Accurate Full Spectrum Test Robust to Simultaneous Non-Coherent Sampling and Amplitude Clipping

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Abstract

For spectral testing of Built-in Self-Test Analog to Digital Converters, it is a very challenging task to precisely control the amplitude and frequency of input sinusoid signal. Amplitude over-range results in clipping ADC output and non-coherent sampling results in spectral leakage. In this paper, a new method is proposed that provides accurate spectral results even when the input to ADC is both over-ranged and non-coherently sampled. This relaxes the condition to have precise control over the input signal and thus decreases the cost. The method includes fundamental identification, removal and residue interpolation to obtain accurate spectral results. Simulations show the functionality and robustness of proposed method with both non-coherency and amplitude over-range. Measurement results of a commercially available 16-bit SAR ADC are used to verify the method for both functionality and robustness.

1. Introduction

With advancements in technology, more complex circuits such as System on Chips (SoC) are designed. Though this approach decreases the design cost by embedding more circuits on a single chip, it increases the cost associated with testing such systems. This is because, it becomes both challenging and expensive to de-embed each block separately for testing. In order to decrease the test cost, there is a strong drive to design Built-in Self-Test (BIST) circuits in which both the testing circuitry and Device under Test (DUT) are on the same chip.

Analog-to-Digital Converters (ADC) are one of the most widely used integrated circuits (IC) in SoCs. ADCs are usually tested for static characteristics such as Integral Non-linearity (INL) and dynamic characteristics such as Total Harmonic Distortion (THD) and Spurious Free Dynamic Range (SFDR) [1-2]. Spectral testing of ADCs is also called AC testing and includes testing of ADCs dynamic characteristics. A method is said to perform Full Spectrum test when the method not only tests for dynamic specifications but also focuses on testing all spectral bins including harmonic and non-harmonic bins. Being able to perform Full Spectrum test is especially important for systems whose SFDR is limited by non-harmonic spurious tones, such as time-interleaved ADCs. The test setup for both spectral test and Full spectrum test is the same as shown in Fig. 1.

To accurately perform spectral testing, the IEEE standard for Digitizing Waveform Recorders (IEEE Std. 1057) [3] and IEEE standard for Terminology and Test Methods for Analog-to-Digital Converters (IEEE Std. 1241) [4] recommend the test setup to satisfy the following five conditions. The first condition is to have an input signal that is at least 3-4 bits more pure than the ADC under test. The second condition is that the relative jitter between the input signal and clock signal should be very less. The third condition is to sample the input signal coherently. The fourth condition is that the amplitude of the input signal should be slightly lower than the ADC input range so that the ADC output is not clipped. The fifth condition is to have sufficient data record length. It can be mentioned that the initial four conditions are challenging to achieve, especially for BIST ADCs. This work aims at relaxing two of these challenging requirements to enable BIST ADC design.

In BIST ADCs, since both the testing circuitry and the ADC are present on the same chip, it would be challenging to achieve precise control over frequency and amplitude of the test input sinusoid signal. Due to imprecise frequency control, coherent sampling cannot be achieved unless a master clock is used. However, using master clock on-chip is not an attractive solution as it increases the silicon area. On the other hand, due to imprecise amplitude control, there could be cases when the input signal to ADC exceeds the input range of ADC. Such cases result in clipped ADC output. Both these situations could occur in BIST ADCs and can result in grossly wrong spectral results if Discrete Fourier Transform (DFT) is performed on such data. As a result, it is important to design a robust method that can accurately test the dynamic characteristics of an ADC even when the input signal is slightly over-ranged and is non-coherently sampled.

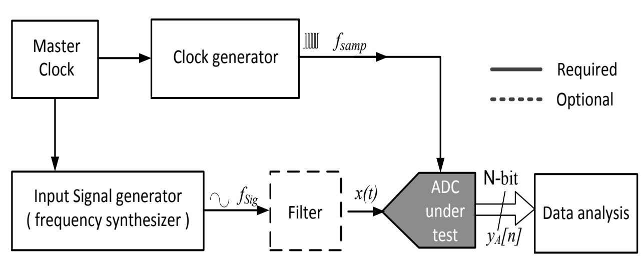

Fig. 1: Setup to test ADC Spectral characteristics.

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In the literature, several methods have been proposed to obtain accurate spectral results when the input is non-coherently sampled. Such methods include windowing method [5], interpolating DFT method [6], singular value decomposition method [7], four parameter sine fitting method [8] and fundamental identification and replacement methods [9-11]. However, in the presence of clipped ADC output, none of the above methods provide accurate spectral results.

The issue of over-ranged input was discussed in the recent past. In [12], a method to identify the fundamental and estimate ENOB when the input is over-ranged was proposed. However, the method cannot be used for high resolution ADCs and also cannot estimate all spectral parameters accurately. In [13], a technique to suppress the spurious noises generated by ADC clipping using interpolation of clipped samples was proposed. The method involves oversampling, polynomial spline and sinc function interpolation which is complex. In [14], oversampling ADC output was used followed by polyphase decomposition to compensate for clipping.

It can be said that none of the methods mentioned above can accurately test spectral characteristics of high resolution ADCs when the input is simultaneously over-ranged and non-coherently sampled. For BIST ADCs to be practical, it is important to develop such a method as it is challenging to achieve precise control over frequency and amplitude of input signal. Other advantage of using such a method is that the whole range of ADC can be tested [12].

In this paper, such a method that can accurately test the spectral characteristics of ADC when the input signal is slightly over-ranged and is not coherently sampled is proposed. The input over-range is limited to 2% of the input range of ADC. This is a valid and practical limit for BIST ADCs as amplitude of the on-chip input signal to ADC can be controlled up to 2% without any challenges. The proposed method involves accurate estimation of the fundamental component and later subtracts the estimated fundamental from the output of ADC to obtain the residue. The residue is then interpolated to obtain accurate information of ADC's harmonics. The performance of the proposed method with respect to jitter and noise would be similar to the method using coherent sampling.

The remainder of the paper is presented as follows. In section II, a brief overview of non-coherent sampling and ADC output clipping is presented. In Section III, a new method is proposed that can accurately estimate spectral characteristics when the ADC output is both clipped and non-coherently sampled. In Section IV, simulations are presented that show the accuracy and robustness of the proposed method. In Section V, measurement results are shown to validate the functionality of proposed method and the advantages and limitation of the method are briefly discussed in Section VI. Section VII concludes the paper.

2. Effect of non-coherent sampling and ADC clipping

Before discussing about the effects of non-coherent sampling and ADC clipping, a brief overview about the ADC input range is presented. For a given N-bit ADC with a gain of 1 and no offset, let T (0 to V_{ADC}) be the total input range of the ADC that covers all the codes from 0 (000...0) to 2^N-1 (111...1) as shown in Fig. 2. For any input below 0, the output is clipped at 0(000...0) and for any input above V_{ADC}, the output is clipped at 2^N-1 (111...1). The linear range of the ADC is the range in which the ADC is tested and is recommended to be operated. In major applications, the total range of ADC is tested. In such cases, T would be the linear range. However, in some applications, ADCs are tested only for a partial range in which they are intended to be applied. As shown in Fig. 2, the linear range of the ADC that needs to be tested is given by TL (= $F_t - F_b$), where F_t and F_b are the top and bottom values of range TL respectively. In such cases, if the input is below F_b (above 0) or above F_t (below V_{ADC}), the ADC provides a valid code and does not clip. Taking spectrum of such output would result in pessimistic results as the tested results do not correspond to the actual linear input range. From this point, in this paper, the term "input range of ADC" corresponds to the linear input range of the ADC that is tested. If T is the input range of ADC that is to be tested, then, $F_t = 2^N-1$ and $F_b=0$.

Let f_{Sig} be the frequency of input signal to ADC, f_{Samp} be the clock frequency, M be the total number of data points recorded to measure the spectral characteristics and J be the total number of periods of the input signal sampled in M points. The four parameters are related by equation (1).

$$J = M \frac{f_{Sig}}{f_{Samp}} \qquad (1)$$

The M point data record is said to be coherently sampled if J in (1) is an integer and, non-coherently sampled if J is not an integer.

Let the input range of ADC under test be [F_b F_t] as shown in Fig. 2. Let X(t) be the time domain representation of analog input to ADC at time t. **X** is ideally a pure sine wave and is given by (2).

$$X(t) = V_{OS} + A\cos\left(2\pi f_{Sig} t + \phi\right) + w(t) \qquad (2)$$

where, A and ϕ are the amplitude and initial phase of the fundamental respectively, V_{os} is the DC level and w(t) is the noise at time t. The conditions to obtain in-range and over-range input signals are given by (3) and (4) respectively.

$$(V_{OS} + A <= F_t) \text{ AND } (V_{OS} - A >= F_b) \qquad (3)$$

$$(V_{OS} + A > F_t) \quad \text{OR} \quad (V_{OS} - A < F_b) \qquad (4)$$

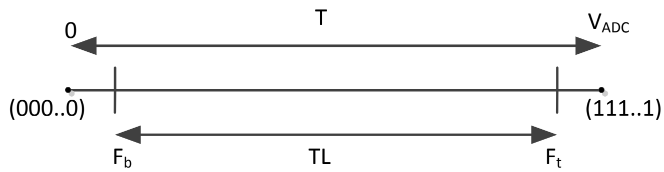

Fig. 2: Figure showing the total range, T (Codes 000…0 to 111..1) and the linear input range, TL (F_t to F_b) of ADC. T = TL if, F_t = 111..1 and F_b = 000..0.

The standard setup to test an ADC is shown in Fig. 1. The amplitude of the input signal is slightly below the ADC input range and the input signal is coherently sampled. Later DFT is performed on the data to obtain a spectrum as shown in Fig. 3. It can be seen that the spectrum is clean and total power of fundamental is present in a single bin corresponding to the frequency and so is the case with all harmonics. As a result, power of each harmonic component can be calculated accurately. This procedure gives accurate spectral results [1,9].

2.1 Non-coherent sampling

If the input signal is within the input range of ADC and is not coherently sampled, J in (1) is not an integer and equation (3) is true. Acquiring such data and taking DFT would result in a spectrum with huge leakage due to non-coherent sampling of fundamental component [9]. Such a spectrum leads to inaccurate test results as described in [9].

2.2 ADC Clipping

If the input signal is coherently sampled and is over-ranged (J in (1) is an integer and equation (4) is true), the output of ADC is clipped. The spectrum of such clipped data is shown in Fig. 4. It can be seen that severe distortion is introduced due to clipping which provides inaccurate spectral results.

2.3 Non-coherent sampling & ADC Clipping

If the input signal to ADC is simultaneously over-ranged and non-coherently sampled, the DFT of the output of such data would result in a spectrum as shown in Fig. 5. The spectrum not only has leakage due to non-coherent sampling but also has higher distortions due to clipped ADC output. The spectrum cannot provide accurate spectral results. As mentioned earlier, such cases could arise in BIST ADCs due to lack of precise control over amplitude and frequency of input. A test method that can accurately estimate all the spectral characteristics of ADC when the input is both over-ranged and non-coherently sampled is required.

From Fig. 5, it can be stated that, when DFT is performed on a non-coherently sampled, slightly clipped ADC output, the leakage and distortion in the spectrum is mainly due to the fundamental component in ADC output. The effect of non-coherent sampling can be eliminated by first accurately estimating the non-coherently sampled, over-ranged fundamental component in ADC output. The estimated fundamental is then clipped and subtracted from the ADC output to obtain the residue. This residue contains the information of harmonics and noise of ADC (at points when the ADC output is not clipped). Fig. 6 shows the spectrum of residue and it can be seen that leakage is eliminated.

The effect of clipping can be removed by constructing a coherently sampled, unclipped fundamental signal and adding the information of harmonics and noise at each code hit by the newly constructed fundamental. This information of harmonics and noise can be obtained by interpolating the residue from ADC output codes that are not clipped. A block diagram showing the summary of proposed method is shown in Fig. 7.

To identify the fundamental in clipped ADC output data, methods described in [10-11] cannot be used as they do not consider the effect of clipping. The methods were used for high resolution spectral testing where the distortion power is negligible compared to that of the fundamental power. However, in presence of clipping, the distortion power is no longer negligible compared to that of the fundamental. So, a new fundamental identification method is required that is valid for clipped data. In the following section, a method to accurately estimate the spectral characteristics of an ADC when an input is non-coherently sampled and is over-ranged is proposed. A new method to identify the fundamental component in a non-coherently sampled, clipped ADC output data is described. The process of interpolating the residue is explained in detail.

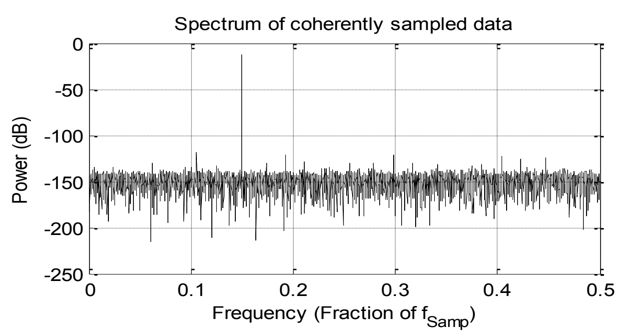

Fig. 3: Spectrum of a coherently sampled, unclipped ADC output data

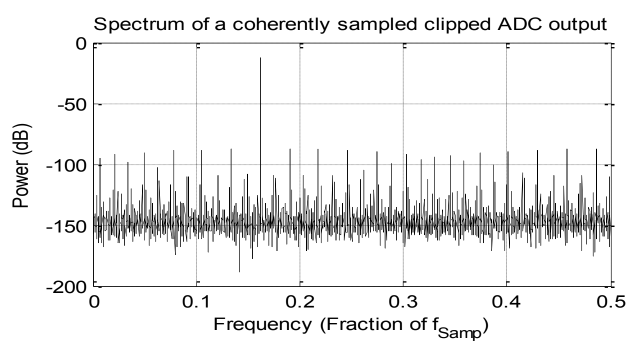

Fig. 4: Spectrum of a coherently sampled, clipped ADC output data

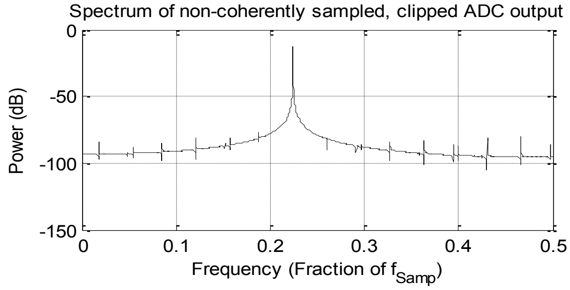

Fig. 5: Spectrum of a non-coherently sampled and clipped ADC output.

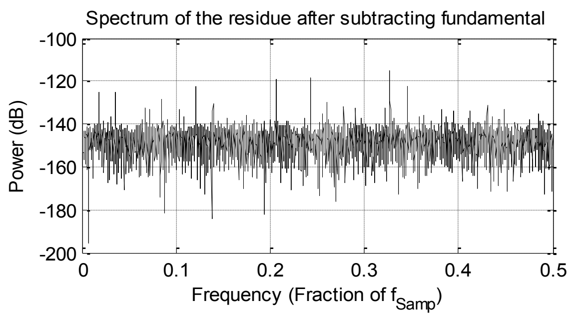

Fig. 6: Spectrum of the residue obtained after subtracting the non-coherently sampled, over-ranged fundamental from ADC output in Fig. 5. Leakage due to non-coherent sampling is eliminated.

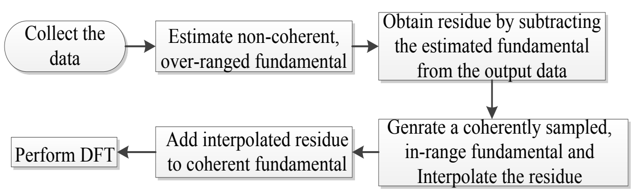

Fig. 7: Summary of proposed method.

3. Proposed Method

Let $x[n]$ be the n[th] sampled point of X(t). Let $y_A[n]$ be the analog interpretation of n[th] sampled digital output of ADC whose gain and offset are corrected. From (1), (2), (3), (4) and noting that the over-range up to 2% is considered, $x[n]$ and $y_A[n]$ can be represented by equations (5-6) respectively.

$$x[n] = V_{OS} + A\cos\left(\frac{2\pi J}{M}n + \phi\right) \tag{5}$$

$$\begin{aligned} y_A[n] &= V_{OS} + A\cos\left(\frac{2\pi J}{M}n+\phi\right) + \sum_{h=2}^{H} A_h \cos\left(\frac{2\pi hJ}{M}n + \phi_h\right) + w[n] && \text{if } 0 \le x[n] \le V_{ADC} \\ &= 0 && \text{if } x[n] \le 0 \\ &= V_{ADC} && \text{if } x[n] \ge V_{ADC} \end{aligned} \tag{6}$$

for n = 0,1,2,….,M-1, w[n] is the noise in n[th] sample, A_h and ϕ_h respectively contain the information of amplitude and phase of h[th] harmonic of ADC such that A_h<<A and $\phi_h \epsilon (0\ 2\pi]$. M is usually selected to be a power of 2 for faster processing of Fast Fourier Transform (FFT). If $y_A[n]$ is non-coherently sampled, J is not an integer and is given as the sum of an integer part J_{int} and a non-integer part δ ($J = J_{int} + \delta$)

Before the fundamental is identified, in order to test ADC within the input range given by [F_b F_t], equation (6) can be changed to (7). This includes clipping the values of $\boldsymbol{y_A}$ that are not in the ADC input range. It should be noted that if $F_t = V_{ADC}$ and $F_b = 0$, equation (6) is equal to equation (7).

$$\begin{aligned} y[n] &= y_A[n] && \text{if } F_b \le y_A[n] \le F_t \\ &= F_b && \text{if } y_A[n] < F_b \\ &= F_t && \text{if } y_A[n] > F_t \end{aligned} \tag{7}$$

The sample waveforms of $\boldsymbol{x}$ and $\boldsymbol{y}$ are shown in Fig. 8. As input, $\boldsymbol{x}$, exceeds the input range of ADC, $\boldsymbol{y}$ is clipped. With this clipped data, the fundamental component is estimated as described in the following section.

3.1 Fundamental Identification

From (6-7), in order to identify the fundamental, it is required to estimate V_{OS}, A, J_{int}, δ and ϕ. All the five parameters can be estimated using both time domain and frequency domain data. It should be noted that there is no information of fundamental or harmonics in the points that are clipped.

3.1.1 Estimate A and V_{OS}

From equations (6-7), it can be seen that $\boldsymbol{y}$ contains J cycles of the input signal. All the points in $\boldsymbol{y}$ and $\boldsymbol{x}$ are folded in to a single cycle as shown in Fig. 9 to obtain $\boldsymbol{y_1}$ and $\boldsymbol{x_1}$ as given by (8-9) respectively. The effect of harmonics is neglected since A_h<<A.

$$x_1[n] = V_{OS} + A\cos\left(\frac{2\pi n}{M} + \phi\right) \tag{8}$$

$$\begin{aligned} y_1[n] &= V_{OS} + A\cos\left(\frac{2\pi n}{M} + \phi\right) && \text{if } F_b \le x_1[n] \le F_t \\ &= F_b && \text{if } x_1[n] \le F_b \\ &= F_t && \text{if } x_1[n] \ge F_t \end{aligned} \tag{9}$$

Let K_t and K_b be the total number of points in $\boldsymbol{y_1}$ that are equal to F_t and F_b respectively. Let φ and ψ be the phases in $\boldsymbol{y_1}$ when the clipping stops at F_t and clipping starts at F_b respectively as shown in Fig. 9. Using K_t and K_b, the values of φ and ψ are obtained from equation (10).

$$\varphi = \frac{K_t - 1}{M}\pi, \quad \psi = \pi - \frac{K_b - 1}{M}\pi \tag{10}$$

Substituting F_t and F_b for $\boldsymbol{y_1}$ in (9) at phases φ and ψ respectively, the values of A and V_{os} can be estimated (11-12).

$$A = \frac{F_t - F_b}{\cos\left(\frac{K_t - 1}{M}\pi\right) + \cos\left(\frac{K_b - 1}{M}\pi\right)} \tag{11}$$

$$V_{OS} = \frac{F_t + F_b + A\left\{\cos\left(\frac{K_b - 1}{M}\pi\right) - \cos\left(\frac{K_t - 1}{M}\pi\right)\right\}}{2} \tag{12}$$

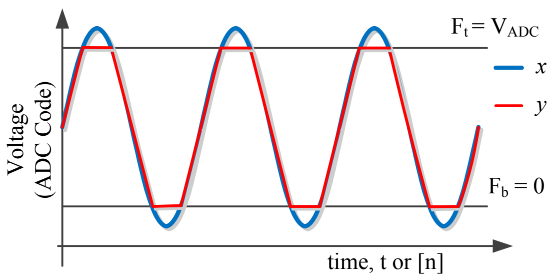

Fig. 8: Plot showing a sample waveform of input to ADC, $\boldsymbol{x}$ (Eq. 5) and modified output of ADC, $\boldsymbol{y}$ (Eq. 7).

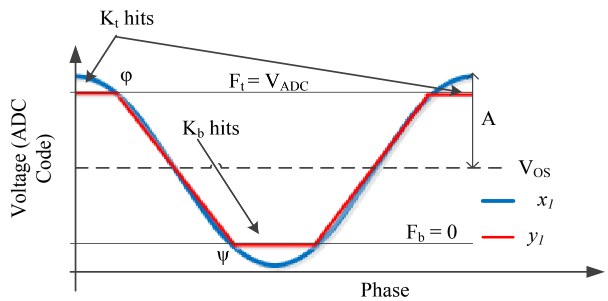

Fig. 9: Plot showing signals $\boldsymbol{x_1}$ and $\boldsymbol{y_1}$ after folding $\boldsymbol{x}$ and $\boldsymbol{y}$ in to one cycle. K_t and K_b are total number of points in $\boldsymbol{y_1}$ (or $\boldsymbol{y}$) equal to F_t and F_b respectively.

It should be noted that it is not required to perform this folding operation in the algorithm. It was used only to elucidate the procedure to estimate A and V_{OS}. From (11-12), to estimate A and V_{OS}, the values of F_t, F_b, K_t and K_b are required. F_t and F_b are known from the input range of ADC while K_t and K_b can be obtained by processing $\boldsymbol{y}$ directly.

3.1.2 Estimate J_{int}, δ and ϕ

The remaining three parameters are estimated using both time and frequency domain data of $\boldsymbol{y}$. DFT is applied on $\boldsymbol{y}$ to obtain Y_k {k^{th} DFT coefficient} given by (13).

$$Y_k = \sum_{n=0}^{M-1} y[n] e^{-j\frac{2\pi k}{M}n}, \quad \text{for } k = 0, 1, 2, ..., M-1 \tag{13}$$

J_{int} is estimated as the bin index that has the maximum power in half spectrum excluding DC component and is given by equation (14).

$$J_{\text{int}} = \underset{1 \le k \le M/2}{\arg\max} |Y_k| \tag{14}$$

From [15], to obtain the initial estimates of δ and ϕ, for M>1024, Y_k is given as (15)

$$Y_k \approx \frac{A}{2M} e^{j\phi} \frac{1-e^{j2\pi(J-k)}}{1-e^{j\frac{2\pi(J-k)}{M}}} \tag{15}$$

Using $Y_{J\text{int}}$, $Y_{J\text{int}+1}$ and $Y_{J\text{int}-1}$, from (15), the initial values of δ and ϕ can be obtained using (16-17) respectively.

$$\delta_0 = \frac{M}{2\pi} imag\left(\ln\left(\frac{\frac{Y_{J_{\text{int}}}}{Y_{J_{\text{int}}+1}} - \frac{Y_{J_{\text{int}}}}{Y_{J_{\text{int}}-1}}}{\frac{Y_{J_{\text{int}}}}{Y_{J_{\text{int}}+1}} - \frac{Y_{J_{\text{int}}}}{Y_{J_{\text{int}}-1}} + e^{j\frac{2\pi}{M}} - e^{-j\frac{2\pi}{M}}}\right)\right) \tag{16}$$

$$\phi_0 = -imag\left(\ln\left(\frac{2MY_{J_{\text{int}}}}{A}\frac{1-e^{j\frac{2\pi\delta_0}{M}}}{1-e^{j2\pi\delta_0}}\right)\right) \tag{17}$$

Since, Y_k corresponds to a non-coherently sampled data, the values of δ and ϕ obtained are not accurate. Let $\Delta\delta$ (=δ-δ_0) and $\Delta\phi$ (=ϕ-ϕ_0) correspond to errors in estimating δ and ϕ using (16-17) respectively. It is required to estimate these errors in order to improve accuracy of estimated δ and ϕ.

Substituting the estimated values in $\boldsymbol{y}$ (eq. 7) gives equation (18) and rearranging the terms, we get (19). The effect of harmonics is neglected and "n" in (18-19) correspond to the samples that are not clipped. The points in $\boldsymbol{y}$ that are close to V_{os} (mid-range codes) are considered in (19) as shown in Fig. 10. These points are considered because, they are linear in the range and give good estimates of $\Delta\delta$ and $\Delta\phi$ when Least squares is applied on (19). After estimating $\Delta\delta$ and $\Delta\phi$, using (11-12), (14), (16-17) the initial estimated fundamental, $\boldsymbol{z_i}$, is given by (20).

$$y[n] = V_{OS} + A\cos\left(\frac{2\pi(J_{\text{int}} + \delta_0 + \Delta\delta)n}{M} + \phi_0 + \Delta\phi\right) \tag{18}$$

$$\frac{2\pi n}{M}\Delta\delta + \Delta\phi = \cos^{-1}\left(\frac{y[n]-V_{OS}}{A}\right) - \phi_0 - \frac{2\pi(J_{\text{int}} + \delta_0)n}{M} \tag{19}$$

$$z_i[n] = V_{OS} + A\cos\left(\frac{2\pi(J_{\text{int}} + \delta_0 + \Delta\delta)}{M}n + (\phi_0 + \Delta\phi)\right) \tag{20}$$

To further improve the accuracy in estimating A, let ΔA be the error in estimating A (eq. 11). ΔA can be estimated by first clipping $\boldsymbol{z_i}$ to obtain $\boldsymbol{z_{ic}}$ and subtracting $\boldsymbol{z_{ic}}$ from ADC output, $\boldsymbol{y}$, to obtain the error signal, $\boldsymbol{ez}$ as shown in (21-22). Later, DFT is applied on $\boldsymbol{ez}$ and the fundamental amplitude in $\boldsymbol{ez}$, ΔA, is estimated using (23) (from eq. (15)). $\tilde{Y}_k$ in (23) is the k^{th} DFT coefficient of $\boldsymbol{ez}$. Now the actual fundamental component in $\boldsymbol{y}$ can be estimated as $\boldsymbol{z}$ using (24) and is shown in Fig. 11a (green dotted plot).

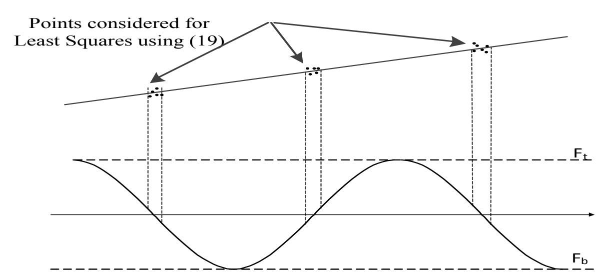

Fig. 10: Plot showing signal **y** and the points considered for Least squares using equation (19). Only output codes around mid-range are considered.

$$\begin{aligned} z_{ic}[n] &= z_i[n] && \text{if } F_b \le z_i[n] \le F_t \\ &= F_b && \text{if } z_i[n] \le F_b \\ &= F_t && \text{if } z_i[n] \ge F_t \end{aligned} \quad (21)$$

$$ez[n] = y[n] - z_{ic}[n] \quad (22)$$

$$\Delta A = 2M \left| \tilde{Y}_{J_{\text{int}}} \right| \left| \frac{1-e^{j\frac{2\pi\delta_0}{M}}}{1-e^{j2\pi\delta_0}} \right| \quad (23)$$

$$z[n] = z_i[n] + \Delta A \cos\left(\frac{2\pi\left(J_{\text{int}} + \delta_0 + \Delta\delta\right)}{M} n + \left(\phi_0 + \Delta\phi\right) \right) \quad (24)$$

Since, non-coherently sampled fundamental is the major source of error, subtracting the estimated fundamental, ***z***, from ADC output, ***y***, eliminates the effect of non-coherent sampling

3.2 Obtain Error (Harmonics + Noise of ADC) Information

The residue, ***e***, obtained after subtracting the estimated fundamental, ***z***, from modified ADC output, ***y***, is shown in Fig. 11b. It can be seen that the shaded regions contain no information about the ADC non-linearity as the input signal is not in the ADC input range. So, the information of harmonics and noise of ADC is present only in the points that are not clipped in ***y***. In order to include the dynamic effects of ADC, the error voltage, ***e***, is separated into two categories, FP and RP which correspond to the error voltage when the input signal is falling and rising respectively as shown in Fig. 11b. The error voltage in FP and RP categories is separated to be ***ef*** and ***er*** respectively. From Fig. 11a and 11b, for each sample in RP, the value of *er[n]* is plotted with respect to corresponding *y[n]* as shown in Fig. 12b. Similarly, *ef[n]* is plotted with respect to *y[n]* as shown in Fig. 12a. With this, error in both phases (Rising and Falling) with respect to the code hit by ADC is obtained. This error contains the information of harmonics and noise of ADC.

3.3 Coherent time domain data reconstruction

With the information of J_{int}, ϕ and ADC range (F_t and F_b), the input signal that is coherently sampled and covers the input range of ADC (without clipping) is obtained and is given by ***v*** in (25). It is required to add the information of harmonics of ADC on to each sample of the coherent fundamental ***v*** to accurately estimate the spectral characteristics.

It should be noted that the codes hit in ADC output, ***y***, are not the same codes that are hit using ***v*** (Coherent, unclipped). To obtain the errors corresponding to each code in ***v***, first, all points in ***v*** are folded into one cycle to get $\boldsymbol{v_1}$ given by (26) as shown in Fig. 13 (blue). Also, shown in Fig. 13 is the folded ADC output $\boldsymbol{y_1}$ (red). The information of error for each sampled point in the falling phase of $\boldsymbol{v_1}$ is obtained by using ***ef*** as shown in Fig. 14 (blue straight lines). For each sampled point in falling phase of $\boldsymbol{v_1}$ ($v_1[c]$), two codes in $\boldsymbol{y_1}$ ($y_1[a]$ and $y_1[b]$) that are in the falling phase and close to $v_1[c]$ are used and interpolation of the two errors at those codes (*ef[a]* and *ef[b]*) is performed to estimate the error at code $v_1[c]$ (given as *ev[c]*). The interpolation equation is as shown in (27). Similarly, the error information for the rising phase of $\boldsymbol{v_1}$ can be obtained by interpolating ***er*** (Red straight lines in Fig. 14).

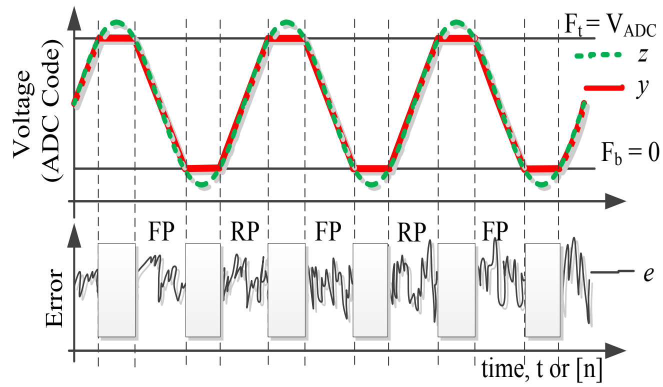

Fig. 11: a): TOP :- Figure showing ADC output, ***y*** and estimated fundamental, ***z***. b) BOTTOM :- Figure showing the error (Harmonics + Noise) information, ***e***, of ADC. (RP: Rising Phase, FP: Falling Phase, Shaded: Neglect). Both axes share the same x-axis (time)

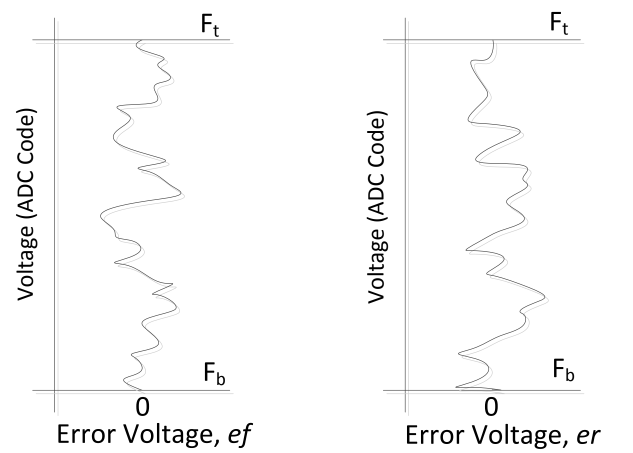

Fig. 12:a) LEFT:- Plot of error in FP, ***ef*** versus ADC output code, ***y***. b) RIGHT:- Plot of error in RP, ***er*** versus ADC output code, ***y***.

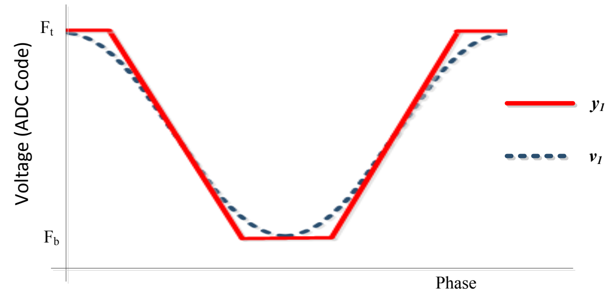

Fig. 13: Figure showing folded ADC output, $\boldsymbol{y_1}$ and folded coherent signal, $\boldsymbol{v_1}$.

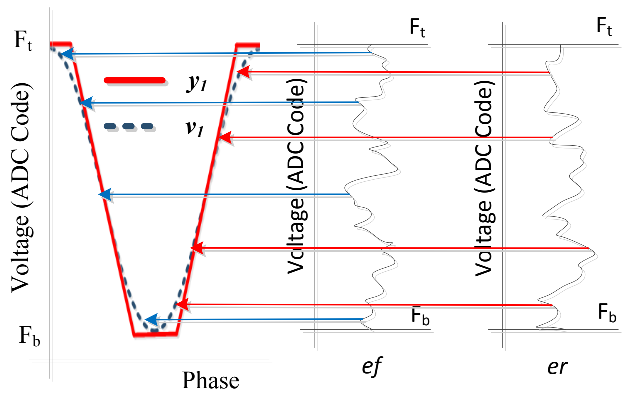

Fig. 14: Figure illustrating the interpolation of error onto $\boldsymbol{v_1}$. Error in $\boldsymbol{ef}$ is interpolated onto points in $\boldsymbol{v_1}$ in falling phase (Here left half of $\boldsymbol{v_1}$) while error in $\boldsymbol{er}$ is interpolated onto points in $\boldsymbol{v_1}$ in rising phase (Here right half of $\boldsymbol{v_1}$)

$$v[n] = \frac{F_t+F_b}{2} + \frac{F_t-F_b}{2}\cos\left(\frac{2\pi J_{\text{int}} n}{M}+\phi\right) \quad (25)$$

$$v_1[n] = \frac{F_t+F_b}{2} + \frac{F_t-F_b}{2}\cos\left(\frac{2\pi}{M}n+\phi\right) \quad (26)$$

$$\text{ev}[c] = \text{ef}[a] + \frac{(\text{ef}[b]-\text{ef}[a])}{(y_1[b]-y_1[a])}(v_1[c]-y_1[a]) \quad (27)$$

The information of error obtained for each code hit by $\boldsymbol{v_1}$ (in both rise and fall phases) is then added to $\boldsymbol{v_1}$ to get $\boldsymbol{f_1}$. $\boldsymbol{f_1}$ is then unfolded into J_{int} cycles to obtain $\boldsymbol{f}$. It should be noted that $\boldsymbol{f}$ contains not only the information of coherent fundamental but also the accurate information of harmonics and noise of ADC. Taking FFT of $\boldsymbol{f}$ would result in a spectrum that looks identical to the spectrum obtained using ideal test setup. Hence, accurate spectral characteristics of ADC can be obtained even with non-coherently sampled and clipped data using proposed method.

It can be argued that the dV/dt effects are not the same for both the signals $\boldsymbol{y}$ and $\boldsymbol{v}$ (from Fig. 13). However, since in this paper, only 2% over-range is considered, it can be mentioned that the dV/dt effect on harmonics' estimation is negligible

3.4 Need for interpolation

It should be noted that the step to perform interpolation for each point on the coherently sampled data ($\boldsymbol{v}$) is important to test high resolution ADCs. If interpolation is not performed, the residue obtained (after clipping and subtracting the estimated fundamental from the clipped ADC output) is directly added to the coherently sampled fundamental, $\boldsymbol{v}$, to obtain final data. DFT is performed on this data to perform spectral test. Let this process be called "Method B". Method B can be used only when the power associated with the clipped points is less than the noise power of the ADC. Hence Method B can be used to test only low resolution ADCs. However, as the resolution of ADC increases, the noise power of ADC decreases and Method B cannot be used. For instance, consider testing a 12-bit ADC and 16-bit ADC with non-coherently sampled, 1.7% over-ranged input signal. The spectral results obtained using three methods are given in Table 1 and Table 2. The first method is the standard method with unclipped and coherently sampled ADC output. The second method is using Method B and the third method is using the proposed method on the same ADC with non-coherently sampled, over-ranged input. From Tables 1 and 2, it can be seen that Method B provides accurate results only for the 12-bit ADC. However, the proposed method can be used to test both low and high resolution ADCs accurately. Hence, it is important to perform residue interpolation to obtain accurate spectral results when an input to ADC is over-ranged and is non-coherently sampled.

TABLE 1: Spectral results of a 12-bit ADC

Method	THD (dB)	SFDR (dB)
Coherent Unclipped (Reference)	-70.3	73.9
Method B (No interpolation)	-70.5	74.7
Proposed Method(With Interpolation)	-70.8	74.3

TABLE 2: Spectral results of a 16-bit ADC

Method	THD (dB)	SFDR (dB)
Coherent Unclipped (Reference)	-93.1	97.7
Method B (No interpolation)	-79.7	84.2
Proposed Method(With Interpolation)	-93.3	97.7

The flow chart to perform accurate ADC spectral testing with non-coherent sampling and over-ranged input using the proposed method is shown in Fig. 15. The steps in solid rectangle can be used when the output is non-coherently sampled while the steps in dotted rectangle can be used when the output is clipped

3.5 Comparison with 4 parameter sine fit

The fundamental identification method in Section A can be replaced with the four parameter sine fit method as described in [4] with slight modification. Both the methods provide accurate estimates of the fundamental. However, the proposed method is more computationally efficient than the four parameter sine fit method. Using four parameter sine fit method, all unclipped points in data are considered to perform non-linear least squares. This includes large data set and several iterations to obtain convergence which consumes large computation time. However, in the proposed method, the time consuming blocks are FFT and linear least squares method. As M is usually selected to be a power of 2, FFT consumes very small amount of time. Since a very small number of points around the mid-range codes are considered for linear least squares (as shown in Fig. 10), this operation also does not consume more time. The other factor which makes the proposed method more time efficient compared to the four parameter sine fit method is that, there is no necessity to perform iterations

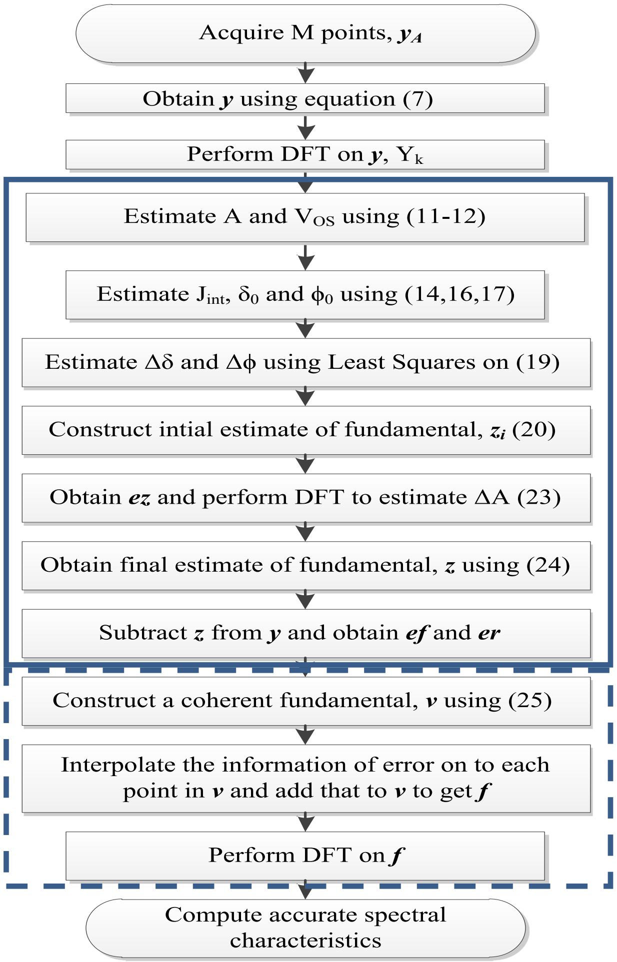

Fig. 15: Flow chart to perform accurate ADC spectral test using proposed method on Non-coherently sampled, clipped ADC output. Solid rectangle steps used when signal is non-coherently sampled. Dotted rectangle steps used when output is clipped.

4. Simulation Results

The accurate functionality and robustness of the proposed method is shown using simulation results in this section.

4.1 Functionality

A 16-bit ADC with INL of 1.5LSB was generated using MATLAB. A total of 8192 points were sampled. The ADC was first tested with a sine wave that is coherently sampled and not clipped. The signal is generated such that it covers the ADC input range without getting clipped. The values of THD and SFDR obtained are considered as the reference values. The same ADC is later fed with an over-ranged, non-coherently sampled input signal. The output is processed using the proposed method and the values of THD and SFDR are compared.

Fig. 16 shows the spectrum of ADC output when it is coherently sampled. The value of J = 3241. It can be seen that there is no leakage in the spectrum and the values of THD and SFDR obtained are listed in Table 3. Fig. 17 shows the spectrum of the same ADC output when a non-coherently sampled, over-ranged input signal is fed to the ADC. The value of J is 3241.199 and over-range is 0.78%. It can be seen that there is both leakage and severe distortion in the spectrum. Later the same ADC output is processed using the proposed method and the spectrum obtained is as shown in Fig. 18. The spectrum is clean and it exactly matches with the spectrum obtained using coherent sampling. The values of THD and SFDR obtained using proposed method are listed in Table 3. From Table 3 and Fig. 18, it can be said that the proposed method accurately estimates the THD and SFDR of a non-coherently sampled, clipped ADC output.

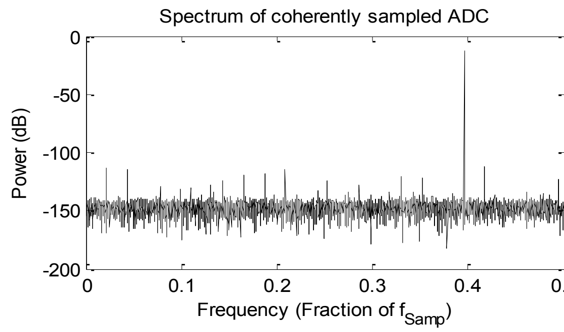

Fig. 16: Spectrum of a coherently sampled, unclipped ADC output (J = 3241)

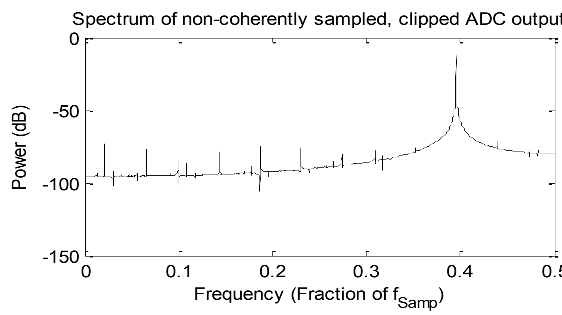

Fig. 17: Spectrum of DFT of a non-coherently sampled, clipped ADC output (J = 3241.199 → δ = 0.199, %over-range = 0.78)

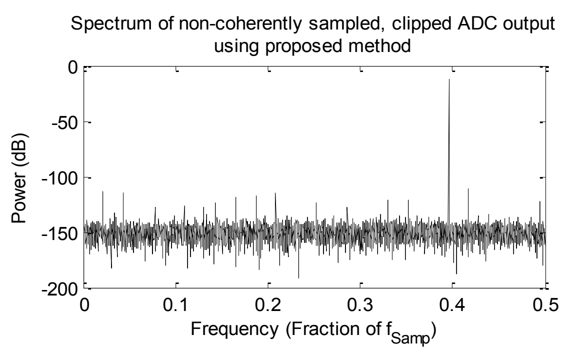

Fig. 18: Spectrum of a non-coherently sampled, clipped ADC output after using proposed method (J=3241.199→ δ = 0.199, % Over-range = 0.78)

TABLE 3: Spectral results of 16-bit ADC (Fig.16, Fig. 18)

Method	**THD (dB)**	**SFDR (dB)**
Coherent + Unclipped + DFT	-93.7	99.2
Non-coherent+Clipped + Proposed	-94.1	99.2

4.2 Robustness

The robustness of proposed method is shown with respect to non-coherent sampling and amplitude clipping up to 2%. A 16-bit ADC was generated using MATLAB with an INL of 1.8 LSB. A total of 500 runs with randomly selected values of δ and over-range were run. The values of δ vary from -0.5 to 0.5 (total range) and over-range percentage was in the range 0 to 2. The data record length for each run was 8192. Fig. 19 shows the errors in estimating the values of THD with any value of δ. Fig. 20 shows the errors in THD with change in percent input over-range. The maximum error obtained in estimating THD is about 0.9dB. This shows that the method is robust to both non-coherent sampling and amplitude over-range up to 2%.

5. Measurement Results

In this section, the proposed method is verified using measurement results from industry labs using a commercially available high resolution ADC.

The ADC that is used is ADS8318 which is a 16-bit Successive Approximation Register (SAR) ADC clocked at 500 kSPS. The input range of ADC is 0 to 5V. Fig. 21 shows the test setup. The input signal is followed by two band pass filters each with center frequency at 10 kHz. The output of the second band pass filter is fed to the input of ADC. A total of 8192 samples are collected.

The input frequency to achieve coherent sampling is given by 10.070800781 kHz which gives a value of J = 165. The blue spectrum in Fig. 22 shows the spectrum obtained using coherently sampled and unclipped data. It can be seen that there is no leakage in the spectrum.

Later, the frequency of input signal is changed to 10.0494768908 kHz and the input amplitude is slightly increased to be about 1.5% more than the ADC input range. As a result, the input signal is both non-coherently sampled (J=164.65, δ ≈ -0.35) and over-ranged (1.5%). The green plot in Fig. 22 shows the spectrum obtained without any correction. As expected, there is huge spectral leakage and higher distortion in the spectrum. The same time domain data is then processed using the proposed method. The red plot in Fig. 22 shows the spectrum obtained using the proposed method. It can be seen that the red spectrum (Non-coherent + Clipped + Proposed method) matches exactly with that of the blue spectrum (Coherent + Unclipped). The values of THD, SFDR and SNR obtained using standard coherent sampling method on unclipped data and proposed method on non-coherently sampled, clipped data are listed in Table 4. From Table 4 and Fig. 22, the accurate functionality of the proposed method with non-coherently sampled and over-ranged input is verified.

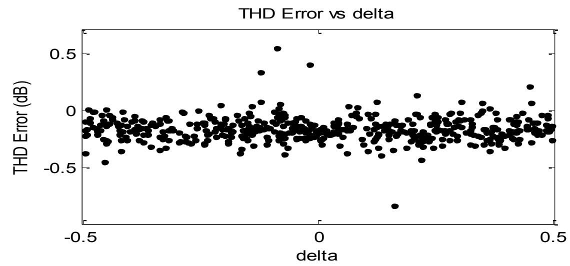

Fig. 19: Error in estimating THD values (in dB) using proposed method over whole range of δ.

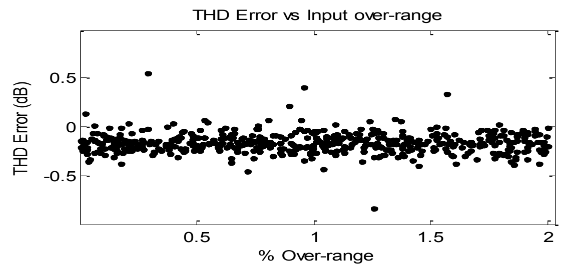

Fig. 20: Error in estimating THD values (in dB) using proposed method for different input over-range amplitudes.

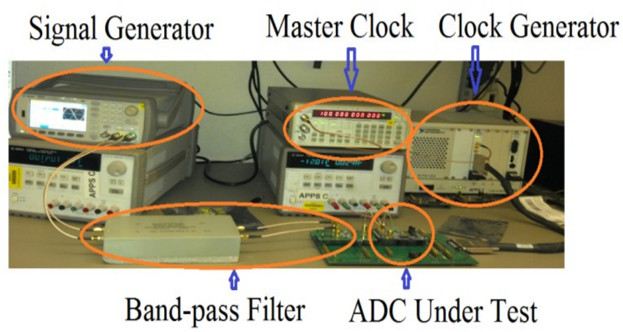

Fig. 21: Test setup for Measurement data of ADS8318.

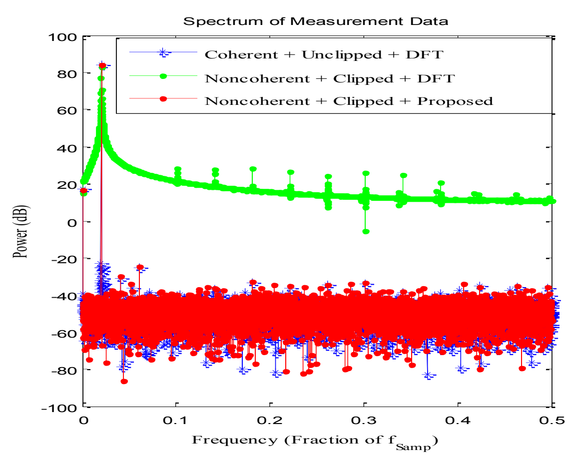

Fig. 22: Spectrums showing the accurate functionality of the proposed method. BLUE: Spectrum with Coherently sampled, unclipped data, GREEN: Spectrum with clipped and non-coherently sampled data, RED: Spectrum with Non-coherently sampled, Clipped data using the proposed method. (δ=-0.35, % over-range = 1.5) (For color version, visit IEEE online or contact authors)

TABLE 4: Spectral results of ADS8318 from Fig. 22

Method	THD (dB)	SFDR (dB)	SNR (dB)
Non-coherent + Clipped + Proposed (Red plot in Fig. 22)	-107.6	109.2	95.7
Coherent + Unclipped + DFT (**Standard**, Blue plot in Fig. 22)	-107.5	108.5	95.8

6. Discussion

From the above discussion, it can be said that the proposed method can relax two of the requirements to perform spectral testing. The advantages include reduction in the cost associated with achieving coherent sampling and with requiring precise amplitude control of the input. However, in the presence of noise and jitter, the method provides results with the same accuracy as the standard coherent sampling method. It was also shown that the proposed method is more computationally efficient compared to the four parameter sine fit method. One of the key salient features of the proposed method is to interpolate the error obtained, to achieve accurate results as shown in section 3.4. As a result, the method provides accurate results compared to the method in [12] which does not involve interpolation. Furthermore, the proposed method can perform full spectrum test unlike the method proposed in [12].

7. Conclusion

A new test method that accurately performs full spectrum test of an ADC with non-coherently sampled and over-ranged input was proposed. This relaxes the requirement to have precise control over frequency and amplitude of input signal for spectral testing. A new computationally efficient method to identify the over-ranged, non-coherently sampled fundamental using time domain and frequency domain data was described. The residue obtained after subtracting the estimated non-coherent fundamental is interpolated onto a coherently sampled signal to obtain accurate spectral results of ADC. Simulation results were presented to show the accurate functionality and robustness of the proposed method for any non-coherency and over-range up to 2%. The proposed method was validated using measurement results of a 16-bit SAR ADC when the output data is non-coherently sampled (δ = -0.35) and is clipped (1.5%). The method can be readily used in applications, where in, it is challenging to obtain precise control over frequency and amplitude of test signal, such as, BIST ADCs.

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