Measuring Mixed-Signal Substrate Coupling
Yves Rolain, Wendy Van Moer, Gerd Vandersteen, and Marc van Heijningen

Abstract—A measurement method is proposed to characterize the substrate coupling between digital and analog sections of a mixed-signal CMOS chip. Induced noise and spurious signals can be measured by a custom-designed analog sensor. This paper proposes a method that, when given such a sensor, allows to measure the crosstalk between digital and analog chip sections. Calibrated sampling scope measurements illustrate the performance of the measurement setup.

Index Terms—High frequency measurements, mixed signal measurements, substrate noise, time domain network analyzer.

I. INTRODUCTION

The ever-growing need for the integration of analog and digital circuitry on a single chip has raised a number of interesting problems in recent years. On the measurement/modeling side, one of the issues is the need for a way to characterize the signal integrity for the analog signal blocks. On-chip disturbances induced by digital circuitry on the analog circuitry will degrade the analog signal quality. Since these disturbances are highly correlated with the previously sent data bits and/or digital clocks, their influence on the bit error rates can be much higher than that of an equal power white noise source. Hence, modeling this correlated perturbation as an uncorrelated noise source may result in the underestimation of the signal degradation.

To measure the correlation between analog perturbation and digital signal activity, a mixed-signal chip containing an analog sensor and a digital excitation source was developed. Basically, the output signal of a large FET that senses the substrate effects induced by the digital ports is fed to a differential amplifier to obtain a larger signal that can be taken off-chip. A differential output stage is used here to avoid common-mode contamination of the analog signal. Further immunity of the analog amplifier against substrate effects is obtained using an external current source for the analog stages. To keep power supply-induced perturbations minimal, the analog and digital dc power supplies are fully separated.

This paper is devoted to the calibrated measurement of the contamination effects. A time-domain network analyzer ([1], [2]) (TDNA) setup based on a digital sampling oscilloscope (DSO) is therefore developed. The measurement setup, its calibration, and the associated data processing are then derived. Finally, the processed measurements are shown and some conclusions are drawn.

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B. Proposed Method

A first, seemingly straightforward approach to characterize the system would be to apply a periodic, band-limited excitation and to evaluate the linear transfer function between excitation signal \( V_{in} \) and the difference \( V_+ - V_- \). However, due to the digital nature of the input, a band-limited test signal is varying too slowly in time and results in an excessive timing jitter.

The alternative that is used here is to excite the digital gates with a clean, broad-band, step-like excitation that matches the level requirements of the logic family used. Measuring the transient waveforms at the digital input and the analog outputs then enables relating excitation \( V_{in} \) and the spurious signal \( V_+ - V_- \). The broad-band nature of the excitation and the device requires the use of a broad-band measurement device. In this paper, a digital sampling oscilloscope is used. This device has the advantage of a very wide bandwidth and has four simultaneously sampled data channels. The main drawback is that due to the sampling principle, the excitation is required to be periodic to allow the measurement. Hence, the inverter string will be excited by a clean square-wave signal that will also be used to trigger the DSO and will have a period that is long enough to ensure that all the device transients are damped in half a period.

The broad-band character of both signal and device (bandwidth close to 1 GHz) implies that, normally, both the reflected and the incident waves have to be measured at the three interaction ports. However, since the measurement device is designed to have measurement channels with a broad band 50 Ω impedance, a simplified setup will be derived. Ideally, one can assume the following:

Assumption 1. Perfect Port Match: The signal ports of the DUT are perfectly matched to 50 Ω over the measurement bandwidth. Measured port voltage and incident voltage wave are therefore equal. As shown in Fig. 1, the signals \( a_{V_+} \), \( a_{V_-} \) and \( a_{IDC} \) are zero by this assumption and, hence, only five signals remain to be measured. A four-channel, time-domain network analyzer (TDNA) is then used to measure these five waveforms. This is possible in a TDNA, under the condition that the reflectometric measurement has nonoverlapping responses for both the impinging waveform \( b_{V_{in}} \) and the reflected waveform \( b_{V_{D}} \).

Assumption 2. Non-overlapping Reflections: There is no overlap between the incident and reflected signal transients at the excitation port \( V_{in} \).

C. Practical Setup

The practical measurement setup consists of a DSO HP54121A, an HP33121 waveform generator to generate the square waves, an HP 4142 dc biasing system to provide the bias voltage, and a Microtech probing station to apply dc bias and to measure the signals. The devices are hooked up as shown in Fig. 4. Each probe allows connecting two signals and two bias voltages simultaneously to the DUT. The probes have a connection footprint as shown in Fig. 5. Calibration of such complex devices is complicated by the asymmetric layout of the signal ports. It will be close to impossible to realize a...
through connection with equal delay for the four signal ports used in the setup.

III. CALIBRATION OF THE MEASUREMENT SETUP

A. Theoretical Considerations

The off-chip calibration is responsible for bringing the calibrated measurement plane to the probe-tips and the connectors of the DUT. Due to the broad-band nature of the system, a full microwave calibration of the setup is required.

Unlike the standard microwave calibration for a frequency domain network analyzer, where a sinusoidal excitation is used, a broad-band, step-like waveform will be applied in this case to the measurement setup, under the assumption that:

Assumption 3: The measurement device is a linear, time-invariant system, a linear relation exists between calibrated, and measured waveforms. To further simplify this relation, one more assumption is required.

Assumption 4: The crosstalk in between measurement ports is negligible. The error correction matrix now reduces to the following form [4]

\[ \begin{bmatrix} a_{mD}^m \\ b_{mD}^m \\ b_{mV_D}^m \\ b_{mV_o}^m \\ b_{mI_D}^m \\ b_{mI_o}^m \end{bmatrix} = \begin{bmatrix} e_{11} & e_{12} & 0 & 0 & 0 & 0 \\ e_{21} & e_{22} & 0 & 0 & 0 & 0 \\ 0 & 0 & c_{33} & 0 & 0 & 0 \\ 0 & 0 & 0 & c_{34} & 0 & 0 \\ 0 & 0 & 0 & 0 & c_{55} & 0 \end{bmatrix} \begin{bmatrix} a_{mD}^e \\ b_{mD}^e \\ b_{mV_D}^e \\ b_{mV_o}^e \\ b_{mI_D}^e \\ b_{mI_o}^e \end{bmatrix} \]  

where a superscript \( m \) denotes a measured and a superscript \( e \), an exact quantity. Since multiple reflections are ruled out by assumption 1 and the square wave excitation \( V_D \) is slow enough to allow damping of the transients within half a period, the measurements of the reflected and incident waveforms at port \( V_D \) are independent; hence, \( e_{12} = e_{21} = 0 \). The full calibration of the setup is hereby reduced to a frequency response calibration, where only the diagonal correction terms are required to be known.

The goal of the setup is to measure the coupling between the digital circuitry and the analog sensor. Since the coupling is a relative quantity, a relative calibration of the setup is sufficient to remove systematic errors. As a consequence, one of the \( e \) functions in the correction matrix may still be chosen arbitrarily.

B. Frequency Response Calibration

To obtain a calibration that is valid over the full bandwidth of the measurement probes, a very wide band excitation signal is required. The HP54121A DSO has a reflectometric mode where a periodically repeated step signal is applied to one of the measurement ports of the device. This step has a bandwidth in excess of 10 GHz, which makes it suitable as a calibration signal for the bandwidth of 1 GHz, as will be used for the device characterization.

Note that, for the calibration measurement, the schematic diagram of Fig. 1 is somewhat modified, since the excitation signal is generated inside the DSO. The modified setup for the signal generation is shown in Fig. 6. Note that no reconnections are required to switch between both setups. The only difference is that during the measurements the square wave generator generates the signal, while the internal generator acts as a load, while, during the calibration, the square wave generator acts as a load and the internal generator drives the line.

1) Theoretical aspects: To determine the error terms required in the calibration, four independent experiments are performed. In a first experiment, a short-circuit is applied at the probe tip of the excitation port \( V_D \), using the setup of Fig. 6. Assuming that the circuit path between the measurement port and the samplers is reciprocal, the incident waveform at the probe tip is

\[ a_{V_D}(\omega) = e_{22}(\omega) b_{mV_D}^m(\omega), \]

\[ a_{V_o}(\omega) = e_{11}(\omega) a_{V_D}(\omega). \]  

If an ideal, short circuit is assumed at the probe tip, \( a_{V_D}(\omega) = -b_{mV_D}^m(\omega) \) and the relation between the measured and the exact reflected waveforms becomes

\[ b_{mV_D}^m(\omega) = -e_{33}(\omega) e_{11}(\omega) b_{mV_D}^m(\omega). \]  

In the next experiments, the \( V_D \) probe is successively connected to the other three ports and a transmission measurement is performed

\[ b_{mV_D}^m(\omega) = e_{33}(\omega) e_{22}(\omega) a_{mV_D}^m(\omega), \]

\[ b_{mV_o}^m(\omega) = e_{44}(\omega) e_{22}(\omega) a_{mV_o}^m(\omega), \]

\[ b_{mI_D}^m(\omega) = e_{55}(\omega) e_{22}(\omega) a_{mI_D}^m(\omega). \]  

In this thru connection experiment, the exact waveforms are set equal. To obtain a calibration plane that is exactly located at the probe tips, the connection between probe tips should be infinitely short.

Given (3) and (4), the transfer functions \( e_{22}(\omega) \) to \( e_{55}(\omega) \) can be determined, under the assumption that the thru connection has zero length. Choosing \( e_{11}(\omega) = 1 \) in the relative calibration

\[ e_{33}(\omega) e_{22}(\omega) = \frac{b_{mV_D}^m(\omega)}{a_{mV_D}^m(\omega)}, \]

\[ e_{44}(\omega) e_{22}(\omega) = \frac{b_{mV_o}^m(\omega)}{a_{mV_o}^m(\omega)}, \]

\[ e_{55}(\omega) e_{22}(\omega) = \frac{b_{mI_D}^m(\omega)}{a_{mI_D}^m(\omega)}, \]

and

\[ e_{22}(\omega) = \frac{-b_{mV_D}^m(\omega)}{a_{mV_D}^m(\omega)}. \]  

2) Data Processing of Step-like Signals: The only remaining problem is to transform the step-like signals as measured in the broad-band experiments to the frequency domain. Since straightforward application of the FFT proves
impossible due to leakage errors, the method proposed in [3] is used to obtain an exact compensation for the presence of the step. This allows easy and exact calculation of the step-like, excitation-based transfer functions.

3) Practical Realization of the On-wafer Connections: Since the signal connections on the probes are grouped two by two and the connections are asymmetrical, the thru connection of these ports poses some layout problems. Cross connection of signal ports on different probes requires that the connecting wires are crossed, while connection of the signal ports on the same probe requires connecting lines that are curved. Hence, it can be argued that these layout differences imply delay difference in the different thru connections and, thereby, falsify the measurement. However, careful design of the calibration substrate to keep all connections short and of equal length reduces this delay uncertainty sufficiently to neglect it.

IV. EXPERIMENTAL RESULTS

A. Verification of the Hypotheses

First, the \( V_2 \) port will be checked for the port match, as imposed by assumption 1. A short-circuit to ground is applied at the probe carrying this signal.

The general shape of the signal is shown in Fig. 7(a). Clearly, there are no major reflections present beside the short-circuit at the probe tip. In the zoomed top of the step (b), a small transient included in the pulse is noted on the left side. On the right-hand side, the reflections on the different connectors and interfaces of the probe itself are found. On the zoomed bottom of the step (c), a small residual reflection appears around 8.7 ns, which is the expected position of the first multiple reflection of the short-circuit. A reflection is indeed present there, whose shape indicates that it is a multiple of the reflection on the probe connection rather than on the DSO port. The relative size of the main versus the parasitic reflection allows considering this reflection as a second-order effect.

B. Calibration Results

The values of the correction functions as a function of the frequency are shown in Figs. 8 and 9. Up to 15 GHz, the amplitude roll-off of the measurement channel is lower than 5 dB. Note the rising standard deviation of the correction coefficient. This is mainly due to the step-like excitation signal, which has a power spectrum that decreases with increasing frequency. Note that above 5 GHz, a ripple appears in the magnitude. This can be caused by the multiple reflections that were also detected in the time-domain waveforms.

The results for the other calibration coefficients are very similar, except for the delay that is much higher. For \( c_{33} \), a delay of 3.97 ns was subtracted before the phase plot shown in Fig. 9(b) was produced and for \( c_{44} \), a delay of 3.92 ns was subtracted. Note the very small difference in behavior between the correction terms, if the delay difference is accounted for.

C. Corrected Noise Coupling Measurements

In the following step, a digital signal is applied to the digital input port \( V_2 \), while the differential signal is measured at the \( V_+ \) and \( V_- \) pins. The digital circuitry is configured such as to enable one seven gate string on the digital chip part.
To ensure that all the transients are damped, the frequency of the applied digital signal is fixed to be 100 KHz. The transients of the rising and the falling edge of the digital signal are shown in Fig. 10. Note that, as could be expected, the transient length of the falling edge transition is quite a bit longer than the rising edge one. Note also that, unlike expected, the magnitudes of the two transient signals are quite different. Up to now, there is no clear cut explanation to this. One possibility is that the presence of an asymmetrical load impedance on the inputs of the differential amplifier introduces an asymmetry in the gain versus frequency characteristic of the differential pair. A second possibility is that the external resistor, which should act as an ideal current source, has a dynamic behavior that degrades the differential operation.

To allow for a valid interpretation of the results, it was decided to perform a kind of “on-the-chip” calibration to minimize the imbalance between signals. Unfortunately, there is no way to inject a known current into the substrate for calibration purposes. Hence, it will be assumed that the imbalance between channels is an artifact introduced by the sensor. It acts as a systematic error and, therefore, can be eliminated through calibration. Since, even visually (see Fig. 10), it is clear that the noise degradation on the $V_-$ is quite a bit larger, it was decided to take the $V_+$ signal as the reference signal and to calibrate the $V_-$ signal to restore the symmetry, as explained below.

First, the dc components $DC_{V_+}$ and $DC_{V_-}$ are stripped from both signals. Next, the RMS value of both signals is evaluated. A gain compensation is then performed to ensure that the RMS value of both signals is equal. The corrected $V_-$, labeled $V_{\text{cmp}}$, is calculated as follows

$$V_{\text{cmp}} = \frac{\text{rms}(V_+ - DC_{V_+})}{\text{rms}(V_- - DC_{V_-})},$$

Finally, to remove the differential delay between the two signals, the maximum of the cross correlation function $\text{xcorr}(V_+, V_-)$ is determined. This determines the delay that has to be added to obtain an optimal alignment in time for the two signals.

To check whether or not the discrepancy between channels can be modeled by a static gain and a linear delay, the transfer function between spectra $V_+(\omega)$ and $V_-(\omega)$ is determined. The result for the rising edge transient is shown in Fig. 11. Note the high signal-to-noise ratio ($\text{SNR} \approx 20$ dB) up to 1 GHz.

Above 350 MHz, the gain of the transfer function is relatively flat. Below that, however, a quite wide resonance and a sharp anti-resonance around 250 MHz are superimposed. Clearly, the dynamic difference between the channels will only approximately be modeled by the proposed static compensation and, hence, a residual common-mode signal will persist.

From the available sensor, a voltage gain compensation of $G = 2.82$ and a delay compensation of 130 ps are used to compensate the measured $V_-$ signal. To control whether or not this compensation is a measurement artifact, the sensor characteristic has been simulated (HSpice simulation) and calculated analytically. Both approaches show that there is, indeed, a gain difference present. In both cases, a gain difference of approximately 2.7 was found. This matches the measured value quite
closely and shows the good accuracy of the proposed measurement approach.

The compensated signals are shown in Fig. 12. Note that the response consists of two different parts: the fast transient in the front is the signal induced by the fast digital gates, while the slower tail at the end is very probably related to the impulse response of the differential amplifier itself. Note that in the corrected waveforms, there are seven transient peaks (seven gates were actually activated) that appear much more clearly separated, while the common-mode signal is smaller. This indicates that the compensation indeed shows some of the expected effect on the measurements.

For the falling edge transients, the situation is more complex, as shown on Fig. 13. There are many more transients present and, therefore, the response is more confused. However, looking closely at the difference response, again a slow and a fast transient can be denoted: the slow transient that is responsible for the periodic appearance of peaks in the second part of the response has a periodicity roughly equivalent to the periodicity of the slow transient in Fig. 12. The presence of many more fast transients tends to prove that the transient response of the gates is much more oscillatory in this case and this lightly damped behavior is responsible for the complex wave shape.

Clearly, the string of gates gives a complex structure to the signals that are measured on the differential amplifier output. It would have been easier to extract a quantitative model for the interaction if it would have been possible to use a single gate to generate the transient.

V. CONCLUSION

A calibrated measurement setup for the on-wafer probing of the substrate coupling between analog and digital parts of a circuit is proposed. The setup uses a time-domain network analyzer that is fully calibrated under the assumption of a perfect port match. Despite some nonideal behavior that complicates the analysis, the proposed sensor is shown to be capable of sensing the analog stray voltage induced by the digital circuitry. To obtain a full, absolute-level characterization, however, there is a need for circuitry to calibrate the on-chip sensor, which was not present in the current hardware.

REFERENCES


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