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A Linearization Strategy for Undersampling Analog-to-Digital Converters

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Abstract- The ability of high performance Radar and Broadband Systems to detect weak targets in presence of strong interferers or clutter is given by their Spurious Free Dynamic Range (SFDR). Although the Signal-to-Noise-Ratio (SNR) necessary for detection may be improved by well-known system processing gains, the dynamic range is ultimately limited by distortion terms caused by nonlinear behaviour of receiver components. The Software Defined Radio (SDR) paradigm assigns the Analog-to-Digital Converter a key role in receiver design. For systems using IF- Subsampling, linearity requirements place a heavy burden on the ADC, as SFDR signifcantly degrades with increasing input frequency. As a consequence, the ADC can only be used at input frequencies fairly below its intrinsic full power bandwidth, restricting the systems IF placement. This contribution discusses the possibility of processing ADC output data in the digital domain to achieve improved linearity. The Volterra series approach of nonlinear systems and its constrained variants are discussed. We will show in detail that for higher input frequencies, dynamic errors cause the harmonic terms to loose their in-phase ability; in higher Nyquist zones a frequency-dependend dynamic phase error has to be considered. Assumptions are backed by an evaluation of coherent data from the LTC2208 (16 Bit, 120 MSPS). A specific correction algorithm incorporating the dynamic phase error will be presented, which yielded 25 dB SFDR improvement in the 7th Nyquist Zone (360-420 MHz). The reproducibility of correction results is considered in some detail.

I. Introduction

Highlighting the importance of Analog-to-digital converters is justified for a number of reasons: The choice of receiver architecture depends directly on ADC type and performance; ADCs have a direct impact on the system's overall performance in terms of usable dynamic range and detectabilty of targets; ADCs are less-understood than purely analog components; ADCs are key components for receivers exploiting the Software Defined Radio (SDR) concept or the integration ability [22]. IF-Subsampling for example was not possible until major advances in ADC technology and performance allowed for it [15]. Especially Radar drove converter requirements for a long time [18][16]. Today two classical ADC application forms can be observed: Sigma-delta converters for Baseband-Sampling and Pipelined converters for IF-Subsampling. This paper is devoted to the latter: IF-Subsampling uses one ADC operating at IF, by applying Nyquist's basic sampling theorem: ADC operation is divided into Nyquist Zones, which are frequency regions of $f_s/2$ width. After sampling, a bandlimited signal originally located in a specific Nyquist Zone will appear between 0 and $f_s/2$, ready for a Digital Downconversion (DDC). Unfortunately, ADC linearity performance in terms of SFDR degrades with increasing analog input frequency due to front-end induced dynamic errors [7]. How high a IF can be placed for proper digitization is limited by SFDR performance. In comparison, SNR degradation due to aperture jitter is not as drastic, due to today's ultra-low ADC internal jitter and high-end external clock sources [19]. The well-known IF-Subsampling technique allows for some architectural benefits as it helps avoiding known drawbacks like I/Q imbalance and other dirty RF effects associated with baseband-sampling like time-varying dc offset [30]. SFDR describes the ability to distinguish strong signals in the presence of weaker ones – a very important Radar requirement. While various processing gains from matched filter, beamforming or pulse-to-pulse-integration can be used to increase a Radar's SNR and thus detect small signals in the presence of noise, SFDR remains a given fact and does ultimately limit the dynamic range, unless nonlinear correction schemes are applied. In the case of Pulse-Doppler-Radar, distortion degrades the detection process and will produce unwanted false targets [29][25].
Identification of a proper model with respect to amplitude and frequency is a necessary step before correction-aimed models can be deployed. Over the last years such error correction methods, which improve SFDR by post-processing digital data showed their relevance: [40] made basic work on identification and correction of nonlinear distortion and [33] gave a state-of-the-art on ADC compensation methods. In [43][44] and many foreground papers of the same research group a ADC correction scheme based on a frequency-dependent,

dynamic INL model combined with static LUT correction was presented. In [41][42] this work was continued and more results were presented. Focusing strictly on slope-dependent dynamic errors in the ADC frontend, a compensation algorithm was published in [39]. A similar approach emerged earlier in [34]. The authors of [37] used wavelet networks to model dynamic nonlinear effects, but did not apply any correction. Recently, a purely static correction for the first Nyquist Zone applied to a F&I ADC was published and its robustness to temperature variations was verified [12]. More general investigations on digital improvement possibilities in Radar and RF systems were made in [26][27][28]. It should be pointed out that Digital Assist has the potential to improve SFDR of analog amplifiers and amplifing mixers preceding the ADC. It can help to build future systems, for example fully digital phased array receivers, with lower power consumption and costs.

II. Theory and Modelling

The simplest form of a nonlinear system is the memoryless power series, which is based on normal polynomials:

$$y = \sum_{i=0}^{L} a_i x^i \tag{1}$$

There is no frequency-dependency and when applying an input sinewave all produced harmonics will be in-phase with the fundamental. Setting amplitude to 1, the DC offset to 0 and limiting the series to 3rd order yields

$$\begin{aligned} x &= \cos(2\pi ft + \varphi_0) \\ y &= \sum_{i=1}^{3} a_i x^i = a_1 \cos(2\pi ft + \varphi_0) + (a_2 \cos(2\pi ft + \varphi_0))^2 + (a_3 \cos(2\pi ft + \varphi_0))^3 \\ y &\cong a_1 \cos(2\pi ft + \varphi_0) + \frac{1}{2} a_2 \cos(2\pi ft \cdot 2 + 2\varphi_0) + \frac{1}{4} a_3 \cos(2\pi ft \cdot 3 + 3\varphi_0) \end{aligned} \tag{2}$$

The link between harmonic powers and polynomial coefficients can be expressed in general [43]. Here, the newly generated frequency components belong to the group of harmonic distortion (HD), while applying a two-tone stimulus causes the production of more terms, which are refered to as intermodulation distortion (IMD). The IMD products for the above limited series appear at f_2-f_1, f_2+f_1 (2nd order IMD) and at $2f_2$-f_1 and $2f_1$-f_2 (3rd order IMD), respectively. Together with harmonics they share a specific amplitude behaviour, known as the m-th order law, e.g. with every 1 dB decrease in fundamental power, 3rd order harmonic and intermodulation distortion terms will decrease by 3 dB. While in the analog world this is used for linearity definitions in terms of interception points. ADCs are not power devices and thus specified in terms of one-tone and two-tone SFDR [1]. To overcome the defects of the traditional power series and to introduce memory to the system, the Volterra series approach to nonlinear systems is applied. The Volterra model captures a wide range of nonlinear systems with weak, fading memory by expanding the traditional linear systems with higher-order impulse responses, called Volterra kernels, and multi-dimensional convolution integrals, called Volterra operators [23]. Now the system's output value will depend on past input values.

$$\begin{aligned} y(x(t)) &= T_1[x(t)] + T_2[x(t)] + T_3[x(t)] + \ldots = \\ &= \int_{-\infty}^{\infty} h_1(\tau_1) \cdot x(t-\tau_1) d\tau_1 + \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_2(\tau_1,\tau_2) x(t-\tau_1) x(t-\tau_2) d\tau_1 d\tau_2 + \\ &+ \int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} h_2(\tau_1,\tau_2,\tau_3) x(t-\tau_1) x(t-\tau_2) x(t-\tau_3) d\tau_1 d\tau_2 d\tau_3 + \ldots \end{aligned} \tag{3}$$

Truncation of the infinite series leads to a discrete, digitally implementable form of the Volterra filter. A graphical depiction of the filter can be found in [40].

$$\begin{aligned} y(x(k)) = &\sum_{m_1=0}^{M_1} h_1(m_1) \cdot x(k-m_1) + \sum_{m_1=0}^{M_1}\sum_{m_2=0}^{M_2} h_2(m_1,m_2) x(k-m_1) x(k-m_2) + \\ &+ \sum_{m_1=0}^{M_1}\sum_{m_2=0}^{M_2}\sum_{m_3=0}^{M_3} h_3(m_1,m_2,m_3) x(k-m_1) x(k-m_2) x(k-m_3) \end{aligned} \tag{4}$$

Unfortunately, the number of coefficients tremendously rises with increasing memory length M and nonlinearity order L. This can be expressed as M + M2 + M3 +…, e.g. for quadratic and cubic nonlinearity only, a memory length of 6 already causes a Volterra filter with 6 + 36 + 216 = 258 coefficients. While being very complex (limiting its practical usage in terms of identification [5] and implementation), the Volterra series theoretically is able to describe a great amount of nonlinear, frequency-dependent effects. In our opinion, ADCs are perfectly suited for Volterra-based correction, because distortion terms are always far away from the fundamental at full scale excitation (at least 50 dBc), which satisfies the condition for a weak nonlinear system, and the dominant distortion terms are 2nd and 3rd order [6]. Quadratic and cubic contributors produce higher frequency components; as will be seen later, it is necessary to add an interpolating section prior to the nonlinearity function. Only then, the frequency dependency of distortion terms can be described without running into aliasing effects. Every general system can be simplified, if deeper knowledge about its structure is available. The assumption of symmetrical kernels allows a decomposition of system nonlinearity and dynamics [32] and the usage of linear filters. A special symmetry is then imposed on upper and lower IMD distortion terms.

$$y(x(k)) = \sum_{i=1}^{L}\sum_{m=0}^{M} h_i(m)(x(k-m))^i = \sum_{m=0}^{M_1} h_1(m)\cdot x(k-m) + \sum_{m=0}^{M_2} h_2(m)(x(k-m))^2 + ... + \sum_{m=0}^{M_L} h_L(m)(x(k-m))^L \quad (5)$$

A parallel structure of static nonlinearities followed by linear filters, is known as a polynomial Hammerstein model. Note that the expression "linear filter" refers to the filter structure and not to the phase characteristica in terms of linear phase FIR etc. It is also possible to reverse the position of the blocks, yielding a polynomial Wiener model. The three systems (Volterra, Hammerstein and Wiener) are summarized in Figure 1.

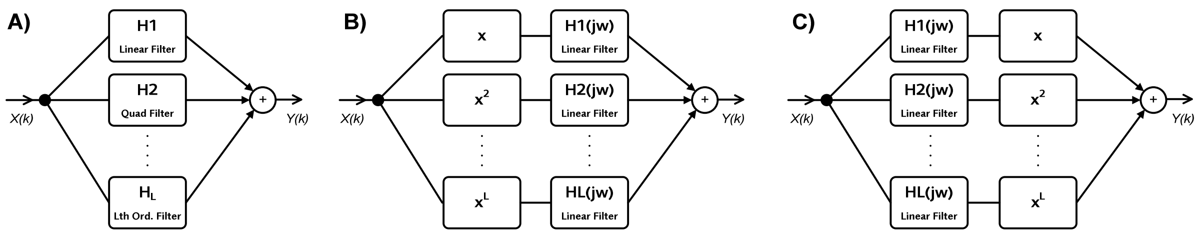

Figure 1. A) Volterra System, B) Polynomial Hammerstein System, C) Polynomial Wiener System

We now address an interesting problem, which will be referered to as dynamic phase error. Dynamic nonlinear distortion causes the distortion terms to fall out of phase with the fundamental [2][9][35][38]. Reasons can be group delay issues in the linear subsystem or other internal memory mechanisms. Recalling the power series from (2), an additional frequency-dependent phase error φ_{Di} has to be considered for each distortion order:

$$y \cong a_1(f)\cos(2\pi ft + \varphi_0 + \varphi_{D1}(f)) + \frac{1}{2}a_2(f)\cos(2\pi ft\cdot 2 + 2\varphi_0 + \varphi_{D2}(f)) + \frac{1}{4}a_3(f)\cos(2\pi ft\cdot 3 + 3\varphi_0 + \varphi_{D3}(f)) \quad (6)$$

If the filters inside the Hammerstein system have a constant group delay, all distortion terms will be in phase with the input signal. But, if we want to correct dynamic errors produced inside a real ADC, our correction filter must be able to capture the phase shift of harmonics. The memory effects causing phase rotation cannot be tackled with real coefficients of a taylor series (this would allow only phase shifts of 0 and +/- 180°); the coefficients have to be complex [3]. Only for very low input frequencies, the dynamic contributions of the ADC frontend can be neglected; unfortunately, when using undersampling, input frequencies are always high. In order to elaborate more of this, we will discuss shortly the Integral Nonlinearity (INL) of ADCs. INL describes the static transfer function of the ADC quantizer in terms of voltage transition levels, after correcting for gain and offset error [31]. Differential nonlinearity (DNL) is related to INL and often found in data sheet specifications; a proper DNL ensures that the ADC quantizer has no missing codes. The INL depends on the ADC architecture and can be calculated from ADC raw data using either the sinusoid histogram method [10], the sinewave fit method or more refined methods [11][12]. The test and its calculation formula is described in [4][7] and will not be reproduced here. Accordingly, there are two main contributors to ADC nonlinearity: static and dynamic [13] factors. This basic ADC structure is shown in Figure 2. A modeling approach refered to as dynamic integral nonlinearity modeling (DINL), used in [41][42][14], separates the INL curve into a static and a frequency-dependent polynomial part.

$$INL(f,k) = INL_{HCF}(k) + INL_{LCF}(f,k) \quad (7)$$

The static part is frequency-independent and called HCF (High Code Frequency Componenent), the dynamic part is frequency-dependent and called LCF (Low Code Frequency Component). For the correction, the LCF part is extracted by applying stepped sine waves across the Nyquist Zone, making a least-square fit to the different data sets and using the frequency-dependent polynomial coefficients for a finite impulse response filter design.

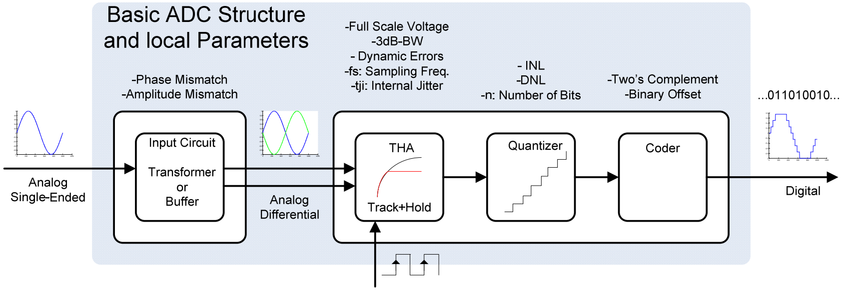

Figure 2. ADC Structure: Cascaded Nonlinearity Contribution inside an Track-and-Hold ADC

As the ADC exhibits memory at higher input frequencies due to a) imperfections in the input circuitry (mismatch/parasitics) and b) inherent nonlinearity of the track-and-hold-amplifier, some effects are essential:

- In addition to their frequency-dependent amplitude behaviour, harmonics encounter a phase rotation
- Classical methods that extract INL from raw data are very insensitive to out-of-phase distortion terms [8], a fact that leads to an underestimation of nonlinear behaviour and reduced improvement for subsequent correction algorithms. In [42], for a given undersampling ADC, a limitation of DINL modelling/correction to the first Nyquist zone was suggested.
- Dynamic and static nonlinearity contributions are overlaid, making them difficult to distinguish, it is therefore more reasonable to think of an ADC as a cascaded, inter-connected nonlinear system [17]. However, for most designed ADCs, static INL is very low and the dynamic contribution at the frontend is the main limiting factor of the SFDR.

III. Correction Scheme

There is a fundamental difference between static and dynamic ADC correction methods [51]: A static correction is sufficient in a system where memory effects can be neglected, but with a static correction being active and the input frequency being increased, the performance will at some point start to degrade significantly. Dynamic correction on the other hand is more complex to achieve but has the advantage of improving linearity for systems with large bandwidth and high input frequencies. In order to compensate nonlinearities, the inverse nonlinearity must be established [23]. Figure 3 shows the preprocessing correction scheme as implemented in MATLAB. The sampling frequency in the nonlinear path is properly increased in order to avoid aliasing. We expanded the Volterra/Hammerstein model up to 3rd order nonlinearity. In a first simulatory step, we used both a manufacturer model [21] and a model designed in our internal project.

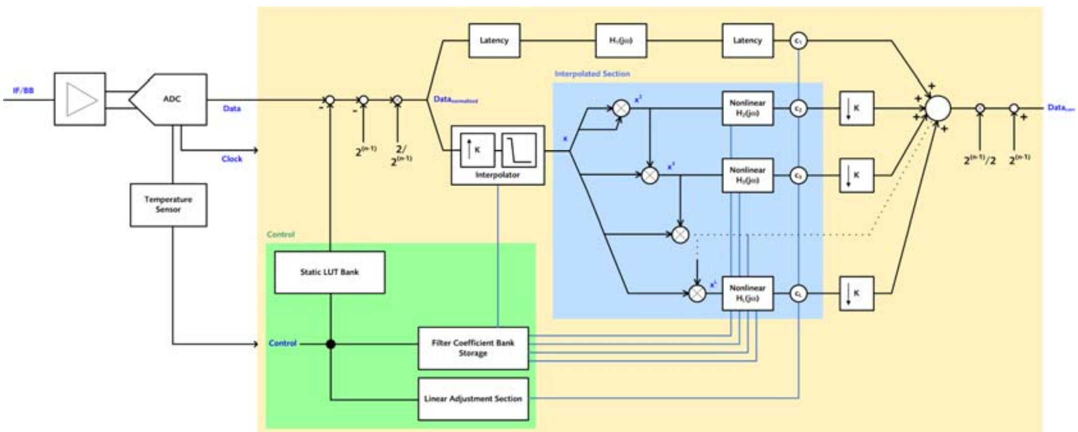

Figure 3: ADC Correction Scheme

Comparing the correction results with our own model, it was observed that ADIsimADC does not capture memory effects that cause the distortion terms to loose their in-phase ability, which is important for dynamic correction at high frequencies. The next chapter presents correction results for a real ADC at frequencies beyond Nyquist.

IV. Correction Results for the LTC2208

The LTC2208 (n: 16 Bit, f_s: 120 MSPS, FPBW: 700 MHz) was studied using four corresponding evaluation boards of Linear Technology. This ADC is ideally suited for dynamic correction, as static INL is very low. For signal generation and clocking we used two Rohde & Schwarz SMA100 A and appropiate K&L tunable bandpass filters. We did measurements up to the 7th Nyquist zone (360-420 MHz) with 16 stepped sinewaves at various input amplitudes. Coherent sampling was ensured. SFDR, HD2 and HD3 showed good agreement with data sheet specifications. However, from the 2nd Nyquist zone on, the histogram and sinefit method showed insensitivity to out-of-phase harmonics, which caused very unpredictable results for the INL (e.g., 2nd order distortion was nearly not detected) and unsatisfactory results after correction. We therefore decided to use harmonic power levels and phases, derived from a 64K FFT, to evaluate nonlinearity in terms of amplitude and degree of phase shift. As the on-board FIFO is written at a random time, the initial phase shift φ_0 is also random. In order to get the true phase error over frequency (φ_{D2} and φ_{D3} in formula 2) a modulo algorithm was written in MATLAB. We present results from one evalboard in the 7th Nyquist zone. Figure 4 shows the dynamic phase error: We superposed measurements from input amplitudes ranging from -1 to -5 dBfs, but observed no notable dependency. Increased measurement error for HD3 is due to the nearby noise floor. To demonstrate the reproducibility of the offline calibration, we superposed additional time-distinct measurements.

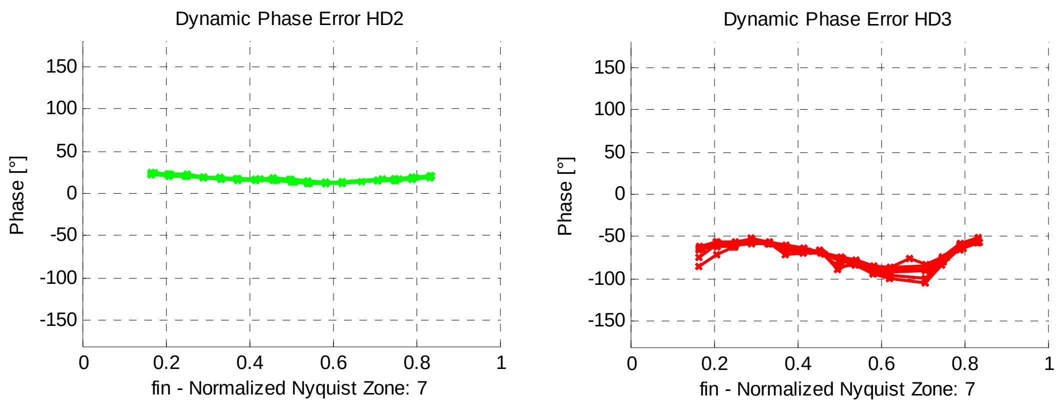

Figure 4. LTC2208 data: Dynamic Phase Error for HD2 and HD3 superposed for input amplitudes ranging from -1 to -5 dBfs, results are derived from time-distinct data sets

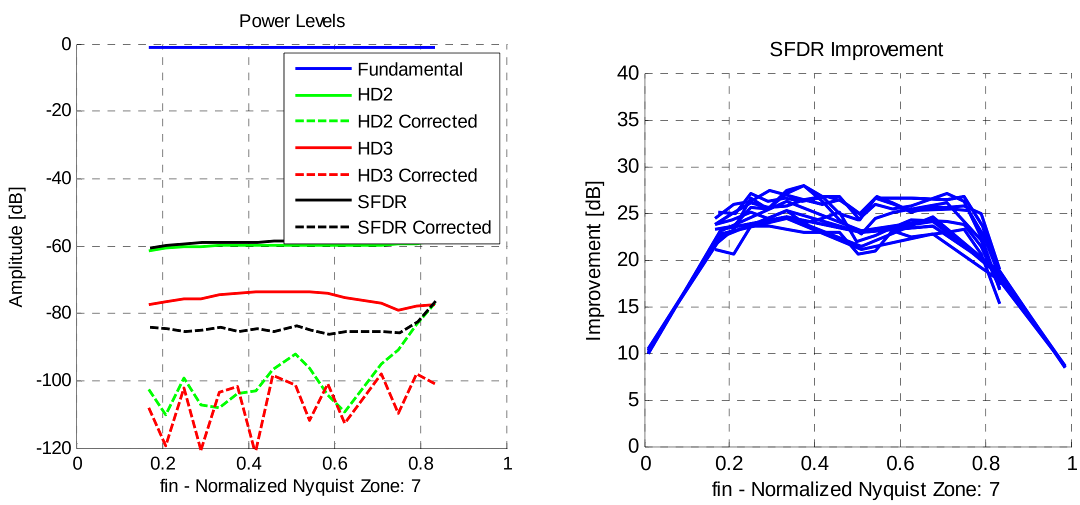

Figure 5. LTC2208 data: (left) SFDR and HD power levels before and after Correction with full scale excitation, (right) relative SFDR Improvement for amplitues ranging from -1 to -5 dBfs derived from time-distinct data sets

As the input circuitry on the evaluation board was intentionally left unoptimized for high frequencies, we measured slightly higher SFDR values in comparison to the data sheet. As mentioned, dynamic nonlinearity arises from the cascadation of input circuitry and track-and-hold amplifier. In order to capture the phase shift of harmonics we designed a nonlinear phase FIR filter based on phase specification derived from the FFT measurements. The filter was added to the nonlinearity branches in Figure 3. The filters must not match the phase shifts perfectly to achieve a satisfying suppression of distortion, which relieves the filter effort. We thus used an linear phase FIR (32 taps) to capture amplitude behaviour of harmonics and a nonlinear phase FIR (32 taps) with arbitrary group delay to capture phase behaviour of harmonics - we refer to detailed filter design methods in [24]. Figure 5 on the left side shows the power levels before and after correction, while the right side presents the relative dynamic range improvement. The training was carried out with full scale excitation. Correction was applied to new data taken at different frequency points. Lower amplitudes encountered similiar improvements, as the m-th order law was satisfied. This can be seen in the right plot of Figure 5, where SFDR improvement is superposed for input amplitudes ranging from -1 to -5 dBfs. Three time-distinct raw data sets were corrected to guarantee reproducibility and robustness. Finally, Figure 6 shows a representative spectrum for input frequency 400 MHz and input amplitude -1 dBfs before and after the correction was applied.

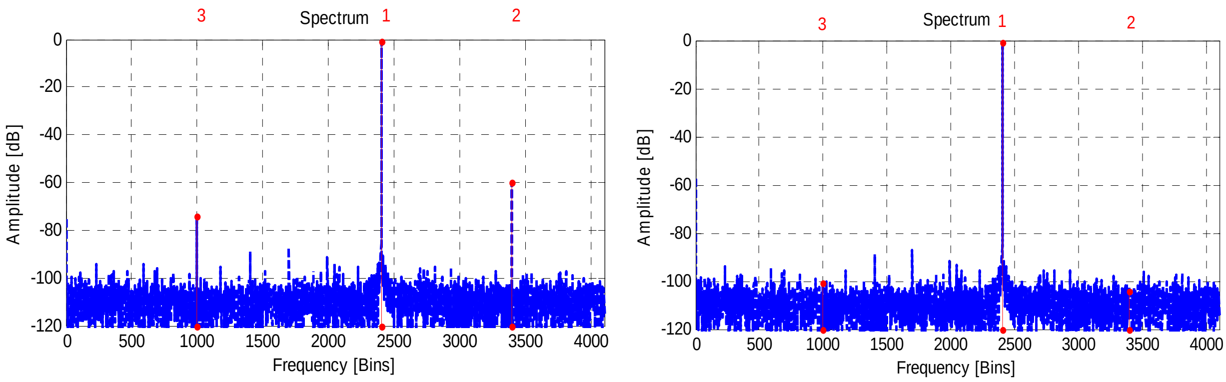

Figure 6: LTC2208: FFT Spectrum before and after correction with input frequency 400 MHz and full scale excitation, with the spectrum being plotted in the 1st Nyquist zone

The performance was boosted beyond data sheet specifications. As was shown for a wide range of frequencies and amplitudes, SFDR of the LTC2208 could be improved by 25 dB. No significant higher harmonics or spurs limited our results, a constant SFDR of 85 dBfs could be achieved in 60% of the 7th Nyquist Zone (corresponding to a 36 MHz bandwith around a centered IF of 390 MHz). An optimization of the input circuitry combined with higher filter effort (increased memory length) could allow further increased linearization results.

V. Conclusion

Digital correction methods have the potential to improve the dynamic range of future Radar and RF receivers. The challenge is to find a nonlinear, dynamic model that captures distortion behaviour of the ADC with respect to frequency and amplitude, while maintaining a justifiable degree of complexity. Once found, the inverted model can be used in digital domain to process ADC data and improve the system's SFDR. After discussing the important role of the ADC inside a Radar receiver architecture, we gave a state-of-the-art on correction principles deduced from relevant literature. The Volterra model and its constraint variants were discussed. Theory of modeling approaches and correction principles were presented. We showed that for IF-subsampling, or higher input frequencies in general, the internal ADC memory effects cause a phase rotation of distortion terms. Also, classical INL methods like the histogram are very insensitive to out-of-phase harmonics. This dynamic phase error must be incorporated into correction models that aim to work in higher Nyquist zones. By using real data from a 120 MSPS LTC2208 we investigated the relative phases of distortion terms using a FFT. A nonlinear phase FIR filter that captures the phase shift of harmonics was designed and added to the correction scheme. Raw data was processed in MATLAB, but the correction interpolators and filters have a realistic size, constrained for a FPGA implementation. For the LTC2208, we achieved up to 25 dB SFDR improvement in the 7th Nyquist zone (input frequencies 360-420 MHz), yielding 85 dBfs SFDR over the whole amplitude range. The reproducibility and robustness of the correction algorithm was demonstrated through measurements. Future work will focus on advanced filter design methods, behaviour of intermodulation distortion, part-to-part nonlinearity variations, robustness to temperature and the hardware implementation of the algorithm.

As the correction principle is not limited to ADC correction, it will be possible to include an analog amplifier or mixer into the receiver's postdistortion concept, refering to predistortion principles and their benefits in transmitters. An increasing number of scientific contributions emerged on the subject over the last 5 years and we expect this trend to continue: Systems can greatly benefit from a holistic approach, that explores the best possibilities from both analog and digital world.

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