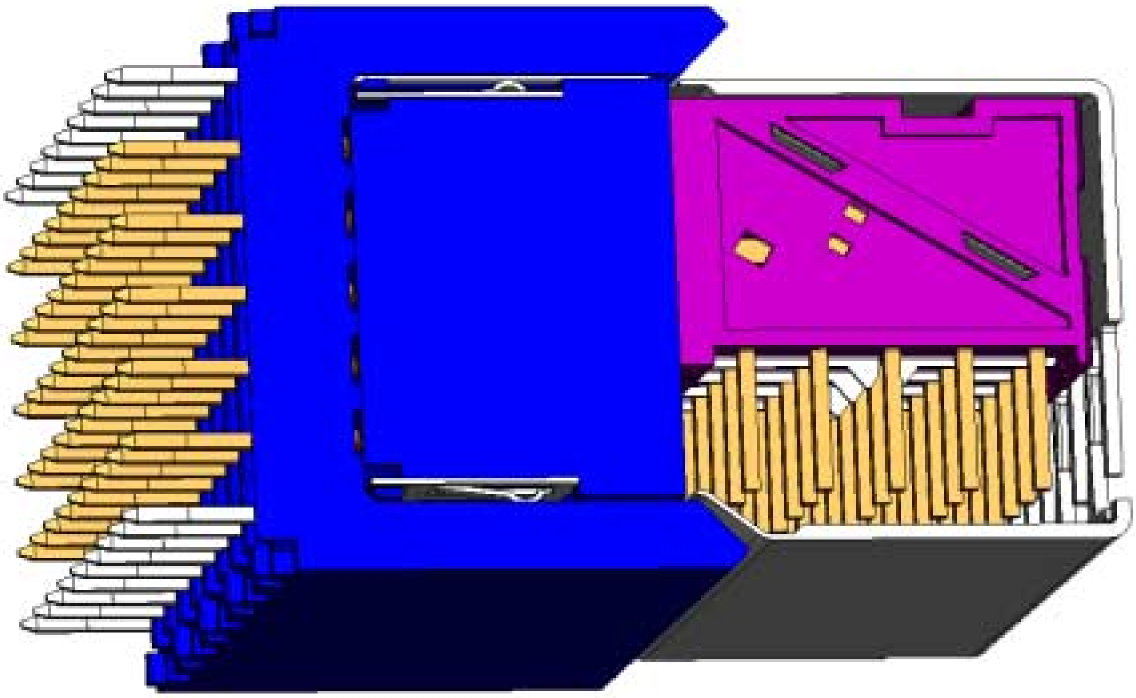
A KYOCERA GROUP COMPANY

Generation of SPICE Simulation Models for ELCO 2mm Hard Metric Connectors



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1 Introduction

The intention of this document is to give help in understanding the ELCO B25 model generation and hardware to model correlation. For this purpose the used terms and definitions like wave impedance are explained at first. The following chapters describe the ELCO B25 model generation from full 3D electromagnetic simulation to the SPICE model generation including an explanation of the SPICE model structure. A further chapter about the test board visualizes the different calibration and test structures which are necessary to verify the accuracy of the SPICE model. A final section "Hardware to model correlation" explains the required measurements and simulations in the time and frequency domain. At last the used software, hardware and literature are shown.

2 Basics

The fundamental criteria's for the use of the ELCO B25 at high data rates are:

- Adaptation to a typical impedance of the signal path to avoid interference by reflections
- Low crosstalk between different signal pins
- Prevention from radiation at high frequencies to the environment
- Insensitivity in opposite to disturbances from outside

For a common basis in discussing the examination results the following terms are explained:

- wave impedance
- S-parameters
- voltage standing wave ratio
- crosstalk
- differential signal transmission

2.1 Wave impedance

The wave impedance and the transfer function of a connector system are crucial for its high frequency behavior. If on the signal path in the connector occurs differences in the wave impedance this would lead at the so called hit points to undesired reflections. This means that one part of a signal with the amplitude V will be reflected and one part will be transferred. These differences in wave impedance are also called mismatch.

2.1.1 Definition of wave impedance

Initially wave impedance is treated in general. According to its definition wave impedance is a complex and frequency dependent value.

$$Z_W = \sqrt{\frac{R'+j\varpi L'}{G'+j\varpi C'}}$$

There are two possibilities to determine wave impedance:

- S-parameters in the frequency domain allows to determine the complex wave impedance after corresponding transformations. However only the complex wave impedance of the entire measurement object is determined. Insights in the wave impedance inside the measurement object are not feasible.

- With the assumption, R' and G' are appropriate small and therefore negligible, the wave impedance will be a real term.

$$Z_W = \sqrt{\frac{L'}{C'}}$$

For connectors this assumption is acceptable at short lengths. Now the real value of the examined object can be determined from the reflection behavior. This method has the advantage that the wave impedance can be displayed as a function of the geometrical position. Thereby an insight in the inner structure of the connector system can be achieved.

2.1.2 Wave impedance as a function of the geometrical position

A voltage source generates a step function with a very steep rising edge. This signal goes through a cable with a wave impedance which is normally about the same as the input impedance of the generator and the measurement object. If there is a change in wave impedance at a certain point, a part of the signal will be reflected.

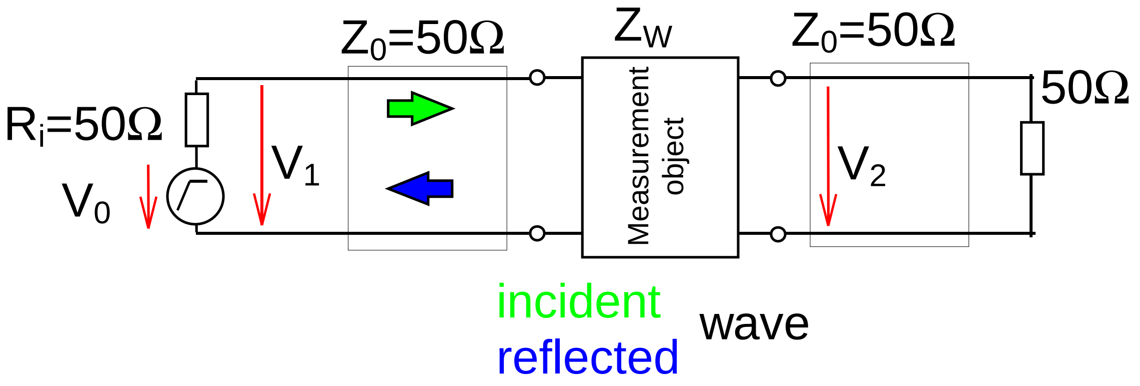


Fig. 1 Principle of a time domain reflectometer

This reflected part overlays the original signal and will be measured as a function of time at the output. The wave impedance as a function of time can be determined from the reflected voltage V_0 (generator voltage V_0 and characteristic internal resistance Z_0 must be known). In a further step, the velocity of propagation of the signal must be known, the wave impedance at a certain position in the connector can be calculated from the wave impedance as a function of time. With this specification – wave impedance as a function of geometrical position – mismatches can be located.

2.1.2.1 Wave impedance as a function of time

This method will be explained in the following equations:

After reflection incident and reflected voltage are added and the summarization of both voltages V_1 can be measured at the generator:

$$V_1 = V_i + V_r \qquad (1)$$

with: V_i = incident wave
V_r = reflected wave

Due to the fact that the inner resistance R_i of the generator and the wave impedance of the connected cable are about the same, they operate as a voltage divider and the amplitude of the incident signal is:

$$V_i = 0.5{*}V_0 \qquad (2)$$

The absolute value of the voltage of the reflected wave is given by the reflection factor of the voltage r_V:

$$V_r = r_v \cdot V_i \qquad (3)$$

This reflection factor itself can be determined from the characteristic wave impedance Z_0 of the measurement system and the wave impedance Z_w of the measurement object. The reflection factor of the voltage at the transition between measurement system and measurement object is:

$$r_V = \frac{Z_W - Z_0}{Z_W + Z_0} \qquad (4)$$

By these four equations the wave impedance of the measured voltage V_1 can be determined after corresponding conversions. Wave impedance is then:

$$Z_W = \frac{V_1 \cdot Z_0}{V_0 - V_1} \quad (5)$$

with: V_1 measured voltage
V_0 source voltage of the generator
Z_0 wave impedance of the measurement system

2.1.2.2 Wave impedance as a function of geometrical position

Due to the fact that an electromagnetic wave has a known finite velocity of propagation, the covered distance of a signal can be determined from a time measurement. If there is a reflection in a distance x from the generator, the time t_x, which the signal needs from the measuring instrument to the place of reflection and back again, is measured. With this known velocity of propagation any reflection can be assigned to a geometrical position.

As normally the insulator of a connector is made up of one specific material only, it can be assumed that the velocity of propagation is determined by the dielectric constant of the insulator.

The place of a reflection is therefore determined by:

$$t_r = 2 \cdot \frac{x_r}{v_{signal}} \quad (6)$$

$$x_r = 0{,}5 \cdot t_r \cdot v_{signal} \quad (7)$$

with $v_{signal} = c_0 \cdot \frac{1}{\sqrt{\varepsilon_r}}$ follows:

$$x_r = 0{,}5 \cdot c_0 \cdot \frac{1}{\sqrt{\varepsilon_r}} \cdot t_r \quad (8)$$

with: c_0 speed of light
ε_r relative dielectric constant of the insulator between inner and outer conductor
x_r place of reflection
t_r measured time of incident and reflected step

2.2 S-Parameter

To describe the behavior of a connector system in frequency domain (signal attenuation, velocity of propagation), parameters derived from the so called “scattered parameters” (S-parameters) are used.
The principle of these S-parameters will be explained on the basis of the following drawing.

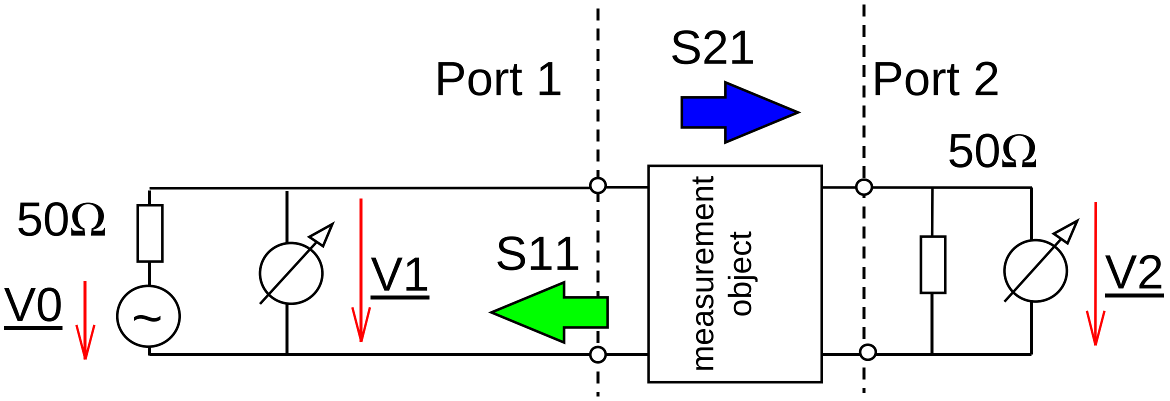


Fig. 2 Principle of S-parameters

At port 1 of the network analyzer a source with 50Ω inner resistance generates a sinusoidal alternating voltage with variable frequency. This electromagnetic wave passes the supply line, with 50Ω wave impedance like the whole measurement system, to the measurement object. If there are reflections in the measurement object caused by differences in wave impedance, one part of the wave will be reflected and another part passes through the measurement object. At port 1 (source) and port 2 (sink or receiver) the magnitude and the phase shift of the reflected and transmitted part of the electromagnetic wave are measured.

These measurement values are called “scattered parameters” or S-parameters in short form. The indices indicates the direction of the S-parameters. In the sketch above only the transmission behavior is plotted, if port 1 is the source and port 2 the sink. Normally the source and the sink are changed at the measurement once again. The results are 4 different S-parameters. The meaning is explained in the following Tab. 1.

Ind.	Meaning	source	sink
S11	reflected part at Port 1	Port 1	Port 2
S21	transmitted part from port 1 to port 2	Port 1	Port 2
S12	transmitted part from port 2 to port 1	Port 2	Port 1
S22	reflected part at Port 2	Port 2	Port 1

Tab. 1 Description of S-parameter

2.3 Voltage standing wave ratio

Another parameter which is used to characterize a connector system is the voltage standing wave ratio. This means: One part of the incident electromagnetic wave will be reflected at a hit point between two wave impedances. This reflected part interferes with the incident wave.
This means, that the incident and reflected wave cancels out each other partially and standing waves are formed. This process is identical with the known interference from the optics (laser) or mechanics (vibrations). Fig. 3 shows 3 different types of standing waves on a lossless conductor.

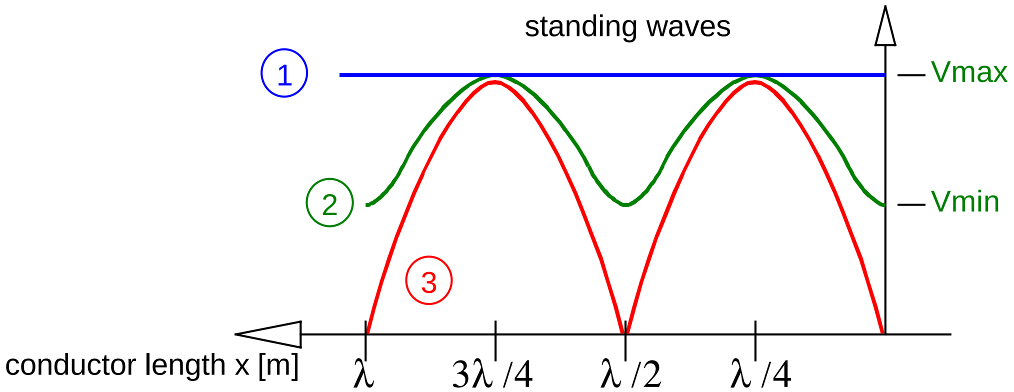


Fig. 3 Standing waves on a lossless conductor

- curve (1) shows ideal case (no standing waves are formed). The reflection factor is 0.
- curve (2) shows standing waves at a partially reflection. The reflection factor in this case is e.g. 0.5.
- curve (3) shows standing waves at a total reflection, with an open end. The refl. factor in this case is 1.

Now the term „voltage standing wave ratio" is introduced. The common abbreviation is the term VSWR. VSWR is the ratio between the maximum and minimum voltage of standing waves.

$$VSWR = \frac{V_{\max}}{V_{\min}} \qquad (9)$$

In Fig. 3 V_{max} and V_{min} for curve (2) is inserted. The three curves in Tab. 2 would result in the following values for VSWR:

Curve	V_{max}	V_{min}	VSWR
1	1	1	1
2	1	0.5	2
3	1	0	∞

Tab. 2 Voltage standing wave ratio

VSWR=1 would be the ideal case for a transmission system because there are no losses by reflections.
The higher VSWR gets, the more reflections occur and the transmission system (connector) gets worse.

To calculate VSWR, equation (9) is normally not used, but the absolute value of the measured reflection factor:

$$\mathrm{VSWR} = \frac{1+|r|}{1-|r|} \qquad (10)$$

Whereby the absolute value of the reflection factor is equal to the absolute value of S11 from a measurement with a network analyzer or a simulation.

2.4 Crosstalk

Crosstalk is differed according to the so called near end and far end crosstalk.
The principle of both types of crosstalk is explained on the basis of the following drawing [1]:

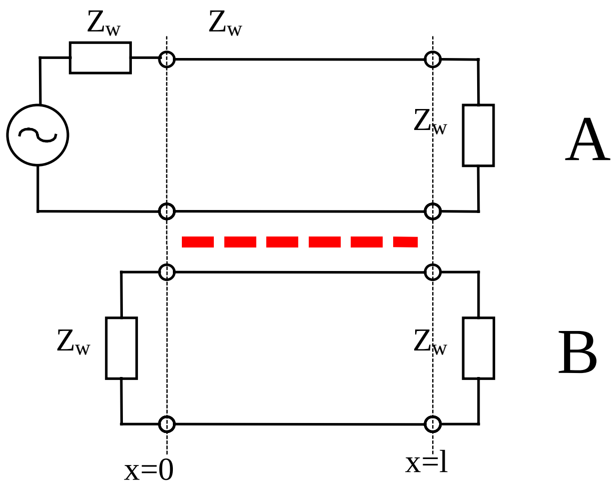


Fig. 4 Crosstalk between conductor A and conductor B

Two conductors are coupled over the length l. For a better understanding it is assumed that both conductors have the same wave impedance Z_w and each is terminated correctly with Z_w.
We look at the coupling mechanism between both conductors, when a signal in conductor A affects conductor B. A distinction is drawn between capacitive and inductive coupling. Both types of coupling and the consequential currents and voltages are shown in Fig. 5.

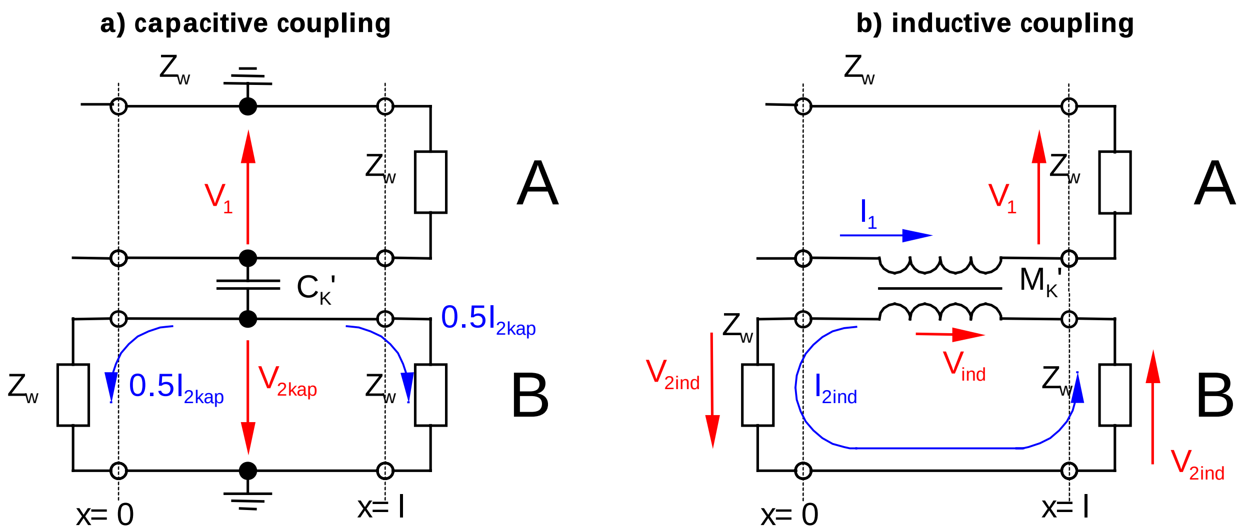


Fig. 5 Coupling mechanism

The capacitive coupling generates a current, which divides itself in two equal parts and flows off over the terminating resistance at x=0 and x=l. The inductive coupling generates a current, which counteracts the induced voltage. Overlaying both effects shows, that the current and voltage at point x=0 and x=l do not have the same size. For the voltage relationship at these both points results:

x=0:

$$\frac{V_2(0)}{V_1(0)} = \frac{1}{2} j\omega \left(Z_w C_K' + \frac{M_K'}{Z_w} \right) \frac{1-e^{-2\gamma l}}{2\gamma} \tag{18}$$

x=l:

$$\frac{V_2(l)}{V_1(l)} = \frac{1}{2} j\omega \left(Z_w C_K' - \frac{M_K'}{Z_w} \right) l \tag{19}$$

with M_K' inductive coupling
C_K' capacitive coupling

The crosstalk at point x=0 is called near end crosstalk and at point x=l far end crosstalk.

Additional to both types of crosstalk:

- Far end crosstalk disappears, if $C_K' = \frac{M_K'}{Z_W^2}$. At short conductor lengths far end crosstalk is at first very small but rises with growing l (see equation 19).
- Near end crosstalk reaches a boundary value unequal zero for growing l (see equation 18).

Normally crosstalk is not given as linear voltage relationship like in equation (18) and (19), but as attenuation. Because both conductors have the same wave impedance, crosstalk attenuation is:

$$crosstalk = 20 \log \left| \frac{V_1}{V_2} \right| dB \tag{20}$$

2.5 Differential signal transmission

Differential signal lines have the advantage in higher disturbance resistance and a definite return path for the current especially at high frequencies. It is advantageous that disturbances which are induced in both differential signal conductors can be eliminated later at the receiver. These chapters are an extension to the chapters 2.1 wave impedance and 2.2 S-parameters.

2.5.1 Wave impedance

There is made a difference between a) single ended conductors and b) differential ended conductors (odd mode) at all performed measurements. For a complete description the wave impedance at even mode will be explained too.

2.5.1.1 Impedance of a single ended configuration

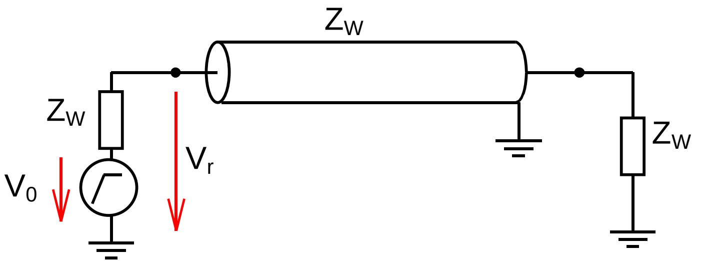


Fig. 6 Impedance single ended configuration

At this measurement only the impedance of a single conductor is measured. This is comparable to measurements e.g. with normal coaxial connectors, which have only one inner conductor and one outer conductor.

2.5.1.2 Impedance of a differential conductor configuration in odd mode

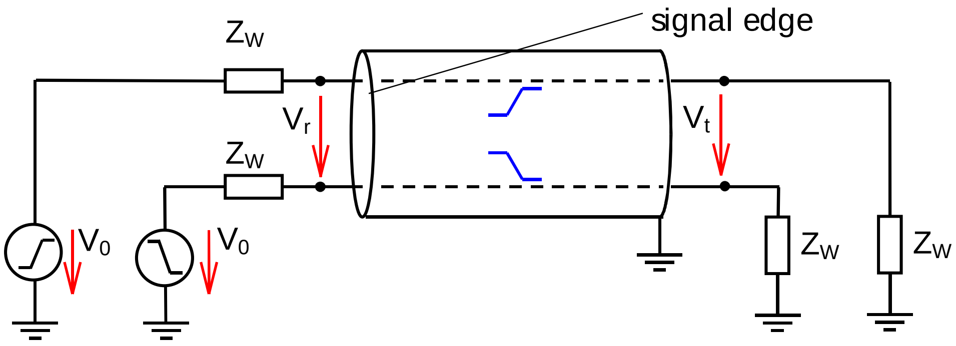


Fig. 7 Differential signal transmission in odd mode

The configuration above describes the principle of differential signal transmissions, like it is used at high speed data transmission systems. The signal source V_0 is positioned between each inner conductor of the cable to ground. The connection to ground at the output is realized with a 50 Ohm resistor. Thereby the current flows through a conductor to the terminating resistance and back through the other resistor. The signal edges in this differential configuration are therefore contrary. Thereby a very powerful electric field between both inner conductor is formed and is different from the wave impedance in even mode. The advantage compared to the single ended configuration is, that there is a defined path for the back flowing current. The result is a significant lower fault liability for data transmissions, especially over long distances and with powerful electromagnetic disturbance sources.

2.5.1.3 Impedance of a differential conductor configuration in even mode

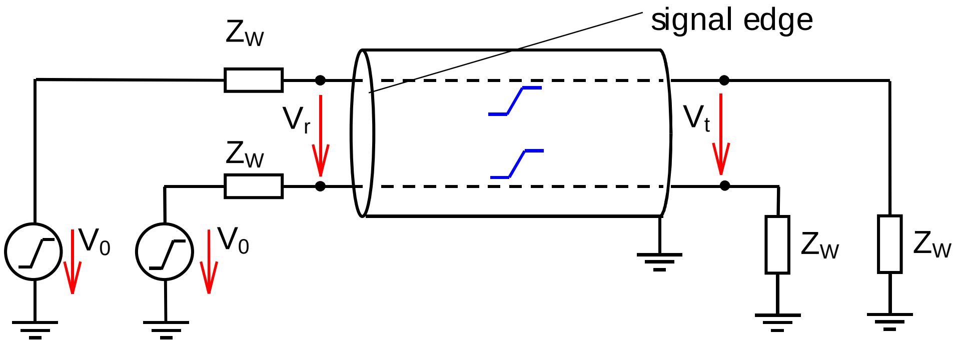


Fig. 8 Impedance in even mode (common mode)

It is spoken of even mode because the signal edges in both conductors have the same rising direction.
Thereby the electrical field between both conductors is getting very low and the measured wave impedance describes the influence of the outer conductor (shielding) on both inner conductors. At even mode the advantages at odd mode are getting lost, because both signals don't differ from each other. On this account even mode makes no sense in practice. In reality a real odd mode cannot be realized. Because e.g. both conductors don't have exactly the same length, a phase shift is produced. This phase shift causes that for a short moment even mode instead of odd mode exists.

2.5.2 *Differential S-parameters in odd mode*

Regarding to chapter 2.2 the S-parameter basics are expanded with the differential mode.
Fig. 9 displays a S-parameter simulation circuit for a differential line. The electromagnetic wave is injected by Port1 and Port2 and divided with the power splitter in two identical signals, with half the power of the original injected signal. One of both signals is inverted in phase by 180 degrees by a phase shifter (the upper red sine wave visualizes the in phase signal and the lower red sine wave the opposite phase signal). The signal is then connected via a transmission line to the measurement object.

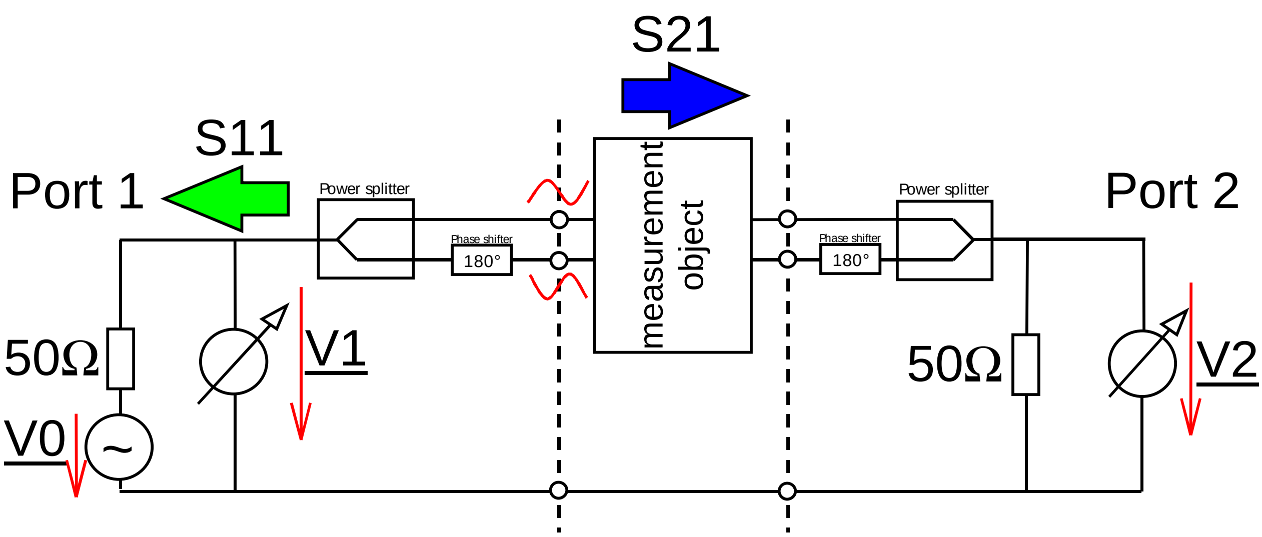


Fig. 9 Circuit for differential S-parameter simulations

3 Model generation

This chapter describes the way from the S-parameter simulation in a fully 3D electromagnetic (EM) simulation program to the SPICE model and the structure of the SPICE model.

3.1 Extraction Flow

In a first step, the 3D geometry information of the connector is imported from a CAD environment to the 3D EM calculation program Microwave Studio from CST to calculate the frequency domain behavior of the ELCO connector. Waveguide ports are applied to the connector pins in the Microwave Studio, similar to network analyzer measurements, before starting the S-parameter simulation. From these S-parameters a SPICE compatible model can be derived and exported. This SPICE model can now be used to simulate circuits including drivers, resistors, .. in Agilent's ADS, which is especially designed for high frequency applications. The following picture shows the correlation with the different steps graphically [2].

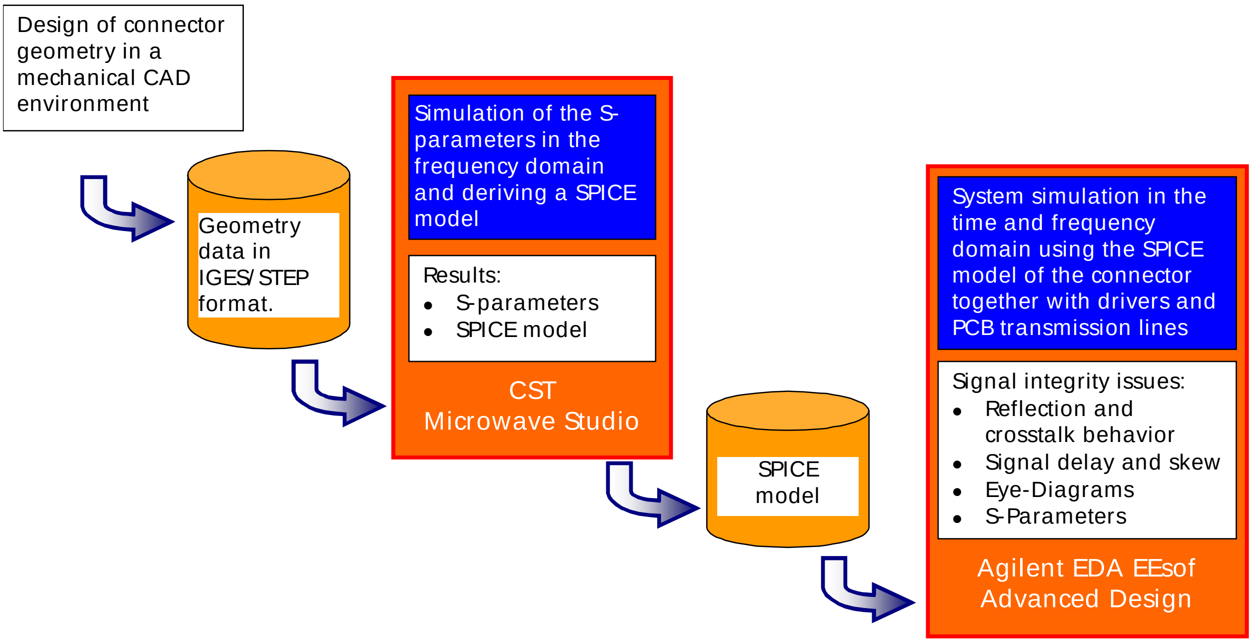


Fig. 10 Model generation flow

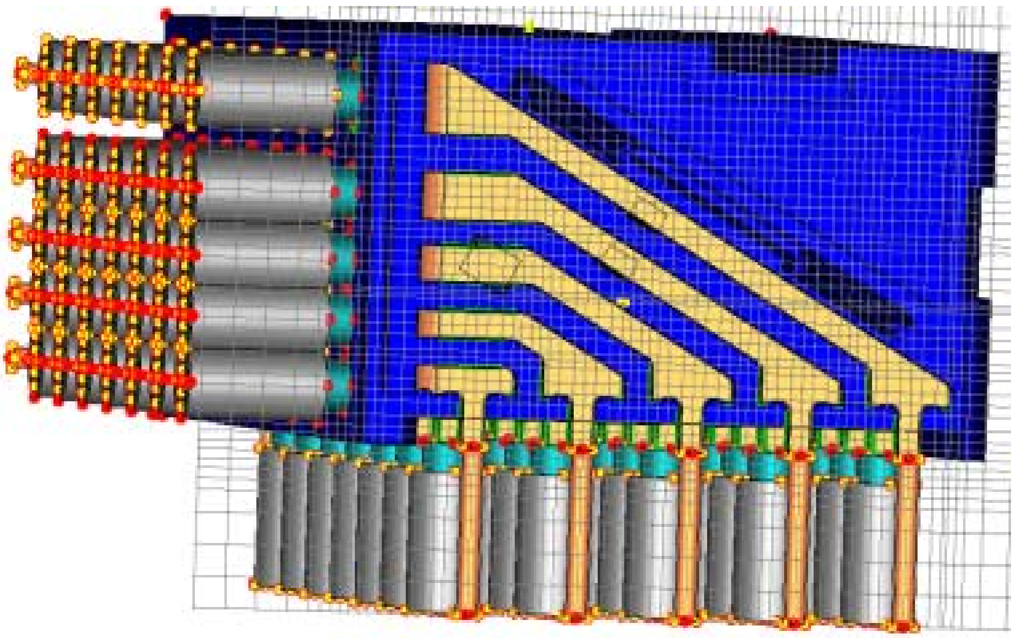
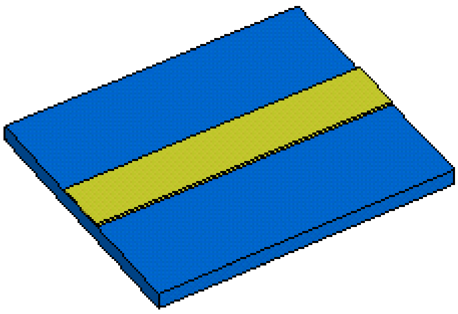


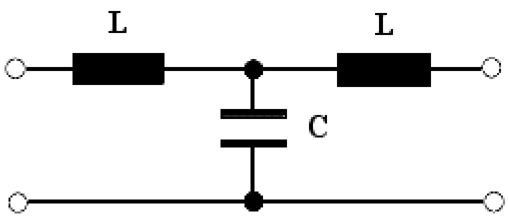
Fig. 11 3D model of the ELCO B25 female connector

3.2 SPICE Parameter Extraction

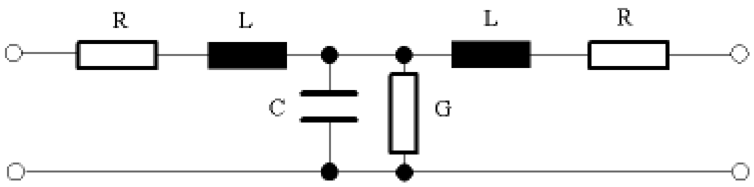
The SPICE parameter extraction in the 3D EM field solver Microwave Studio from CST is used to derive a SPICE compatible network model consisting of lumped R, L, C, G, K elements from previously calculated S-parameters [3].

The parameter extraction is best suited for a network of coupled transmission lines as it occurs in connectors, IC packages, stripline networks, etc. Based on a fixed topology of the model, the lumped element parameters are adjusted such that the lumped element network response matches closely with the response of the distributed system.
The following page explains the topology of the model and how the network parameters are calculated from the S-parameters which are assumed to be calculated previously.
A single transmission line
In a first step let us consider a simple loss free transmission line (here a microstrip line) which can be modeled quite accurately by a T shaped network model, if the transmission line is electrically short (less than a tenth part of a wavelength).



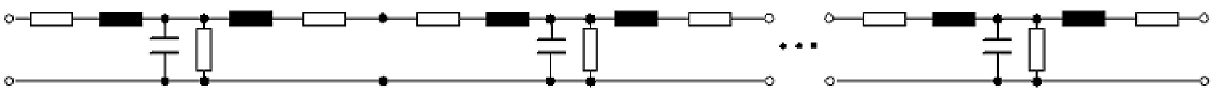


For lossy structures a serial resistance can be added to the inductance and a parallel resistance can be added to the capacity. The resulting model then is as follows:



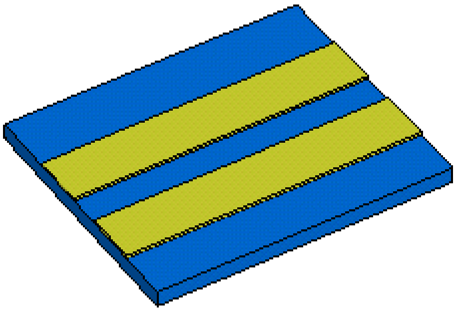
If the transmission line is electrically longer than approximately a tenth part of a wavelength, it is important that the lumped network model can also describe the wave propagation along the lines. In such cases a cascaded transmission line network can be used, where the number of cascades can be specified. A large number of cascades increases the complexity of the model but also improves the accuracy of the approximation. As a general guideline, one cascade for each tenth part of a wavelength should be used. Thus, if the transmission line length is about one wavelength, ten cascades should be used to obtain a sufficient accuracy.

The following picture shows the cascaded transmission line model:

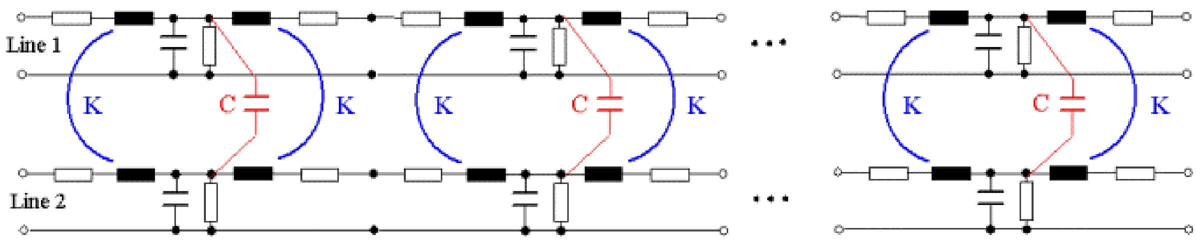


A coupled network of transmission lines

So far only a single transmission line has been considered. In the next step the investigation is extended to a network of two parallel transmission lines (here microstrip lines) as shown in the following picture:



The crosstalk between coupled transmission lines can be described with coupling capacitances (C) and coupled ideal transformers (K). The following picture shows the topology of a coupled network of two transmission lines:



Example cut out for the model topology of a SPICE connector model with more than 2 lines:

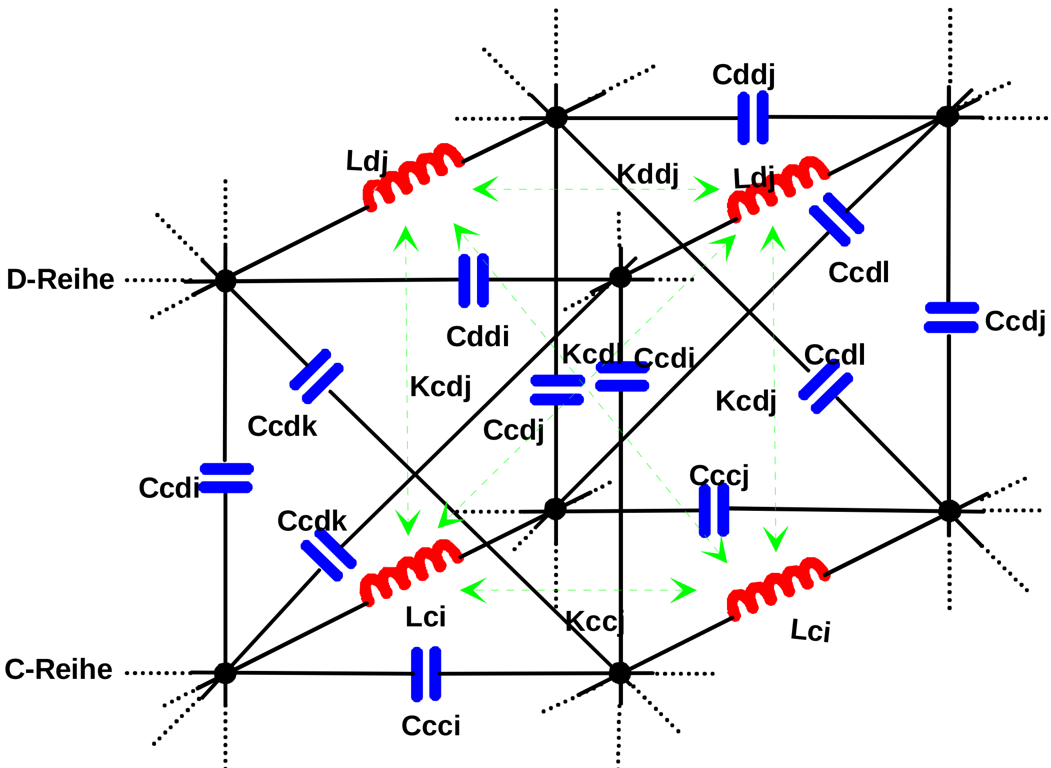


Fig. 12 SPICE model topology [4]

3.3 Modular structure of the SPICE model

In the following an abstract of the main circuit of a simulation model for a connector system is displayed, where the essential components of the structure of a simulation model are explained.

☞ **comment lines start with ' * '**

```
*
* ---------------------------------------------------------------------------
* ELCO 2mm Connector Ground Standard Version
* ---------------------------------------------------------------------------
*
* Version: 1.0
* Date:    11.11.03
*
*
* Parts
* ------
* 27-7200-125-100-001 : 2mm Female Connector 125 Contacts Right Angle Female
* 17-7200-125-100-011 : 2mm Male Connector 125 Contacts Vertical Male
*
* Model layout:
* -------------
*
* multi pair model with crosstalk effects
* 7 columns of 5 signals (see also sketch below)
* 5-segment model, valid for edge rates of 150 ps (10% - 90%) or slower equivalent to f=2GHz
*
* PCB termination
* ---------------
*
* Proper model performance requires that the user add appropriate plated through hole (PTH)
* capacitance to represent the connector interface to the daughter card and backplane. PTH
* capacitance is NOT included in the connector model due to the different possible board
* thickness. A typical values for this PTH capacitance is roughly 1pF per pin.
* For a more accurate results, it is recommended to perform an EM simulation * of the actual
* interface configuration.
*
* Pin assignment
* --------------
*
* The connection nodes of the model are named as follows:
* - the first block of pins is assigned to the simulated signals of the receptacle contact tails
* - the second block of pins is assigned to the simulated signals of the vertical header contact
*   tails
*
*
*
* Usage in a SPICE simulation circuit
* -----------------------------------
*
* The connector subcircuit call statement syntax is:
*
*  x_connector_elco_2mm_standard
*  + f_a1   f_a2   f_a3   f_a4   f_a5   f_a6   f_a7
*  + f_b1   f_b2   f_b3   f_b4   f_b5   f_b6   f_b7
*  + f_c1   f_c2   f_c3   f_c4   f_c5   f_c6   f_c7
*  + f_d1   f_d2   f_d3   f_d4   f_d5   f_d6   f_d7
*  + f_e1   f_e2   f_e3   f_e4   f_e5   f_e6   f_e7
*  +
*  + m_a1   m_a2   m_a3   m_a4   m_a5   m_a6   m_a7
*  + m_b1   m_b2   m_b3   m_b4   m_b5   m_b6   m_b7
*  + m_c1   m_c2   m_c3   m_c4   m_c5   m_c6   m_c7
*  + m_d1   m_d2   m_d3   m_d4   m_d5   m_d6   m_d7
*  + m_e1   m_e2   m_e3   m_e4   m_e5   m_e6   m_e7
*  +
*  + elco_2mm_standard
*
* Replace the symbolic node names with the real node names in your circuit in the same order !!
*
* Notes to pin assignment:
* ------------------------
* - f_a1, f_a2, ..., m_e7 are signal pins.
*
```

```
*            model boundary -->|
*                              |
*                              |
*     _________________________
*    |                         |
*    |                         |
*    |                         |
*    |                      e  |- -- f_e1, f_e2, f_e3, f_e4, f_e5, f_e6, f_e7 (signal pins row e)
*    |                         |
*    |                      d  |- -- f_d1, f_d2, f_d3, f_d4, f_d5, f_d6, f_d7 (signal pins row d)
*    |                         |
*    |                      c  |- -- f_c1, f_c2, f_c3, f_c4, f_c5, f_c6, f_c7 (signal pins row c)
*    |                         |
*    |                      b  |- -- f_b1, f_b2, f_b3, f_b4, f_b5, f_b6, f_b7 (signal pins row b)
*    |                         |
*    |                      a  |- -- f_a1, f_a2, f_a3, f_a4, f_a5, f_a6, f_a7 (signal pins row a)
*    |                         |
*    |                         |
*    |                         |
*    |                      ___|
*  _ |                     | _
* | \|                     |/ |
* | ||                     || |
* | ||                     || |
* | ||_____________________|| |         model boundary
* | |_______________________| |         |
* |                           |         |
* |     e   d   c   b   a     |         |
* |___________________________| ________V_
*       |   |   |   |   |
*       |   |   |   |   |
*       |   |   |   |   ------- -- m_a1, m_a2, m_a3, m_a4, m_a5, m_a6, m_a7  (signal pins row a)
*       |   |   |   ----------- -- m_b1, m_b2, m_b3, m_b4, m_b5, m_b6, m_b7  (signal pins row b)
*       |   |   --------------- -- m_c1, m_c2, m_c3, m_c4, m_c5, m_c6, m_c7  (signal pins row c)
*       |   ------------------- -- m_d1, m_d2, m_d3, m_d4, m_d5, m_d6, m_d7  (signal pins row d)
*       ----------------------- -- m_e1, m_e2, m_e3, m_e4, m_e5, m_e6, m_e7  (signal pins row e)
*
* -------------------------------------------------------------------------------------
*    SPICE code
* -------------------------------------------------------------------------------------
*
```

definition of the subcircuits connector for the whole connector

```
.Subckt elco_2mm_standard
```

definition of subcircuit pins

```
+ 1011 1012 1013 1014 1015 1016 1017
+ 1021 1022 1023 1024 1025 1026 1027
+ 1031 1032 1033 1034 1035 1036 1037
+ 1041 1042 1043 1044 1045 1046 1047
+ 1051 1052 1053 1054 1055 1056 1057
+ 2011 2012 2013 2014 2015 2016 2017
+ 2021 2022 2023 2024 2025 2026 2027
+ 2031 2032 2033 2034 2035 2036 2037
+ 2041 2042 2043 2044 2045 2046 2047
+ 2051 2052 2053 2054 2055 2056 2057

x1
```

connecting x1(daughtercard_elco_2mm) with Input and x2 (backplane_elco_2mm)

```
+ 1011 1012 1013 1014 1015 1016 1017
+ 1021 1022 1023 1024 1025 1026 1027
+ 1031 1032 1033 1034 1035 1036 1037
+ 1041 1042 1043 1044 1045 1046 1047
+ 1051 1052 1053 1054 1055 1056 1057
+ 3011 3012 3013 3014 3015 3016 3017
+ 3021 3022 3023 3024 3025 3026 3027
+ 3031 3032 3033 3034 3035 3036 3037
+ 3041 3042 3043 3044 3045 3046 3047
+ 3051 3052 3053 3054 3055 3056 3057
+ daughtercard_elco_2mm
```

```
x2
```
☞ **connecting x2(daughtercard_elco_2mm) with x2 (backplane_elco_2mm) and Output**
```
+ 3011 3012 3013 3014 3015 3016 3017
+ 3021 3022 3023 3024 3025 3026 3027
+ 3031 3032 3033 3034 3035 3036 3037
+ 3041 3042 3043 3044 3045 3046 3047
+ 3051 3052 3053 3054 3055 3056 3057
+ 2011 2012 2013 2014 2015 2016 2017
+ 2021 2022 2023 2024 2025 2026 2027
+ 2031 2032 2033 2034 2035 2036 2037
+ 2041 2042 2043 2044 2045 2046 2047
+ 2051 2052 2053 2054 2055 2056 2057
+ backplane_elco_2mm
* -----
```

Structure of a subcircuit for one connector segment

In the following an abstract of a subcircuit of the daughtercard connector module is displayed, where the essential components of its is explained.

☞ **definition of the subcircuit ' daughtercard_elco_2mm '**
```
.Subckt daughtercard_elco_2mm
```
☞ **definition of the subcircuits pins**
```
+  11  21  31  41  51  61  71
+  81  91 101 111 121 131 141
+ 151 161 171 181 191 201 211
+ 221 231 241 251 261 271 281
+ 291 301 311 321 331 341 351
+  12  22  32  42  52  62  72
+  82  92 102 112 122 132 142
+ 152 162 172 182 192 202 212
+ 222 232 242 252 262 272 282
+ 292 302 312 322 332 342 352

* CST circuit Connector Daughtercard
```
☞ **definition of the single elements**
```
* Resistance of one connector segment
* -------------------------------
R1 11 360 0.001
R2 362 363 0.001
:

* Inductance of one connector segment
* -------------------------------
L1 360 361 3.14731e-010
L2 361 362 3.14731e-010
:

* Capacitances in x-direction of one connector segment
* -------------------------------
C36 361 365 8.70754e-015
:

* Magnetic coupling of one connector segment
* -------------------------------
K1 L1 L3 0.376341
K2 L2 L4 0.376341
:
```

☞ **end of subcircuit 'SEG1' definition**
```
.ends
```

✐ definition of the subcircuit ' backplane_elco_2mm '

```
.subckt backplane_elco_2mm
```

✐ definition of the subcircuits pins

```
+  11  21  31  41  51  61  71
+  81  91 101 111 121 131 141
+ 151 161 171 181 191 201 211
+ 221 231 241 251 261 271 281
+ 291 301 311 321 331 341 351
+  12  22  32  42  52  62  72
+  82  92 102 112 122 132 142
+ 152 162 172 182 192 202 212
+ 222 232 242 252 262 272 282
+ 292 302 312 322 332 342 352

* CST circuit Connector Backplane
```

✐ definition of the single elements

```
R1 11 360 0.001
R2 362 363 0.001
L1 360 361 1.54097e-009
:
```

✐ end of subcircuit ' backplane_elco_2mm ' definition

```
.ends


.ends
```

4 Test board

The purpose of this test board is to verify the generated SPICE ELCO B25 model. Therefore a comparison between different SPICE model simulations and corresponding measurements with this test board are made. After an accurate calibration very extensive measurements can be performed.
The following sections describe the layout of this test board in detail.

Fig. 13 Assembled test board

4.1 Top Layer

On the Top Layer the ELCO connectors and calibration structures for SOLT (Short Open Load Thru) and TRL (Thru Reflect Line) Calibration are fitted. Please refer to the measurement instructions and the manufacturing documents for further information.

There are 12 test structures (6 single line and 6 differential line test structures):

OPEN
SHORT
LOAD
THRU
LINE1
LINE2

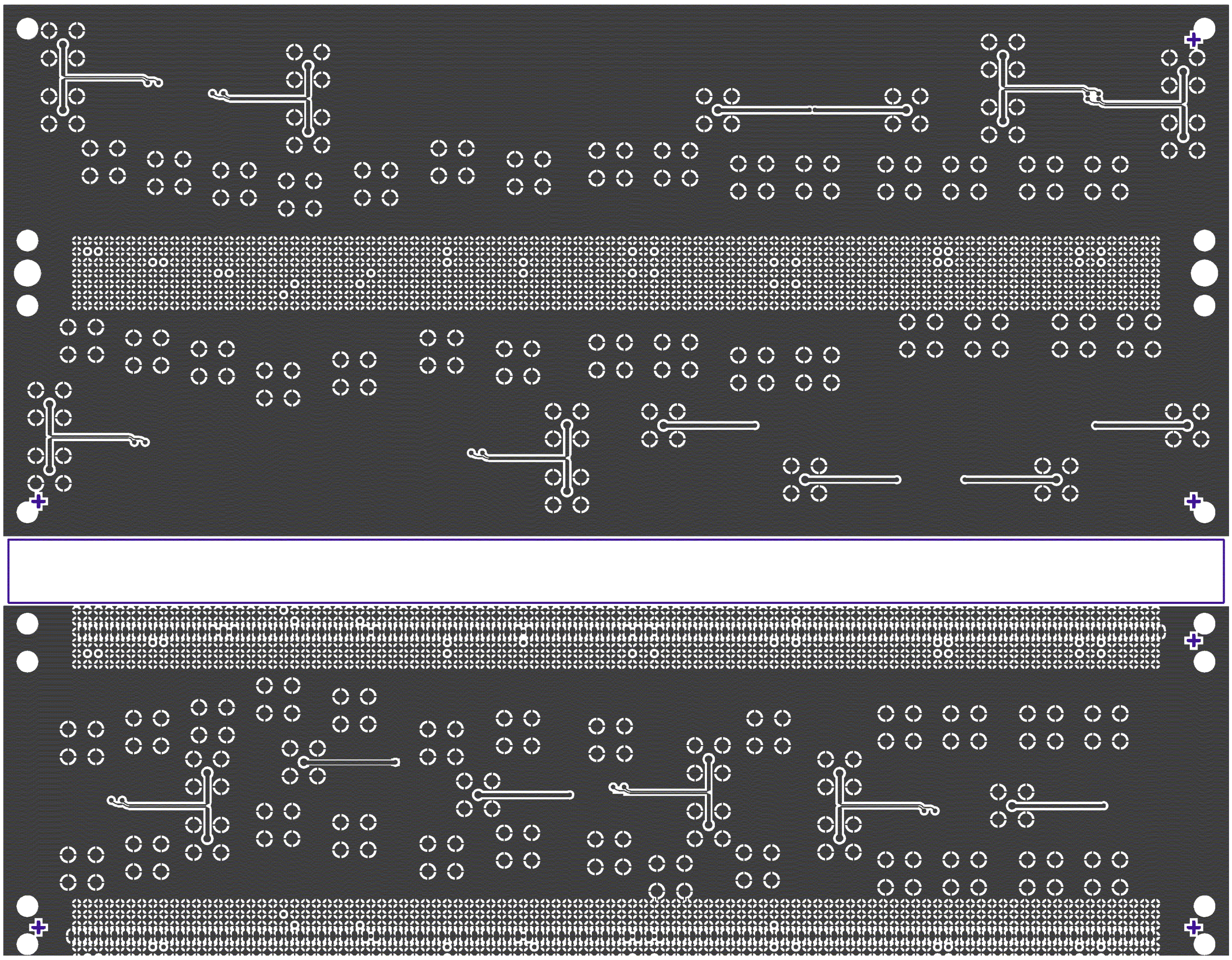


Fig. 14 Backplane (upper PCB) and daugthercard (lower PCB) with test structures on top layer

4.2 Bottom Layer

The Bottom Layer contains 63 different test setups with TDR, S-parameter, near and far end crosstalk measurements for single and differential line. Please refer to the measurement instructions and the manufacturing documents for additional information.

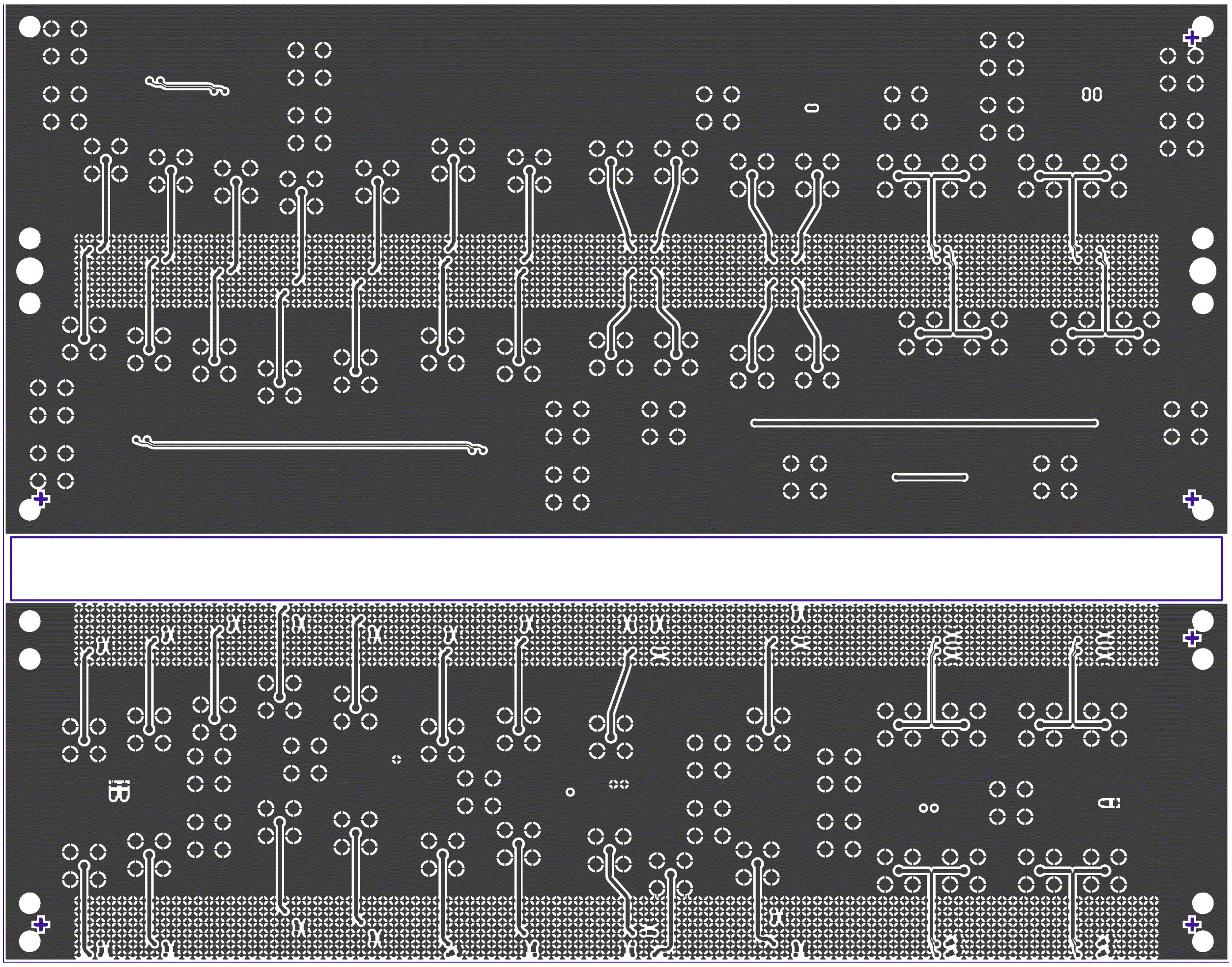


Fig. 15 Backplane (upper PCB) and daugthercard (lower PCB) with test structures on bottom layer

5 Hardware to model correlation

To determine the typical characteristics of a connector system the following two measurement and simulation methods are used:

- S-parameter measurements in frequency domain with the aid of a network analyzer
- Reflection behavior measurement with a time domain reflectometer (TDR)

This is necessary to compare the measurements with the generated ELCO B25 SPICE model. With this comparison the quality and accuracy of the ELCO B25 SPICE model can be verified.

5.1 Time domain

5.1.1 *Measurement*

Fig. 16 shows the measurement setup for a time domain reflectometer. The measurement object is connected with the step function generator via a high precision measurement cable. The test object (ELCO connector) is connected with the step function generator and the measurement inputs via short conductors and SMA connectors. The TDR is controlled via the software IC-CAP from Agilent Technologies. The principle of this measurement is explained in chapter 2.1.

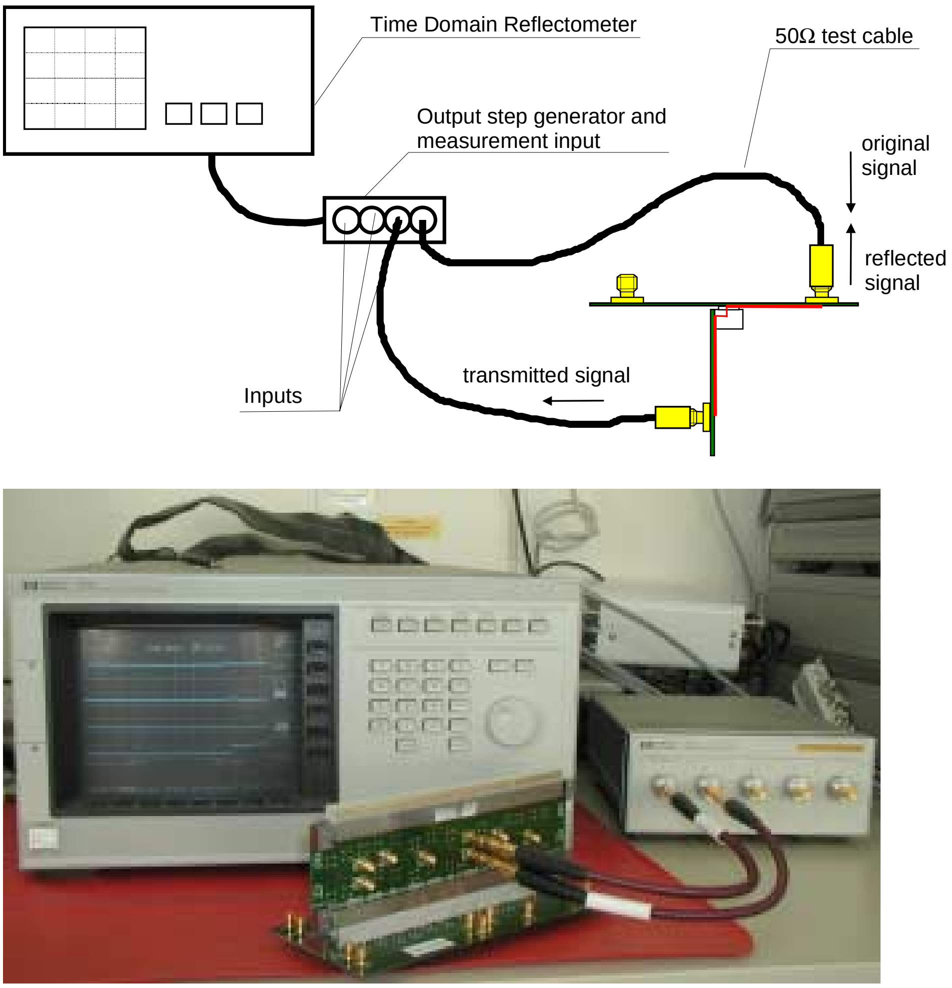
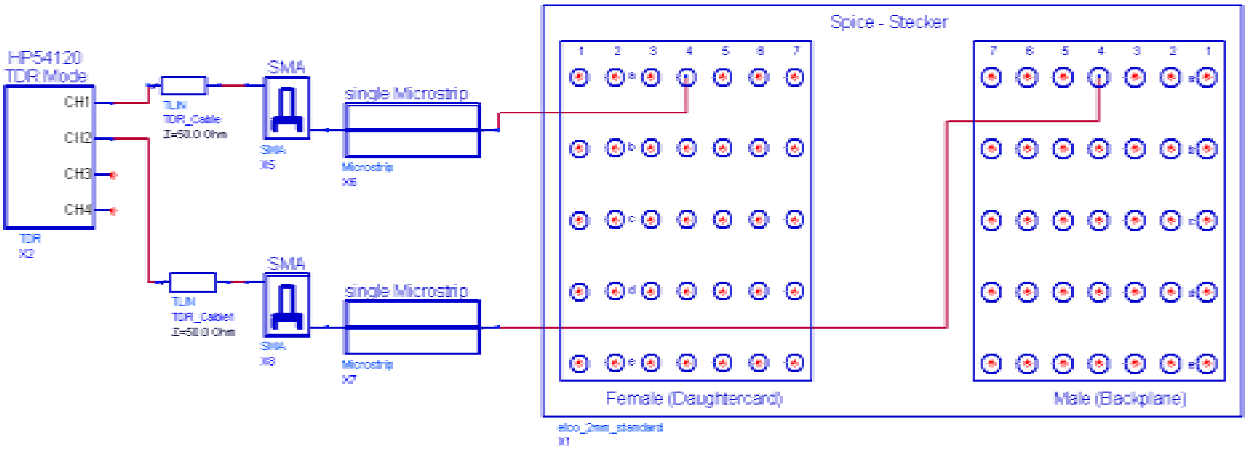


Fig. 16 TDR measurement setup

5.1.2 *Simulation*

The wave impedance of a connector is determined by the described time domain reflectometer method from the reflection behavior of the measurement object in time domain like in Fig.1. For this purpose the TDR measurement circuit is reproduced in the circuit simulator. The measurement object, the ELCO connector, is represented through its SPICE model. The impedance is simulated with the Advanced Design System (ADS) from Agilent Technologies. This simulator is especially developed for high frequency applications which is capable to use S-parameters for time domain calculations.

Single:



Differential:

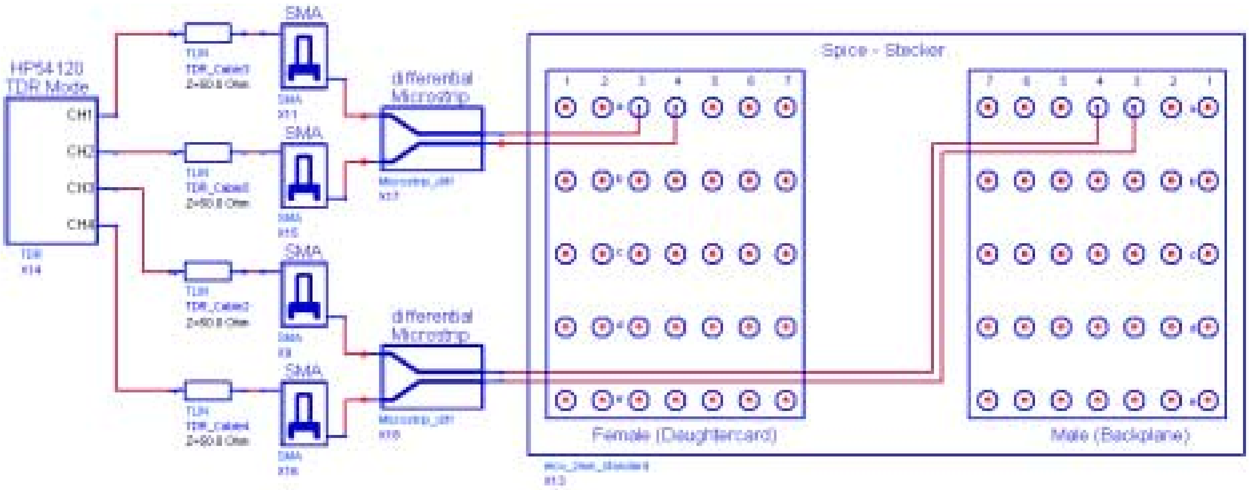


Fig. 17 Simulation setup for transmission and reflection behavior

5.1.3 Examples

Measured impedance of the test board in Ω

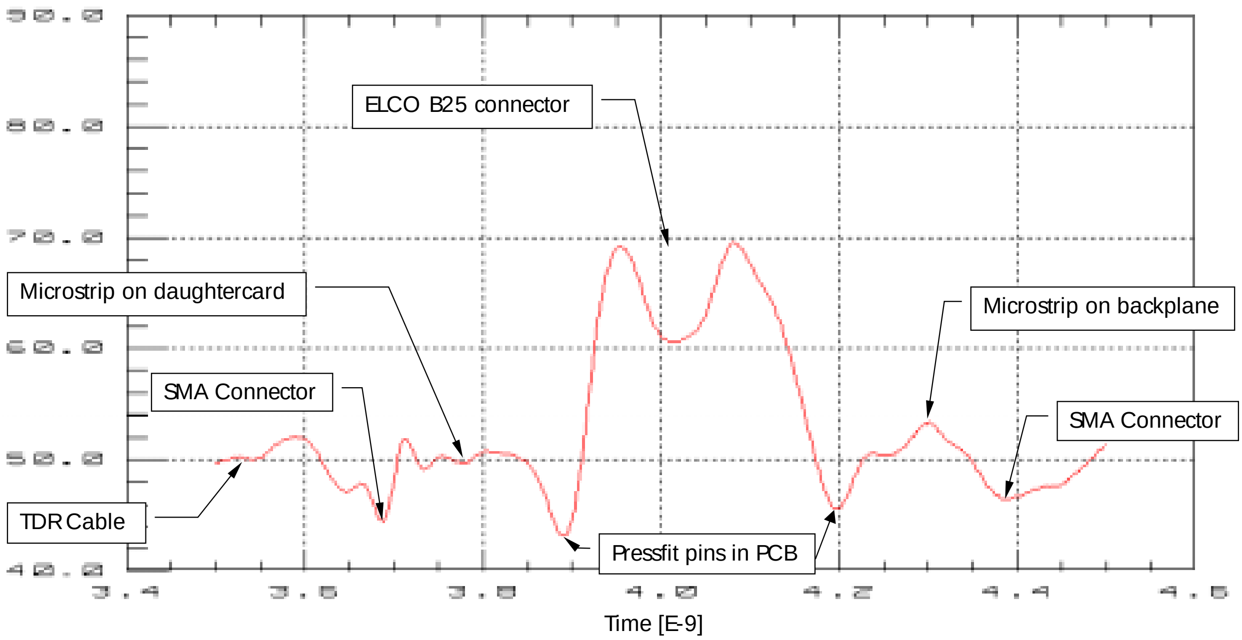
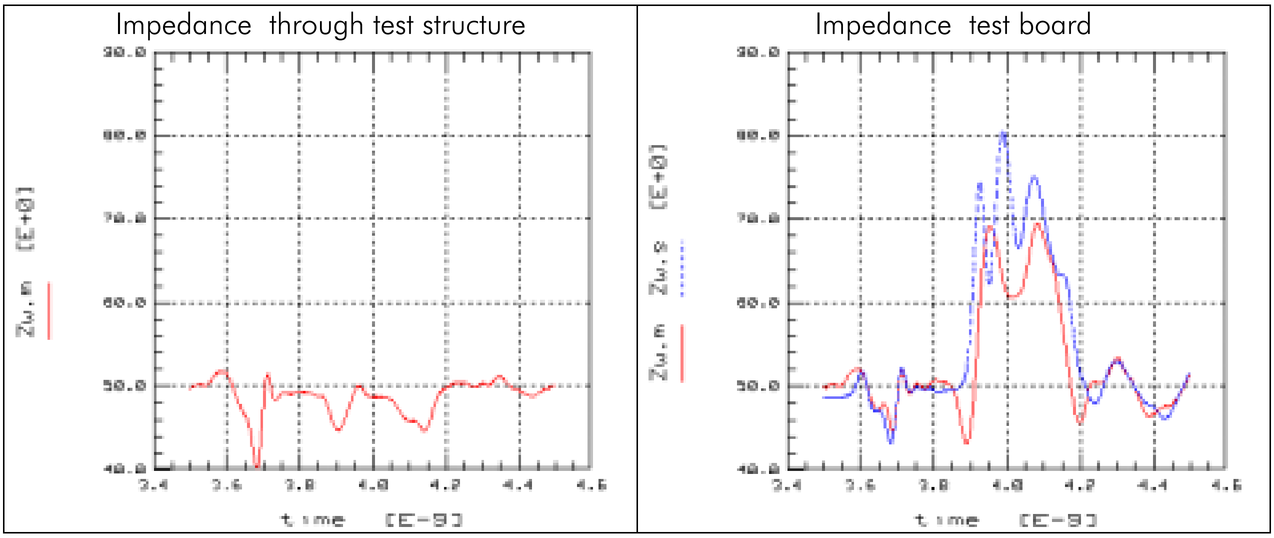


Fig. 18 Impedance curve with explanations

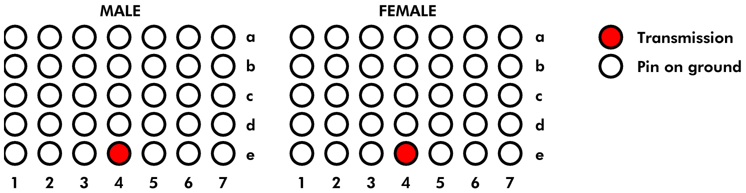
This diagram shows the typical impedance of the signal path from the first SMA connector on the daughter card to the second SMA connector on the backplane. It is simulated in the same way than a Time Domain Reflectometer measurement. The y-axis directly shows the impedance in [Ω].
Please note, that due to multiple reflections, the impedance profile is only correct at the beginning of the transmission path !

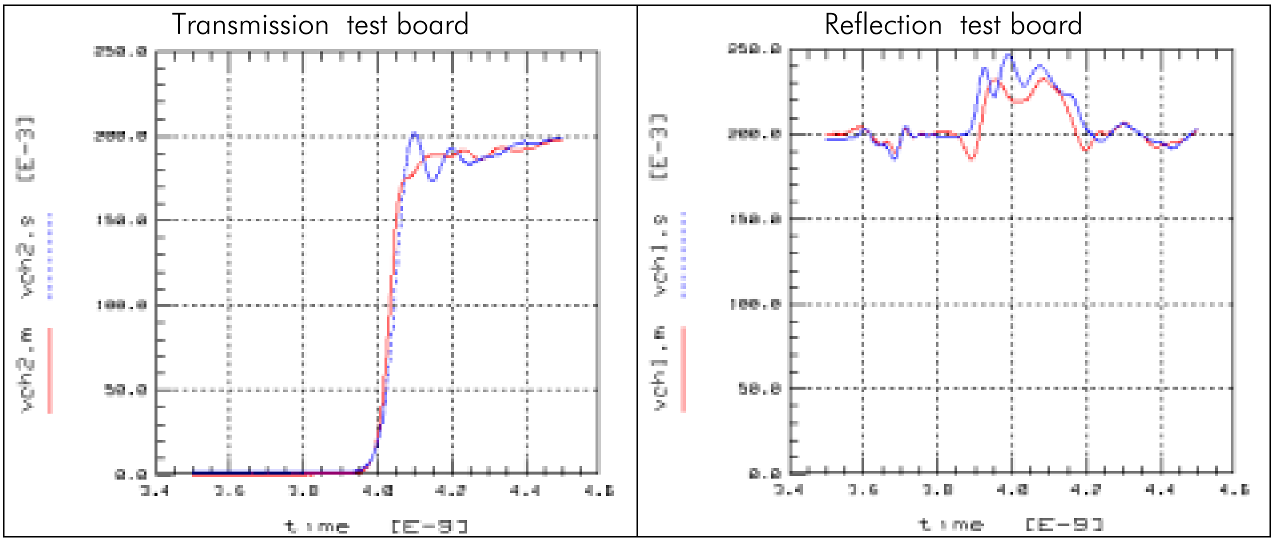
The diagrams below show the impedance in [Ω] of the through test structure (SMA Connector – Microstrip – SMA Connector) and the measured (red line) and simulated (blue line) impedance in [Ω] of the test board with SMA Connector – Microstrip – ELCO B25 Connector – Microstrip – SMA Connector (see diagram above for detailed information).



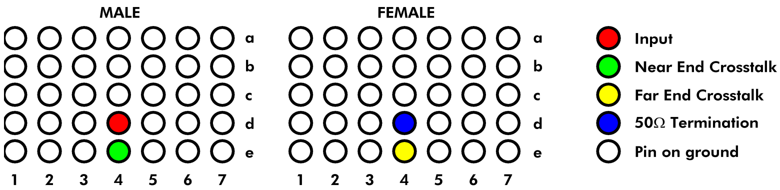
Single test structure:

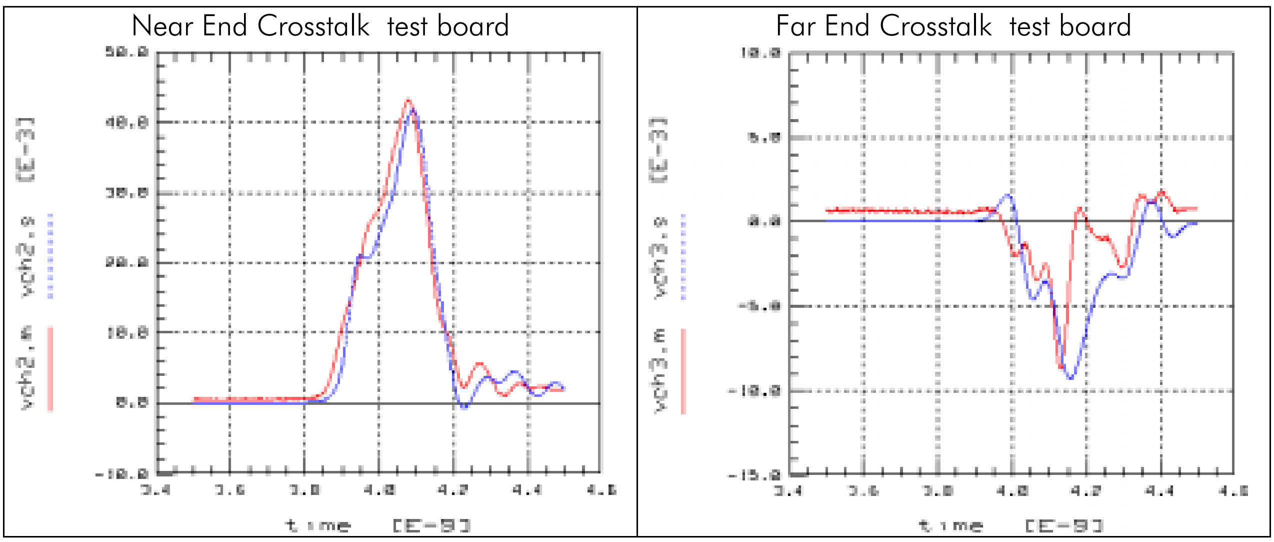
The following diagrams show the measured (red line) and simulated (blue line) transmission and reflection behavior of a single test structure with the ELCO B25 connector at row e in [mV].





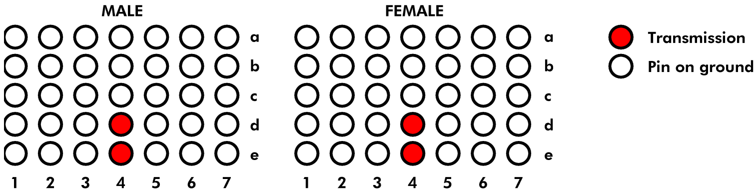
The next diagrams show the measured (red line) and simulated (blue line) Near and Far End Crosstalk of a single test structure with the ELCO B25 connector between row d and e in [mV].

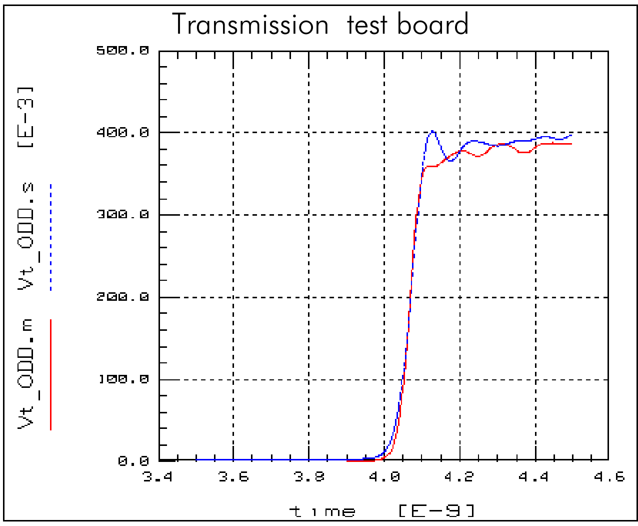


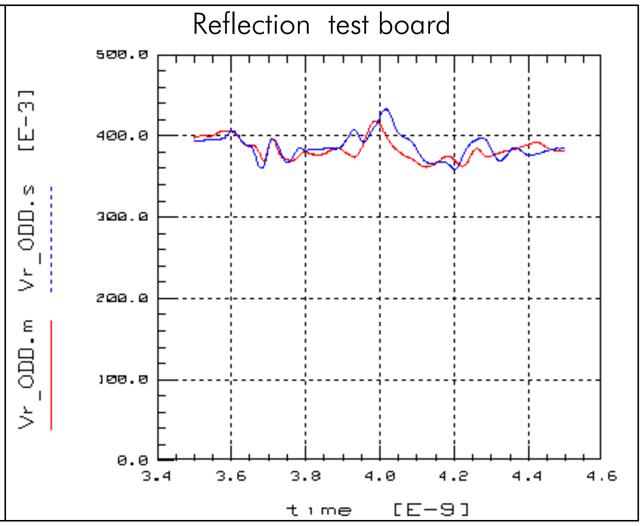


Differential test structure:

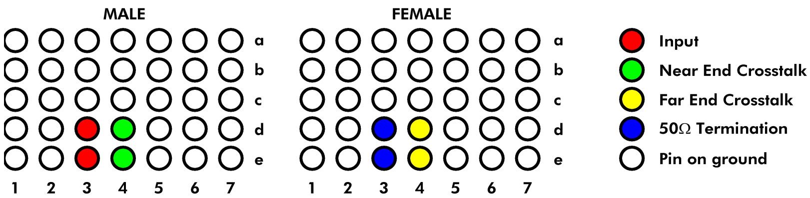
The following diagrams show the measured (red line) and simulated (blue line) transmission and reflection behavior of a differential test structure with the ELCO B25 connector in [mV].

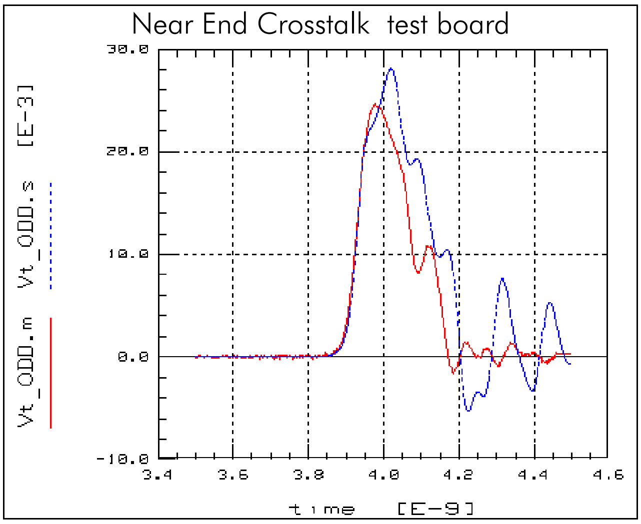






The next diagrams show the measured (red line) and simulated (blue line) Near End Crosstalk of a differential test structure with the ELCO B25 connector in [mV].





5.2 Frequency domain

5.2.1 Measurement

The principle measurement setup for S-parameter measurements with a network analyzer (NWA) is shown in Fig.19. The NWA can be adjusted with a calibration kit to the calibration plane. The actual test object, the ELCO connector, is connected with short conductors and SMA connectors on the test board to the cables of the NWA. This means, that the measurement reproduces the behavior of the measurement object plus the influences of the test board. With the de-embedding procedure these unwanted effects (conductor, SMA connector) can be eliminated.

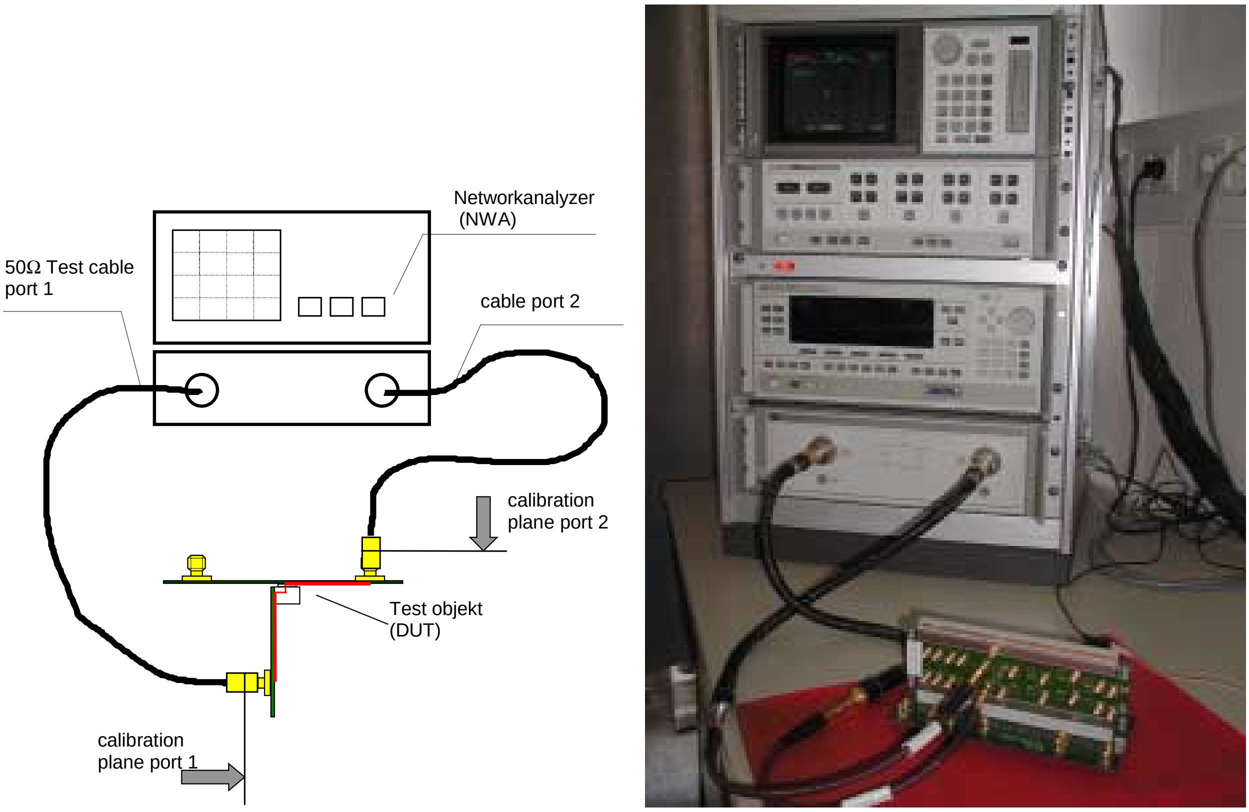


Fig. 19 NWA measurement setup

5.2.1.1 De-embedding

To access the measurement object with a measurement instrument, SMA connectors as well as lossy transmission lines on the PCB are connected to its contacts. The goal of the described de-embedding procedure is to cancel the negative influence of these elements.

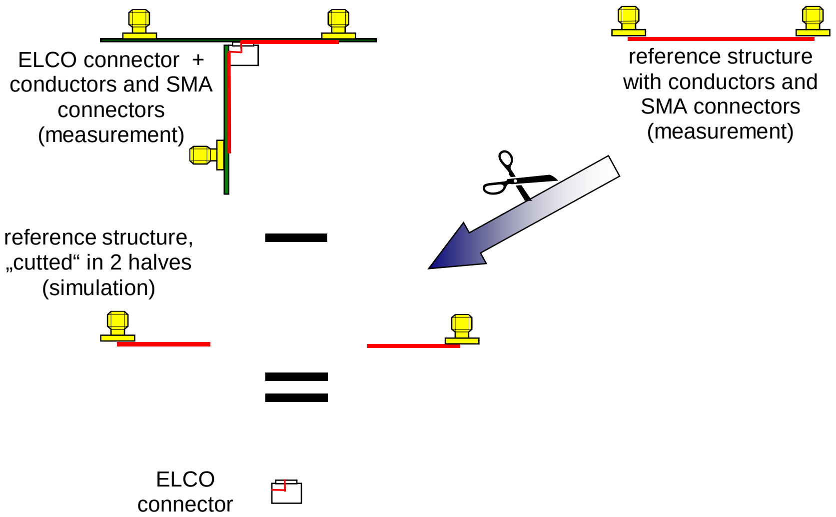


Fig. 20 De-embedding principle

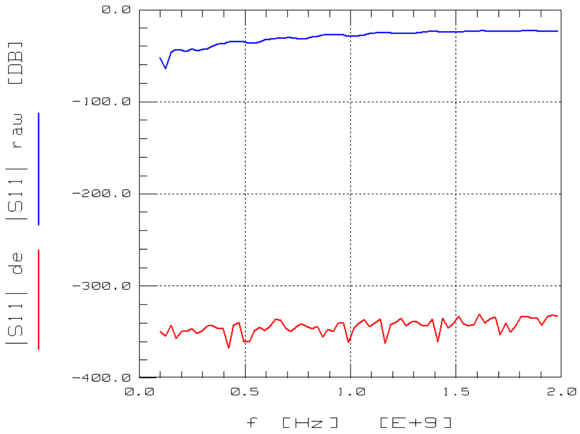
For TRL calibration different standards (TRL = Through Reflect Line) are measured and afterwards used for de-embedding the measurement object. The following drawing shows a list of the standards and their notation:

standard	notation	comment
	SHORT (= REFLECT)	Half reference conductor, shorted at the end. This generates maximum reflection.
	THROUGH	Reference conductor. The calibration plane is located exactly in the middle of the reference conductor after de-embedding.
	LINE	One or several conductors (dependent on the frequency range) with the same layout as the reference conductor.

Tab. 3 Standards for TRL de-embedding

The TRL procedure is included as a function in the software IC-CAP from Agilent Technologies, which is used for measuring and analyzing the S-parameters. A exact derivation can be found in [5].

The quality of the de-embedding procedure can be examined by de-embedding the reference structure from itself.
In ideal case this would result in the value 1 (=0dB) for transmission |S21| and the value 0 (=- ∞dB) for reflection |S11|. Both diagrams below show the typical results for a de-embedding:



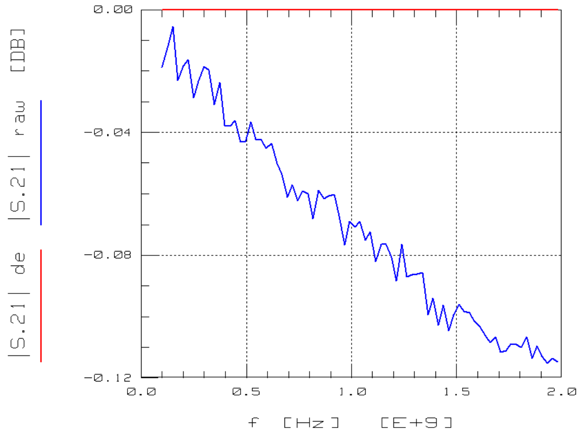
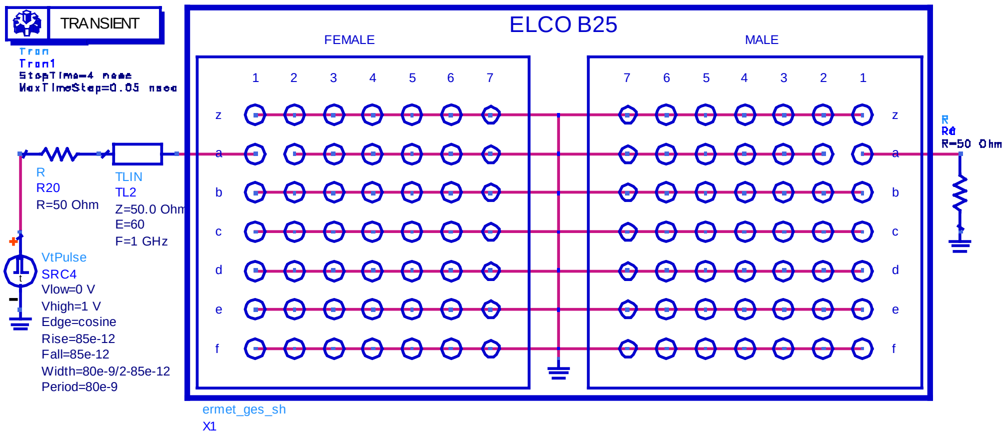


Fig. 21 |S11| and |S21| of a reference structure before and after de-embedding

5.2.2 *Simulation*

The corresponding simulation circuit to the frequency measurement is displayed below. This is for example a single and a differential line setup.

Single:



Differential:

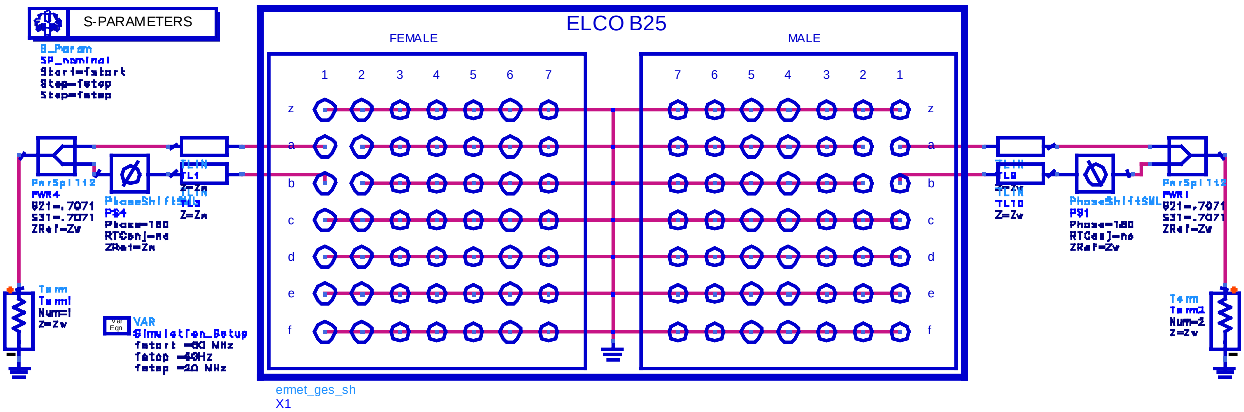
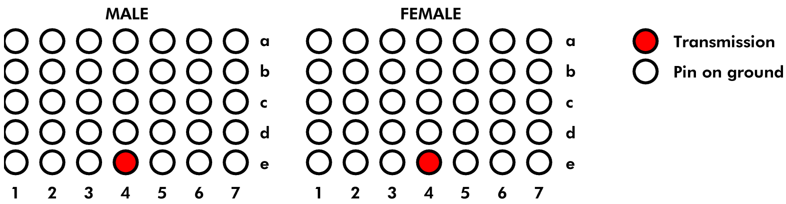


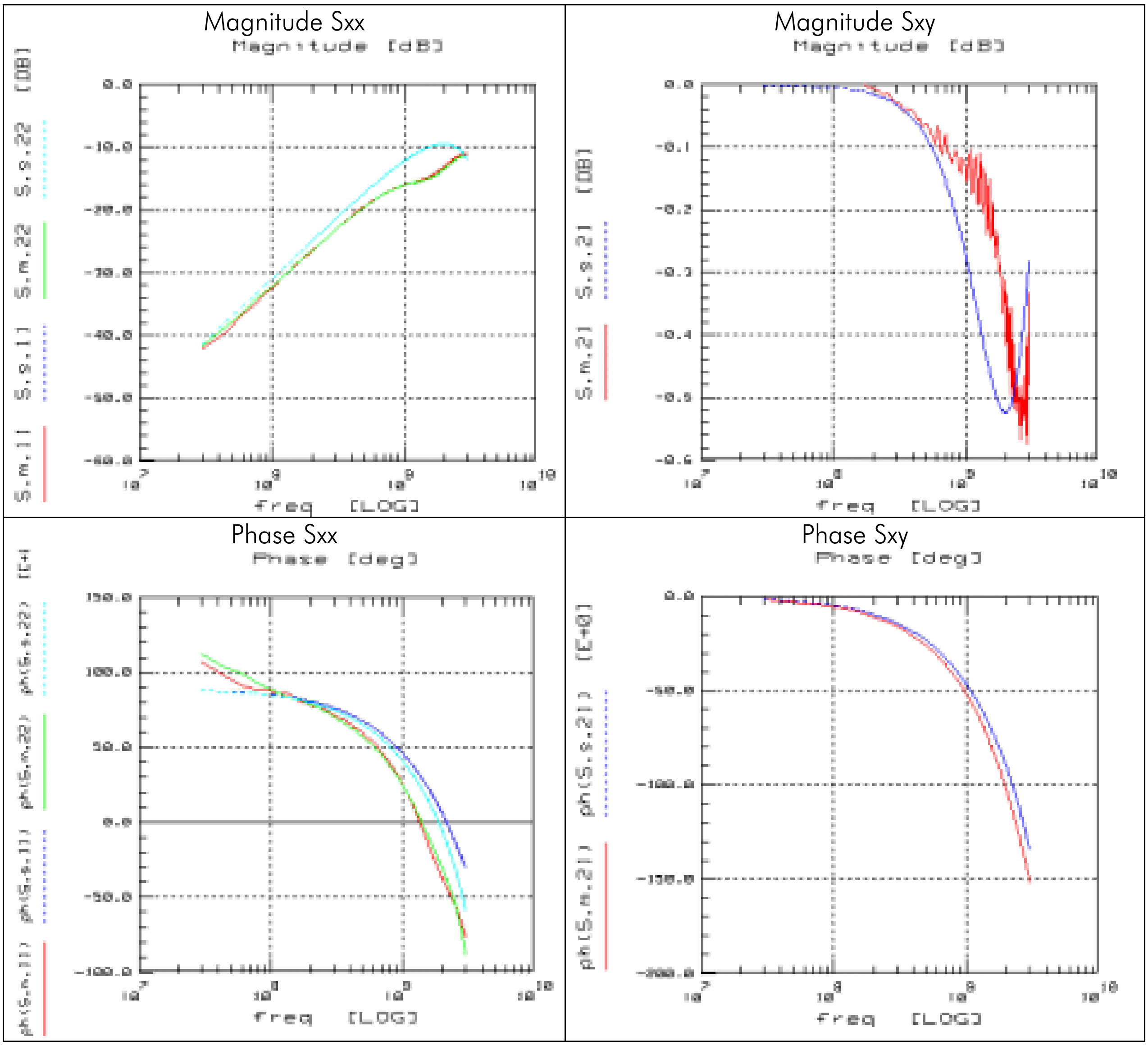
Fig. 22 Simulation setup for single and differential frequency behavior

5.2.3 *Examples*

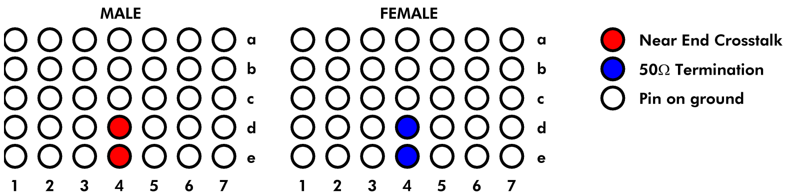
Single test structure:

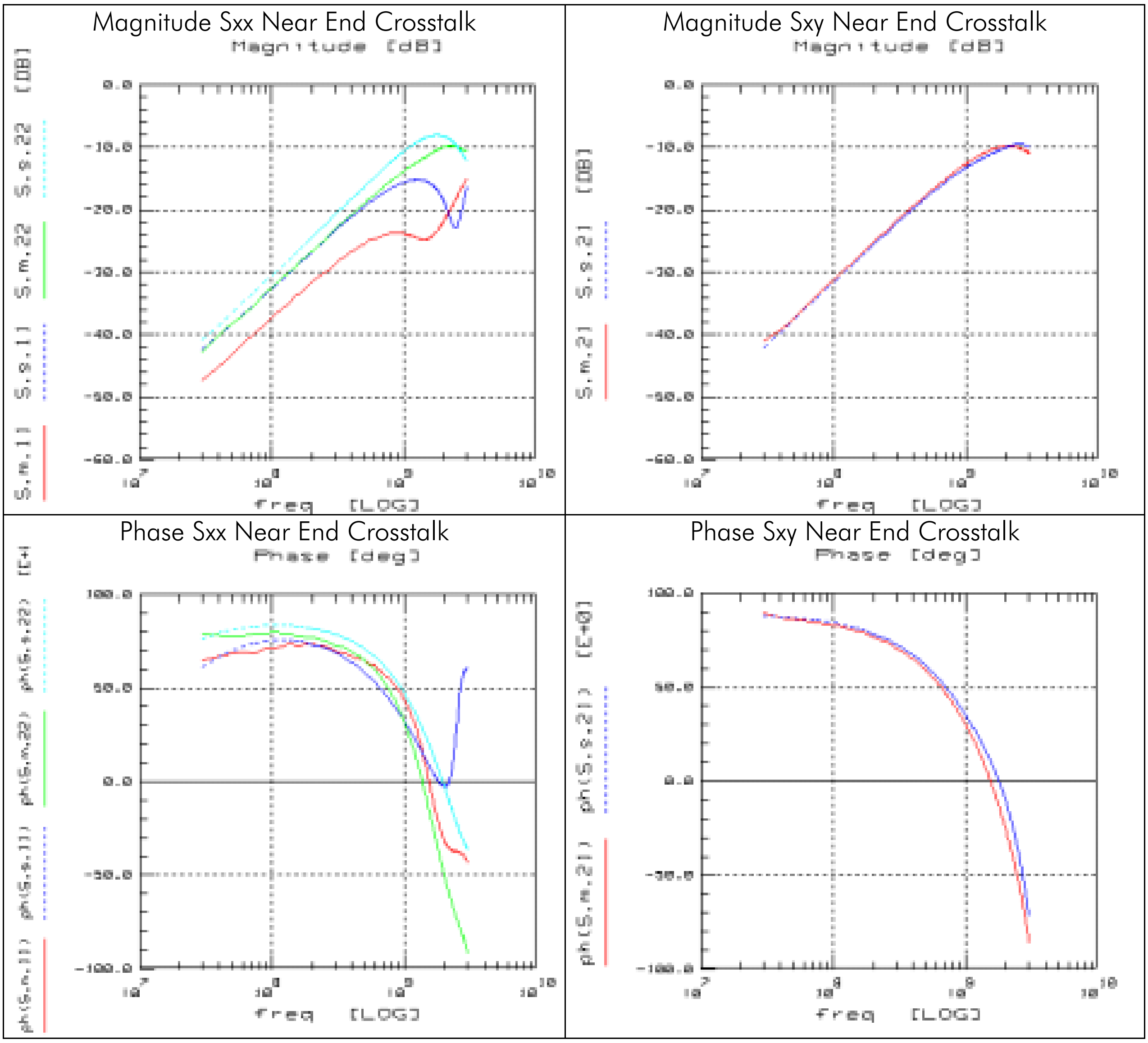
The following plots show the measured (red line) and simulated (blue line) magnitude and the phase of transmission and reflection at the following configuration:





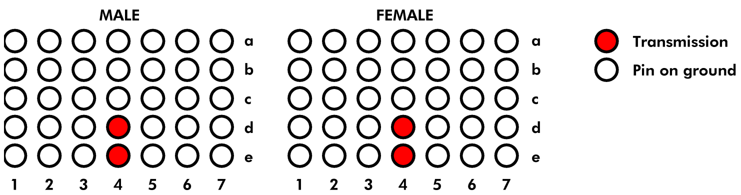
The next plots show the measured (red line) and simulated (blue line) magnitude and the phase of near end crosstalk at the following configuration:

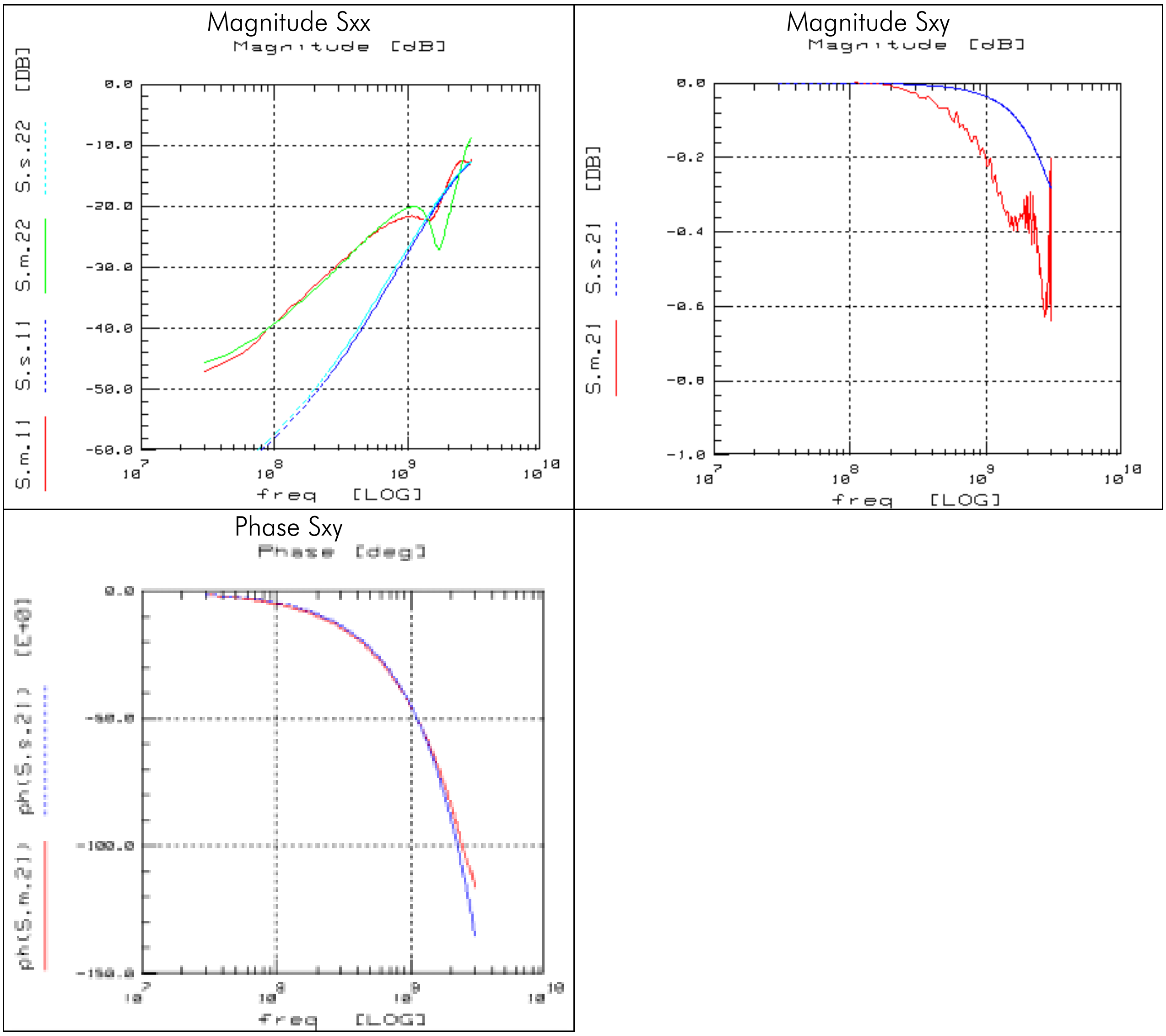




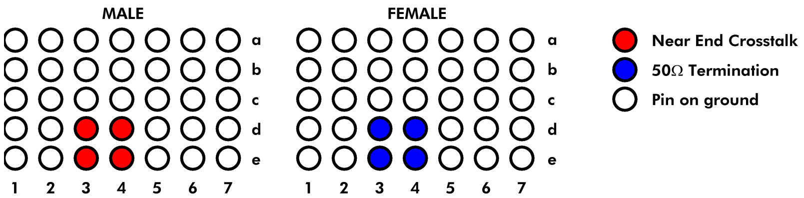
Differential test structure:

The following plots show the measured (red line) and simulated (blue line) magnitude and the phase of transmission and reflection at the following configuration:

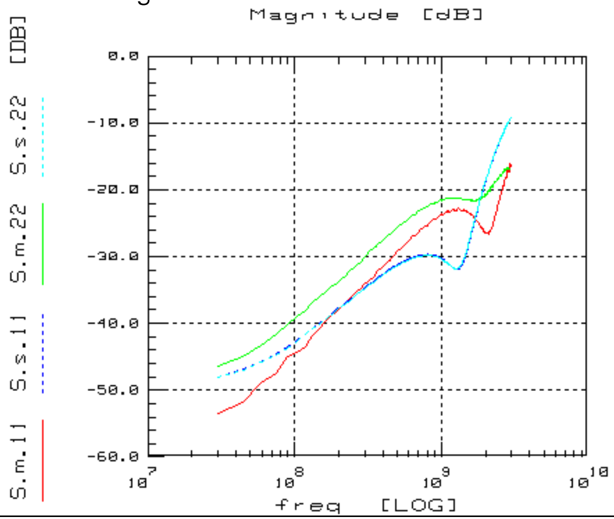




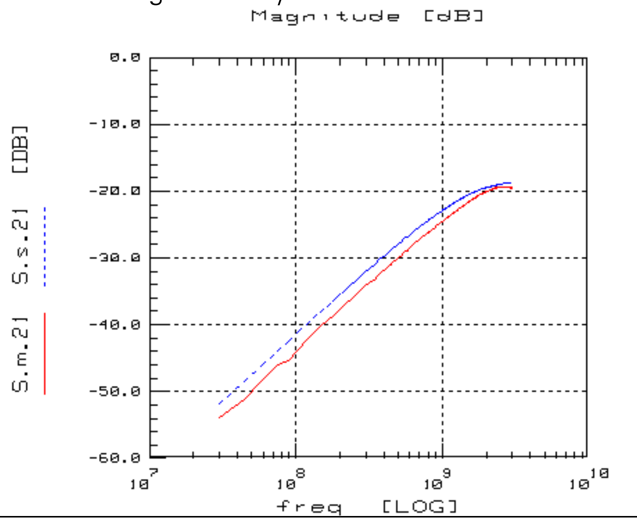
The next plots show the measured (red line) and simulated (blue line) magnitude and the phase of near end crosstalk at the following configuration:



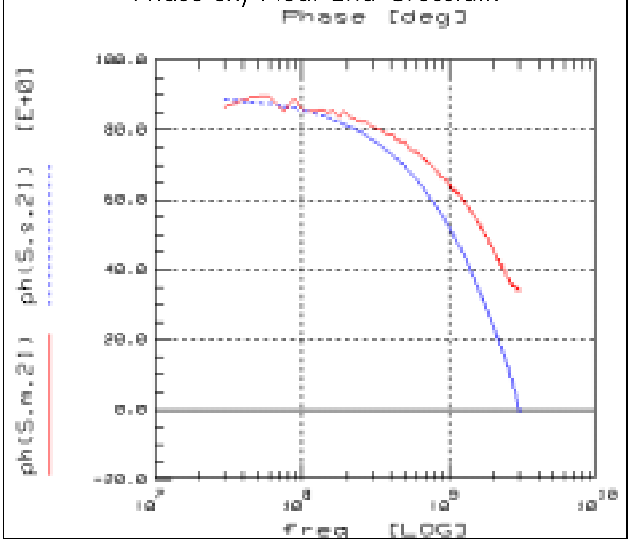
Magnitude Sxx Near End Crosstalk



Magnitude Sxy Near End Crosstalk



Phase Sxy Near End Crosstalk



6 Addendum

Tools and software used for measurement, simulation and PCB design:

NWA HP 8510	Hewlett Packard
NWA Calibration Kit	Hewlett Packard
TDR HP 54120	Hewlett Packard
Microwave Studio	CST
ADS	Agilent Technologies
IC-CAP	Agilent Technologies
Test board	ELCO
Power PCB	Mentor Graphics

Literature source:

[1]	E. Herter, W. Lörcher	Nachrichtentechnik		1987
[2]	T. Gneiting	Interconnect modeling for high speed data communications	EPFL Lausanne	2000
[3]		"Online Manual"	CST Microwave Studio	2003
[4]	T. Gneiting, H. Katzier	Modellierung von Gehäusen und Steckern	Hewlett Packard Workshop	1998
[5]	H.J. Eul, B.Shiek	Thru Match Reflect: One Result of a Rigorous Theory for De-embedding and Network Analyzer Calibration	Proc. 18th European Microwave Conference	1998