Spectral Prediction for Specification-Based Loopback Test of Embedded Mixed-Signal Circuits

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***Abstract*— Loopback testing of Mixed-signal SOCs provides a low-cost test solution, but suffers from fault masking, resulting in serious yield loss and low test accuracy. This paper presents an efficient loopback test methodology which enables individual characterization of dynamic performance of Devices Under Test (DUTs) in loopback mode. DUTs are loop-backed externally on a loadboard (DUT board), and a simple filter and an analog adder on the loadboard produce a composite loopback response. Characteristic parameters are extracted from these loopback responses, and a non-linear regression technique based on spectral predictors is used to predict various performance parameters such as Gain, SNR, THD and SINAD. The spectral predictor provides more accurate and reliable prediction compared to a time-domain approach. Both simulation and hardware measurements are presented to validate the proposed technique.**

I. Introduction

The continuing demand for data and telecommunication applications is driving the need for integrating many analog and mixed-signal functions into a single system-on-chip (SOC). While SOCs offer significant benefits by integrating several different system functions on a single die, they pose significant challenges for testing. A conventional module-level test may not be attractive for highly integrated mixed-signal SOCs, due to limited test resources of ATEs and limited I/O accessibility to each individual module [1].

Several methods have been proposed to overcome this hurdle of testing SOCs. Notable among them, are the Loopback-based Built-In-Self-Test (BIST) schemes [2] [3] [4]. In the loopback configuration, a Device Under Test (DUT) is placed in a mode that loops the output of one signal path back into the input of another signal path. Instead of running sequential tests for individual analog or mixed-signal blocks, the loopback response is measured, and compared against the nominal value with tolerance bands to make a pass/fail decision. Therefore, by removing conventional sequential tests for individual blocks, a significant reduction in test time can be achieved.

This method, however, suffers from fault masking caused by the uncorrelated interaction between non-functionally related components in loopback mode. The combination of seriously degraded components in one of the functional paths and overqualified components in another functional path, may result in misinterpretation of the loopback response. Therefore the loopback response does not directly represent the functional specifications of individual blocks or paths under test. The pass/fail decision based on the pure loopback test may lead to serious yield loss and low test accuracy since it is not made in the performance parameter domain where the specifications are clearly defined. At the same time, it may not be possible to perform the compensation tests for marginal DUTs [5] due to the difficulties of directly evaluating functional specifications. In addition, its utilization in diagnosis may be limited to locating an error to the loop rather than to a functional block. Hence, the extraction of specifications of each functional path has become a crucial issue in loopback test.

While analog and mixed-signal BIST techniques using loopback are well researched, there is not much in the literature on the fault masking issue. Yu et al. [6] proposed a loopback technique which characterizes the noise of ADC and DAC in loopback mode, by monitoring internal nodes of delta-sigma modulators. The usefulness of this approach, however, is limited to certain types of mixed-signal circuits such as a Delta-Sigma converter, and it only provides the characteristics related to noise.

To overcome such limitations and improve the accuracy, we propose an *efficient loopback test technique using a simple analog filter and spectral predictors*. A two-tone test input is applied to a DUT in loopback mode, and an off-chip analog filter and analog adder produce a composite loopback response which can be decomposed into loopback responses of different weights. *Characteristic parameters* are obtained from their spectral representations, and are used as predictors for a mapping function derived using a non-linear regression technique. We assume that the filter specification can be readily identified, since it is implemented on the loadboard.

The organization of the rest of this paper is as follows. The issues involved in conventional loopback test are introduced in Section II. The basic concepts of our method are discussed in Section III. In Section IV, the simulation and hardware measurements to illustrate the effectiveness of our method are presented, and we conclude this paper in Section V.

II. Issues in Loopback Test

In order to better understand the fault masking issue in loopback tests, assume that a unit magnitude signal $\cos(\omega t)$ is applied to a non-linear mixed-signal circuit such as the DAC channel shown in Fig. 1. we use t notation in digital domain

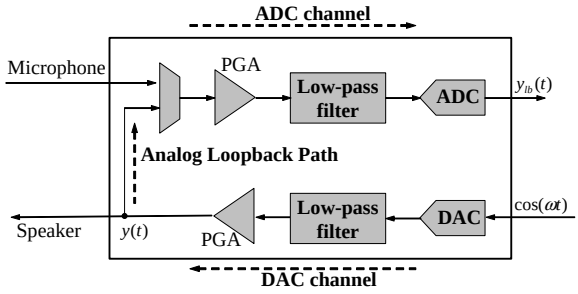

Fig. 1. Conventional Analog Loopback Path of a Mixed-signal Circuit

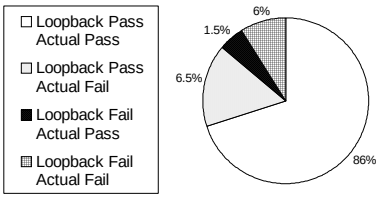

Fig. 2. Test Escape and Yield Loss of Conventional Loopback Test

for simplicity. Harmonic distortion considered is up to the third order. The output response $y(t)$ can be approximated by a Taylor expansion as follows [7].

$$\begin{aligned} y(t) &= \alpha_1 \cos(\omega t) + \alpha_2 \cos^2(\omega t) + \alpha_3 \cos^3(\omega t) \\ &= (\alpha_1 + \frac{3\alpha_3}{4}) \cos(\omega t) + \frac{\alpha_2}{2} \cos(2\omega t) + \frac{\alpha_3}{4} \cos(3\omega t) \end{aligned} \quad (1)$$

where α_1, α_2 and α_3 are the first, second and third Taylor coefficients, and $y(t)$ can be expanded with harmonic coefficients as shown in the second row in Equation 1.

As shown in Fig. 1, suppose this distorted and noisy analog signal $y(t)$ be loopbacked to the ADC channel, then the loopback response $y_{lb}(t)$ can be expressed as

$$\begin{aligned} y_{lb}(t) = (\beta_1\alpha_1) \cos(\omega t) + (\beta_1\alpha_2 + \beta_2\alpha_1^2) \cos^2(\omega t) \\ + (\beta_1\alpha_3 + 2\beta_2\alpha_1\alpha_2 + \beta_3\alpha_1^3) \cos^3(\omega t) \end{aligned} \quad (2)$$

where β_1, β_2 and β_3 are the first, second and third harmonic coefficients of the ADC channel respectively.

The common way to quantify the non-linearity of analog circuits is to identify α_1, α_2, etc., or the sum of the powers of harmonics [7]. However, it can be observed from Equation 2 that the characterization of harmonic coefficients (or Taylor coefficients) of $y_{lb}(t)$ cannot provide the solution for α_i, unless the values of β_i are exactly identified.

In addition to the harmonic distortion, noise is a key parameter in mixed-signal circuits. Assume that $N_\alpha(f)$ is the Power Spectral Density (PSD) of a pure noise input to the ADC channel, which is generated by the DAC channel, and $N_\beta(f)$ is the PSD of the ADC channel. Further assume that the noise of the DAC and the ADC channel are uncorrelated. The output referred noise of the loopback can be expressed as follows.

$$\overline{v_{lb}^2} = \int_0^\infty K N_\alpha(f) df + \int_0^\infty N_\beta(f) df \quad (3)$$

where K is the overall gain of the DAC and ADC channel. As can be seen from Equation 3, $\overline{v_{lb}^2}$ is the sum of noise of the ADC and the DAC channel, and any excessive value of noise in one channel can be masked by an overqualified performance of the other channel.

A. Example of issues

Consider a simple analog loopback configuration shown in Fig. 1. A set of 2200 DUT ensembles was generated by introducing statistical variations into the noise and distortion that affect the SINAD of the ADC and the DAC channel, as well as the SINAD of the loopback response. The mean and standard deviation of the distribution of the obtained SINAD were approximately 66dB and 1.3dB respectively. In order to determine test limits of the loopback response, a set of 200 DUT ensembles was then simulated in loopback and normal mode, and the SINAD of individual ADC and DAC, and loopback response were measured. The only specification parameter considered for the classification was SINAD. The SINAD limit of the ADC/DAC was set to 64dB (2dB tolerance bands). The lower SINAD limit of the loopback response is determined by the minimum loopback SINAD, which passes both the ADC and the DAC test. Table I summarizes the specification limits for the ADC, the DAC and the loopback response.

TABLE I
DAC-ADC SPECIFICATIONS USED FOR CLASSIFICATION

	Specification Limits		
	ADC	**DAC**	**Loopback Response**
SINAD	64dB	64dB	62.6dB

The specification limit of the loopback response shown in Table I was compared against the measured SINAD of the remaining 2000 DUT ensembles for classification. Fig. 2 shows the ratio of the misclassification for the conventional loopback test. 6.5% of DUTs which passed the loopback test turned out to be faulty in either the ADC or the DAC. This is due to the fault masking, which may cause a significant reduction in test quality. In addition, 1.5% of DUTs which passed the individual tests failed the loopback test.

This result demonstrates that the loopback test alone cannot provide thorough production test and diagnosis. Therefore, the extraction of the characteristics of individual channels from the loopback response is required to improve the test quality.

III. Proposed Approach

The previous section discussed issues related to the fault masking problem found in loopback test, and the difficulties in extracting the performance parameters of each signal path from the loopback response. The following sections will discuss how a two-tone test input and an analog filter are used to extract characteristic parameters of a DUT, and how the dynamic performance parameters of individual channels can be predicted from the characteristic parameters using a non-linear regression technique based on spectral predictors.

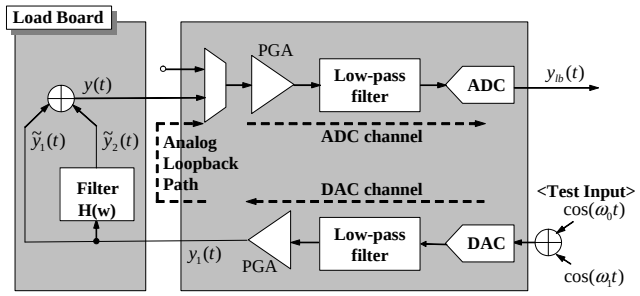

Fig. 3. Proposed Loopback Test Scheme

A. Harmonic Distortion Parameter Extraction

Fig. 3 shows the proposed test architecture for testing the ADC and DAC channels using the loopback scheme. This loopback path is connected externally to the die on a loadboard [8]. A simple low-order filter and an analog adder are implemented on the loadboard, hence they can be readily debugged and characterized. A two-tone cosine wave of frequency ω_0 and ω_1, is applied to the DAC using either an on-chip DSP or an external digital ATE. The response of the Taylor series to a two-tone signal can be determined by expanding and using trigonometric identities. That is,

$$\begin{aligned} y_1(t) = &\sum_k \alpha_k \cos(k\omega_0 t) + \sum_l \alpha_l^{'} \cos(l\omega_1 t) \\ &+ \sum_{m,n} \eta_{m,n} \cos((m\omega_0 \pm n\omega_1)t) + \alpha_0 \end{aligned} \tag{4}$$

where α_k and $\alpha_l^{'}$ are harmonic coefficients of the DAC channel for frequency ω_0 and ω_1, and α_0 is DC value. $\eta_{m,n}$ are Intermodulation Distortion (IMD) coefficients. We assume that ω_0 and ω_1 are as close as possible to each other, such that the harmonic coefficients of $cos(k\omega_0 t)$ and $cos(l\omega_1 t)$ are identical, i.e., $\alpha_k = \alpha_l^{'}$. In addition to the fundamental and harmonic components, the interaction of ω_0 and ω_1 produces IMD. ω_0 and ω_1 are chosen to be mutually prime, thus the IMD components do not overlap with the fundamental and harmonic components, i.e., $m\omega_0 \pm n\omega_1 \neq k\omega_0$ and $m\omega_0 \pm n\omega_1 \neq l\omega_1$ for any integer k, l, m and n [9]. Since the fundamental and distortion components are only of interest (as characteristic parameters), the IMD components will be ignored in deriving the following equations. In addition, α_0 will be neglected, since the filter or PGA in the ADC channel will cancel the DC terms. Therefore, Equation 4 can be re-written as follows.

$$\tilde{y_1}(t) = \sum_k \alpha_k (\cos(k\omega_0 t) + \cos(k\omega_1 t)) \tag{5}$$

The filter in Fig. 3 produces the secondary loopback input $\tilde{y_2}$ from $\tilde{y_1}$, which can be expressed as

$$\begin{aligned} \tilde{y_2}(t) = &\sum_k |H(k\omega_0)| \, \alpha_k \cos(k\omega_0 t + \theta_k) \\ &+ \sum_l |H(l\omega_1)| \, \alpha_l \cos(l\omega_1 t + \phi_l) \end{aligned} \tag{6}$$

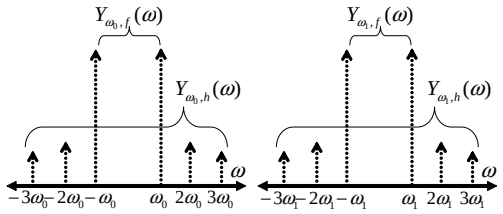

Fig. 4. Fundamental and Harmonic components of $Y(\omega)$

where $H(\omega)$ is the transfer function of the filter which is

$$\begin{aligned} H(\omega)|_{w=k\omega_0} &= \gamma_k e^{j\theta_k} \\ H(\omega)|_{w=k\omega_1} &= \zeta_k e^{j\phi_k} \end{aligned} \tag{7}$$

Since in general the roll-off slope of the filter in the frequency domain is much sharper than the variation of distortion coefficients of the ADC and DAC, we assume that the values of γ_k and ζ_k are different even when ω_0 and ω_1 are close to each other. We further assume that the specifications of the filter are characterized prior to the test, and γ_k, ζ_k, θ_k and ϕ_k are known values.

As shown in Fig. 3, $\tilde{y_2}$ is added to $\tilde{y_1}$, and consequently we obtain the final input to the ADC channel $y(t)$. That is,

$$\begin{aligned} y(t) = &\sum_k (\alpha_k \cos(k\omega_0 t) + \gamma_k \alpha_k \cos(k\omega_0 t + \theta_k)) \\ &+ \sum_l (\alpha_l \cos(l\omega_1 t) + \zeta_l \alpha_l \cos(l\omega_1 t + \phi_l)) \end{aligned} \tag{8}$$

As shown in Fig. 4, the frequency domain representation of $y(t)$, $Y(\omega)$ can be divided into four parts, fundamental and harmonic components of ω_0 and ω_1 as follows.

$$Y(\omega) = Y_{\omega_0,f}(\omega) + Y_{\omega_0,h}(\omega) + Y_{\omega_1,f}(\omega) + Y_{\omega_1,h}(\omega) \tag{9}$$

We use Equation 1 to model the ADC channel with Taylor coefficients β_k, and the loopback response y_{lb} can be expressed as

$$y_{lb}(t) = \beta_1 y(t) + \beta_2 y^2(t) + \beta_3 y^3(t) \tag{10}$$

$y^2(t)$ in Equation 10 can be expressed as a convolution of $Y(\omega)$ in the frequency domain, which is

$$\mathcal{F}(y^2(t)) = Y^{'}(\omega) = Y(\omega) \otimes Y(\omega) \tag{11}$$

Since the convolution of ω_0 and ω_1 produces IMD in which we are not interested, these terms are ignored. Then Equation 11 becomes

$$\begin{aligned} Y^{'}(\omega) = &(Y_{\omega_0,f}(\omega) + Y_{\omega_0,h}(\omega)) \otimes (Y_{\omega_0,f}(\omega) + Y_{\omega_0,h}(\omega)) \\ &+ (Y_{\omega_1,f}(\omega) + Y_{\omega_1,h}(\omega)) \otimes (Y_{\omega_1,f}(\omega) + Y_{\omega_1,h}(\omega)) \end{aligned} \tag{12}$$

and can be expanded to

$$\begin{aligned} Y^{'}(\omega) = &\{Y_{\omega_0,f}(\omega) \otimes Y_{\omega_0,f}(\omega) + 2Y_{\omega_0,f}(\omega) \otimes Y_{\omega_0,h}(\omega) \\ &+ Y_{\omega_0,h}(\omega) \otimes Y_{\omega_0,h}(\omega)\} \\ &+ \{Y_{\omega_1,f}(\omega) \otimes Y_{\omega_1,f}(\omega) + 2Y_{\omega_1,f}(\omega) \otimes Y_{\omega_1,h}(\omega) \\ &+ Y_{\omega_1,h}(\omega) \otimes Y_{\omega_1,h}(\omega)\} \end{aligned} \tag{13}$$

The convolution of harmonic components $Y_h(\omega) \otimes Y_h(\omega)$ may be negligible. For example, the multiplication of the second and third harmonics will produce trivial magnitudes. While they can be included for better accuracy, it will be ignored for computational simplicity. Then, Equation 13 becomes

$$\begin{aligned} Y^{'}(\omega) = & Y_{\omega_0,f}(\omega) \otimes Y_{\omega_0,f}(\omega) + 2Y_{\omega_0,f}(\omega) \otimes Y_{\omega_0,h}(\omega) \\ & + Y_{\omega_1,f}(\omega) \otimes Y_{\omega_1,f}(\omega) + 2Y_{\omega_1,f}(\omega) \otimes Y_{\omega_1,h}(\omega) \end{aligned} \tag{14}$$

Similarly, the frequency response $Y^{''}(\omega)$ of $y^3(t)$ can be obtained, and the frequency response $Y_{lb}(\omega)$ of $y_{lb}(t)$ is

$$Y_{lb}(\omega) = \beta_1 Y(\omega) + \beta_2 Y^{'}(\omega) + \beta_3 Y^{''}(\omega) \tag{15}$$

Consequently, the frequency response of the fundamental and harmonic components of $y_{lb}(t)$ can be expressed as

$$\begin{aligned} Y_{lb}(\omega_0) &= \beta_1\hat{\alpha_1} + \beta_2/2(\hat{\alpha_1}\hat{\alpha_2}) \\ &\quad + \beta_3/8(3\hat{\alpha_1}^3 + 2\hat{\alpha_1}^2\hat{\alpha_3} + 2\hat{\alpha_1}^2\hat{\alpha_2}) \\ Y_{lb}(2\omega_0) &= \beta_1\hat{\alpha_2} + \beta_2/4(\hat{\alpha_1}^2 + 2\hat{\alpha_1}\hat{\alpha_3}) \\ &\quad + \beta_3/4(3\hat{\alpha_1}^2\hat{\alpha_2} + \hat{\alpha_1}\hat{\alpha_2}\hat{\alpha_3}) \\ Y_{lb}(3\omega_0) &= \beta_1\hat{\alpha_3} + \beta_2/2(\hat{\alpha_1}\hat{\alpha_2}) \\ &\quad + \beta_3/8(\hat{\alpha_1}^3 + 4\hat{\alpha_1}^2\hat{\alpha_3} + 2\hat{\alpha_1}\hat{\alpha_2}^2) \\ Y_{lb}(\omega_1) &= \beta_1\check{\alpha_1} + \beta_2/2(\check{\alpha_1}\check{\alpha_2}) \\ &\quad + \beta_3/8(3\check{\alpha_1}^3 + 2\check{\alpha_1}^2\check{\alpha_3} + 2\check{\alpha_1}^2\check{\alpha_2}) \\ Y_{lb}(2\omega_1) &= \beta_1\check{\alpha_2} + \beta_2/4(\check{\alpha_1}^2 + 2\check{\alpha_1}\check{\alpha_3}) \\ &\quad + \beta_3/4(3\check{\alpha_1}^2\check{\alpha_2} + \check{\alpha_1}\check{\alpha_2}\check{\alpha_3}) \\ Y_{lb}(3\omega_1) &= \beta_1\check{\alpha_3} + \beta_2/2(\check{\alpha_1}\check{\alpha_2}) \\ &\quad + \beta_3/8(\check{\alpha_1}^3 + 4\check{\alpha_1}^2\check{\alpha_3} + 2\check{\alpha_1}\check{\alpha_2}^2) \end{aligned} \tag{16}$$

where $\hat{\alpha_k}$ is $\alpha_k + \gamma_k\alpha_k e^{j\theta_k}$, and $\check{\alpha_k}$ is $\alpha_k + \zeta_k\alpha_k e^{j\phi_k}$.

In summary, $Y_{lb}(k\omega_0)$ and $Y_{lb}(k\omega_1)$ (characteristic parameters) can be calculated by performing Fast Fourier Transform (FFT) on the measured loopback response. Therefore, given γ_k, ζ_k, θ_k and ϕ_k, we can estimate α_k and β_k from Equation 16 and the harmonic distortion of the ADC and DAC channel can be quantified separately.

B. Noise Parameter Extraction

In our proposed scheme shown in Fig. 3, the output referred noise of y_{lb} in Equation 3 can be re-written as follows.

$$\begin{aligned} \overline{v_{lb}^2} = & \int_0^\infty KN_\alpha(f)df + \int_0^\infty N_\beta(f)df \\ & + \int_0^\infty KN_\alpha(f)\left|H(f)\right|^2 df + \int_0^\infty KN_H(f)df \end{aligned} \tag{17}$$

where $N_H(f)$ is the PSD of the filter, and we use the f notation rather than ω for simplicity. The evaluation of the integral in Equation 17 is often difficult in a real implementation. We use the *equivalent noise bandwidth* [10]. The equivalent noise bandwidth f_N is chosen to give the same total output noise voltage as the original block when the same input noise voltage is applied. As shown in Fig. 5, the frequency response of the filter $|H(f)|^2$ can be approximated as a simple gain-frequency characteristic with abrupt band edges, $|H^{'}(f)|^2$, where the magnitude is the same as the low frequency gain of $|H(f)|^2$. Thus, Equation 17 becomes

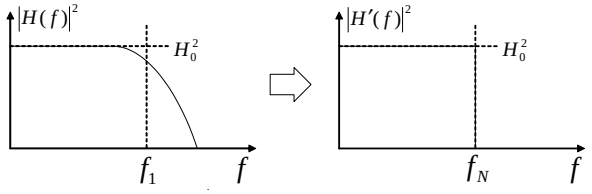

Fig. 5. Equivalent Noise Bandwidth (f_1 is 3dB cutoff frequency)

$$\begin{aligned} \overline{v_{lb}^2} = & \int_0^{f_N} KN_\alpha(f)\left|H^{'}(f)\right|^2 df \\ & + \int_0^\infty (KN_\alpha(f) + N_\beta(f) + KN_H(f))df \end{aligned} \tag{18}$$

$\overline{v_{lb}^2}$ in Equation 18 can be divided into two components with respect to the frequency ranges. $\overline{v_a^2}$ is total noise of y_{lb} in the frequency range [0 f_N] and $\overline{v_b^2}$ is the total noise in the frequency range [f_N ∞].

$$\begin{aligned} \overline{v_a^2} &= K(1+H_0^2)\int_0^{f_N} N_\alpha(f)df + \int_0^{f_N}(N_\beta(f) + KN_H(f))df \\ \overline{v_b^2} &= K\int_{f_N}^\infty N_\alpha(f)df + \int_{f_N}^\infty (N_\beta(f) + KN_H(f))df \end{aligned} \tag{19}$$

$\overline{v_a^2}$ and $\overline{v_b^2}$ are readily computed from the measured loopback response, and given K, $H(f)$, and $N_H(f)$, $\int_0^{f_N} N_\alpha(f)df$, $\int_0^{f_N} N_\beta(f)df$, $\int_{f_N}^\infty N_\alpha(f)df$ and $\int_{f_N}^\infty N_\beta(f)df$ can be calculated. Consequently, the noise parameters of the ADC and DAC channel can be determined from Equation 19.

C. Non-linear Regression using a Spectral Predictor

In a real implementation, it may not be feasible to solve non-linear equations analytically. The problem, in fact, becomes more acute if higher order harmonic distortion is considered such as $Y_h(\omega) \otimes Y_h(\omega)$, 4th harmonic, 5th harmonic, etc. In addition, uncertainty of analog signals may cause an error in measurement, and the error can propagated into the evaluation of the equations. Therefore, we use a non-linear regression technique based on spectral predictors to solve the derived non-linear equations.

Multivariate Adaptive Regression Splines (MARS) has been developed to solve this problem, with the purpose of predicting the values of the response variables from a set of predictor variables [11]. This technique has become particularly popular in the area of alternate test [12] for analog and RF circuits, and it is used to find non-linear mapping functions between signatures and specifications of a DUT [13]. This approach has been mainly based on the transient response to a transient stimulus, however it may suffer from certain limitations which are summarized as follows.

1) This approach requires measurements to be either synchronized or precisely triggered in order to provide

TABLE II

MEAN AND STANDARD DEVIATION OF PREDICTION ERRORS OF SIMULATIONS (SNR(dB), THD(dB), SINAD(dB), GAIN(%))

Parameter	DAC channel		ADC channel	
	Mean	STD.	Mean	STD.
SNR	0.54dB	0.41dB	0.32dB	0.28dB
THD	0.88dB	0.78dB	0.87dB	0.73dB
SINAD	0.50dB	0.48dB	0.50dB	0.42dB
GAIN	3.3%	2.9%	3.0%	2.5%

the stationary sampling points while constructing the regression equation and predicting specifications.

2) The transient-response approach generally requires large predictor variables to increase the efficiency of prediction, resulting in extensive computational overhead.

In order to solve these problems, the frequency response is used as the predictor variable through spectral sampling. Spectral powers at frequency bins which are used in Equations 16 and 19, are used as predictor variables for MARS. Therefore, the $Y_h(\omega) \otimes Y_h(\omega)$ which we ignored in Equation 13, are included. The advantages of using this technique are as follows.

1) Synchronization issues related to our method can be resolved using a weighting function such as a windowing which is a popular way to solve stationary problems resulting from the undesirable effects of FFT [9].
2) This method requires smaller predictor dimensions than the transient response-based approach.

This spectral approach requires an additional FFT process to obtain input predictors, resulting in computational overhead. However, this overhead can be compensated by using smaller predictor dimensions.

D. Alternate Test Scheme

In addition to the two-tone test scheme described thus far, it is also possible to apply two separate single tone signals. Two responses with the different weights, $Y_{\omega_0}(\omega)$ and $Y_{\omega_1}(\omega)$ are generated separately. The two loopback responses are analyzed in a similar way, and the characteristic parameters can be derived. This test scheme may be more effective than the two-tone scheme when the harmonics are relatively small compared to noise; therefore their identification is difficult. This may occur in the two-tone test scheme since the power of harmonics tends to spread out to IMD bins, which we ignored for analysis.

IV. EXPERIMENTAL RESULTS

In this section, we present both simulation and hardware measurements to validate the proposed technique. A MATLAB simulation using a two-tone input will first be presented, and hardware measurements using two single tone inputs performed on a commercial broadband modem (AD9865) will be presented.

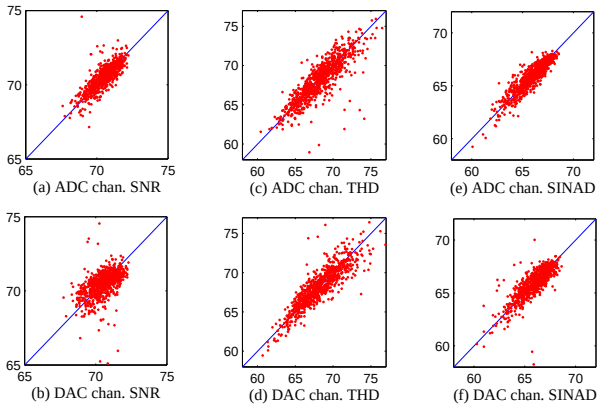

Fig. 6. Comparison of actual and predicted values (simulation data)

A. Simulation Results

The block diagram shown in Fig. 3 is a typical transmit and receive channel of a mixed-signal circuit. In test mode, the DAC channel output was looped back to the ADC channel through the analog bandpass filter which was modeled as a second-order filter. A two-tone sine-wave whose harmonics and intermodulation harmonics do not overlap with each other was applied to the DUT. A set of 1000 DUT ensembles was generated by introducing statistical variations with normal distribution in values of gain, random noise and harmonic distortion. Combinations of these values were also generated randomly. 200 ensembles (training sets) were simulated in normal and test mode, and the performance parameters of the DAC and ADC channel, and spectral predictors which were used to derive the mapping functions were then obtained. The mapping functions were used to predict the performance parameters of the remaining DUT ensembles (validation sets). The performance parameters considered in this paper were the gain, SNR, THD and SINAD.

Fig. 6 shows plots of the predicted versus the actual performance parameters for the proposed method. Table II summarizes the mean and variance of prediction errors. The result indicates that prediction errors were less than 1.7dB, and the prediction errors of the gain were within 6%.

B. Hardware Measurements

Hardware measurements were performed on a commercial broadband modem mixed-signal front-end IC (Analog Devices AD9865) [14], and its block diagram is shown in Fig. 7. The AD9865 is a mixed-signal front end IC for transceiver application requiring Tx and Rx path functionality with data rates up to 80MSPS. The Tx signal path consists of a bypassable 2x/4x low-pass interpolation filter, a 10-bit TxDAC, and driver. The gain of the Tx signal path can be configured by the TxDAC and the driver. The receive path is composed of a tunable 3-pole low-pass filter, a programmable gain amplifier, and a 10-bit ADC. The lowpass filter cutoff frequency can be set over a 15MHz to 35MHz range or simply bypassed. In our experiment, the filter was bypassed in normal functional

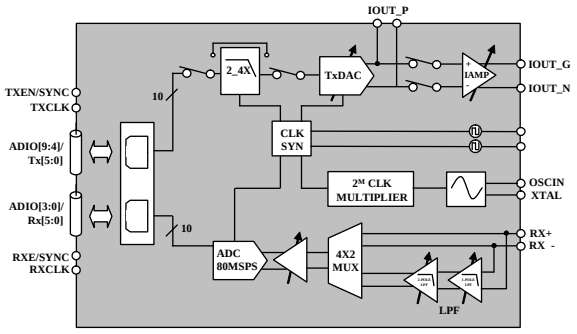

Fig. 7. AD9865 Block Diagram

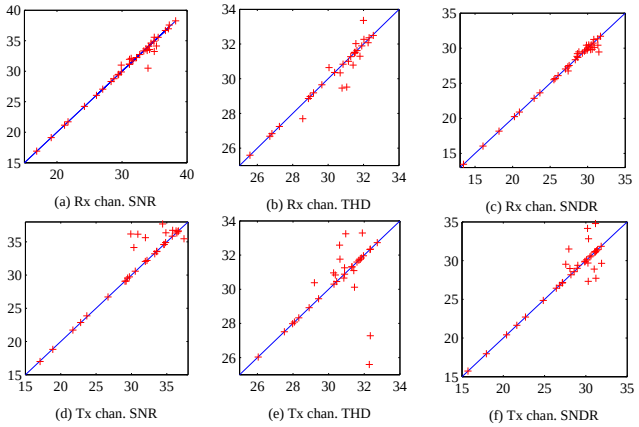

Fig. 8. Comparison of actual and predicted values (hardware measurements)

mode, and turned on in a test mode, generating a weighted response to emulate the on-board filter in Fig. 3. Unlike the simulation setup in the previous section, $Y_{\omega_0}(\omega)$ and $Y_{\omega_1}(\omega)$ in Equation 9 were generated separately. In order to inject a fault in the AD9865, we performed measurements under various stress conditions by sweeping the power supplies, the input amplitude and re-configuring the Rx/Tx gain, and by combining them. This resulted in 7dB of variations in THD, and 20dB in SNR. 48 DUTs were used for the training set, and 40 DUTs were used for the validation. Fig. 8 shows the predicted value of the performance parameters for 40 validation DUTs, and Table III summarizes the statistics of the prediction errors. As shown in Table III, the prediction errors were within 2.4dB except for the SNR of the Tx channel. We observed that one DUT produced more than +10dB error in predicting the SNR of the Tx channel. Also, it shows that the prediction error is increased by approximately 1dB compared to the simulation results. This is primarily due to the measurement error, which we observed to be approximately $0.6dB$.

V. Conclusions

This paper presented an efficient loopback test technique using spectral predictors and a simple analog filter on a DUT interface board. It enables the characterization of performance parameters of individual channels in loopback mode. Two test schemes were introduced, and each scheme was validated through either the hardware measurements or simulations. These experimental results show that the dynamic performance parameters of each channel were characterized accurately with a prediction error of 2.4dB. The proposed test scheme can be readily implemented on a conventional loopback scheme with an external connection and a low-cost analog filter. It does not depend on the type of circuits, and thus can be applied to general mixed-signal circuits.

TABLE III

Mean and Standard deviation of prediction errors for Hardware Measurements (SNR(dB), THD(dB), SINAD(dB))

Parameter	Tx channel		Rx channel	
	Mean	STD.	Mean	STD.
SNR	0.32dB	0.67dB	1.35dB	2.43dB
THD	0.31dB	0.57dB	0.63dB	1.40dB
SINAD	0.25dB	0.41dB	0.94dB	1.57dB

VI. Acknowledgements

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References

[1] S. Ozev and A. Orailoglu, "System-Level Test Synthesis for Mixed-Signal designs," *IEEE Trans. on Circuits and Systems II*, vol. 48, no. 6, pp. 588–599, June 2001.

[2] S. K. Sunter and N. Nagi, "A Simplified Polynomial-Fitting Algorithm for DAC and ADC BIST," in *Proc. International Test Conference*. IEEE, November 1997, pp. 389–395.

[3] S. Bhattacharya and A. Chatterjee, "A built-in loopback test methodology for RF transceiver circuits using embedded sensor circuits," in *Proc. Asian Test Symposium*. IEEE, November 2004, pp. 68–73.

[4] M. Toner and G. Roberts, "A frequency response, harmonic distortion, and intermodulation distortion test for BIST of a sigma-delta ADC," *IEEE Trans. on Circuits and Systems II*, vol. 43, no. 8, pp. 608–613, AUG. 1996.

[5] A. Halder, S. Bhattacharya, and A. Chatterjee, "Automatic multitone alternate test generation for RF circuits using behavioral models," in *Proc. International Test Conference.* IEEE, 2003, pp. 665–673.

[6] H. Yu, J. A. Abraham, S. Hwang, and J. Roh, "Efficient loop-back testing of on-chip ADCs and DACs," in *Proc. Asia and South pacific Design Automation Conference*. IEEE, July 2003, pp. 651–656.

[7] B. Razavi, *Design of Analog CMOS Integrated Circuits*. New York: McGRAW Hill, 2001.

[8] J. Rivoir, "Low-Cost Analog Signal generation using A pulse-density modulated digital ATE channel," in *Proc. Asian Test Symposium*. IEEE, November 2004, pp. 290–295.

[9] M. Burns and G. W. Roberts, *An Introduction to Mixed-Signal IC Test and Measurement*. New York: Oxford University Press, 2001.

[10] P. R. Gray, P. J. Hurst, S. H. lewis, and R. G. meyer, *Aanalysis and Design of Analog Integrated circuits*. New York: John wiley and Sons Inc, 2001.

[11] J. H. Friedman, "Multivariate Adaptive Regression Splines," *The annals of Statistics*, vol. 19, no. 1, pp. 1–141, 1991.

[12] S. S. Akbay and A. Chatterjee, "Built-In test of RF components using mapped feature extraction sensors," in *Proc. VLSI Test Symposium*. IEEE, May 2005, pp. 243–248.

[13] A. Raghunathan, H. Shin, and J. A. Abraham, "Prediction of analog performance parameters using oscillation based test," in *Proc. VLSI Test Symposium.* IEEE, Apr 2004, pp. 963–971.

[14] Analog Devices, "Data Sheets for Broadband Modem Mixed-Signal Front End," www.analog.com/UploadedFiles/DataSheets/286389915AD9865a.pdf.