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Base-Band Impedance Control and Calibration for On-Wafer Linearity Measurements

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Abstract *— This paper introduces a direct and accurate method for controlling and measuring the on-wafer device terminations at the base-band / envelope frequency, using an extension of a conventional network analyzer setup. The base-band impedance can be adjusted manually as well as electronically and is able to (over)_compensate the losses in the measurement setup. This facilitates on-wafer base-band terminations ranging from negative to high Ohmic values. The proposed measurement techniques are particularly useful when characterizing active devices for their linearity.*

Key Words — Load-pull measurement, calibration, intermodulation distortion, on-wafer measurement, base-band impdance termination.

I. Introduction

The effect of the base-band (envelope frequency) impedance on the linearity of active devices has been the subject of many publications [1], [2], [3]. In contrast with this attention in literature, is the lack of commercial solutions for the control and calibrated measurement of these base-band impedances. This is even more pronounced when considering on-wafer active device characterization setups. Here, the use of conventional bias tees in combination with controlled DC sources for the biasing of the Device under Test (DUT) yields undefined base-band impedances. Although not problematic in small-signal measurements, large signal linearity characterization, like a two-tone test, requires information on the impedance at the difference frequency while controlling its value is desirable in many situations. For this reason, a method has been developed for the control and calibrated measurement of the envelope / base-band device terminations in an on-wafer environment. The proposed solution is integrated as part of a harmonic load-pull system [4] and is particularly useful for nonlinear device characterization. While [4] is focused on the implementation of the RF signal paths, this paper descries in detail the implementation of the base-band impedance control with its associated real-time calibration.

II. Impedance Measurement and Control

The most convenient way to control the base-band impedances is to combine the circuit with the biasing paths of the DUT. The solution described in this paper uses custom bias tees with a very low inductance in the bias path. As can be seen from **Figure 1**, the DC source is AC decoupled from the DUT by a high valued choke.

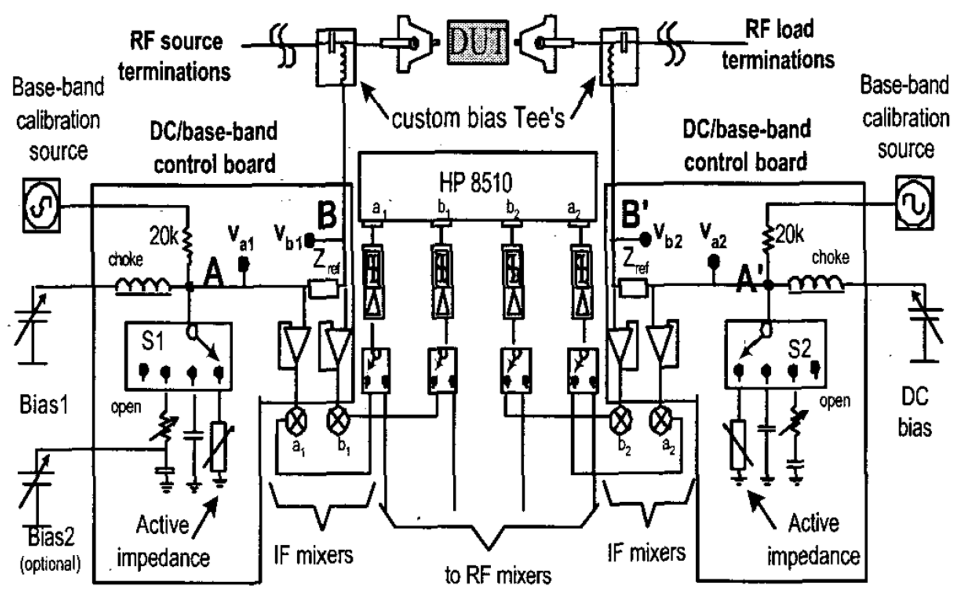

Figure 1. Proposed base-band /envelope impedance measurement and control.

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Since this choke provides a very high impedance even at low frequencies, the base-band impedance Z_{BB} can be selected through a DIL switch (S1 & S2). Currently three selections can be made for the base-band impedance termination but this can be easily extended at a later time.

The first choice is an AC ground, which is quite often favorable when working with FET devices in highly linear operation. The second selection provides an ohmic base-band termination through a potentiometer. This type of loading is often desireable at the input of bipolar transistors when using out-of-band terminations for maximum linearity [2].

Note that optionally, the device can be biased through this potentiometer enforcing $Z_{DC} = Z_{BB}$ (See Bias2 in **Figure 1**). This allows a more straightforward comparison with practical circuit implementations.

The third switch selection provides an electronically controlled base-band impedance which can force ohmic terminations ranging from negative up to high ohmic values. This newly developed circuit facilitates on-wafer AC short conditions for the base band by compensating the reference resistor (Z_{ref}) and other ohmic losses in the connecting cables and probes. Its value can be changed electronically as well as manually. For the moment we skip complex base-band impedances since they will cause asymmetry in the low and high IM3 products [2], [3],[6], [7]. However, there is no practical limitation to add complex base-band impedances at a later stage.

Note that the proposed implementation provides base-band loading conditions for all nonlinear mixing products that fall in this band, starting from DC. This means that it is also possible to measure the nonlinear behaviour of this of the DUT using realistic (modulated) input signals. This in contrast with signal injection solutions [3], which only work for one single frequency.

The reference resistor (Z_{ref}) is included to measure the AC signals at point A and B. These two signals are measured with a network analyzer (NWA). Before directing these signals to the NWA, they are buffered and upconverted to 20 MHz, which is the IF of the HP8510 network analyzer. Custom bias tees with very low inductance in the bias path connect the circuit to the DUT. Finally, for calibration purposes, a synchronized source is connected through a high valued resistor (20 kΩ) to point A. The DC / base-band control board prototype, which combines the above is shown in **Figure 2**.

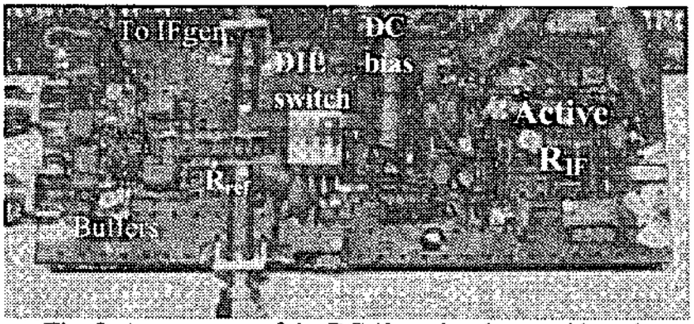

Figure 2. A prototype of the DC/base-band control board.

The use of separate mixers for the base-band signal requires switches in the 20 MHz IF path of the NWA to combine the measurement setup with the conventional RF test set. This results in a slightly more complicated NWA system, but still yields a single instrument solution. By using high quality Agilent switches no notable effect of the switching action on both the bias-band as well as the RF calibration could be observed.

III. Calibration of the Impedance Measurement

In the development of the base-band impedance measurement / calibration technique, two major requirements had to be fulfilled: first the possibility to provide AC shorted conditions on-wafer, which requires the use of an active circuit to compensate for the ohmic losses in the signal path (cables, probes, etc.). Second, we want to provide variable base-band impedances during the measurement. The combination above implies the use of a real-time like measurement technique for the impedances. we can achieve this functionality by making use of two modes of operation, namely; the calibration mode and a measurement mode.

A. Calibration Mode:

For the calibration mode, the schematic of **Figure 1** can be simplified to that of **Figure 3**. In this mode the calibration source is activated while the DIL switch

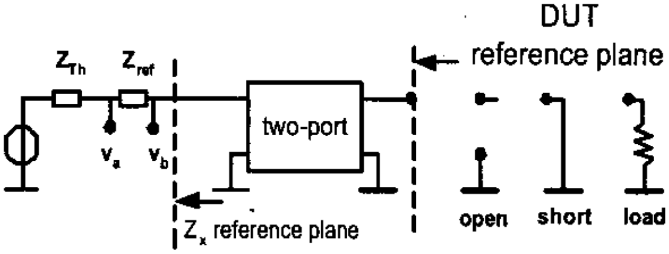

Figure 3. Equivalent schematic of the calibration mode.

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is set to an open in order to avoid a short-circuit condition for the calibration signal. In this configuration the calibration source will provide the signal for the readings of the AC voltages v_a and v_b, which will depend on the standards connected at the DUT reference plane. Consequently by measuring these known standards, we can achieve information on the connecting two-port network between the reference resistor and the DUT and also take into account the gain and phase differences in the measurement paths of v_a and v_b by the NWA.

The calibration source with its parasitic loadings can be replaced, for the sake of simplicity, by a Thèvenin voltage source (V_{Th}) with internal impedance Z_{Th} (**Figure 3**). It will be shown that neither the value of the V_{Th} nor the value of Z_{Th} is important for our calibrations. To be fully compatible with the RF calibration procedure, avoiding the increase in calibration complexity we limit ourselves to the use of the classical short, open and load conditions at the DUT reference plane. This will yield a linear system of equations from which the three unknowns can be solved. Consequently, we can solve for a complex correction factor α, which represents the relation between the ratio v_a/v_b, and the ratio a/b measured by the NWA.

The two leftover unknowns allow us to derive the information of the two-port connecting network under the assumption of a symmetrical network ($y11 = y22$ & $y12 = y21$) since this network consists mainly out of 50 ohm line connections. In practice, it works out that the symmetry assumption is valid under the constraint that the custom bias tees exhibit a low inductance in their bias path. With the above in mind we can write for the ratio of the NWA signals **η**:

$$\eta = \frac{a}{b} = \alpha \frac{v_a}{v_b} = \alpha \frac{Z_{ref} + Z_{in2\,port}}{Z_{in2\,port}} \qquad (1)$$

Where:

a, b	=	Measured waves from NWA.
α	=	Complex correction factor.
$Z_{in2port}$	=	Input impedance of the two-port.
Z_{ref}	=	Reference impedance.

Note that $Z_{in2port}$ for a symmetrical network is given in terms of the admittance parameters as:

$$Z_{in2\,port} = \frac{y_{11} + y_L}{y_{11}^2 - y_{21}^2 + y_{11} y_L} \qquad (2)$$

With y_L being the load admittance at the DUT reference plane. Substitution of the short, open and load conditions for y_L in eq. 2 yields:

$$Z_{in2\,port\_short} = \frac{1}{y_{11}} \qquad (3)$$

$$Z_{in2\,port\_open} = \frac{y_{11}}{y_{11}^2 - y_{21}^2} \qquad (4)$$

$$Z_{in2\,port\_load} = \frac{y_{11} + 1/R_L}{y_{11}^2 - y_{21}^2 + y_{11}/R_L} \qquad (5)$$

With R_L being the resistance of the load standard. Substitution of (3) to (5) in (1) yields our system of equations which can be used to solve for the three unknowns α, $y11$ and $y21^2$. This results in:

$$\alpha = \frac{R_L \eta_S (\eta_L - \eta_O)}{Z_{ref}(\eta_S - \eta_L) + R_L(\eta_L - \eta_O)} \qquad (6)$$

$$y_{11} = \frac{\eta_S - \eta_L}{R_L(\eta_L - \eta_O)} \qquad (7)$$

$$y_{21}^2 = \frac{(\eta_S - \eta_L)(\eta_S - \eta_O)[Z_{ref}(\eta_S - \eta_L) + R_L(\eta_L - \eta_O)]}{R_L^2 Z_{ref} \eta_S (\eta_L - \eta_O)^2} \qquad (8)$$

With:

η_S	=	a/b, short condition.
η_O	=	a/b, open condition.
η_L	=	a/b, load condition.

This concludes the calibration procedure.

B. Measurement Mode:

In this mode the calibration source is switched-off and we measure an active device under two-tone conditions, or with a modulated signal. These signals driving the DUT will generate (due to the always present non-linearities) a base-band signal which is used for the impedance measurement. The impedance Z_{source} experienced by the DUT at the envelope frequency is now given by the input impedance of the black box at the DUT plane (see also **Figure 4**) for

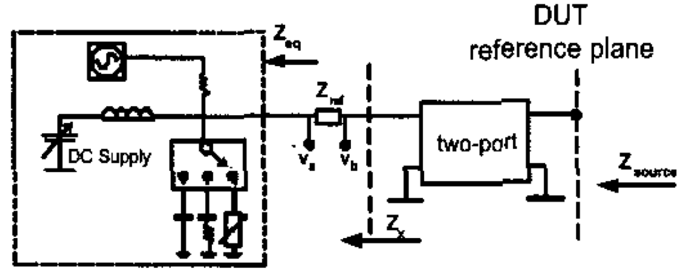

Figure 4. Equivalent schematic measurement mode.

which we can write:

$$Z_{source} = \frac{y_{11} + 1/Z_x}{y_{11}^{\ 2} - y_{21}^{\ 2} + y_{11}/Z_x} \qquad (9)$$

with: $Z_x = Z_{ref} + Zeq$ being the loading impedance of the black box. This loading is equal to the sum of the reference impedance and the equivalent impedance Zeq of the connected bias/control circuit. This remaining "unknown" impedance is given by the corrected NWA ratio of the AC voltages over the reference impedance, namely:

$$Z_{eq} = \frac{Z_{ref}\eta}{\alpha - \eta} \qquad (10)$$

With η being the a/b ratio given by the NWA in the DUT measurement mode.

Note that with the procedure above we have achieved a simple calibration and measurement technique for the base-band/envelope terminations which requires no Load-Thru, (SOLT) RF calibration technique. Furthermore, since Zeq is measured continuously during the DUT measurement mode, a real-time-like calibration is achieved. This allows free tuning of the base-band loading conditions and, at the same time, mesuring its actual value in an accurate manner.

IV. Bias Tees

Conventional bias tees are optimized for their wide band behavior and typically use a high inductance in the bias path. This disqualifies them when using IF impedance control. For this reason we have designed custom bias tees which have ben optimized for our application at 1.95 GHz to have low bias path inductance and low insertion loss for the fundamental and harmonic frequencies (Nx1.96 GHz). **Figure 5** shows such a bias tee.

The measured response is given in **Figure 6**. As can be seen, the use of a 1/2-wavelength stub results in a repeated transmission band at each harmonic

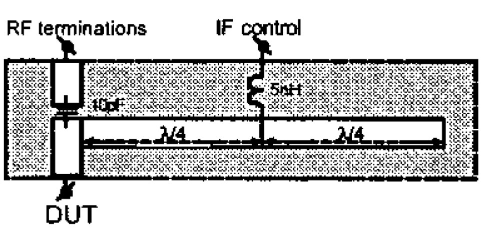

Figure 5. Custom harmonic bias tee.

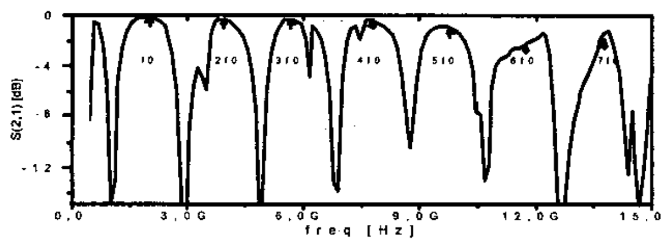

Figure 6. s_{21} of the harmonic bias tee.

frequency. The losses at the fundamental frequency are limited to 0.2 dB making it compatible with load-pull requirements. Since the IF connection is made on a virtual ground for the fundamental frequency, its inductance can be kept very low, since the lowest frequency where the inductance must present a high impedance is at the 2nd harmonic (3.92 GHz). Also the DC-blocking capacitor can be chosen small since the first pass-band is the fundamental frequency. This will reduce the capacitive loading of the bias/control circuit.

A. Alternative Solution

With the above mentioned bias tee design, there are high reflection coefficients outside the frequencies of interest. For some active devices this can yield stability problems. For this reason, an alternative solution is proposed that circumevents the use of bias tees. This solution is particularly useful in harmonic load-pull setups and is based on the use of several traveling wave directional filters [8], [9], [10] to separate the harmonics. From **Figure 7** it can be seen that there is a direct DC/base-band signal path from the DUT to the control board. In this way, the RF part of the system is completely isolated from the low frequency path.

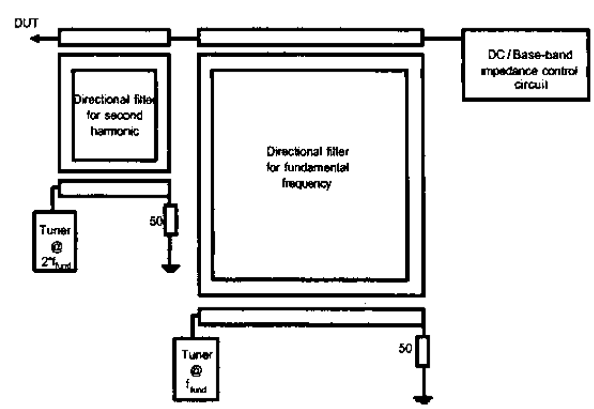

Figure 7. Harmonic multiplexer with base-band impedance control and bias circuit.

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V. Measured Results/System Performance

The calibration was performed as discussed previously. After changing to measurement mode using a bipolar DUT, the base-band impedance seen at the DUT reference plane could be measured. By using the DC/base-band control board with various passive and active terminations, loads ranging from negative to high ohmic values were obtained. The terminations provided to the DUT are measured with the previously described technique at 0.1, 0.5, 1, 2.5 and 5 MHz. The results are given in **Figure 8**.

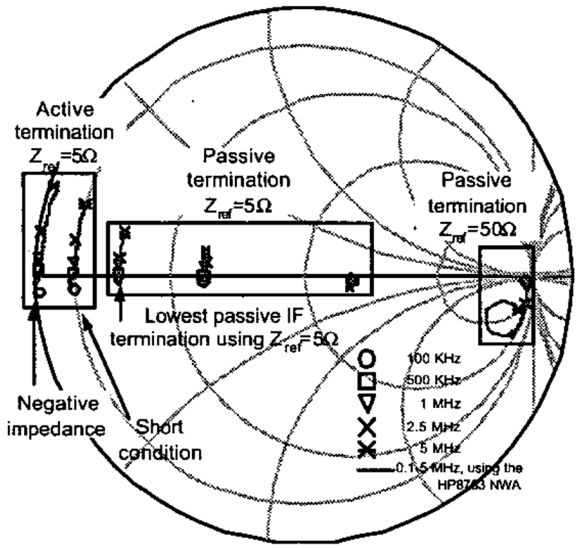

Figure 8. Various base-band impedances seen by the DUT, measured using an independent HP8753 NWA as reference..

To check the accuracy of the implemented calibration procedure, an HP8753 NWA was connected to the testports of the system by using a through standard on a calibration substrate. In this way, the actual base-band source and load reflection coefficients seen at the DUT reference plane can be measured independently and compared against the calibrated values obtained by the system. This comparison is given in **Figure 8**. Note the excellent agreement between these measurements.

From **Figure 8** we can also observe that the lowest obtainable impedance level with a passive termination is limited by the value of the reference impedance plus the ohmic losses in the measurement system. The use of an active termination allows compensation of these losses. This active termination can provide impedances ranging from a few Ohms negative up to very high impedances. Another issue is the choice of the reference resistor. A low value yields high accuracy for low impedances, while a high value gives the best results for high ohmic terminations. In practice one wants to keep the value of the reference resistor low in order to make the loss compensation by the active circuit less constrained.

A. Frequency Dependence of Γ

When using wide-band signals, the phase variation of the reflection coefficient with frequency at the DUT reference plane should be minimized in order to provide a realistic termination. For this reason the path between the control circuit and the DUT should be minimized. In our experimental setup we have already eliminated most of the cable connections except for a 25 cm coax cable to the probes. Note that this short cable is already responsible for almost 5° phase variation of Γ over 5 MHz bandwidth. This consideration yields to the conclusion that by using an even more compact construction of our setup, we can most likely reduce the phase variation of Γ by a factor of two. This more compact construction could be obtained by combining the DC/base-band control circuit with either the custom bias tees or the alternative solution based on directional filters.

B. DUT Measurement Results

As a final illustration of the importance of base-band impedance tuning for device linearity, consider **Figure 9**. Here the third order intermodulation distorion

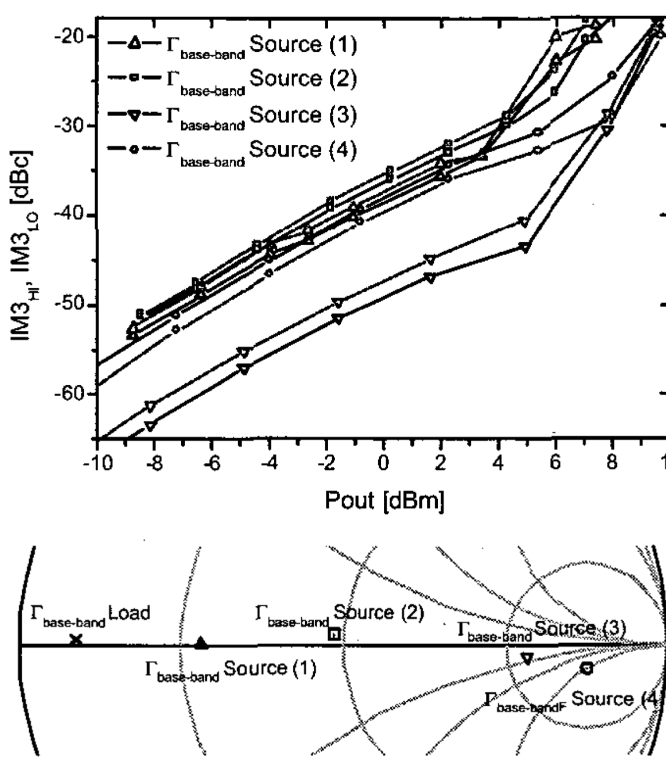

Figure 9. IM3 components (upper) at various base-band loading conditions (lower).

components (IM3) of an Infineon bipolar SiGe transistor (B7HF) with an emitter area of $1.55 \times 20 \times 2$ μm^2 is given as function of output power for several base-band loading conditions at the input of the device.

At the fundamental frequency, the DUT was terminated for a conjugate match at the input and the output was matched for maximum gain. The second harmonic loading conditions were an open for the input and a short for the output.

For these specific device terminations, a clear optimum for the source base-band termination ($\Gamma = 0.56$ or $Z_S = 177\ \Omega$) shows a minimum for the IM3 components. Also note the good symmetry of the low and high IM3 products which only differ by a few dB in magnitude.

VI. Conclusions

A single instrument concept is introduced for the real time calibrated measurement and control of the base-band source and load impedances for use in device linearity characterization.

VII. Acknowledgment

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