In-fixture calibration of an S-parameter measuring system by means of time domain reflectometry

IN-FIXTURE CALIBRATION OF AN S-PARAMETER MEASURING SYSTEM BY MEANS OF TIME DOMAIN REFLECTOMETRY[1]

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ABSTRACT

We present a technique which resorts to the time domain capabilities of a vector network analyzer and to the network synthesis tools, in order to perform an in-fixture calibration of the S-parameter measurement system directly to the ports of the device under test. The effects of the customer's non ideal fixtures can be removed without requiring the insertion of standard components or particular loads, which can affect the calibration effectiveness. The inaccuracies due to the precision of the actual loads and to the connection repeatability are also avoided.

Some simulation results demonstrate the very good capability of the technique. Experimental tests were also carried out on an actual microstrip transistor fixture, showing a very satisfactory launcher modeling and de-embedding.

INTRODUCTION

For modern microwave vector network analyzers it is possible to perform accurate calibrations up to the instrument coaxial connectors, by using standard coaxial components.

Measurements on active components or other devices, whose terminals are not of coaxial type, require suitable transitions and interface fixtures between the device and the measuring system. These fixtures introduce measuring errors [1] that are very difficult to evaluate.

Some techniques, both in the frequency [2,3,4] and the time domain [9,10], have been proposed that satisfactory overcome these problems, but do not solve them

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completely; in facts there remains the uncertainty of the connections repeatability (especially with MMIC fixtures); furthermore at higher frequencies the characteristic impedance of the delay line is subjected to variations that influence the transition identification in a way that is difficult to evaluate.

An ideal calibration procedure should be capable of characterizing the fixture without relying on any sort of assembly/disassembly for the insertion of the standard loads.

The procedure that we describe meets the above requirements, because, after calibrating the network analyzer up to the coaxial ports [7], it consists in measuring only the empty fixture.

The facilities offered by the network analyzer HP 8510 [8] are essential for the performance of the proposed techniques, which consists in isolating single echoes of the time domain response of the fixture.

This instrument performs frequency measurements of the S-parameters on a frequency band, that can be very wide, depending on the instrument hardware; the sampled frequency measurements are stored into one of the instrument memories, and may be acted upon with mathematical tools, such as the *chirp-z* transform, in order to get the impulse response of the device under test [5].

The reflectance echoes of the time response may be gated, stored in different memories, and acted upon with the same tools.

If the first echo is isolated as said before, it can be viewed as the input impulse response of a two port whose output port is matched. Reverting to the frequency domain, the reflectance corresponding to this echo may be viewed as the input scattering parameter of this two port.

Assuming that it is a lossless network, circuit theory allows to identify almost completely its whole scattering matrix, a part a delay on the input and on the output side; these delays may be modeled with lines, although the output line is of less interest, because its effects may be included into the delay that separates one echo from the next one.

By matching the model input reflectance phase with the actual one, by means of the insertion of a suitable length of line, it is possible to obtain a complete model of the discontinuity that caused the first echo.

This model can be de-embedded from the initial measurements, and the procedure started again on the first residual echo.

The effects of the discontinuity de-embedding may be easily checked by examining the residual time domain response; the errors, due to the approximations implied by this procedure, may be checked in the frequency domain by examining the effect of de-embedding on the input reflectance of the gated discontinuity; we did not experience any difficulty in obtaining residual errors lower than -40 dB.

DISCONTINUITY MODELING

As said before, by means of the gating facilities of the network analyzer, it is possible to isolate the first discontinuity echo, whose *chirp-z* transform represents the input scattering parameter S_{11} of the two-port, that models the discontinuity, and that may be reasonably assumed to be lossless.

Another reasonable assumption is that this model is made up with a lumped part preceded by a length of line, which is needed to match the parameter overall phase shift; some experimental tests carried out on typical microstrip launchers convinced us to limit the lumped part complexity to the second order.

Since the phase shift introduced by the aforementioned line does not influence the magnitude, the latter can be approximated independently, and in particular it is possible to carry out the approximation procedure on the characteristic function F_c which is tied to the squared magnitude of S_{11} by the equation:

$$|F_c(j\omega)|^2 = \frac{|S_{11}(j\omega)|^2}{1 - |S_{11}(j\omega)|^2}$$

This change of the approximation function is quite recommended, since the characteristic function is subjected to less stringent realizability conditions compared to S_{11}.

With the second order assumption, and excluding unusual topologies that are very unlikely apt to model the kind of discontinuity we are dealing with, the mathematical form of the characteristic function squared magnitude is:

$$|F_c(j\omega)|^2 = \frac{A\omega^4 + B\omega^2 + C}{\omega^{2n}}$$

where n determines the function type, that is lowpass, bandpass, or highpass, and A and partially B determine the function degree [11].

The above squared magnitude may be approximated by means of a weighted least squares algorithm, subjected to simple conditions, in order to guarantee that it remains non negative for every ω.

After some simple algebra, the characteristic function is obtained:

$$F_c(s) = \frac{as^2 + bs + c}{s^n} = \frac{h(s)}{f(s)}$$

and the characteristic equation may be solved:

$$g(s) = \text{Hurwitz part of } [h(s)h(-s) + f(s)f(-s)]$$

and eventually the whole scattering matrix of the lumped part of the discontinuity two-port model is obtained. At this point the phase shift of the lumped part of S_{11} is compared with the gated reflectance phase, and matched with a length of line that introduces a delay τ.

Once we have the complete model of the discontinuity, we can de-embed its effects by means of well known circuit theory procedures that use the transmission matrix of the model: in details, all this corresponds to computing the inverse of the transmission matrix of the discontinuity model, and to compute its input reflectance when the output is loaded on the measured reflectance.

EXPERIMENTAL RESULTS

The procedure described in the previous section has been thoroughly tested on simulated data obtained from the analysis of known circuits, and the parameters of the original circuits were constantly obtained again within a precision of a few percent, when the frequency data were limited to just 201 samples between 130 MHz and 23.13 GHz.

Among others we tested a complicated circuit, that included three lumped discontinuities separated by two lines, and exhibited a very bad reflectance never better than – 7 dB on the entire frequency band.

The final residual reflectance after the complete de–embedding was lower than –50 dB up to 10 GHz, while it increased up to –10 dB at the upper frequency end; the residual echo (which should have been ideally zero) was hardly noticeable when plotted together with the original one.

We applied this technique in order to model a user built microstrip fixture, that was intended for measuring packaged MESFETs, figure 1. This fixture may be viewed as a four port circuit: ports 1 and 2 are connected to the network analyzer through the coaxial connectors, while the other ports 3 and 4 are those where the device under test should be connected; the empty fixture exhibits between ports 3 and 4 a capacitive coupling (corresponding to the gap between the line ends) which is highly reflective.

After calibrating the network analyzer to the coaxial ports by means of the usual twelve term calibration, the scattering matrix of the empty fixture was measured.

The time domain response associated with the input reflectance is shown in figure 2, and a similar behaviour is exhibited at the output port; the launcher echo is clearly visible, while the second large echo is produced by the capacitive gap; multiple reflections are responsible of the ripples that appear especially after the gap echo.

The first echo is gated out by means of the gating facility of the network analyzer, and modeled with the procedure described above; a first attempt to model the launcher with a first order low pass function matched the discontinuity reflectance magnitude in a way that was not considered sufficiently satisfactory, while a second order low pass function was found adequate to match the reflectance magnitude with acceptable accuracy; after this a line length was determined in order to match the total reflectance phase. The Smith chart plot of the actual and approximated

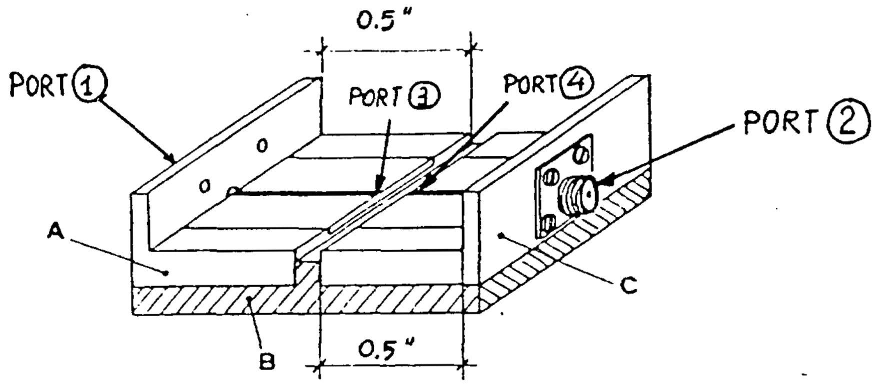

Figure 1: Sketch of the microstrip fixture

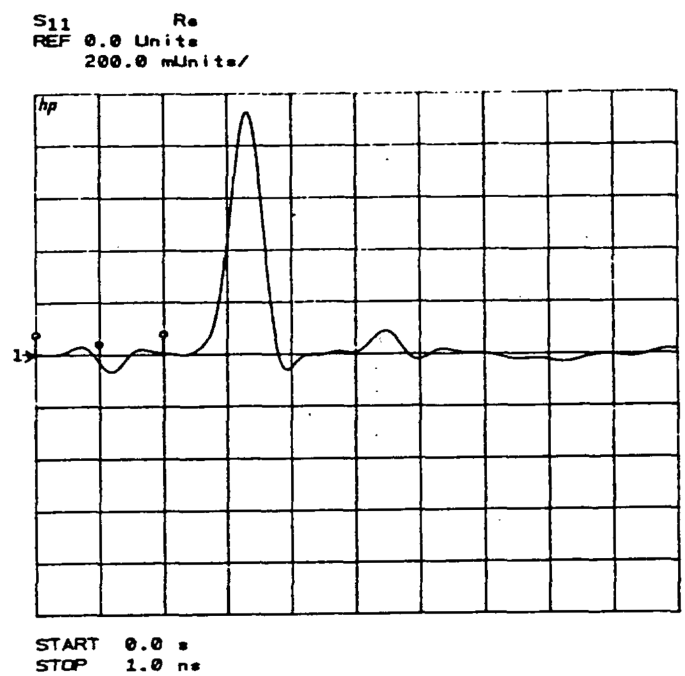

Figure 2: Time domain response at port 1.

reflectance of the first discontinuity is shown in figure 3.

As a by product of this technique a circuit model of the microstrip launcher can be synthesized, as it is shown in figure 4.

The transmission matrix of the launcher model can be now used to de–embed the launcher effects from the fixture scattering parameters; at the input side the residual reflectance corresponds to a time domain response as shown in figure 5, where the echo of the first discontinuity is almost completely canceled, and the gap echo is shifted to the left by the amount of the de–embedded delay.

Similar steps can be performed on the output side of the fixture. Repeating the whole procedure eventually the entire fixture can be adequately modeled, figure 6.

For what concerns the highly reflective gap discontinuity it must be noticed that numerical problems may arise, which may reduce the approximation accuracy. Nevertheless this fact is assumed to be of less importance, since the significant point is to determine the last line length in order to shift the reference planes to ports 3 and 4; in facts the gap discontinuity is external to the fixture model because it will be substituted by the device under test.

For this reason we are developing a different procedure in order to approximate the final internal discontinuity by generating its whole scattering matrix, instead of working with just one reflectance.

CONCLUSION

We presented a technique suitable for modeling microstrip fixtures in order to allow to de–embed the effects of the discontinuities, and to shift the reference planes of an ideal reflectometer to the physical ports of the device under test. The procedure allows to calibrate the network analyzer to these ports without using microstrip components as standard loads, and without assembling/disassembling the measuring fixture.

The results that we obtained are very satisfactory, and the procedure seems to be applicable to other kinds of fixtures, also those that are built directly by the user for his specific applications.

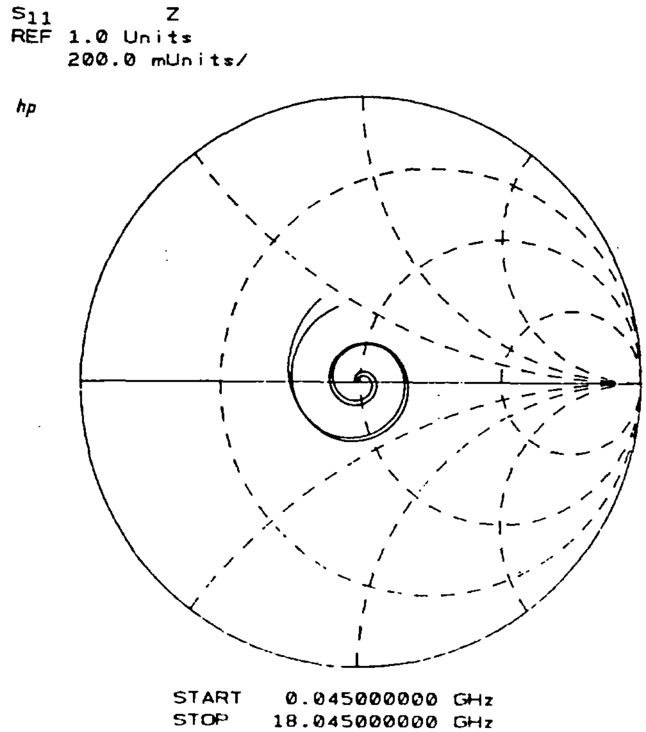

Figure 3: Smith chart plot of the gated and the modeled reflectance.

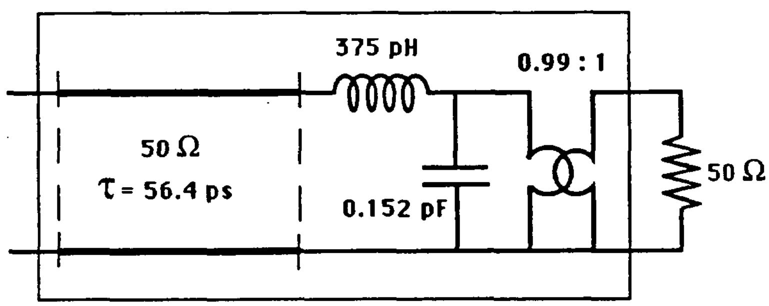

Figure 4: Circuit model of the launcher.

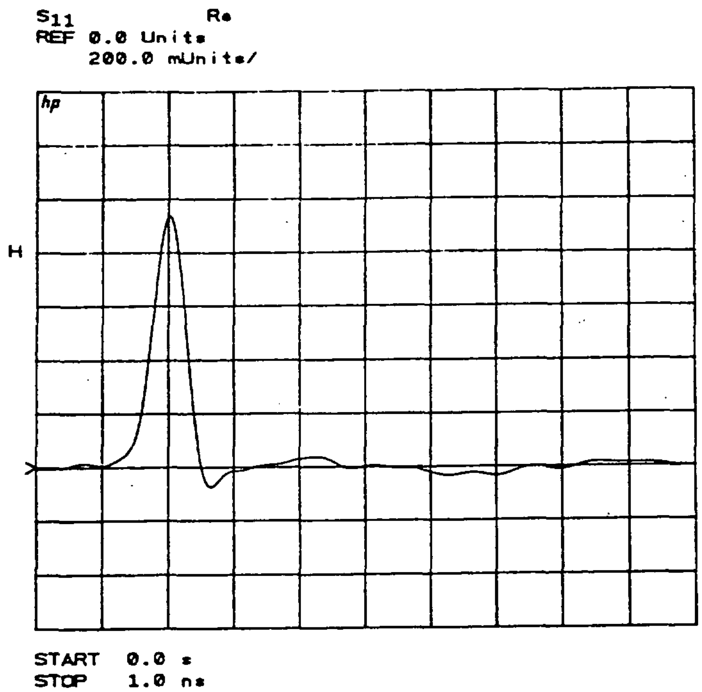

Figure 5: Time domain response without the effects of the launcher.

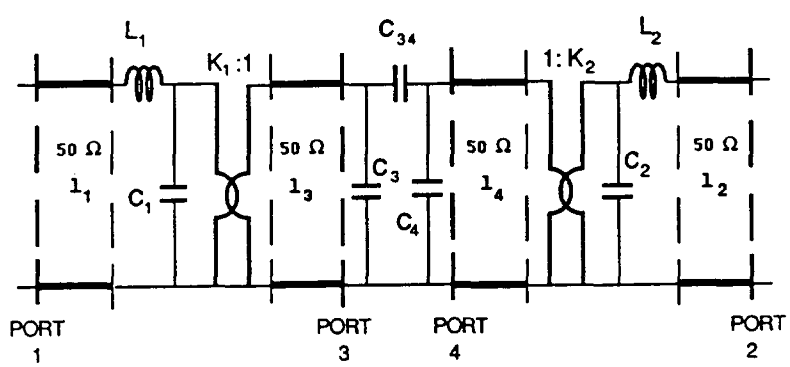

Figure 6: Four port model of the fixture.

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