SAMPLING JITTER IN FULL-DUPLEX RADIO TRANSCEIVERS: ESTIMATION AND MITIGATION

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ABSTRACT

In this paper, the problem of sampling-jitter estimation in OFDM full-duplex radio transceivers is addressed. The sampling-jitter estimation is very challenging, because the effects are very small, but still high enough to be a problem in full-duplex radios. The proposed sampling-jitter estimation method is based on estimating only the most significant frequency components of the sampling jitter. The performance of the estimation algorithm is evaluated with simulations by studying how much self-interference cancellation is achievable when the self-interference is mitigated using the estimates of the sampling jitter samples.

Index Terms— Full-duplex radio technology, sampling jitter, estimation, mitigation, OFDM

1. INTRODUCTION

One of the recent breakthroughs in the area of wireless communications is so-called full-duplex radio (FDR) technology. The principle is old and simple: to transmit and receive signals simultaneously at the same frequency band with a single transceiver. However, the coupling of the own signal from the transmitter part of the transceiver to the receiver has made the practical implementation very challenging [1]. The self-interference (SI) that stems from the coupling comes from the own transmitter antenna few tens of centimeters away from the receiver antenna whilst the useful signal might come from tens of meters or even kilometers away. Therefore, the power of the SI signal can be tens of decibels higher than the power of the useful signal [1].

On the positive side is that the receiver part of the full-duplex transceiver always knows the transmitted samples from its own transmitter part. It also has access to the to-be-transmitted analog signal waveform. Therefore, cancelling the SI signal at the receiver is possible, even perfectly in theory. However, various distortions and changes that the transmitter components, communications channel and receiver components cause to the SI signal are unknown, and therefore make the perfect SI cancellation only a dream. For example, in the previous literature it has been shown that oscillator phase noise [2] and sampling jitter [3] significantly limit the maximum attainable SI cancellation.

This paper focuses on the effect of the sampling jitter in FDR transceivers. There is only one previous work that has considered sampling jitter in FDR in [3]. They observed that the sampling jitter effect on the SI cancellation is significant, if the FDR transceiver is used in a network utilizing a modern orthogonal frequency division multiplexing (OFDM) based communications standard. They also proposed a sampling-jitter mitigation method. However, they only considered sampling jitter mitigation with perfectly known sampling jitter. In this paper, also the sampling jitter estimation is addressed. It has been previously addressed for direct-sampling receivers in [4] and [5].This paper proposes a sampling-jitter estimation method to be used in FDR transceivers. This is the first ever sampling-jitter estimation technique considered for FDR transceivers. This paper also shows with simulations how well the proposed sampling jitter estimation technique works in different conditions and with different parameters, when it is combined with the mitigation technique proposed in [3].

The structure of this paper is as follows. In Section 2, the basic principle of the used FDR transceiver is explained and the corresponding signal models are derived. Furthermore, the used sampling-jitter mitigation technique is described. The proposed sampling-jitter estimation technique is then discussed in Section 3. In Section 4, the performance of the proposed sampling-jitter estimation algorithm is thoroughly studied with simulations. In Section 5, the works is concluded.

2. SAMPLING JITTER IN FULL-DUPLEX RADIOS

In this section, the used OFDM FDR-transceiver model is described in detail while the signal model, and the combined sampling-jitter mitigation and digital SI cancellation method are explained. The used FDR transceiver model is depicted in Fig. 1.

2.1. The full-duplex radio transceiver model

The used transceiver structure is depicted in Fig. 1 and is based on the one proposed in [1]. The SI is cancelled in two stages, in the analog radio-frequency (RF) domain and in the digital domain. The analog domain SI cancellation is here called analog linear cancellation (ALC) and the digital domain cancellation is called digital linear cancellation (DLC). In the ALC, the to-be-transmitted RF signal is fed through a tunable attenuation and delay component that mimics the main multipath component of the SI channel. The output is then subtracted from the received signal, effectively cancelling most of the SI signal coupling through the main component of the multipath channel. In the DLC, the known samples from the transmitter are fed through a tapped delay line that to mimics the whole effective SI multipath channel with ALC. In this paper, the DLC is improved to take also the sampling jitter into account. The modified DLC is explained in subsection 2.3.

2.2. Signal model with sampling jitter prior the DLC

In the used FDR transceiver model, inverse discrete Fourier transform (IDFT) is taken from the OFDM subcarrier symbols X_k in

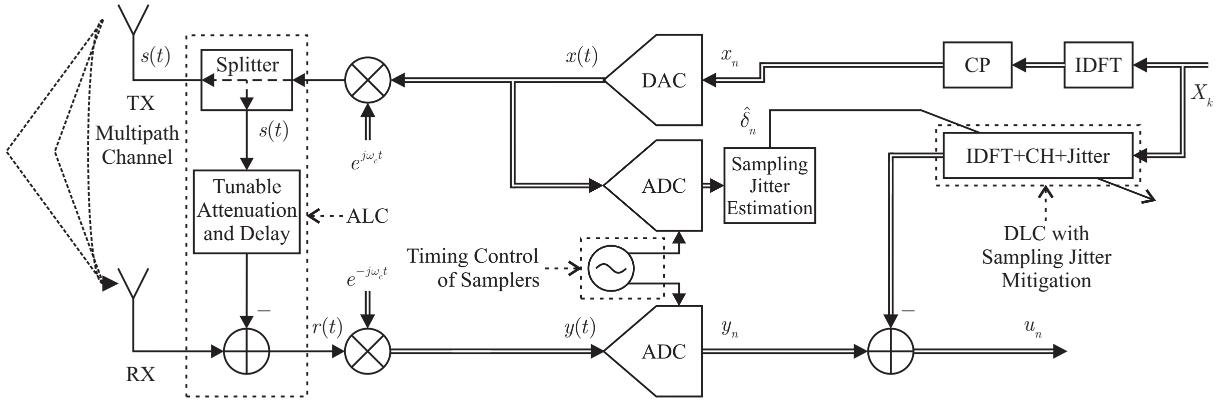

Fig. 1. OFDM full-duplex radio transceiver model with sampling jitter estimation and mitigation. ALC is analog linear cancellation, DLC is digital linear cancellation, CP is cyclic prefix addition, IDFT is inverse discrete Fourier transform, and DAC and ADC are digital-to-analog and analog-to-digital converters, respectively.

blocks of N symbols, and the cyclic prefix is added for each block. The resulting samples are denoted by x_n. These samples are then converted to an analog signal $x(t)$, which is then upconverted to the RF signal $s(t) = x(t)\exp\left(j\omega_c t\right)$, where ω_c is the RF carrier frequency. This signal is transmitted, and after experiencing SI channel with impulse response $h(t)$, it is received by the received part. Now, the ALC is done, and the signal at this point can be written as

$$\begin{aligned} r(t) &= h * s(t) - s_{ALC}(t) + n(t) \\ &= h_{ALC} * s(t) + n(t). \end{aligned} \tag{1}$$

Here, $s_{ALC}(t)$ is the signal at the output of the tunable attenuation and delay unit of the ALC, $h_{ALC}(t)$ is the impulse response of the SI channel when the ALC is taken into account in the main multipath component, $n(t)$ is the signal that consists of the received useful signal, additive white Gaussian noise and all the distortions, and $a * b(t)$ denotes the convolution between $a(t)$ and $b(t)$.

After downconversion at the receiver, the signal is written as

$$\begin{aligned} y(t) &= e^{-j\omega_c t}\left\{h_{ALC} * s(t)\right\} + n'(t) \\ &= h^c_{ALC} * x(t) + n'(t). \end{aligned} \tag{2}$$

Here, $n'(t)$ is the baseband equivalent of $n(t)$, and $h^c_{ALC}(t)$ is the channel impulse response when the phase differences between the upconverting and downconverting oscillators for different multipath components are taken into account. After the downconversion, the signal is sampled with sampling frequency F_s. The sampled signal can be written as

$$\begin{aligned} y_n &= y(nT_s + \delta_n) \\ &= \sum_{m=0}^{P-1} h^c_{m,ALC}\, x\left[(n-m)T_s + \delta_n\right] + n'(nT_s + \delta_n), \end{aligned} \tag{3}$$

where, $T_s = 1/F_s$ is the sampling interval. Notice that here the effective channel is modelled with discrete-time impulse response $h^c_{n,ALC}$, and $P-1$ is the maximum delay spread of the channel in samples.

2.3. Signal model after DLC with sampling jitter mitigation

The sampling jitter mitigation is done with the principle proposed in [3] and depicted in Fig. 1. The idea is to generate the DLC signal using the transmitted subcarrier modulated symbols. It is done so that the subcarrier modulated symbols are OFDM modulated in a way that the sampling jitter and the effective SI channel are also taken into account in the process. This signal is then subtracted from the received samples. This was shown in [3] to be very efficient way to cancel sampling jitter with perfectly known sampling jitter samples. In this paper, we do not assume perfectly known sampling jitter though.

In (3), the SI signal is a convolution between the SI channel after ALC and signal x_n, but the convolved signal is in addition corrupted by the sampling jitter. Therefore, the DLC with sampling-jitter mitigation explained above is done as

$$\begin{aligned} u_n &= y_n - \sum_{m=0}^{P-1} \hat{h}^c_{m,ALC}\, x\left[(n-m)T_s + \hat{\delta}_n\right] \\ &\approx n'(nT_s + \delta_n), \end{aligned} \tag{4}$$

where $\hat{h}^c_{n,ALC}$ is the estimate of the effective SI channel with ALC, and $\hat{\delta}_n$ is the estimate of the sampling jitter for the n th sample. Two of the main problems at this point are the estimation of the sampling jitter $\hat{\delta}_n$ and then based on that computing value for $x\left[(n-m)T_s + \hat{\delta}_n\right]$. The former problem is tackled in the next section. In general, computing value for $x\left[(n-m)T_s + \hat{\delta}_n\right]$ can be made for example by interpolation of the known samples x_n, but it is either computationally relatively complex or not accurate enough, because $\hat{\delta}_n$ is always very small value and we are now interested in relatively accurate value of $x\left[(n-m)T_s + \hat{\delta}_n\right]$ even for a slightest change in $\hat{\delta}_n$. Fortunately, we can model the sampling jitter accurately in the following way during the OFDM sample generation.

With subcarrier symbols X_k to be transmitted, an OFDM signal is generated by taking IDFT from the subcarrier symbols in blocks of N symbols. Each block therefore consists of N samples and a block is called OFDM symbol. Then, a cyclic prefix is added and the signal is converted to a time-domain signal with digital-to-analog converter (DAC). The signal waveform during b th OFDM symbol at this point without the cyclic prefix can be written as

$$x_b(t) = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} X_{k,b} e^{j2\pi kt/T}, \tag{5}$$

where T is the OFDM symbol length in seconds (without the cyclic prefix.) When the sampling, the sampling jitter and the sample offset m (as in $x\left[(n-m)T_s + \hat{\delta}_n\right]$) are then modeled into this signal, it can be written as

$$x\left[(n-m)T_s+\hat{\delta}_n\right]=\frac{1}{\sqrt{N}}\sum_{k=-N/2}^{N/2-1}X_{k,m}e^{j2\pi k\left[(n-m)T_s+\hat{\delta}_{n,b}\right]/T} \\ =\frac{1}{\sqrt{N}}\sum_{k=-N/2}^{N/2-1}X_{k,m}e^{j2\pi k(n-m)/N}e^{j2\pi kF_s\hat{\delta}_{n,b}/N}. \quad (6)$$

So we are able to model the sampling jitter and the time delay into the OFDM signal natively during OFDM symbol generation. Then, the signal used in the DLC can be constructed according to (4) using (6). Note that because of handling OFDM signals now, the indexing is done so that $n\in\left[0,N-1\right]$ for every OFDM symbol separately, thus changing $\hat{\delta}_n$ into $\hat{\delta}_{n,b}$ and so on. In the notation, added index b always refers to b th OFDM symbol.

3. SAMPLING-JITTER ESTIMATION

The basic idea of the proposed sampling-jitter estimation method is depicted in Fig. 1. The time-domain baseband OFDM waveform from the transmitter (from now on called a reference signal) is fed to an extra analog-to-digital converter (ADC). The sampling frequency is the same as in the main ADC in the receiver part, and the timing of the sampler part is controlled with exactly the same clock, thus making the sampling jitter realizations the same. The sampling-jitter estimation is done with digital signal processing from the samples of the reference signal. Naturally, the sampling jitter could also be estimated from the received signal without the extra ADC, but unfortunately the effect of the sampling jitter is so small, that it is impossible to be accurately recovered because of the noise and the useful signal are present at the receiver part.

To get started, the baseband signal at the transmitter during the b th OFDM symbol $x_b(t)$ is fed to the extra ADC, controlled with the same clock as the ADC of the receiver part with sampling jitter $\delta_{n,m}$, to generate the reference signal

$$x_{n,b,ref}=x_b(nT_s+\delta_{n,b})=\frac{1}{\sqrt{N}}\sum_{k=-N/2}^{N/2-1}X_{k,b}e^{\frac{j2\pi kF_s\delta_{n,b}}{N}}e^{\frac{j2\pi kn}{N}}. \quad (7)$$

From this form, it is very burdensome to estimate the sampling jitter or even derive closed-form discrete Fourier transform (DFT) (signal after OFDM demodulation). However, by studying (7), we learn that the sampling-jitter effect has multiplier $2\pi kF_s/N$, and it therefore has higher and higher effect as the frequency of the subcarrier index (kF_s) (also the frequency of the subcarrier) increases. Therefore, the lower frequency subcarriers do not suffer from sampling jitter almost at all, whilst the effect is very high on the higher frequency components. Motivated by this, we use only the higher frequency subcarriers in the sampling-jitter estimation. We also notice from the multiplier $2\pi kF_s/N$ that even though the sampling jitter has different level of effect on every subcarrier, the effect is almost the same on adjacent subcarriers (e.g., the multiplier difference between 285th and 300th subcarriers is only around 5 %, so if we study the 285th subcarrier, the effect varies maximum 5 % for 15 subcarriers on the both sides). Therefore, we can make an approximation that the length-N DFT of $x_{n,b,ref}$ for a set of subcarrier near a high-frequency subcarrier, say a th subcarrier, is (when index z belongs to this set of high-frequency subcarriers)

$$X_{z,b,ref}\approx \mathrm{F}_z\left\{e^{\frac{j2\pi aF_s\delta_{n,b}}{N}}\right\}*X_{z,b}=J_{z,a,b}*X_{z,b} \\ =\sum_{l=-N/2}^{N/2-1}J_{l,a,b}X_{z-l,b}. \quad (8)$$

Here, $J_{k,a,b}$ is the DFT ($\mathrm{F}_k\{\cdot\}$) of $\exp\left(j2\pi a\,\mathrm{F}_s\delta_{n,b}/N\right)$. This is a simple form, and only applies approximately for subcarriers z that are near subcarrier a (called from now on 'the set of high-frequency subcarriers' (SHFS)). In the sampling-jitter estimation technique only the SHFS are used, so the DFT carried out in (8) can be done for only the SHFS subcarriers, thus making the technique computationally relatively simple. Notice now that for these subcarriers the sampling-jitter effect has been approximated to be similar as the effect of the phase-noise [6]. Therefore, using the idea in [6] is possible now for sampling-jitter estimation for SHFS.

Now, if the sampling clock is generated by a realistic oscillator, the values at very low frequency components (low values of k) of $J_{k,a,b}$ dominate the values at other frequency components. Therefore, it is from estimation point-of-view interesting to rewrite (8) in form

$$X_{z,b,ref}\approx\sum_{l=-u}^{u}J_{l,a,b}X_{z-l,b}+c_{z,b}. \quad (9)$$

Here, $c_{k,b}$ is the sampling jitter contribution of the higher frequency components of $J_{k,a,b}$. Now, we make a selection to only estimate the low-frequency components (from $-u$ to u) of the sampling jitter, because they are anyway dominating. Therefore, $c_{k,b}$ are noise from the estimation point of view. We now assume that subcarriers from α to γ belong to SHFS group and rewrite (9) for only SHFS group as

$$\begin{bmatrix} X_{\alpha+u,b,ref} \\ X_{\alpha+u+1,b,ref} \\ \vdots \\ X_{\gamma-u,b,ref}\end{bmatrix}\approx\begin{bmatrix} X_{\alpha+2u,b} & X_{\alpha+2u-1,b} & \cdots & X_{\alpha,b} \\ X_{\alpha+2u+1,b} & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ X_{\gamma,b} & \cdots & \cdots & X_{\gamma-2u,b}\end{bmatrix}\begin{bmatrix} J_{-u,a,b} \\ J_{-u+1,a,b} \\ \vdots \\ J_{u,a,b}\end{bmatrix}+\begin{bmatrix} c_{\alpha+u,b} \\ c_{\alpha+u+1,b} \\ \vdots \\ c_{\gamma-u,b}\end{bmatrix}. \quad (10)$$

Notice from (10) that we are actually using the SHFS subcarriers and u subcarriers from the both sides of them in this form and eventually in the estimation. The (10) can be written in a matrix form as $\mathbf{X}_{b,u,ref}\approx\mathbf{X}_{b,u}\mathbf{J}_{b,a,u}+\mathbf{C}_b$, and on this we can use the well-known least squares (LS) algorithm to get an estimate for $\mathbf{J}_{b,a,u}$ as

$$\hat{\mathbf{J}}_{b,a,u}=\left(\mathbf{X}_{b,u}^H\mathbf{X}_{b,u}\right)^{-1}\mathbf{X}_{b,u}^H\mathbf{X}_{b,u,ref}. \quad (11)$$

These are the estimates of the most significant frequency components (from $-u$ to u) of $\exp\left(j2\pi a\,\mathrm{F}_s\delta_{n,b}/N\right)$, therefore the sampling-jitter estimates are obtained after IDFT, taking argument of the result and dividing the result with $2\pi a\,\mathrm{F}_s/N$.

The validity of all the approximations used in this section is eventually proven to be reasonable in the next section by the performance of the sampling-jitter estimation algorithm, since no such approximations are used in the simulated signals.

4. SIMULATIONS

4.1. Simulator and simulation parameters

The simulator is built based on the transceiver structure presented in Section 2. The ALC and DLC are done with perfect channel knowledge, but the channel estimation error is modelled by adding white Gaussian noise to the known channel impulse response taps. Antenna separation is assumed to be 30 dB, ALC level is assumed to be 30 dB (the channel estimation error is set so that the main multipath component is suppressed enough that the SI power is decreased by 30 dB by the ALC). The DLC is done with perfect channel knowledge, so that the sampling jitter is the only thing limiting the DLC performance. This gives good understanding how sampling jitter affects the performance and how well the algorithm performs. The power-delay profile of the SI channel is 0 dB, −65 dB, −67 dB, −65 dB, −71 dB and −72 dB for delays of 0, 1, 2, 4, 6 and 9 samples, respectively. This is selected based on the measurement for full-duplex relay in [7], but the results are modified to fit better for FDR transceiver.

In the simulations OFDM signal is used. The parameters are selected so that the system resembles 3GPP Long Term Evolution (LTE) downlink with 10 MHz effective bandwidth (1024 subcarriers, 300 on both sides of zero subcarrier are 16QAM subcarrier modulated and other are zero subcarriers, with 15.36 MHz sampling frequency done at OFDM native sample rate [8].)

The sampling jitter is generated with phase-locked loop (PLL) oscillator model described in detail in [9] and [10], with parameters $L_f = -40\ \mathrm{dBc/Hz}$ at 1 kHz offset from the oscillation frequency and $L_w = -120\ \mathrm{dBc/Hz}$ at 1 MHz offset from the oscillation frequency. The generated jitter is then normalized to the desired root-mean-square (RMS) level of the sampling jitter.

In the sampling-jitter estimation algorithm, the used SHFS are non-zero subcarriers with the highest positive frequencies. The results can be improved by using also the negative frequencies, but to limit the complexity of the algorithm we opt to use only the positive frequency subcarriers.

4.2. Simulation results and analysis

The simulations results are given in Fig. 2, Fig. 3 and Fig. 4. In the results the total achievable SI cancellation (ALC+DLC) is given as a performance indicator. The 'Proposed Alg.' is the performance of the proposed algorithm and reference result 'CPE Est.' is case when only DC-bin (mean within each OFDM symbol period) of the sampling jitter is estimated with the proposed algorithm.

It is good to keep in mind the required level of SI cancellation for modern communications standards. For example, for 3GPP LTE, maximum UE power is $23\ \mathrm{dBm}$ [8]. After antenna separation of 20-30 dB, the power is −7-3 dBm. In the worst case, the required sensitivity level is at around −110 dBm [11]. Therefore, the required SI cancellation (ALC+DLC) is around 106 to 120 dB.

In Fig. 2, the varied parameter is the RMS value of the sampling jitter, the amount of SHFS is fixed at 80, and the parameter u is fixed at 20. As can be seen, the proposed algorithm boosts the SI cancellation by extra 20 dB. This corresponds to around 10-times effective decrease in the RMS value of the sampling jitter. Good enough SI cancellation is achievable without sampling-jitter mitigation in the lower studied sampling jitter regions. However, the higher studied regions are of more interest, since the lower region is achievable only with very high quality (and expensive, power consuming and/or bulky) samplers.

In Fig. 3 and Fig. 4, parameters of the sampling-jitter estimation algorithm are varied. In Fig. 3, the value of u is varied, and the amount of SHFS is fixed at 80. We can see that changing the value of u has only slight effect on the SI cancellation performance after value of around 7 (increasing it increases the complexity). However, when the value is too high, the amount of SHFS is not sufficient and the performance drops dramatically. In Fig. 4, the amount of SHFS is varied. Value of u is fixed at 5. From the results it is clear that after high enough value of SHFS, increasing it does not have any practical effect on the performance since the it is enough for reliable estimation. Also, used approximations hold best with small number of SHFS. The SHFS should be sufficient for the selected value of u. The use of low value of u and low amount of SHFS results in good performance with low computational complexity.

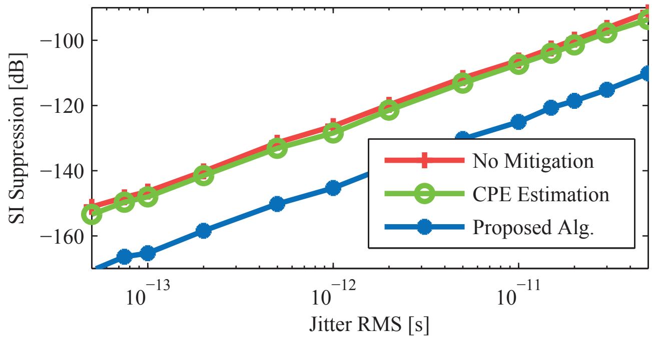

Fig. 2. SI suppression given as a function of the root-mean-square (RMS) value of the sampling jitter. $u = 20$ and $\mathrm{SHFS} = 80$.

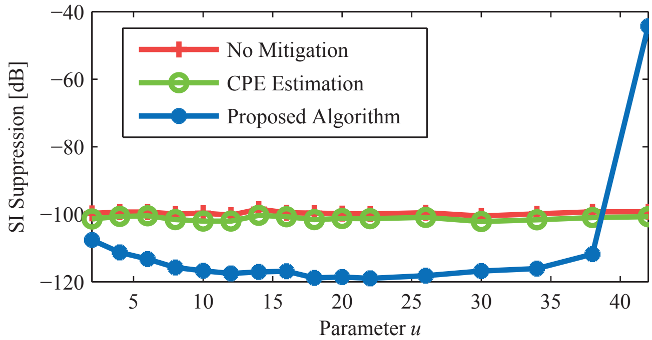

Fig. 3. SI suppression given as a function of the parameter u. Jitter RMS is 20 ps and $\mathrm{SHFS} = 80$.

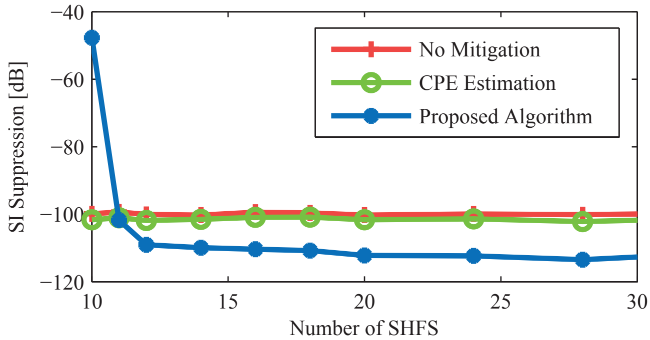

Fig. 4. SI suppression given as a function of number of SHFS. Jitter RMS is 20 ps and $u = 5$.

5. CONCLUSION

Sampling jitter is a problem in full-duplex radio transceivers. In this paper, a sampling-jitter estimation algorithm was proposed. When used in aid of digital cancellation of the self-interference, the sampling-jitter effects lowered to such a level, that the sampling jitter should not be a problem with practical sampling-jitter levels even in demanding modern communications standards.

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