellation task in direct RF sub-sampling type radios. The proposed jitter estimation method is based on carefully injecting or superimposing an additional known reference signal to the received signal at the sampler input. Proper digital signal processing methods are then devised and applied to estimate the sampling jitter realizations from the obtained jittered samples. Using these jitter estimates, combined with proper jitter modelling, the jitter effects can then be efficiently removed from the actual received signal. Careful performance analysis of the overall estimation-cancellation scheme is also carried out using computer simulations with 3GPP LTE type multicarrier signals, assuming also different amounts of RF filtering prior to RF sub-sampling stage. In the performance simulations, both additive white Gaussian noise and extended ITU-R Vehicular A multipath radio channel types are considered.**

Keywords-sampling jitter; mitigation; dirty-RF; RF-DSP; direct RF sampling; bandpass sampling; sub-sampling

I. INTRODUCTION

In sampling of high-frequency signals, or signals with powerful neighbouring channels, relatively high requirements are set for the timing accuracy of the sampling process [7], [9], [15], [16]. In traditional radio receivers, relatively low-frequency signals are sampled because of signal down-conversion prior to sampling. In addition, most of the interference near the interesting frequencies is attenuated. These result into relatively relaxed requirements for the dynamics and timing accuracy of the sampling and A/D-conversion circuitry in current receivers. However, when emphasizing radio flexibility and re-configurability, more and more of the selectivity filtering is moved to digital domain [4]-[7]. Similarly from the frequency-translations point of view, applying sampling to higher-frequency signals in the overall receiver signal-processing chain is one of the main trends currently. Under these working assumptions, timing inaccuracies in the sampling process, called sampling jitter, can become a severe problem [2], [4]-[8], [15], [16]. Thus, for the future radio receivers, understanding, modelling and mitigation of the sampling jitter is a very interesting and important topic.

In the so-called direct RF sampling (DRFS) receiver architecture, very high-frequency signals are sampled in conditions with possibly high-power interferers present [6], [7], [13]. DRFS is a receiver architecture in which the sampling of the incoming signal is done already at the radio frequencies (RF). DRFS reduces the amount of analogue components in the receiver, thus emphasizing re-configurability and also potentially minimizing size, power consumption and costs of the receiver compared to more traditional radios. However, the DRFS concept still has many practical implementation issues to be solved [7], [13], and it is thus not commonly considered feasible for mobile terminal receivers with today's implementation technology. The most notable problem with the DRFS approach is that it indeed poses very high demands for the quality of the sampling process. With current technologies, the combined requirements of timing accuracy and relatively high sampling frequency and resolution for the used sampling and A/D-conversion circuitry, result into relatively high power consumption. High power consumption in turn is one of the main concerns in mobile terminals. To circumvent this, digital signal processing (DSP) methods are developed in this paper to lower the quality requirements for the sampling process.

In the literature, DSP-based estimation and mitigation of sampling jitter is not too widely researched yet. This is partially because jitter is inherently a random process varying from sample instant to another and thus its estimation and mitigation at sample level is challenging. Also in most of today's receiver architectures, sampling takes place at fairly low frequencies, and thus jitter is not a major concern. However, as indicated earlier, this might not be the case in near future, since minimizing the amount of analogue parts in receivers while increasing flexibility becomes more and more emphasized. Furthermore, DSP capabilities of the mobile devices are continuously increasing. Recently, in [11], use of phase noise mitigation techniques [14] in mitigation of sampling jitter in bandpass sampling receivers was proposed for orthogonal frequency division multiplexing (OFDM) systems. In addition, the authors of [8] have proposed a technique to remove sampling jitter effects from general narrowband signals with help of a reference tone. In this paper, the idea of carefully designing and injecting a reference tone on top of the received signal at the sampler input is further developed. This paper first proposes an efficient technique to estimate the sampling jitter realizations with the help of such reference tone. The estimates of the jitter realizations are then used on-line to mitigate the effects of the sampling jitter from the

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actual received sampled waveform, without the limitations of narrowband signals and small jitter values as was assumed in [8]. Special emphasis in modelling and algorithm development is put to the DRFS receiver case where only partial selectivity is implemented at RF prior to sampling.

After this Section, the paper layout is as follows. Section II proposes the method to estimate sampling jitter values with the help of reference tone. Section III then proposes ways to mitigate the sampling jitter in DRFS system by using the jitter estimates. In Section IV the simulation environment is described and the corresponding simulation results and analysis are given in Section V. Conclusions for the work in this paper is finally given in Section VI.

II. Sampling Jitter Estimation Using a Sampled Reference Tone

In this Section, we propose an efficient technique to estimate the sampling jitter realizations from sampled jittered signal for direct sub-sampling radios. This is done by using the idea of injecting a reference tone to the incoming signal prior to the sampling circuitry [8], combined with proper digital processing of the jittered samples. These jitter estimates are then used with further digital signal processing to cancel the jitter effects from the actual received waveform samples in Section III.

A. Jittered Sub-Sampling Signal Model

First, let $r(t)$ denote the incoming received RF signal, written here as a general bandpass signal of the form

$$\begin{aligned} r(t) &= z_I(t)\cos[2\pi f_c t] - z_Q(t)\sin[2\pi f_c t] \\ &= A_r(t)\cos[2\pi f_c t + \theta_r(t)] \end{aligned}, \tag{1}$$

where $z(t) = z_I(t) + jz_Q(t) = A_r(t)e^{j\theta_r(t)}$ denotes the composite baseband equivalent of the overall received signal. Notice that depending on the amount of RF filtering, this may contain also several neighbouring channels, in addition to the target signal. Now the idea is to inject a reference tone, say $A_{ref}\cos(2\pi f_{ref}t)$, on top of $r(t)$ where f_{ref} denotes the reference tone frequency. This is illustrated in Fig. 1. Here sinusoidal reference signal is used, but in principle also other kinds of known signals could be used, if considered beneficial. Now, with suitable reference tone injected, the sampled signal is given by

$$\begin{aligned} r_n &= r(nT_s + \zeta_n) + A_{ref}\cos\left[2\pi f_{ref}(nT_s + \zeta_n)\right] \\ &= z_I(nT_s + \zeta_n)\cos\left[2\pi f_c(nT_s + \zeta_n)\right] \\ &\quad - z_Q(nT_s + \zeta_n)\sin\left[2\pi f_c(nT_s + \zeta_n)\right] \\ &\quad + A_{ref}\cos\left[2\pi f_{ref}(nT_s + \zeta_n)\right] \end{aligned}, \tag{2}$$

where $t_n = nT_s + \zeta_n$ denote the jittered sample instants, $F_s = 1/T_s$ is the nominal sampling frequency, and ζ_n denote jitter. Taking next the sub-sampling principle into account, which aliases the signal into lower frequencies in a controlled manner, this sample stream can also be written as

$$\begin{aligned} r_n &= z_I(nT_s + \zeta_n)\cos\left[2\pi f_{IF}nT_s + 2\pi f_c\zeta_n\right] \\ &\quad - z_Q(nT_s + \zeta_n)\sin\left[2\pi f_{IF}nT_s + 2\pi f_c\zeta_n\right], \\ &\quad + A_{ref}\cos\left[2\pi f'_{ref}nT_s + 2\pi f_{ref}\zeta_n\right] \end{aligned} \tag{3}$$

where f_{IF} and f'_{ref} denote the aliased centre-frequency and (possibly) aliased reference frequency, respectively. Depending on the choice of the reference frequency f_{ref} relative to sampling frequency F_s, $f'_{ref} < f_{ref}$ (controlled aliasing) or $f'_{ref} = f_{ref}$ (no aliasing).

B. Sampling Jitter Estimation

Based on (3), the main effect of jitter is that it causes phase noise to the composite I/Q carriers of the incoming signal as well as to the reference tone. Now, given that the aliased reference frequency f'_{ref} is outside the band of the aliased received signal (at f_{IF}), we can use complex digital mixing and lowpass filtering (LPF) to isolate the jittered reference tone. First we write (3) as

$$\begin{aligned} r_n &= \\ &\mathrm{Re}\{[z_I(nT_s + \zeta_n) + jz_Q(nT_s + \zeta_n)]e^{j2\pi f_{IF}nT_s}e^{j2\pi f_c\zeta_n}\} \\ &+ A_{ref}\,\mathrm{Re}\{e^{j2\pi f'_{ref}nT_s}e^{j2\pi f_{ref}\zeta_n}\} \\ &= \mathrm{Re}\{z(nT_s + \zeta_n)e^{j2\pi f_{IF}nT_s}e^{j2\pi f_c\zeta_n}\} \\ &\quad + A_{ref}\,\mathrm{Re}\{e^{j2\pi f'_{ref}nT_s}e^{j2\pi f_{ref}\zeta_n}\} \end{aligned}. \tag{4}$$

So at the output of a complex digital mixer running at f'_{ref}, followed by LPF, we have

$$\begin{aligned} y_n &= \mathrm{LPF}\left(r_n \times e^{-j2\pi f'_{ref}nT_s}\right) \\ &\simeq e^{j2\pi f_{ref}\zeta_n} \end{aligned}. \tag{5}$$

This represents a complex exponential where jitter is seen essentially as phase noise. Strictly speaking, the last expression in (5) holds only approximately, because due to jitter, the spectra of all the frequency components in the sampled signal in (4) are spread around their original frequency contents. However, with practical sampling clocks, it is expected that there is some correlation between consecutive jitter values and thus most of the energy of the sampled signal components is still located at the original frequencies. This is, of course, also influenced by the choice of the reference frequency relative to the actual received signal. Finally, based on (5), estimates of the sampling jitter values ζ_n can be obtained as

$$\hat{\zeta}_n = \arg\{y_n\}/(2\pi f_{ref}). \tag{6}$$

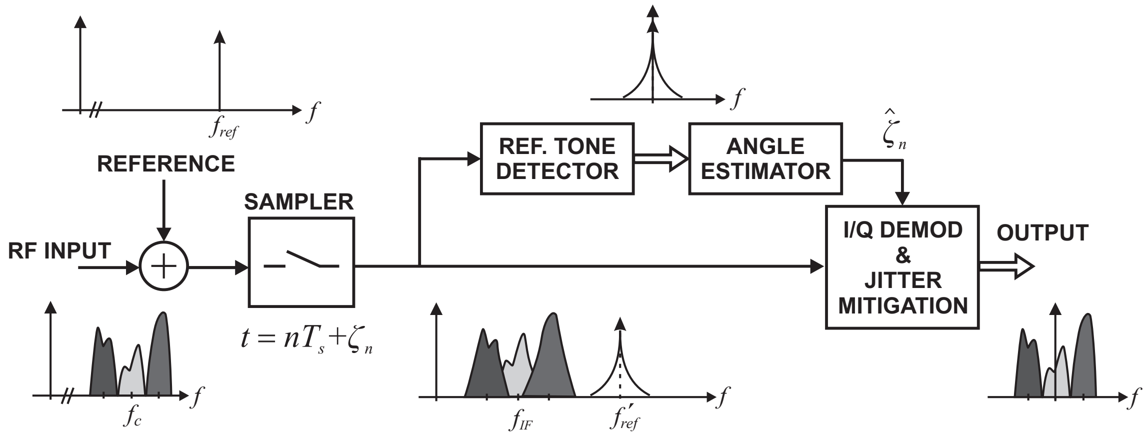

Fig. 1. General illustration of super-imposing a reference tone at sampler input, which is then used for digital jitter estimation and cancellation from the actual received signal in a direct-sampling radio. For illustration purposes, only three received channels are shown, and the target signal is depicted in the lightest grey.

C. Discussion

Like discussed already shortly above, in the estimation process, there are some essential limitations that should be carefully understood. The reference frequency f_{ref} must be selected so that it is possible to sufficiently separate (filter) the reference tone from the actual incoming signal in the jittered sampling. In practice, the amount of this separation depends on (i) the frequency separation between $r(t)$ and f_{ref} (after aliasing), (ii) the amount of jitter and (iii) the power ratio between $r(t)$ and the reference tone. In the selection of f_{ref}, it should also be kept in mind that jitter has heavier effect on high-frequency signals. Thus, the higher the selected reference frequency is, the easier it is, from practical signal processing point of view, to detect the phase behaviour of the isolated reference tone in (5). It should also be acknowledged that the reference signal itself must be very accurately known, and it is likely that the selection of f_{ref} affects also the generation accuracy of the reference. Furthermore, not only the frequency of the reference tone, but also its amplitude is important in the design. On one hand, if the dynamic range in the incoming signal is high, then also high number of bits is needed in the digital domain signal processing. On the other hand, it is easier to estimate the sampling jitter if the reference tone is strong (compared to $r(t)$), so from the estimation point of view, more powerful the reference tone the better. Thus a proper compromise is needed in dimensioning the reference signal power in practice. Concrete examples will be given in Section IV.

III. SAMPLING JITTER CANCELLATION

This Section proposes an efficient way to use the obtained jitter estimates $\hat{\zeta}_n$ in (6) for the jitter mitigation of the incoming signal.

A. Basic Cancellation Processing

The cancellation builds upon the received signal modelling in (4). Applying first digital I/Q down-conversion, from IF to zero frequency, and lowpass filtering yields

$$\begin{aligned} z_n &= \mathrm{LPF}\left(r_n \times e^{-j2\pi f_{IF} n T_s}\right) \\ &\simeq [z_I(nT_s + \zeta_n) + jz_Q(nT_s + \zeta_n)]e^{j2\pi f_c \zeta_n} \\ &= z(nT_s + \zeta_n)e^{j2\pi f_c \zeta_n} \end{aligned} \quad . \tag{7}$$

Above, it is assumed that the (jittered) reference tone is essentially suppressed by the LPF. Now, if the original RF centre-frequency f_c is much higher than the corresponding signal bandwidth, the jitter contribution on the composite modulating I and Q components (z_I and z_Q) is much lower than on the carrier components [12]. Thus we can approximate $z_I(nT_s + \zeta_n) \approx z_I(nT_s)$ and $z_Q(nT_s + \zeta_n) \approx z_Q(nT_s)$, and write (7) as

$$\begin{aligned} z_n &\approx [z_I(nT_s) + jz_Q(nT_s)]e^{j2\pi f_c \zeta_n} \\ &= z(nT_s)e^{j2\pi f_c \zeta_n} \end{aligned} \quad . \tag{8}$$

This basically means that the dominant jitter effect on the sub-sampled RF signal can be approximated as phase noise in the composite carriers of the original bandpass signal. Similar conclusion is also drawn in [11].

Based on the above modelling, jitter mitigation (using the jitter estimates $\hat{\zeta}_n$) can be carried out by multiplying the observed complex sample stream z_n with $\exp(-j2\pi f_c \hat{\zeta}_n)$. Assuming perfect jitter estimates ($\hat{\zeta}_n = \zeta_n$), this fully removes the jitter stemming from the original bandpass nature of the sampled signal, i.e.,

$$\begin{aligned} \hat{z}_n &= z_n \times e^{-j2\pi f_c \hat{\zeta}_n} \approx z(nT_s + \zeta_n) \\ &\approx z(nT_s) \end{aligned} \quad . \tag{9}$$

B. Further Interpretations for Multichannel Signals

In the previous Subsection, no explicit assumptions other than $z(nT_s + \zeta_n) \approx z(nT_s)$ were made on the nature of the incoming signal. Here we build further understanding and interpretations in the case where the overall baseband equivalent signal $z(t)$ of the incoming RF signal contains also neighbouring channels. This is important because jitter is generally most challenging in such scenarios. In an example case of one neighbouring channel on both sides, the overall baseband equivalent is of the form

$$z(t) = z_c(t) + z_{i,1}(t)e^{j2\pi f_{i,1}t} + z_{i,2}(t)e^{j2\pi f_{i,2}t}. \tag{10}$$

Here $z_c(t)$ denotes the complex envelope of the target signal (centred originally at f_c), and $z_{i,1}(t)$ and $z_{i,2}(t)$ denote the complex envelopes of the neighbouring channels located originally at offsets $f_{i,1} > 0$ and $f_{i,2} < 0$ from f_c , respectively. Under these assumptions, the compensated signal in (9) can be written as

$$\begin{aligned}\hat{z}_n &= z_n \times e^{-j2\pi f_c \hat{\zeta}_n} \\ &\approx z(nT_s + \zeta_n) \\ &= z_c(nT_s + \zeta_n) + z_{i,1}(nT_s + \zeta_n)e^{j2\pi f_{i,1}nT_s}e^{j2\pi f_{i,1}\zeta_n} \\ &\quad + z_{i,2}(nT_s + \zeta_n)e^{j2\pi f_{i,2}nT_s}e^{j2\pi f_{i,2}\zeta_n}\end{aligned} . \tag{11}$$

Assuming now that the individual complex envelopes $z_c(t)$, $z_{i,1}(t)$ and $z_{i,2}(t)$ change slowly relative to jitter values (which is always a valid assumption given jitter values in the tens of pico-second range and bandwidths in the order of few or few tens of MHz), we have $z_c(nT_s + \zeta_n) \approx z_c(nT_s)$, $z_{i,1}(nT_s + \zeta_n) \approx z_{i,1}(nT_s)$ and $z_{i,2}(nT_s + \zeta_n) \approx z_{i,2}(nT_s)$. Thus the signal in (11) can be simply written as

$$\begin{aligned}\hat{z}_n &= z_n \times e^{-j2\pi f_c \hat{\zeta}_n} \\ &\approx z_c(nT_s) + z_{i,1}(nT_s)e^{j2\pi f_{i,1}nT_s}e^{j2\pi f_{i,1}\zeta_n} \\ &\quad + z_{i,2}(nT_s)e^{j2\pi f_{i,2}nT_s}e^{j2\pi f_{i,2}\zeta_n}\end{aligned} . \tag{12}$$

Based on (12), the jitter stemming from the target signal is fully removed and also the jitter effects due to neighbouring channels are heavily reduced. The remaining jitter noise in the compensated signal in (12) stemming from the neighbouring channels is now relative to $\exp(j2\pi f_i \zeta_n)$ which is vanishingly small with practical jitter values ζ_n and neighbouring channel frequency offsets f_i .

IV. Simulation Environment

This Section demonstrates the applicability and performance of the proposed jitter estimation and cancellation technique. For reference, also the technique described in [8] by Rutten et al. is implemented. As an example radio system, 3GPP LTE -like system [1] is used. Thus OFDM waveform with 1024 subcarriers, of which 600 are active, is deployed with 16QAM subcarrier modulation and 15 kHz subcarrier spacing. Used RF centre-frequency is 2.6 GHz. After transmitter, the transmit waveform travels through either plain additive white Gaussian noise (AWGN) or extended ITU-R Vehicular A multipath [10] channel. Cases without and with neighbouring channels are simulated.

As a model of the sampling clock, and thereon of the jitter spectrum, the phase locked loop (PLL) based oscillator described in [11] and [14] is deployed. Similar oscillators are also described, e.g., in [2]. In the neighbouring channel studies, three sinusoidal interferers at 7.495 MHz, 7.5 MHz and 7.505 MHz offset (above) from the desired signal carrier frequency together with a bandpass-noise type interferer with approximately 4 MHz bandwidth at 9.5 MHz offset (below) from the desired signal carrier frequency are considered, as illustrated in Fig. 2. Thus the total incoming signal bandwidth is around 19-20 MHz. Stemming from the target centre-frequency of 2.6 GHz and the total bandwidth, we use a sub-sampling frequency of F_s = 16×1024×15 kHz = 245.76 MHz. This aliases the RF signal from 2.6 GHz centre-frequency to 11×245.76 MHz – 2600 MHz = 103.36 MHz IF frequency in a controlled manner. At the sampler input, the received signal is injected with a reference tone of 36 MHz. An example of aliased spectrum after sub-sampling is shown in Fig. 2 for interferer-free case and the case with adjacent channel interferers. After sub-sampling, the proposed jitter estimation and cancellation methods are applied as described in Sections II and III. Then, the digital selectivity filtering is applied and the OFDM waveform is detected and the symbol-error rates (SER) are evaluated as performance indicators.

Two kinds of reference power allocation schemes are assumed in the simulations. In both schemes the adjacent channel signals are considered to be 0 to 40 dB above the interesting OFDM signal. For the first scheme, the reference tone is constantly 40 dB above the interesting OFDM signal. In the second scheme, the reference tone power is allocated so that it has the same power level as the interferers. The latter case might be more practical one, as in such scheme the reference tone does not increase the dynamic requirements of the A/D-circuitry. In practice, the reference tone power can be allocated in connection with the automatic gain control mechanism of the A/D-circuitry, where the power of the signals at the sampler input is calculated anyway.

V. Simulation Results and Analysis

Simulation results with AWGN channel can be seen in Fig. 3. In the results, power levels of the interferers and the reference tone are 40 dB above the interesting signal. The results show that both the proposed technique as well as the reference technique from [8] improve the performance (reduce jitter noise) greatly compared to case without compensation. For interferer-free case, the proposed technique gives practically ideal performance over the whole evaluated SNR region with fixed 20 ps RMS jitter. However, small decrease in performance is visible in the more challenging case with the neighbouring channels included (+int. curves in the figures).

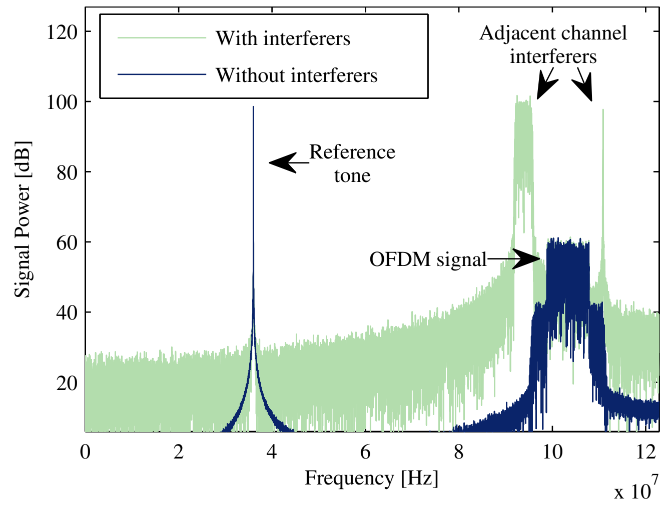

Fig. 2. Example spectra of the reference tone, the OFDM signal and the interferers after aliasing by sub-sampling at 245.76 MHz with 20 ps RMS jitter. The original RF signal centre-frequency is 2.6 GHz while the reference tone frequency is 36 MHz.

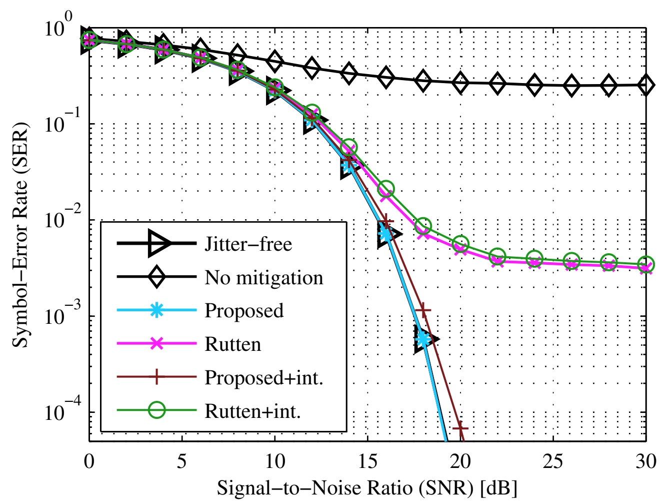

Fig. 3. SER given as a function of received SNR with fixed RMS jitter of 20 ps in AWGN channel.

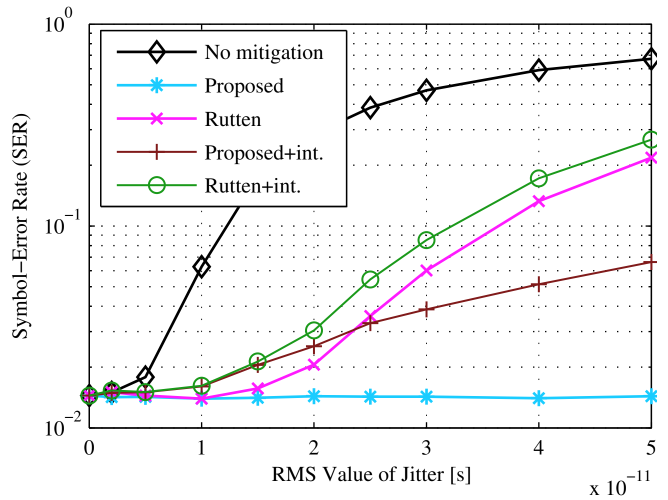

Fig. 4. SER given as a function of sampling jitter RMS with fixed received SNR of 26 dB in extended ITU-R Vehicular A multipath channel.

The results for extended ITU-R Vehicular A multipath channel case are depicted in Fig. 4 for an example fixed received SNR of 26 dB over wide jitter RMS region. Here as well as above, the power of the interferers and the reference tone are 40 dB above the level of useful signal. The reference technique from [8] performs nicely for lower than 15 ps jitter RMS. Its performance starts to decrease very fast after the 15 ps point. This is natural, because a low-jitter assumption was made in derivation of the reference technique in [8] (in AWGN case, already 5 to 10 ps RMS jitter started to affect the performance of the reference technique heavily.) With the neighbouring channels included, the proposed technique performs again very well giving near-ideal performance with lower than 10 ps RMS jitter. With jitter RMS over 10 ps, some decrease in performance is seen, as the jitter noise due to interferers gets higher and higher contaminating also the reference tone band. With the reference technique from [8], the performance is again limited to lower jitter values. Anyway, the proposed technique can again efficiently push down the signal distortion up to 20 ps RMS jitter range.

In Fig. 5, Fig. 6 and Fig. 7, the results are given as a function of the power level of the adjacent channel interferers. This is very interesting because in DRFS receiver, the selectivity is one of the main design concerns. In Fig. 5 the interference and the reference tone power levels are the same. There we see that when the reference tone level is lowered to the level of the OFDM signal, there is significant performance drop. Still, the proposed technique manages to give clear performance increase compared to the case without mitigation. From interference free cases, we see that when the reference tone is less than 10 dB over the useful signal, the spread of the OFDM signal lowers reference tone detection quality. When comparing the results in Fig. 5 to the results in Fig. 6, we see that the spread of the interferers has undesired effect also on the detection of the reference tone. Fig. 6 also shows that when using the reference tone fixed at 40 dB above the useful signal, the performances of the mitigation techniques are very good until the power of the adjacent channel interferers gets high enough to deteriorate the quality of the reference tone due to spreading. Similar conclusion can be made for the case of extended Vehicular A multipath channel, as depicted in Fig. 7. Multipath channel does not change the relative results too much, because most of the errors in the simulations are caused by the sampling process in the receiver.

VI. Conclusions

In direct RF sub-sampling radios, sampling jitter is one of the main practical issues from the practical implementation point of view. This paper proposes digital signal processing methods to estimate the sampling jitter realization sample-by-sample basis. The estimation is done by injecting a reference

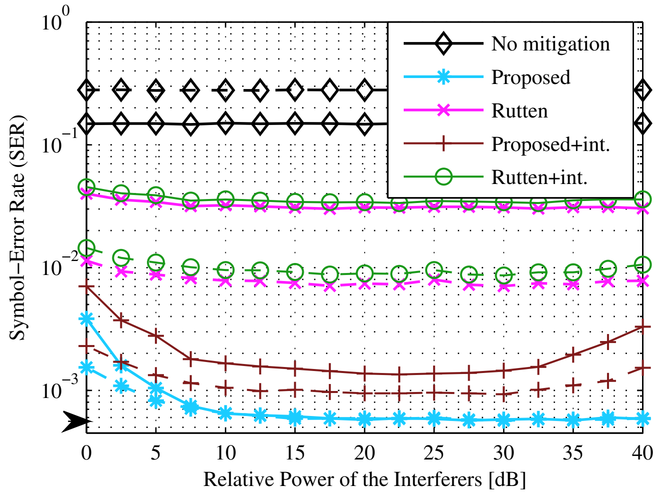

Fig. 5. SER given as a function of relative power of the interferers in AWGN channel with received SNR of 18 dB. The reference tone power is the same as the power of the interferers. RMS jitter is 20 ps (dashed) / 50 ps (solid). The arrow on SER axis shows the jitter-free performance.

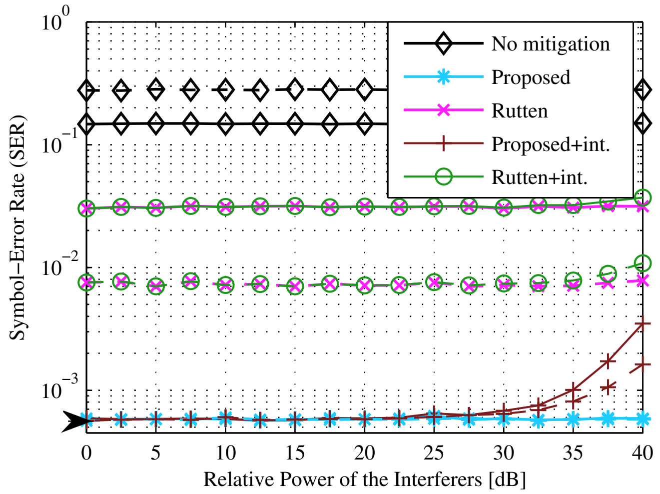

Fig. 6. SER given as a function of relative power of the interferers in AWGN channel with received SNR of 18 dB. The reference tone power is 40 dB above the useful signal power. RMS jitter is 20 ps (dashed) / 50 ps (solid). The arrow on SER axis shows the jitter-free performance.

tone to the sampler input and applying signal processing to that signal after the sampling. Furthermore, an efficient digital signal processing structure, that utilizes these jitter estimates to reduce the jitter effects from the actual received signal, was also proposed with special emphasis on DRFS receiver. Based on the performance simulations, the proposed estimation-cancellation approach can efficiently mitigate sampling jitter also in the challenging case of strong neighbouring channels entering the sub-sampling radio. Simulations showed that the performance of the technique is very good compared to the performance of state-of-the-art technique. One of the most interesting parameters in the proposed technique is the reference tone power, as it affects the dynamic range requirements in the sampling and also dictates the mitigation performance of the proposed technique.

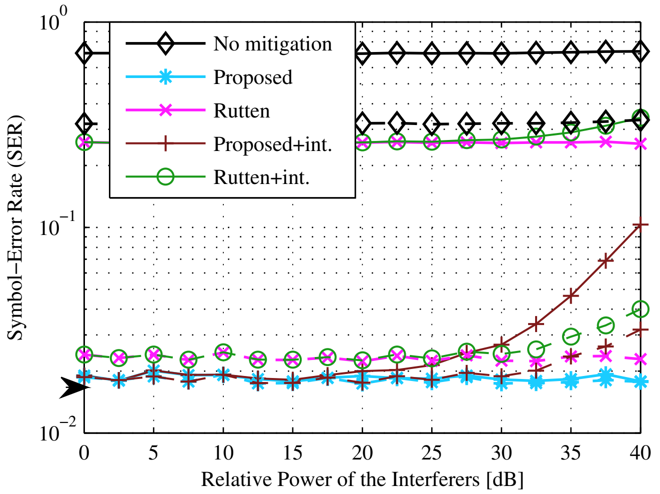

Fig. 7. SER given as a function of relative power of the interferers in extended ITU-R Vehicular A multipath channel with SNR of 26 dB. The reference tone power is 40 dB above the useful signal power. RMS jitter is 20 ps (dashed) / 50 ps (solid). The arrow on SER axis shows the jitter-free performance.

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