# Jitter and Decision-level Noise Separation in A/D Converters

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## Abstract

*Gaussian aperture jitter leads to a reduced Signal-to-Noise-Ratio of A/D converters. Also other noise sources, faults and nonlinearities affect the digital output signal. A measurement setup for a new off-chip diagnosis method, which systematically separates the jitter-induced errors from the errors caused by these other factors, is described. Deterministic errors are removed via a subtracting technique. High-level ADC simulations and measurements have been carried out to determine relations between the size of the jitter or decision-level noise and the remaining random errors. By carrying out two tests at two different input frequencies and using the simulation results, errors induced by decision-level noise can be removed.*

**Keywords:** jitter, decision-level noise, ADC testing, diagnosis.

## 1. Introduction

In A/D converters, the aperture time determines the time difference between the sample command and the actual time the analogue input signal is sampled and converted into a digital output signal. Variations in this aperture time are called aperture time uncertainty or aperture jitter [**Plassche 94**]. Gaussian aperture jitter originates from thermal Gaussian distributed noise sources within the ADC and Gaussian distributed noise sources from the environment (interference) [**Raczkowycz 96**]. Therefore, aperture jitter is likely to consist of both Gaussian noise components and some periodic components, depending on the environment in which the A/C converter is used.

In [**Harris 90**] it is shown that sinusoidal jitter components lead to sidebands of the signal in the output spectrum, which can be described by Bessel functions. Further it is shown that Gaussian jitter leads to a reduction in the signal-to-noise ratio of A/D converters.

Other noise sources, nonlinearities and faults in A/D converters also affect the output signal. In this paper, the systematic separation of the effects of aperture jitter from the effects of these other factors is considered. Only Gaussian jitter is taken into consideration. For detection of sinusoidal jitter components the results of [**Harris 90**] can be used.

In section 2, the different factors that influence the output signal of full-flash A/D converters are categorised. Our separation method is described in section 3. Deterministic error, like faults and nonlinearities, are removed via a subtracting technique. By carrying out tests at different input frequencies, also errors caused by jitter and decision-level noise can be separate too. The method has been evaluated using simulations on high level descriptions [**Rosing 98a**, **Rosing 98b**]. A measurement set-up has been made to prove the validity of the method proposed and will be described in section 4.

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## 2. Different factors affecting the output signal of A/D converters

The different factors that influence the output signal of A/D converters and their effects can be categorised, as shown in Table 1. A distinction is made between random sources and deterministic sources. Contrary to conventional noise sources, which are random in nature, quantisation "noise" is a deterministic "noise" source. This is caused by the fact that the size of the quantisation error is mathematically determined by the value of the input signal at the sample moment. However, often these quantisation errors lead to random noise on the output signal. In that case quantisation noise can not be distinguished from random noise sources.

| **Category of sources** | **Noise, error or distortion source** | **Effects** |
|---|---|---|
| Random noise sources | Aperture jitter | Noise on the output signal |
| | Noise at the decision levels | |
| Deterministic "noise" source | Quantisation noise | |
| Deterministic distortion sources | INL | Harmonic distortion components on the output signal |
| | Nonlinearities in the comparators | |
| Deterministic error source | DNL | Distribution of samples over the bins changes |
| Unknown | Faulty bias signals | Further research is needed to evaluate the effects |

**Table 1:** *Sources that determine the amount of noise and harmonic distortion on the output signal*

Aperture jitter and noise on the decision levels of A/D converters are both assumed to be white and therefore random noise sources. Nonlinearities in the comparators and Integral Nonlinearity (INL) of the converter lead to harmonic distortion in the output spectrum. In [**Kim 94**] the size of the INL is derived from the size of the harmonics in the output spectrum. In a histogram of output data, bins with Differential Nonlinearity (DNL) contain a number of samples that deviates from the ideal number. Therefore the amount of DNL of every bin can be determined using a code density measurement [**Doernberg 84**].

Both the described difference in origin of the different sources and their different effect on the output signal are used in section 3 to separate jitter from the other sources.

## 3. Separation of jitter from other error sources

### 3.1 Basics of the separation method

The separation method that is used to determine the size of the aperture jitter from the output data uses two steps, being:

1. The effects of deterministic sources are filtered out of the output data. Only random noise information remains (see section 3.2).
2. Aperture jitter is separated from the noise on the decision levels (see section 3.3).

As is explained in both sections 3.2 and 3.3 coherent sampling [**Mahony 87**] will be used in our test method.

In a coherent test the following equation holds:

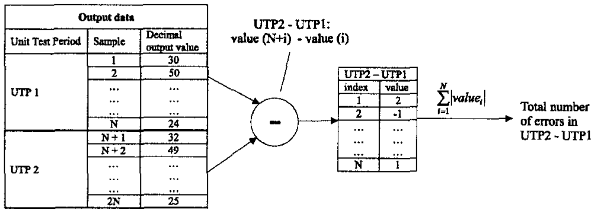
$$F_{in} = \frac{M}{N} \cdot F_{sample} \text{ , with M, N relatively prime} \quad \textbf{(1)}$$

where $F_{sample}$ : the sample frequency
$F_{in}$ : the frequency of the input signal
$M$ : the number of periods of the input signal captured during the test period
$N$ : the total number of samples taken

After M cycles of the input waveform (and thus after N samples) the sample sequence repeats itself. This is the shortest valid test window, called Unit Test Period (UTP).

### 3.2 Separation of random noise sources from deterministic noise and distortion sources

A test can be performed that captures two UTPs and thus captures 2N samples over 2M periods of the input signal. In such a test the set of output values obtained from the second UTP is essentially equal to the set of values obtained from the first UTP. This fact is used in [**Langard 94**], where a test is performed that contains two periods of a beat frequency test. This is a form of coherent testing in which M=N+1. The output values of the second period of the beat frequency are point to point subtracted from the output values of the first period. This is shown in Figure 1. Deterministic error sources cause the same errors in the first period as in the second period. Therefore, all deterministic error sources as mentioned in Table 1 are removed by this operation. However, random noise causes differences in the two data sets. The result after subtraction is therefore an array of errors caused by random noise sources.



| Output data | | |
|---|---|---|
| Unit Test Period | Sample | Decimal output value |
| UTP 1 | 1 | 30 |
| | 2 | 50 |
| | ... | ... |
| | ... | ... |
| | ... | ... |
| | N | 24 |
| UTP 2 | N + 1 | 32 |
| | N + 2 | 49 |
| | ... | ... |
| | ... | ... |
| | ... | ... |
| | 2N | 25 |

| UTP2 – UTP1 | |
|---|---|
| index | value |
| 1 | 2 |
| 2 | -1 |
| ... | ... |
| ... | ... |
| ... | ... |
| N | 1 |

**Figure 1:** *Method for separation of random sources from deterministic sources*

For every error in the table that contains the data after the subtraction of the two UTPs, there is a certain chance that it occurs. Therefore, there will be a relation between the total amount of errors after the subtraction and the size of the random noise. This relation is evaluated using high level ADC descriptions in section 4. The error values are added after the absolute value of each error is taken to avoid cancellation.

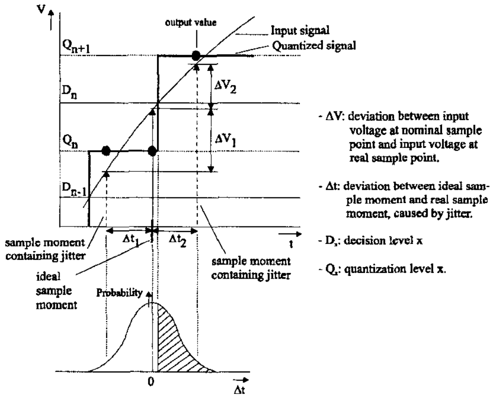
A disadvantage of a beat frequency method is that the input frequency has to be $\frac{N+1}{N}$ times the sample frequency, which puts high quality demands to the waveform generator in test system when high-speed A/D converters are being tested. Therefore, coherent sampling at a lower frequency is preferable and will be used in the separation method presented in the next subsection.

### 3.3 Separation of jitter induced errors from decision level noise induced errors

In a full-flash A/D converter the comparators perform both the quantisation and the sample and hold function at the same time. In Figure 2 the effect of aperture jitter on the output signal of a full-flash A/D converter is shown. The output values are indicated by dots. The solid arrow points at the value of the input signal at the nominal sample moment. The output value is now the digital code corresponding to the quantisation level $Q_n$. The dotted arrows depict sample moments containing jitter. For the sample moment with a positive delay, this leads to a faulty output value. In the other case the output value is unaffected.

The chance that jitter leads to the incorrect output value corresponding to the upper quantisation level is indicated by the shaded area in the Gaussian jitter distribution. The error chance depends on the time difference between the sample moment and the moment the decision level is crossed.

The deviation in sample time (Δt) causes a deviation in the voltage at the sample moment (ΔV) equal to the time deviation multiplied by the slope of the input signal.



- ΔV: deviation between input voltage at nominal sample point and input voltage at real sample point.
- Δt: deviation between ideal sample moment and real sample moment, caused by jitter.
- $D_x$: decision level x
- $Q_x$: quantization level x.

**Figure 2**: *the effect of aperture jitter on the output signal of a full-flash A/D converter*

Therefore the error chance depends both on the slope, and therefore also the frequency, of the input signal and the voltage distance between the input voltage at the sample moment and the decision-level voltage. The time probability distribution causes a voltage distribution with a standard deviation equal to:

$$\sigma V_{jitter} = \frac{dV(t_s)}{dt} \cdot \sigma t_{jitter}, \quad (2)$$

where $t_s$ denotes the sample time point

When a sine wave, $V_{in} = A \cdot \sin(2 \cdot \pi \cdot f_{in} \cdot t_s)$, is used as the input signal this gives:

$$\sigma V_{jitter} = 2 \cdot \pi \cdot f_{in} \cdot A \cdot \cos(2 \cdot \pi \cdot f_{in} \cdot t_s) \cdot \sigma t_{jitter}, \quad (3)$$

This distribution is maximum for points at the zero crossing of the input signal and minimum for points at the maxima and minima of the input signal.

On the other hand, the chance of an error caused by decision-level noise depends only on the distance of a sample to a decision-level and does not depend on the slope of the input signal.

If the samples taken in one UTP are reordered in ascending (or descending) order of voltage at the sample point, a single period of the input waveform is obtained [**Mahony 87**]. The resulting sample points are equivalent to the sample points when a single UTP is captured. Therefore, the sample values and thus also the distance of the samples to the decision levels does not depend on M, but only on N. The propagation of decision-level noise to errors in the output signal is therefore independent of M (for a certain N).

To separate aperture jitter induced errors from errors caused by decision-level noise the test shown in figure 1 has to be carried out twice:

1. The first time an input signal with a low frequency has to be used. If this frequency is chosen sufficiently low, no errors or only a small amount of errors are induced by aperture jitter. The amount of decision-level noise can now be determined from the amount of errors in the output signal.
2. The second time the test is carried out, a higher input frequency has to be used. Both the decision level noise and the aperture jitter now cause errors in the output signal. If the amount of errors is significantly higher then during the first period, the jitter propagation is dominant. The size of the aperture jitter can then be determined using the relation between the size of the jitter and the amount of errors. When the number of errors shows a less high increase, a relation between the total random noise and the amount of errors in (UTP2 - UTP1) has to be found.

The propagation of both decision-level noise and aperture jitter to errors in the output signal is minimum for converters with a small number of bits. Therefore, diagnosis of aperture jitter will be most difficult in these converters. In a 6 bits full-flash A/D converter the bin size is relatively large. The chance that noise on the decision-levels leads to errors larger then 1 LSB is therefore assumed to be negligible. The chance of a jitter-induced error is highest when both the standard deviation of the jitter and the frequency of the input signal are high. If, for example, an input signal of 20 MHz is applied to a 6-bits converter (A = 32 LSB) with a clock jitter of 100 ps. The maximal standard deviation of the voltage distribution is found to be $\sigma_{Vjitter}$=0.4 LSB using equation (3).

| **distance of sample to a decision level in number of σ of the noise** | **error change** | **# errors (N= 32768)** |
|---|---|---|
| 0.5 | 3.085E-01 | 10,108.90 |
| 1.0 | 1.587E-01 | 5,200.28 |
| 1.5 | 6.681E-02 | 2,189.23 |
| 2.0 | 2.275E-02 | 745.47 |
| 2.5 | 6.210E-03 | 203.16 |

**Table 2:** *Error changes and errors for 32768 ($2^{15}$) samples calculated using a numerical analysis procedure*

Note however that using equation (3) gives rather optimisic numbers since only at the zero crossing of the sine wave we have maximum probability of jitter effects.

A high-level simulation program has been written [**Rosing 98a**, **Rosing 98b**] to determine the relation between the size of the aperture jitter and the remaining random errors and between the size of the decision-level noise and these errors. Using these results the size of the decision-level noise can be determined from the first diagnostic session, that uses a low frequent input signal. The amount of errors found in the second section, that uses an input signal with a higher frequency, is used to determine if jitter-induced noise is dominant to decision level noise. If this is the case, the size of the jitter can be derived from this session.

## 4. Experimental Measurement set-up

The primary goal of our measurement set-up (see Figure 3) is to verify the predicted relations as indicated by equation (3). We use an IMS-ATS1 200MHz digital tester as master, which controls a VXI rack containing an HP E1445A arbitrary waveform generator (AWG). The digital tester generates a 10 MHz clock, to which artificially jitter is added via a comparator using a Rohde & Schwarz noise generator (R&S SUF BN4150). This jittering clock is used as sample clock by a 6 bit ADC (Philips ACE3A) under test. The input of the ADC is a sine wave generated by the AWG, which is synchronised with the 10 MHz using a Brüel & Kjær synchronisation module (BK3154) to be able to perform coherent testing over a sufficient long number of periods.

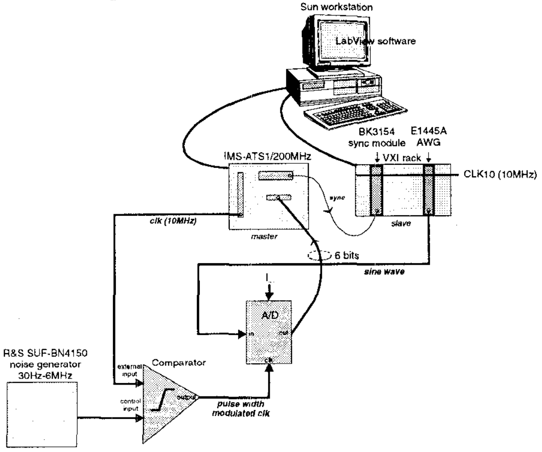


**Figure 3:** *Experimental Measurement set-up*

To be able to measure a reasonable number of errors among the number of samples N=2048 taken, we chose to use artificially high $\sigma t_{jitter}$ of 10 ns (=10%).

Using equation 1 and 3 and choosing appropriate periods M we obtained Table 3 shows the input frequency $F_{in}$ needed and the $\sigma V_{jitter}$.

| M | $F_{in}$ (Hz) | $\sigma V_{jitter}$ (LSB) |
|---|---|---|
| 101 | 493 164.0625 | 0.992 |
| 97 | 473 632.8125 | 0.952 |
| 89 | 434 570.3125 | 0.874 |
| 83 | 405 273.4375 | 0.815 |

**Table 3:** *Part of the measurements to be carried out*

In principle with this measurement set-up we should be able to prove the theory, but unfortunately we were not yet able to obtain convincing results. The main limiting factor is the stability of the reference current used by the comparators. This causes extra errors at the minima and maxima of the sine wave and hides the theoretical effect described.

## 5. Conclusions and further work.

A method has been described that enables off-chip diagnosis of aperture jitter. In this method jitter-induced errors are systematically separated from errors caused by other sources. The separation is carried out using two diagnostic sessions at two different input-signal frequencies. Within each session two Unit Test Periods of a coherent test are captured. Subtraction of the two essentially equal data sets obtained within a session removes all deterministic errors.

In [**Langard 94**] the same subtraction techniques where used to remove deterministic errors from the output data. However, in that method the aperture jitter must be high enough to cause errors of several LSBs. It can therefore only be applied to ADCs using a high number of bits. Our method can also be applied to ADCs with a lower number of bits. It is shown how it is possible to measure aperture jitter for a 6-bits ADC. Further, using our method it is possible to measure aperture jitter using a lower frequency of the input signal.

A measurement set-up has been made to validate the method introduced. In near future we will improve our set-up and obtain data to prove our method.

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