

TECHNICAL SPECIFICATION



CONSOLIDATED VERSION

**Metallic communication cable test methods –
Part 4-1: Electromagnetic compatibility (EMC) – Introduction to electromagnetic
screening measurements**



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screening measurements**

INTERNATIONAL
ELECTROTECHNICAL
COMMISSION

ICS 33.100.10

ISBN 978-2-8322-8341-7

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Metallic communication cable test methods – Part 4-1: Electromagnetic compatibility (EMC) – Introduction to electromagnetic screening measurements

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INTERNATIONAL ELECTROTECHNICAL COMMISSION

METALLIC COMMUNICATION CABLE TEST METHODS –

Part 4-1: Electromagnetic compatibility (EMC) – Introduction to electromagnetic screening measurements

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IEC TS 62153-4-1 edition 1.1 contains the first edition (2014-01) [documents 46/465/DTS and 46/492/RVC] and its amendment 1 (2020-05) [documents 46/726/DTS and 46/748/RVDTS].

In this Redline version, a vertical line in the margin shows where the technical content is modified by amendment 1. Additions are in green text, deletions are in strikethrough red text. A separate Final version with all changes accepted is available in this publication.

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Technical specifications are subject to review within three years of publication to decide whether they can be transformed into International Standards.

IEC TS 62153-4-1, which is a technical specification, has been prepared by IEC technical committee 46: Cables, wires, waveguides, R.F. connectors, R.F. and microwave passive components and accessories.

This first edition of technical specification IEC TS 62153-4-1 constitutes a technical revision. This edition includes the following significant technical changes with respect to IEC TR 62153-4-1:

- a) comparison of the frequency response of different triaxial test set-ups to measure the transfer impedance of cable screens;
- b) background of the shielded screening attenuation test method (IEC 62153-4-4);
- c) background of the shielded screening attenuation test method for measuring the screening effectiveness of feed-throughs and electromagnetic gaskets (IEC 62153-4-10);
- d) background of the shielded screening attenuation test method for measuring the screening effectiveness of RF connectors and assemblies (IEC 62153-4-7).

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

A list of all parts of the IEC 62153 series, under the general title: *Metallic communication cable test methods*, can be found on the IEC website.

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

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METALLIC COMMUNICATION CABLE TEST METHODS –

Part 4-1: Electromagnetic compatibility (EMC) – Introduction to electromagnetic (EMC) screening measurements

1 Scope

This part of IEC 62153 deals with screening measurements. Screening (or shielding) is one basic way of achieving electromagnetic compatibility (EMC). However, a confusingly large number of methods and concepts is available to test for the screening quality of cables and related components, and for defining their quality. This technical specification gives a brief introduction to basic concepts and terms trying to reveal the common features of apparently different test methods. It is intended to assist in correct interpretation of test data, and in the better understanding of screening (or shielding) and related specifications and standards.

2 Normative references

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

IEC 60096-1:1986, *Radio-frequency cables – Part 1: General requirements and measuring methods*¹

IEC 60096-4-1, *Radio-frequency cables – Part 4: Specification for superscreened cables – Section 1: General requirements and test methods*¹

IEC 60169-1-3, *Radio-frequency connectors - Part 1: General requirements and measuring methods - Section Three: Electrical tests and measuring procedures: Screening effectiveness*

IEC 61196-1:2005, *Coaxial communication cables - Part 1: Generic specification - General, definitions and requirements*

IEC 61726, *Cable assemblies, cables, connectors and passive microwave components - Screening attenuation measurement by the reverberation chamber method*

IEC 62153-4-2, *Metallic communication cable test methods - Part 4-2: Electromagnetic compatibility (EMC) - Screening and coupling attenuation - Injection clamp method*

IEC 62153-4-3, *Metallic communication cable test methods - Part 4-3: Electromagnetic compatibility (EMC) - Surface transfer impedance - Triaxial method*

IEC 62153-4-4, *Metallic communication cable test methods - Part 4-4: Electromagnetic compatibility (EMC) - Shielded screening attenuation, test method for measuring of the screening attenuation as up to and above 3 GHz*

IEC 62153-4-5, *Metallic communication cables test methods - Part 4-5: Electromagnetic compatibility (EMC) - Coupling or screening attenuation - Absorbing clamp method*

¹ This publication has been withdrawn.

IEC 62153-4-6, *Metallic communication cable test methods - Part 4-6: Electromagnetic compatibility (EMC) - Surface transfer impedance - Line injection method*

IEC 62153-4-7, *Metallic communication cable test methods - Part 4-7: Electromagnetic compatibility (EMC) - Test method for measuring the transfer impedance and the screening - or the coupling attenuation - Tube in tube method*

IEC 62153-4-9, *Metallic communication cable test methods – Part 4-9: Electromagnetic compatibility (EMC) – Coupling attenuation of screened balanced cables, triaxial method*

IEC 62153-4-10, *Metallic communication cable test methods - Part 4-10: Electromagnetic compatibility (EMC) - Shielded screening attenuation test method for measuring the screening effectiveness of feed-throughs and electromagnetic gaskets double coaxial method*

IEC/TR 62152:2009, *Transmission properties of cascaded two-ports or quadripols – Background of terms and definitions*

EN 50289-1-6: 2002, *Communication cables – Specifications for test methods Part 1-6: Electrical test methods – Electromagnetic performance*

CISPR 25, *Vehicles, boats and internal combustion engines – Radio disturbance characteristics – Limits and methods of measurement for the protection of on-board receivers*

3 Symbols interpretation

This clause gives the interpretation of the symbols used throughout this specification.

α_1, α_2	attenuation constants of primary and secondary circuit
a_s	screening attenuation
a_{sn}	normalized screening attenuation with phase velocity difference not greater than 10 % and 150 Ω characteristic impedance of the injection line ($Z_s=150 \Omega$ and $ \Delta v/v_1 =10 \%$ or $\epsilon_{r1}/\epsilon_{r2n}=1,21$)
c_0	velocity of light in free space $c_0 = 3 \times 10^8$ m/s
C_T	through capacitance of the braided cable
CUT	cable or component under test
E	e.m.f.
f	frequency
f	far end
f_c	cut-off frequency
f_{cf}	far end cut-off frequency
f_{cn}	near end cut-off frequency
Φ_1	the total flux of the magnetic field induced by the disturbing current I_1
Φ'_{12}	the direct leaking magnetic flux
Φ''_{12}	complete magnetic flux in the braid
I_1, U_1	current and voltage in the primary circuit (feeding system)
I_F	current coupled by the feed through capacitance to the secondary system (measuring system)

ε_{r1}	relative permittivity of the injection line (feeding system)
ε_{r2}	relative permittivity of the cable (measuring system)
L	cable length, coupling length
L_1	(external) inductance of the outer circuit
L_2	(external) inductance of the inner circuit
M'_{12}	mutual inductance related to direct leakage of the magnetic flux Φ'_{12}
M''_{12}	mutual inductance related to the magnetic flux Φ''_{12} (or $\frac{1}{2} \Phi''_{12}$) in the braid
	$M'_{12} = \frac{\Phi'_{12}}{j\omega I_1}$ and $M''_{12} = \frac{1}{2} \cdot \frac{\Phi''_{12}}{j\omega I_1}$
M_T	effective mutual inductance per unit length for braided screens
	$M_T = M'_{12} - M''_{12}$
	where M'_{12} relates to the direct leakage of the magnetic flux and M''_{12} relates to the magnetic flux in the braid [24]
n	near end
P_1	sending power
P_{2f}	far end measured power
P_{2n}	near end measured power
P_r	radiated power in the environment of the cable, which is comparable to $P_{2n} + P_{2f}$ of the absorbing clamp method of 12.4 of IEC 61196-1:1995
P_s	radiated power in the normalised environment of the cable under test ($Z_s = 150 \Omega$ and $ \Delta v/v_1 = 10\%$ or $\varepsilon_{r1}/\varepsilon_{r2n} = 1,21$)
R	load resistance of secondary circuit (input resistance of receiver)
R_T	screen resistance per unit length
T	coupling transfer function
T_f	far end transfer function
T_n	near end transfer function
U'_{12}	the disturbing voltage induced by Φ'_{12}
U''_{rh}	the disturbing voltage induced by $\frac{1}{2} \Phi''_{12}$ of the right hand lay contribution
U''_{lh}	the disturbing voltage induced by $\frac{1}{2} \Phi''_{12}$ of the left hand lay contribution
U''_{12}	is equal to U''_{rh} and U''_{lh} (= the disturbing voltage induced by $\frac{1}{2} \Phi''_{12}$)
v	phase velocity
v_1	phase velocity of the "primary" system (feeding system)
v_2	phase velocity of the "secondary" system (measuring system)
v_{r1}	relative phase velocity of the "primary" system (feeding system)
v_{r2}	relative phase velocity of the "secondary" system (measuring system)
Z_1	characteristic impedance of the "primary" system (feeding system or line (1))
Z_2	characteristic impedance of the cable under test (CUT) (measuring system or line (2))
Z_{1f}	terminating impedance of the line (1) in the far end
Z_{2n}	terminating impedance of the line (2) in the near end

Z_{2f}	terminating impedance of the line (2) in the far end (in a matched set-up) $Z_{1f} = Z_1$ and $Z_{2n} = Z_{2f} = Z_2$ $Z_{12} = \sqrt{Z_1 Z_2}$
Z_a	surface impedance of the braided cable
Z_F	capacitive coupling impedance per unit length
Z_f	capacitive coupling impedance
Z_T	surface transfer impedance per unit length
Z_{Th}	transfer impedance of a tubular homogeneous screen per unit length
Z_t	surface transfer impedance
Z_{TEn}	effective transfer impedance ($= Z_F + Z_T $) per unit length in the near end
Z_{TEf}	effective transfer impedance ($= Z_F - Z_T $) per unit length in the far end
$Z_{TEn,f}$	effective transfer impedance ($= Z_F \pm Z_T $) per unit length in the near end or in the far end
Z_{TE}	effective transfer impedance ($= \max Z_{TEn}, Z_{TEf} $) per unit length
Z_{te}	effective transfer impedance ($= \max Z_f \pm Z_t $)
Z_{ten}	normalized effective transfer impedance of a cable ($Z_1 = 150 \, \Omega$ and $ v_1 - v_2 / v_2 \leq 10 \, \%$ velocity difference in relation to velocity of CUT)

4 Electromagnetic phenomena

It is assumed that if an electromagnetic field is incident on a screened cable, there is only weak coupling between the external field and that inside, and that the cable diameter is very small compared with both the cable length and the wavelength of the incident field. The superposition of the external incident field and the field scattered by the cable yields the total electromagnetic field $(\vec{E}_t, \vec{H}_t)$ in Figure 1. The total field at the screen's surface may be considered as the source of the coupling: electric field penetrates through apertures by electric or capacitive coupling; also magnetic fields penetrate through apertures by inductive or magnetic coupling. In addition, the induced current in the screen results in conductive or resistive coupling.

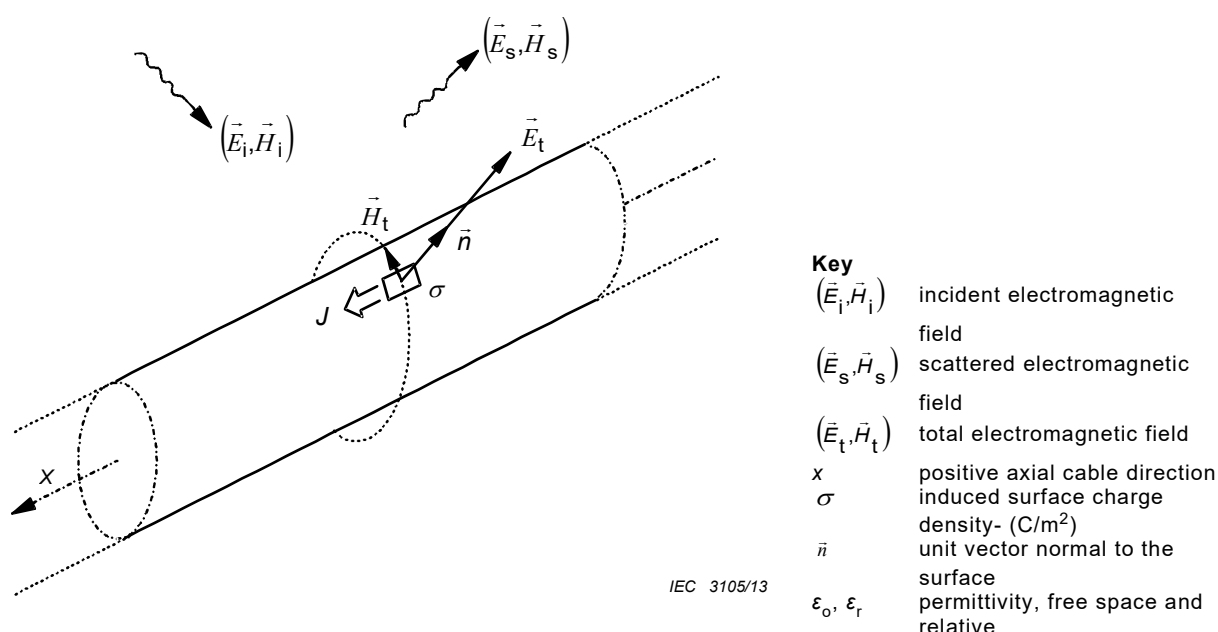


Figure 1 – Total electromagnetic field $(\vec{E}_t, \vec{H}_t)$

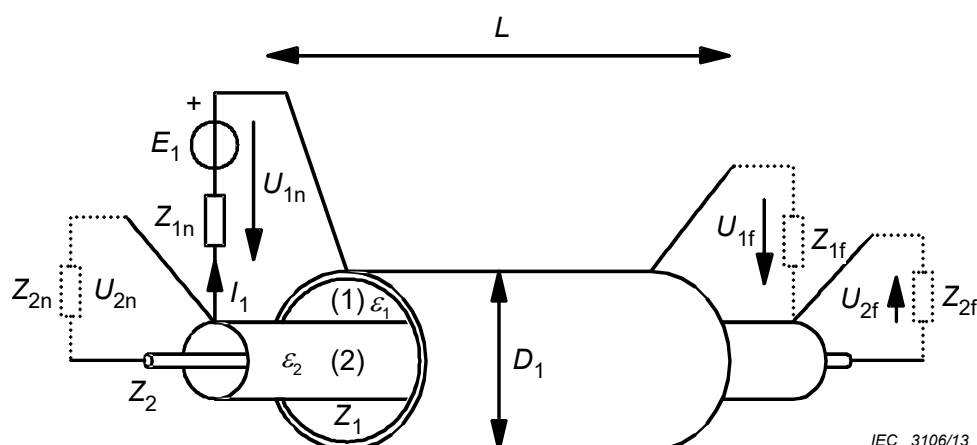
$$(\vec{E}_t, \vec{H}_t) = (\vec{E}_i, \vec{H}_i) + (\vec{E}_s, \vec{H}_s) \quad (1)$$

$$J = \vec{n} \cdot \vec{H}_t \quad (2)$$

$$\sigma = \vec{n} \cdot \vec{E}_t \epsilon_0 \epsilon_r \quad (3)$$

where the symbols are described in the key of Figure 1.

As the field at the surface of the screen is directly related to density of surface current and surface charge, the coupling may be assigned either to the total field $(\vec{E}_t, \vec{H}_t)$ or to the surface current- and charge- densities (J and σ). Consequently, the coupling into the cable may be simulated by reproducing, through any suitable means, the surface currents and charges on the screen. Because the cable diameter is assumed to be small, the higher modes may be neglected and it is possible to use an additional coaxial conductor as the injection structure, as shown in Figure 2.



Key (for Figures 2,3,4,5)

(1), (2)	outer circuit (1), tube, respectively inner circuit (2), cable
$Z_{1,2}$	characteristic impedance of the outer circuit (1), tube, respectively inner circuit (2), cable
$\varepsilon_{1,2}$	dielectric permittivity of the outer circuit (1), tube, respectively inner circuit (2), cable
$\beta_{1,2}$	phase constant of the outer circuit (1), tube, respectively inner circuit (2), cable
$\lambda_{1,2}$	wave length of the outer circuit (1), tube, respectively inner circuit (2), cable
L	coupling length
D_1	diameter of injection cylinder-tube
V	voltmeter
A	ammeter
Z_{1n}, Z_{1f}	load resistance at the near end, respectively far end of the outer circuit (1), tube
Z_{2n}, Z_{2f}	load resistance at the near end, respectively far end of the inner circuit (2), cable
E_1	EMF of the generator
I_1, I_2	current in the outer circuit (1), tube, respectively inner circuit (2), cable
U_{1n}, U_{1f}	voltage at the near end, respectively far end of the outer circuit (1), tube
U_{2n}, U_{2f}	voltage at the near end, respectively far end of the inner circuit (2), cable

Figure 2 – Defining and measuring screening parameters – A triaxial set-up

Figure 2 shows the concept of a triaxial set-up. The outer circuit (1) is formed by an injection cylinder-tube and the screen under test, with an characteristic impedance Z_1 . The inner circuit (2) is formed by the screen under test, and centre conductor, with an characteristic impedance Z_2 . The screening at the ends of circuit (2) is not shown. Observe the conditions Z_{1f} , Z_{2n} , Z_{2f} and λ in Figure 3 and Figure 4. Also note that diameter of the injection cylinder tube (D_1) shall be much smaller than the coupling length (L).

5 The intrinsic screening parameters of short cables

5.1 General

The intrinsic parameters refer to an infinitesimal length of cable, like the inductance or capacitance per unit length of transmission lines. Assuming electrically short cables, with $L \ll \lambda$ which will always apply at low frequencies, the intrinsic screening parameters are defined and can be measured as indicated in the subclauses 5.2 and 5.3.

5.2 Surface transfer impedance, Z_T

As shown in Figure 3, where Z_{1f} and Z_{2f} are zero, the surface transfer impedance (Z_T in Ω/m) is given:

$$Z_T = \frac{U_{2n}}{I_1 \cdot L} \quad (4)$$

where

Z_T is the transfer impedance, U_{2n} is the voltage at the near end of the inner circuit (2),

L is the coupling length I_1 is the current in the outer circuit (1).

The dependence of Z_T on frequency is not simple and is often shown by plotting $\log Z_T$ against $\log$ frequency. Note that the phase of Z_T may have any value, depending on braid construction and frequency range.

NOTE In circuit (2) of Figure 3, the voltmeter and short circuit may also be interchanged.

5.3 Capacitive coupling admittance, Y_C

As shown in Figure 4, where Z_{1f} and Z_{2f} are open circuit, the capacitive coupling admittance (Y_C in S/m) is given by:

$$Y_C = j \cdot \omega C_T = \frac{I_2}{U_{in} \cdot L} \quad (5)$$

where

Y_C is the coupling admittance C_T is the through capacitance;

ω is the radian frequency; j is the imaginary operator

L is the coupling length I_2 is the current in the inner circuit (2).

The through capacitance C_T is a real capacitance and has usually a constant value up to 1 GHz and higher (with aperture $a \ll \lambda$).

While Z_T is independent of the characteristics of the coaxial circuits (1) and (2), C_T is dependent on those characteristics. There are two ways of overcoming this dependence:

- a) The normalized through elastance K_T (with units of m/F) derived from C_T is independent of the size of the outer coaxial circuit (2), but it depends on its permittivity:

$$K_T = C_T / (C_1 \cdot C_2) \quad (6)$$

$$K_T \sim 1 / (\epsilon_{r1} + \epsilon_{r2}) \quad (7)$$

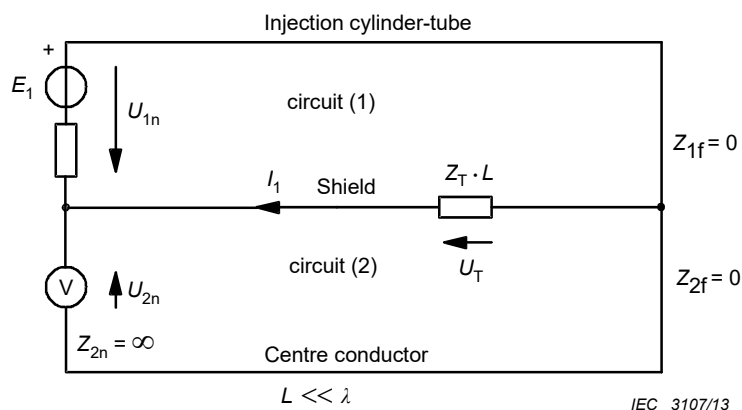
where C_1 and C_2 are the capacitance per unit length of the two coaxial circuits.

- b) The capacitive coupling impedance Z_F (with units of Ω/m) again derived from C_T is also independent of the size of the outer coaxial circuit (2) and, for practical values of ϵ_{r1} , is only slightly dependent on its permittivity:

$$Z_F = Z_1 Z_2 Y_C = Z_1 Z_2 j \omega C_T \quad (8)$$

$$Z_F \sim \sqrt{(\epsilon_{r1} \cdot \epsilon_{r2})} / (\epsilon_{r1} + \epsilon_{r2}) \quad (9)$$

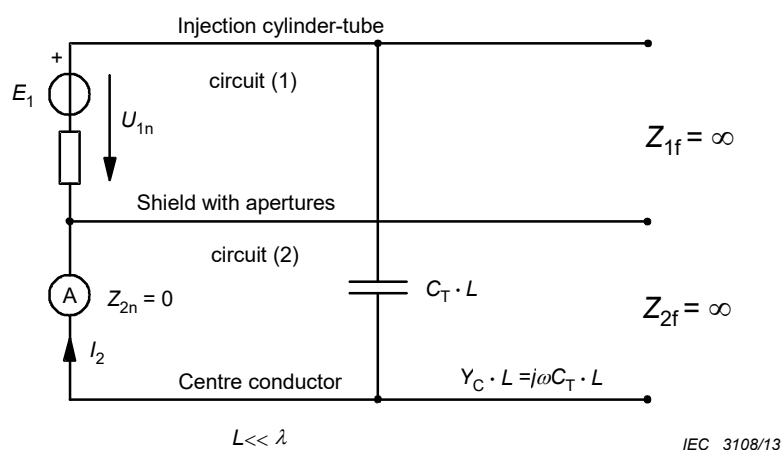
Compared with Z_T , Z_F is usually negligible, except for open weave braids. It may, however, be significant when Z_{2n} and $Z_{2f} \gg Z_2$ (audio circuits).



Key

See Figure 2.

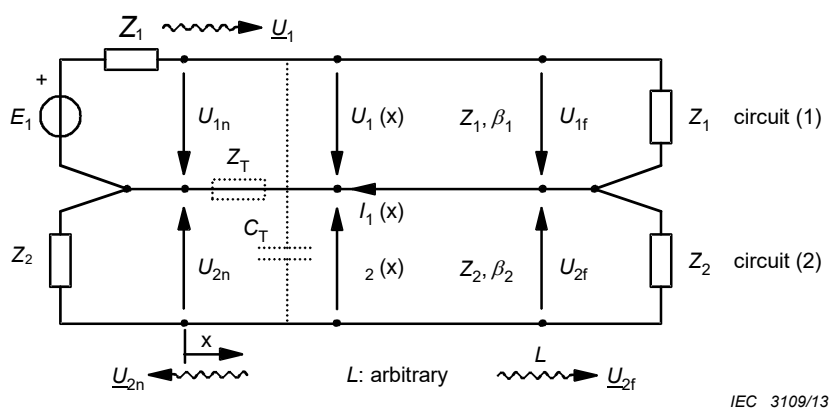
Figure 3 – Equivalent circuit for the testing of Z_T



Key

See Figure 2.

Figure 4 – Equivalent circuit for the testing of $Y_C = j \omega C_T$



Key

See Figure 2.

NOTE Z_T and C_T are distributed (not correctly shown here). The loads Z_1 , Z_2 at the ends may represent matched receivers.

Figure 5 – Electrical quantities in a set-up that is matched at both ends

5.4 Injecting with arbitrary cross-sections

A coaxial outer circuit (2) has been assumed so far in this report, but it is not essential because of the invariance of Z_T and Z_F . Using a wire in place of the outer cylinder, the injection circuit (2) becomes two-wire with the return via the screen of the cable under test. Obviously the charge and current distribution become non-uniform, but the results are equivalent to coaxial injection, especially if two injection lines are used opposite to each other, and may be justified for worst-case testing. Note that the IEC line injection test uses a wire.

5.5 Reciprocity and symmetry

Assuming linear shield materials, the measured Z_T and Z_F values will not change when interchanging the injection circuit (1) and the measuring circuit (2). Each of the two conductors of the two-line circuit can be interchanged, but in practice the set-up will have to take into account possible ground loops and coupling to the environment.

5.6 Arbitrary load conditions

When the circuit ends of Figure 3 and Figure 4 are not ideally a short or open circuit, Z_T and Z_F will act simultaneously. Their superposition is noticeable in the low frequency coupling of the matched circuit (1) and circuit (2) (see Figure 5 and Table 1).

6 Long cables – coupled transmission lines

The coupling over the whole length of the cable is obtained by summing up (integrating) the infinitesimal coupling contributions along the cable while observing the correct phase. The analysis utilizes the following assumptions and conventions:

- matched circuits considered with the voltage waves $\underline{U}_1$, $\underline{U}_{2n}$, $\underline{U}_{2f}$, see Figure 5,
- representation of the coupling, using the normalized wave amplitudes $\underline{U}/\sqrt{Z} [\sqrt{\text{Watt}}]$, instead of voltage waves. i.e. the coupling transfer function, in the following denoted by "coupling function", will be defined as

$$T_n = \frac{\underline{U}_{2n} / \sqrt{Z_2}}{\underline{U}_1 / \sqrt{Z_1}} \quad (10)$$

$$T_f = \frac{\underline{U}_{2f} / \sqrt{Z_2}}{\underline{U}_1 / \sqrt{Z_1}} \quad (11)$$

The square of the coupling transfer function, $|T|^2$, is the ratio of the power waves travelling in circuits (2) and (1). Due to reciprocity and assuming linear screen (shield) materials, T is reciprocal, i.e. invariant with respect to the interchange of injection and measuring circuits (1) and (2). The quantity $|1/T|^2$ or in logarithmic quantities

$$a_s = -20 \times \log_{10} |T| \quad (12)$$

may be considered as the "screening attenuation" of the cable, specific to the set-up.

Performing the straight forward calculations of coupled transmission line theory, the coupling function T , given in Table 1, is obtained. The term $S\{L \cdot f\}$ is the "summing function" S , being dependent on L and f . (The wavy bracket just indicates that the product $L \cdot f$ is the argument of the function S and not a factor to S). S represents the phase effect, when summing up the infinitesimal couplings along the line, and is:

$$S_{n,f}\{L \cdot f\} = \frac{\sin \frac{\beta L \pm}{2}}{\frac{\beta L \pm}{2}} \exp\left(-j \cdot \frac{\beta L +}{2}\right) \quad (13)$$

$$\beta L + = (\beta_2 + \beta_1) \cdot L \quad (14)$$

$$\beta L \pm = (\beta_2 \pm \beta_1) \cdot L \quad (15)$$

$$\beta L \pm = 2\pi L f \cdot (1/\nu_2 \pm 1/\nu_1) \quad (16)$$

$$\beta L \pm = 2\pi L f \cdot (\sqrt{\varepsilon_{r2}} \pm \sqrt{\varepsilon_{r1}})/c \quad (17)$$

subscript $\pm$ refers to near/far end respectively; i.e. + indicates the near end and – indicates the far end;

+ refers to both near/far ends.

Note that weak coupling, i.e. $T \ll 1$, has been assumed. This case, including losses, is given in [1]².

Equation (18) and the representation in Table 1 illustrate the contributions of the different parameters to the coupling function T :

$$T_{n,f} = (Z_F \pm Z_T) \cdot \frac{1}{\sqrt{Z_1 \cdot Z_2}} \cdot \frac{L}{2} \cdot S_{n,f}\{L \cdot f, \varepsilon_{r1}, \varepsilon_{r2}\} \quad (18)$$

² Figures in square brackets refer to the bibliography.

Table 1 – The coupling transfer function T (coupling function)^a

Set-up parameters ^b $(Z_1), L, \varepsilon_{r1}$ $T_n = (Z_F \pm Z_T) \cdot \frac{1}{\sqrt{Z_1 \cdot Z_2}} \cdot \frac{L}{2} \cdot S_n \{L \cdot f, \varepsilon_{r1}, \varepsilon_{r2}\}$	
Intrinsic screen parameters "Low-frequency coupling", short cables ^c	Cable parameters ^b $(Z_2, L), \varepsilon_{r2}$ "HF-effect", cut-off $(L \cdot f)_c$ Length + frequency effect
^a T^2 is the power coupling from circuit (1) to circuit (2). The stacked subscripts _f are associated to the stacked operation symbols $\pm$ in the obvious way: upper subscript $\rightarrow$ upper operation, lower subscript $\rightarrow$ lower operation. ^b ε_{r1} and ε_{r2} contained in S as parameters. ^c for $L \ll \lambda$: $S\{L \cdot f\} \rightarrow 1$.	

Note especially the following points.

- There may be a directional effect ($T_n \neq T_f$) in the whole frequency range if Z_F is not negligible. (But Z_F is usually negligible except with loose, single braid shields.)
- Up to a constant factor, T is the quantity directly measured in a set-up.
- For low frequencies, i.e. for short cables ($L \ll \lambda$), the trivial coupling formula is obtained that is directly proportional to L :

$$T_n = (Z_F \pm Z_T) \cdot \frac{1}{Z_{12}} \cdot \frac{L}{2} \quad (19)$$

where

$$Z_{12} = \sqrt{Z_1 \cdot Z_2}$$

- The summing function $S\{L \cdot f\}$ is presented in Figure 6.
- $S\{L \cdot f\}$ has a $\sin(x)/x$ behaviour. A cut-off point may be defined as $(L \cdot f)_c$:

$$(L \cdot f)_{c_n} = \frac{c}{\pi \left| \sqrt{\varepsilon_{r1}} \pm \sqrt{\varepsilon_{r2}} \right|} \quad (20)$$

- The exact envelope of $S\{L \cdot f\}$ is

$$\text{Env} \left| S_n \left\{ L \cdot f \right\} \right|_f = \frac{1}{\sqrt{1 + \frac{(L \cdot f)^2}{(L \cdot f)_{cn}^2}}} \quad (21)$$

g) The first minimum (zero) of $S\{L \cdot f\}$ occurs at

$$(L \cdot f)_{\min} = \pi(L \cdot f)_c \quad (22)$$

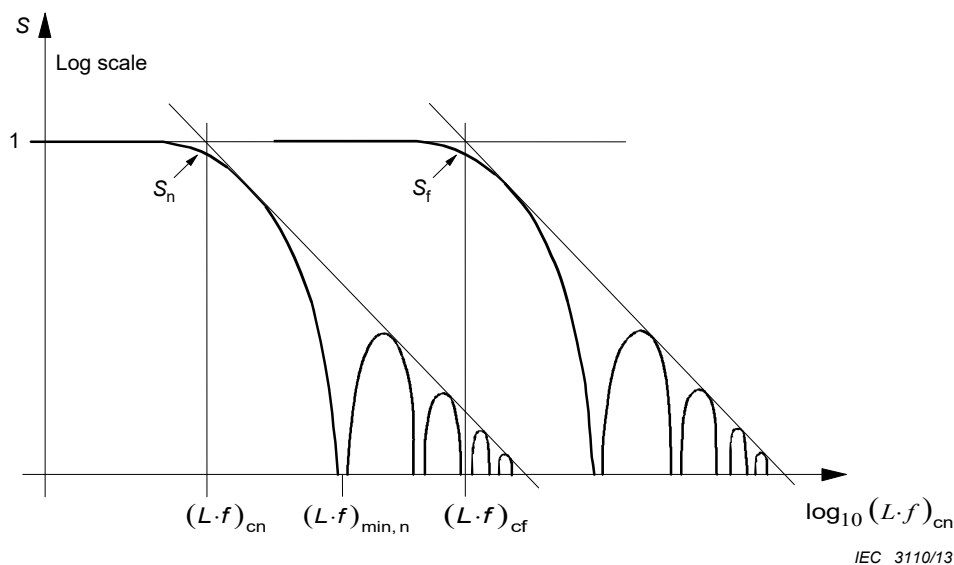
h) As seen from Equations (13) and (21), below the cut-off points $(L \cdot f)_{cn}$ is $S\{L \cdot f\} \approx 1$ and above them it starts to oscillate and its envelope drops asymptotically 20 dB/decade,

$$\text{Env} \left| S_n \left\{ L \cdot f \right\} \right|_f \approx \frac{\left((L \cdot f)_{cn} \right)}{(L \cdot f)} \quad (23)$$

i) S is symmetrical in L and f , i.e. L and f are interchangeable. For a fixed length a cut-off frequency f_c and vice versa, for a fixed frequency a cut-off length L_c may be defined. Substituting c/λ_o for f , we obtain the cut-off length as

$$L_{cn} = \frac{\lambda_o}{\pi \left| \sqrt{\varepsilon_{r1}} \pm \sqrt{\varepsilon_{r2}} \right|} \quad (24)$$

- j) The effect of S in the frequency range ($L = \text{constant}$) is illustrated in Figure 8. The coupling function is proportional to Z_T , only if $f < f_c$. Note also the typical values indicated for f_c .
- k) The minima and maxima of S are not resonances, they are due to cancelling and additive effects of the coupling along the line.
- l) The far end cut-off frequency is significantly influenced by the permittivity of the outer system (ε_{r1}). Selecting $\varepsilon_{r1} \rightarrow \varepsilon_{r2}$ we obtain $(L \cdot f)_{cf} \rightarrow \infty$, i.e. no cut-off at the far end. Due to practical aspects (tolerances, homogeneity, etc.), an ideal phase matching ($\varepsilon_{r1} \equiv \varepsilon_{r2}$) is not feasible.
- m) The effects of Z_T and Z_F on the coupling transfer functions T_n and T_f are shown in Figure 8.
- n) The total effect of L on the coupling is not contained in S alone, but in the product $L \cdot S\{L \cdot f\}$. The product $L \cdot S$ is presented in Figure 12 for $f = \text{constant}$. The coupling function T which can be measured in a set-up is proportional to L if $L < L_c$. However, for appropriately long cables ($L > L_c$), the maximum coupling is independent of L and we obtain a length independent shielding attenuation above the cut-off point $(L \cdot f)_c$. But we should remember that $(L \cdot f)_c$ as well as A_s are still dependent on the set-up parameters (ε_{r1}, Z_1) .

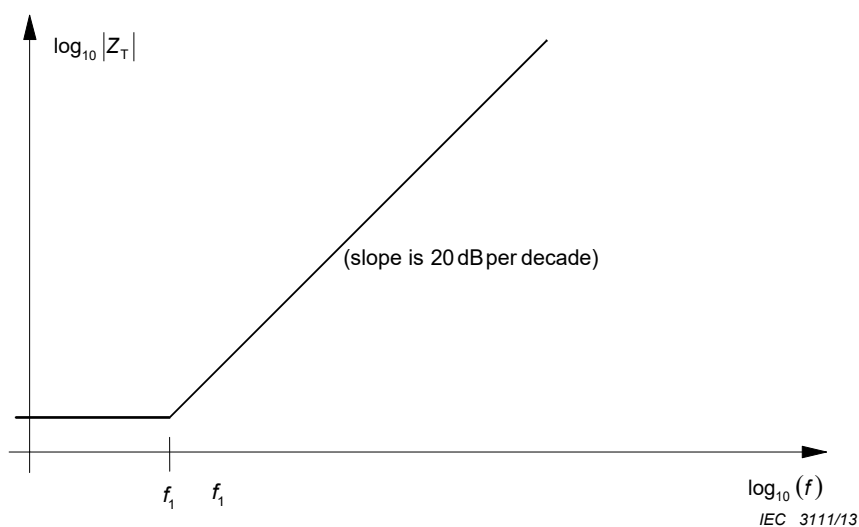


Key

$(L \cdot f)_{cn,f}$ cut-off point at near (n) respectively far (f) end
 $S_{n,f}$ summing function at the (n) respectively far (f) end

NOTE $S_f > S_n$ above near end cut-off, yielding a directive effect.

Figure 6 – The summing function $S\{L \cdot f\}$ for near and far end coupling



Key

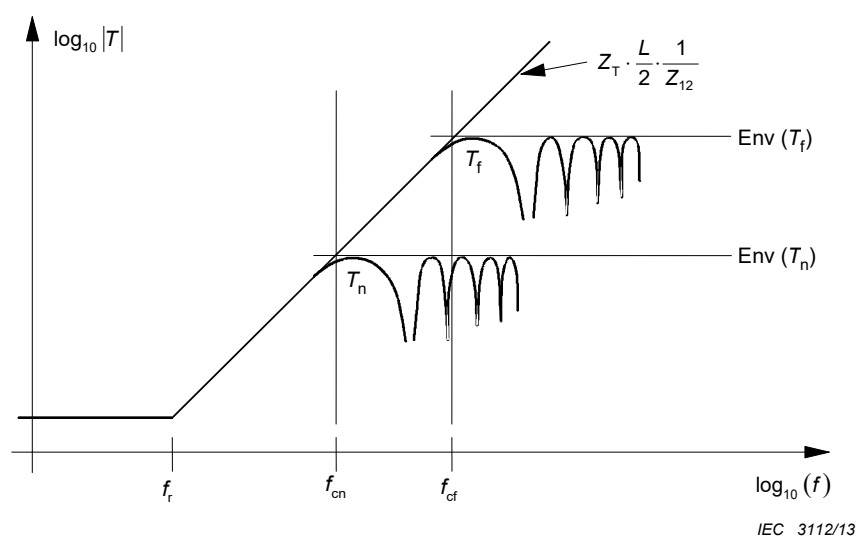
$\log_{10}|Z_T|$ magnitude of the transfer impedance drawn on a logarithmic scale

$\log_{10}(f)$ frequency drawn on a logarithmic scale

f_1 frequency of the intersection of the DC resistance of the screen and the 20dB slope at higher frequencies

Figure 7 – Transfer impedance of a typical single braid screen

Figure 8 gives the result of adding (on a log scale) the frequency responses from Figure 6 and Figure 7. It is assumed the cable has a negligible capacitive coupling impedance Z_F ($Z_F \ll Z_T$).

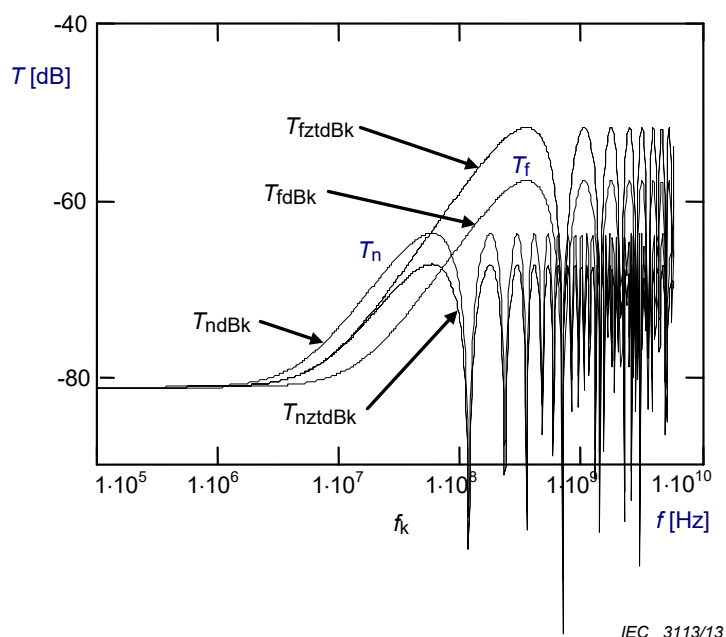


Key

- $T_{n,f}$ coupling transfer function at the (n) respectively far (f) end
 $Env(T_{n,f})$ envelope of the coupling transfer function at the (n) respectively far (f) end
 $f_{cn,f}$ cut-off frequency at the (n) respectively far (f) end

Example: $\epsilon_{r1} = 1$ (set-up), $\epsilon_{r2} = 2,2$ (cable), $L = 1$ m; results in $f_{cn} = 40$ MHz; $f_{cf} = 200$ MHz

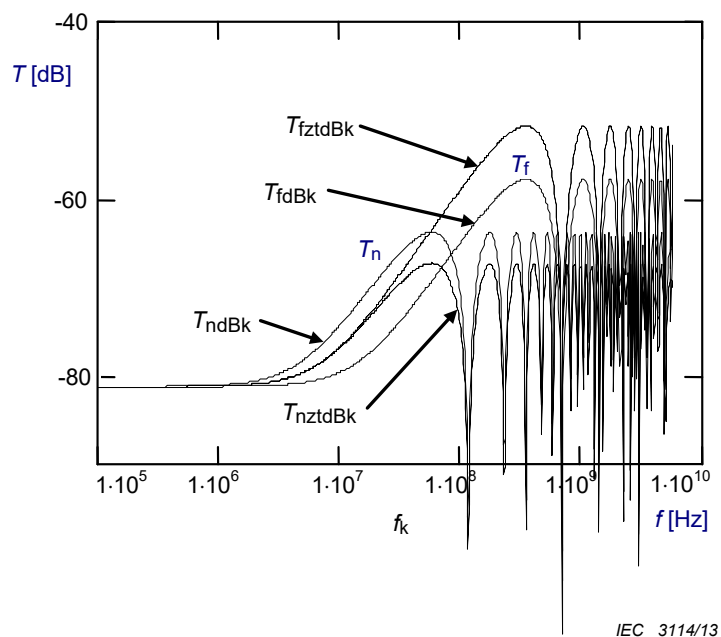
Figure 8 – The effect of the summing function on the coupling transfer function of a typical single braid screen cable



In calculations the following parameters are used:

Z_T (d.c.) = 15 mΩ/m and Z_T (10 MHz) = 20 mΩ/m increasing 20 dB/decade (see Figure 7), cable length 1 m, and velocities of the outer and inner line: $v_1 = 200$ Mm/s and $v_2 = 280$ Mm/s corresponding to a velocity difference of 40 %.

Figure 9 – Calculated coupling transfer functions T_n and T_f for a single braid – $Z_F = 0$

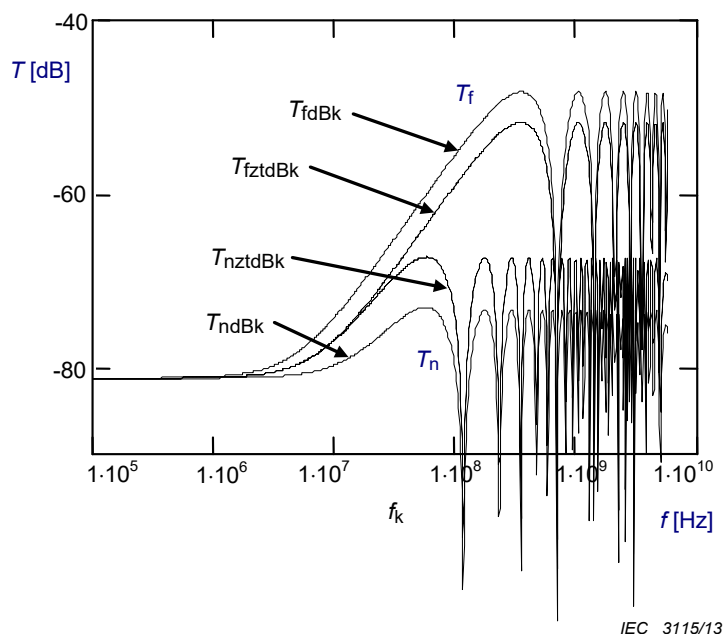


T_n is 3,5 dB higher and T_f is 6 dB lower than in reference Figure 9 because

$$T_n \sim |Z_F + Z_T| = 1,5 \times Z_T \text{ and}$$

$$T_f \sim |Z_F - Z_T| = 0,5 \times Z_T$$

Figure 10 – Calculated coupling transfer functions T_n and T_f for a single braid – $\text{Im}(Z_T)$ is positive and $Z_F = +0,5 \times \text{Im}(Z_T)$ at high frequencies



T_f is 3,5 dB higher and T_n is 6 dB lower than in reference Figure 9 because

$$T_f \sim |Z_F - Z_T| = 1,5 \times |Z_T| \text{ and}$$

$$T_n \sim |Z_F + Z_T| = 0,5 \times |Z_T|$$

Figure 11 – Calculated coupling transfer functions T_n and T_f for a single braid – $\text{Im}(Z_T)$ is negative and $Z_F = -0,5 \times \text{Im}(Z_T)$ at high frequencies

In Figure 9, $Z_F = 0$ and Z_T is positive.

In Figure 10 and Figure 11, Z_F is significant ($Z_F = (1/2) \times Z_T$).

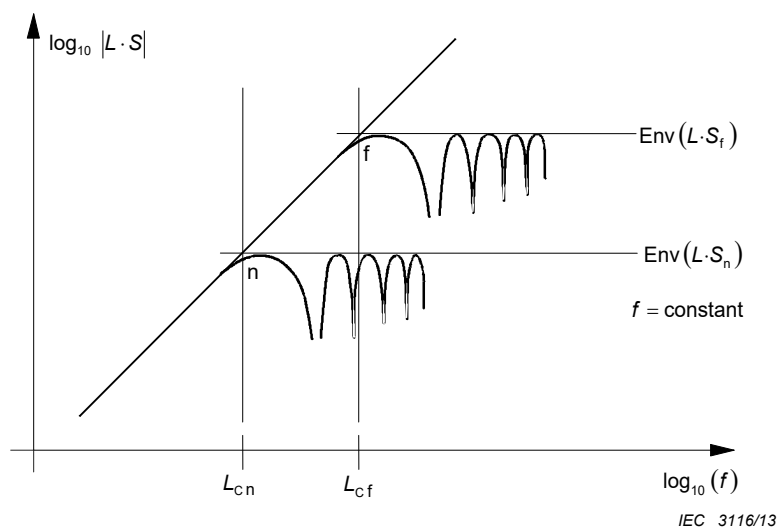
In Figure 11, the imaginary part of Z_T is negative at high frequencies.

The following notes apply to Figure 9 to Figure 11.

NOTE 1 T_n for near-end, T_f for far-end and dB means that $T_{n,f}$ are calculated in dB ($20 \times \log_{10} |T_{n,f}|$).

NOTE 2 T_n dB: near-end when $Z_F = (1/2) \times Z_T$ and T_{nzt} dB: near-end when $Z_F = 0$.

NOTE 3 T_f dB: far-end when $Z_F = (1/2) \times Z_T$ and T_{fzt} dB: far-end when $Z_F = 0$.



NOTE 1 For $L > L_c$, the maximum value of T is attained, i.e. the maximum coupling (or the screening attenuation) is not dependent on L .

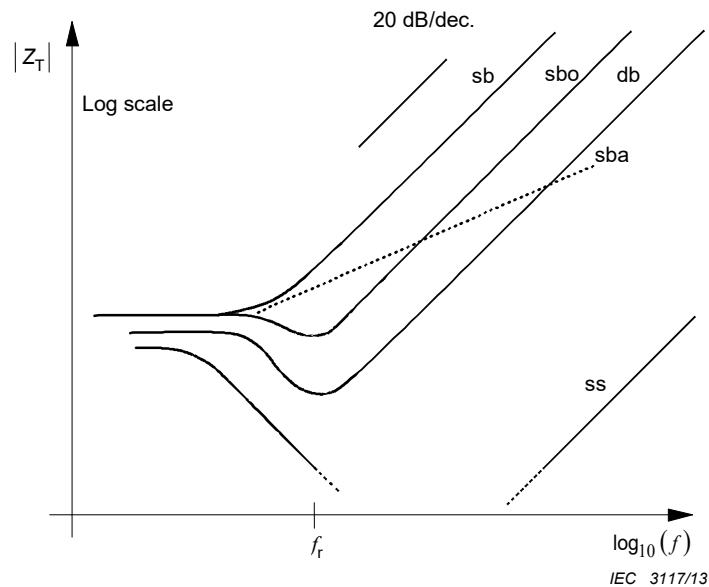
NOTE 2 L_{cf} strongly depends on ϵ_{r1} .

NOTE 3 See also Table 1 and list item n)

Figure 12 – $L \cdot S$: the complete length dependent factor in the coupling function T

7 Transfer impedance of a braided wire outer conductor or screen

Typical transfer impedances of cables with braided wire screens are shown in Figure 13. The constant Z_T value at the low-frequency end is equal to the DC resistance of the screen, the 20 dB per decade rise at the high-frequency end is due to the inductive coupling through the screen and the dip at the middle frequencies is caused by eddy currents or skin effect of the braid. Some braided cables may behave anomalously having less than a 20 dB per decade rise at high frequencies. By using an extrapolation of 20 dB per decade we are in most cases on the conservative side. This extrapolation can be used up to several GHz.



Key

- f_r : typically 1....10 MHz
- sb: single braid
- sbo: single braid optimized
- sba: single braid 'anomalous'
- db: double braid
- ss: superscreen

Figure 13 – Transfer impedance of typical cables

An electrically short piece of braided coaxial cable (2) is considered to be placed in a triaxial arrangement as in Figure 2.

It is assumed that the outer circuit (1) is the disturbing one. As stated, a braided cable has a transfer impedance Z_T that increases proportionally to frequency at high frequencies, because of the leakage of the magnetic field through holes in the braid.

The total flux of the magnetic field induced by the disturbing current I_1 is Φ_1 . A part of it, Φ'_{12} leaks directly through the holes and includes a disturbing voltage U'_2 in the inner circuit. However, a part Φ''_{12} of Φ_1 flows in the braid and complicates the mechanism of the total magnetic leakage by the following additional phenomenon.

The braiding wires alternate between the outer and inner layer. It means that the inner and outer braid wires are likewise ingredients of both the inner (2) and outer (1) circuit of Figure 14.

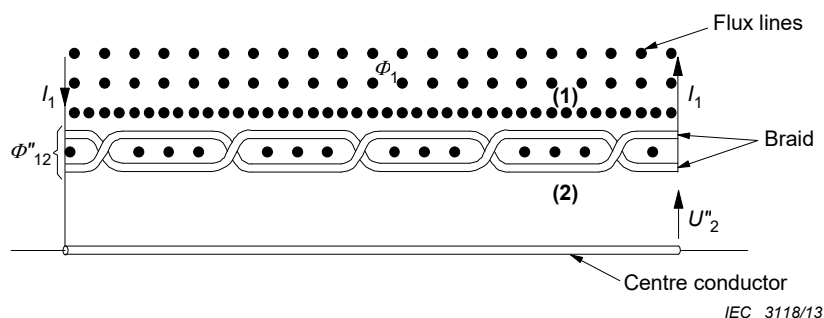


**Figure 14 – Magnetic coupling in the braid –
Complete flux**

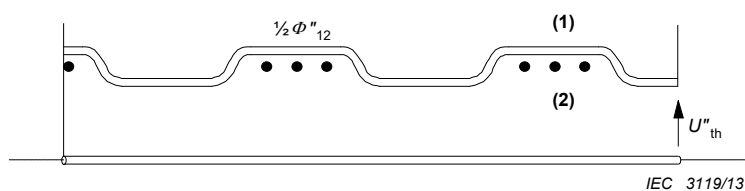


**Figure 15 – Magnetic coupling in the braid –
Left-hand lay contribution**

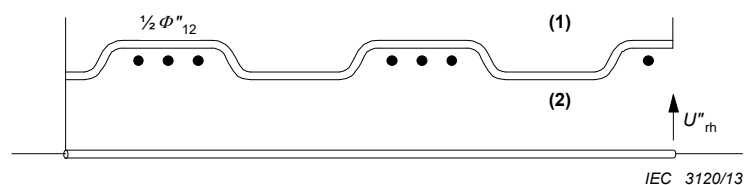


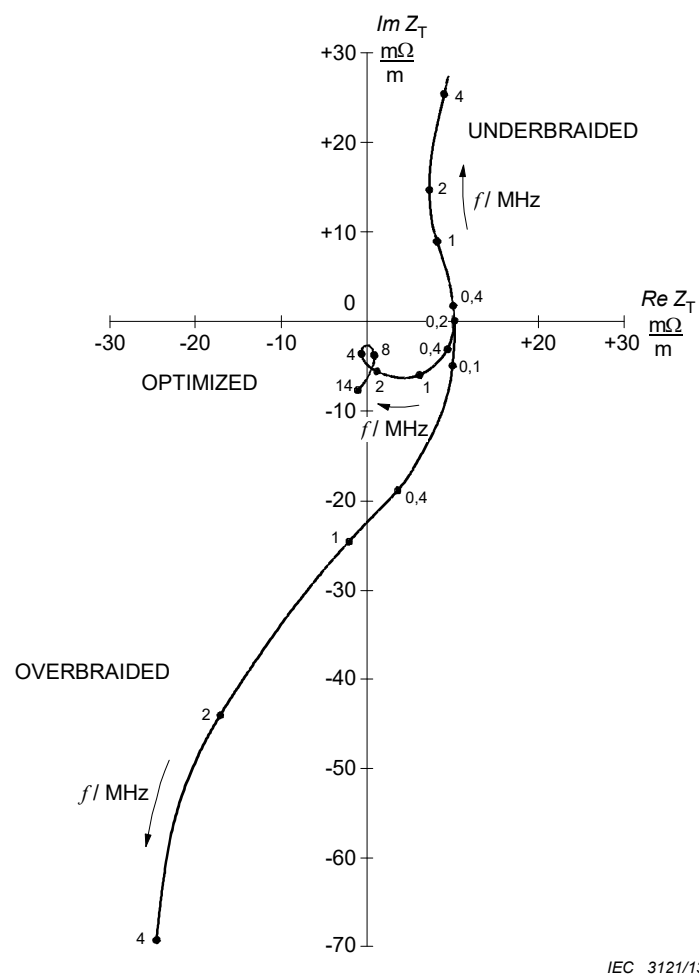
**Figure 16 – Magnetic coupling in the braid –
Right-hand lay contribution**

Therefore it is necessary and unavoidable that Φ''_{12} is partly also in the inner circuit (see Figure 14). Both the left hand (lh) (see Figure 15) and right hand (rh) lay (see Figure 16) of the braiding wires bring into the inner circuit (2) an equal disturbing voltage U''_2 induced by $\Phi''_{12} / 2$. The voltages are in parallel:

$$U''_{rh} = U''_{lh} = U''_2 = \frac{1}{2} j \omega \Phi''_{12} \quad (25)$$

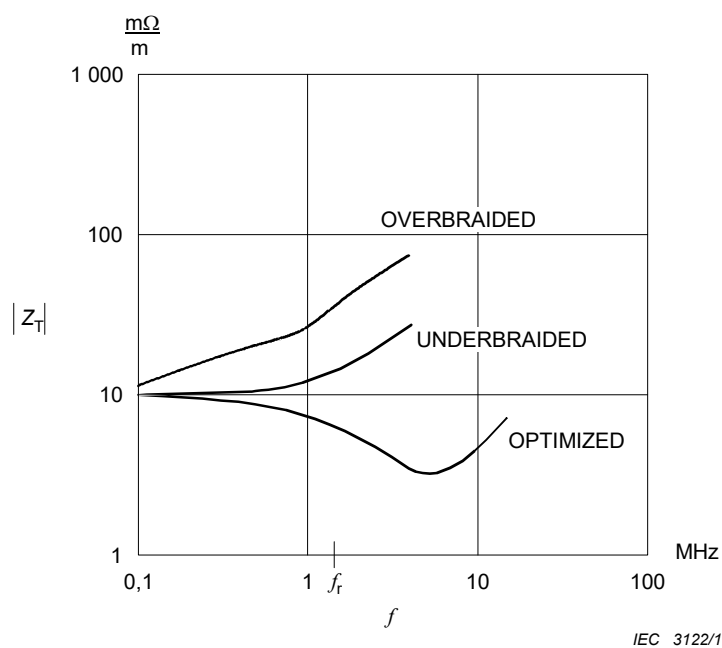
This phenomenon is similar to the "magnetic part" of the coupling through a homogeneous screen.

The two induced disturbing voltages oppose each other.



IEC 3121/13

Figure 17 – Complex plane, $Z_T = Re Z_T + j Im Z_T$, frequency f as parameter



IEC 3122/13

Figure 18 – Magnitude (amplitude), $|Z_T(f)|$

In Figure 17 and Figure 18, the d.c., resistance Z_T (d.c.), is set to the value of 10 mΩ/m.

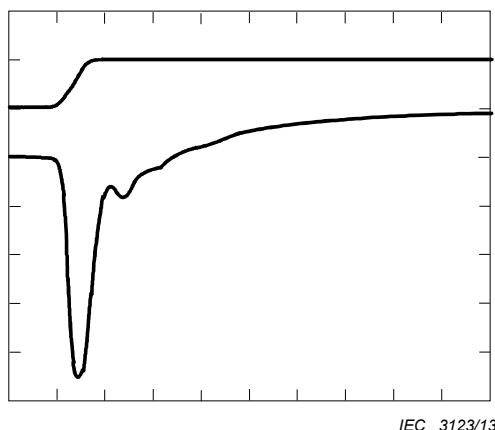


Figure 19a – Overbraided cable

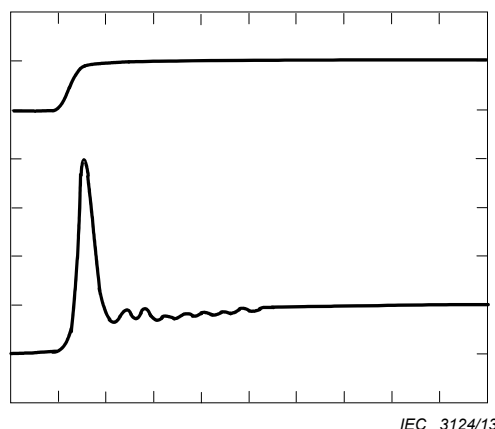


Figure 19b – Underbraided cable

Top trace: Injection step current (100 mA/div)

Time base: 50 ns/div

Amplifier gain: 30 dB, therefore Z_T (time) = 12,5 mΩ/m/div

Lower trace: The height of the spike corresponds to

a) – Z_T (3 MHz) = $-4,7 \times 12,5 \text{ m}\Omega/\text{m} = -59 \text{ m}\Omega/\text{m}$;

b) – Z_T (3 MHz) = $+4 \times 12,5 \text{ m}\Omega/\text{m} = +50 \text{ m}\Omega/\text{m}$.

Figure 19 – Typical Z_T (time) step response of an overbraided and underbraided single braided outer conductor of a coaxial cable

Braid optimization is based on these important physical facts. Both leakage phenomena can be described by mutual inductances:

$$M'_{12} = \frac{\Phi'_{12}}{j\omega I_1} \quad (26)$$

$$M''_{12} = \frac{1}{2} \times \frac{\Phi''_{12}}{j\omega I_1} \quad (27)$$

Clearly it is possible to make braided-wire screens where either M'_{12} or M''_{12} are dominant or where they cancel each other. Therefore, underbraided, overbraided or optimized braids may be considered. Figure 17 shows measured transfer impedances in the complex plane of such screens and the main transfer impedance components of a braided screen can be observed. From the optimized case, it can be concluded that at low frequencies the braid behaves approximately as a homogeneous tubular screen. The same can be concluded from Figure 18 where the transfer impedance amplitudes are shown as a function of frequency, but from it cannot be seen directly if the screen is underbraided or overbraided.

The transfer impedance of a braided wire screen consists of the following three main components (mentioned above).

- a) At low and medium frequencies, the tubular screen coupling behaviour (Z_{Th}) varies with eddy currents and decreasing Z_T . In [2] it is stated that a good approximation for Z_{Th} is a tubular homogeneous screen [3] with the thickness of one wire diameter and the same d.c. resistance as the braid.
- b) The mutual inductance M'_{12} is related to direct leakage of the magnetic flux Φ'_{12} .
- c) The mutual inductance M''_{12} (negative) is related to the magnetic flux Φ''_{12} in the braid.

By adding these components, a good approximation is obtained for the transfer impedance Z_T of a braided wire screen:

$$Z_T \approx Z_{Th} + j\omega (M'_{12} - M''_{12}) \quad (28)$$

and the first approximation of the equivalent circuit is shown in Figure 20a.

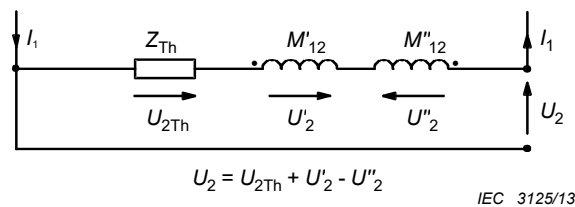


Figure 20a – Contributions to the transfer impedance

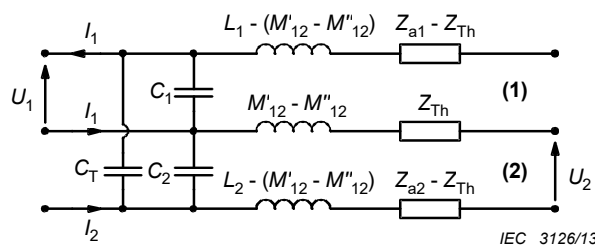


Figure 20b – Significant elements of circuits (1) and (2)

Figure 20 – Z_T equivalent circuits of a braided wire screen

A more complete equivalent circuit where the through capacitance C_T and surface impedances Z_a of the braided cable are incorporated is shown in Figure 20b. L_1 and L_2 are the (external) inductances of the outer and inner circuit.

Many attempts have been made to calculate the transfer impedance of a braided coaxial cable. Most of the literature ([2], [4], [5]) have concentrated on models of braided screens and calculation of direct leakage of the magnetic field induced by I_1 , and of M'_{12} . Satisfactory results have been achieved.

There exists very little literature ([6], [7]) on M''_{12} but the matter has been studied by experts of standardization bodies. Especially the calculation and stability of M''_{12} have been shown to be very problematic because of so many uncertain and unstable parameters, e.g. the resistance of the crossover points of the wires, which have an effect on the magnetic field distribution in the braid. Also the pressure of the jacket has an effect on the small space between the right hand lay and left-hand lay of the braided wires. Not to mention the number of wire ends per carrier and the braid angle and the tightness and optical coverage of the braid.

After understanding the magnetic coupling mechanisms, it is not surprising that the transfer impedances of braided wire screens vary considerably and are unstable for many braid and cable constructions whether or not they are optimized. It is also clear that a perforated tube cannot be used as a model for a braided screen.

It is clear that a loose highly optimized braid can have a very unstable Z_T during bending, twisting and/or pressing. An overbraided screen with a high filling factor or optical cover normally has a (pure) negative transfer impedance at high frequencies because of a large M''_{12} coupling through the mutual "space" between the left and right lays of the braid in comparison with a small leakage through the braid M'_{12} . Pressure on the jacket would improve the screening performance by diminishing the mutual "space" and decrease the Z_T .

The manufacture of a good stable optimized cable requires the control of braid parameters such as:

- braid angle, tension (and lubricant) of the strands;
- number of strand in a spindle;
- wire diameter;
- plating;
- pressure of the jacket on the braid in manufacturing.

8 Test possibilities

8.1 General

A number of test procedures are used to test cables for their screening properties, some of which will be found in IEC standards. Each procedure has benefits for some users which for historical reasons may not be widely appreciated. Table 2 summarizes the test procedures available, some of which will be discussed here, with special reference to their applicability to cables, cable assemblies and connectors.

8.2 Measuring the transfer impedance of coaxial cables

All tests listed in Table 2 can be used on coaxial cables, but if a single test is needed to cover frequencies above and below 100 MHz, tests 1, 4, 7, 9 and 10 can be dismissed. Of the others, those with 's' under 'grouping' (column 3) have better intrinsic isolation between measuring and injection circuits, while in those with 'o' under grouping the injection circuit is unscreened. The difference is the line interchange referred to in 4.5 above. One benefit of an unscreened injection line is that better access may be obtained for inspection of the cable under test, which may be useful if the sample is in any way flawed. The two test methods with unscreened injection lines are test 3 and test 8. The latter, with its wide frequency coverage is recommended for future testing.

8.3 Measuring the transfer impedance of cable assemblies

Even with a restricted frequency range, many of the tests listed in Table 2 are not suited to tests on cable assemblies. Tests 1, 4, and 6 are unsuitable because an electrically short sample may be needed to achieve the upper frequencies, while test 10 is still limited to frequencies above 100 MHz. Tests with screened injection wires (test 2 and test 5) are difficult to set up due to the varying cross section of the assembly, a difficulty which also applies to test 3. Such objections leave tests 7, 8 and 9. To set against its low (effective) upper frequency limit, with test 7 it is easy to distinguish between connector and cable contributions, so it is ideal in a diagnostic role. Test 9 works only above 30 MHz, which may be restrictive. Test 8 will require several measurements on each sample, as it is unreasonable to assume that a cable assembly has circular symmetry.

It is only fair to state that in any frequency domain test on cable assemblies where signal phase is not recorded, a test is only valid if the sample length is not varied (tests carried out on a sample of one length cannot be used to assess a sample of another length – whether it be longer or shorter). Of the transfer impedance tests being discussed, only test 7 can be used in this way.

Multi-conductor cable assemblies are more complex, because the 'core' cannot be considered to be coaxial. A test for such cable assemblies has not yet been addressed.

8.4 Measuring the transfer impedance of connectors

In principle, all the tests in Table 2 can be used on coaxial connectors.

As with tests on cable assemblies, there is much benefit to be gained from using a test with an unscreened injection circuit, though other tests will remain in the standard, because they have become accepted. If it is possible to distinguish the screening of a connector from that of the attached cable, this will considerably ease the test procedure.

Multi-pin connectors are far more numerous and varied than coaxial connectors. However, non-circular connectors cannot be tested by the means implied by the test procedures of Table 2, though by suitable variation test 7 and test 10 would become appropriate. This problem is under study.

NOTE These methods give only an outline for measurement of symmetrical multicore cables, multipin connectors and cable assemblies made with these components.

The problems to be addressed come from the fact that:

- a) a connector is electrically short, while the parameters of a cable are distributed, and it may be electrically long;
- b) multi-core cables rarely have circular symmetry. This applies both physically and to the signal paths on their conductors;
- c) most multi-pin connectors have no circular symmetry; nor are they equally spaced from other conductors, which might couple to them;
- d) economics will dictate that a cable assembly test should apply to other assemblies using the same components, even though of differing overall length.

8.5 Calculated maximum screening level

It is important to know the exact theoretical limitation of the test equipment. By knowing the limitations, it is possible to calculate the maximum measurable screening effectiveness. This should be calculated to check the strengths and weaknesses of the test setup or even to optimize the test setup.

The following test equipment specifications are required for the calculation:

- minimum input (noise floor);
- maximum input;

- amplification/attenuation;
- maximum output.

Figure 21 gives a comparison of the different signal levels in a generic test setup. The maximum screening is the difference between the maximum obtainable input signal to the DUT and the minimum detectable signal from the DUT, in this case 131 dB. The noise floor level (*NL*) of the measuring system must be low enough to allow the measurement. In this case, lower than –122 dBm. Measurements at the noise floor result in a maximum error of 3 dB. When measured 6 dB above the noise floor, the error is only about 1 dB.

The triaxial tube column is divided in two to show both the loss in the tube and the actual maximum screening.

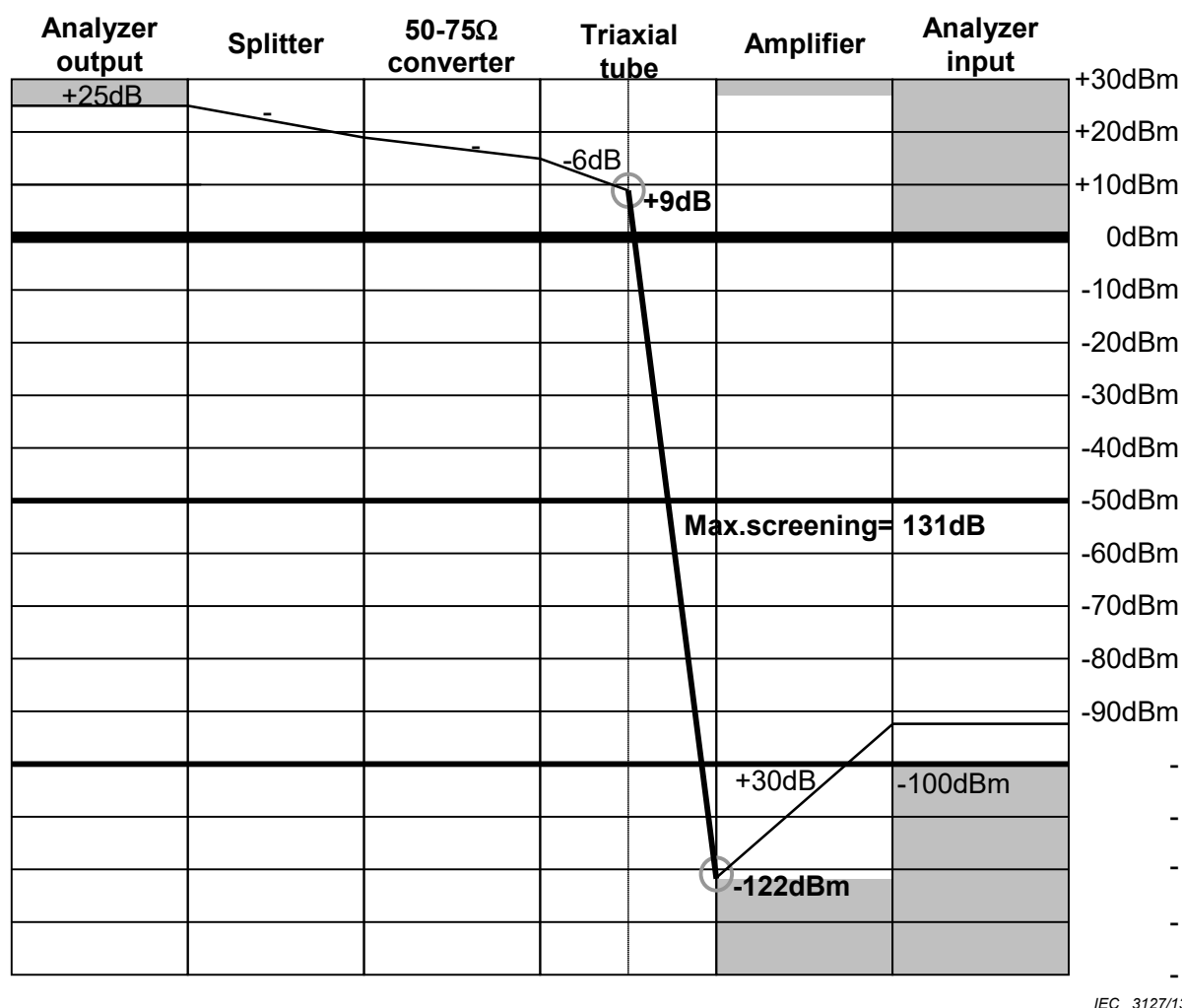


Figure 21 – Comparison of signal levels in a generic test setup

Taking into consideration the noise level of 1 Hz bandwidth at room temperature being –173 dBm, (increase $10 \times \log(\text{bandwidth})$ dB) and adding the noise figure of the amplifier, we get the theoretical noise level of the test setup. Assuming that the amplifier in the Figure 21 example has a noise figure of 11 dB, we can then calculate that the bandwidth (Δf) of the network analyzer shall be smaller than 10 kHz.

This can be expressed as a general formula:

$$NL = (-173 + F + 10 \times \log_{10} \Delta f) \quad (29)$$

where

NL is the noise floor level of receiving side of the measuring system in dBm;

F is the noise figure of the pre amplifier in dB;

Δf is the bandwidth of the receiver in Hz.

Table 2 – Screening effectiveness of cable test methods for surface transfer impedance Z_T

Short title	Reference	Grouping (see Note 1)	Frequency range		Injection N or F (see Note 2)	Advantages or shortcomings
			Possible	Actually used		
1 IEC triaxial	IEC 62153-4-3	kf s	d.c. to 50 MHz	10 kHz to 30 MHz	F	Rigid test rig or flexible (milked on braid)
2 Terminated triaxial (Simons)	Figure A5 of IEC 60096-1:1986 [32]	m s	10 kHz to 1 GHz	100 kHz to 500 MHz	N F	Flexible test jig relies on ferrites
3 Braid injection (Fowler)	[9]	m o	d.c. to 500 MHz	10 kHz to 500 MHz	N F	Flexible test needs good screening on measuring system
4 Quadaxial	[10]	m s	100 kHz to 50 MHz	100 kHz to 1 GHz	N	Deep resonances make use above 50 MHz theoretically impossible. The test has been used for assessing screening at frequencies up to 1 GHz
5 Matched T triaxial (Staegar)	IEC 60169-1-3 [11]	m s	1 kHz to 12 GHz	100 MHz to 10 GHz 10 kHz to 100 MHz	N F	Rigid test jig needs good screening
6 ERA triaxial (Smithers)	[12]	kf s	d.c. to 400 MHz	10 kHz to 300 MHz	F	Very short CUT requires amplifier or phase locked loop
7 Line injection (time domain)	IEC 60096-4-1 [33] [13]	m o	d.c. to 100 MHz	1 kHz to 80 MHz (note 3)	N F	Very easy to use. Needs good screening in measuring amplifier
8 Line injection (frequency domain)	IEC 62153-4-6 [14]	m o	d.c. to 20 GHz	10 kHz to 3 GHz	N F	Flexible and cheap measuring set-up, equipment needs to be well shielded
9 Open screening attenuation test method (absorbing clamp)	IEC 62153-4-5	m o	30 MHz to 2.5 GHz	30 MHz to 1 GHz 300 MHz to 2,5 GHz	N F	Poor sensitivity. Measuring of a_s is dependent on the surroundings
10 Reverberation chamber method	IEC 61726 [15]	kn kf	0,1 GHz →	0,3 GHz to 40 GHz	N & F	Flexible in use, but a complex and expensive computer controller with sophisticated test software needed
11 Shielded screening attenuation test method	IEC 62153-4-4 [16] (note 4)	m s	d.c. to 5 GHz	10 kHz to 3 GHz	F	High-sensitivity measurements can be made without a screened room. Transfer impedance and screening attenuation can be measured with one set-up
12 Open multipin connector screening test method	[17]	o	d.c. to 1 GHz	10 kHz to 700 MHz	N	Low cost and flexible

Short title	Reference	Grouping (see Note 1)	Frequency range		Injection N or F (see Note 2)	Advantages or shortcomings
			Possible	Actually used		
13 Coupling attenuation measurements of balanced cables, cable-assemblies, connecting hardware 13.1 Current clamp injection method 13.2 Shielded triaxial test method 13.3 Absorbing clamp method	IEC 62153-4-2 IEC 62153-4-9 [22] IEC 62153-4-5 IEC 62153-4-11 IEC 62153-4-12 IEC 62153-4-13 IEC 62153-4-7		50 MHz to 1 GHz d.c. to 3 GHz 50 MHz to 2,5 GHz	50 MHz to 1 GHz d.c. to 1 GHz 50 MHz to 2,5 GHz		High sensitivity but a screened room is recommended High-sensitivity measurements can be made without a screened room Poor sensitivity
14 Shielded screening attenuation, test method for measuring the transfer impedance Z_T and the screening attenuation as of RF connectors up to and above 3 GHz; tube in tube method		m s	d.c. to 20 GHz	d.c. to 3 GHz		High-sensitivity measurements can be made without a screened room Transfer impedance and Screening attenuation with one test set-up
15 Shielded screening attenuation test method for measuring the screening effectiveness of feedtroughs and electromagnetic gaskets – double coaxial method	IEC 62153-4-10	m s	d.c. to 4 GHz	d.c. to 3 GHz		High-sensitivity measurements can be made without a screened room

NOTE 1 Grouping by condition of 'primary circuit':
kn = short circuit at near end;
kf = short circuit at far end;
m = matched with characteristic impedance;
o = open on unscreened;
s = screened or shielded.

NOTE 2 N denotes near end feeding of primary relative to secondary circuit. F denotes far end feeding of primary relative to secondary circuit.

NOTE 3 Effective frequencies tested. Actually pulse with $T_R = 3,5$ ns and duration up to 160 μ s.

NOTE 4 Secondary circuit near end short circuited.

9 Comparison of the frequency response of different triaxial test set-ups to measure the transfer impedance of cable screens

9.1 General

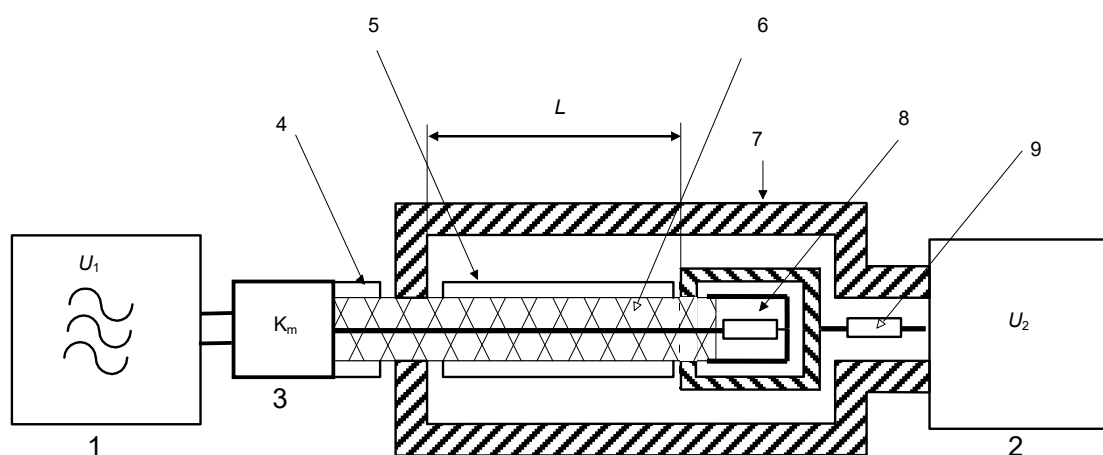
Different triaxial test set-ups for the measurement of the transfer impedance exist as described in EN 50289-1-6 and the IEC 62153-4 series. All of them are based on the same principle but are using different load conditions. In one method for example the cable under test is matched, while in the other the cable is short circuited at the far end. Furthermore, generator and receiver may be interchanged in the different set-ups. The following investigation analyses the frequency response of the different set-ups and their influence on the cut-off frequency up to which the transfer impedance could be measured.

9.2 Physical basics

9.2.1 Triaxial set-up

9.2.1.1 General

The triaxial set-up is of the “triple coaxial” form, see Figure 22 and Figure 23. A short length of the screen under test forms both, the inner conductor of the outer system and at the same time the outer conductor of the inner system. The coupling between the two coaxial systems is caused by the transfer impedance and the capacitive coupling admittance of the screen. The matching circuit, load resistor and series resistor are used to change the load conditions of the set-up. Also the generator and receiver may be interchanged between the different methods.

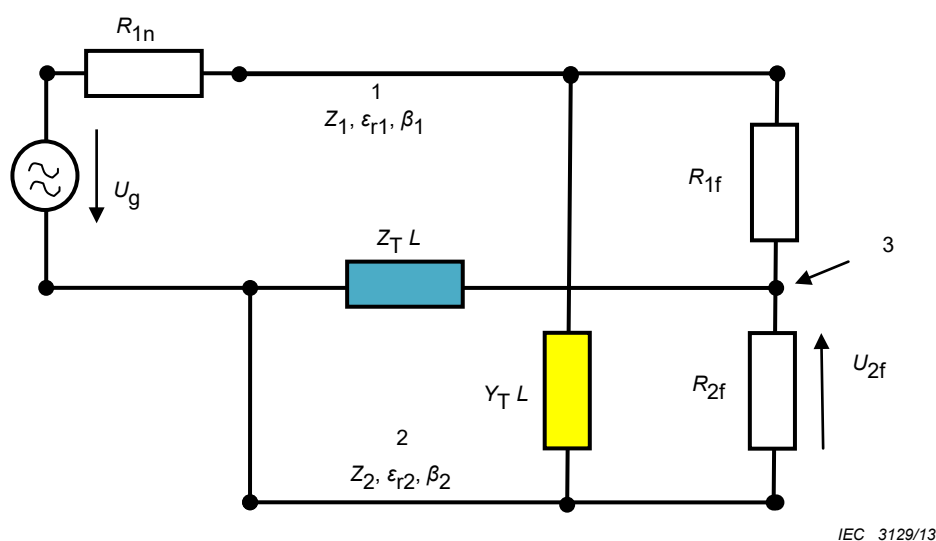


IEC 3128/13

Key

- | | |
|---|------------------------|
| 1 Signal generator | 6 Cable screen |
| 2 Calibrated receiver or network analyzer | 7 Tube |
| 3 Matching circuit | 8 Terminating resistor |
| 4 Cable under test | 9 Series resistor |
| 5 Cable sheath | |

Figure 22 – Triaxial set-up for the measurement of the transfer impedance Z_T



Key

- 1 inner circuit, cable
- 2 outer circuit, tube
- 3 screen
- $Z_{1,2}$ characteristic impedance of the inner circuit, cable, respectively outer circuit, tube
- $\varepsilon_{1,2}$ dielectric permittivity of the inner circuit, cable, respectively outer circuit, tube
- $\beta_{1,2}$ phase constant of the inner circuit, cable, respectively outer circuit, tube
- L coupling length
- Z_T transfer impedance
- Y_T capacitive coupling admittance
- R_{1n} load resistance at the near end of the inner circuit, cable. Equal to the output impedance of the generator respectively input impedance of the receiver including an eventually used feeding resistor
- R_{1f} load resistance at the far end of the inner circuit, cable. Depending on the used method either equal to the characteristic impedance of the cable or a short circuit.
- R_{2f} load resistance at the far end of the outer circuit, tube. Equal to the output impedance of the generator respectively input impedance of the receiver including an eventually used feeding resistor
- U_g EMF of the generator
- U_{2f} voltage at the far end of the outer circuit

Figure 23 – Equivalent circuit of the triaxial set-up

9.2.1.2 Load conditions of the different set-ups

EN 50289-1-6 is using a method, where the cable under test and the far end of the secondary circuit are matched. The signal is fed to the cable under test and the disturbing voltage is measured at the far end of the outer circuit. A simplified method is to neglect the matching resistor at the far end of the outer circuit, which results in a higher dynamic range.

IEC 61196-1 describes two methods:

Method 1: Feeding through a resistance, where the signal is fed via a resistance into the outer circuit and the disturbing voltage is measured at the far end of the cable under test.

Method 2: Direct feeding, where the signal power is fed directly into the outer circuit and the disturbing voltage is measured at the far end of the cable under test.

With the revision of IEC 61196-1, the standard IEC 62153-4-3 has been published which also describes several methods:

Method A “Matched-Short” is equal to EN 50289-1-6.

Method B “Short-Short” is the double short circuited method, where the load resistance of the cable is replaced by a short circuit, thus having two short circuits in the set-up. One is at the near end of the outer circuit (between the cable screen and the tube) and the other is at the far end of the cable. The advantage of this method is the simplification of the sample preparation. A short circuit is easier to make than to solder a resistor, especially if the sample is a multi-conductor cable. Furthermore, the measurement sensitivity is improved. Compared to the “matched-short” method, the dynamic range is improved by about 16 dB. In the “milked on braid” method, an additional braid, the measuring braid, is pulled over the cable sheath instead of using the measuring tube. The advantage is that the sample could be bent under test, however the preparation is more laborious than with the measuring tube.

The load conditions of the different methods are given in Table 3. The impedance of the outer circuit, Z_2 is varying with the diameter of the screen under test. Using the measuring tube Z_2 is in general higher, and in the “milked on braid” method Z_2 is lower, than the input impedance of the receiver.

Table 3 – Load conditions of the different set-ups

Method	Generator	Receiver	R_{1n}/Z_1	R_{1f}/Z_1	Z_2/R_{2f}
EN 50289-1-6					
Standard	IC	OC	1	1	0,71
simplified	IC	OC	1	1	1...5 depending on the tube diameter
IEC 61196-1					
Method 1: feeding through a resistance	OC	IC	1	1	0,71
Method 2: direct feeding	OC	IC	1	1	1...5 depending on the tube diameter
IEC 62153-4-3 Double short circuit methods					
With tube	OC	IC	1*	0	1...5 depending on tube diameter
With milked on braid	IC	OC	1*	0	0,1...0,4 depending on screen and sheath diameter of the cable
IC: inner circuit (cable under test)					
OC: outer circuit (tube)					
* only if the cable impedance is equal to the generator impedance. For other cable impedances, the value may vary, e.g. 0,67 for cables with an impedance of 75 Ω .					

9.2.2 Coupling equations

The equations for the coupling between the inner circuit and outer circuit for any load conditions are described in [18] and [19]. By taking into account the short circuit at the near end of the outer circuit (between the cable screen and the measuring tube), neglecting the attenuation of the disturbing and disturbed line, assuming non ferromagnetic materials and introducing further variables, the following equations are defined.

$$\frac{u_{2f}}{u_q} = \frac{L}{R_{lf} + R_{ln}} \cdot [Z_T \cdot g + Z_F \cdot h] \quad (30)$$

$$g = \frac{1}{N} \cdot \frac{1}{1-n^2} \cdot \frac{j}{x} \cdot \{ r \cdot [\cos x - \cos nx] - j \cdot n \cdot \sin nx + j \cdot \sin x \} \quad (31)$$

$$h = \frac{1}{N} \cdot \frac{1}{1-n^2} \cdot \frac{j}{x} \cdot \{ n \cdot r \cdot [\cos x - \cos nx] - j \cdot \sin nx + j \cdot n \cdot \sin x \} \quad (32)$$

$$N = \left\{ \cos x + \frac{j \cdot \sin x}{r + w} \cdot [1 + r \cdot w] \right\} \cdot \{ \cos nx + j \cdot v \cdot \sin nx \} \quad (33)$$

$$x = \beta_1 \cdot L = 2\pi \cdot \frac{L}{\lambda_1} \quad (34)$$

$$n = \frac{\beta_2}{\beta_1} = \frac{\lambda_1}{\lambda_2} = \sqrt{\frac{\epsilon_{r2}}{\epsilon_{r1}}} \quad (35)$$

$$r = \frac{R_{lf}}{Z_1} \quad (36)$$

$$v = \frac{Z_2}{R_{2f}} \quad (37)$$

$$w = \frac{R_{ln}}{Z_1} \quad (38)$$

where

$Z_{1,2}$	is the characteristic impedance of the inner circuit (cable) respectively outer circuit (tube);
$\epsilon_{1,2}$	is the dielectric permittivity of the inner circuit (cable) respectively outer circuit (tube);
$\beta_{1,2}$	is the phase constant of the inner circuit (cable) respectively outer circuit (tube);
$\lambda_{1,2}$	is the wave length in the inner circuit (cable) respectively outer circuit (tube);
L	is the coupling length;
Z_T	is the transfer impedance;
Y_T	is the capacitive coupling admittance;
$R_{1,n}$	is the load resistance at the near end of the inner circuit (cable). Equal to the output impedance of the generator respectively input impedance of the receiver including an eventually used feeding resistor;
$R_{1,f}$	load resistance at the far end of the inner circuit (cable). Depending on the used method either equal to the characteristic impedance of the cable or a short circuit.

The factors g and h (see Equations (31) and (32)) describe the frequency response of the test set-up. At low frequencies, when $\lambda \gg L$, the factors g and h are equal to 1. However, with increasing frequency, the factors g and h start to oscillate and thus also the measurement results. The maximum frequency to which the transfer impedance could be measured without oscillations, caused by the set-up, is defined as the 3 dB deviation from the linear interpolation of the measurement results. Or in other words, the maximum frequency is reached when the factor g respectively h becomes $>\sqrt{2}$ respectively $<1/\sqrt{2}$.

9.3 Simulations

9.3.1 General

For the following investigations, simulations have been chosen rather than a pure mathematical solution because they are easier to grasp and clearly illustrate the differences in the set-ups given in Table 4. In general, the capacitive coupling can be neglected compared to the magnetic coupling ($Z_F \ll Z_T$). i.e. the cut-off frequency is mainly determined by the frequency behaviour of the factor g . Thus the following simulations are limited to the factor g .

Due to the reciprocity of the materials, it is possible to interchange the generator and receiver without changing the results. Thus the standard EN 50289-1-6 method gives the same results as IEC 61196-1, method 1: “feeding through a resistance” and the simplified EN 50289-1-6 method gives the same results as IEC 61196-1, method 2: “direct feeding”.

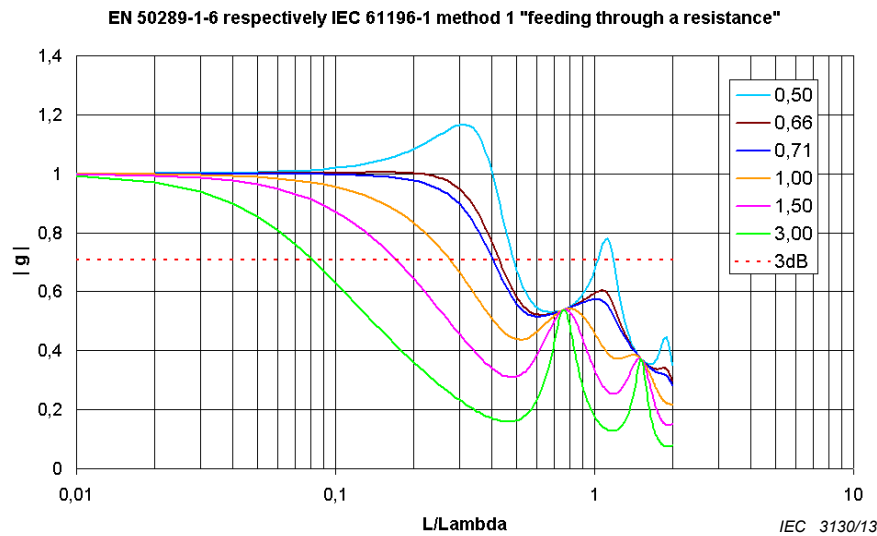
Table 4 – Parameters of the different set-ups

Method	$w=R_{1n}/Z_1$	$r=R_{1f}/Z_1$	$v=Z_2/R_{2f}$	$n=\sqrt{\varepsilon_{r2}}/\sqrt{\varepsilon_{r1}}$
EN 50289-1-6, IEC 62153-4-3 method A				
Standard	1	1	0,71	0,66 (0,45)...0,91
Simplified	1	1	1...5 depending on the tube diameter	
IEC 61196-1				
Method 1: feeding through a resistance	1	1	0,71	0,66 (0,45)...0,91
Method 2: direct feeding	1	1	1...5 depending on the tube diameter	
IEC 62153-4-3 Double short circuit methods				
With tube	1 ^a	0	1...5 depending on tube diameter	0,66 (0,45)...0,91
With milked on braid	1 ^a	0	0.1...0,4 depending on screen and sheath diameter of the cable	1,02...2,0
^a only if the cable impedance is equal to the generator impedance. For other cable impedances, the value may vary, e.g. 0,67 for cables with an impedance of 75 Ω.				

In the tube methods, the factor n is given by the dielectric permittivity of the cable (inner circuit) as the dielectric permittivity of the outer circuit is nearly independent on the sheath material and can be assumed to be 1. However, in the “milked on braid method”, the factor n is dependent on both the dielectric permittivity of the cable insulation and the sheath, as the “measuring braid” is directly put on the sheath of the sample. The values for the factor n are given for typical insulation materials (PE, foam PE, PTFE ...). The values in brackets are given for an insulation material of PVC, which may be used in multi-pair/conductor cables. For the “milked on braid” method, typical combinations of insulation and sheath materials (PE/PVC, PE/LSZH, PTFE/FEP...) are taken into account, resulting in a value $n > 1$.

9.3.2 Simulation of the standard and simplified methods according to EN 50289-1-6, IEC 61196-1 (method 1 and 2) and IEC 62153-4-3 (method A)

In EN 50289-1-6, IEC 61196-1 method 1: “feeding through a resistance” and IEC 62153-4-3 method A: “Matched-Short”, the factor $v=Z_2/R_{2f}$ is specified at $1/\sqrt{2}$. The following simulations show that this factor is a good compromise with respect to the maximum frequency to which the transfer impedance could be measured.

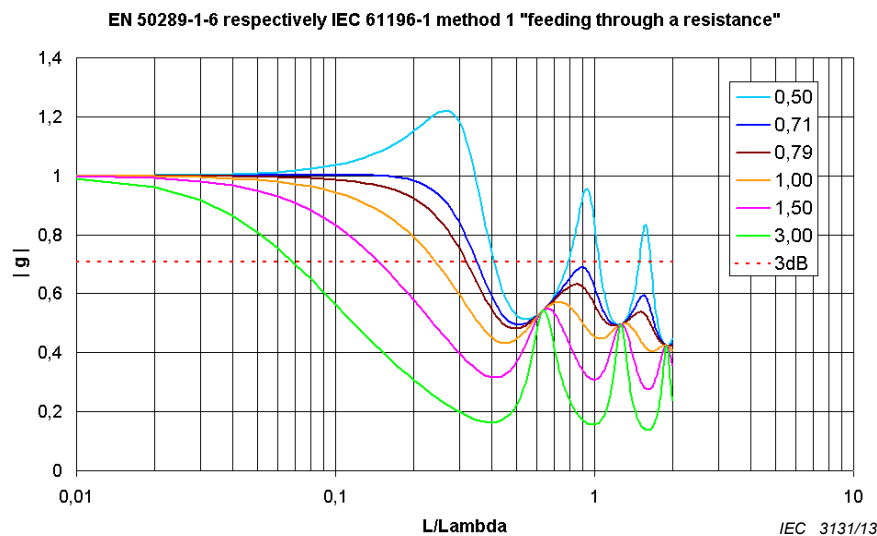


Key

Coloured lines correspond to indicated factors of $v=Z_2/R_{2f}$.

Simulation parameters		
ϵ_{r1}	ϵ_{r2}	n
2,3 (solid PE)	1,0	0,569

Figure 24 – Simulation of the frequency response for g

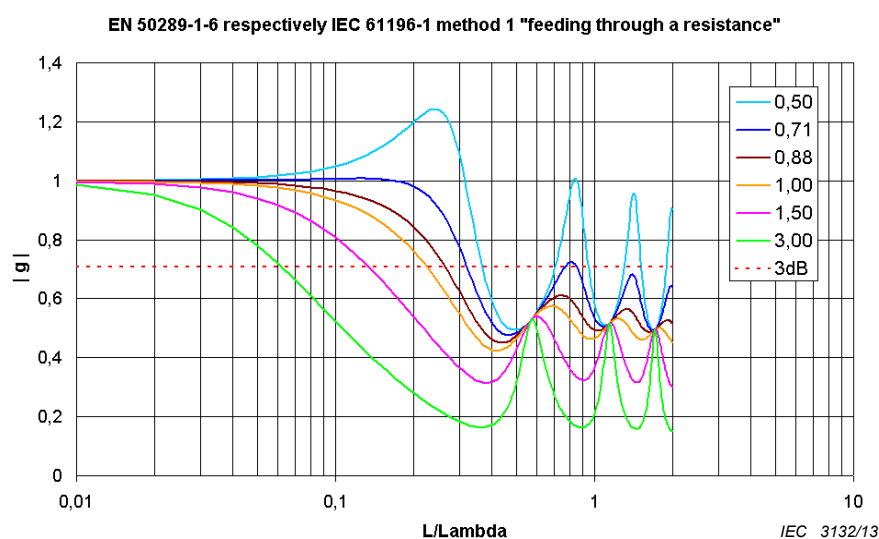


Key

Coloured lines correspond to indicated factors of $v=Z_2/R_{2f}$.

Simulation parameters		
ϵ_{r1}	ϵ_{r2}	n
1,6 (foam PE)	1,0	0,791

Figure 25 – Simulation of the frequency response for g

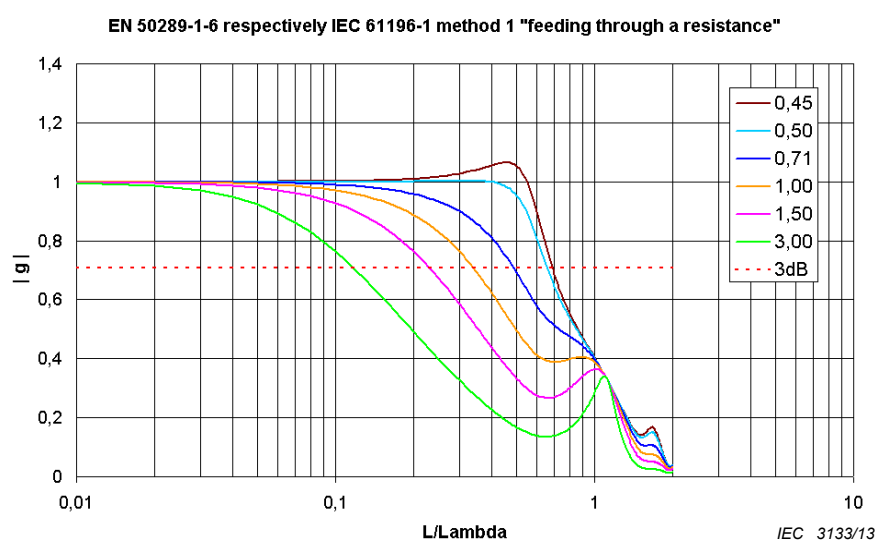


Key

Coloured lines correspond to indicated factors of $\nu = Z_2/R_{2f}$.

Simulation parameters		
ϵ_{r1}	ϵ_{r2}	n
1,3 (foam PE)	1,0	0,877

Figure 26 – Simulation of the frequency response for g



Key

Coloured lines correspond to indicated factors of $\nu = Z_2/R_{2f}$.

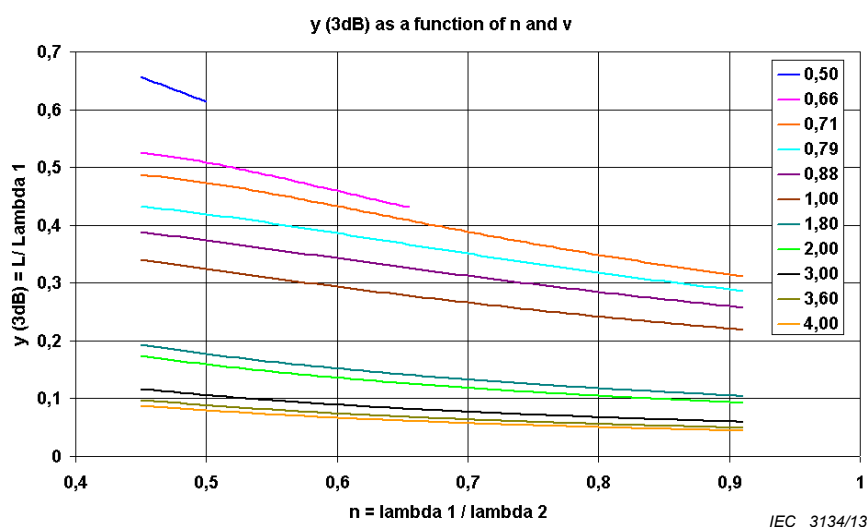
Simulation parameters		
ϵ_{r1}	ϵ_{r2}	n
5 (PVC)	1,0	0,447

Figure 27 – Simulation of the frequency response for g

The highest frequencies (respectively shortest wavelengths) are obtained if the factor $\nu = 1/\sqrt{2}$ respectively $\nu = n$, whichever is smaller. In Figure 24 and Figure 27, the highest frequency is obtained for $\nu = n$ ($=0,659$ respectively $0,447$). But in Figure 25 and Figure 26, the highest frequency is obtained for $\nu = 1/\sqrt{2} = 0,71$. Below that value, the factor g overshoots, i.e. becomes higher than one. Above that value, the cut-off frequency is decreasing.

Figure 28 gives the calculated, by iteration, 3 dB cut-off wavelength (L/λ_1) at which the factor $|g|$ becomes $1/\sqrt{2}$. The graph is given as a function of the factor $n = \sqrt{\epsilon_{r2}}/\sqrt{\epsilon_{r1}}$ and for different factors $v = Z_2/R_{2f}$. The curves show a linear behaviour and could be interpolated by straight line.

This has been done in Figure 29 for $v=1/\sqrt{2}$, $v=1$, $v=1,8$ and $v=3,6$. The factor $v=1/\sqrt{2}$ corresponds to the set-up according to EN 50289-1-6, IEC 61196-1 method 1 “feeding through a resistance” and IEC 62153-4-3 method A “Matched-Short”. The other values of the factor v correspond to the simplified set-up, i.e. direct feeding. For common diameters of the measuring tube (around 40 mm) and common cable screen diameter (2 mm to 9 mm), the impedance in the outer circuit is 90 Ω to 180 Ω and $v=1,8\ldots 3,6$.

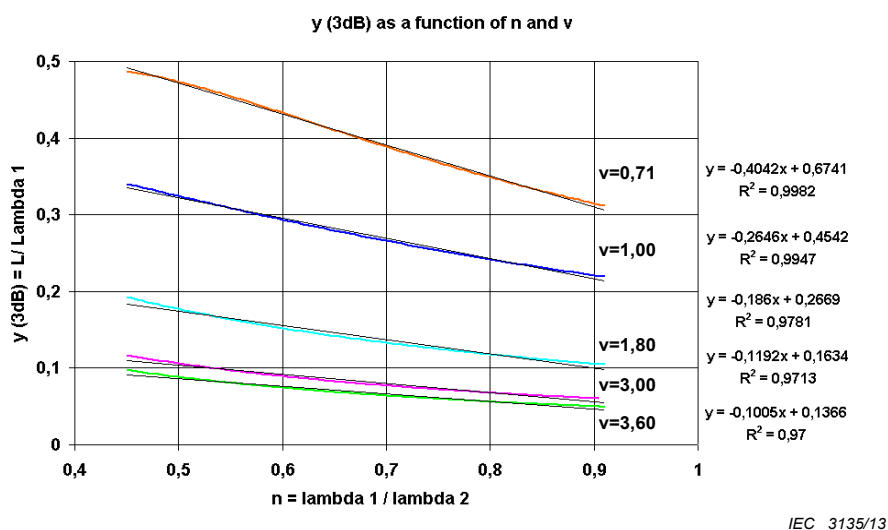


Key

Coloured lines correspond to indicated factors of $v = Z_2/R_{2f}$.

Figure 28 – Simulation of the 3 dB cut off wavelength (L/λ_1)

The graphs for $v=0,5$ and $v=0,66$ are only given for n up to 0,5 respectively 0,66 because otherwise the factor g overshoots as described above.



Key

Coloured lines correspond to indicated factors of $v = Z_2/R_{2f}$.

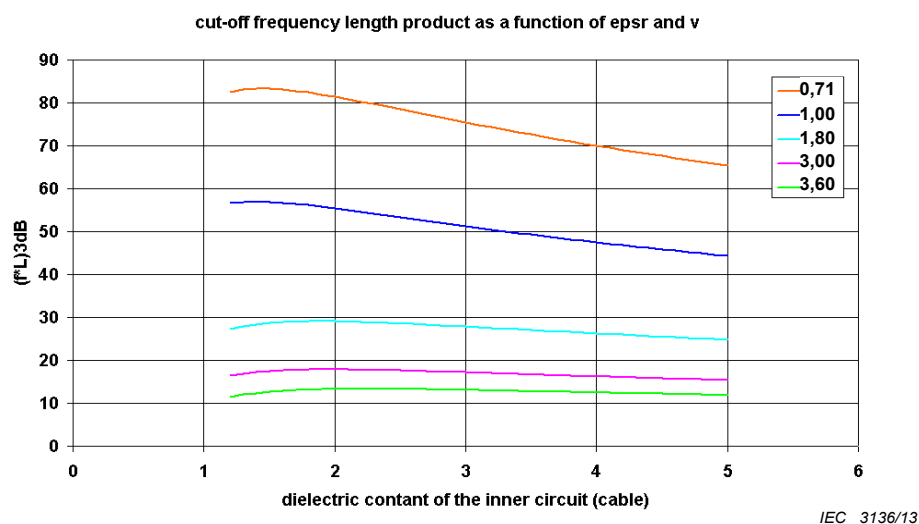
Figure 29 – Interpolation of the simulated 3 dB cut off wavelength (L/λ_1)

The linear interpolation equation is used to derive an equation to calculate the cut-off frequency length product up to which the transfer impedance could be measured in a given triaxial test set-up.

Table 5 – Cut-off frequency length product

Triaxial test set-up	ν	Cut-off equation
EN 50289-1-6 IEC 61196-1 method 1 “feeding through a resistance” IEC 62153-4-3 method A “matched-short”	$\nu=1/\sqrt{2}$	$(f \cdot L)_{3\text{ dB}} \approx \left[\frac{200}{\sqrt{\epsilon_{r1}}} - \frac{120}{\epsilon_{r1}} \right] \cdot \text{MHz} \cdot \text{m}$
Simplified EN 50289-1-6 IEC 61196-1 method 2 “direct feeding”	$\nu=1$	$(f \cdot L)_{3\text{ dB}} \approx \left[\frac{135}{\sqrt{\epsilon_{r1}}} - \frac{80}{\epsilon_{r1}} \right] \cdot \text{MHz} \cdot \text{m}$
	$\nu=1,8$	$(f \cdot L)_{3\text{ dB}} \approx \left[\frac{80}{\sqrt{\epsilon_{r1}}} - \frac{55}{\epsilon_{r1}} \right] \cdot \text{MHz} \cdot \text{m}$
	$\nu=3$	$(f \cdot L)_{3\text{ dB}} \approx \left[\frac{50}{\sqrt{\epsilon_{r1}}} - \frac{35}{\epsilon_{r1}} \right] \cdot \text{MHz} \cdot \text{m}$
	$\nu=3,6$	$(f \cdot L)_{3\text{ dB}} \approx \left[\frac{40}{\sqrt{\epsilon_{r1}}} - \frac{30}{\epsilon_{r1}} \right] \cdot \text{MHz} \cdot \text{m}$

The equations given in Table 5 are drawn in the graphs of Figure 30.



Key

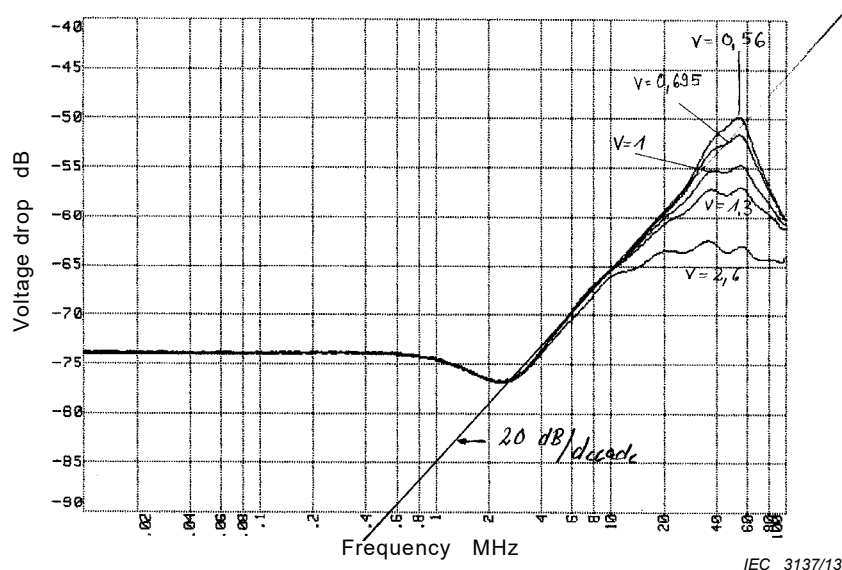
Coloured lines correspond to indicated factors of $\nu=Z_2/R_{2f}$.

Figure 30 – 3 dB cut-off frequency length product as a function of the dielectric permittivity of the inner circuit (cable)

For example, if a cable with a PE insulation – dielectric permittivity of, $\epsilon_{r1} = 2,3$, and a screen diameter of 3,5 mm is measured in a triaxial set-up according to EN 50289-1-6 or IEC 61196-1 method 1 “feeding through a resistance” with $\nu=0,71$, then the cut-off frequency length product is about 80 MHz.m. Therefore for a coupling length of 0,5 m, the maximum frequency to which the transfer impedance could be measured is around 160 MHz.

If the same cable is measured in a triaxial set-up according to IEC 61196-1 method 2 “direct feeding” or the simplified set-up according to EN 50289-1-6 where $v=3$, then the cut-off frequency length product is about 18 MHz·m. For a coupling length of 0,5 m, the maximum frequency to which the transfer impedance could be measured is around 36 MHz.

Figure 31 and Figure 32 show the measurement results of the normalised voltage drop – i.e. the attenuation caused by the series resistor has been taken into account – in the triaxial set-up for different factors of v . Both figures show the results of the same screen design, however one with a solid PE insulation ($\epsilon_{r1}=2,3$), the other with a foam PE insulation ($\epsilon_{r1}=1,6$). The measurement results confirm the simulations. From the equations given in Table 5 one obtains cut-off frequency length products for $v=3$ of about 18 MHz·m and for $v=1$ of about 55 MHz·m for both the solid PE and the foam PE. This is also found from the measurement results.



IEC 3137/13

Key

Indicated lines correspond to factors of $v=Z_2/R_{2f}$

Measurement set-up parameters				
ϵ_{r1}	ϵ_{r2}	n	Z_2	L
2,3 (PE)	1,0	0,659	130 Ω	1 m

Figure 31 – Measurement result of the normalised voltage drop of a single braid screen on a solid PE dielectric in the triaxial set-up