RF-BIST: Loopback Spectral Signature Analysis

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Abstract

Built-In Self-Test (BIST) becomes important also for more complex structures like complete front-ends. In order to bring down the costs for the test overhead, Spectral Signature Analysis at system level seems to be a promising concept. Investigations that have been carried out are targeted on the most challenging problems: Generation of the Test Signature, Evaluation of the Signature Response, Implementation of the concept and Verification by Simulation. From investigations it can be concluded that the concept is suitable especially in the case of transceiver-type DUT.

1 Introduction

Microelectronic realisation of systems in mobile telecommunications is characterised presently by a rapidly growing integration density and more complex design structures. On the other hand the enormous cost pressure must be also taken into account. In this context the reduction of costs for test overhead is an important aspect. For that purpose Built-In-Self-Test (BIST) is applied as the fundamental approach. This means integrating suitable test structures on the chip.

A special challenge is the application of BIST in the RF- and mixed signal domain. In contrast to the test of digital systems, analog systems have only a few inputs and outputs and their internal states exhibit low time constants. From that follows, that a test with a high coverage is possible with small effort. The trade-off here is the generation of the test stimulus.

This paper is concerned mainly with the problem of test stimulus generation. The object of investigations is the so called RF front-end.

This text is structured as follows. Section 2 shortly outlines the selection of a test strategy. Section 3 describes a theoretical approach to evaluate the test signature response. In section 4 our proposal for an universal front-end BIST is emphasised. Section 5 presents a simulation environment for generating and analysing signature test stimuli and for modelling different devices under test with adjustable non-idealities. Conclusions are drawn in section 6.

2 Selection of a suitable Test Strategy

Possible strategies [1] to implement BIST for an RF front-end can be divided in two categories.

Well known from the literature is the separate test of different single building blocks of the RF front-end [11]. In this case every block corresponds to a device under test (DUT) and a special test signature will be formed according to the DUT's requirements. An appropriate processing of the DUT's signature response is also required.

The advantage of this principle is a high test coverage due to special test stimulus that take into account all test conditions of the specific DUT. The test signature can be specially designed.

The main disadvantage is the high test overhead. It results from the necessity to design a special test set-up for each block to be tested. Moreover, it will be assumed that all building blocks used to generate the test stimulus and to convert the response are ideal. Therefore any failure in these building blocks will cause corresponding failures in the test.

The second strategy [7], [8] consists in testing the whole transceiver front-end as a complete system (Fig. 1). In this case the, DUT's structure corresponds to a chain of building blocks connected in a loop at the antennas.

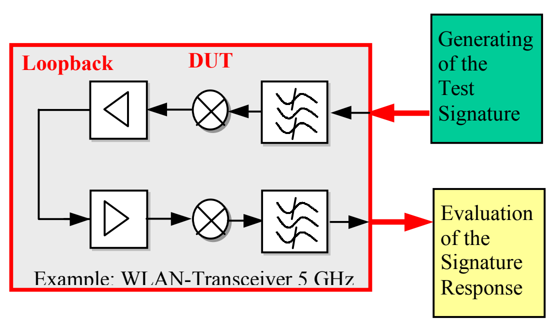

Figure 1: System testing

This principle is known from point-to-point radios as loopback technique [9]. The test signature will be injected in the transmitter's baseband interface and the signature response of the DUT will be evaluated on the receiver's

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baseband interface. Hence, all blocks of the transceiver's RF front-end are included in the DUT.

The main advantage of this principle, compared to block testing, is the lower effort. The very small test overhead is not depending on the architecture and the technology of the DUT. Therefore the results of the test will not be strongly influenced by failures in the test overhead. This principle has also a higher flexibility. Adapting the test signature to different requirements can easily carried out by changing the algorithm used for signature generation in the DSP. This allows flexibility with different architectures or technologies [3].

There are also disadvantages of this principle. Mainly, the test coverage is lower due to the fact, that the complete transceiver is tested as a whole. It could be possible to not detect bad spectral properties of the transmitted signal due to the masking by the receiver's selectivity. Furthermore, because of the higher complexity of the test signature generation, an additional DSP is required [2]. In addition both, transmitter and receiver, must be already implemented on silicon.

The argument of low costs for test overhead becomes increasingly important in the case of a quick production test in the high frequency range, such as a 5 GHz-system.

Extensions towards to implement an optimisation of the signal path using BIST is also possible. In contrast to known concepts for block-orientated self-correction with low operating frequency, the system approach allows to optimise building blocks with high operation frequencies.

3 Signature Generation and Analysis

Among other things test cost reduction is possible by scaling down the quantity of tests and employing low cost testers. In principle such requirements can be fulfilled by applying a complex and optimised test stimulus to the device under test (DUT). A sufficiently sophisticated post processing system to analyse and evaluate the DUT response is also required.

Testing methodologies of this type are called signature testing. Compared to conventional specification testing, signature testing has the following advantages:

- Multiple DUT specifications can be computed using a single response acquisition
- Reduced overhead due to the single test stimulus and the single test configuration.
- Test instruments are less complicated and cheaper.

In the following a multitone OFDM signal [10] will be used as spectral test signature. The power spectral density of this signal is depicted in figure 2.

The motivation for this choice is the fact, that a general characterisation of each block in a RF-Front-End is possible by its impulse response. In the frequency domain this corresponds to the characterisation of the block by a transfer function. For RF front-ends, a good approximation to the impulse in time domain as test signature stimulus is bandwidth-limited noise in frequency domain [6]. Therefore the proposal for the general test signature is a multitone OFDM signal

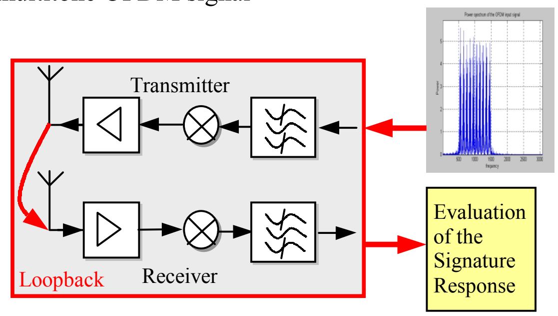

Figure 2: OFDM test signature

3.1 Synthesis of a Test Signature

An approach to generate a time domain test signature that allows the detection of different conventional measured specification parameters is presented in [4].

Our investigations are focused in particular on WLAN-Front-Ends. Therefore the DUT is a transceiver, for which the concept of modelling by transfer-functions can be used.

Usually the practical performance of a transceiver is characterised by means of some special parameters like gain, noise figure, IIP3 etc. These parameters are not depending on the concrete realisation of the circuit. They are abstract, characterising a block on its behavioural level. In this sense these parameters are closely related to transfer-functions. This underlines our motivation to apply the spectral signature analysis as a suitable BIST-concept and to use an OFDM signal as a test stimulus [5].

Let $X(f)$ be the test stimulus. It is of length n and consists of m discrete spectral lines at frequencies f_j, $j = 0, 1, ..., m-1$ different from zero, equidistantly spaced on the frequency axis. The first elements in $X(f)$ different from zero has the index wl. Corresponding with $wu = wl+m-1$ we get

$$X(f) = (0, \ldots, 0, X(f_{wl}), \ldots, X(f_{wu}), 0, \ldots, 0)$$

or short $X = (0, \ldots, 0, X_{wl}, \ldots, X_{wu}, 0, \ldots, 0)$ (1)

When feeding this signal, X to the input of the DUT we get the response, $Y(f)$.

$$Y(f) = (Y(f_0), \ldots, Y(f_{n-1})) = (Y_0, \ldots, Y_{n-1}) \quad (2)$$

with n spectral lines on discrete frequencies f_i, $i = 0, 1, ..., n-1$. Generally $n \geq m$ because of the non linear behaviour of the DUT resulting in intermodulation products.

In the case of the spectral signature analysis the property of interest concerns the relation between integrated power at the input and the output of the DUT.

More precisely, the ratio between the input X_j at frequency f_j, and the resulting output Y_i at any other frequency f_i, is of interest. This ratio is known as the transfer factor $T_{i,j} = \frac{Y_i}{X_j}$. Using transfer factors of this kind we have the possibility of easy modelling the frequency spreading behaviour of a nonlinearity in module of the amplifier type. Regarding the definitions in (1) and (2) we can define a matrix A_k with transfer factors as matrix elements corresponding to k^{th} non-linear stage

$$A_k = \begin{bmatrix} T_{0,0} & \cdots & T_{0,n-1} \\ \vdots & \ddots & \vdots \\ T_{n-1,0} & \cdots & T_{n-1,n-1} \end{bmatrix} \tag{3}$$

In principle the matrix A_k describes how the energy is transmitted from the n^{th} frequency input to the signature response Y.

It follows that in the case of an ideal and distortion-free amplifier with constant gain over frequency there is no spreading of energy from one frequency to another. Therefore all elements beside of the main diagonal in A_k are zero.

When the module of interest in the l^{th} stage is of a non-ideal and non-linear mixer type we get the matrix M_l.

$$M_l = \begin{bmatrix} T_{0,0} & \cdots & T_{0,n-1} \\ \vdots & \ddots & \vdots \\ T_{n-1,0} & \cdots & T_{n-1,n-1} \end{bmatrix} \tag{4}$$

An ideal mixer with constant conversion gain over frequency and ideal sideband suppression shifts energy only to the frequency difference between the output frequency of the l^{th} local oscillator and the discrete frequency of the input signal. Therefore all elements except parallels of the main diagonal are zero.

We consider now the filter as the module of interest. In this case the selectivity of the filter is the most important parameter. This characteristic can be modelled assuming that the filter is a linear component.

Therefore the matrix F_j can be simplified. Only elements in the main diagonal are non-zero. They corresponds to the complex value of the transfer function at frequencies $f_0, \ldots, f_{n-1}$:

$$F_j = \begin{bmatrix} T(f_0) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & T(f_{n-1}) \end{bmatrix} \tag{5}$$

In practice the behaviour of the filter is non-linear to a certain amount. In this case it is possible to model the non-linear filter by a serial of a non-linear amplifier and a linear filter.

We look now at the DUT's architecture typical for a transceiver (fig. 3). Exciting the DUT with the spectral signature X the corresponding signature response Y will be formed by all modules included in the chain.

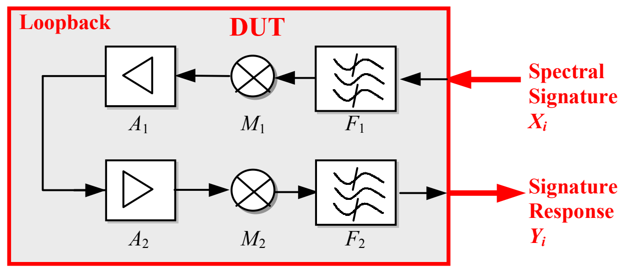

Figure 3: Chain of stages in a typical architecture

Assuming that the modules of the chain are connected trough non-reactive paths, there is a simple approach to model the whole chain using the models for the components developed above. In this case it is possible to define the dependencies of the Y from the X by multiplying the matrices.

Because of the non-linear properties of the amplifiers and the mixers, the sequence of matrices in the corresponding product is depending on the architecture of the transceiver chain. Therefore it holds:

$$Y = f(X) = F_2 M_2 A_2 A_1 M_1 F_1 X \tag{6}$$

The expression (6) could be simplified. Since all matrices are of the order (n,n), it is possible to merge all factors

$$T = F_2 M_2 A_2 A_1 M_1 F_1 \tag{7}$$

The matrix T is valid for a defined level of X.

The frequency domain approach introduced here is advantageous in comparison to the approach described in [4] because here is no need to calculate very big matrices like there for each element of the circuit. In our approach we need only to know the transmission matrices of all stages of the system for the finite number of discrete frequencies of interest. In respect of an optimisation of the signal path extension, only sensitivity matrices of stages corresponding to tuneable parameters are of interest.

3.2 Analysis of the Test Signature Response

The analysis of the stimulus response is made by calculating the "distillation quality" of the receiver. That means the capability of the receiver, to separate wanted signals inside of the wanted frequency band from all other unwanted signals.

The wanted components of the spectral signature response inside of the wanted band are determined by the sum

$$Y_w = \sum_{i=wl}^{wu} Y_i \; . \tag{8}$$

For the unwanted components of the signature response holds

$$Y_{uw} = \sum_{i=0}^{wl-1} Y_i + \sum_{i=wu+1}^{n-1} Y_i \tag{9}$$

In the following, the so called global channel selectivity *GCS* in the frequency domain will be defined. It is the ratio between the power of the wanted signal inside the wanted frequency band $f_{wl}...f_{wu}$ and the sum of the power of the wanted and the unwanted signal:

$$GCS = 10\lg\left(\frac{Y_w}{Y_w + Y_{uw}}\right) \tag{10}$$

This definition has two advantages. It takes into account filters out of tune characterised by slopes of the in-band frequency response. Also it takes into account interferers and spurs outside the wanted band coming from intermodulations.

The so called disturbance figure, *S*, calculates in terms of an I/O relation, the degradation of the ratio between wanted and unwanted spectral components, or in other terms the ratio between the global channel selectivity at the input GCS_{in} and the output GCS_{out}, if the test signature passes the DUT.

$$S = GCS_{in} - GCS_{out} \tag{11}$$

If it can be assumed, that it is possible to generate a ideal test signature at the baseband input of the transmitter, then, $GCS_{in} = 0$ dB.

Because of the continuous property of *S*, instead of a more "digital" parameter like *BER*, this figure can very effectively act as the actual value in a control loop necessary for self-correction.

The separation between the wanted and the unwanted spectral components is possible because of the orthogonality of the OFDM carriers. By using the FFT in the OFDM demodulator each carrier will be separated from all others and its amplitude and phase will be determined. This is the principle of the Fourier Voltmeter (FVM) [1], [2].

In [5] it is shown for CDMA, how to use the measurement of the pilot channel strength as the wanted signal and the total signal strength for calculating SNR, noise figure and other parameters. To adopt this principle for OFDM signals means, that measurement of pilot and total channel strength will be replaced by the measurement of each carrier amplitude using FVM for BIST and BISC purposes.

4 Implementation

We propose the block diagram depicted in figure 7 (at the end of this paper) in order to implement the loopback spectral signature analysis for a transceiver front-end.

The RF loopback [9] is marked by the dotted optional offset mixer. The usage of that mixer depends on the frequency planning of the target application. An attenuator can be also necessary. Beside this, the block diagram depicts a dotted IF loopback. The IF loopback is an option known from the literature, but its application is limited to cases, where the same IF is used in transmitter and receiver. In general, using more then one loopback can be very useful for a self correction process.

The optional detection path in the dotted box shows, that for certain reasons additional effort for the detection of the test signature responses may be necessary. This is especially the case, if the demands, for instance of the spectral purity, at the transmit antenna are very high and low adjacent channel leakage is required for a transceiver system. In such a case the loopback test of a system of transceiver type is not sufficient in respect of spectral purity. Additional effort must be spent to test and adjust the compatibility with other transceivers not included in the test set-up.

The path drawn in the dashed BISC loop box shows the additional effort for a self-correction control loop.

The loopback spectral signature test allows the on-wafer testing for SoC solutions. When the RF front-end is separated from the baseband chip, only on-board testing is possible. In both cases the input test signal can be generated by the baseband chip, if available, or by the tester that emulates the signal generation algorithm of the DSP.

5 Verification

To demonstrate the functionality of the proposed method a transceiver chain has been tested. The input signal is a multi carrier signal centred on 325 Hz with the bandwidth 450 Hz and a carrier spacing equal to 50 Hz. The input signal is converted up in frequency by an image reject mixer and filtered with a second order Butterworth band-pass filter. The mixer and the filter models the transmitter. The receiver is modelled by another second order band-pass filter, a quadrature mixer used to generate I and

Q paths, a four multiplier image reject mixer and a sixth order Chebyshev low-pass filter for channel selection. The simulated transceiver chain is presented in figure 4.

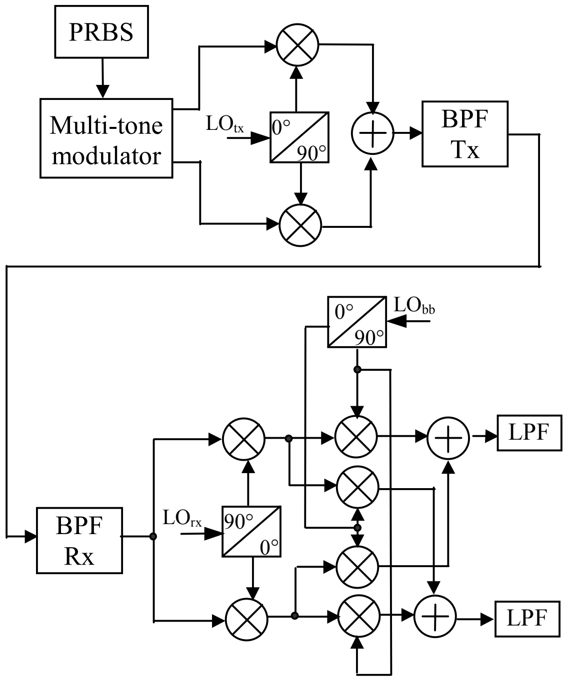

Figure 4: The simulated transceiver chain

Several non-idealities can be introduced in the simulated chain. The most important are non-linearity, amplitude and phase mismatch in the quadrature mixers and frequency errors of the filters. As example the effect of the filter non-linearity on the *SNR* and *GCS* has been simulated. The power spectral densities of the signals at the input and output of the chain are presented in figure 5.

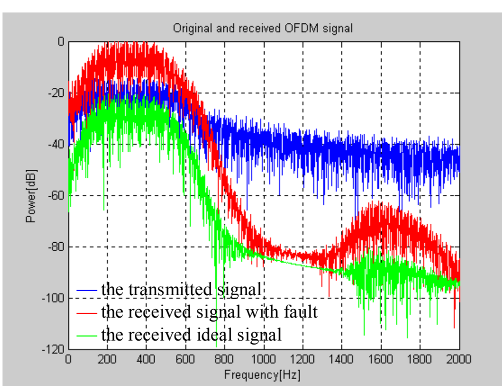

Figure 5: The input signal, the output signal for linearity error of the band-pass filter and the expected ideal output

The *SNR* and *GCS* has been measured against the input signal power. The measured values are given in table 1.

Input power [dBm]	SNR [dB]	GCS [dB]
5	73	-23
2	65	-9
1	59	-1
0.5	53	6
0.1	39	19
0.05	33	23
0.01	19	17
$5 \cdot 10^{-3}$	13	12
$1 \cdot 10^{-3}$	-1	-1
$5 \cdot 10^{-4}$	-7	-7
$1 \cdot 10^{-4}$	-21	-21
$5 \cdot 10^{-5}$	-27	-27
$1 \cdot 10^{-5}$	-35	-35

Table 1. SNR and GCS variation with the input signal level

The corresponding graphical representation is given in figure 6.

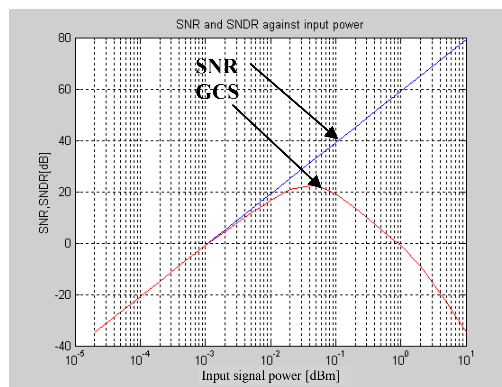

Figure 6: SNR and GCS variation with the input signal power

6 Conclusions

In this paper a method called "Spectral Signature Analysis" has been presented. The method consists of two parts: Generation of a test signature and analysis of the signature response of the DUT. This principle allows an optimisation of the signature for self-correction. A block diagram and a simulation environment realised with MATLAB have been presented also. Simulation results show that the proposed method is not only suitable for BIST of RF Front-Ends, but also for BISC.

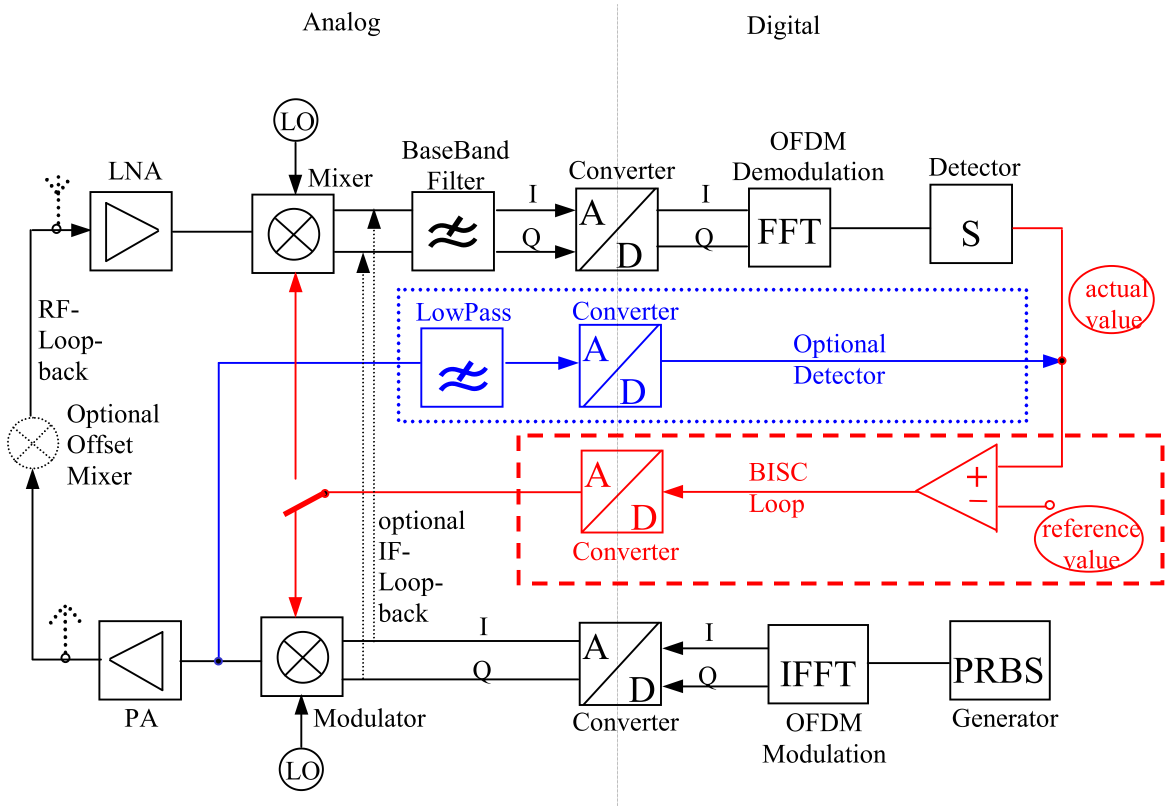

Figure 7: Block diagram

Acknowledgement

This work has been supported by the German Government (BMBF) under Grant No. 01M3040. In addition, this work has been supported by Nokia Research Center in Bochum, Infineon AG in Munich and Melexis GmbH in Erfurt.

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