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TECHNICAL REPORT

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Multicore and symmetrical pair/quad cables for digital communications – Part 1-5: Correction procedures for the measurement results of return loss and input impedance

IEC/TR 61156-1-5:2013(E)

IEC Central Office
3, rue de Varembé
CH-1211 Geneva 20
Switzerland

Tel.: +41 22 919 02 11
Fax: +41 22 919 03 00
info@iec.ch
www.iec.ch

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CONTENTS

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MULTICORE AND SYMMETRICAL PAIR/QUAD CABLES FOR DIGITAL COMMUNICATIONS –

Part 1-5: Correction procedures for the measurement results of return loss and input impedance

FOREWORD

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IEC/TR 61156-1-5, which is a technical report, has been prepared by subcommittee SC46C: Wires and symmetric cables, of IEC technical committee TC46: Cables, wires, waveguides, R.F. connectors, R.F. and microwave passive components and accessories.

The text of this technical report is based on the following documents:

Enquiry draft	Report on voting
46C/973/DTR	46C/979/RVC

Full information on the voting for the approval of this technical report can be found in the report on voting indicated in the above table.

This publication has been drafted in accordance with the ISO/IEC Directives, Part 2.

A list of all parts in the IEC 61156 series, published under the general title *Multicore and symmetrical pair/quad cables for digital communications*, can be found on the IEC website.

The committee has decided that the contents of this publication will remain unchanged until the stability date indicated on the IEC web site under "http://webstore.iec.ch" in the data related to the specific publication. At this date, the publication will be

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MULTICORE AND SYMMETRICAL PAIR/QUAD CABLES FOR DIGITAL COMMUNICATIONS –

Part 1-5: Correction procedures for the measurement results of return loss and input impedance

1 Scope

This part of IEC 61156 describes correction procedures for the measurement results of return loss and input impedance.

2 Normative references

The following documents, in whole or in part, are normatively referenced in this document and are indispensable for its application. For dated references, only the edition cited applies. For undated references, the latest edition of the referenced document (including any amendments) applies.

IEC 61156-1, *Multicore and symmetrical pair/quad cables for digital communications – Part 1: Generic specification*

IEC/TR 61156-1-2, *Multicore and symmetrical pair/quad cables for digital communications – Part 1-2: Electrical transmission characteristics and test methods of symmetrical pair/quad cables*

IEC/TR 62152, *Transmission properties of cascaded two-ports or quadripols – Background of terms and definitions*

IEC 62153-1-1, *Metallic communication cables test methods – Part 1-1: Electrical – Measurement of the pulse/step return loss in the frequency domain using the Inverse Discrete Fourier Transformation*

ASTM D4566:1998, *Standard Test Methods for Electrical Performance Properties of Insulations and Jackets for Telecommunications Wire and Cable*

3 Acronyms

CUT	cable under test
FRL	fitted return loss
GRL	gated return loss
IFDT	Inverse discrete Fourier transformation
OSRL	open short return loss
PRL	parasitic inductance corrected return loss
RL	return loss
SRL	structural return loss

4 Return loss measurements

4.1 General

The return loss of a transmission line (cable) can be obtained by different test methods, each having certain advantages and/or disadvantages and therefore giving not exactly the same results. At higher frequencies, the measured return loss is strongly influenced by the cable end preparation (stray inductances and capacitances play an important role) leading to an underestimated cable performance. Under laboratory conditions, one might be able to minimize this negative effect; however under general industrial conditions using automated test systems, one might need mathematical procedures to eliminate the effect of the cable end preparation.

The results of return loss measurements may also depend on the sample length. Therefore, for balanced cables according to IEC 61156 series, the specified sample length is at least 100 m (if not specified otherwise).

4.2 Return loss (RL)

The return loss is obtained by measuring the scattering parameter $S11$ using a test apparatus of which the test port has the same reference impedance than the cable under test, and where the far end of the cable is terminated with its reference impedance (See IEC/TR 62152, IEC 61156-1). The return loss takes into account the structural variations along the cable and the mismatch between the reference impedance and the (mean) characteristic impedance of the pair. If the (mean) characteristic impedance of the pair is different from the reference impedance, one gets, especially at lower frequencies (where the round trip attenuation is low), multiple reflections that are overlaid to the structural and junction reflections. Therefore, return loss RL is also referenced as operational return loss.

NOTE The impedance of a homogenous transmission line is a complex quantity where the real part is decreasing with frequency and tending to an asymptotic value, the so called characteristic impedance and the imaginary part is also decreasing from negative values to zero. This “normal” behaviour of a transmission line will also generate multiple reflections.

As an example, Figure 1 shows the operational return loss under different conditions. The blue line shows the return loss of a pair having a characteristic impedance equal to the reference impedance but taking into account that the impedance is varying with frequency (see right-hand graph). The red line shows the return loss of a pair having a characteristic impedance that is different from the reference impedance (110 Ω vs. 100 Ω). For both lines, we observe periodic variations that are caused by multiple reflections between the junctions at the near and far end. The green line shows a simulation of a pair having a frequency independent characteristic impedance which is equal to the reference impedance.

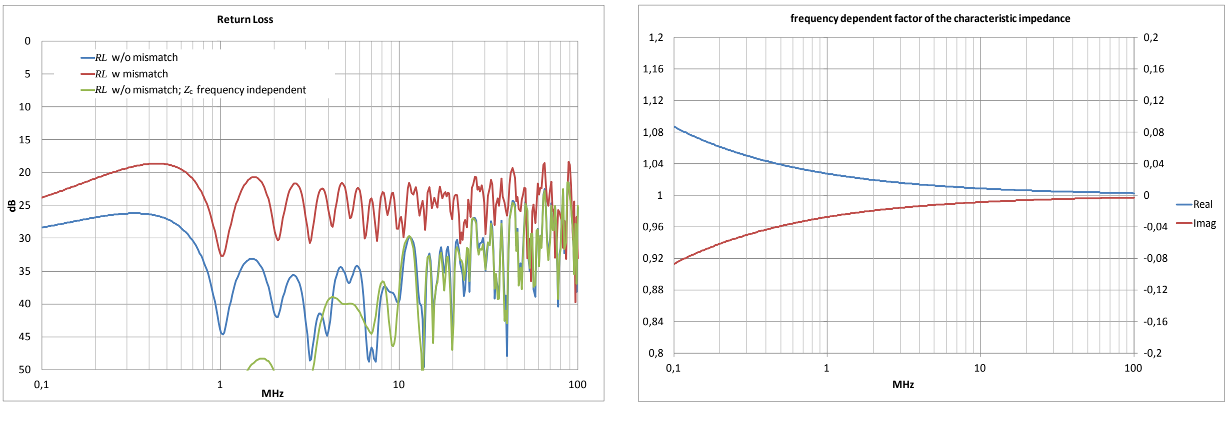

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Figure 1 – Return loss with and without junction reflections

4.3 Open/short return loss (*OSRL*)

A way to avoid in the measurement of return loss multiple reflections due to a mismatch between the characteristic impedance (asymptotic value at high frequencies) of the CUT and the reference impedance is to use a CUT terminated in its nominal impedance and having a very long test length such that the round trip attenuation of the CUT is at least 40 dB at the lowest frequency to be measured. For standard LAN cables, this would result in a CUT length of roughly 1 000 m for the lowest frequency of 1 MHz.

Another way (when long CUT length is not available) is to measure the characteristic impedance (open/short method, see IEC/TR 62152) and to calculate the return loss. As the characteristic impedance is obtained from the measurement of the open and short circuit impedance, it is proposed to name such obtained return loss open/short return loss.

This open/short return loss includes the effect of structural variations and the mismatch at the near end (including the effect due to a frequency-dependent characteristic impedance), but it does not take into account multiple reflections.

Figure 2 shows the difference between operational return loss and open/short return loss. The left-hand graph shows the results of a pair having a characteristic impedance which is different from the reference impedance (110 Ω vs. 100 Ω). The right-hand graph shows the results of a pair having a characteristic impedance which is equal to the reference impedance (100 Ω). We recognize that the open/short return loss does not take into account multiple reflections.

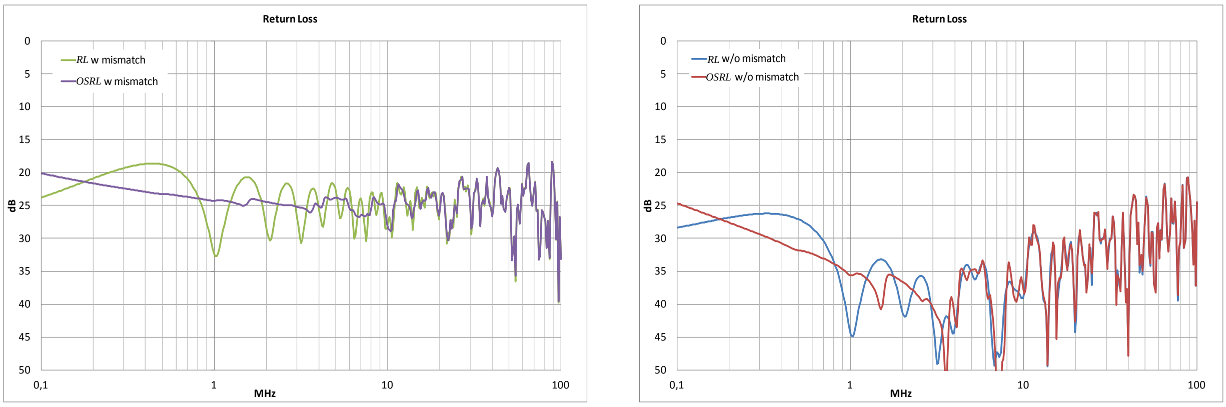

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Figure 2 – Return loss and open/short return loss

4.4 Structural return loss (*SRL*)

The structural return loss is the return loss where only structural variations along the cable are taken into account. The mismatch effects at the input and output of the transmission line (including the effect due to a frequency-dependent characteristic impedance) have been eliminated (see IEC/TR 62152). The structural return loss cannot be measured directly but is calculated from the measurement of the characteristic impedance (open/short method).

$$SRL = 20 \cdot \log_{10} \left| \frac{Z_{\text{CM}} - Z_{\text{C}}}{Z_{\text{CM}} + Z_{\text{C}}} \right| \qquad (1)$$

where

Z_{CM} is the (complex) mean characteristic impedance obtained from the measurement of the open and short circuit impedance;

Z_C is the (complex) characteristic impedance obtained from a curve fitting of the real and imaginary part of Z_{CM}.

The left-hand graph of Figure 3 shows the operational return loss, open/short return loss and structural return loss of a CUT having a characteristic impedance of 110 Ω. We clearly see the differences between them. The operational return loss takes into account all effects (structural variations, mismatch effects at the input and output). The open/short return loss does not take into account mismatch effects at the output (i.e. no multiple reflections). Whereas the structural return loss only takes into account structural variations along the cable.

The right-hand graph shows the real and imaginary part of the mean characteristic impedance (obtained from the measurement of the open and short circuit impedance) and it's fitting.

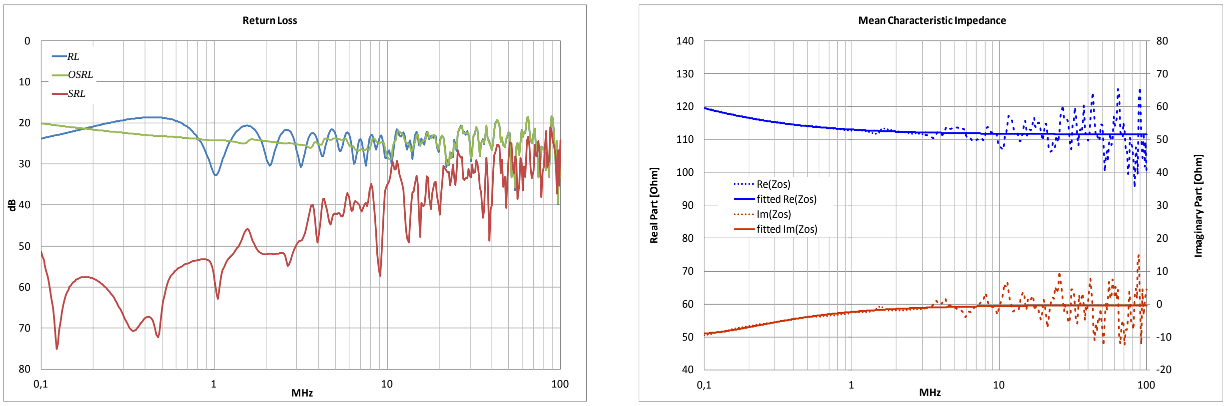

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Figure 3 – Return loss, open short return loss and structural return loss

5 Correction procedures

5.1 General

From the transmission line theory we know that the characteristic impedance is decreasing with frequency and approaching an asymptotic value (assuming dielectrics having an almost frequency-independent dielectric constant). Therefore an asymptotic value is assumed in the described correction procedures.

However measurements at higher frequencies often show an increase of the characteristic impedance. This is due to the cable end preparation and related stray inductances and capacitances. In fact we can consider the CUT as a cascade of 3 transmission lines, the two cable ends and the cable, or even 5 lines taking into account the test fixtures.

Figure 4 shows the results of an S/FTP cable obtained with an automated test system. On the left-hand side we see the input impedance (magnitude) and its fitting and on the right-hand side the operational return loss. One can recognize the above-described effect of increasing impedance and related decreasing return loss. Thus correction procedures are needed.

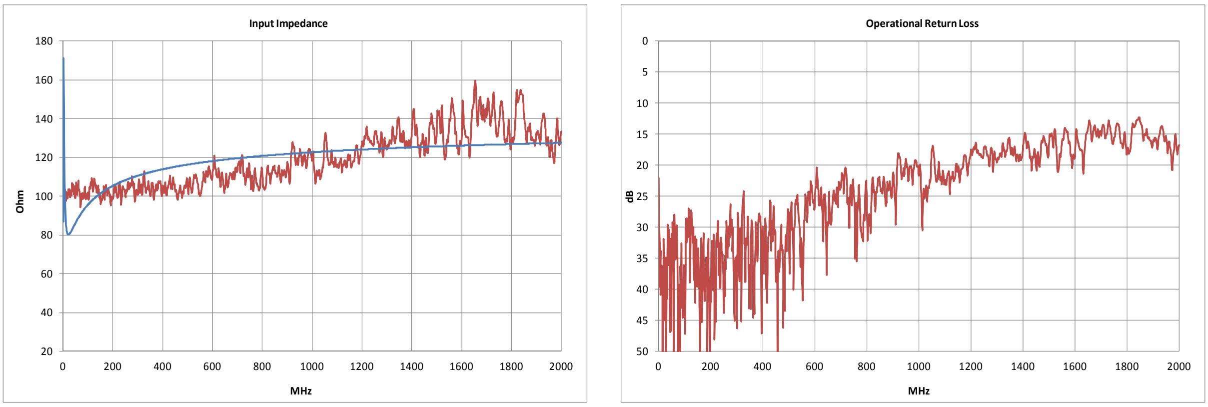

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Figure 4 – Input impedance (magnitude) and operational return loss up to 2 GHz

5.2 Parasitic inductance corrected return loss (*PRL*)

The characteristic impedance of an ideal cable is at high frequencies constant and only has a real part. Measuring the input impedance Z_i of a cable in contrast often shows a steady increase with frequency. The same effect can be seen measuring the return loss. This is mainly due to a parasitic inductance L_{str} in the zone of the connection to the measurement equipment. The correction procedure is based on the equivalent circuit shown in Figure 5.

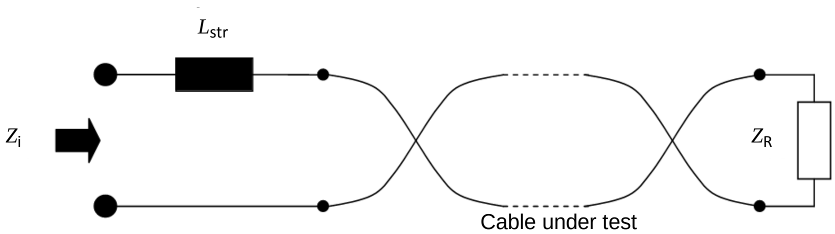

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Figure 5 – Equivalent circuit for the corrective calculation

The stray inductance L_{str} is calculated from the imaginary part of the complex input impedance Z_i at specific frequencies. The input impedance is then corrected using the stray inductance. From the corrected input impedance, a corrected return loss can be calculated.

The calculation of the corrected input impedance and the return loss data is done in the following steps:

For frequencies higher than 30 MHz up to the half of the maximum measurement frequency f_{max}, the value of stray inductance is calculated for every measurement frequency point:

$$L_{str} = \frac{\mathrm{Im}\,(Z_i)}{2\pi f} \tag{2}$$

where

f is the frequency, in Hz, and 30 MHz $\leq f \leq f_{max}/2$;

L_{str} is the stray inductance;

Z_i is the measured input impedance.

Then the mean value $L_{str,m}$ and the standard deviation L_{stad} of the stray inductance are calculated:

$$L_{\mathrm{str,m}} = \frac{1}{N}\sum_{n=1}^{N} L_{\mathrm{str,n}} \tag{3}$$

where

$L_{\mathrm{str,n}}$ is the stray inductance at the frequency corresponding with n;

N is the number of frequency points between 30 MHz and $f_{\mathrm{max}}/2$.

$$L_{\mathrm{stad}} = \sqrt{\frac{1}{N}\sum_{n=1}^{N}\left(L_{\mathrm{str,n}} - L_{\mathrm{str,m}}\right)^2} \tag{4}$$

The corrected input impedance $Z_{\mathrm{i,corr}}$ is calculated for every measurement frequency point:

$$Z_{\mathrm{i,corr}} = Z_{\mathrm{i}} - j2\pi f L_{\mathrm{str,m}} \tag{5}$$

The corrected return loss RL_{corr} is calculated for every measurement frequency point:

$$RL_{\mathrm{corr}} = 20\log\left|\frac{Z_{\mathrm{i,corr}} - Z_{\mathrm{R}}}{Z_{\mathrm{i,corr}} + Z_{\mathrm{R}}}\right| \tag{6}$$

where

Z_{R} is the reference impedance of the system.

To ensure the calculation is valid, the mean value stray inductance, the standard deviation of the stray inductance and the ratio between these two values shall be below the following limits:

a) $L_{\mathrm{str,m}} \leq$ 24 nH

b) $L_{\mathrm{str,m}} \geq$ 0 nH

c) $L_{\mathrm{stad}} \leq$ 8 nH

d) $L_{\mathrm{stad}} / L_{\mathrm{str,m}} \leq$ 0,8

NOTE These conditions have been found by experimentations. Values outside these specified limits indicate that the measurement error is not only due to a stray inductance but other effects. For example a negative value of $L_{\mathrm{str,m}}$ indicates a capacitive effect. For high frequencies or when the sample preparation is not well done one can no longer assume a concentrated stray inductance but has to take into account the line transformation effect of the cable ends (see also fitted return loss procedure).

5.3 Gated return loss (*GRL*)

A way to eliminate mathematically the effects of the test fixtures and sample preparation is to use gating. The operational return loss is measured in the frequency domain and the results are converted to the time domain using an inverse discrete Fourier transformation (IDFT), see also IEC 62153-1-1. In the time domain one can observe the reflection peaks at the cable ends. The gate is set on the near end side just after the peak and on the right hand side just before the peak. After activating the gate, the peaks are eliminated and transforming back into the frequency domain results in the gated return loss.

The IFDT described in IEC 62153-1-1 describes the time domain low pass mode which has some specific requirements for the frequency range. The measured frequencies must be set so that the start frequency is equal to the frequency step between two successive frequencies. The advantage of the time domain low pass mode is that it allows determining the location and type of default.

However, in general, the frequency step between two successive frequencies is different from the start frequency. In this case, one has to use the time domain band pass mode. It allows identifying the location of the default but does not indicate whether the default is capacitive, inductive or resistive.

The used frequency range and number of measurement points results in the resolution and a maximum length that can be measured.

$$CUT_{\mathrm{max}} = \frac{1}{2} \cdot \frac{n-1}{f_{\mathrm{stop}} - f_{\mathrm{start}}} c_0 \cdot v_{\mathrm{r}} \tag{7}$$

$$\Delta x = \frac{1}{2} \frac{c_0 \cdot v_{\mathrm{r}}}{f_{\mathrm{stop}} - f_{\mathrm{start}}} \tag{8}$$

NOTE The factor ½ is due to the reflection measurement where the signal is propagating forward to the default location and then backward to the input.

where

Δx is the minimum resolution (physically in the cable);

n is the number of points in the frequency sweep;

CUT_{max} is the maximum physical length contained in the results;

c_0 is the speed of light in free space;

v_{r} is the relative propagation constant;

f_{start} is the lowest frequency of the measurement;

f_{stop} is the highest frequency of the measurement.

Let's consider an S/FTP cable having a relative propagation constant of 70 % ("worst case"). If the measurement is done in the frequency range from 1 MHz to 2 000 MHz using 1601 frequency points (maximum value available for common network analysers), we obtain:

CUT_{max}=84 m Δx=53 mm

This means that only samples with a length of up to 84 m can be measured and corrected by gating.

The graphs of Figure 6 show the measured and gating corrected results for an S/FTP cable in the frequency range up to 2 000 MHz. The upper two graphs show the two set points for the gating. After the gate is turned on, the values left respectively right from the set points are set to zero. The lower two graphs show the return loss and input impedance (magnitude) before and after gating correction. After gating, the impedance is well centred around a constant value and the return loss has significantly improved.

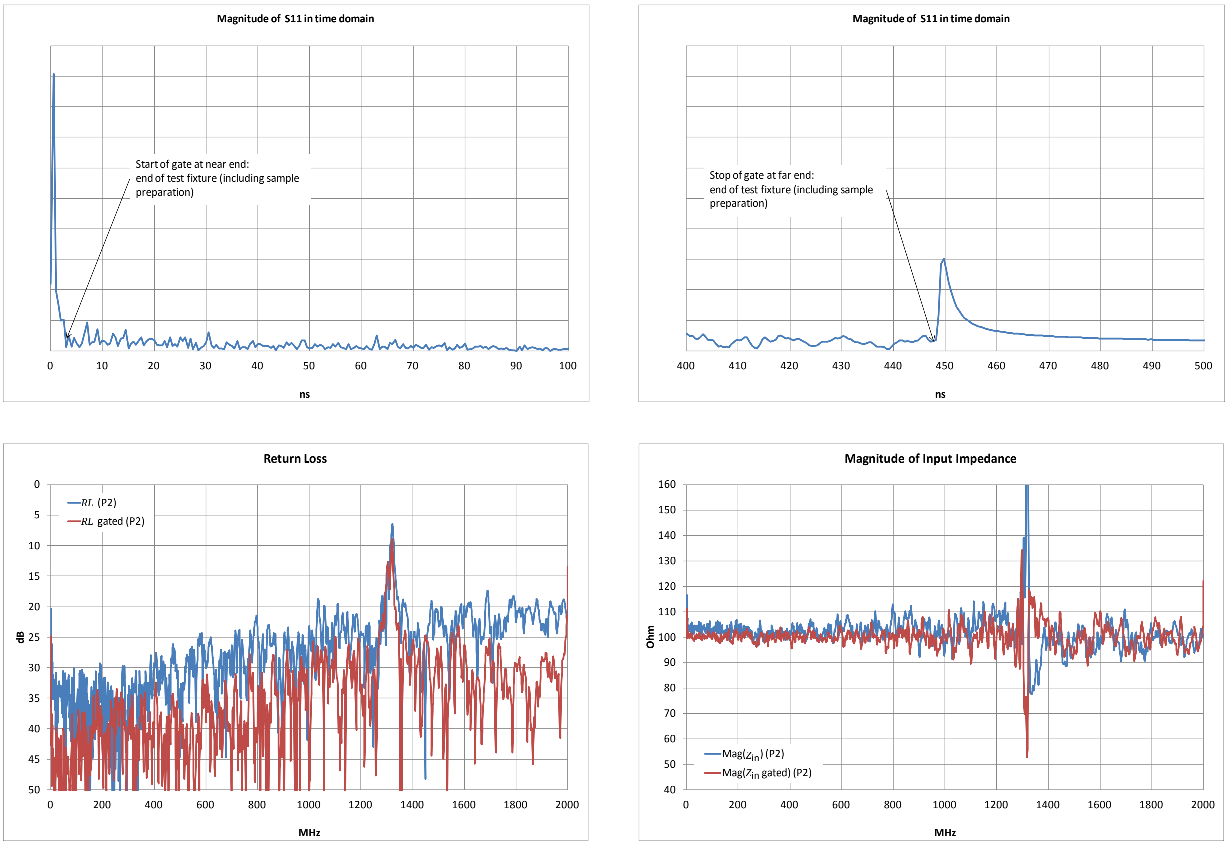

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Figure 6 – Return loss with gating correction

The difficulty with gating is to identify the end of the test fixtures (including the sample preparation) and thus where to set the beginning and end of the gate. Gating also distorts the end points of the gated frequency response and thus vector network analysers use a built in correction algorithm[1].

Gating also eliminates the effect of multiple reflections caused by a mismatch between the cable impedance and the reference impedance. A further study is given in Annex A.

5.4 Fitted return loss (*FRL*)

The correction procedure for fitted return loss is based on transmission line theory dealing with characteristic impedance discussed in 5.1 wherein an asymptotic value for characteristic impedance is assumed.

For the measurement up to and above 2 GHz, one may use a balun-less setup using a multiport (4 port) vector network analyser (VNA) with the option of modal decomposition (fixture simulator, soft balun). A full (4) port calibration is performed using an E-Cal module or equivalent methods, i.e. the test fixture needed to connect the CUT to the VNA is not included in the calibration routine of the VNA.

1 Further information is found in Agilent application note 1287-12 “Time Domain Analysis Using a Network Analyser.

The first step of the correction procedure is thus to eliminate the fixture by performing an $S11$ measurement of the fixture under open, short and load conditions. The fixture is corrected using the one port error model of a VNA. In a second step, the $S11$ of the CUT is measured, where the CUT is terminated with the reference impedance.

NOTE If the measurement is done with baluns, the fixture may be included in the calibration procedure of the VNA; in this case, the separate correction of the fixture is not necessary.

$$S11_{\text{CUT}} = \frac{S11_{\text{meas}} - EDF}{ESF \cdot \left(S11_{\text{meas}} - EDF\right) + ERF} \tag{9}$$

$$EDF = S11_{\text{fix,load}} \tag{10}$$

$$ESF = \frac{2 \cdot EDF - S11_{\text{fix,short}} - S11_{\text{fix,open}}}{S11_{\text{fix,short}} - S11_{\text{fix,open}}} \tag{11}$$

$$ERF = \left(EDF - S11_{\text{fix,short}}\right) \cdot \left(1 + ESF\right) \tag{12}$$

where

$S11_{\text{meas}}$ is the measured $S11$ of the CUT without fixture correction;

$S11_{\text{CUT}}$ is the $S11$ of the CUT after fixture correction;

$S11_{\text{fix,load}}$ is the $S11$ of the test fixture terminated with the reference impedance;

$S11_{\text{fix,short}}$ is the $S11$ of the test fixture terminated with a short circuit;

$S11_{\text{fix,open}}$ is the $S11$ of the test fixture terminated with an open circuit;

EDF is the effective directivity error;

ESF is the source match error;

ERF is the reflection signal path tracking error;

In the next step we calculate the input impedance after fixture correction and apply a fitting on the real and imaginary part of the measured input impedance. Investigations have shown that a proper fitting (for the purpose of the correction procedure) is best achieved using a two-parameter fitting (least square fit, see also IEC/TR 61156-1-2 or ASTM D4566) for frequencies up to 100 MHz.

$$Zin_{\text{CUT}} = Z_{\text{ref}} \cdot \frac{1 + S11_{\text{CUT}}}{1 - S11_{\text{CUT}}} \tag{13}$$

$$Zfit_{\substack{\text{real}\\\text{imag}}} = K_{0,\substack{\text{real}\\\text{imag}}} + \frac{K_{1,\substack{\text{real}\\\text{imag}}}}{\sqrt{f}} \tag{14}$$

where

Zin_{CUT} is the CUT complex input impedance after fixture correction;

$Zfit_{\substack{\text{real}\\\text{imag}}}$ is the fitted real respectively imaginary part of the input impedance after fixture correction;

Z_{ref} is the reference impedance of the set-up, i.e. the load resistance used during calibration;

$S11_{\text{CUT}}$ is the CUT complex $S11$ after fixture correction;

$K_{0,\substack{\text{real}\\\text{imag}}}$, $K_{1,\substack{\text{real}\\\text{imag}}}$	are the respective real and imaginary least squares fit coefficients, where $f_{\text{start}} \le f \le 100$ MHz;
f	is the frequency.

Now we calculate the real and imaginary part of the input impedance variation around the real and imaginary part of the fitted impedance respectively. The real and imaginary part of this residual impedance is then fitted (least square fit) using a polynomial function (equation 16).

NOTE It might be helpful to use the frequency in GHz rather than MHz (or Hz) to perform the matrix calculations used in the least square fit.

$$Z_{\text{residual},\substack{\text{real}\\\text{imag}}} = Zin_{\text{CUT},\substack{\text{real}\\\text{imag}}} - Zfit_{\substack{\text{real}\\\text{imag}}} \tag{15}$$

$$Zfit_{\text{residual},\substack{\text{real}\\\text{imag}}} = \sum_{n=1}^{10} a_{\text{n},\substack{\text{real}\\\text{imag}}} \cdot f^{\,n} \tag{16}$$

where

$Z_{\text{residual},\substack{\text{real}\\\text{imag}}}$	is the real respectively imaginary part of the residual impedance;
$Zfit_{\text{residual},\substack{\text{real}\\\text{imag}}}$	is the fitted real respectively imaginary part of the residual impedance;
$Zin_{\text{CUT},\substack{\text{real}\\\text{imag}}}$	is the CUT real and imaginary part of the complex input impedance after fixture correction;
$Zfit_{\substack{\text{real}\\\text{imag}}}$	is the fitted real respectively imaginary part of the input impedance after fixture correction;
$a_{\text{n},\substack{\text{real}\\\text{imag}}}$	are the respective real and imaginary least squares fit coefficients as indicated in equation (16);
f	is the frequency in GHz.

In the last step we obtain the fully corrected input impedance and return loss.

$$Zin_{\text{corr}} = Zin_{\text{CUT}} - Zfit_{\text{residual}} \tag{17}$$

$$RL_{\text{corr}} = 20 \cdot \log\left|\frac{Zin_{\text{corr}} - Z_{\text{ref}}}{Zin_{\text{corr}} + Z_{\text{ref}}}\right| \tag{18}$$

where

Zin_{corr}	is the input impedance (complex) after full correction;
Zin_{CUT}	is the CUT complex input impedance after fixture correction;
$Zfit_{\text{residual}}$	is the fitted complex residual impedance;
Z_{ref}	is the reference impedance of the set-up, i.e. the load resistance used during calibration;
RL_{corr}	is the fitted return loss after full correction.

The following graphs illustrate the different steps of the correction procedure. The measurements were done in a frequency range of 1 MHz to 2 000 MHz (1 601 points, linear sweep) on a 50 m sample of standard S/FTP CAT7A cable. The measurement was balun-less using a multiport VNA with mixed mode scattering parameter capabilities.

Figure 7 shows the return loss and input impedance before any correction. We obtain a poor return loss at high frequencies. The real and imaginary part of the input impedance show some big oscillations which are not related to the cable itself.

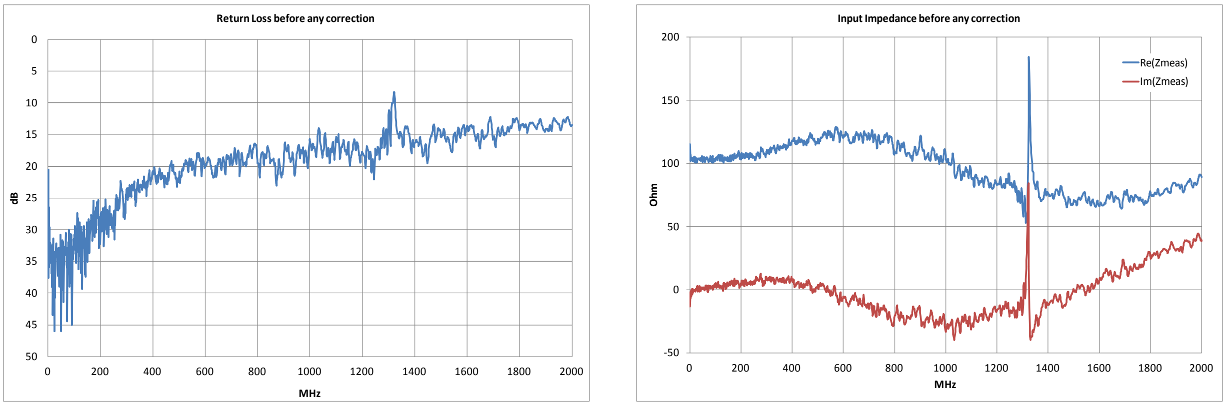

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Figure 7 – Return loss and input impedance before any correction

After elimination of the effect of the fixture (using the open, short and load method), the return loss is improved by some 5 dB and the oscillations of the real and imaginary part of the input impedance are removed. However, we now observe an important increase of the imaginary part and a slight decrease of the real part of the input impedance (see Figure 8).

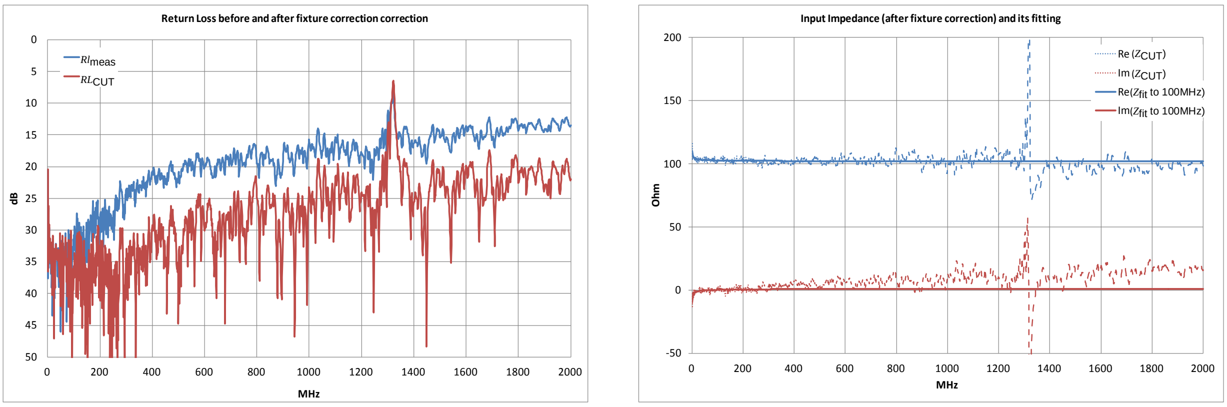

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Figure 8 – Return loss and input impedance (with fitting) after fixture correction

These abnormal tendencies become more visible if we analyse the residual impedance, which is the variation of the input impedance (real and imaginary parts) around the fitted input impedance (see Figure 9).

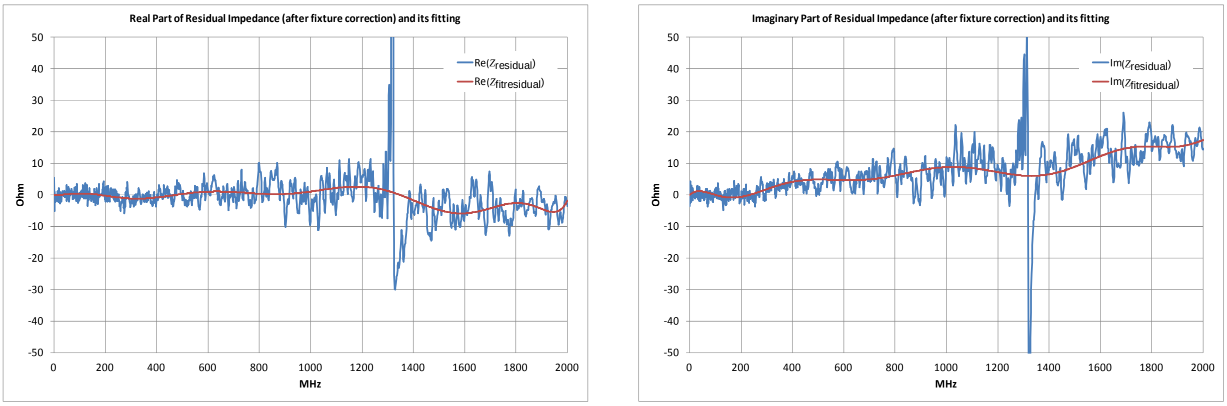

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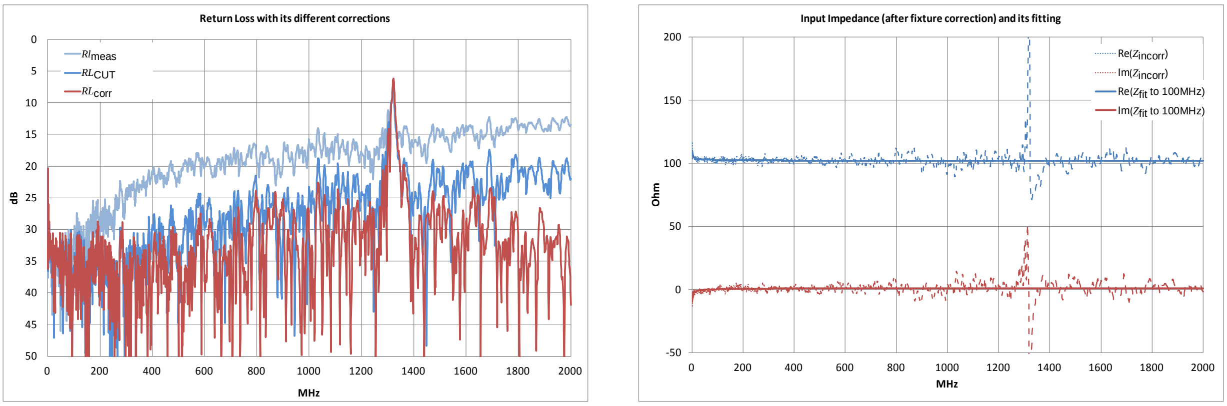
Figure 9 – Residual impedance (after fixture correction) and its fitting

To remove these abnormal tendencies, we subtract the fitted residual impedance from the input impedance. This results in a corrected input impedance which has no abnormal tendencies. From the corrected impedance we obtain the corrected return loss, which at high frequencies is improved by up to approximately 10 dB compared to the results after fixture correction (see Figure 10).

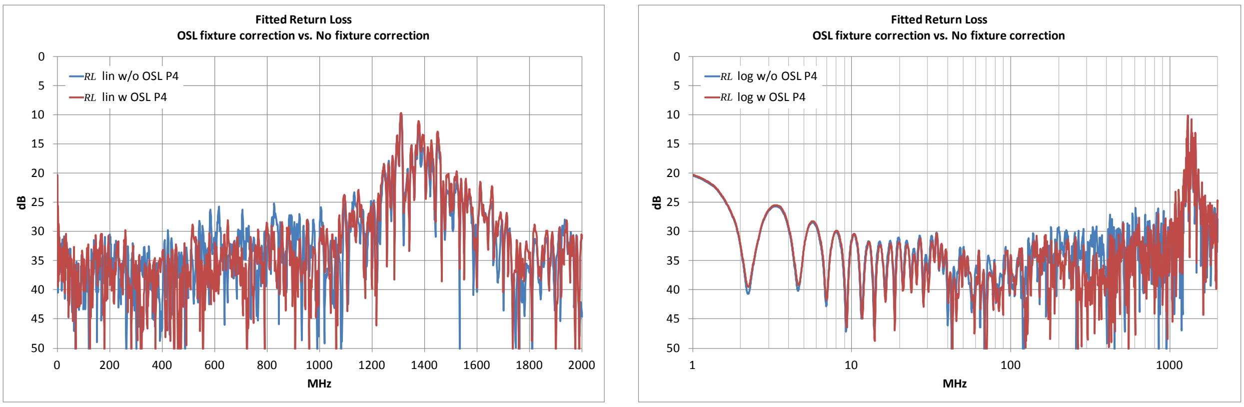


IEC 1621/13

Figure 10 – Return loss and input impedance after complete correction

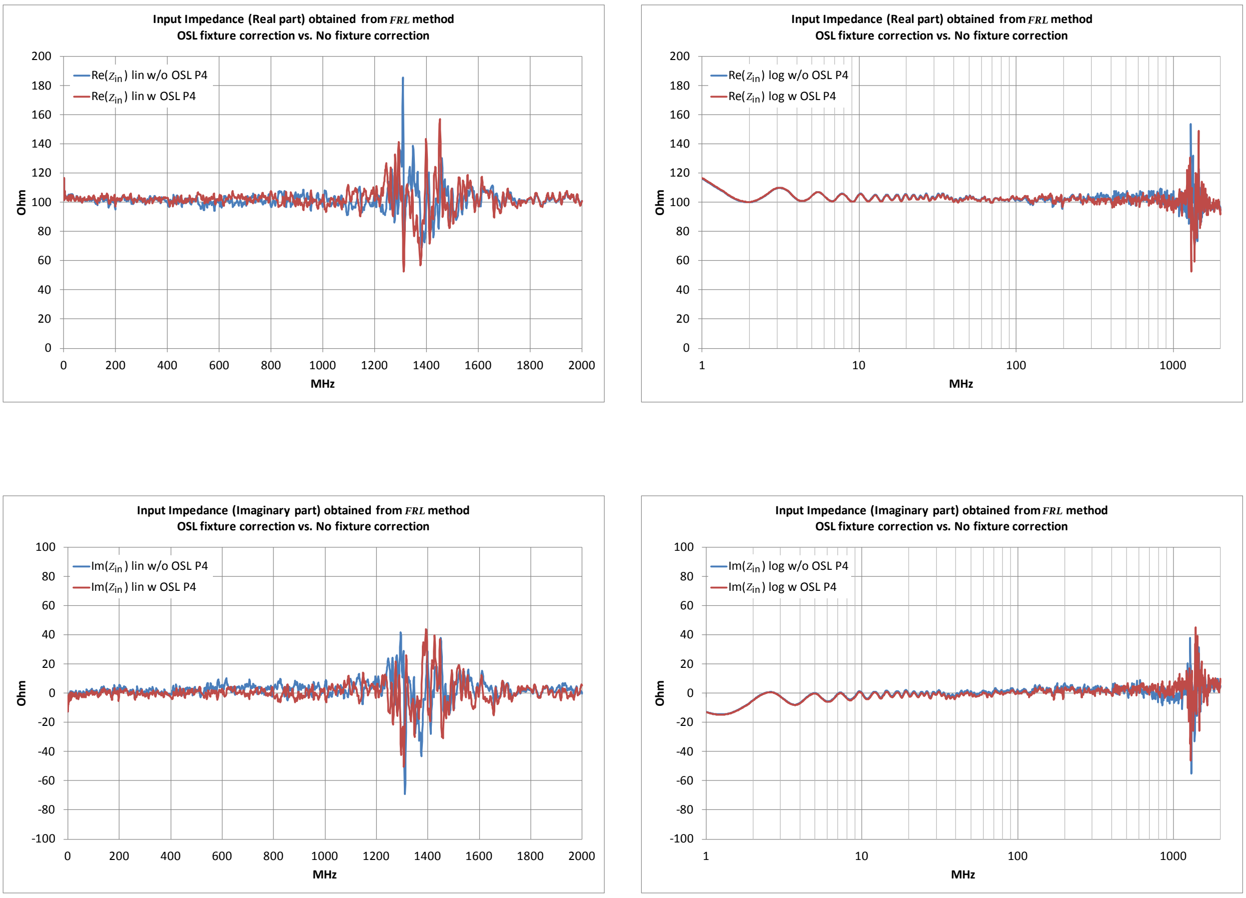
In principle, this correction procedure might be simplified by omitting the first correction of the test fixture (the open/short/load method). In fact, the “fitting” correction may be sufficient to eliminate in one step both effects, the one of the fixture and the one of the sample preparation. In addition to the simplification, one also gets rid of the uncertainty of the quality of the “balanced” open, short and load “standards”.

Figure 11 and Figure 12 compare the results of the return loss and input impedance (real and imaginary part) obtained from the fitted return loss method with fixture correction and without fixture correction for a S/FTP CAT7A cable. The agreement between both results is pretty good.



IEC 1622/13

Figure 11 – Return loss, with fixture correction vs. without fixture correction



IEC 1623/13

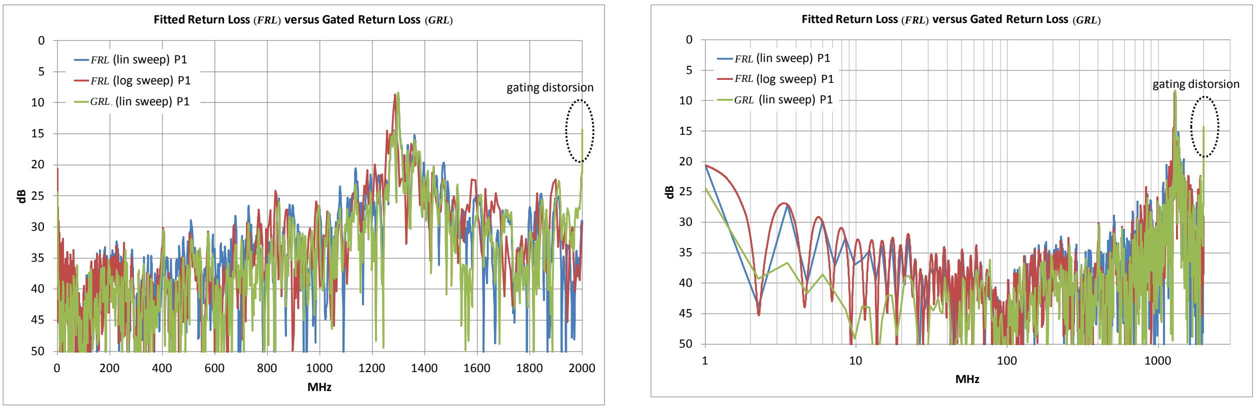
Figure 12 – Input impedance (real and imaginary part), with fixture correction vs. without fixture correction

Annex A
(informative)

Comparison of gated return loss (*GRL*) with fitted return loss (*FRL*)

The following graphs compare measurement results of return loss and input impedance comparing the fitting and gating correction procedure. Left-hand graphs show a linear frequency scale and right-hand graphs a logarithmic frequency scale. The results from the gating correction have been obtained in a linear sweep (a logarithmic sweep is not possible when using FFT). The results from the fitting correction procedure have been obtained with both a linear sweep (blue line) and logarithmic sweep (red line).

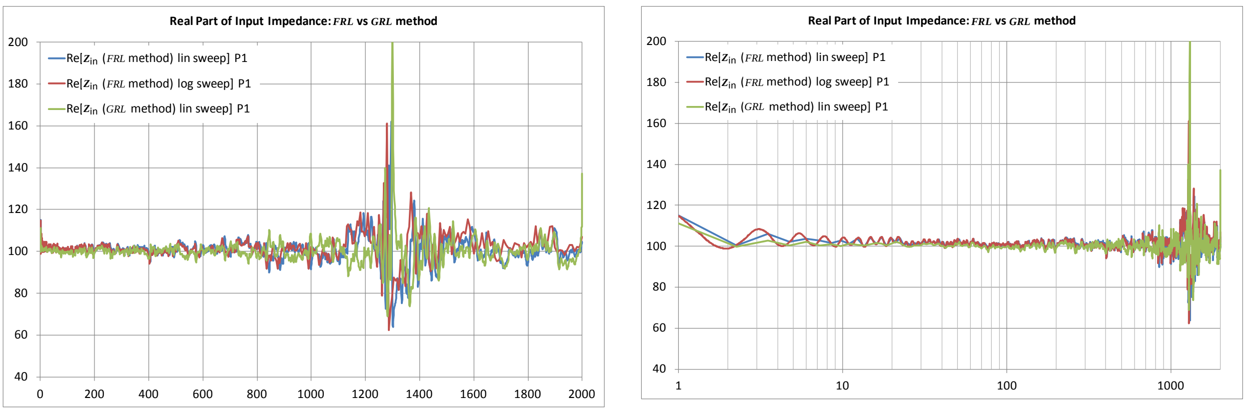
For the return loss measurements (see Figure A.1), we find a pretty good agreement between both correction procedures for frequencies above some 100 MHz. For the lower frequency range, we observe an important difference. This is due to the fact that gating eliminates the end (junction) effects and thus multiple reflections.



IEC 1624/13

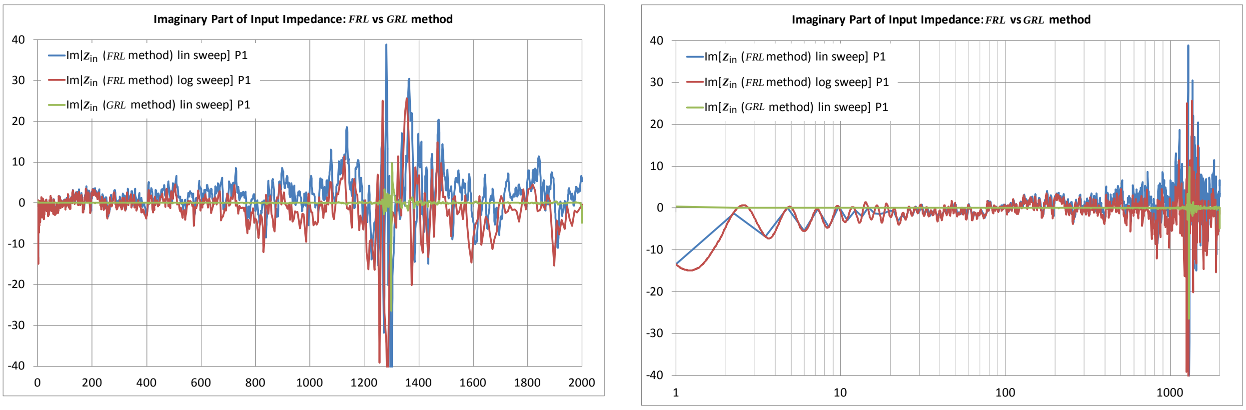
Figure A.1 – Return loss, fitting vs. gating correction

Comparing the input impedance obtained with both correction procedures reveals a good concordance of the real part over the whole frequency range (see Figure A.2), but a poor concordance for the imaginary part (see Figure A.3). The gating correction almost eliminates the imaginary part, i.e. using gating we lose the phase information.



IEC 1625/13

Figure A.2 – Real part of input impedance, fitting vs. gating correction



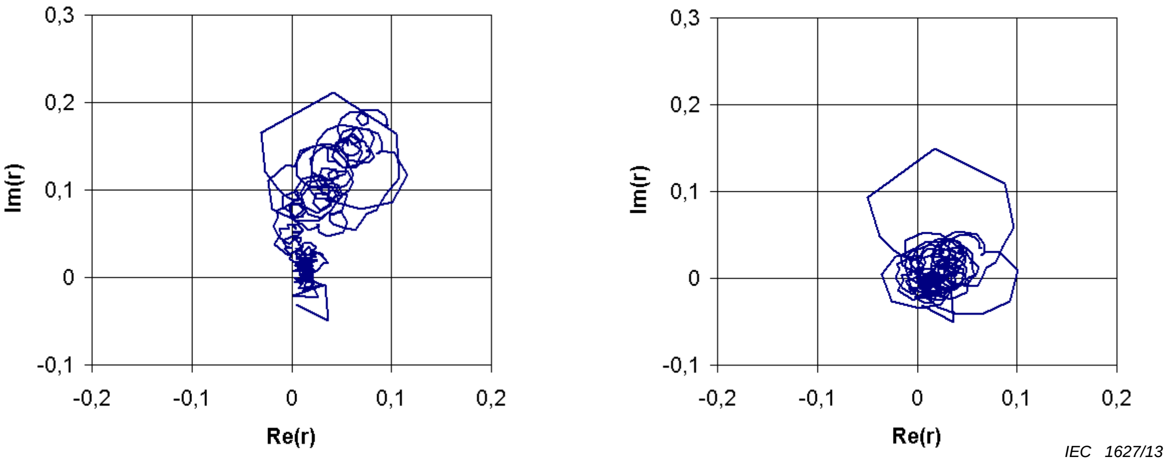
IEC 1626/13

Figure A.3 – Imaginary part of input impedance, fitting vs. gating correction

Annex B
(informative)

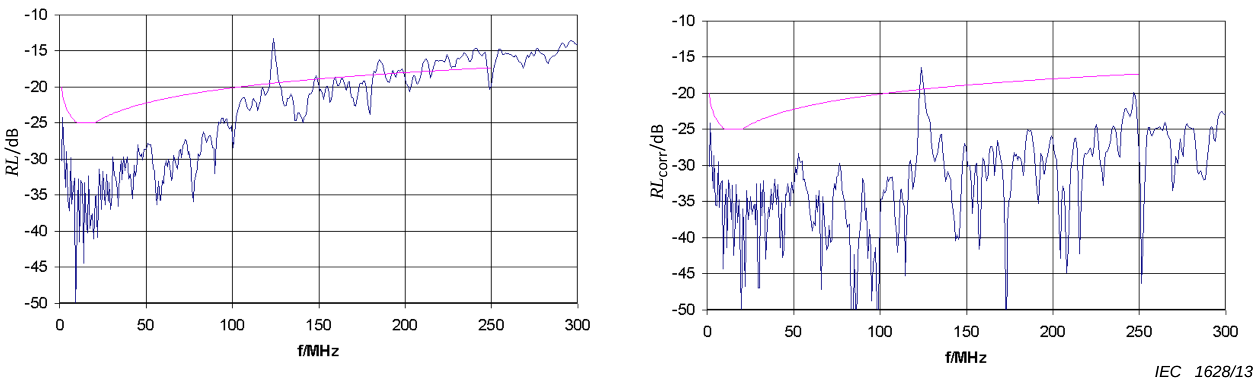
Influence of the correction technique on return loss peaks

Depending on the random point a cable is cut to measure return loss and input impedance, a possible return loss peak has at its maximum a specific phase angle. Therefore, the input impedance of the cable at the frequency can be affected by the stray inductance (see Figure 5) in a way that either the maximum peak value can be increased (the peak is inductive and the stray inductance is added) or decreased (the peak is capacitive and compensated by the stray inductance)[2]. This is shown in the following examples.



The reflection peak is inductive.

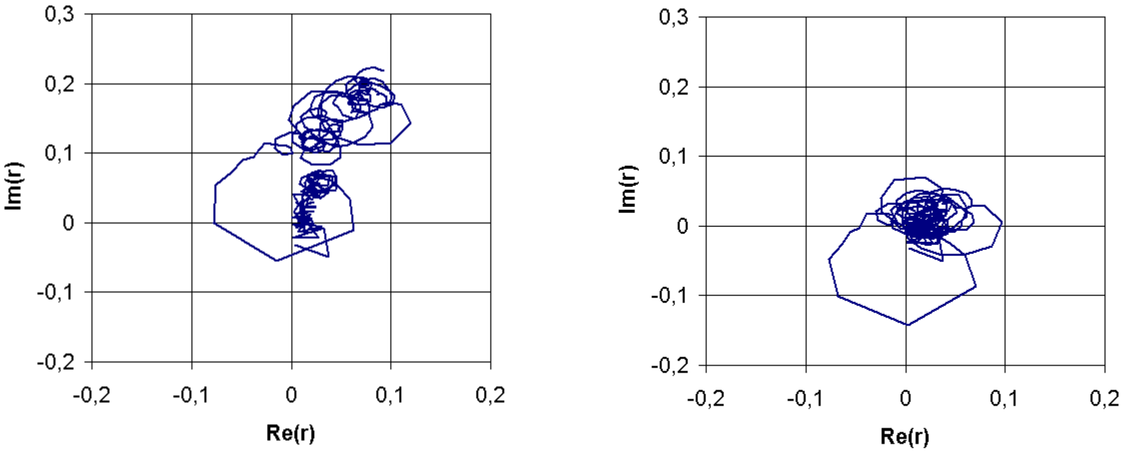
Figure B.1 – Reflection coefficient of a Cat.6 data cable in polar coordinates without and with *PRL*-correction



The correction decreases the peak value to the correct height.

Figure B.2 – Return loss traces corresponding to Figure B.1

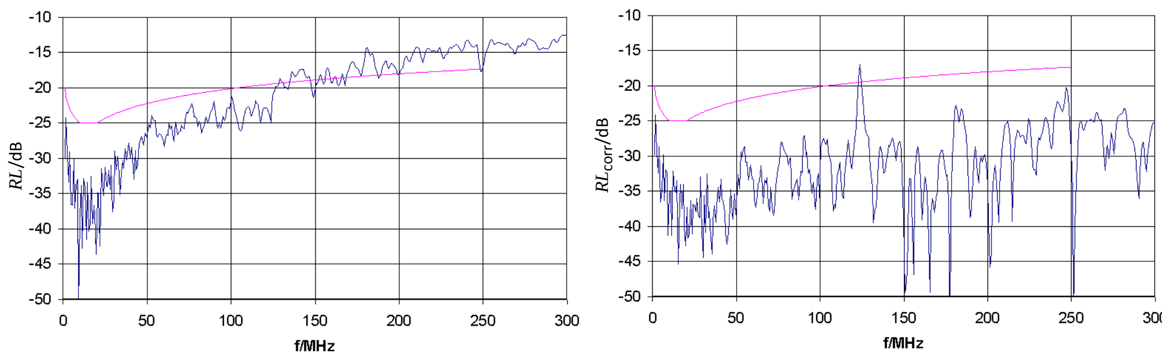
2 PFEILER, C.; WAßMUTH, A.: A Correction Technique for Reflection Measurement of Data Cables; proceedings of the 53rd IWCS 2004, pp. 23 - 26



IEC 1629/13

The reflection peak is capacitive.

Figure B.3 – Reflection coefficient of a Cat.6 data cable in polar coordinates without and with PRL-correction



IEC 1630/13

The correction increases the peak value to the correct height.

Without correction, the peak is invisible as the stray inductance compensates the capacitive peak.

Figure B.4 – Return loss traces corresponding to Figure B.3

From these figures, it can be seen that a correction technique is necessary to evaluate return loss peaks correctly.

INTERNATIONAL
ELECTROTECHNICAL
COMMISSION

3, rue de Varembé
PO Box 131
CH-1211 Geneva 20
Switzerland

Tel: + 41 22 919 02 11
Fax: + 41 22 919 03 00
info@iec.ch
www.iec.ch