LRM and LRRM CALIBRATIONS WITH **AUTOMATIC** DETERMINATION OF LOAD INDUCTANCE

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Abstract

Two new techniques are presented in an effort to achieve greater accuracy and better repeatability in the on-wafer calibration of vector network analyzers. The first is a method of determining an inductance value for the match standard during the calibration process so that only its resistance, not its reactance. needs to be known. The second is a new type of calibration. LRRM (line-reflect-reflect-match), which is a variation of LRM with several possible advantages. Also. a simple series resistance-inductance model for a coplanar load IS experimentally investigated to 40 GHz and found to provide a good description of the load behavior.

Introduction

When performing an LRM calibration'. 2. 3 at the wafer level with small (50 micron) coplanar resistors as impedance standards the goal is to obtain a known, frequency independent reference impedance. This requires that an accurate model of the match standard be known.

For coplanar 50-ohm loads it will be shown that a simple series resistance-inductance circuit appears to be a good model up to 40 GHz. The inductance accounts for the overlap-dependent parasitic reactances of the load, and thus is not easily calculated from first principles.

A new method will be presented for determining this inductance during the calibration process without making additional measurements. This inductance can then be employed. along with a frequency independent resistance, to complete the electrical description of the load, allowing recalculation of the error coefficients. Thus only the resistance of the load needs to be precisely defined. so that the amount of overlap on the match standard becomes Unimportant.

A variation on the LRM technique with several possible advantages. line-reflect-reflect-match (LRRM0. will be described.

Measurements for this work were made with an HP 8510 network analyzer, Cascade 150jrm pitch GSG' probes, and a Cascade LRM impedance standard substrate (ISS). Figure 1 shows the geometry of the ISS standards. ,

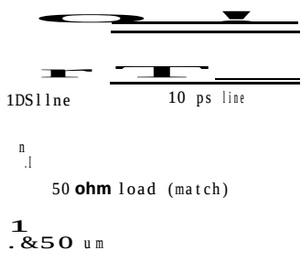


Fig. 1 ISS standards geometry

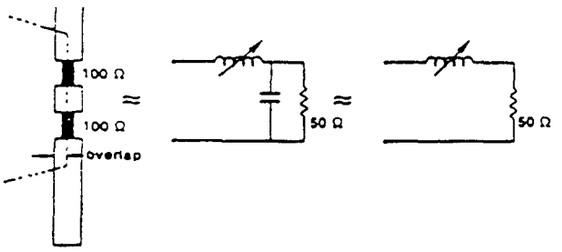
Reference reactance and the load

Present implementations of the LRM calibration initially solve an error model that leaves reference planes for both ports at the middle of the line. Thus the 'error adapter" for each port includes one half of the line along with the transition at the probe tip. The reference planes can be moved to each probe tip according to a physical model for the thru standard. and the resulting error adapters will still include the desired probe tip transition.

Reference 4 describes the effects a change in overlap has on measurements. As the overlap increases. the amount of series inductance seen by the probe decreases. while the shunt capacitance at the transition remains largely unchanged. This combination of capacitance and variable inductance at the probe tip constitutes what may be described as a **reference** ***reactance.'***

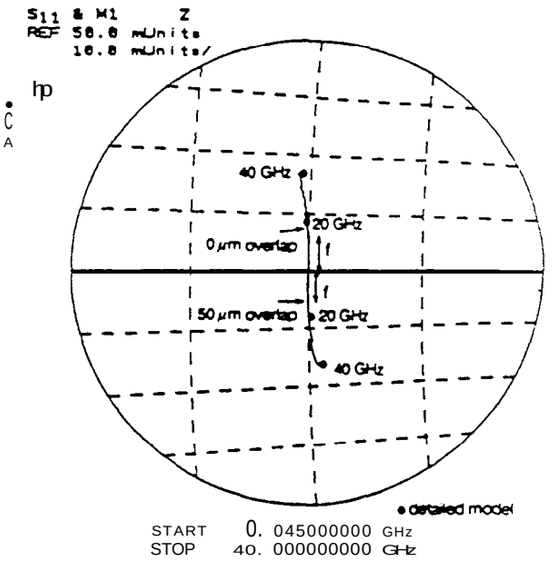
Ideally. DUTs and calibration standards should both be contacted so that they have identical reference reactances. This is most easily accomplished by using DUTs with pads similar to the coplanar lines of the thru standard and contacting them with identical overlaps. thus ensuring proper removal of the transition.

Evaluation of coplanar load For LRM calibrations, the reference impedance is determined solely by the definition of the match standard, therefore it is important that the behavior of the match (load) be understood. Figure 2 shows **a** simplified equivalent circuit of a coplanar 50 ohm load. As overlap varies, the capacitance across the resistor stays nearly constant, while the inductance changes significantly due to the change in the length of the current path. Because the resistor is 50 ohms and wC and wL are both small, the capacitance will behave to first order as negative series inductance. Thus the capacitance may be accounted for by the inductor, leaving simply a variable inductor in series with the resistor. The value of this inductance includes differences between the reference reactances of the load and the thru standards.



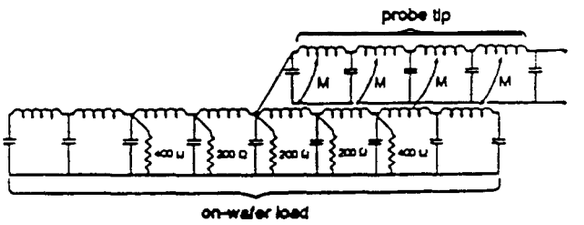
Flg. 2 Simple coplanar load model

The appropriateness of this model is demonstrated in figure 3. Here a load was contacted with a 25 micron overlap as an LRM calibration standard, and was defined lo have zero load inductance. It was then measured with 0 and 50 micron overlap. The primary effect is a change in the reactance, as would be suggested by the above simple model. Thus. a variable inductor provides a reasonable description of the effects of a varying overlap. The small changes in resistance are only 0.25 ohms for each case, or 0.01 ohms per micron at 40 GHz.



Flg. 3 Change In load Impedance wlth overlap

A more detailed model was constructed in hopes of understanding the small change in resistance. This model accounted for the distributed nature of the load, as well as the coupling between the signal line of the probe and the pads beneath the probe. The model is seen in figure 4 for the case of 25 microns overlap. Similar models were used for 0 and 50 microns of overlap.



Flg. 4 **Detalled model of probe contacting load**

To obtain the qualitative behavior of this model. all element values were estimated, except the mutual inductances. Mutual inductances were chosen so that reactances of the 0 and 50 micron cases agreed with the measurement. The final results (fig. 3) were computed with EESOF's Touchstone software and show a change in resistance similar to that seen it-i the measurements, suggesting that this effect can be attributed to the distributed nature of the load and the beneath-the-probe coupling.

The above tests say nothing about the absolute value of the load resistance, only its change with overlap and frequency. To gain confidence that its resistance is relatively constant with frequency, a load was measured with an LRL calibration implemented on the HP 8510 network analyzer, using a 1 picosecond line and a 10 picosecond line (fig. 5) on the LRM impedance standard substrate.

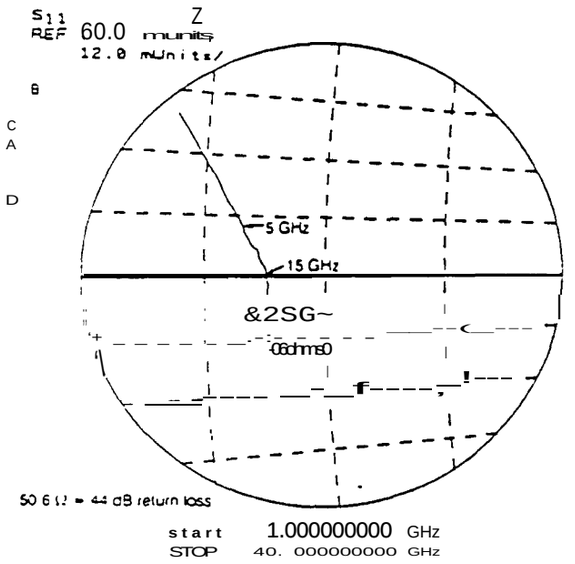


Fig. % Load measured wlth TRL calibration

The anomalous low-frequency behavior of figure 5 is caused by the effects of losses on the line impedance" that were not accounted for in the TRL calibration, and these effects may have been aggravated by a line standard too short for frequencies below about %GHz. The high-frequency line impedance deviated from the load resistance by about 1.5 ohms.

Disregarding the low frequency effects, the change in resistance of the load relative to the Z_0 of the line standard is about 0.6 ohms. For a deviation from 50 ohms this would correspond to a 44 dB return loss. This small change with frequency between the load resistance and the line impedance suggests that either both are nearly constant or that both vary in a similar manner. The former explanation seems more likely since it is improbable that completely different structures would behave similarily.

LRRM

A new and slightly different version of the LRM method. the LRRM (line-reflect-reflect-match) method uses a line. two unknown reflects (e.g.. both undefined shorts and undefined opens), and a match. The match needs to be measured on only one Of the two ports, and as with the LRM technique it actually needn't be a perfect match at all-it simply needs to be an arbitrary but defined one-port measurement. The LRM method can be viewed as being the special case of the LRRM method for which the match is chosen as one of the reflect standards.

Three possible advantages arise from the LRRM method. Unless the same load is used for both pons of an LRM calibration, any small discrepancies in resistance will result in slight inconsistencies in the resulting calibration. Misalignment between resistive material and metal pattern also could result in asymmetric loads. such that the port 1 probe might see a different reactance than the port 2 probe. even though they had the same overlap. Using a load on only one pon avoids the above two potential problems with LRM. Lastly, advantage may arise from using both the open and short for reflect standards. since they are the most widely separated on the Smith chart.

Calibrations with the LRRM technique produce results nearly identical to carefully performed traditional LRM calibrations. A new Cascade program performs both LRM and LRRM calibrations. Raw data is read from the network analyzer, then error coefficients are computed and stored into the network analyzer memory. Measurements with each calibration technique were made on a 40 picosecond line. Figure 6 shows the reflection coefficient of the line measured as an open stub. as well as the transmission through the line for each calibration. The results are essentially identical. verifying the LRRM procedure. The phases measured with LRM and LRRM also agreed.

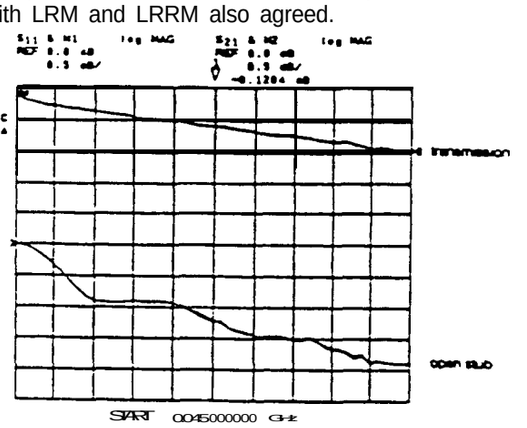


Fig. 6 Comparison of LRM and LRRM cailbratlons-A 40 ps line measured as a two-port (S21) and as a one-port open stub

An additional verification of the LRRM calibration is seen in figure 7. The match calibration standard was measured on port 1 only, and assumed to be perfect. Shown is a plot of the same 50 ohm load measured on port 2 after de-embedding. This result shows the reference impedance for the port 2 reflection measurements is essentially the same as that for the pon 1 measurements, even though the impedance standard was only measured on port 1 during calibration. The L-R-R measurements have successfully “transferred” the reference impedance to port 2.

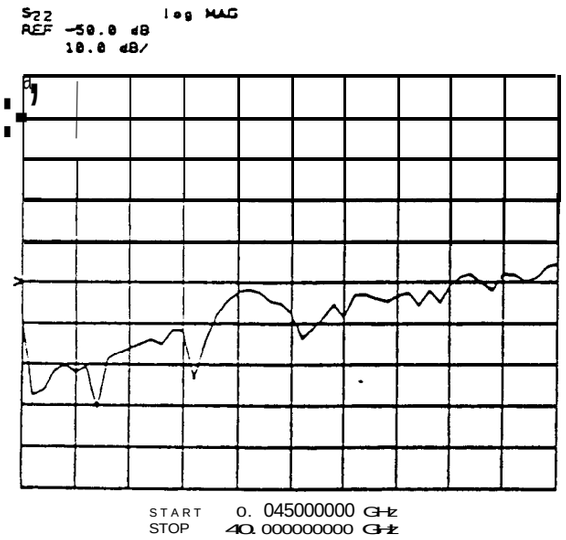


Fig. 7 Load measured on port 2. Calibration was LRRM with the load standard measured on port 1 only, and deflned to be a perfect match.

Determination of load inductance

Reference (1) describes how errors are revealed in the resulting calibration if an LRRM calibration is performed where the match is assumed to have zero reactance but actually doesn’t. For calibration on a 1 picosecond thru. if the load standard actually had a negative reactance, then the resulting de-embedded probes-in-air open will appear to have a negative conductance. not a conductance of zero. This can be corrected by two methods; either the overlap on the load is changed in subsequent calibrations until an open of zero conductance is achieved, or the load inductance is determined and its value incorporated into the match standard definition for recalculation of the error coefficients. The latter approach is more desirable. requiring no additional measurements if a suitable open was used for one of the reflect measurements

The Cascade program uses this latter approach, which “automatically” accounts for load inductance. It reads the raw calibration measurements from an HP 8510 network analyzer and makes an initial determination of error coefficients, assuming the load to have no inductance. It then determines the resulting de-embedded value of open admittance, from which it calculates the load inductance definition which would have produced an open conductance of zero. **The** values obtained for load inductance are averaged over all frequencies. The result is then used with the load resistance (assumed frequency independent) to recalculate the error coefficients, which are then stored into the HP 8510 network analyzer.

Determination of load Inductance Of the seven independent equations that are solved during two-pon de-embedding (ignoring isolation and switch terminations), six are generated by the L-R-R measurements. The seventh is generated by the match. For the case where the match is measured on pon 1 and the error adapters are described with ABCD parameters, it happens that the ratio C/D = C’ of the pon 1 adapter is directly proportional to the defined admittance of the match standard.

$$C_a = Y_{def}^{match}$$

It also happens that the measured (or de-embedded) value of a one-pon DUT is directly proponional to C’:

$$Y_{meas}^{DUT} = a\,C$$

Thus if the match is improperly defined by the factor

$$\frac{dY_{def}^{match}}{Y_{actual}^{match}}$$

then C’ and hence $Y^{dut}/_{meas}$ will be wrong by the same factor:

$$Y_{meas}^{dut} = \frac{Y_{def}^{match}}{Y_{actual}^{match}} \cdot Y_{actual}^{DUT} \qquad (1)$$

This can be viewed as a distortion of the Smith chart as illustrated in reference (2).

For the case where the match is defined to be a pure resistance

$$Y_{def}^{match} = \frac{1}{R + j0}$$

but physically has an admittance,

$$Y_{actual}^{match} = \frac{1}{R + j\omega L}$$

then for the open admittance, from (1),

$$Y_{meas}^{open} = G_{meas} + jB_{meas} = \left(\frac{R + j\omega L}{R}\right) Y_{actual}^{open} \quad (2)$$

Now, if one assumes for the open,

$$Y_{actual}^{open} = 0 + jB_{actual}$$

then the match inductance can be found from (2):

$$L = \frac{-G_{meas} \cdot R}{B_{meas} \cdot \omega} \quad (3)$$

Thus the correct match inductance has been found from the initially measured (erroneous) admittance of the open.

The previous method will give a value of inductance at individual frequencies and can be outlined as follows:

1. Assume an initial value for the match inductance (L = 0) and calculate error terms.
2. Calculate the de-embedded admittance of the open.
3. Calculate the load inductance from equation (3), which would have produced an open with zero conductance.
4. Recalculate the error terms based on the new value for load inductance.

For a sweep over multiple frequencies, the values for inductance may be averaged before doing step 4. A weighted average is desirable due to the sensitivity to errors at lower frequencies (B_{meas} approaches zero in equation [3]). For the present work, frequencies below 10 GHz were weighted with factor zero, and frequencies above 10 GHz were weighted proportionately with frequency.

Figure 8 shows the result of the above procedure. A set of measurements were made for an LRRM calibration and the load was purposefully contacted with excessive overlap, making it look slightly capacitive. The top trace resulted from a definition of zero load inductance, the bottom trace resulted from the above procedure and represents a conductance of zero as assumed, thus verifying the procedure. Both calibrations used the same calibration standard measurements.

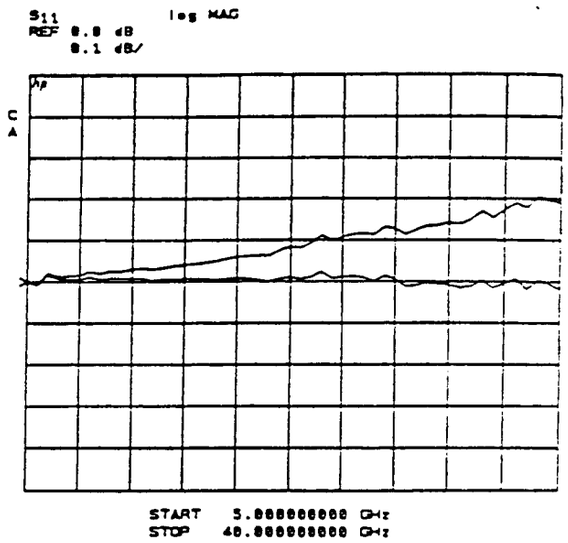


Fig. 8 Open reflection coefficient, with and without automatic accounting for load inductance.

The load inductance determined during the calibration was −11.5 pH. This corresponds to an error in the load definition of j2.9 ohms at 40 GHz for the case where the load was assumed to have zero inductance. This error was avoided with the new procedure.

Magnitude of open reflectton coefflclent The question arises as lo what the open conductance should be-does radiation loss, dielectric loss, or some other mechanism affect it? The LRM method cannot be used to answer this because the load reactance is not (easity) known from physical principles.

LRL standards, on the other hand, have no unknown quantities such as the load reactance that would seriously affect the open conductance. Hence the open conductance can be measured directly with an LRL calibration. This was done using the 1 ps line and 10 ps line on the LRM ISS **as** standards. Because the same 1 ps fine is used for both the LRL and LRM calibrations. the open (probes in air) is an identical DUT for each calibration. Great care was taken to use equal overlaps on each line. Figure 9 shows the result. indicating that zero conductance is a good assumption through 40 GHz.

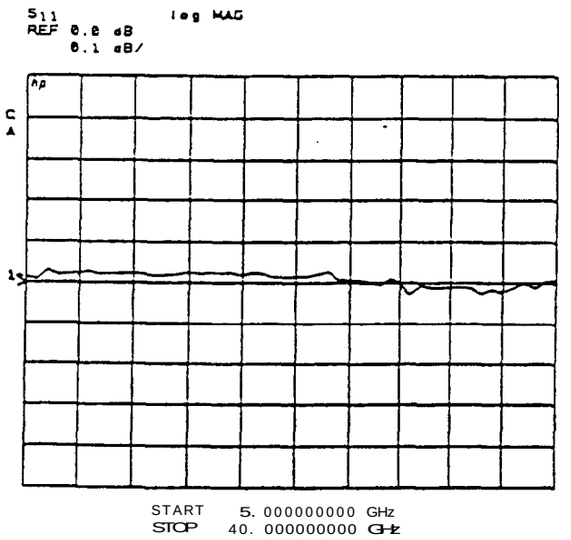


Fig. 9 Open reflection coefficient measured *with* LRL calibration.

Conclusion

A simple series resistance-inductance model for a coplanar load was shown to provide a good description of the load through 40 Ghz. Thus the dc resistance of the load can be assumed equal to the RF resistance over this range. An algorithm was presented, which automatically calculates the load's inductance when used as the match standard in an LRM type of calibration. The LRRM calibration technique has several possible advantages and also can be used with the load inductance algorithm.

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