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Edition 1.0 2023-11

TECHNICAL SPECIFICATION

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

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

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INTERNATIONAL ELECTROTECHNICAL COMMISSION

MULTICORE AND SYMMETRICAL PAIR/QUAD CABLES FOR DIGITAL COMMUNICATIONS –

Part 1-2: Electrical transmission characteristics and test methods of symmetrical pair/quad cables

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IEC TS 61156-1-2 has been prepared by subcommittee 46C: Wires and symmetric cables, of IEC technical committee 46: Cables, wires, waveguides, RF connectors, RF and microwave passive components and accessories. It is a Technical Specification.

This first edition cancels and replaces the first edition of IEC TR 61156-1-2 published in 2009 and Amendment 1:2014. This edition constitutes a technical revision.

This edition includes the following significant technical changes with respect to the previous edition (TR):

a) typos and editorial corrections;

b) the scope was updated;

c) Figure 14: ports swapped between port 2 and port 3;

d) new figures for balunless testing.

The text of this Technical Report is based on the following documents:

Draft	Report on voting
46C/1247DTS	46C/1259e/RVDTS

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

The language used for the development of this Technical Specification is English.

This document was drafted in accordance with ISO/IEC Directives, Part 2, and developed in accordance with ISO/IEC Directives, Part 1 and ISO/IEC Directives, IEC Supplement, available at www.iec.ch/members_experts/refdocs. The main document types developed by IEC are described in greater detail at www.iec.ch/publications.

A list of all parts of the IEC 61156 series, 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 document will remain unchanged until the stability date indicated on the IEC website under webstore.iec.ch in the data related to the specific document. At this date, the document will be

- reconfirmed,
- withdrawn,
- replaced by a revised edition, or
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MULTICORE AND SYMMETRICAL PAIR/QUAD CABLES FOR DIGITAL COMMUNICATIONS –

Part 1-2: Electrical transmission characteristics and test methods of symmetrical pair/quad cables

1 Scope

This part of IEC 61156 specifies symmetrical pair/quad electrical transmission characteristics and test methods present in IEC 61156-1:2002 (Edition 2) and not carried into IEC 61156-1:2007 (Edition 3). It details characteristic impedance test methods and function fitting procedures, the open/short-circuit method and the background of unbalance attenuation measurement.

It is extended by a description of the balunless measurements technique, which is an amendment to the former technical report and is improved and incorporated into this new edition. The complete document is transferred into a technical specification.

2 Normative references

The following documents are referred to in the text in such a way that some or all of their content constitutes requirements of this document. For dated references, only the edition cited applies. For undated references, the latest edition of the referenced document (including any amendments) applies.

IEC 60050-726, *International Electrotechnical Vocabulary (IEV) – Part 726: Transmission lines and waveguides*

IEC 61156-1:2007, *Multicore and symmetrical pair/quad cables for digital communications – Part 1: Generic specification*

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

3 Terms, definitions, symbols, units and abbreviated terms

3.1 Terms and definitions

For the purposes of this document, the terms and definitions given in IEC 60050-726, IEC 61156-1, IEC TR 62152 and the following apply.

ISO and IEC maintain terminological databases for use in standardization at the following addresses:

- ISO Online browsing platform: available at https://www.iso.org/obp
- IEC Electropedia: available at http://www.electropedia.org/

3.1.1
single-ended
measurement with respect to a fixed potential, usually ground

3.2 Symbols, units and abbreviated terms

For the purposes of this document, the following symbols, units and abbreviated terms apply.

Transmission line equation electrical symbols and related terms and symbols:

R	pair resistance (Ω/m)
L	pair inductance (H/m)
G	pair conductance (S/m)
C	pair capacitance (F/m)
α	attenuation coefficient (Np/m)
β	phase coefficient (rad/m)
γ	propagation coefficient (Np/m, rad/m)
v_{P}	phase velocity of cable (m/s)
v_{G}	group velocity of cable (m/s)
τ_{P}	phase delay time (s/m)
τ_{G}	group delay time (s/m)
Z_{C}	complex characteristic impedance (Ω)
$\angle Z_{\mathrm{C}}$	angle of the complex characteristic impedance in radians
Z_{∞}	high frequency asymptotic value of the complex characteristic impedance (Ω)
l	length (m)
j	imaginary denominator
Re	real part operator for a complex variable
Im	imaginary part operator for a complex variable
ω	radian frequency (rad/s)
f	frequency (Hz)
R'	first derivative of R with respect to ω
C'	first derivative of C with respect to ω
L'	first derivative of L with respect to ω
R_0	DC resistance of a round solid wire with radius r (Ω/m)
R_{C}	constant with frequency component of resistance which is about 1/4 of the DC resistance (Ω/m)
R_{S}	square-root of frequency component of resistance (Ω/m)
L_{E}	external (free space) inductance (H/m)
L_{I}	internal inductance whose reactance equals the surface resistance at high frequencies (H/m)
σ	specific conductivity of the wire material (S/m)
ρ	resistivity of the wire material (Ω/m^2)
μ	permeability of the wire material (H/m)
r	radius of the wire (m)
δ	skin depth (not to be confused with the dissipation factor tan δ) (m) $\delta = \frac{1}{\sqrt{\pi f \mu\sigma}}$

tan δ	dissipation factor $\tan\delta = \dfrac{G}{\omega C}$
q	forward echo coefficient at the far end of the cable at a resonant frequency
p	reflection coefficient measured from the near end of the cable at a resonant frequency, $p = 10^{\frac{-PSRL}{20}} = \left\| \dfrac{Z_{\mathrm{CM}} - Z_{\mathrm{C}}}{Z_{\mathrm{CM}} + Z_{\mathrm{C}}} \right\|$
A_{Q}	forward echo attenuation at a resonant frequency (dB) $A_{\mathrm{Q}} = -20\log\|q\|$
$PSRL$	structural return loss at a resonant frequency (dB) $PSRL = -20\log\|p\|$
K	$K = 2\alpha l - 1 \text{ when } 2\alpha l \gg 1 (\mathrm{Np})$
A_{Q}	$A_{\mathrm{Q}} = 2 \times PSRL - 20\log(2\alpha l - 1)$ (dB) where 2 αl is in Np
Z_{OC}	complex measured open circuit impedance (Ω)
Z_{SC}	complex measured short circuit impedance (Ω)
Z_{CM}	complex characteristic impedance as measured (with structure) (Ω) $Z_{\mathrm{CM}} = \sqrt{Z_{\mathrm{OC}} \times Z_{\mathrm{SC}}}$
Z_{MEAS}	complex measured impedance (open or short) (Ω)
Z_{IN}	input impedance of the cable when it is terminated by Z_{L} (Ω)
Z_{OUT}	output impedance of the cable when the input of the cable is terminated by Z_{G} (Ω)
Z_{N}	nominal (reference) impedance of the link and/or terminals (the system) between which the cable is operating (Ω)
Z_{R}	(nominal) reference impedance that is used in measurement. Normally (for actual return loss results), $Z_{\mathrm{R}} = Z_{\mathrm{N}}$. When using a return loss measurement to approximate SRL, it is practical to choose Z_{R} to give the best balance in the given frequency range (Ω)
Z_{T}	terminated impedance measurement made with the opposite end of the cable pair terminated in the reference impedance Z_{R} (Ω)
ς	reflection coefficient measured in the terminated measurement method $\varsigma = \dfrac{Z_{\mathrm{R}} - Z_{\mathrm{C}}}{Z_{\mathrm{R}} + Z_{\mathrm{C}}}$
Z_{G}	termination at the cable input when defining the output impedance of the cable Z_{OUT} (Ω)
Z_{L}	termination at the cable output when defining the input impedance of the cable Z_{IN} (Ω)
L_0, L_1, L_2, L_3	least squares fit coefficients for angle of the complex characteristic impedance
K_0, K_1, K_2, K_3	least squares fit coefficients of the complex characteristic impedance
$\|Z_{\mathrm{C}}\|$	fitted magnitude of the complex characteristic impedance (Ω)
$\|Z_{\mathrm{CM}}\|$	measured magnitude of the complex characteristic impedance (Ω)
$\angle(V_{\mathrm{1N}})$	input angle relative to a reference angle in radians

$\angle(V_{1F})$	output angle relative to the same reference angle in radians
k	multiple of 2π radians
m	indicator of matrix parameter
S_{11}	reflection coefficient measured with an S parameter test set
RL	return loss (dB)
SRL	structural return loss (dB)
CUT	cable under test

Unbalance attenuation electrical symbols:

TA	transverse asymmetry
LA	longitudinal asymmetry
R_1, R_2	resistance of one conductor per unit length (Ω)
L_1, L_2	inductance of one conductor per unit length (H)
C_1, C_2	capacitance of one conductor to earth (F)
G_1, G_2	conductance of one conductor to earth (S)
α_u	unbalance attenuation (dB)
T_u	unbalance coupling transfer function
Z_{com}	characteristic impedance of the common-mode circuit (Ω)
Z_{diff}	characteristic impedance of the differential-mode circuit (Ω)
Z_{unbal}	unbalance impedance (Ω)
x	length coordinate (m)
γ_{com}	propagation factor of the common-mode circuit (Np/m, rad/m)
γ_{diff}	propagation factor of the differential-mode circuit (Np/m, rad/m)
α_{diff}	operational differential-mode attenuation of the cable (dB)
α_{com}	operational common-mode attenuation of the cable (dB)
ΔR	resistance unbalance of the sample length (Ω)
ΔL	inductance unbalance of the sample length (H)
ΔC	capacitance unbalance to earth (F)
ΔG	conductance unbalance to earth (S)
S	summing function
U_{diff}	voltage in the differential-mode circuit (V)
U_{com}	voltage in the common-mode circuit (V)
n, f	index to designate the near end and far end, respectively

4 Basic transmission line formulae

4.1 Overview

A review of the relationships between the propagation coefficient, the complex characteristic impedance and the primary parameters R, L, G and C is useful here. Characteristic impedance is commonly thought of as being a magnitude quantity. While this concept can suffice for high frequency applications, this quantity is actually a complex one consisting of real and imaginary components or magnitude and angle. The associated propagation coefficient is readily viewed as being complex, consisting of the real attenuation and imaginary phase coefficient components. The four secondary components are readily related to the primary components. Frequency dependence of these parameters is also developed.

The cable pair parameters are represented as frequency domain dependent quantities. The measurement methods are based on frequency domain techniques. Measurement methods based on time domain techniques and combinations of time and frequency while useful in many cases are not covered here. The present-day availability of excellent frequency domain equipment such as the network analysers and impedance meters support the frequency domain approach.

4.2 Complex characteristic impedance and propagation coefficient formulae

4.2.1 General

The frequency domain of the complex characteristic impedance Z_C relates to the primary parameters as:

$$Z_C = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{1}$$

The propagation coefficient, γ, relates to the primary parameters as:

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \tag{2}$$

4.2.2 Propagation coefficient

4.2.2.1 Attenuation and phase coefficients

Formula (2) is separated into its real and imaginary parts, the attenuation coefficient α and the phase coefficient β:

$$\alpha = \sqrt{-\frac{1}{2}\left(\omega^2 LC - RG\right) + \frac{1}{2}\sqrt{\left(R^2 + \omega^2 L^2\right)\left(G^2 + \omega^2 C^2\right)}} \tag{3}$$

$$\beta = \sqrt{\frac{1}{2}\left(\omega^2 LC - RG\right) + \frac{1}{2}\sqrt{\left(R^2 + \omega^2 L^2\right)\left(G^2 + \omega^2 C^2\right)}} \tag{4}$$

Further, by factoring out $\omega\sqrt{LC}$ we obtain:

$$\beta = \omega\sqrt{LC}\sqrt{\frac{1}{2}\left(1 - \frac{R}{\omega L} - \frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}} \tag{5}$$

It can be shown that:

$$\alpha\beta = \omega\sqrt{LC}\left(\frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}}\right) \tag{6}$$

4.2.2.2 Formulae useful at high frequencies

From Formulae (5) and (6) we can solve for α and thus obtain for α and β the following expressions, valid within the entire frequency range:

$$\alpha = \frac{\frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}}}{\sqrt{\frac{1}{2}\left(1 - \frac{R}{\omega L} - \frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}}} \tag{7}$$

$$\beta = \omega\sqrt{LC}\sqrt{\frac{1}{2}\left(1 - \frac{R}{\omega L} - \frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}} \tag{8}$$

Formulae (7) and (8) are well suited for evaluation of high frequencies.

4.2.2.3 Formulae useful at low frequencies

For low frequency evaluations, the expressions given by Formulae (9) and (10) are suitable.

$$\alpha = \sqrt{\frac{\omega RC}{2}}\sqrt{\left(\frac{G}{\omega C} - \frac{\omega L}{R}\right) + \sqrt{\left(1 + \frac{\omega^2 L^2}{R^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}} \tag{9}$$

$$\beta = \sqrt{\frac{\omega RC}{2}}\sqrt{\left(\frac{\omega L}{R} - \frac{G}{\omega C}\right) + \sqrt{\left(1 + \frac{\omega^2 L^2}{R^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}} \tag{10}$$

4.2.3 Complex characteristic impedance

4.2.3.1 Real and imaginary parts

The complex characteristic impedance Z_C can also be separated into its real and imaginary parts as developed in Formulae (11) and (12).

$$Z_C = \text{Re}\, Z_C + j\,\text{Im}\, Z_C = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = \frac{\alpha + j\beta}{G + j\omega C} \tag{11}$$

$$Z_C = \frac{\frac{1}{\omega C}\left[\left(\beta + \frac{G}{\omega C}\alpha\right) - j\left(\alpha - \frac{G}{\omega C}\beta\right)\right]}{1 + \frac{G^2}{\omega^2 C^2}} \tag{12}$$

4.2.3.2 Formulae useful at high frequencies

After substituting Formulae (7) and (8) into Formula (12), the real and imaginary parts of the complex characteristic impedance are obtained as given in Formulae (13) and (14) respectively. These are well suited for simplification (see 4.3) at high frequencies:

$$\text{Re}\, Z_C = \frac{\sqrt{\frac{L}{C}}\left[\frac{1}{2}\left(1 - \frac{R}{\omega L}\frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}\right]}{\left(1 + \frac{G^2}{\omega^2 C^2}\right)\sqrt{\frac{1}{2}\left(1 - \frac{R}{\omega L}\frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}}} \tag{13}$$

$$-\text{Im}\, Z_C = \frac{\frac{R}{2\omega\sqrt{LC}} + \frac{G}{2\omega C}\sqrt{\frac{L}{C}} - \frac{G}{\omega C}\sqrt{\frac{L}{C}}\left[\frac{1}{2}\left(1 - \frac{R}{\omega L}\frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}\right]}{\left(1 + \frac{G^2}{\omega^2 C^2}\right)\sqrt{\frac{1}{2}\left(1 - \frac{R}{\omega L}\frac{G}{\omega C}\right) + \frac{1}{2}\sqrt{\left(1 + \frac{R^2}{\omega^2 L^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}}} \tag{14}$$

4.2.3.3 Formulae useful at low frequencies

On the other hand, by substituting Formulae (9) and (10) into Formula (12), the real and imaginary parts given in Formulae (15) and (16) respectively are obtained. These are useful for simplification in the low frequency range:

$$\text{Re}\, Z_C = \frac{\sqrt{\frac{R}{2\omega C}}\left[\sqrt{\frac{\omega L}{R} - \frac{G}{\omega C} + \sqrt{\left(1 + \frac{\omega^2 L^2}{R^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}} + \frac{G}{\omega C}\sqrt{\frac{G}{\omega C} - \frac{\omega L}{R} + \sqrt{\left(1 + \frac{\omega^2 L^2}{R^2}\right)\left(1 + \frac{G^2}{\omega^2 C^2}\right)}}\right]}{\left(1 + \frac{G^2}{\omega^2 C^2}\right)} \tag{15}$$

$$-\mathrm{Im}\,Z_C = \frac{\sqrt{\frac{R}{2\omega C}}\left[\sqrt{\frac{G}{\omega C}-\frac{\omega L}{R}+\sqrt{\left(1+\frac{\omega^2 L^2}{R^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}}-\frac{G}{\omega C}\sqrt{\frac{\omega L}{R}-\frac{G}{\omega C}+\sqrt{\left(1+\frac{\omega^2 L^2}{R^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}}\right]}{\left(1+\frac{G^2}{\omega^2 C^2}\right)} \tag{16}$$

4.2.4 Phase and group velocity

The phase propagation time (per unit length) is:

$$\tau_P = \frac{\beta}{\omega} \tag{17}$$

By introducing β from Formulae (8) and (10), we obtain:

$$\tau_P = \sqrt{LC}\sqrt{\frac{1}{2}\left(1-\frac{R}{\omega L}\frac{G}{\omega C}\right)+\frac{1}{2}\sqrt{\left(1+\frac{R^2}{\omega^2 L^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}} \tag{18}$$

and

$$\tau_P = \sqrt{\frac{RC}{2\omega}}\sqrt{\left(\frac{\omega L}{R}-\frac{G}{\omega C}\right)+\sqrt{\left(1+\frac{\omega^2 L^2}{R^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}} \tag{19}$$

The group propagation time (per unit length) is:

$$\tau_G = \frac{d\beta}{d\omega} \tag{20}$$

$$\tau_G = \frac{\beta}{\omega}+\frac{1}{2}\left(\frac{L'}{L}+\frac{C'}{C}\right)\beta+\frac{\omega^2 LC}{4\beta}\left[\left(-\frac{G}{\omega C}+\frac{\frac{R}{\omega L}\left(1+\frac{G^2}{\omega^2 C^2}\right)}{\sqrt{\left(1+\frac{R^2}{\omega^2 L^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}}\right)\frac{d\left(\frac{R}{\omega L}\right)}{d\omega}\right] \tag{21}$$

$$\left[+\left(-\frac{R}{\omega L}+\frac{\frac{G}{\omega C}\left(1+\frac{R^2}{\omega^2 L^2}\right)}{\sqrt{\left(1+\frac{R^2}{\omega^2 L^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}}\right)\frac{d\left(\frac{G}{\omega C}\right)}{d\omega}\right]$$

The phase and group velocities are, respectively,

$$v_\mathrm{P}=\frac{1}{\tau_\mathrm{P}} \tag{22}$$

$$v_\mathrm{G}=\frac{1}{\tau_\mathrm{G}} \tag{23}$$

The above expressions are accurate and valid within the whole frequency range. If C and $G/(\omega C)$ can be regarded as frequency independent coefficients, then we obtain:

$$\tau_\mathrm{G}=\frac{\beta}{\omega}+\frac{\beta}{2}\frac{L'}{L}+\frac{C}{4\beta}\left[-\frac{G}{\omega C}+\frac{\frac{R}{\omega L}\left(1+\frac{G^2}{\omega^2 C^2}\right)}{\sqrt{\left(1+\frac{R^2}{\omega^2 L^2}\right)\left(1+\frac{G^2}{\omega^2 C^2}\right)}}\right]\left(-R+R'\omega-\frac{L'R}{L}\omega\right) \tag{24}$$

The above expressions, which are valid within the entire frequency range, can be simplified into approximate expressions, which are valid at high or low frequencies only.

4.3 High frequency representation of secondary parameters

The high frequency representations of the formulae are useful over a broad range of frequencies extending from voice frequency on up because of the range of values for the dissipation factor. $G/(\omega C)$ = tan δ < 0,03 (< 3 %) even for PVC insulated cables up to 1,5 MHz and for the polyethylene (PE), insulation is exceedingly small at about 0,000 1 (0,01 %). This results in approximations, which in practice are valid for the whole frequency range as follows:

$$\mathrm{Re}\,Z_\mathrm{C}\approx\sqrt{\frac{L}{C}}\sqrt{\frac{1}{2}+\frac{1}{2}\sqrt{1+\frac{R^2}{\omega^2 L^2}}} \tag{25}$$

$$-\mathrm{Im}\,Z_\mathrm{C}\approx\frac{R}{2\omega C\,\mathrm{Re}\,Z_\mathrm{C}}-\frac{G}{\omega C}\mathrm{Re}\,Z_\mathrm{C}+\frac{\frac{G}{2\omega C}\frac{L}{C}}{\mathrm{Re}\,Z_\mathrm{C}} \tag{26}$$

$$\alpha \approx \frac{R}{2\mathrm{Re}\,Z_\mathrm{C}} + \frac{G\left(\sqrt{\frac{L}{C}}\right)^2}{2\mathrm{Re}\,Z_\mathrm{C}} \tag{27}$$

$$\beta \approx \omega C \mathrm{Re}\,Z_\mathrm{C} \tag{28}$$

$$\tau_\mathrm{P} \approx \sqrt{LC} \tag{29}$$

$$\tau_\mathrm{G} \approx \frac{\beta}{\omega} + \frac{\beta}{2}\frac{L'}{L} + \frac{C}{4\beta}\left(-\frac{G}{\omega C} + \frac{\frac{R}{\omega L}}{\sqrt{\left(1+\frac{R^2}{\omega^2 L^2}\right)}}\right)\left(-R + R'\omega - \frac{L'R}{L}\omega\right) \tag{30}$$

When also $R/(\omega L) < 0{,}1$, which is true for high frequencies (f > 1 MHz for 0,5 mm wire), the formulae holding better than about 1 % accuracy can be further simplified as shown below.

$$\mathrm{Re}\,Z_\mathrm{C} \approx \sqrt{\frac{L}{C}} \tag{31}$$

$$-\mathrm{Im}\,Z_\mathrm{C} \approx \frac{R}{2\omega C \mathrm{Re}\,Z_\mathrm{C}} - \frac{G}{\omega C}\mathrm{Re}\,Z_\mathrm{C} \approx \sqrt{\frac{L}{C}}\left(\frac{R}{2\omega L} - \frac{G}{2\omega C}\right) \tag{32}$$

$$\alpha \approx \frac{R}{2\mathrm{Re}\,Z_\mathrm{C}} + \frac{G}{2}\mathrm{Re}\,Z_\mathrm{C} \approx \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} \tag{33}$$

$$\beta \approx \omega C \mathrm{Re}\,Z_\mathrm{C} \approx \omega\sqrt{LC} \tag{34}$$

$$\tau_\mathrm{P} \approx \sqrt{LC} \tag{35}$$

$$\tau_G \approx \tau_P + \frac{\beta}{2}\frac{L'}{L} + \frac{C}{4\beta}\left(-\frac{G}{\omega C} + \frac{R}{\omega L}\right)\left(-R + R'\omega - \frac{L'R}{L}\omega\right) \quad (36)$$

4.4 Frequency dependence of the primary and secondary parameters

4.4.1 Resistance

The high frequency resistance (surface resistance) of a solid round wire for frequencies where the wire radius r is greater than twice the skin depth δ can be regarded as consisting of two parts where one is constant and the other $f^{0,5}$ dependent.

$$R = R_C + R_S = R_C + \rho\sqrt{\omega} \approx R_0\left(\frac{1}{4} + \frac{r}{2\delta}\right) \quad (37)$$

$$\rho = \frac{R_S}{\sqrt{\omega}} = \frac{R_0{}^r}{4}\sqrt{2\mu\sigma} \quad (38)$$

The above is true for a solid wire alone. In a pair, the proximity effects and the presence of other pairs and screen contribute both to the resistance and inductance. These effects can increase the R by about 15 % at 1 MHz and follow also approximately the square-root of frequency law. Also, the constant component of resistance while often neglected is about 15 % of the frequency dependent component at 1 MHz for a 0,5 mm diameter copper pair.

4.4.2 Inductance

The total inductance consists also of two main components such that

$$L \approx L_E + L_I = L_E + \frac{R_S}{\omega} = L_E + \frac{\rho}{\sqrt{\omega}} \quad (39)$$

The external free space inductance is reduced by the proximity effect of the pair and the free space limiting effects of the nearby shield and/or other pairs. These inductive components are negative and fairly frequency independent at high frequencies.

4.4.3 Complex characteristic impedance

The complex characteristic impedance at high frequency asymptotic value Z_∞ is given by Formula (40).

$$Z_\infty = \sqrt{\frac{L_E}{C}} \quad (40)$$

The high frequency impedance formulae are given by Formulae (41) and (42):

$$\operatorname{Re} Z_C \approx \sqrt{\frac{L}{C}} \approx Z_\infty \left(1+\frac{R_S}{2\omega L_E}\right) = Z_\infty + \frac{\rho}{2\sqrt{L_E C}\sqrt{\omega}} \tag{41}$$

$$-\operatorname{Im} Z_C \approx \sqrt{\frac{L}{C}}\left(\frac{R}{2\omega L}-\frac{G}{2\omega C}\right) \approx \frac{R_C+\rho\sqrt{\omega}}{2\omega\sqrt{L_E C}\left(1+\frac{\rho}{2L_E\sqrt{\omega}}\right)} - \sqrt{\frac{L_E}{C}}\left(1+\frac{\rho}{2L_E\sqrt{\omega}}\right)\frac{\tan\delta}{2}$$
$$\approx \frac{R_C}{2\omega\sqrt{L_E C}} + \frac{\rho}{2\sqrt{L_E C}\sqrt{\omega}} - \frac{Z_\infty}{2}\left(1+\frac{L_1}{L_E}\right)\tan\delta \approx \frac{\rho}{2\sqrt{L_E C}\sqrt{\omega}} - \frac{Z_\infty}{2}\tan\delta \tag{42}$$

4.4.4 Attenuation coefficient

Using the above approximations with Formulae (31) through (36) results in the remaining formulae of this subclause:

$$\alpha \approx \frac{\left(R_C - \frac{\rho^2}{2L_E}\right)}{2Z_\infty} + \frac{\rho\sqrt{\omega}}{2Z_\infty} + \frac{\rho\sqrt{\omega}\tan\delta}{4Z_\infty} + \frac{\omega\sqrt{L_E C}\tan\delta}{2} \tag{43}$$

which is of the form:

$$\alpha \approx A + B\sqrt{\omega} + C\omega \tag{44}$$

where A, B and C are constants.

The first term of Formula (44) indicates that at the low end of the high frequency range the attenuation increases a little more slowly than the square-root-law. The first $\omega^{0,5}$ term in Formula (43) which is dominant in the high frequency attenuation formula also appears in the phase coefficient, Formula (45).

$$\beta \approx \omega\sqrt{LC} \approx \omega\sqrt{L_E C}\left(1+\frac{R}{2\omega L_E}\right) \approx \omega\sqrt{L_E C} + \frac{\rho\sqrt{\omega}}{2Z_\infty} \tag{45}$$

4.4.5 Phase delay and group delay

The phase and group delay are given in Formulae (46) and (47) respectively:

$$\tau_P \approx \sqrt{LC} = \sqrt{L_E C}\left(1+\frac{R}{2\omega L_E}\right) \approx \sqrt{L_E C} + \frac{\rho}{2Z_\infty\sqrt{\omega}} \tag{46}$$

$$\tau_{\mathrm{G}} = \tau_{\mathrm{P}} + \frac{\beta}{2}\frac{L'}{L} + \frac{C}{4\beta}\left(-\frac{G}{\omega C} + \frac{R}{\omega L}\right)\left(-R + R'\omega - \frac{L'R}{L}\omega\right) \approx \left(1 - \frac{R}{4\omega L}\right) - \frac{R}{8\omega L}\left(\frac{R}{\omega L} - \frac{G}{\omega C}\right) \approx$$
$$\tau_{\mathrm{P}}\left(1 - \frac{R}{4\omega L}\right) \approx \sqrt{L_{\mathrm{E}}C}\left(1 + \frac{R}{4\omega L_{\mathrm{E}}}\right) \approx \sqrt{L_{\mathrm{E}}C} + \frac{\rho}{4\sqrt{\omega}Z_{\infty}} \qquad (47)$$

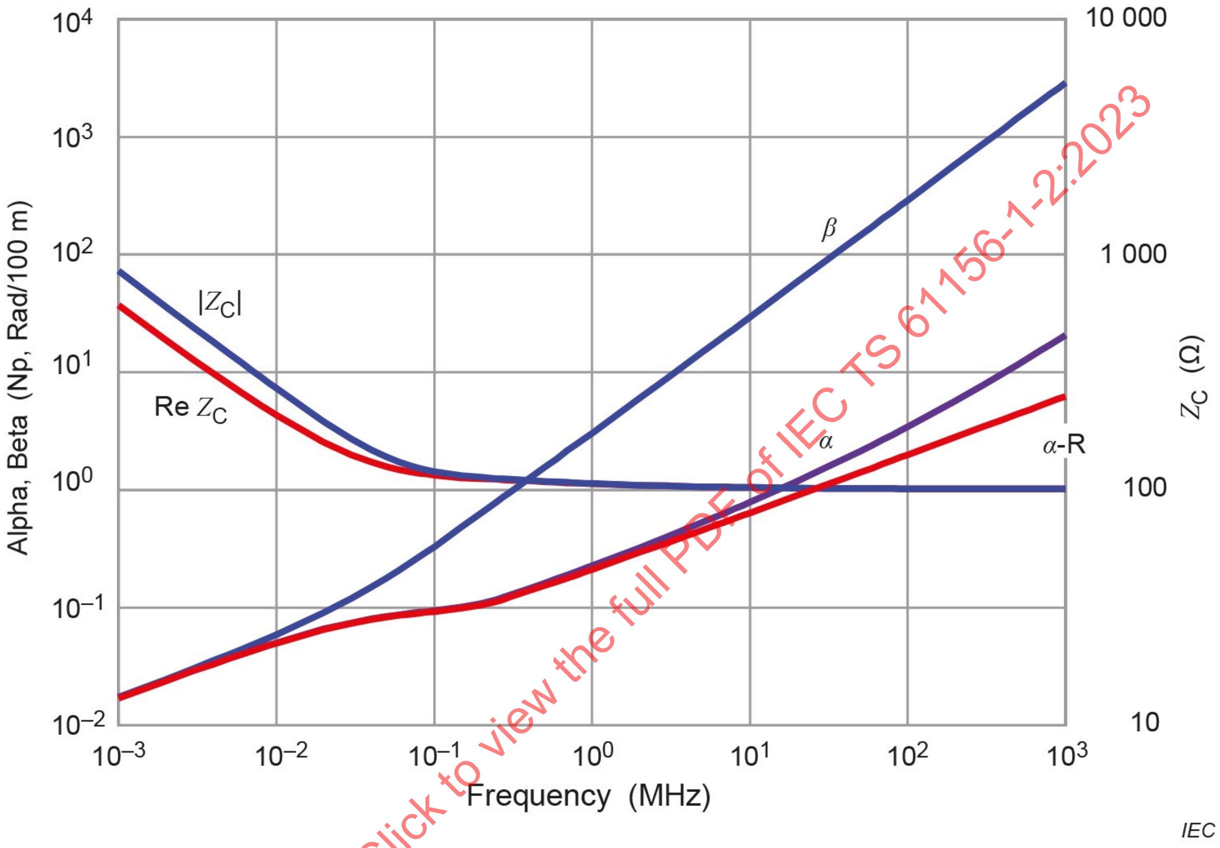


Figure 1 – Secondary parameters extending from 1 kHz to 1 GHz

Figure 1 shows the secondary parameters of a UTP pair with 0,5 mm conductors versus frequency. At voice frequencies, the attenuation and phase coefficients are equal. At these frequencies, the absolute value of the complex characteristic impedance and the real part of the complex characteristic impedance differ by the square root of 2. At frequencies above 100 kHz, attenuation is much less than the phase coefficient on the Nepers and radians scale, and the complex characteristic impedance is mostly real. The total attenuation (Alpha) differs from the conductor attenuation (Alpha-R) by the dielectric component of attenuation for this example, where the dissipation factor is assumed to be 0,01.

5 Measurement of the complex characteristic impedance

5.1 General

The complex characteristic impedance Z_C of a homogeneous cable pair is defined as the quotient of a voltage wave and current wave which are propagating in the same direction, either forwards (f) or backwards (r). For homogeneous cables (with no structural variations), the complex characteristic impedance can be measured directly as the quotient of voltage U and current I at the cable ends.

$$Z_C = \frac{U_f}{I_f} = \frac{U_r}{I_r} \tag{48}$$

A number of methods for obtaining complex characteristic impedance are described. Some of these methods offer convenience (at the cost of accuracy in portions of the frequency range). Others offer capability beyond what is currently needed for routine product inspection but are useful in laboratory evaluation where measurement throughput is not as critical.

The open/short circuit single-ended impedance measurement made with a balun in 5.2 is viewed as the reference method for obtaining the data. Alternative methods are listed below:

a) Complex characteristic impedance determined from phase coefficient and capacitance measurements (see 5.4);
b) terminated cable impedance measurements (see 5.5);
c) extended open/short impedance measurements excluding balun performance (see 5.6);
d) extended open/short impedance measurements made without a balun (see 5.7);
e) open/short impedance measurements at low frequencies with a balun (see 5.8);
f) impedance measurements obtained by modal decomposition technique (see 5.9).

It is intended that impedance measurements will be performed using sufficiently closely spaced frequencies so that impedance variation is represented. Either a linear sweep or a logarithmic sweep can be used depending on whether the high end or low end, respectively, of the desired frequency range is to be more fully represented. Typically, several hundred points (such as the available 401 points) are required depending on frequency range and cable length.

The balun used for connecting the symmetric cable pair to the coaxial port on the test instrument shall have a pass-band frequency range adequate for the desired measurement range. It shall be capable of transforming from the instrument port impedance to the nominal pair impedance. The three-step impedance measurement calibration is performed at the secondary (pair side) of the balun.

Function fitting Z^m (discussed in 5.3) of the impedance data is useful for separating structural effects from the complex characteristic impedance when such effects are substantial. When function fitting is used, the concept is that measurements from nearby frequencies aid in the interpretation of the values obtained at a particular frequency. Function fitting of the impedance magnitude or real part results in high values (typically 0,5 Ω or less) because of the positive and negative deviations not being symmetrical on the impedance scale. Function fitting can be conducted on the S-parameter values, which are linear responses, if more rigorous results (both impedance and SRL) are desired.

5.2 Open/short circuit single-ended impedance measurement made with a balun (reference method)

5.2.1 Principle

Open and short circuit measurements made with a balun from one end of a symmetric cable pair is the reference method for obtaining complex characteristic impedance values. The complex characteristic impedance is the geometric mean of the product of the open and short circuit measured values and is defined as:

$$Z_C = \sqrt{Z_{OC} Z_{SC}} \tag{49}$$

When the cable is not homogenous, an impedance inclusive of structural effects is obtained:

$$Z_{CM} = \sqrt{Z_{OC} Z_{SC}} \tag{50}$$

where Z_{CM} is the complex characteristic impedance together with structure (input impedance), expressed in ohms (Ω).

Formula (49) represents the complex characteristic impedance, Z_C, when structural effects are negligible. The fitting of the open/short impedance data with a complex characteristic impedance such as function of frequency can be employed to obtain Z_C from the input impedance, Z_{CM}, Formulae (50) when structural effects are substantial. Formulae (49) and (50) (and this measurement technique) are valid for frequencies extending from low values, where the cable length is only a fraction of a wavelength, to high frequencies where cable length represents many wavelengths.

5.2.2 Test equipment

A network analyser (together with an *S*-parameter unit) or an impedance meter can be used to obtain the data. Figure 2 shows the main components of an impedance measurement circuit where the generator and receiver are parts of the network analyser. An *S*-parameter unit, where the key component is the reflection bridge, is used with a network analyser to separate the reflected signal from the incident signal. A balun with the appropriate frequency range, impedance transformation (such as 50 Ω to 100 Ω for 50 Ω equipment and 100 Ω pair) and balanced at least as well as the pair under test facilitates making measurements on symmetric pairs under balanced conditions. Three terminating conditions, open, short and the nominal load resistance, are used as appropriate for the type of measurement being made (open, short or terminated).

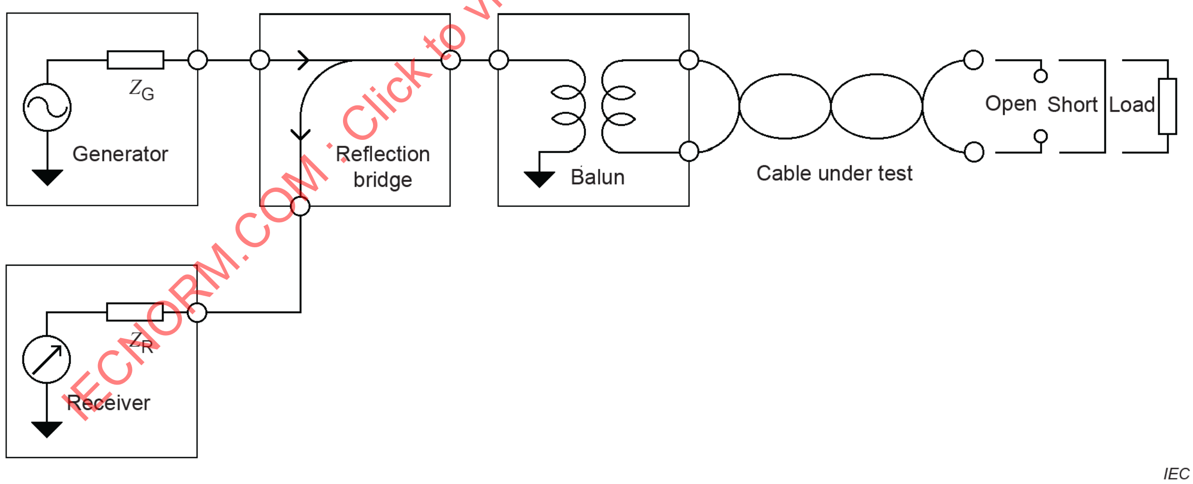


Figure 2 – Diagram of cable pair measurement circuit

5.2.3 Procedure

A three-step calibration procedure using the same open, short and load terminations as used for the actual measurements is conducted at the secondary of the balun with the cable pair disconnected. Upon completing the 3-step calibration procedure at the secondary of the balun, the network analyser is capable of directly measuring the complex reflection coefficient (*S*-parameter) or impedance of a cable pair. An internal 3-step calibration procedure including calculations is provided by most network analysers when an *S*-parameter unit is used. The method presented in 5.6 covers a similar 3-step calibration procedure by using the F-matrix principle where all the quantities are stated as impedances. This method is useful when the network analyser is not suitably equipped, in which case the computations can be accomplished external to the analyser.

The measured impedance (open or short) is computed from the reflection coefficient measurements S_{11} by means of Formula (51) either by the network analyser or by a computer (on acquired data):

$$Z_{\text{MEAS}} = Z_{\text{R}} \frac{1+S_{11}}{1-S_{11}} \quad (51)$$

5.2.4 Expression of results

Conceptually, several approaches are possible. On the one hand, the input impedance consisting of the combined complex characteristic impedance and structural effects can be viewed as needing to meet a broader single requirement (such as the 85 Ω to 115 Ω range) over the specified frequency range. Alternatively, a narrower range (such as a 95 Ω to 105 Ω range) can be viewed as being a requirement for the asymptotic component of function fitted complex characteristic impedance. In this case, RL or SRL specifications are used to control structural effects. The advantage of a broad single requirement in many instances is measurement simplification.

The advantage of separating the two effects is that of obtaining quantitative information for the two effects. The requirements for the impedance and structural effects are given in the relevant cable specification.

5.3 Function fitting the impedance magnitude and angle

5.3.1 General

Function fitting of the impedance data is useful for separating structural effects from the complex characteristic impedance when such effects are substantial. Where function fitting is used, the concept is that measurements from nearby frequencies aid in the interpretation of the values obtained at a particular frequency. Function fitting of the impedance magnitude or real part results in high values (typically 0,5 Ω or less), because of the positive and negative deviations not being symmetrical on the impedance scale. Function fitting can be conducted on the S-parameter values, which are linear responses, if more rigorous results (both impedance and SRL) are desired.

5.3.2 Impedance magnitude

5.3.2.1 Function fitting the magnitude of the complex characteristic impedance

While function fitting can be applied to the real and imaginary components of Z_{C}, the usual situation is that interest in the magnitude is greater than interest in the two separate components or the angle. The impedance magnitude tracks the real component closely at high frequencies where the imaginary component is small.

Function fitting of the impedance magnitude or real part results in fairly high values (typically 0,5 Ω or less), because of the positive and negative deviations not being symmetric on the impedance scale. Function fitting can be conducted on the S-parameter values, which is a linear response scale, if more rigorous results (both impedance and SRL) are desired.

This method differs from smoothing in that a complex characteristic impedance like function (based on transmission theory) is used to fit the measured data (obtained from Formula (50) or terminated impedance data). The function is stated as follows.

The fitted complex characteristic impedance magnitude is calculated with a least squares curve fit to Z_C, based on Formula (52):

$$\left|Z_C\right| = K_0 + \frac{K_1}{f^{\frac{1}{2}}} + \frac{K_2}{f} + \frac{K_3}{f^{\frac{3}{2}}} \tag{52}$$

Where terminated cable impedance data is used instead of open/short data, round-trip loss of measured length should be sufficiently large (in the 10 dB to 20 dB range for desired accuracies in the 5 Ω to 1,5 Ω range respectively when maximum deviation is 15 Ω – see 5.5).

Discreet point data equally spaced according to the log of frequency is advantageous for function fitting in that it results in appropriate weighting of the lower and upper ends of a multi-decade frequency sweep. Linear frequency spacing with logarithmic weighting can be used in the calculations when log of frequency spacing leads to concern about undersampling at high frequencies. Plotting the data versus the log of frequency is helpful here (as it is in network theory). The function fitting for individual data sets can readily be accomplished by importing ASCII format data obtained from the network analyser directly into a spreadsheet program and using the built-in regression procedures. Optimised software for analysing numerous data sets is desirable for use in a production setting.

The terms of the right-hand side of Formula (52) diminish in importance from left to right. The first two terms have strong theoretical basis. The constant term has the strongest basis in that it represents the space (external) inductance (largest component of inductance) and the capacitance of the pair (see Clause 4). The second term is significant in that it represents the component of complex characteristic impedance resulting from the internal inductance. The last two terms are supplied to provide for second order effects such as the capacitance decreasing with frequency, as with polar insulation materials or the effects of a shield. In the latter case, the low frequency end function fitting range is limited to frequencies where slope is increasing with frequency (2nd derivative positive).

The fit coefficients are calculated from Formula (53) where all summations are performed over N data points.

$$\begin{bmatrix} \sum_{i=1}^{N} \left|Z_{CM}\right| \\ \sum_{i=1}^{N} \frac{\left|Z_{CM}\right|}{\sqrt{f_i}} \\ \sum_{i=1}^{N} \frac{\left|Z_{CM}\right|}{f_i} \\ \sum_{i=1}^{N} \frac{\left|Z_{CM}\right|}{f_i^{3/2}} \end{bmatrix} = \begin{bmatrix} N & \sum_{i=1}^{N} \frac{1}{\sqrt{f_i}} & \sum_{i=1}^{N} \frac{1}{f_i} & \sum_{i=1}^{N} \frac{1}{f_i^{3/2}} \\ \sum_{i=1}^{N} \frac{1}{\sqrt{f_i}} & \sum_{i=1}^{N} \frac{1}{f_i} & \sum_{i=1}^{N} \frac{1}{f_i^{3/2}} & \sum_{i=1}^{N} \frac{1}{f_i^{2}} \\ \sum_{i=1}^{N} \frac{1}{f_i} & \sum_{i=1}^{N} \frac{1}{f_i^{3/2}} & \sum_{i=1}^{N} \frac{1}{f_i^{2}} & \sum_{i=1}^{N} \frac{1}{f_i^{5/2}} \\ \sum_{i=1}^{N} \frac{1}{f_i^{3/2}} & \sum_{i=1}^{N} \frac{1}{f_i^{2}} & \sum_{i=1}^{N} \frac{1}{f_i^{5/2}} & \sum_{i=1}^{N} \frac{1}{f_i^{3}} \end{bmatrix} \times \begin{bmatrix} K_0 \\ K_1 \\ K_2 \\ K_3 \end{bmatrix} \tag{53}$$

5.3.2.2 Obtaining log spaced data

Choose to acquire equally spaced data points on a log frequency basis when possible. This approach provides better weighting emphasis for data spanning several decades. Most network analysers offer this type of sweep. Convert the data being fitted to log spacing by interpolation when it is equally spaced on a linear frequency scale. Alternatively, use $1/f$ weighting (this means weighting a 10 MHz data point by 0,1 when a 1 MHz data point is weighted by 1) in performing the summations to simulate log frequency spaced data points.

The 4th order system of equations and unknowns is solved by the computer, by using determinants or matrix inversion techniques.

5.3.2.3 Fewer terms

Depending on the measurement frequency range and the amount of structural variation, usage of one or more of the higher order terms is not justifiable. The contributions from the higher order terms are intended to be second order. Where the data spans one decade or less, only the first two terms (or only the constant term) can be justified. The resultant function fit is considered valid if it has a negative slope at low frequencies, is asymptotic at higher frequencies and is free of oscillation with frequency.

Two or three terms can be sufficient when the data spans only one or two decades of frequency. This is accomplished by discarding one or more lower rows of Formula (53) and the same number of rightmost columns of the square matrix. While a four-term fit is indicated by Formulae (52) and (53), in some cases fewer terms can suffice. It is shown in 4.4.2 that just associating the inductance variation of a cable pair with frequency, calls for the first two terms of Formula (52). This is particularly true when the low frequency range of the data being fitted extends below about 3 MHz. If the capacitance is changing with frequency as it does when polar dielectric material is present, more terms are justified.

Four criteria indicate use of fewer terms – check or have the computer program determine if the fitted function obtained by solving Formula (53) meets the following set of four criteria:

a) The fitted function, except when it is only a constant, has negative slope for frequencies below 3 MHz.
b) The 10 MHz fitted value is within the impedance range of +5 to −2 of the high frequency asymptote (fitted constant value).
c) The area under the fitted function supplied by the frequency dependent terms on a log frequency basis, exclusive of the constant area, is positive (constant component is not above the data).
d) The sum of the negative areas (those due to negative coefficients) is less than the total area due to the frequency dependent terms.

If all four criteria are not met, the number of terms in the function (Formula (52)) shall be reduced by one by omitting the highest order term. Otherwise, data spanning a wider range of frequencies and resulting in a better fit shall be obtained and fitted. The fit for impedance magnitude shall have a monotonic downward behaviour with increasing frequency and approach a high frequency asymptote to a reasonable extent.

5.3.2.4 Compute and plot fitted results

Compute values for the magnitude of the complex characteristic impedance, according to coefficients obtained from the fit at the desired frequencies and plot the results and/or tabulate the fitted results at specification frequencies as desired.

5.3.3 Function fitting the angle of the complex characteristic impedance

This is useful when the complex characteristic impedance is to be specified as a complex quantity. The fitting equation for the angle of the complex characteristic impedance, $\angle Z_C$, is given in Formula (54). The formula should contain the same powers of frequency as those being used for the magnitude of the complex characteristic impedance.

$$\angle Z_C = L_0 + \frac{L_1}{f^{\frac{1}{2}}} + \frac{L_2}{f} + \frac{L_3}{f^{\frac{3}{2}}} \qquad (54)$$

The coefficients for the impedance angle can be calculated with Formula (53).

Plot the results as desired.

NOTE This procedure is necessary only if the angle of the complex characteristic impedance is of interest or if structural return loss (SRL) is being calculated at frequencies low enough to result in a significant angle (degrees).

5.4 Complex characteristic impedance determined from measured phase coefficient and capacitance

5.4.1 General

The mean complex characteristic impedance (homogeneous line) at any frequency can be obtained from the ratio of propagation coefficient to shunt admittance. At high frequencies, the asymptotic value of the real part of Z_C can be obtained by dividing delay by capacitance. This method is expedient for dielectric materials which do not change with frequency (non-polar) permitting a readily obtained low frequency value of capacitance to represent the high frequency range but is more difficult to apply when the capacitance changes with frequency as it does for polar dielectric materials. It results in complex characteristic impedance values free of structural effects. Justification for this method is supplied in Clause 4.

5.4.2 Formulae for all frequencies case and for high frequencies

Complex characteristic impedance Z_C can be expressed as the propagation coefficient divided by the shunt admittance as given in Formula (55). This relationship holds at any frequency. Complex characteristic impedance is readily separated into the real and imaginary components when $G << \omega C$.

$$Z_C = \frac{\alpha + j\beta}{G + j\omega C} \approx \frac{\beta}{\omega C} - \frac{j\alpha}{\omega C} \qquad (55)$$

At high frequencies, where the imaginary component of impedance is small, and the real component and magnitude are the same, Formula (55) can be written as:

$$Z_C = \frac{\beta}{\omega C} = \frac{\tau_P}{C} \qquad (56)$$

$$-\mathrm{Im} Z_C \approx \frac{\alpha}{\omega C} \approx \mathrm{Re} Z_C - Z_\infty = \frac{\beta}{\omega C} - Z_\infty \qquad (57)$$

$$Z_\infty = \sqrt{\frac{L_E}{C}} \qquad (58)$$

5.4.3 Procedure for the measurement of the phase coefficient

5.4.3.1 General

The phase coefficient measurement procedure, in the situation where the complex characteristic impedance is desired, is similar to that outlined for attenuation measurement in IEC 61156-1:2007, 6.3.3.

5.4.3.2 Phase coefficient

The phase coefficient of a pair of conductors is a measure of the phase shift incurred by a sinusoidal signal as it propagates over a length of pair and is affected by the materials and geometry of the insulated conductors.

The phase coefficient, β, relates to the measurements as:

$$\beta = \angle\left(V_{1F}\right) - \angle\left(V_{1N}\right) + 2\pi k \tag{59}$$

The phase coefficient can be obtained as a result of the same measurement procedure used to obtain the attenuation (see IEC 61156-1:2007, 6.3.3) by using a network analyser (which measures vector quantities). For balanced pairs, the transmit and receive ports of the measurement instrument shall supply balanced voltage with respect to ground and balanced currents (commonly accomplished with a balun). Pairs under test shall be terminated in their nominal impedance ±1 %.

5.4.3.3 Determining multiplier *k*

The multiplier k in Formula (59) may be determined either by examining the analyser display or numerically with the aid of a computer.

5.4.3.4 Determining *k* by examination

To determine the multiplier k, examine the analyser display and interpret the acquired data over the range of frequencies as appropriate. The phase meter or network analyser normally yields only the difference between the first and second terms shown on the right-hand side of Formula (59). Figure 3 shows the total phase and the sawtooth representation obtained from a network analyser. When a network analyser is used, a trace of the phase coefficient cycling through the 2π radians (360°) range is displayed on a VNA display, facilitating the determination of k. A frequently used technique in the interactive mode is to start at a low frequency where $k = 0$, by counting the number of 2π to 0π traversals to obtain the value for k.

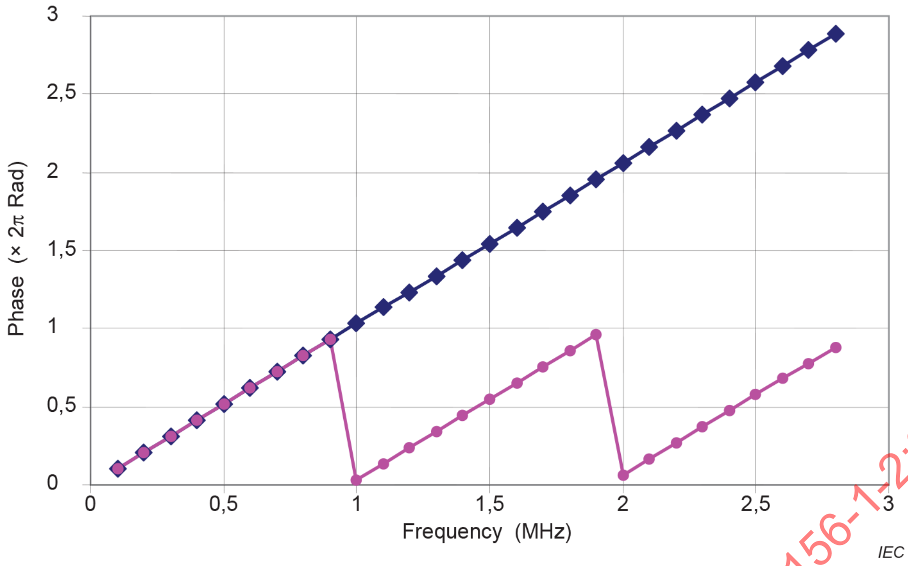


Figure 3 – Determining the multiplier of 2π radians to add to the phase measurement

5.4.3.5 Determining k numerically

Determine k numerically by acquiring the phase information obtained with the network analyser digitally using an interface with a digital computer as was done with the points plotted in Figure 3. Follow the data acquisition with a program procedure which starts by establishing a starting slope from several points in the $k = 0$ (multiple of 2π radian) frequency region. Let the program continue by examining each remaining point in succession. If the point is not within 2π radians of the continuous phase line being established, increment k until it is. This approach works even when intermediate values of k are passed over once the correct starting slope is established.

5.4.3.6 Obtaining total phase from the length function

To obtain the total phase, use the procedure called the "length" function, which is built into many network analysers. This internal procedure subtracts the specified length, which can be expressed as seconds of delay (actually a constant time frequency), from the internally established total delay and displays it. The phase trace is conveniently kept within the 0π to 2π (or alternately –π to +π) range over the whole frequency range by supplying the appropriate length value to the analyser.

5.4.4 Phase delay

Phase delay is a measure of the amount of time a simple sinusoidal signal is delayed when propagating through the length of a pair or cable. As with the phase coefficient, it is affected by the materials and geometry of the insulated conductors.

Formula (60) is used to calculate the phase delay τ_P, as a function of frequency from the phase coefficient β, measured in 5.4.3.2.

$$\tau_P = \frac{\beta}{\omega} \qquad (60)$$

5.4.5 Phase velocity

Phase velocity (reciprocal of phase delay) is a measure of the velocity with which a sinusoidal signal propagates through a cable and is normally reported in units of distance per second such as m/s.

Formula (61) is used to calculate the phase velocity v_P, as a function of frequency from the phase coefficient β, measured in 5.4.3.2.

$$v_P = \frac{\omega}{\beta} \tag{61}$$

NOTE Phase velocity is sometimes reported as a ratio consisting of the phase velocity divided by the velocity of light in a vacuum (c). It is then reported as, for example, 0,71 c, meaning 0,71 × speed of light. A variation is to report it as a percentage such as 71 %.

5.4.6 Procedure for the measurement of the capacitance

The capacitance of the same length as that measured for the phase coefficient (delay) shall be measured between the two conductors of the pair in accordance with IEC 61156-1:2007, 6.2.5.

5.5 Determination of the complex characteristic impedance using the terminated measurement method

A single terminated impedance measurement can be made in place of the open and short circuit measurements when the terminating impedance is sufficiently similar to the impedance being measured (within 15 Ω) and when the roundtrip loss of the measured length is sufficiently large (at least 10 dB). This measurement is useful when the convenience of using the network analyser in a stand-alone mode is desired. Use of this method is with the understanding that the open and short circuit method is the reference method.

Understanding the difference between the measured terminated impedance and the open/short circuit impedance is facilitated by the following formulae. The formula for the terminated input impedance Z_T is:

$$Z_T = Z_C \frac{1 + \varsigma e^{-2\gamma l}}{1 - \varsigma e^{-2\gamma l}} \tag{62}$$

where the reflection coefficient ς is given by:

$$\varsigma = \frac{Z_R - Z_C}{Z_R + Z_C} \tag{63}$$

Z_R and Z_C are the terminating impedance (usually a resistance) and the actual complex characteristic impedance, respectively. Having a closely matched termination or sufficient roundtrip attenuation is adequate for making the terminated measurement yield results close to those obtained by the open and short circuit method.

Formula (62) can be restated as follows:

$$Z_T - Z_C = (Z_R - Z_C)e^{-2\gamma l}\left(\frac{Z_T + Z_C}{Z_R + Z_C}\right) \quad (64)$$

Formula (62) indicates that a 15 Ω difference between the termination resistor and the cable impedance is reduced to a maximum error of approximately 5 Ω with a round trip loss of 10 dB. A 20 dB round trip loss ensures that a 15 Ω impedance difference is reduced to a minimal 1,5 Ω error.

5.6 Extended open/short circuit method using a balun but excluding the balun performance

5.6.1 Test equipment and cable-end preparation

The equipment required for the impedance and S-parameter measurement is that defined in 5.2. For this balanced form of measurement, the termination condition for other pairs and a shield, if present, is of little consequence. These conductors are close to ground even when permitted to float because of the pair twist of the pair under test. Letting these conductors float is acceptable.

5.6.2 Basic formulae

Complex characteristic impedance and the propagation coefficient are expressed in Formula (65) and Formula (66) respectively:

$$Z_C = \sqrt{Z_R{}^2\left(\frac{Z_{itr} - Z_{itf}}{Z_{itr} - Z_{its}}\right)^2\left(\frac{Z_{itcf} - Z_{its}}{Z_{itcf} - Z_{itf}}\right)\left(\frac{Z_{itcs} - Z_{its}}{Z_{itcs} - Z_{itf}}\right)} \quad (65)$$

$$\gamma = \alpha + j\beta = \frac{1}{l}\tanh^{-1}\sqrt{\left(\frac{Z_{itcf} - Z_{its}}{Z_{itcf} - Z_{itf}}\right)\left(\frac{Z_{itcs} - Z_{its}}{Z_{itcs} - Z_{itf}}\right)} \quad (66)$$

where

Z_{itf} is the input impedance measured by leaving the balanced output of the balun open (Ω);

Z_{its} is the input impedance measured by shorting the balanced output of the balun (Ω);

Z_{itr} is the input impedance measured by terminating the balanced output of the balun in a non-inductive, resistive load (Z_R Ω) which value is balanced to ±1 % (Ω);

Z_{itcf} is the input impedance measured by connecting the balanced output of the balun in a twisted pair with far end of the pair open (Ω);

Z_{itcs} is the input impedance measured by connecting the balanced output of the balun in a twisted pair with far end of the pair shorted (Ω).

5.6.3 Measurement principle

This principle describes an extended single end, open/short circuit method using a balun, but excluding the balun performance. The input impedance measurements are implemented by means of an impedance bridge or network analyser and S-parameter test set (see Figure 4 and Figure 5).

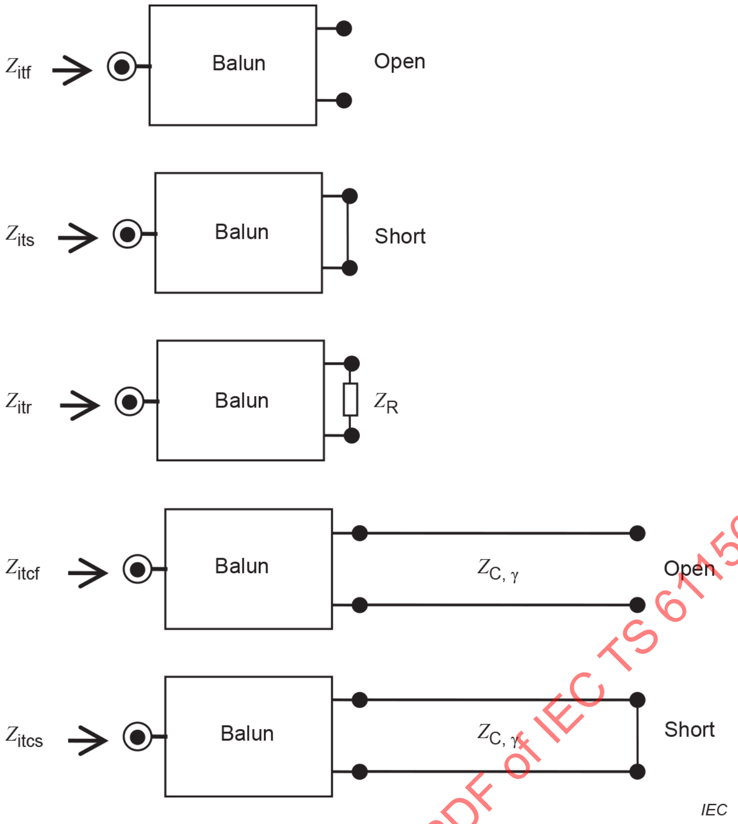


Figure 4 – Measurement configurations

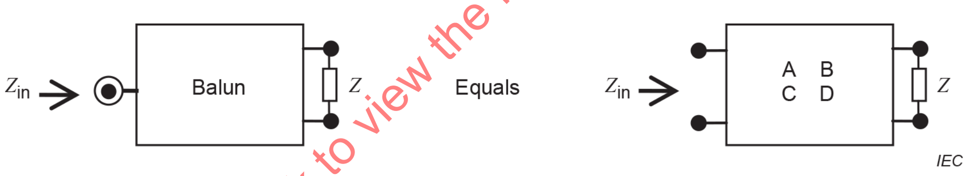


Figure 5 – Measurement principle with four terminal network theory

$$Z_{\text{in}} = \frac{AZ + B}{CZ + D} \tag{67}$$

where

Z_{in} is the input impedance;

Z is the load impedance such as open, short, termination, cable pair open or cable pair shorted.

$$Z_{\text{itf}} = Z_{\text{in}|Z=\infty} = \frac{A}{C}, A = Z_{\text{itf}} C \tag{68}$$

$$Z_{\text{its}} = Z_{\text{in}|Z=0} = \frac{B}{D}, B = Z_{\text{its}} D \tag{69}$$

$$Z_{\text{itr}} = Z_{\text{in}|Z=R} = \frac{AR+B}{CR+D} \tag{70}$$

$$Z_{\text{itcf}} = Z_{\text{in}|Z=Z_{\text{if}}} = \frac{AZ_{\text{if}}+B}{CZ_{\text{if}}+D} \tag{71}$$

$$Z_{\text{itcs}} = Z_{\text{in}|Z=Z_{\text{is}}} = \frac{AZ_{\text{is}}+B}{CZ_{\text{is}}+D} \tag{72}$$

where

Z_{if} is the impedance presented by cable pair with far end open (Ω);

Z_{is} is the impedance presented by cable pair with far end shorted (Ω).

Substituting Formula (68) and Formula (69) into Formula (70),

$$\frac{D}{C} = \frac{R\left(Z_{\text{itf}} - Z_{\text{itr}}\right)}{Z_{\text{itr}} - Z_{\text{its}}} \tag{73}$$

From Formula (71),

$$Z_{\text{if}} = \frac{B - Z_{\text{itcf}}D}{Z_{\text{itcf}}C - A} \tag{74}$$

From Formula (72),

$$Z_{\text{is}} = \frac{B - Z_{\text{itcs}}D}{Z_{\text{itcs}}C - A} \tag{75}$$

Finally:

$$Z_{\text{C}}^2 = Z_{\text{if}} Z_{\text{is}} = \left(\frac{B - Z_{\text{itcf}}D}{Z_{\text{itcf}}C - A}\right)\left(\frac{B - Z_{\text{itcs}}D}{Z_{\text{itcs}}C - A}\right) = \left(\frac{D}{C}\right)^2 \left(\frac{Z_{\text{itcf}} - Z_{\text{its}}}{Z_{\text{itcf}} - Z_{\text{itf}}}\right)\left(\frac{Z_{\text{itcs}} - Z_{\text{its}}}{Z_{\text{itcs}} - Z_{\text{itf}}}\right) =$$
$$R^2 \left(\frac{Z_{\text{itr}} - Z_{\text{itf}}}{Z_{\text{itr}} - Z_{\text{its}}}\right)^2 \left(\frac{Z_{\text{itcf}} - Z_{\text{its}}}{Z_{\text{itcf}} - Z_{\text{itf}}}\right)\left(\frac{Z_{\text{itcs}} - Z_{\text{its}}}{Z_{\text{itcs}} - Z_{\text{itf}}}\right) \tag{76}$$

$$\tanh^2 \gamma l = \frac{Z_{is}}{Z_{if}} = \left(\frac{Z_{itcf} - Z_{itf}}{Z_{itcf} - Z_{its}}\right)\left(\frac{Z_{itcs} - Z_{its}}{Z_{itcs} - Z_{itf}}\right) \quad \alpha \neq \beta \tag{77}$$

5.7 Extended open/short circuit method without using a balun

5.7.1 Basic formulae and circuit diagrams

Complex characteristic impedance and the propagation coefficient are defined by Formula (78) and Formula (79) respectively:

$$\frac{1}{Z_C} = \sqrt{\left(Y_{ff} - \frac{1}{4}Y_{uf}\right)\left(Y_{fs} - \frac{1}{4}Y_{us}\right)} \tag{78}$$

$$\gamma = \alpha + j\beta = \frac{1}{l}\tanh^{-1}\sqrt{\frac{\left(Y_{ff} - \frac{1}{4}Y_{uf}\right)}{\left(Y_{fs} - \frac{1}{4}Y_{us}\right)}} \tag{79}$$

where

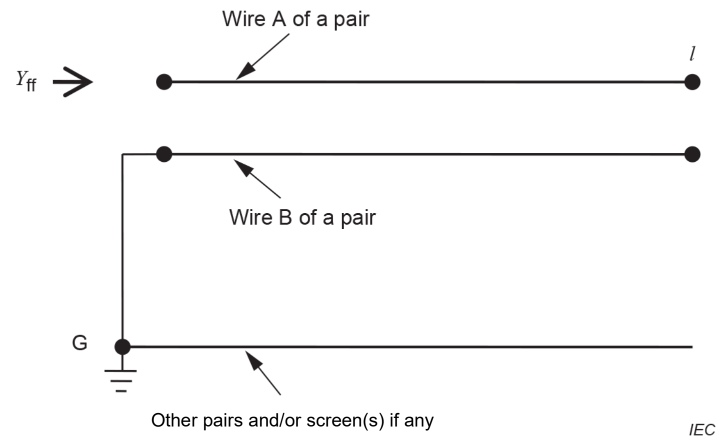
Y_{ff} is the admittance measured with measurement mode a (S);

Y_{fs} is the admittance measured with measurement mode b (S);

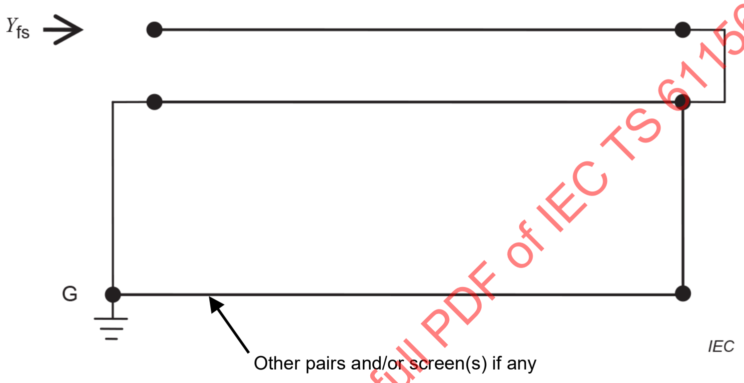
Y_{uf} is the admittance measured with measurement mode c (S);

Y_{us} is the admittance measured with measurement mode d (S).

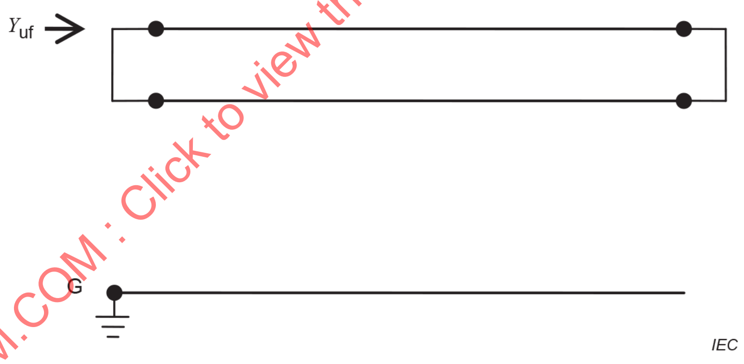
The measurement configurations are given in Figure 6.



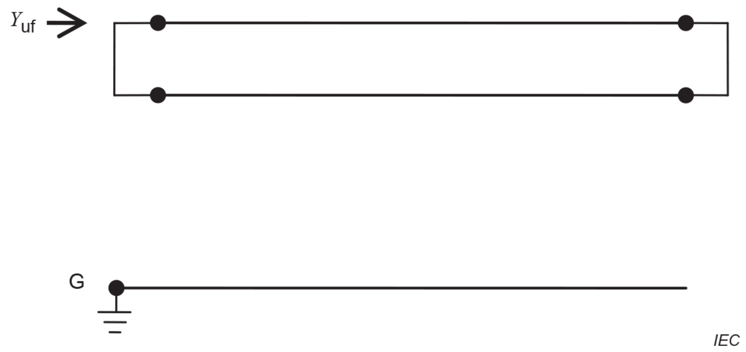
a) Measurement mode a: Y_{ff}



b) Measurement mode b: Y_{fs}



c) Measurement mode c: Y_{uf}



d) **Measurement mode d: Y_{us}**

Key

→ connecting inner conductor of unbalanced type measuring equipment

G connecting outer conductor of unbalanced type measuring equipment

The above set of four admittance measurement configurations assumes the pair is perfectly balanced. Generally, some degree of unbalance is present. This method can be used without additional measurements if the pair unbalance is less than 1 %.

Figure 6 – Admittance measurement configurations

5.7.2 Measurement principle

The measurement principle is given in Figure 7. The input admittance measurements are implemented by means of an impedance bridge or network analyser and S-parameter test set.

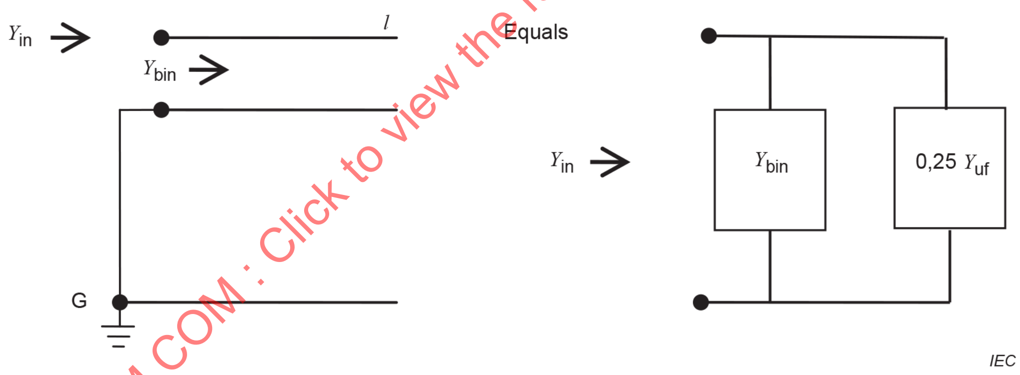


Figure 7 – Admittance measurement principle

For the open circuit case, the measured admittance is given by:

$$Y_{in} = Y_{bin} + \frac{1}{4} Y_u \tanh \gamma_u l = Y_{bin} + \frac{1}{4} Y_{uf} \tag{80}$$

where

γ_u is the unbalanced (common mode) propagation coefficient;

Y_u is the unbalanced (common mode) characteristic admittance;

Y_{bin} is the input admittance of the balanced circuit (open or short).

$$Y_{ff} = Y_{in}\big|_{Y_{bin}=Y_f} = Y_f + \frac{1}{4}Y_{uf} \tag{81}$$

$$Y_{fs} = Y_{in}\big|_{Y_{bin}=Y_s} = Y_s + \frac{1}{4}Y_{us} \tag{82}$$

where

Y_f is the balanced open circuit admittance;

Y_s is the balanced short circuit admittance.

From Formula (81),

$$Y_f = \frac{1}{Z_f} = Y_{ff} - \frac{1}{4}Y_{uf} \tag{83}$$

From Formula (82),

$$Y_s = \frac{1}{Z_s} = Y_{fs} - \frac{1}{4}Y_{us} \tag{84}$$

$$\frac{1}{Z_C} = Y_C = \sqrt{Y_f Y_s} = \sqrt{\left(Y_{ff} - \frac{1}{4}Y_{uf}\right)\left(Y_{fs} - \frac{1}{4}Y_{us}\right)} \tag{85}$$

$$\gamma = \alpha + j\beta = \frac{1}{l}\tanh^{-1}\sqrt{\frac{\left(Y_{ff} - \frac{1}{4}Y_{uf}\right)}{\left(Y_{fs} - \frac{1}{4}Y_{us}\right)}} \tag{86}$$

5.8 Open/short impedance measurements at low frequencies with a balun

For the measurement of the complex characteristic impedance of a cable, the open/short-circuit method can be applied, especially in the frequency range up to 1 MHz. An impedance measuring set with an accuracy of ±2 % is recommended.

The measurement is conducted at the relevant frequency by connecting the pair (or one side of the quad) at one end through a balun to the test set. At the other end, the conductors should be isolated (open-circuited) or short-circuited.

In the open-circuited condition:

$$Z_{\text{CO}} = R_{\text{L}} e^{j\psi_L} \tag{87}$$

In the short-circuited condition:

$$Z_{\text{CC}} = R_{\text{K}} e^{j\psi_K} \tag{88}$$

The modulus of the complex characteristic impedance is:

$$|Z| = [R_{\text{L}} R_{\text{K}}]^{\frac{1}{2}} \tag{89}$$

$$\text{Arg}|Z| = \frac{1}{2}(\psi_{\text{L}} + \psi_{\text{K}}) \tag{90}$$

The attenuation constant is derived from:

$$\alpha = \frac{8{,}686}{2l} \times \text{arctanh}\left[\frac{2\sqrt{\frac{R_{\text{K}}}{R_{\text{L}}}}}{1+\frac{R_{\text{K}}}{R_{\text{L}}}} \times \cos\left[\frac{1}{2}(\phi_{\text{K}} - \phi_{\text{L}})\right]\right] \quad \text{(dB/km)} \tag{91}$$

where l is the length of the cable under measurement (km).

The phase constant is derived from:

$$\beta = \frac{1}{2l}\left[\arctan\left[\frac{2\sqrt{\frac{R_{\text{K}}}{R_{\text{L}}}}}{1-\frac{R_{\text{K}}}{R_{\text{L}}}} \times \sin\left[\frac{1}{2}(\Phi_{\text{K}} - \Phi_{\text{L}})\right]\right] + n \times \pi\right] \quad \text{(rad/km)} \tag{92}$$

As the function arctan is ambiguous, the value of n needs to be determined. In practice, the following formula gives, in most cases, the exact value of n:

$$n = \text{integer}\left[\left|\left(\frac{1}{\pi}\right)\left(b - \frac{2\pi f Z_{\text{C}} C_3}{500}\right)\right| + 0{,}2\right] \tag{93}$$

where

C_3 is the mutual capacitance of the test specimen (nF).

$$b = \arctan\left[\frac{2\sqrt{\frac{R_K}{R_L}}}{1-\frac{R_K}{R_L}} \times \sin\left[\frac{1}{2}(\phi_K - \phi_L)\right]\right] \tag{94}$$

The phase velocity is derived from:

$$v = 2\pi f/\beta \tag{95}$$

5.9 Complex characteristic impedance and propagation coefficient obtained from modal decomposition technique

5.9.1 General

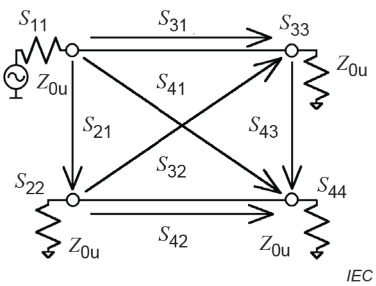
This more involved method results in data for the complex characteristic impedance and propagation coefficient if desired. Furthermore, it yields data for the unbalanced (common) mode as well as cross modal coupling. All combinations of S-parameters are measured using a conventional unbalanced instrument without the use of baluns, with other conductor ends terminated. The balanced- and unbalanced-mode components (impedance element of the matrix) are derived from the measured S-matrix by a mathematical operation ("mathematical balun" or "virtual balun").

5.9.2 Procedure

The procedure is as follows:

a) Calibrate the network analyser system. Full-2-port calibration is recommended.

b) Measure each element of the S-matrix of the Formula (96), e.g., S_{11}, S_{31} (S_{13}), and S_{33} are measured by connecting the one end of the conductor of the pair to the other port of the network analyser. All the rest of the ends of the conductors of the twisted pair, which are terminated to the receptacle of the standard connectors respectively, should be terminated with the standard dummy of the network analyser.



$$\begin{matrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{matrix} \tag{96}$$

c) Transform the S – matrix into the Z – matrix (Y – matrix) using the following formulae.

$$Z = Z_{0u}[E+S][E-S]^{-1} \tag{97}$$

$$Y = \frac{1}{Z_{0u}}[E-S][E+S]^{-1} \tag{98}$$

where

E is the unit matrix of 4 × 4;

Z_{0u} is the system impedance of a scalar value.

d) Once the impedance matrix is obtained, the complex characteristic impedance and the propagation coefficient for the balanced mode are calculated by Formulae (99) and (100):

$$Z_C = 2\sqrt{\frac{Z_{11}-2Z_{21}+Z_{22}}{Y_{11}-2Y_{21}+Y_{22}}} \tag{99}$$

$$\gamma = \frac{1}{2l}\ln\left(\frac{\frac{1}{2}\sqrt{(Z_{11}-2Z_{21}+Z_{22})(Y_{11}-2Y_{21}+Y_{22})}+1}{\frac{1}{2}\sqrt{(Z_{11}-2Z_{21}+Z_{22})(Y_{11}-2Y_{21}+Y_{22})}-1}\right) \tag{100}$$

5.9.3 Measurement principle

This method uses the modal decomposition theory, which has been established in the field of analysing a multiconductor system.

Notation of secondary coefficient: The secondary coefficient is expressed using an impedance matrix Z and an admittance matrix Y. The transmission line system illustrated in Figure 8 is presumed linear and symmetrical to show simple expression.

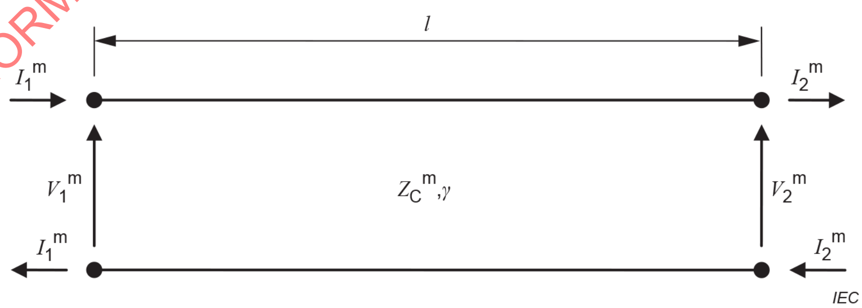


Figure 8 – Transmission line system

$$\begin{bmatrix} V_1^m \\ I_1^m \end{bmatrix} = \begin{bmatrix} \cosh(\gamma l) & Z_C^m \sinh(\gamma l) \\ \frac{\sinh(\gamma l)}{Z_C^m} & \cosh(\gamma l) \end{bmatrix} \begin{bmatrix} V_2^m \\ I_2^m \end{bmatrix} \tag{101}$$

When modified, the second part of the matrix formula is:

$$V_2^{\mathrm{m}} = \frac{Z_{\mathrm{C}}^{\mathrm{m}}}{\sinh(\gamma l)} I_1^{\mathrm{m}} - Z_{\mathrm{C}}^{\mathrm{m}} \coth(\gamma l) I_2^{\mathrm{m}} \tag{102}$$

Substituting this into Formula (102), the following impedance matrix is derived:

$$\begin{bmatrix} V_1^{\mathrm{m}} \\ V_2^{\mathrm{m}} \end{bmatrix} = \begin{bmatrix} Z_{\mathrm{C}}^{\mathrm{m}} \coth(\gamma l) & \dfrac{Z_{\mathrm{C}}^{\mathrm{m}}}{\sinh(\gamma l)} \\ \dfrac{Z_{\mathrm{C}}^{\mathrm{m}}}{\sinh(\gamma l)} & Z_{\mathrm{C}}^{\mathrm{m}} \coth(\gamma l) \end{bmatrix} \begin{bmatrix} I_1^{\mathrm{m}} \\ -I_2^{\mathrm{m}} \end{bmatrix} = \begin{bmatrix} Z_{11}^{\mathrm{m}} & Z_{21}^{\mathrm{m}} \\ Z_{21}^{\mathrm{m}} & Z_{11}^{\mathrm{m}} \end{bmatrix} \begin{bmatrix} I_1^{\mathrm{m}} \\ -I_2^{\mathrm{m}} \end{bmatrix} \tag{103}$$

Similarly, the admittance expression is derived:

$$\begin{bmatrix} I_1^{\mathrm{m}} \\ -I_2^{\mathrm{m}} \end{bmatrix} = \begin{bmatrix} \dfrac{\coth(\gamma l)}{Z_{\mathrm{C}}^{\mathrm{m}}} & \dfrac{-1}{Z_{\mathrm{C}}^{\mathrm{m}} \sinh(\gamma l)} \\ \dfrac{-1}{Z_{\mathrm{C}}^{\mathrm{m}} \sinh(\gamma l)} & \dfrac{\coth(\gamma l)}{Z_{\mathrm{C}}^{\mathrm{m}}} \end{bmatrix} \begin{bmatrix} V_1^{\mathrm{m}} \\ V_2^{\mathrm{m}} \end{bmatrix} = \begin{bmatrix} Y_{11}^{\mathrm{m}} & Y_{21}^{\mathrm{m}} \\ Y_{21}^{\mathrm{m}} & Y_{11}^{\mathrm{m}} \end{bmatrix} \begin{bmatrix} V_1^{\mathrm{m}} \\ V_2^{\mathrm{m}} \end{bmatrix} \tag{104}$$

Thus, we can get the secondary constants $Z_{\mathrm{C}}^{\mathrm{m}}$ and γ as:

$$Z_{\mathrm{C}}^{\mathrm{m}} = \sqrt{\frac{Z_{11}^{\mathrm{m}}}{Y_{11}^{\mathrm{m}}}}, \gamma = \frac{1}{l} \coth^{-1} \sqrt{Z_{11}^{\mathrm{m}} Y_{11}^{\mathrm{m}}} = \frac{1}{2l} \ln \left(\frac{\sqrt{Z_{11}^{\mathrm{m}} Y_{11}^{\mathrm{m}}} + 1}{\sqrt{Z_{11}^{\mathrm{m}} Y_{11}^{\mathrm{m}}} - 1} \right) \tag{105}$$

Because Z_{11}^{m} can be obtained by measuring the ratio of V_1^{m} to I_{11}^{m} with the other terminal opened, that is, by letting I_2^{m} = 0,

$$Z_{11}^{\mathrm{m}} = \left. \frac{V_1^{\mathrm{m}}}{I_1^{\mathrm{m}}} \right|_{I_2^{\mathrm{m}}=0} = Z_{\mathrm{C}}^{\mathrm{m}} \coth(\gamma l), Y_{11}^{\mathrm{m}} = \left. \frac{I_1^{\mathrm{m}}}{V_1^{\mathrm{m}}} \right|_{V_2^{\mathrm{m}}=0} = \frac{1}{Z_{\mathrm{C}}^{\mathrm{m}}} \coth(\gamma l) \tag{106}$$

thus, $Z_{11}^{\mathrm{m}^{\mathrm{m}}\mathrm{open}}$ and $Y_{11}^{\mathrm{m}^{\mathrm{m}}\mathrm{short}}$. This shows that Formula (105) is identical to those which are well known to us as formulae for the open/short method.

For the case of a twisted pair cable, the impedance and the admittance matrix in the modal domain shall be derived.

5.9.4 Scattering matrix to impedance matrix

5.9.4.1 General

The impedance and admittance matrices of the modal domain of the balanced mode can calculate the secondary constants of the pair.

The following three steps are required:

a) measure the scattering parameters of multi-conductor circuit;

b) calculate the impedance and admittance matrix (Z-matrix and Y-matrix respectively) from the scattering matrix (S-matrix); and

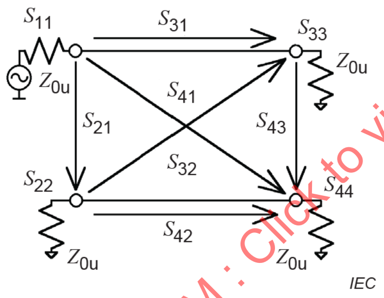
c) calculate the impedance and admittance of the balanced mode according to the modal decomposition theory.

5.9.4.2 Step 1: S-matrix measurement

The measurement is as follows.

a) Calibrate the network analyser system. Full 2-port calibration is recommended.

b) Measure each element of the S-matrix of Formula (107), e.g., S_{11}, S_{31}, S_{13}, and S_{33} are measured by connecting the respective end of the wires of the pair to the ports of the network analyser. All the rest of the ends of the conductors of the twisted pair, which are terminated to the receptacle of the standard connectors respectively, should be terminated with the standard dummy of the network analyser.



$$\begin{matrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{matrix} \quad (107)$$

5.9.4.3 Step 2: Transform S–matrix into Z–matrix

Transform the S – matrix into the Z – matrix (Y – matrix) using the following formulae:

$$Z = Z_{0u}\left[E+S\right]\left[E-S\right]^{-1}; Y = \frac{1}{Z_{0u}}\left[E-S\right]\left[E+S\right]^{-1} \quad (108)$$

where E is a unit matrix of 4 × 4, z_{0u} is the system impedance of a measuring equipment and is defined as a scalar value (typically 50 Ω system).

5.9.4.4 Step 3: Modal decomposition

According to the modal decomposition theory, the impedance matrix Z^{m} and the admittance matrix Y^{m} for a twisted pair cable can be obtained from the multi-conductor line circuit impedance (Z) and admittance (Y) as follows.

$$Z^{\mathrm{m}} = P^{-1}ZQ, Y^{\mathrm{m}} = Q^{-1}YP \tag{109}$$

where the diagonalising matrices P and Q are 4 × 4 real matrices and given as follows:

$$P = \begin{bmatrix} \frac{1}{2} & 1 & 0 & 0 \\ -\frac{1}{2} & 1 & 0 & 0 \\ 0 & 0 & \frac{1}{2} & 1 \\ 0 & 0 & -\frac{1}{2} & 1 \end{bmatrix}, Q = \begin{bmatrix} 1 & \frac{1}{2} & 0 & 0 \\ -1 & \frac{1}{2} & 0 & 0 \\ 0 & 0 & 1 & \frac{1}{2} \\ 0 & 0 & -1 & \frac{1}{2} \end{bmatrix} \tag{110}$$

When the line circuit is assumed to be linear, the matrices are symmetrical, and their expressions become:

$$Z^{\mathrm{m}} = \begin{bmatrix} Z_{11} - 2Z_{21} + Z_{22} & \frac{Z_{11} - Z_{22}}{2} & Z_{31} - Z_{32} - Z_{41} + Z_{42} & \frac{Z_{31} + Z_{41} - Z_{32} - Z_{42}}{2} \\ \frac{Z_{11} - Z_{22}}{2} & \frac{Z_{11} + 2Z_{21} + Z_{22}}{4} & \frac{Z_{31} - Z_{41} + Z_{32} - Z_{42}}{2} & \frac{Z_{31} + Z_{41} + Z_{32} + Z_{42}}{4} \\ Z_{31} - Z_{32} - Z_{41} + Z_{42} & \frac{Z_{31} + Z_{32} - Z_{41} - Z_{42}}{2} & Z_{33} - 2Z_{43} + Z_{44} & \frac{Z_{33} - Z_{44}}{2} \\ \frac{Z_{31} - Z_{32} + Z_{41} - Z_{42}}{2} & \frac{Z_{31} + Z_{32} + Z_{41} + Z_{42}}{4} & \frac{Z_{33} - Z_{44}}{2} & \frac{Z_{33} + 2Z_{43} + Z_{44}}{4} \end{bmatrix} \tag{111}$$

$$Z_{11}^{\mathrm{m}} = Z_{11} - 2Z_{21} + Z_{22} \tag{112}$$

$$Y^{\mathrm{m}} = \begin{bmatrix} \frac{Y_{11} - 2Y_{21} + Y_{22}}{4} & \frac{Y_{11} - Y_{22}}{2} & \frac{Y_{31} - Y_{41} - Y_{32} + Y_{42}}{4} & \frac{Y_{31} + Y_{41} - Y_{32} - Y_{42}}{2} \\ \frac{Y_{11} - Y_{22}}{2} & Y_{11} + 2Y_{21} + Y_{22} & \frac{Y_{31} - Y_{41} + Y_{32} - Y_{42}}{2} & Y_{31} + Y_{41} + Y_{32} + Y_{42} \\ \frac{Y_{31} - Y_{32} - Y_{41} + Y_{42}}{4} & \frac{Y_{31} + Y_{32} - Y_{41} - Y_{42}}{2} & \frac{Y_{33} - 2Y_{43} + Y_{44}}{4} & \frac{Y_{33} - Y_{44}}{2} \\ \frac{Y_{31} - Y_{32} + Y_{41} - Y_{42}}{2} & Y_{31} + Y_{32} + Y_{41} + Y_{42} & \frac{Y_{33} - Y_{44}}{2} & Y_{33} + 2Y_{43} + Y_{44} \end{bmatrix} \tag{113}$$

$$Y_{11}^{\mathrm{m}} = \frac{Y_{11} - 2Y_{21} + Y_{22}}{4} \tag{114}$$

The following formulae are derived from Formula (105).

$$Z_C^m = \sqrt{\frac{Z_{11}^m}{Y_{11}^m}} = 2\sqrt{\frac{Z_{11} - 2Z_{21} + Z_{22}}{Y_{11} - 2Y_{21} + Y_{22}}} \tag{115}$$

$$\gamma = \frac{1}{l}\coth^{-1}\sqrt{Z_{11}^m Y_{11}^m} = \frac{1}{2l}\ln\left(\frac{\sqrt{Z_{11}^m Y_{11}^m}+1}{\sqrt{Z_{11}^m Y_{11}^m}-1}\right) = \frac{1}{l}\coth^{-1}\left\{\left(Z_{11} - 2Z_{21} + Z_{22}\right)\left(\frac{Y_{11} - 2Y_{21} + Y_{22}}{4}\right)\right\}^{\frac{1}{2}} = \frac{1}{2l}\ln\left(\frac{\frac{1}{2}\sqrt{\left(Z_{11} - 2Z_{21} + Z_{22}\right)\left(Y_{11} - 2Y_{21} + Y_{22}\right)}+1}{\frac{1}{2}\sqrt{\left(Z_{11} - 2Z_{21} + Z_{22}\right)\left(Y_{11} - 2Y_{21} + Y_{22}\right)}-1}\right) \tag{116}$$

5.9.5 Expression of results

When the secondary transmission parameters deal with frequency domain data and show that the data varies versus frequency, the least squares function fit method is used to extract the secondary transmission parameters as theoretic ideal parameters of the transmission line.

6 Measurement of return loss and structural return loss

6.1 General

Return loss and *SRL* are both useful for quantifying the level (amount) of the reflected signal. Return loss combines the effects of reflections due to both the deviation from the nominal impedance (such as 100 Ω) and structural effects. It is specified when system performance is the primary interest.

While return loss characterises the performance of the channel or link, *SRL* is used to represent the structural effects of the cable medium itself relative to Z_C and is useful for cable evaluation.

6.2 Principle

The same measurement principles apply as in 5.2. Many network analysers yield return loss in a direct manner as a menu item. The circuit given in Figure 5 is suitable for the *RL* and *SRL* measurements. Where calibration of the network analyser and *S*-parameter unit is performed relative to the reference impedance, the return loss, *RL*, is given by Formula (117):

$$RL = -20\log\left|S_{11}\right| \tag{117}$$

Stated in terms of the impedances, the return loss, *RL*, is given by Formula (118):

$$RL = -20\log\left|\frac{Z_T - Z_R}{Z_T + Z_R}\right| \tag{118}$$

NOTE Open/short circuit data is not appropriate for return loss since both ends of the circuit must be terminated with the reference impedance. The difference between the Z_R used here and the Z_{CM} used for *SRL* is obviously small when roundtrip loss is large enough to render the distant-end reflection negligible.

The SRL is obtained by Formula (119), where Z_C is the fitted complex characteristic impedance being used as the reference value.

$$SRL = -20\log\left|\frac{Z_{CM} - Z_C}{Z_{CM} + Z_C}\right| \tag{119}$$

7 Propagation coefficient effects due to periodic structural variation related to the effects appearing in the structural return loss

7.1 General

The complex characteristic impedance Z_C of a cable is defined as the quotient of a voltage wave (U) and current wave (I) which are propagating in the same direction, forwards (f) or backwards (r). For homogeneous cables with no structural variations, the complex characteristic impedance can be measured directly as the quotient of voltage and current at the cable ends.

$$Z_C = \frac{U_f}{I_f} = \frac{U_r}{I_r} \tag{120}$$

The other characteristics which are important for a cabling system are the input and output impedances and the corresponding return losses and the structural return loss of the cable. These characteristics include structural variation in the cable. They are measured by the S_{11} and S_{22} parameters of the cable, as described in the following.

Important cable-related parameters, which for their part describe the quality of the cable as a transmission medium, are the complex characteristic impedance Z_C and the structural return loss SRL.

System-related parameters are the input impedance and the return loss at the input and output of the cable, which are related to the scattering parameters S_{11} and S_{22}. The insertion loss is also a system-related parameter which is denoted by S_{21}.

The transmission (propagation) coefficient:

$$\gamma = \alpha + j\beta \tag{121}$$

is only cable related. It has already been discussed in Clause 4.

7.2 Formula for the forward echoes caused by periodic structural inhomogeneities

The reflected signals down the line have normally little direct effect on the transmission but through double reflections they influence the forward transmission causing forward echoes at resonant spike frequencies.

With periodic inhomogeneities extending throughout the line, the forward echo coefficient q can be calculated from Formula (122) when the measured periodic structural return loss $PSRL$ coefficient is p at a resonant frequency.

$$|q|_{max} = K|p|^2_{max} \qquad (122)$$

where

$$K = \frac{2\alpha l - 1 + e^{-2\alpha l}}{\left(1 - e^{-2\alpha l}\right)^2} \qquad (123)$$

When 2 αl >> 1 (Np):

$$K \approx 2\alpha l - 1 \qquad (124)$$

The above is only related to the cable and the cable length.

Also, to be considered is the forward echo caused by the mismatch between the generator impedance Z_G, and the input impedance Z_{IN}, and between the load impedance Z_L and the output impedance Z_{OUT} of the cable.

Return losses RL are defined by Formulae (125) and (126):

$$RL_{IN} = -20\log\left|\frac{Z_{IN} - Z_G}{Z_{IN} + Z_G}\right| \qquad (125)$$

$$RL_{OUT} = -20\log\left|\frac{Z_{OUT} - Z_L}{Z_{OUT} + Z_L}\right| \qquad (126)$$

The echo attenuation A_E from these two reflections is:

$$A_E = 2\alpha l + RL_{IN} + RL_{OUT} \qquad (127)$$

The total echo attenuation A_{TOT} of the repeater or regenerator section is:

$$A_{TOT} = -10\log\left(10^{\frac{-A_Q}{10}} + 10^{\frac{-A_E}{10}}\right) \qquad (128)$$

If Z_G and Z_L are taken as reference impedances in the scattering parameters measurement, then:

$$S_{11} = \frac{(Z_{IN} - Z_G)}{(Z_{IN} + Z_G)} \quad (129)$$

$$S_{22} = \frac{(Z_{OUT} - Z_L)}{(Z_{OUT} + Z_L)} \quad (130)$$

The composite loss (same as insertion loss A_I if Z_G = Z_L) is:

$$A_C = -20\log|S_{21}| \quad (131)$$

Observe that the cable attenuation:

$$\alpha l \neq A_C \text{ or } A_I \quad (132)$$

For a homogenous cable, the composite loss (attenuation) is:

$$A_C = \alpha l + 20\log\left|\frac{Z_G + Z_C}{2\sqrt{Z_G Z_C}}\right| + 20\log\left|\frac{Z_L + Z_C}{2\sqrt{Z_L Z_C}}\right| + 20\log\left|l - r_1 r_2 e^{-2(\alpha + j\beta)l}\right| \quad (133)$$

where

$$r_1 = \frac{(Z_G - Z_C)}{(Z_G + Z_C)} \quad (134)$$

$$r_2 = \frac{(Z_L - Z_C)}{(Z_L + Z_C)} \quad (135)$$

8 Unbalance attenuation

8.1 General

Symmetric pairs can be operated in the differential mode (balanced) (see Figure 9) or the common mode (unbalanced) (see Figure 10). In the differential mode, one conductor carries the current and the other conductor carries the return current. The return path (common mode) should be free of any current.

In the common mode, each conductor of the pair carries half of the current and the return path carries the sum of both these currents. All pairs not under test and any screens, if present, represent the return path for the common-mode voltage.

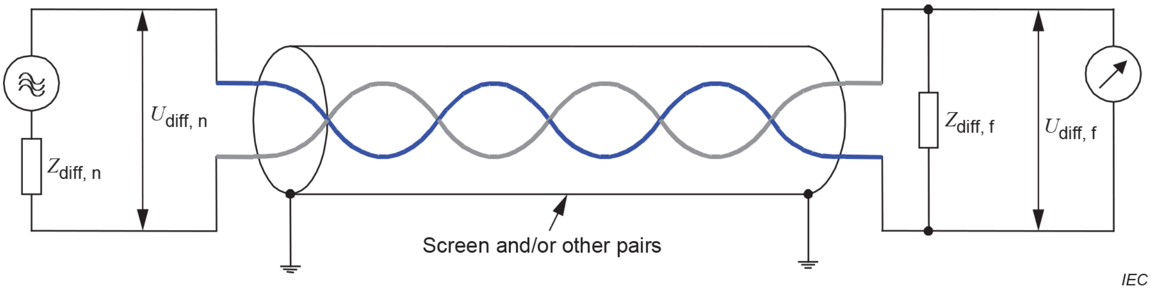


Figure 9 – Differential-mode transmission in a symmetric pair

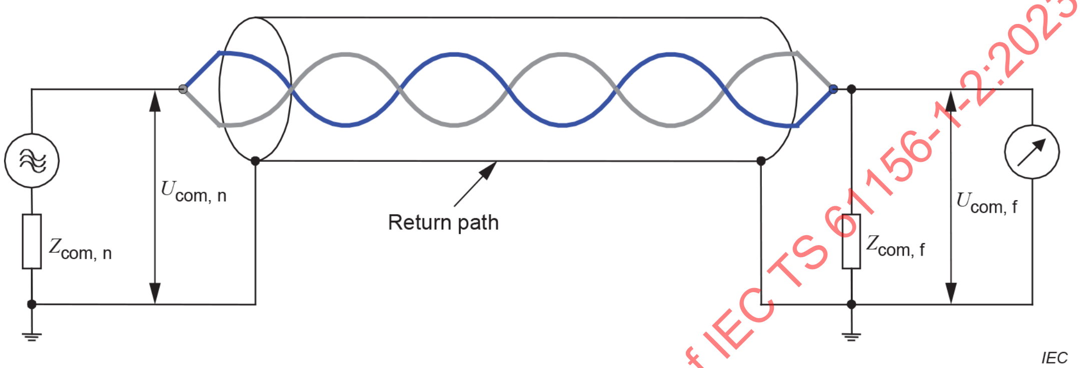


Figure 10 – Common-mode transmission in a symmetric pair

Under ideal conditions, both modes are independent of one another. In reality, both modes influence each other. Differences in the diameter of the insulation, unequal twisting and different distances of the conductors to the screen are some reasons for the unbalance of a pair. The asymmetry is caused by the transverse-asymmetry and by the longitudinal asymmetry. The transverse asymmetry, TA, is caused by longitudinally distributed unbalances to earth of the capacitance and conductance. The longitudinal asymmetry, LA, is caused by the inductance and resistance unbalances between the two conductors of the pair.

8.2 Unbalance attenuation near end and far end

Unbalance attenuation is measured as the logarithmic ratio of the common-mode power to the differential-mode power at the near end and at the far end of the cable. The unbalance attenuation is also often referred to as conversion loss:

- LCL longitudinal conversion loss
- $LCTL$ longitudinal conversion transfer loss
- TCL transverse conversion loss
- $TCTL$ transverse conversion transfer loss

Additionally, the equal level unbalance attenuation far end is defined as follows:

- $EL\ LCTL$ equal level longitudinal conversion transfer loss
- $EL\ TCTL$ equal level transverse conversion transfer loss

The equal level unbalance attenuation is defined as an output-to-output measurement of the logarithmic ratio of the common-mode power to the differential-mode power or vice versa. The output-to-output measurements correspond to the difference of the input-to-output measurement and the respective attenuation:

$$EL\,LCTL = LCTL - \alpha_{\text{com}} \qquad (136)$$

$$ELTCTL = TCTL - \alpha_{\text{diff}}$$

As it is not a common practice to measure the output-to-output ratios directly, the above differences are used to determine the equal level unbalance attenuation. The measurement of the common-mode attenuation of balanced cables is prone to error, and the differential attenuation of the cables needs to be measured anyway. Therefore, the measurement of the equal level unbalance attenuation far end is limited here to the equal level transverse conversion transfer loss.

The unbalance attenuation near end or far end is related to the conversion losses as indicated in Table 1 and Table 2, respectively.

Table 1 – Unbalance attenuation at near end

Power fed at the near end into the differential-mode and coupled power measured at the near end in the common mode	*TCL*
Power fed at the near end into the common-mode and coupled power measured at the near end in the differential mode	*LCL*

Table 2 – Unbalance attenuation at far end

Power fed at the near end into the differential-mode and coupled power measured at the far end in the common mode	*TCTL*
Power fed at the near end into the common-mode and coupled power measured at the far end in the differential mode	*LCTL*
Same as *TCTL* but the measured common-mode power is related to the differential-mode power at the far end (equal level)	*EL TCTL*

Table 3 indicates the common- and differential-mode circuit of the input, and the receive signal for the different types of unbalance attenuation.

Table 3 – Measurement set-up

Unbalance attenuation		**Set-up**			
		Near end		**Far end**	
		Common-mode circuit	**Differential-mode circuit**	**Common-mode circuit**	**Differential-mode circuit**
Near end	*TCL*	Receiver	Generator	–	–
	LCL	Generator	Receiver	-	–
Far end	*TCTL*	–	Generator	Receiver	–
	LCTL	Generator	–	–	Receiver

Using the concept of operational attenuation, the generator and receiver on one port of the network are interchangeable without any change in the results. Therefore, the measurements of *TCL* are identical to those of *LCL*.

However, the measurement of *LCTL* or *TCTL* is inherently a two-port measurement. Therefore, the measurements of *LCTL* are only identical to those of *TCTL* if the longitudinal distribution of the unbalances is homogeneous, and if the velocity of propagation of differential- and common-mode signals is identical. In this case, the twisted pair corresponds to a reciprocal and impedance symmetrical two-port network.

If differential-mode transmission is considered, then the loss due to conversion of the differential-mode signal into common-mode signal only is of interest. This yields an additional advantage. Feeding the power into the differential-mode ports of the balun yields the benefit that the balun then represents a matched generator, which avoids the need of any additional matching pads.

The differential-mode impedance of multiple pair cables is a well-known design parameter. However, the common-mode impedance depends upon the design of the cable and is influenced primarily by the insulation thickness, the dielectric constant of the insulation, the proximity and number of neighbouring pairs and finally by the presence of shields. Thus, the common-mode impedance of nominally 100 Ω cables can vary within the range of 25 Ω to 75 Ω depending on cable construction. For STP (individually screened twisted pair) cables, it is approximately 25 Ω. For FTP (common screened twisted pair) cables, it is approximately 50 Ω. For UTP (unscreened twisted pair) cables, it is approximately 75 Ω.

The baluns used for measuring match the input impedance of the S-parameter test set to the differential-mode impedance of the cable under test (CUT). It is, however, impractical to measure first for each cable the common-mode impedance to match it then to the corresponding common-mode impedance terminations used on the balun. But in practice, as the vector network analyser (or equivalent instrument) used in the measurement has 50 Ω ports in general and the common mode impedance of the balun is also 50 Ω, the common mode impedance of the termination shall be 50 Ω unless otherwise specified. This is to simplify measurements when the common mode impedance of the test specimen varies from 25 Ω to 75 Ω. For cables with a nominal impedance of 100 Ω, the 50 Ω termination is presented by the input impedance of the network analyser. This proceeding entails due to eventual impedance mismatches a variation of the unbalance attenuation due to the reflected signal. Thus, a return loss of 10 dB yields an uncertainty of about ±1 dB.

8.3 Theoretical background

The transverse asymmetry, TA, is caused by longitudinally distributed unbalances to earth of the capacitance and conductance. The longitudinal asymmetry, LA, is caused by the inductance and resistance unbalances between the two conductors of the pair.

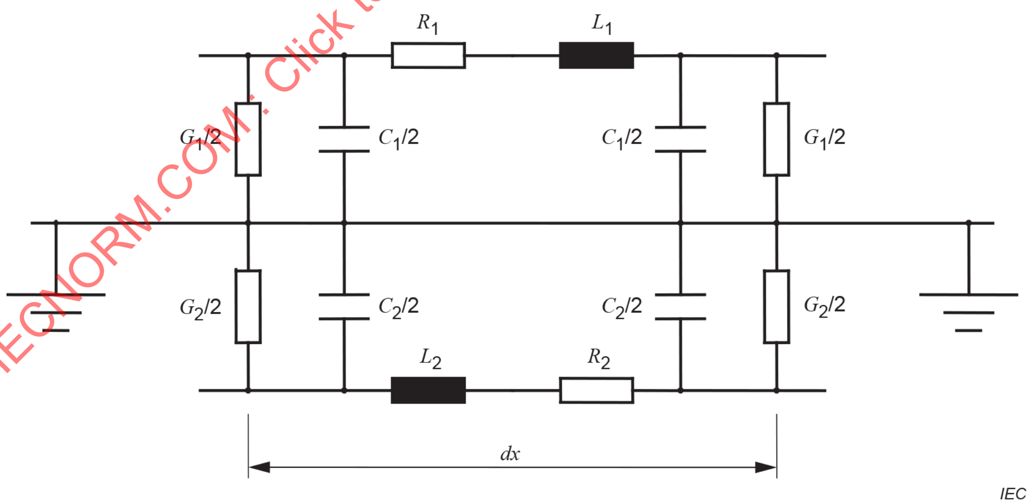


Figure 11 – Circuit of an infinitesimal element of a symmetric pair

With the Formula (137) below, the unbalance of a symmetric pair can be expresses (see Figure 11):

$$TA = (G_2 + j\omega C_2) - (G_1 + j\omega C_1) \tag{137}$$

$$LA = (R_2 + j\omega L_2) - (R_1 + j\omega L_1)$$

The coupling between the differential- and common-mode circuit is expressed by:

$$\alpha_{\substack{\text{u, n}\\ \text{u, f}}} = 20\log\left|T_{\substack{\text{u, n}\\ \text{u, f}}}\right| \quad (138)$$

where

$$T_{\substack{\text{u, n}\\ \text{u, f}}} = \sqrt{\frac{U_{\text{com}}}{U_{\text{diff}}}} \quad (139)$$

With the definition of an unbalance impedance:

$$Z_{\text{unbal}} = \sqrt{Z_{\text{diff}} \times Z_{\text{com}}} \quad (140)$$

The terms for the unbalance coupling functions represent, in principle, the same coupling transfer functions as for the coupling through screens or the coupling between lines (crosstalk). Hence, they can be formally written down as:

$$T_{\text{u,n}} = \frac{1}{4} \times \frac{1}{Z_{\text{unbal}}} \times \int_{x=0}^{x=l} \left[TA(x) \times Z_{\text{unbal}}^2 + LA(x)\right] \times e^{-(\gamma_{\text{diff}} + \gamma_{\text{com}}) \times x} dx \quad (141)$$

$$T_{\text{u,f}} = \frac{1}{4} \times \frac{1}{Z_{\text{unbal}}} \times e^{-\gamma_{\text{com}} \times l} \int_{x=0}^{x=l} \left[TA(x) \times Z_{\text{unbal}}^2 - LA(x)\right] \times e^{-(\gamma_{\text{diff}} - \gamma_{\text{com}}) \times x} dx \quad (142)$$

When $\gamma \times \ell \approx 0$, the unbalance coupling functions can be separated into Formula (143) for the unbalances of the primary parameters:

$$\left|T_{\text{Conductance}}\right| = \frac{Z_{\text{unbal}}}{4} \times \Delta G \qquad \left|T_{\text{Capacitance}}\right| = \frac{Z_{\text{unbal}}}{4} \times \omega\Delta$$

$$\left|T_{\text{Resistance}}\right| = \frac{Z_{\text{unbal}}}{4} \times \Delta R \qquad \left|T_{\text{Inductance}}\right| = \frac{Z_{\text{unbal}}}{4} \times \omega\Delta L \quad (143)$$

Formulae (141) and (142) represent, in principle, the same coupling transfer functions compared to the coupling through the screen or the crosstalk between lines. The integral can only be solved if the distribution of the capacitance, resistance and inductance unbalances along the cable length are known. For longitudinally constant unbalances, the transfer function gives comparable results as for the coupling through cable screens, or the crosstalk between lines.

$$T_{\substack{\mathrm{u,\,n}\\ \mathrm{u,\,f}}} = \left(TA\,Z_{\mathrm{unbal}}^{2} \pm LA\right)\frac{1}{Z_{\mathrm{unbal}}}\frac{l}{4}S_{\substack{\mathrm{n}\\ \mathrm{f}}} \,. \qquad (144)$$

The phase effect, when summing up the infinitesimal couplings along the line is expressed by the summing function S. When the cable attenuation is neglected, S can be expressed by the following formula.

$$S_{\substack{\mathrm{n}\\ \mathrm{f}}} = \frac{\sin\left(\beta_{\mathrm{diff}} + \beta_{\mathrm{com}}\right)\times\frac{l}{2}}{\left(\beta_{\mathrm{diff}} + \beta_{\mathrm{com}}\right)\times\frac{l}{2}} \times e^{-\left(j\left(\beta_{\mathrm{diff}}+\beta_{\mathrm{com}}\right)\times\frac{l}{2}\right)} \qquad (145)$$

For high frequencies, the asymptotic value becomes

$$S_{\substack{\mathrm{n}\\ \mathrm{f}}} = \frac{2}{\left(\beta_{\mathrm{diff}} \pm \beta_{\mathrm{com}}\right)\times l} \qquad (146)$$

and for low frequencies, the summing function becomes

$$\left|S_{\substack{\mathrm{n}\\ \mathrm{f}}}\right| \rightarrow 1 \qquad (147)$$

In practice, we have small systematic couplings together with statistical couplings. Thus, $T_{\mathrm{u,n}}$ increases by approximately 15 dB per decade. Figure 12 shows the calculated coupling transfer function for a cable length of 100 m and a capacitance unbalance to earth, which consists of a constant part of 0,4 pF/m and a random ±0,4 pF/m longitudinal variation.

The relative dielectric permittivity of the differential- and common-mode circuit is here assumed to be 2,3. The magnetic coupling and the cable attenuation have been neglected. Figure 13 shows the measured coupling transfer function for a length of 100 m of a Twinax[1] cable with 105 Ω, and with a braided screen. The conductors are PE insulated and have an inner PE sheath. The resultant velocity difference is, therefore, zero.

1 Twinax is an example of a suitable product available commercially. This information is given for the convenience of users of this document and does not constitute an endorsement by IEC of this product.

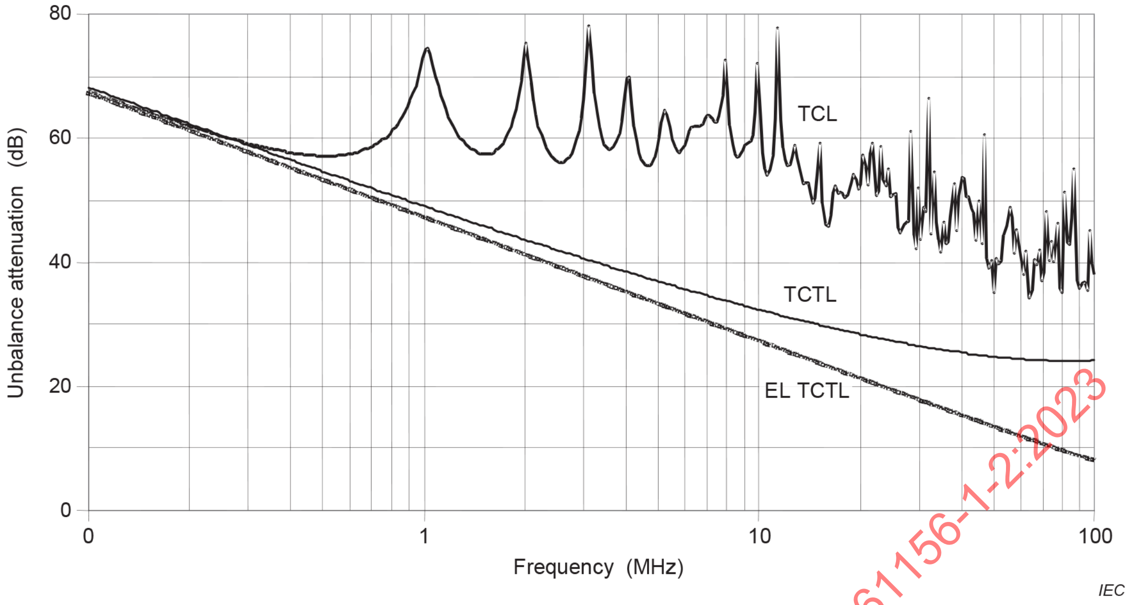


Figure 12 – Calculated coupling transfer function for a capacitive coupling of 0,4 pF/m and random ±0,4 pF/m (ℓ =100 m; εr1 = εr2 = 2,3)

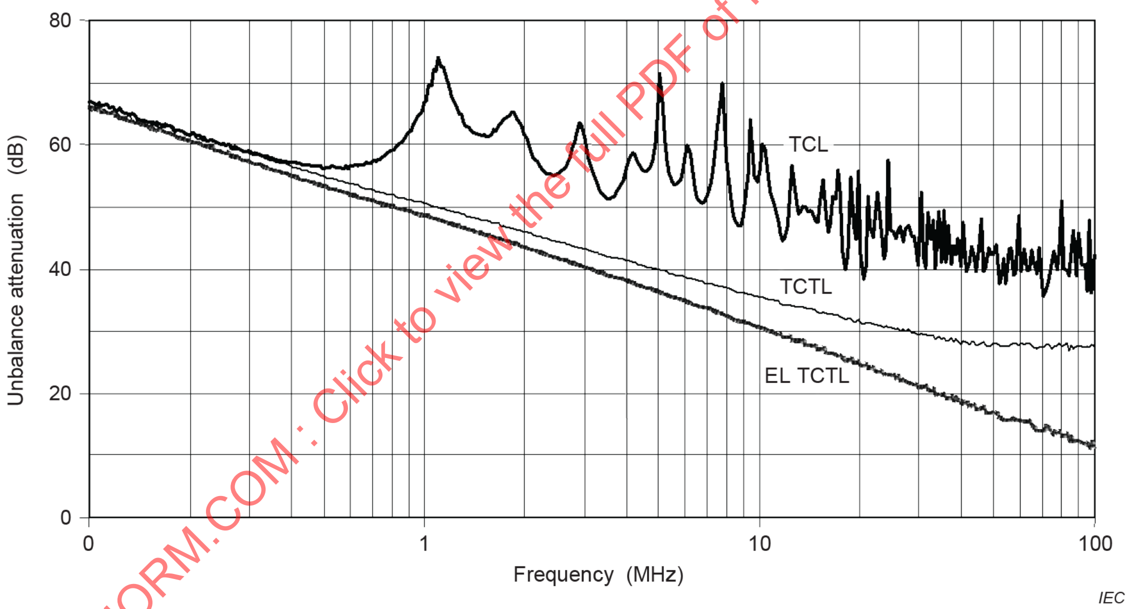


Figure 13 – Measured coupling transfer function of 100 m Twinax 105 Ω

9 Balunless test method

9.1 Overall test arrangement

9.1.1 Test instrumentation

The test procedures hereby described require the use of a vector network analyser or similar test equipment. The analyser shall have the capability of full 4-port calibration and should include isolation calibrations. The analyser shall cover at least the full frequency range of the cable or cabling under test (CUT). Not all given parameters below are applicable to all cables or cabling which could be evaluated according to this document.

Measurements shall be taken using a mixed mode test set-up, which is often referred to as an unbalanced, modal decomposition or balun-less setup. This allows measurements of balanced devices without use of an RF balun in the signal path. With such a test set-up, all balanced and unbalanced parameters can be measured over the full frequency range.

Such a configuration allows testing with both a common or differential mode stimulus and responses, ensuring that intermodal parameters can be measured without reconnection.

A 16-port network analyser is required to measure all combinations of a 4 pair device without external switching; however, the network analyser shall have a minimum of 2 ports to enable the data to be collected and calculated.

It should be noted that the use of a 4-port analyser will involve successive repositioning of the measurement ports in order to measure any given parameter.

A 4-port network analyser is recommended as a minimum number of ports, as this will allow the measurement of the full 16 term mixed mode S-parameter matrix on a given pair combination without switching or reconnection in one direction.

In order to minimise the reconnection of the CUT for each pair combination, the use of an RF switching unit is also recommended.

Each conductor of the pair or pair combination under test should be connected to a separate port of the network analyser, and results are processed either by internal analysis within the network analyser or by an external application.

Reference loads and through connections are needed for the calibration of the set-up. Requirements for the reference loads are given in 9.1.5. Termination loads are needed for termination of pairs, used and unused, which are not terminated by the network analyser. Requirements for the termination loads are given in 9.1.7.

9.1.2 Measurement precautions

To assure a high degree of reliability for transmission measurements, the following precautions are required:

a) Consistent and stable resistor loads shall be used throughout the test sequence.
b) Cable and adapter discontinuities, as introduced by physical flexing, sharp bends and restraints shall be avoided before, during and after the tests.
c) Consistent test methodology and termination resistors shall be used at all stages of transmission performance qualifications.
d) The relative spacing of conductors in the pairs shall be preserved throughout the tests to the greatest extent possible.
e) The balance of the cables shall be maintained to the greatest extent possible by consistent cable lengths, pair twisting and lay up of the screen to the point of load.
f) The sensitivity to setup variations for these measurements at high frequencies demands attention to details for both the measurement equipment and the procedures.

9.1.3 Mixed mode S-parameter nomenclature

The test methods specified in this document are based on a balun-less test set-up in which all terminals of a device under test are measured and characterised as single-ended (SE) ports, i.e., signals (RF voltages and currents) are defined relative to a common ground. For a device with 4 terminals, a diagram is given in Figure 14.

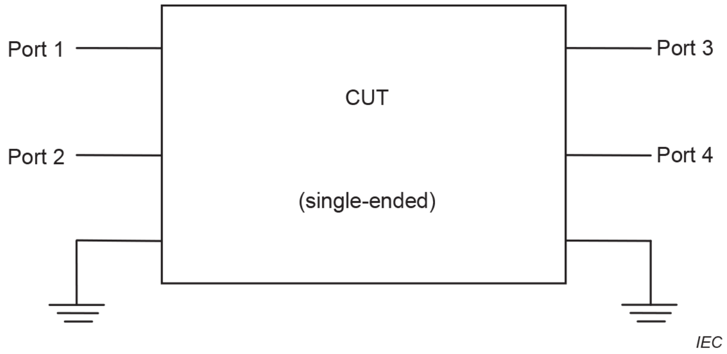


Figure 14 – Diagram of a single-ended 4-port device

The 4-port device in Figure 14 is characterised by the 16 term SE S-matrix given in Formula (148), in which the S-parameter S_{ba} expresses the relation between a single-ended response on port "b" resulting from a single ended stimulus on port "a".

$$S = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{34} & S_{44} \end{bmatrix} \quad (148)$$

For a balanced device, each port is considered to consist of a pair of terminals (= a balanced port) as opposed to the SE ports defined above, see Figure 15.

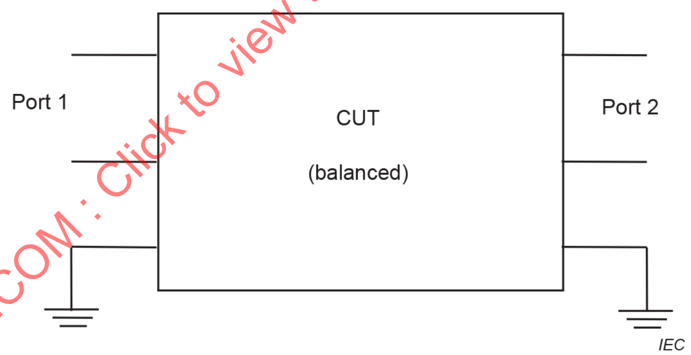


Figure 15 – Diagram of a balanced 2-port device

In order to characterise the balanced device, both the differential mode and the common mode signals on each balanced port shall be considered. The device can be characterised by a mixed mode S-matrix that includes all combinations of modes and ports, e.g., the mixed mode S-parameter S_{DC21} that expresses the relation between a differential mode response on port 2 resulting from a common mode stimulus on port 1. Using this nomenclature, the full set of mixed mode S -parameters for a balanced 2-port device can be presented as in Table 4.

Table 4 – Mixed mode S-parameter nomenclature

		Differential mode stimulus		Common mode stimulus	
		Port 1	Port 2	Port 1	Port 2
Differential mode response	Port 1	S_{DD11}	S_{DD12}	S_{DC11}	S_{DC12}
	Port 2	S_{DD21}	S_{DD22}	S_{DC21}	S_{DC22}
Common mode response	Port 1	S_{CD11}	S_{CD12}	S_{CC11}	S_{CC12}
	Port 2	S_{CD21}	S_{CD22}	S_{CC21}	S_{CC22}

A 4-terminal device can be represented both as a 4-port SE device as in Figure 14 characterised by a single ended S-matrix (Formula (148)) and as a 2-port balanced device as in Figure 15 characterised by a mixed mode S-matrix (see Table 4). As applying a SE signal to a port is mathematically equivalent to applying superposed differential and common mode signals, the SE and the mixed mode characterisations of the device are interrelated. The conversion from SE to mixed mode S-parameters is given in Annex A. Making use of this conversion, the mixed mode S-parameters can be derived from the measured SE S-matrix.

9.1.4 Coaxial cables and interconnect for network analysers

Assuming that the characteristic impedance of the network analyser is 50 Ω, coaxial cables used to interconnect the network analyser, switching matrix (if present) and the test fixtures shall be of 50 Ω characteristic impedance and of low transfer impedance (double screen or more).

These coaxial cables shall be as short as possible. (It is recommended that they do not exceed 1 000 mm each.)

The screens of each cable shall be electrically bonded to a common ground plane, with the screens of the cable bonded to each other at multiple points along their length.

To optimise dynamic range, the total interconnecting cable insertion loss shall be minimised. (It is recommended that the interconnecting cable loss does not exceed 3 dB at 1 000 MHz.)

9.1.5 Reference loads for calibration

The N-connector should be seen as an example. Other connectors can be used for similar purposes such as e.g., SMA-connectors. Some test equipment even uses no standardised fixtures.

To perform a one port calibration of the test equipment, a short circuit, an open circuit and a reference load are required. For a 2-port calibration, additionally to the above, a thru artefact is required. These devices shall be used to obtain a calibration.

The reference load shall be calibrated against a calibration reference, which shall be a 50 Ω load, traceable to an international reference standard. One 50 Ω reference load shall be calibrated against the calibration reference. The reference load for calibration should be placed in an N-type connector according to IEC 61169-16 or an SMA-connector according to IEC 61169-15, meant for panel mounting, which is machined-flat on the back side (see Figure 16). For frequencies higher than 1 GHz, a SMA-connector should be used.

The load shall be fixed to the flat side of the connector. A network analyser shall be calibrated, 1-port full calibration, with the calibration reference. Thereafter, the return loss of the reference load for calibration shall be measured. The verified return loss shall be ≥46 dB at frequencies up to 100 MHz and ≥40 dB at frequencies above 100 MHz and up to the limit for which the measurements are to be conducted.

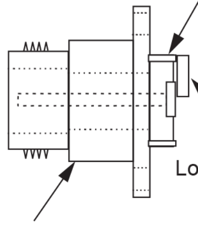


Figure 16 – Solution for calibration of reference loads

For short and open, the inductance and capacitance shall be minimised.

9.1.6 Calibration

Isolation measurements should be used as part of the calibration.

The calibration shall be equivalent to a minimum of a full 4-port SE calibration for measurements where the response and stimulus ports are the same (S_{xx11} and S_{xx22}) or different (S_{xx12} and S_{xx21}).

An individual calibration shall be performed for each signal path used for the measurements. If a complete switching matrix and a 4-port network analyser test set-up is used, a full set of measurements for a 4-pair device (i.e., 16 single-ended ports) will require 28 separate 4-port calibrations, although many of the measurements within each calibration are in common with other calibrations. A software or hardware package can be used to minimise the number of calibration measurements required.

The calibration shall be applied in such a way that the calibration plane shall be at the ends of the fixed connectors of the test fixture.

The calibration can be performed at the test interface using appropriate calibration artefacts, or at the ends of the coaxial test cable using coaxial terminations.

Where calibration is performed at the test interface, open, short and load measurements shall be taken on each SE port concerned, and through and isolation measurements should be taken on every pair combination of those ports.

Where calibration is performed at the end of the coaxial test cables, open, short and load measurements shall be taken on each port concerned, and through and isolation measurements should be taken on every pair combination of those ports. In addition, the test fixture shall then be de-embedded from the measurements. The de-embedding techniques shall incorporate a fully populated 16 port S-matrix. It is not acceptable to perform a de-embedded calibration using only reflection terms (S_{11}, S_{22}, S_{33}, S_{44}) or only near-end terms (S_{11}, S_{21}, S_{12}, S_{22}).

De-embedding using reduced term S-matrices can be used for post processing of results.

9.1.7 Termination loads for termination of conductor pairs

9.1.7.1 General

When this document is used for the measurement of performance against standards, the differential mode terminations applied to the device under test (DUT) shall provide the differential mode and common mode reference termination impedances specified in standards for the cabling system where the DUT is used.

50 Ω wires to ground terminations shall be used on all active pairs under test. 50 Ω differential mode to ground terminations shall be used on all inactive pairs and on the opposite ends of active pairs for near-end crosstalk (*NEXT*) and far-end crosstalk (*FEXT*) testing. Inactive pairs for return loss testing shall be terminated with 50 Ω differential mode to ground terminations. See Figure 17.

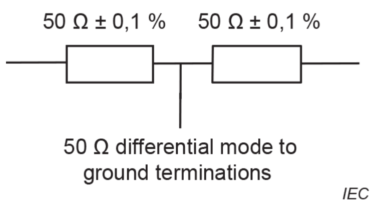


Figure 17 – Resistor termination networks

Small geometry chip resistors should be used for the construction of resistor terminations. The two 50 Ω DM terminating resistors shall be matched to within 0,1 % at DC, and 2 % at 1 000 MHz (corresponding to a 40 dB return loss requirement at 1 000 MHz). The length of connections to impedance terminating resistors shall be minimised. Use of soldered connections without leads is recommended.

9.1.7.2 Verification of termination loads

The performance of impedance matching resistor termination networks shall be verified by measuring the return loss of the termination and the residual *NEXT* between any two resistor termination networks at the calibration plane.

For the return loss measurement, a 2-port SE calibration is required using a reference load verified according to 9.1.5.

After calibration, connect the resistor termination network and perform a full 2-port SE *S*-matrix measurement. The measured SE *S*-matrix shall be transformed into the associated mixed mode *S*-matrix to obtain the *S*-parameters S_{DD11} and S_{CC11} from which the differential mode return loss RL_{DM} and the common mode return loss RL_{CM} are determined. The return loss of the resistor termination network shall meet the requirements of Table 5.

For the residual *NEXT* measurement, a 4-port SE calibration is required. After calibration, connect the resistor termination networks and perform a full 4-port SE *S*-matrix measurement. The measured *S*-matrix shall be transformed into the associated mixed mode *S*-matrix to obtain the *S*-parameter S_{DD21} from which the residual *NEXT* of the terminations, $NEXT_{residual\_term}$, is determined. The residual *NEXT* shall meet the requirements of Table 5.

For the *TCL* measurement, a 2-port SE calibration is required using a reference load verified according to 9.1.5.

After calibration, connect the resistor termination network and perform a full 2-port SE *S*-matrix measurement. The measured SE *S*-matrix shall be transformed into the associated mixed mode *S*-matrix to obtain the *S*-parameter S_{CD11} from which the differential mode *TCL* is determined. The *TCL* of the resistor termination network shall meet the requirements of Table 5.

Table 5 – Requirements for terminations at calibration plane

Parameter	Frequency kHz	Requirement up to maximum frequency (f [MHz])
SE port (50 Ω) return loss (dB)	$1 \leq f \leq f_{max}$	≥74-20 log(f) dB 40 dB max 20 dB min
DM port (100 Ω) return loss (dB)		≥74-20 log(f) dB 40 dB max 20 dB min
DM port to port residual *NEXT* (dB)		≥140-20 log(f) dB 104 dB max 80 dB min
DM port *TCL* of loads (dB)		≥ 60-10 log(f) dB 50 dB max 20 dB min

9.1.8 Termination of screens

The screen or screens of the CUT and the measurement cables shall be fixed to the ground plane as close as possible to the calibration plane.

9.1.9 Calibration

A full 4-port SE calibration shall be performed at the calibration planes in accordance with 9.1.6. Reference loads used for calibration shall be in accordance with 9.1.5.

9.1.10 Establishment of noise floor

The noise floor of the set-up shall be measured. The level of the noise floor is determined by white noise, which can be reduced by increasing the test power and by reducing the bandwidth of the network analyser, and by residual crosstalk within the test fixture.

The noise floor shall be measured by terminating the test ports of the test fixture with resistor termination networks and performing a full SE S-matrix measurement. The measured SE S-matrix is transformed into the associated mixed mode S-matrix to obtain the S-parameter S_{DD21} from which the noise floor is established. The noise floor shall be established for all conductor pair combinations.

The noise floor shall be at least 10 dB better than any specified limit for the crosstalk. If the measured value is closer to the noise floor than 10 dB, this shall be reported.

For high crosstalk values, it can be necessary to screen the terminating resistors.

9.2 Cabling and cable measurements

9.2.1 Insertion loss and *EL TCTL*

9.2.1.1 Object

When this document is used for the measurement of performance against standards, the differential mode terminations applied to the DUT shall provide the differential mode and common mode reference termination impedances specified in standards for the cabling system where the DUT is used.

The object of this test is to measure the insertion loss (*IL*) and equal-level transverse conversion transfer loss (*EL TCTL*) of a cable or cabling pair. Insertion loss is defined as the attenuation that is caused by the cable or cabling pair. *EL TCTL* is defined as the unbalance attenuation at far end.

9.2.1.2 Cable and cabling insertion loss and *EL TCTL*

Cable or cabling shall be evaluated for insertion loss at least in one direction and *EL TCTL* in both directions. In case one of the ends does not pass the applicable *RL* requirements, insertion loss shall be evaluated in both directions too.

9.2.1.3 Test method

Insertion loss is evaluated from the mixed mode parameter S_{DD21} and *EL TCTL* is evaluated from the mixed mode parameter S_{CD21} for each conductor pair. The mixed mode *S*-parameters are derived by transformation of the SE *S*-matrix.

9.2.1.4 Test set-up

The test set-up consists of a 4-port network analyser and two test fixtures. An illustration of the test set-up, which also shows the termination principles, is shown in Figure 18. Resistor termination networks in accordance with 9.1.7 shall be applied for all inactive pairs.

NOTE As a minimum, a 2-port network analyser is sufficient, but this requires many reconnections which reduces the measurement accuracy.

9.2.1.5 Measurement

The CUT shall be arranged in an appropriate test set-up according to Figure 18, including proper termination of the active, inactive pairs and screen. A full SE *S*-matrix measurement shall be performed. The measured SE *S*-matrix shall be transformed into the associated mixed mode *S*-matrix to obtain the *S*-parameter S_{DD21} from which insertion loss is determined and S_{CD21} from which *TCTL* is determined.

$$IL(f) = -20 \times \log_{10}\left(\left|S_{\text{DD21}}\right|\right) = -20 \times \log_{10}\left(\left|\frac{1}{2}\left(S_{31} - S_{41} - S_{32} + S_{42}\right)\right|\right) \tag{149}$$

$$TCTL(f) = -20 \times \log_{10}\left(\left|S_{\text{CD21}}\right|\right) = -20 \times \log_{10}\left(\frac{1}{2}\left(S_{31} + S_{41} - S_{32} - S_{42}\right)\right) \tag{150}$$

$$ELTCTL(f) = TCTL(f) - IL(f) = -20 \times \log_{10}\left(\left|\frac{S_{31} + S_{41} - S_{32} - S_{42}}{S_{31} - S_{41} - S_{32} + S_{42}}\right|\right) \tag{151}$$

Evaluate all conductor pairs and record the results.

The calibration and reference planes could be both at the interface between the cable and the fixture.

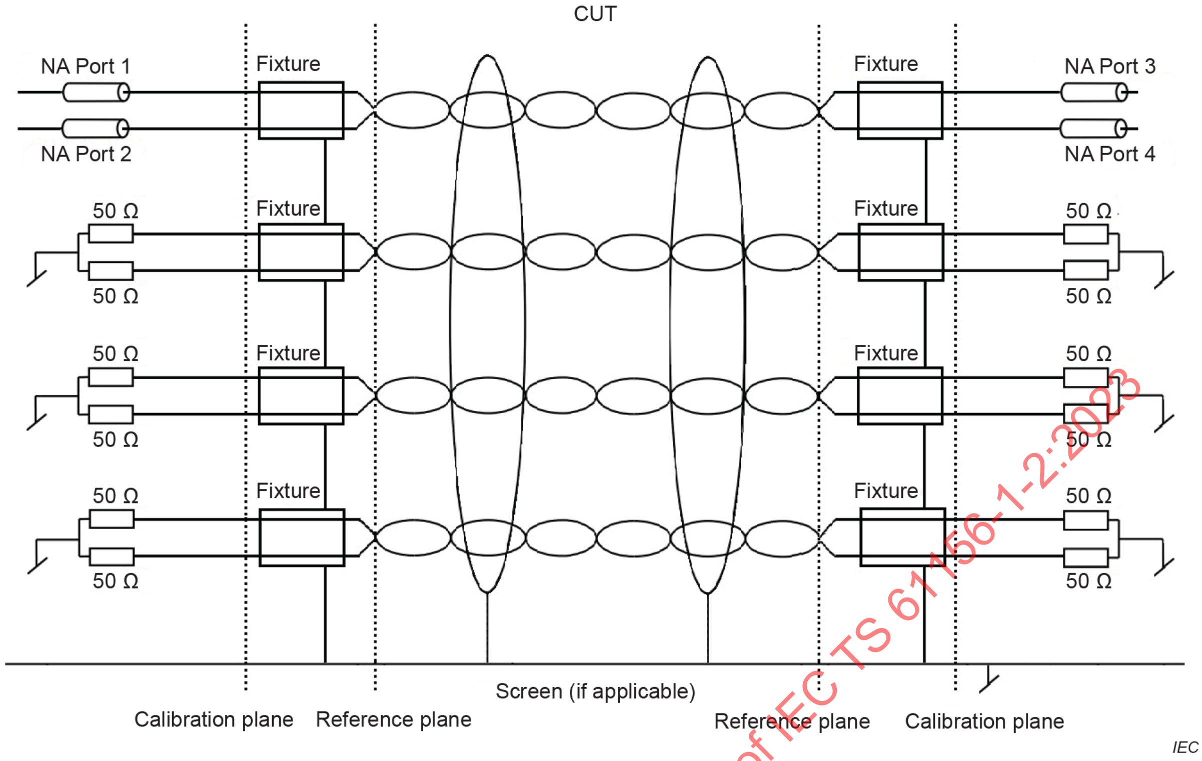


Figure 18 – Insertion loss and *EL TCTL*

9.2.1.6 Test report

The test results shall be reported in graphical or table format with the specification limits shown on the graphs or in the table at the same frequencies as specified in the relevant detail specification. Results for all pairs shall be reported. It shall be explicitly noted if the test results exceed the test limits.

9.2.1.7 Accuracy

As there is no definition of accuracy in this document and there is no procedure defined to determine the accuracy, the accuracy requirement is for further studies.

9.2.2 *NEXT*

9.2.2.1 Object

When this document is used for the measurement of performance against standards, the differential mode terminations applied to the DUT shall provide the differential mode and common mode reference termination impedances specified in standards for the cabling system where the DUT is used.

The objective of this test procedure is to measure the magnitude of the electric and magnetic coupling between the near ends of a disturbing and disturbed pair of a cable or cabling pair combination.

9.2.2.2 Cable or cabling *NEXT*

Cable or cabling shall be evaluated for *NEXT* in both directions.

9.2.2.3 Test method

NEXT is evaluated from the mixed mode parameter S_{DD21} for all conductor pair combinations. The mixed mode *S*-parameters are derived by transformation of the measured SE *S*-matrix.

9.2.2.4 Test set-up

The test set-up consists of a 4-port network analyser and two test fixtures. An illustration of the test set-up, which also shows the termination principles, is shown in Figure 19. Resistor termination networks in accordance with 9.1.7 shall be applied for all inactive pairs.

NOTE As a minimum, a 2-port network analyser is sufficient, but this requires many reconnections which reduces the measurement accuracy.

9.2.2.5 Measurement

The CUT shall be arranged in an appropriate test set-up according to Figure 18, including proper termination of the active, inactive pairs and screen. A full SE *S*-matrix measurement shall be performed. The measured SE *S*-matrix shall be transformed into the associated mixed mode *S*-matrix to obtain the *S*-parameter S_{DD21} from which *NEXT* is determined.

$$NEXT = -20 \times \log_{10}\left(\left|S_{DD21}\right|\right) = -20 \times \log_{10}\left(\left|\frac{1}{2}\left(S_{31} - S_{41} - S_{32} + S_{42}\right)\right|\right) \qquad (152)$$

The test shall be performed from both ends of the cable or cabling. Evaluate all conductor pair combinations and record the results.

The calibration and reference planes could be both at the interface between the cable and the fixture.

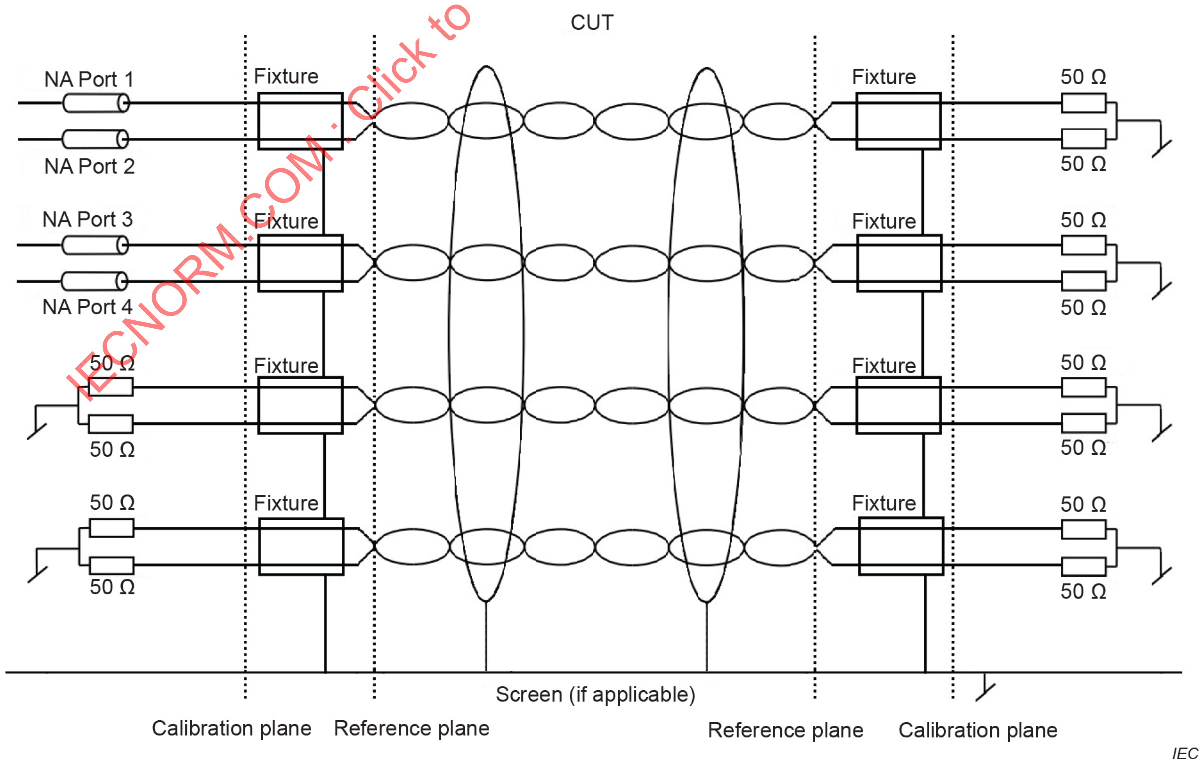


Figure 19 – *NEXT*