REAL TIME MEASUREMENTS OF HIGH RESOLUTION MIXED-SIGNAL CIRCUITS FOR SELF AWARE EMBEDDED SYSTEM

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Abstract: In this paper we discuss a methodology for efficient real-time measurements of high-resolution mixed-signal circuits implemented on the IC. The methodology could be used for real time built-in self-tests of a fail-safe mixed-signal integrated circuits and as a measurement part of a self-aware algorithm and methodology for integrated mixed-signal circuits. We show that a pseudo-random noise signal is a good option for the signal source and that the methodology leads to the efficient and cost-effective measurements in real time. The measurement is running in parallel to the main signal processing. The method is theoretically analyzed and verified using Matlab models and simulations. As an example the response of high precision, high order Σ-Δ ADC with most important non-ideal effects is compared to the response of a bit-true model of a reference digital circuit. The differences are demonstrated using simple area-efficient cross-correlation algorithm that can be implemented in software or in digital hardware.

1 INTRODUCTION

Continuous advances in IC processing technologies offer a possibility to produce integrated circuits with increased complexity and performances for reduced cost. In addition, integrated circuits are more and more composed of heterogeneous embedded systems, with different kind of digital, analogue and mixed-signal circuits and sensors integrated on the same IC. In future, this number will increase and the complexity of all modules will increase as well. It is thus essential, that modules are built in such a way that monitoring their own states is possible, which means that the system is capable to measure some of its performance parameters and act according to that. Monitoring is an essential part of any self-adaptive and/or self-aware system. This is new and difficult topics for digital systems (Santambrogio et al., 2010) and completely new for embedded analogue and mixed-signal circuits. The problem of self-awareness of analogue circuits lies in the fact that this circuits are not flexible as digital circuits are, it is very difficult to measure their characteristics without expensive measurement equipment and without precision generators and usually they do not have any "built-in intelligence". The problem is even harder if high resolution mixed-signal embedded modules like Σ-Δ A/D converters are involved because they are complicated analogue structures, which are difficult to design and almost impossible to measure without high resolution instruments. The modules may also operate at high frequency, while their power consumption is limited to a minimum. Fortunately, modern heterogeneous systems always consists of digital hardware and/or software signal processing, which provide the opportunity for evaluation of monitored parameters; a precision analogue signal source or appropriate replacement is still needed if someone wants to measure the parameters of the analogue module. For systems where human life may be in danger in the case of the failure, the self awareness and thus the measurements of the most important parameters must be executed in real time, that is in parallel to the real operation of the system. Example of such self-aware system is for example electronic stability system in passenger car (Strle, 2007) where the measurement channels and the sensors are

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monitored in real time in parallel to the real operation. Many other examples that require fail-safe operation exist, therefore, it make sense to develop a methodology and a general frame-work for RTBIST (real time measurements; we name it real-time-built-in-self-test) of a mixed signal circuits. Such real time measurements are the basis for self-awareness of embedded analogue and mixed-signal circuits. To be able to measure characteristics of embedded mixed-signal module, a high precision generator is needed that is simple to built, requires small silicon area for the implementation and needs little power for the operation. Such generator is one bit pseudo random (PRN) signal generator described in subsection 2.2.

The rest of the paper is organized as follows. In section 2 the principles of measurements using pseudo random signal are explained. Section 3 deals with measurements of high resolution Σ-Δ modulator using PRN source. Section 4 introduces a real time measurements of important parameters that is going on in parallel to the normal operation, while section 5 deals with implementation of efficient classification circuit. Section 6 presents one example and section 7 concludes the article.

2 PRN MEASUREMENTS

The first step to reach or achieve self awareness is to measure important parameters of embedded mixed-signal circuit that can be any combination of interconnected analogue and/or mixed-signal and digital modules. Generally, to measure characteristics of such circuit a precision signal generator is needed with parameters better than the circuit to be measured. Such signal generator is not available on chip and would be very difficult, demanding and expensive to built. In addition, measured results must be evaluated with high precision and as fast as possible. Fortunately, on a mixed signal VLSI circuit a DSP is usually available and if designed properly it can execute efficiently the algorithms needed for the evaluation of the performance measurements in real time.

One possibility to measure the performances of a high resolution analogue or mixed-signal circuit is to compare the response of the analogue LTI system (continuous or discrete time) with digital discrete time system having the same architecture and equal coefficients. The system to be measured is high precision analogue or mixed-signal circuit implemented on the chip, while reference LTI system can be implemented on or off the chip in hardware or in software. If one bit pseudo random noise (PRN) generator with appropriate autocorrelation and approx. white PSD is used as a signal source a high precision and high linearity can be achieved easily and on a very small silicon area.

2.1 Theory of PRN Measurements

The easiest way to measure analog Linear-time-invariant (LTI) system (discrete time or continuous time) is to measure its transfer function in frequency domain. Such measurements require precision sine-wave generator and narrow bandwidth signal or spectrum analyzer or calculation of the FFT coefficients. Usually a precision sine-wave generator is not available on-chip and the measurement need a long time because the transfer function must be measured at several different frequencies and amplitudes. Theoretically, the time needed for the measurements could be reduced measuring the response $h(n)$ to the unit delta pulse $\delta(t)$. In this way, all information of the LTI system would be obtained in one measurement. Unfortunately, the method is difficult to use because it requires huge dynamic range of a system or the response is covered by the noise and it is therefore not practical.

Applying Pseudo Random Noise signal (PRN) with appropriate amplitude and approx. white spectrum and Gaussian probability density function (PDF) to the input of an LTI system (Couch, 1993 and Pan, 1997) provides the opportunity to measure the response $h(n)$ of the LTI system. If mixed-signal LTI is running in parallel to the reference (digital) LTI as suggested on Figure 1 the difference between two LTI systems could be measured efficiently. Input signals are the same for both systems (with possibly slightly different gain) and have noise-like properties. They are shaped by two, generally different deterministic transfer functions $h_1(n)$ and $h_2(n)$. Both responses are exactly the same if transfer functions are the same. The first response $y_1[n]$ corresponds to analog or mixed-signal discrete or continuous time system while $y_2[n]$ corresponds to “exact” digital discrete time system. Both have exactly the same architecture and equal coefficients. The later is always nominal because it is implemented in digital hardware or software with sufficient word-lengths that the quantization noise of the calculations could be neglected.

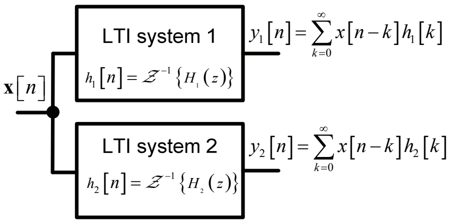

Figure 1: LTI systems driven by signal x[n].

Analog discrete time system described by $h_1[n]$ may deviates from nominal because of catastrophic faults (short or open circuits, etc.) and/or parametric faults: spread of parameters owing to process parameter changes, matching and temperature variations, ageing. Cross-correlation between input signal and response of each system is proportional to the impulse-responses $h_x[m]$ according to (1) and (2) if input signal $x[n]$ has noise-like properties (5). We used discrete time convolution (3) to calculate response of discrete time LTI system to an arbitrary input, (4) to calculate cross-correlation and (5) to

$$\Phi_{xy_1}[m]=\sigma_x^2 h_1[m]+\mu_x^2\sum_{k=0}^{\infty}h_1[k] \tag{1}$$

$$\Phi_{xy_2}[m]=\sigma_x^2 h_2[m]+\mu_x^2\sum_{k=0}^{\infty}h_2[k] \tag{2}$$

calculate auto-correlation of time-shifted white noise signal. *E{.}* is the expectation operator**.**

$$y[n]=\sum_{k=0}^{\infty}x[n-k]h[k] \tag{3}$$

$$\Phi_{xy}[m]=E\{x[n]y[n+m]\} \tag{4}$$

$$E\{x[n]x[n+m-k]\}=\mu_x^2+\sigma_x^2\delta[m-k] \tag{5}$$

Cross-correlations given by (1) and (2) are proportional to appropriate impulse response $h_i(n)$ and variance of the noise if the mean value of the noise signal μ_x^2 approaches zero, which happens for sufficiently long pseudo random sequence (Zepernick, 2005).

The difference of cross-correlations of LTI systems is proportional to the difference of impulse responses according to (6) if μ_x^2 is sufficiently small, which happens for long period of PRN.

$$\Delta_{\Phi_{x,y1},\Phi_{x,y2}}[m]=\Phi_{x,y1}[m]-\Phi_{x,y2}[m]\cong \cong\sigma_x^2\{h_1[m]-h_2[m]\} \tag{6}$$

In the ideal case, both responses are the same and the difference is zero (7):

$$\Delta_{\Phi_{x,y1},\Phi_{x,y2}}[m]=0,\ m=1,2,... \tag{7}$$

If we assume that digital system LTI_2 has nominal impulse response $h_2[n]=h[n]$, while analogue system described by LTI_1 has real impulse response that deviates from the nominal by $\varepsilon[m]$ then: $h_1[n]=h[n]+\varepsilon[n]$. The difference of cross-correlations is proportional to the deviation $\varepsilon[m]$ of the responses and the variance σ_x^2 of the noise (8).

$$\Delta_{\Phi_{x,y1},\Phi_{x,y2}}[m]\cong\sigma_x^2\varepsilon[m] \tag{8}$$

Calculation efficiency can be improved by calculating first the difference of both responses and then the cross-correlations between noise source and the difference of the responses. For linear systems the results are the same as before (9):

$$\begin{aligned}\Phi_{x,\Delta}[m]&=\\ &=E\{x[n](y_1[n+m]-y_2[n+m])\}=\\ &\cong\sigma_x^2(h_1[m]-h_2[m])=\sigma_x^2\cdot\varepsilon[m]\end{aligned} \tag{9}$$

It is not possible to calculate the expectation operator from the definition (4), but for ergodic random signals the time-average operation is equal to expectation calculation (Hayes, 1996) (10):

$$\begin{aligned}&E\{x[n]y[n+m]\}=\\ &=\lim_{N\to\infty}\frac{1}{N}\sum_{n=0}^{N-1}x[n]\cdot y[n+m]\end{aligned} \tag{10}$$

For sufficiently large N the estimate of the mean of the cross-correlation $\bar{\Phi}_{x,\Delta}[m]$ (11) is equal to the cross-correlation (12). This result provides the opportunity for efficient calculation of cross-correlation coefficients.

$$\begin{aligned}\bar{\Phi}_{x,\Delta}[m]&=\frac{1}{N}\sum_{n=0}^{n=N-1}x[n]\cdot\Delta[n+m]=\\ &=\frac{1}{N}\sum_{n=0}^{n=N-1}x[n]\cdot\sum_{k=0}^{\infty}x[n+m-k]\cdot\varepsilon[k]\\ &=\sum_{k=0}^{\infty}\varepsilon[k]\cdot\frac{1}{N}\sum_{n=0}^{n=N-1}x[n]\cdot x[n+m-k]\end{aligned} \tag{11}$$

$$E\{\bar{\Phi}_{x,\Delta}[m]\}=\Phi_{x,\Delta}[m] \tag{12}$$

2.2 PRN as a Signal Source

All results in previous subsection are based on the assumption that the input signal has real “white noise” properties. In reality, such signal does not exist and even approximation is hard to implement. Good approximation, which is easy to built and requires small silicon area is pseudo random noise signal (PRN) that under certain conditions possesses appropriate characteristics (Zepernick, 2005):

- The spectrum is discrete with approximate “white” power spectrum density (PSD). The spectral components exist at frequencies $f_i = f_{clk} / \left(2^N - 1\right)$ for $i = 0...\left(2^N - 1\right)$; the sequence is periodic and repeated every $\left(2^N - 1\right)$ clock cycles,
- The period is sufficiently long so, that μ_x^2 approaches zero. Inside period the signal appears random,
- All states have approx. equal probability, while state 0 is not allowed,
- The PRN can be single or multi bit, dependent on the application but the linearity and accuracy requirements must be maintained,
- The sequence with appropriate autocorrelation properties $R_{x,x+\tau} = 1$ for $\tau = 0$, otherwise $R_{x,x+\tau} = 0$ must be used,
- The PRN source must be simple with small silicon area required for the implementation.

Several useful implementations of PRN exist (Zepernick, 2005). Converting binary pseudo random signal into analogue voltage is accomplished using 1 bit D/A converter that is inherently linear, very accurate and very simple for the implementation. Eventual inaccuracy of the gain coefficient can be corrected during production calibration phase.

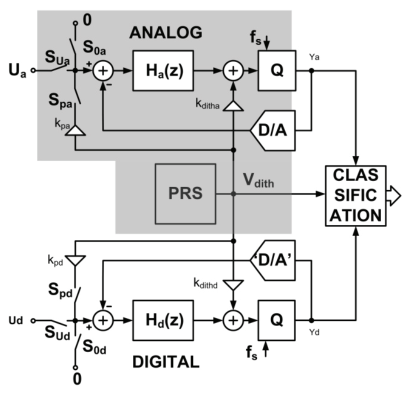

Figure 2: PRN measurements of a modulator.

3 PRN MEASUREMENTS

Measurement of high precision and high resolution A/D converter requires a precision measurement system. For fail-safe electronic system the status of the system and also the status of the ADC must be checked in parallel to the normal operation; any deviation from optimum behaviour must be noticed immediately, therefore important parameters must be measured constantly without the presence of high precision measurement system. One possibility is shown on Figure 2, where basic idea for efficient measurements of a Σ-Δ modulator is presented. Analogue modulator in grey box is implemented on silicon together with PRN generator (PRS source) that fulfils conditions defined in subsection 2.2. We can assume that after calibration the constants k_{ditha} and k_{pa} are exact. The PRN signal is used in any case as a dither signal to prevent limit cycles (Reefman, 2005) and to linearize the quantizer (Widrow, 2008), so it is already present in the circuit. If a multi-bit internal D/A converter is needed, than appropriate linearization or dynamic element matching technique must be used (Jiang, 2007). Dither signals are always connected to the modulator’s dither inputs, usually in front of the internal quantizer. The second modulator, outside the gray area is a reference digital modulator that can be implemented on the chip or off the chip in digital hardware or software algorithm; it should have exactly the same architecture and exactly the same coefficients as the analogue modulator that we want to measure.

Two measurements are generally needed. At first, both inputs are connected to zero (U_a=0, U_d=0), while PRS is connected to dither inputs through k_{dith} and therefore, the noise transfer function or the difference of NTFs is measured. If PRN signal is connected to the inputs through coefficient k_{px} and at the same time to dither inputs through k_{dithx}, the signal transfer function or the difference of STFs is measured. The digital modulator has exactly the same structure as the analogue modulator with equivalent coefficients, but it is built with digital hardware or software with sufficient word-lengths to render negligible any quantization noise owing to fixed-point arithmetic. The digital modulator together with classification block can be implemented outside the chip as a hardware or software module or in case of RTBIST it can be included on-chip together with analogue modulator. It is estimated, that in modern 90nm CMOS technology the area of digital modulator is only one half of the area of the analogue modulator. In both cases, the digital modulator is running in parallel to the analogue modulator using the same or synchronized PRN signal. Both bit-streams and PRN signals are monitored with classification block that calculates cross-correlation coefficients and decides if analogue modulator fits the requirements despite the changes caused by process parameters, matching effects, temperature drift, ageing etc. The behaviour of digital modulator is assumed to be stable, while analogue modulator is subject to changes and this changes we want to measure.

ΣΔ modulators are non-linear systems so the theoretical background described in subsection 2.1 could be used only if the module is linearized. This is achieved by adding dither signal to the input of the modulator's quantizer; in this way eventual limit-cycles are de-correlated (Reefman, 1997) and the operation is linearized (Widrow, 2008). In that case the modulator can be approximated as linear system in z domain (Hamoui, 2004) according to (13), (14), (15) and (16):

$$Y_a(z) = STF_a(z)\left[U_a(z) + R(z)\right] + NTF_a(z)\left[Q_a(z) + P(z)\right] \qquad (13)$$

$$Y_d(z) = STF_d(z)U_d(z) + NTF_d(z)\left[Q_d(z) + P(z)\right] \qquad (14)$$

The relations between $H(z)$, $STF(z)$ and $NTF(z)$ are given in (15) and (16) for both modulators, where index x=a stands for analogue and x=d for digital modulator. $R(z)$ represents input-referred circuit noise of the analogue modulator and $P(z)$ represents pseudo-random noise used for dither.

$$STF_x(z) = \frac{H_x(z)}{1 + H_x(z)} = 1 - NTF_x(z) \qquad (15)$$

$$NTF_x(z) = \frac{1}{1 + H_x(z)} \qquad (16)$$

It is assumed that digital and analogue quantization noises are not equal $Q_a(z) \neq Q_d(z)$ because internal states might be different even if applied input signals are exactly the same. We need to do two tests to be able to measure both transfer functions (STF and NTF). For the first test the input signals are set to zero ($u_a[n] = 0$, $u_d[n] = 0$), the PRN is applied to both dither inputs so $NTF_a(z)$ and $NTF_d(z)$ or their difference can be determined. The cross correlation can be determined using (17), (18) and (5). We assumed, that $p[n]$ is not correlated to any other noise source: $r[k]$, $q_a[k]$ and $q_d[k]$.

$$\Phi_{p,y_{a1}}[m] = \sigma_{xp}^2 \cdot ntf_a[m] + \mu_{xp}^2 \sum_{k=0}^{\infty} ntf_a[k] \qquad (17)$$

$$\Phi_{p,y_{d1}}[m] = \sigma_{xp}^2 \cdot ntf_d[m] + \mu_{xp}^2 \sum_{k=0}^{\infty} ntf_d[k] \qquad (18)$$

If sufficiently long PRS sequence is used, the mean value of PRS signal is approaching zero $\mu_{xp}^2 = 0$ and cross correlation become proportional to corresponding impulse response of the noise-transfer function. The difference of both cross-correlations is proportional to deviation of analogue noise-transfer function from digital noise-transfer function according to (19) and (20):

$$\Delta_{\Phi_{p,ya1},\Phi_{p,yd1}}[m] \doteq \sigma_{xp}^2 \varepsilon_{ntf}[m] \qquad (19)$$

$$\varepsilon_{ntf}[m] = ntf_a[m] - ntf_d[m] \qquad (20)$$

For test 2, the PRN signal is connected to both inputs in addition to both dither inputs, so cross-correlations are ((21), (22)).

$$\Phi_{p,ya2}[m] = \sigma_{xp}^2 \left\{stf_a[m] + ntf_a[m]\right\} + \ldots \qquad (21)$$

$$\Phi_{p,yd2}[m] = \sigma_{xp}^2 \left\{stf_d[m] + ntf_d[m]\right\} + \ldots \qquad (22)$$

Moreover, the difference is (23):

$$\Delta_{\Phi_{p,ya2},\Phi_{p,yd2}}[m] \doteq \sigma_{xp}^2 \left\{\varepsilon_{stf}[m] + \varepsilon_{ntf}[m]\right\} \qquad (23)$$

From (19) and (23) one can estimate deviation of signal and noise transfer functions of analog modulator from the reference modulator and since we know $\varepsilon_{ntf}[m]$ it is easy to calculate $\varepsilon_{stf}[m]$.

4 RTBIST

Real time monitoring of some performance parameters of the mixed-signal circuit in parallel to the normal operation comes from the fail-safe system requirements and the name RTBIST (**R**eal-**T**ime-**B**uilt-**I**n-**S**elf-**T**est) reflects that. We could also name it real-time-self-aware (RTSA) system taking into considerations that higher level functions are implemented in dedicated hardware or software. Here, we are dealing only with methodology, which is based on ideas presented in sections 2 and 3 with the following differences:

- As explained in section 3 both modulators are implemented on chip,
- The LTI_1 (analogue modulator) process the input analogue signal and in parallel the small amount of PRN, while digital modulator with the same

architecture and equal coefficients, process only the PRN noise.

- During production the cross correlation coefficients of the PRN and the difference of the responses for NTFs and STFs at nominal conditions are calculated and stored.
- During real measurements, the NTF is not measured explicitly because it can be calculated from the relation between STF and NTF given in (15): $NTF_x(z) = 1 - STF_x(z)$.
- For S-C implementation we know that the gain factor for the PRN connected to the input is very accurate (0.1%) and has a very low temperature coefficient.
- The correlation between PRN and input signal is assumed negligible.
- The amplitude of PRN signal is small so that only a very small fraction of a mixed-signal circuit dynamic range is consumed by the test source PRN.

5 CLASSIFICATION

A classification circuit can be implemented according to Figure 3, which closely follows equation (11). The average is replaced by moving average or the first order filter.

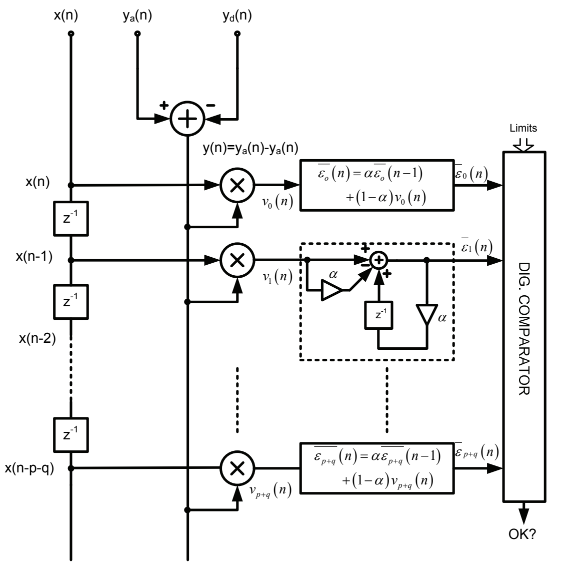

Figure 3: Possible classification circuit.

The signal PRN is delayed instead of $y[n]$ to simplify the hardware: this is possible for ergodic signals; PRS signal is simply delayed using $p+q$ bit shift register because only $p+q+1$ samples of unit step response are needed for LTI systems having p poles and q zeroes, so that the transfer function is correctly represented (Hayes, 1996). Multiplication is performed by simple exchange of the signs as dictated by the PRN. In addition, for one-bit modulators the bit-streams $y_a(n)$ and $y_d(n)$ are also 1 bit, so the result of the subtraction is within $y(n) = y_a(n) - y_b(n) \in \{-2, 0, 2\}$; the multiplication circuitry is very simple. Each product $v_i(n)$ is then LP- filtered. It requires a small portion of FPGA and/or little silicon area. If needed, a higher order digital-averaging filter could be built. The digital comparator then decides if all results are within the limits L_i:

$$\varepsilon_i(n) \le L_i; \quad n = 1, 2, ... \tag{24}$$

How many samples do we need for reliable classification? Is the number of correlation coefficients sufficient? What time do we need for the measurement? What is the probability of classifying correctly? The answers to these questions are not simple and are still under consideration. Available time and hardware resources are limited. In addition, the accuracy of a decision and its speed conflict with each other. For our self-aware and/or fail-safe system we need the information about the system behaviour as fast as possible, thus, appropriate selection of conflicting parameters (time, accuracy of classification, hardware resources available, etc.) must be optimized for particular application. For example, a decision that is more accurate needs more time or a more elaborate averaging process and higher order filters. Fast decision usually leads to poor accuracy of classification, which could be improved by appropriate higher order filtering which in turn needs more hardware resources.

6 EXAMPLES

A fifth-order, single-loop, discrete-time Σ-Δ modulator implemented in S-C technique with one-bit internal quantizer and feed-forward structure to reduce power consumption is used as an example. Over-sampling frequency is 32MHz and the band of interest is from 20kHz up to 400kHz. The problem of efficient testing of such modulator is that we cannot measure internal signals because we would destroy the operation (internal S-C stages have capacitances from 0.5pF down to 50fF) and the power consumption is restricted. To verify presented methodology we have built Simulink models of the modulators and classification circuit. For LTI_1 the most important analog performances of opamps (kT/C noise, thermal noise A0, GBW, offset, slew-rate, non-linearity) and quantizer (offset, hysteresis, noise, latency) are modeled. Capacitor ratios can be perturbed according to the technology and size of unit capacitor. The digital modulator and classification circuit are with bit-true models that calculate in real time eleven cross-correlation coefficients $\varepsilon[0]$ through $\varepsilon[10]$ using first order moving average filter with pole: $\alpha = 1 - 2^{-14}$. Moving average filter needs approx. 2ms for the transient; the results after that time could be used for the comparison. We have run several simulations, trying to imitate different problems of analog modulator related to the production spread as well as to temperature drift and other possible problems. A summary of simulation results is presented on Figure 4 for noise transfer function and on Figure 5 for the signal transfer function. The following experiments were simulated: (a) nominal circuit with no kT/C noise, no op-amp noise, and ideal op-amp characteristics, (b) real op-amp characteristics inside allowed ranges, (c) allowed kT/C and op-amp noise with other conditions as before, (d) the same as (c) with kT/C 10 times bigger (out of specs), (e) the same as (c) but slew-rate of first op-amp 2 times lower than the min allowed, (f) the same as (c) but one capacitor changed by 30%, (g) limit of the 1st amplifier reduced to 0.4V from 0.5V, (h) Monte-Carlo run with capacitor ratio changes $3\sigma = 5\%$ proportional to unit cap size (the spread is intentionally exaggerated to get some out-of-range results). On both figures, cross-correlation coefficients are plotted for different experiments marked with triangles. We can see that in both cases some of the results are out of the limits marked with dots inside the squares. The limits were defined according to the specifications using Matlab simulations.

7 CONCLUSIONS

In this article a possibility for real-time measurements of high resolution mixed-signal circuits have been investigated. The basic idea and the theory behind it have been explored for simple LTI system as well as for more complex mixed-signal module (Σ-Δ modulator). A block diagram of efficient measurements has been presented together with possible implementation of classification circuitry. The aim of presented methodology is to pave the way for real time built-in self test (RTBIST) of such embedded modules and to real time self aware (RTSA) methodology measurements.

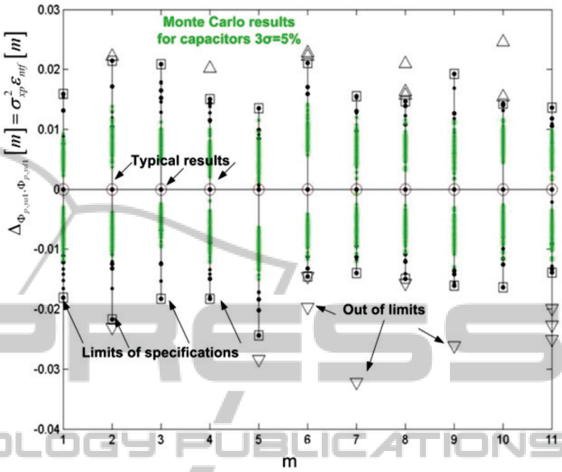

Figure 4: Classification result for NTF.

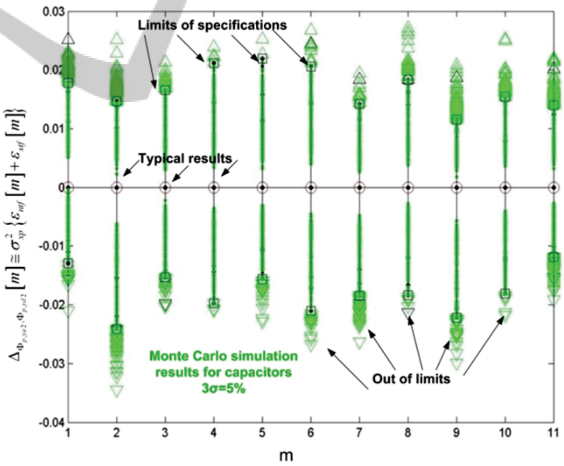

Figure 5: Classification results for STF.

The investigation is far from finished. We believe that only basic steps were analyzed in this work. Many other problems still need to be investigated, like for example: optimization of classification algorithm according to required speed and accuracy, influence of properties and length of PRN sequence to the accuracy and speed of measurements, influence of nonlinearities.

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