IEC/TR 62153-4-1

Edition 1.0 2007-11

TECHNICAL REPORT

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

IEC/TR 62153-4-1:2007(E)

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INTERNATIONAL ELECTROTECHNICAL COMMISSION

PRICE CODE XA

ICS 33.100; 33.120.10

ISBN 2-8318-9364-X

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METALLIC COMMUNICATION CABLE TEST METHODS –

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

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

This publication cancels and replaces IEC/TR 61917, published in 1998.

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

Enquiry draft	Report on voting
46/199/DTR	46/253/RVC

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

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

The committee has decided that the contents of this publication will remain unchanged until the maintenance result 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

- reconfirmed,
- withdrawn,
- replaced by a revised edition, or
- amended.

A bilingual version of this publication may be issued at a later date.

INTRODUCTION

Screening 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.

IEC/TR 62153-4-1 provides a brief introduction to basic concepts and terms trying to reveal the common features of apparently different test methods. It should assist in correct interpretation of test data, and in the better understanding of screening (or shielding) and related specifications and standards.

METALLIC COMMUNICATION CABLE TEST METHODS –

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

1 Scope

IEC/TR 62153-4-1, which is a technical report, gives a brief introduction to basic concepts and terms that reveal the common features of various test methods.

2 Normative references

The following referenced documents are indispensable for the application 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 60096-4-1:1990, *Radio-frequency cables – Part 4: Specification for superscreened cables – Section 1: General requirements and test methods*

IEC 60169-1-3:1988, *Radio frequency connectors – Part 1: General requirements and measuring methods – Section 3: Electrical tests and measuring procedures – Screening effectiveness*

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

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

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

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

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

IEC 62153-4-7, *Metallic communication cables 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*[1]

EN 50289-1-6, *Communication cables – Specification for test methods – Electrical test methods – Electromagnetic performance*

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[1] To be published.

3 List of symbols

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
c	velocity of light
C_T	through capacitance of the braided cable
CUT	cable or component under test
E	EMF
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
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 ½ Φ''_{12}) in the braid

$$M'_{12} = \frac{\Phi'_{12}}{j\omega I_1} \quad \text{and} \quad M''_{12} = \frac{1}{2} \cdot \frac{\Phi''_{12}}{j\omega I_1}$$

n	near end
P_1	sending power
P_{2f}	far end measured power
P_{2n}	near end measured power
T	coupling transfer function
T_f	far end transfer function

T_n near end transfer function

$T_{n,f} = T_{\substack{n\\f}}$

U'_2 the disturbing voltage induced by Φ'_{12}

U''_{rh} the disturbing voltage induced by ½ Φ''_{12} of the right hand lay contribution

U''_{lh} the disturbing voltage induced by ½ Φ''_{12} of the left hand lay contribution

U''_2 is equal to U''_{rh} and U''_{lh} (= the disturbing voltage induced by ½ Φ''_{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

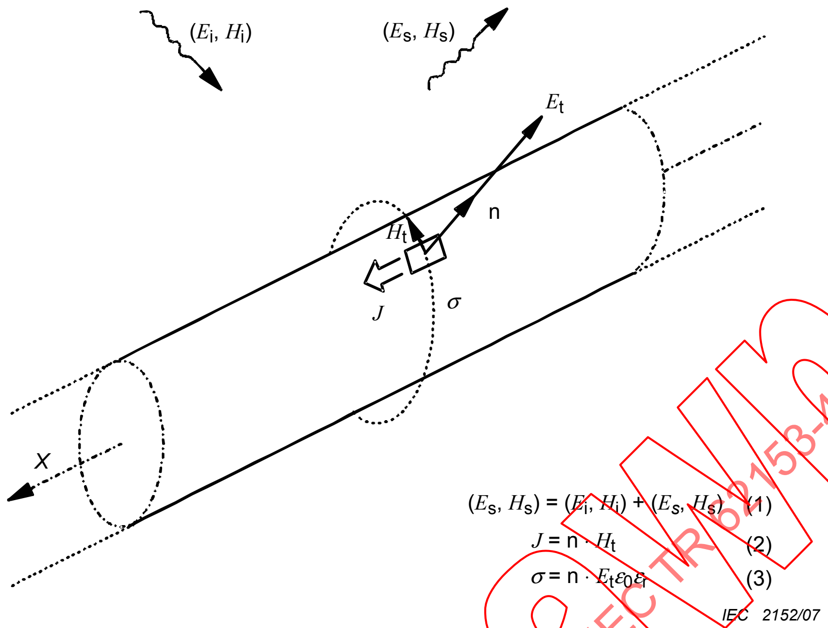
Z_{te} effective transfer impedance (= max $| Z_f \pm Z_t |$)

Z_{ten} normalized effective transfer impedance of a cable (Z_1 = 150 Ω 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 (E_t, H_t, in Figure 1). The total field at the screen's surface may be considered as the source of the coupling: the electric field penetrates through apertures by *electric* or capacitive coupling; also magnetic fields penetrate through apertures by *inductive* or magnetic coupling.

Additionally, the induced current in the screen results in *conductive* or resistive coupling.

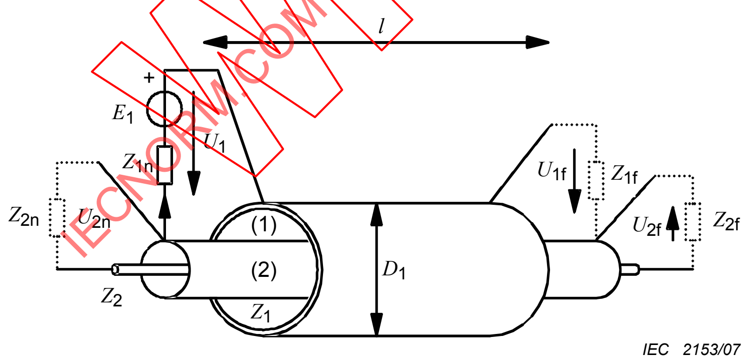


Key

n unit vector normal to surface

Figure 1 – Incident (i), scattered (s) and resulting total electromagnetic fields (E_t, H_t) with induced surface current- and charge- densities J (A/m) and σ (C/m²)

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 (E_t, H_t) or to the surface current- and charge- densities (J and σ). Consequently, the coupling can be simulated into the cable by reproducing through any means the surface currents and charges on the screen. Because a cable of small diameter is assumed, higher modes can be neglected and an additional coaxial conductor can be used as the injection structure, as shown in Figure 2.



Concept of a triaxial set-up

1) Outer circuit, formed by injection cylinder and screen, characteristic impedance Z_1,

2) Inner circuit, formed by a screen, and centre conductor, characteristic impedance Z_2; screening at the ends not shown.

Conditions Z_{1f}, Z_{2n}, Z_{2f} and λ are observed in Figure 3a and Figure 3b.

NOTE 1 $D_1 << l$.

NOTE 2 Both ends of circuit (2) must be well screened.

Figure 2 – Defining and measuring screening parameters – Triaxial set-up

5 Intrinsic screening parameters of short cables

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 $l << \lambda$ which will always apply at low frequencies, the intrinsic screening parameters are defined and can be measured as follows:

5.1 Surface transfer impedance, Z_T

As shown in Figure 2 and Figure 3a (where Z_{1f} and Z_{2f} are zero):

$$Z_T = U_2 / I_1 \cdot l) \qquad (\Omega/\text{m}) \qquad (4)$$

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 3a the voltmeter and short-circuit can be interchanged.

5.2 Capacitive coupling admittance, Y_c

As shown in Figure 2 and Figure 3b (where Z_{1f} and Z_{2f} are open circuit):

$$Y_C = j\omega C_T = I_2 /(U_1 \cdot l) \qquad (\text{mho/m}) \qquad (5)$$

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

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

a) The *normalized through elastance* K_T derived from C_T is independent of the size of the outer coaxial circuit, but it depends on its permittivity:

$$K_T = C_T /(C_1 C_2) \qquad (\text{m/F}) \qquad K_T \sim 1/(\varepsilon_{r1} + \varepsilon_{r2}) \qquad (6)\ (7)$$

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

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

$$Z_F = Z_1 Z_2 Y_C = Z_1 Z_2 j\omega C_T \quad (\Omega/\text{m}) \quad Z_F \sim \sqrt{(\varepsilon_{r1} \cdot \varepsilon_{r2})} /(\varepsilon_{r1} + \varepsilon_{r2}) \qquad (8)\ (9)$$

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

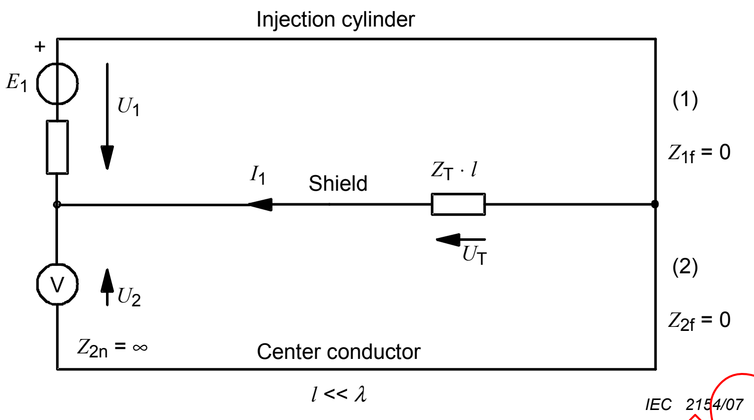


Figure 3a – Equivalent circuit for the definition and possible testing of Z_T

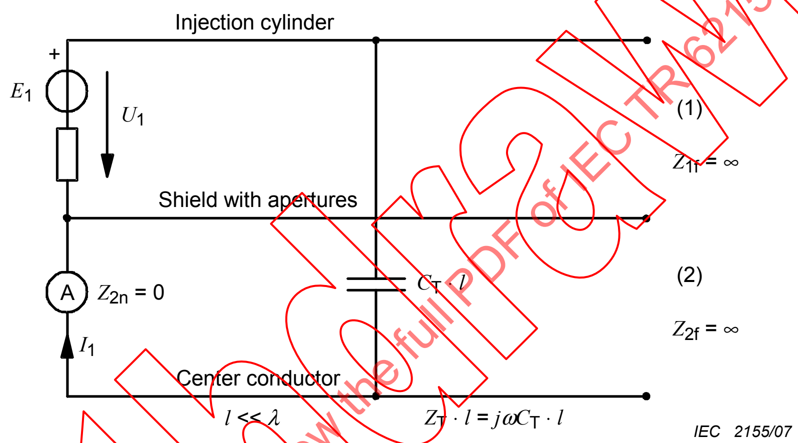
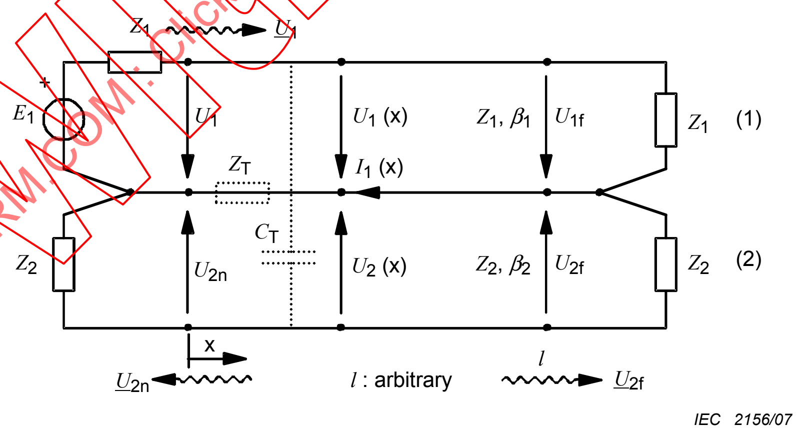


Figure 3b – Equivalent circuit for the definition and possible testing of $Y_c = j\ \omega C_T$



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

Figure 3c – Definition of electrical quantities in a set-up that is matched at all ends

Figure 3 – Defining and measuring screen parameters – Equivalent circuits

5.3 Injecting with arbitrary cross-sections

A coaxial outer circuit 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 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.

5.4 Reciprocity and symmetry

Assuming linear shield materials, the measured Z_T and Z_F values will not change when interchanging injection (1) and measuring (2) circuits. 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.

5.5 Arbitrary load conditions

When the circuit ends of Figure 3a and Figure 3b are not ideally short or open circuit, Z_T and Z_F will act simultaneously. The superposition is noticeable in the low-frequency coupling of the matched circuits (Figure 3c 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 U_1, U_{2n}, U_{2f}, see Figure 3c,

– representation of the coupling, using the normalized wave amplitudes $U/\sqrt{Z}\left[\sqrt{Watt}\right]$, instead of voltage waves, i.e. the *coupling transfer function*, in the following denoted by "*coupling function*", will be defined as

$$T_n = \frac{\underline{U}_{2n}/\sqrt{Z_2}}{\underline{U}_1/\sqrt{Z_1}}, \qquad T_f = \frac{\underline{U}_{2f}/\sqrt{Z_2}}{\underline{U}_1/\sqrt{Z_1}} \qquad (10)\ (11)$$

NOTE 1 $|T|^2$ 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).

NOTE 2 The quantity $|1/T|^2$, or in logarithmic quantities

$$A_s = -20\ \log_{10}|T|, \qquad (12)$$

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 $S\{l\,f\}$ 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:

$$S_{\substack{n\\f}}\{l\,f\} = \frac{\sin\frac{\beta l \pm}{2}}{\frac{\beta l \pm}{2}} \exp\left\{-j\frac{\beta l \pm}{2}\right\} \qquad (13)$$

with

$$\begin{aligned}\beta l \pm &= (\beta_2 \pm \beta_1) \cdot l = 2\pi l f \{1/v_2 \pm 1/v_1\} \\ &= 2\pi l f \left(\sqrt{\varepsilon_{r\,2}} \pm \sqrt{\varepsilon_{r\,1}}\right)/c\end{aligned} \qquad \text{(14a) (14b) (14c)}$$

subscript ± refers to near/far end respectively

+ refers to both near/far ends

Note that weak coupling, i.e. T << 1, has been assumed. This case, including losses, is given in [20 Halme, Szentkuti][2].

NOTE 3 Equation (15) and representation in Table 1 visualizes the contributions of the different parameters to the coupling function T:

$$T_{\substack{n\\f}} = (Z_F \pm Z_T) \cdot \frac{1}{\sqrt{Z_1 \cdot Z_2}} \cdot \frac{l}{2} \cdot S_{\substack{n\\f}}\{l \cdot f, \varepsilon_{r1}, \varepsilon_{r2}\} \qquad (15)$$

Note especially the following points:

a) 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.)

b) Up to a constant factor, T is the quantity directly measured in a set-up.

c) For low frequencies, i.e. for short cables ($l << \lambda$), the trivial coupling formula is obtained that is directly proportional to l :

$$T_{\substack{n\\f}} = (Z_F \pm Z_T) \cdot \frac{1}{Z_{12}} \cdot \frac{l}{2} \quad \text{with} \quad Z_{12} = \sqrt{Z_1 \cdot Z_2} \qquad \text{(15) (16a) (16b)}$$

d) The summing function $S\{l \cdot f\}$ is presented in Figure 4. Note also that:

e) $S\{l \cdot f\}$ has a sin(x)/x behaviour. A cut-off point may be defined as $(l \cdot f)_C$:

$$(l \cdot f)_{C_{\substack{n\\f}}} = \frac{c}{\pi \left|\sqrt{\varepsilon_{r1}} \pm \sqrt{\varepsilon_{r2}}\right|} \qquad (17)$$

f) The exact envelope of $S\{l \cdot f\}$ is

$$\text{Env}\left|S_{\substack{n\\f}}\{l \cdot f\}\right| = \frac{1}{\sqrt{1 + \dfrac{(l \cdot f)^2}{(l \cdot f)^2_{\substack{cn\\f}}}}} \qquad (18)$$

2 Figures in square brackets refer to the bibliography.

Table 1 – The coupling transfer function T (coupling function)[a]

Set-up parameters [b] $(Z_1), l, \varepsilon_{r1}$ $T_{\substack{n\\f}} = (Z_F \pm Z_T) \cdot \frac{1}{\sqrt{Z_1 \cdot Z_2}} \cdot \frac{l}{2} \cdot S_{\substack{n\\f}}\{l \cdot f, \varepsilon_{r1}, \varepsilon_{r2}\}$ Intrinsic screen parameters — Cable parameters [b] $(Z_2, l), \varepsilon_{r2}$ "Low-frequency coupling", short cables [c] — "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 $\substack{n\\f}$ are associated to the stacked operation symbols ± in the obvious way: upper subscript → upper operation, lower subscript → lower operation. [b] ε_{r1} and ε_{r2} contained in S as parameters. [c] for $l \ll \lambda$: $S\{l\, f\} \rightarrow 1$.

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

$$(l \cdot f)_{\min} = \pi (l \cdot f)_{\mathrm{C}} \tag{19}$$

h) As seen from Equations (13) and (18), below the cut-off points $(l \cdot f)_{\substack{\mathrm{cn}\\\mathrm{f}}}$ is $S\{l \cdot f\} \approx 1$ and above them it starts to oscillate and its envelope drops asymptotically 20 dB/decade,

$$\mathrm{Env}\left| S_{\substack{\mathrm{n}\\\mathrm{f}}}\{l \cdot f\} \right| = \frac{\left((l \cdot f)_{\substack{\mathrm{cn}\\\mathrm{f}}} \right)}{(l \cdot f)} \tag{20}$$

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

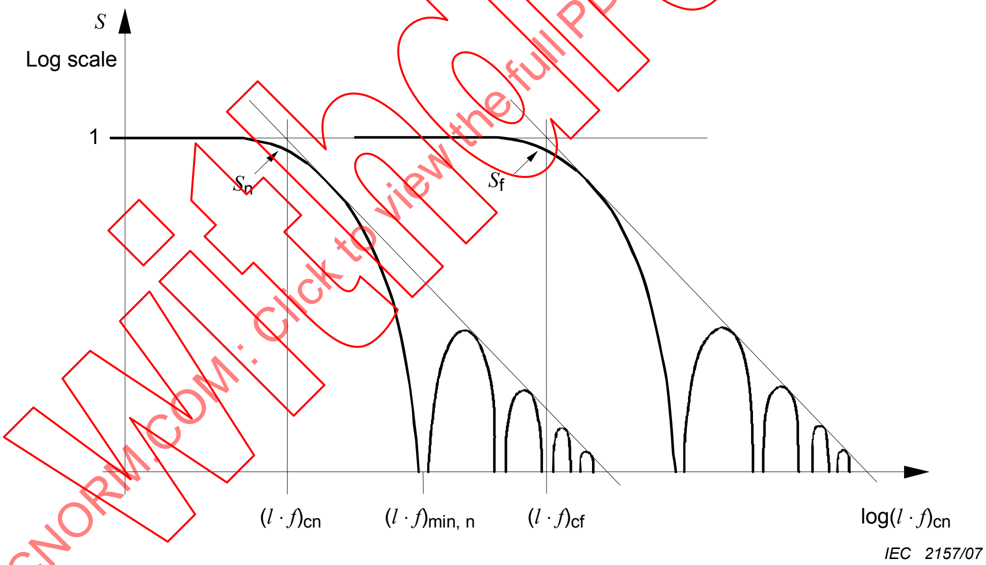
$$l_{C_{\mathrm{n} \atop \mathrm{f}}} = \frac{\lambda_0}{\pi \left| \sqrt{\varepsilon_{r1}} \pm \sqrt{\varepsilon_{r2}} \right|} \tag{21}$$

j) The effect of S in the frequency range (l = constant) is illustrated in Figure 5. 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 but 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}). In selecting $\varepsilon_{r1} \to \varepsilon_{r2}$ we obtain $(l \cdot f)_{Cf} \to \infty$, 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 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 $l \times S$ is presented in Figure 7 for f = 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 a length of independent shielding attenuation is obtained 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).

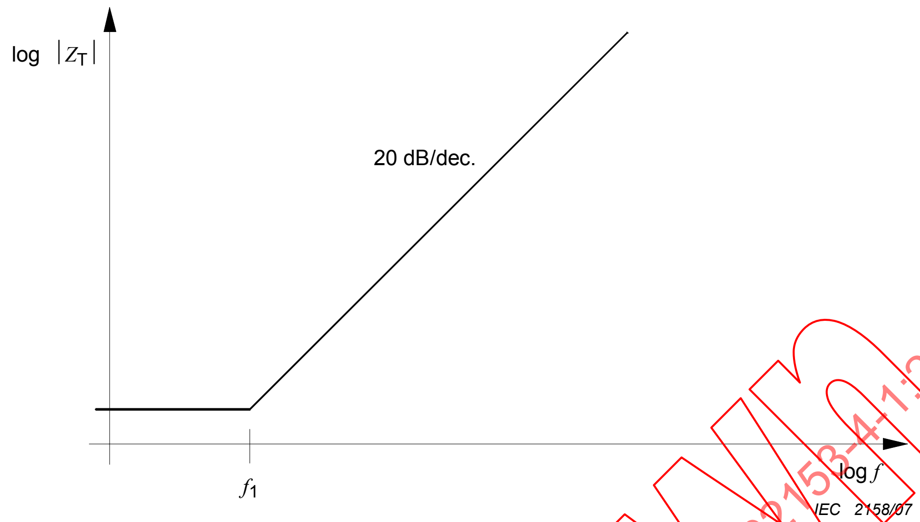


NOTE $S_f > S_n$ above near end cut-off, yielding a directive effect.

Key

$(l \times f)_c$ cut-off point

Figure 4 – Summing function $S\{l \cdot f\}$ for near (n) and far (f) end coupling



$$Z_T(f_1 = 10\,\text{MHz}) = 20\ \text{m}\Omega/\text{m}$$

Figure 5a – Transfer impedance of a typical single braid screen

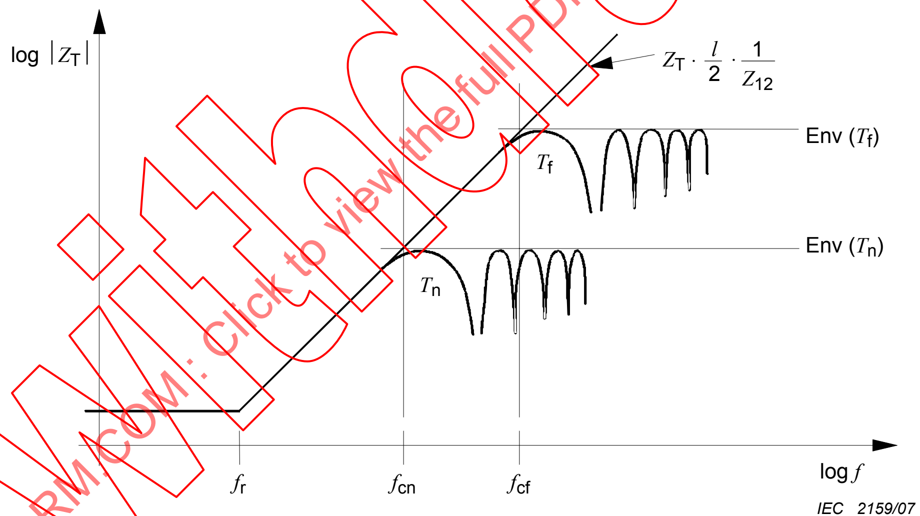


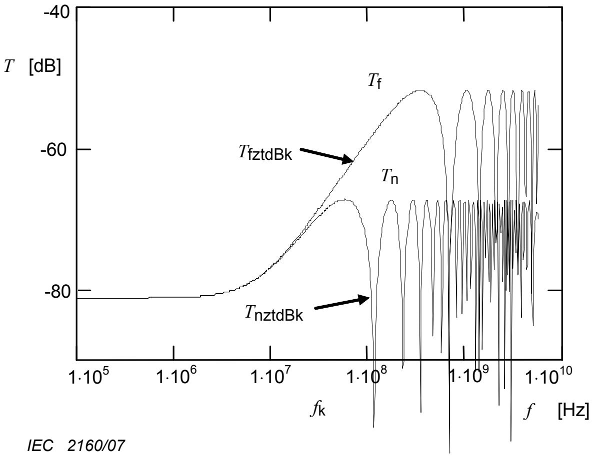
Figure 5b – Coupling transfer function for the same cable with negligible Z_F(Z_F << Z_T): frequency responses of Figure 4 and Figure 5a added on log scale

NOTE The cut-off effect for $f > f_c$.

EXAMPLE: $\varepsilon_{r1} = 1$ (set-up), $\varepsilon_{r2} = 2.2$ (cable),

$l = 1\ \text{m} \rightarrow f_{Cn} = 40\,\text{MHz}, \quad f_{Cf} = 200\,\text{MHz}$

Figure 5 – The effect of the summing function

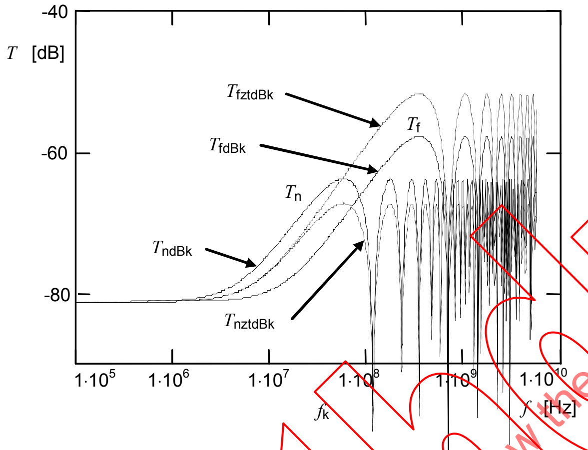


IEC 2160/07

Figure 6a – Calculated coupling transfer functions T_n and T_f for a single braided when Z_F = 0

– In calculations the used parameters are:

Z_T (d.c.) =15 mΩ/m and Z_T (10 MHz) = 20 mΩ/m increasing 20 dB/decade (see Figure 5a), cable length 1 m, and velocities of the outer and inner line: v_1 = 200 Mm/s and v_2 = 280 Mm/s corresponding a velocity difference of 40 %.



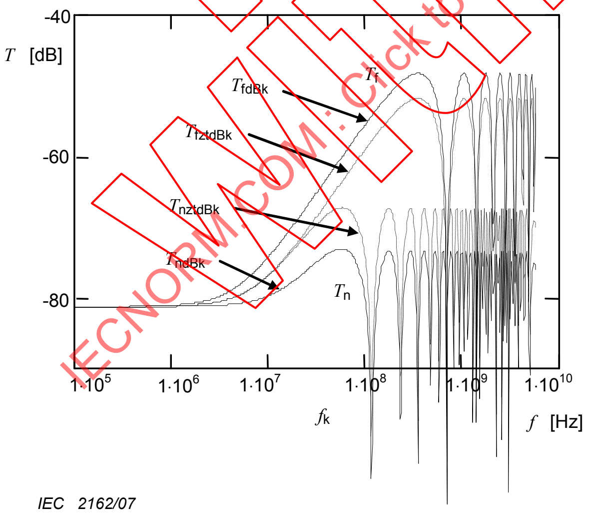
IEC 2161/07

Figure 6b – As Figure 6a but Im(Z_T) is positive and Z_F = +0,5*Im (Z_T) at high frequencies:

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

$T_n \sim | Z_F + Z_T |$ = 1,5*Z_T and

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



IEC 2162/07

Figure 6c – As Figure 6a but Im(Z_T) is negative and Z_F = –0,5*Im(Z_T) at high frequencies:

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

$T_f \sim | Z_F - Z_T |$ = 1,5* $| Z_T |$ and

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

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

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

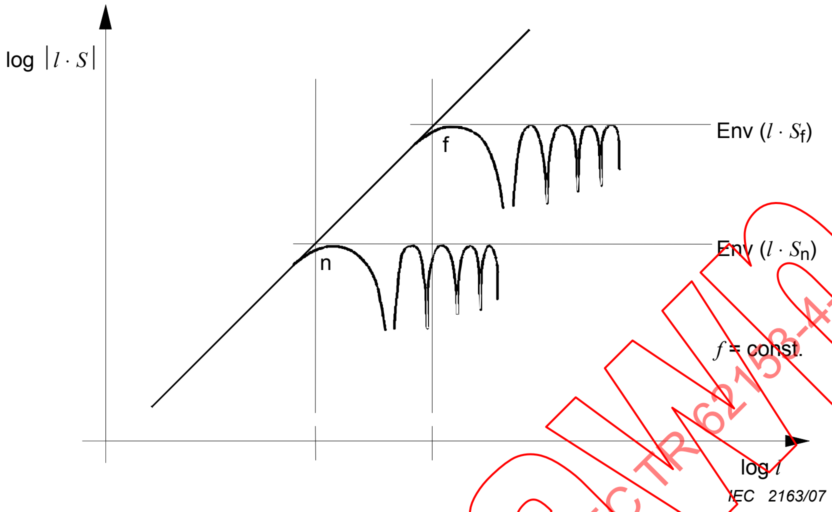
NOTE 3 T_f dB: far-end when $Z_F = (1/2) \cdot Z_T$ and T_{fzt} dB: far-end when Z_F = 0.

Figure 6 – The effects of the Z_T and Z_F to the coupling transfer functions T_n and T_f

– In Figure 6a, Z_F = 0.

– In Figure 6b and Figure 6c, Z_F is significant ($Z_F = (1/2) \cdot Z_T$).

– In Figure 6b Z_T is positive and Figure 6c negative at high frequencies.



NOTE 4 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 5 l_{cf} strongly depends on ε_{r1}.

Figure 7 – $l \times S$: complete length dependent factor in the coupling function T (see Table 1)

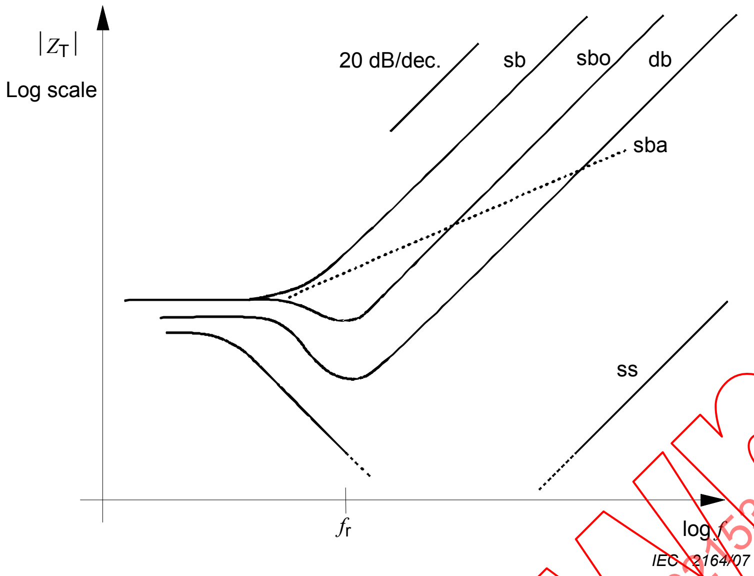
7 Transfer impedance of a braided-wire outer conductor or screen

Typical transfer impedances of cables with braided-wire screens are shown in Figure 8. The constant Z_T value at the low-frequency end is equal to the DC resistance of the screen, the 20 dB/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/decade rise at high frequencies. By using an extrapolation of 20 dB/decade, this remains in most cases on the conservative side. This extrapolation can be used up to several GHz.

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 (1) and outer (2) circuit of Figure 9a.



Key

f_r	typically 1 .... 10 MHz
sb	single braid
sbo	single braid optimized
sba	single braid ‘anomalous’
db	double braid
s	superscreen

Figure 8 – Transfer impedance of typical cables

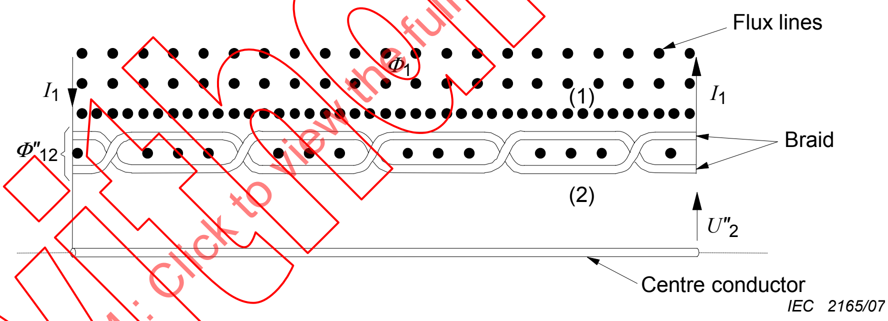


Figure 9a – Complete flux

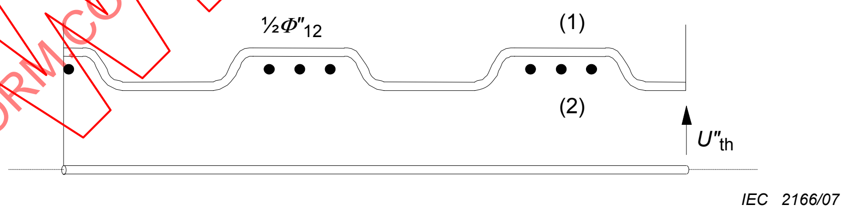


Figure 9b – Left-hand lay contribution

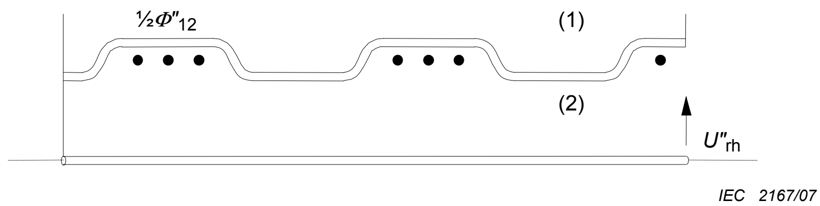


Figure 9c – Right-hand lay contribution

Figure 9 – Magnetic coupling in the braid

Therefore it is necessary and unavoidable that Φ''_{12} is partly also in the inner circuit, Figure 9b. Both the right hand (rh) and left hand (lh) lay of the braiding wires bring into the inner circuit (2) an equal disturbing voltage U''_2 induced by Φ''_{12} / 2. The voltages are in parallel:

$$U''_{rh} = U''_{lh} = U''_2 = \tfrac{1}{2} j\omega\Phi''_{12} \tag{21}$$

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

The two induced disturbing voltages oppose each other.

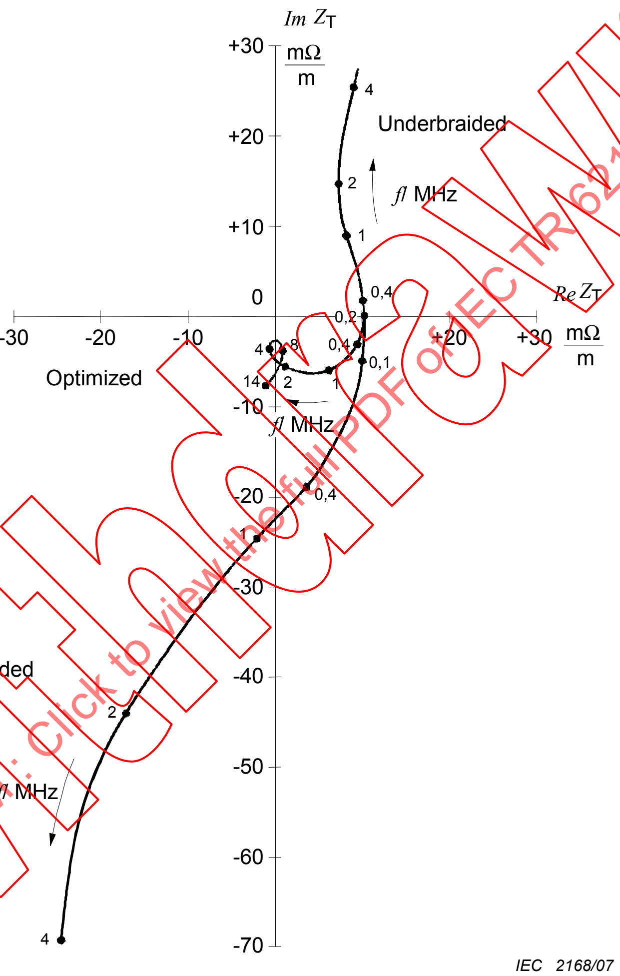


Figure 10a – Complex plane, $Z_T = Re\, Z_T + j\, Im\, Z_T$, frequency f as parameter

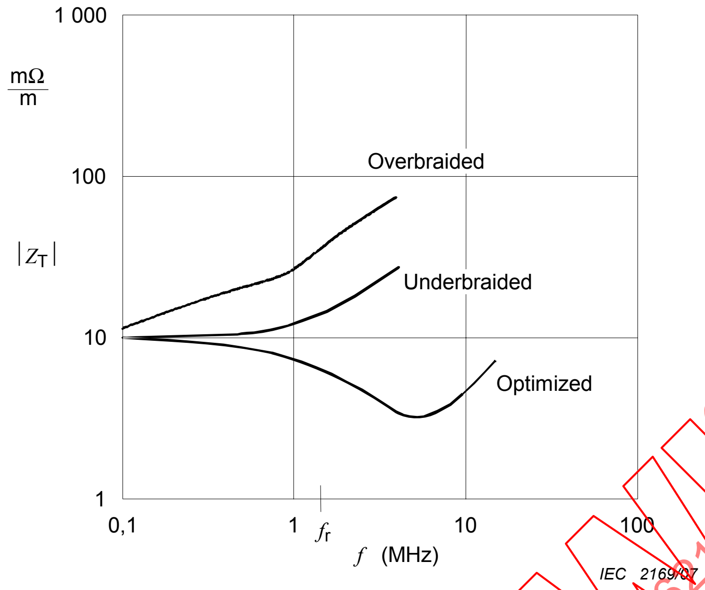


Figure 10b – Magnitude (amplitude), | Z_T (f) |

Figure 10 – Measured transfer impedance Z_T (d.c. resistance Z_T (d.c.) set to the value of 10 mΩ/m

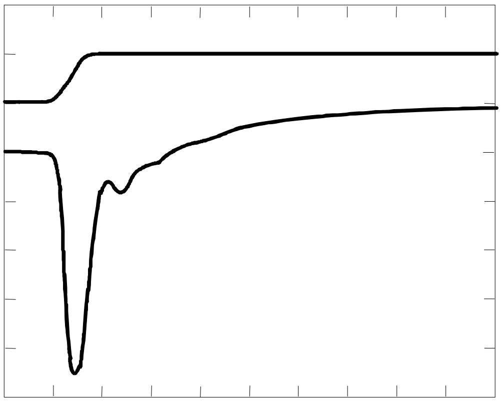

IEC 2170/07

Figure 11a – Overbraided cable

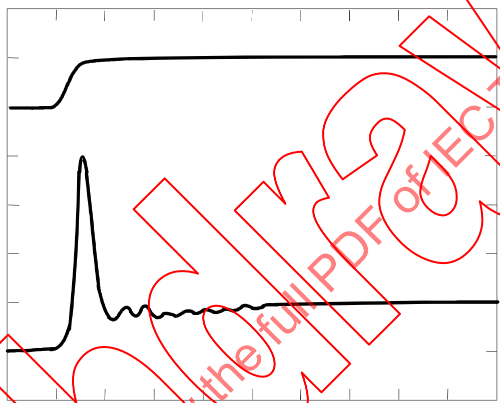

IEC 2171/07

Figure 11b – Underbraided cable

Key
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: height of spike corresponds to:

- Figure 11a – Z_t(3 MHz) = –4,7 × 12,5 mΩ/m = –59 mΩ/m
- Figure 11b – Z_t(3 MHz) = +4,7 × 12,5 mΩ/m = +50 mΩ/m

Figure 11 – 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:

$$M'_{12} = \frac{\Phi'_{12}}{j\omega I_1} \quad (22)$$

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

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 10a 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 10b where the transfer impedance amplitudes are shown as a function of frequency, but from Figure 10b it cannot be seen directly if the screen is underbraided or overbraided. Figure 11 gives the typical time domain characteristics of under- and over-braided, single braid coaxial cable.

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 [14 Vance] it is stated that a good approximation for Z_{Th} is a tubular homogeneous screen [5 Schelkunoff] 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

$$Z_T \approx Z_{Th} + j\,\omega\,(M'_{12} - M''_{12}) \tag{24}$$

and the first approximation of the equivalent circuit is shown in Figure 12a.

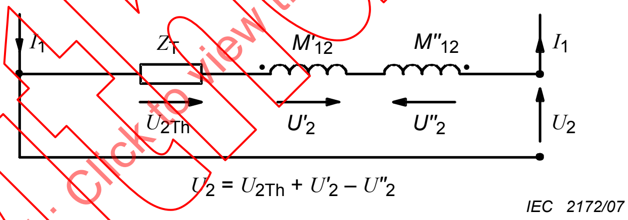


Figure 12a – Contributions to the transfer impedance

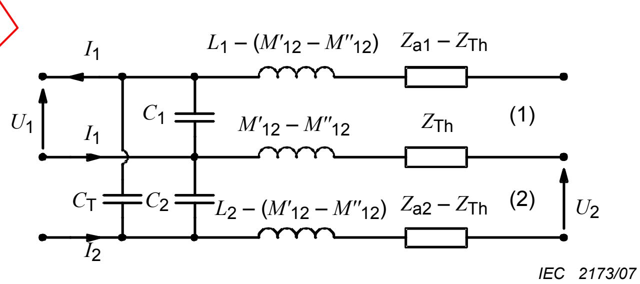


Figure 12b – Significant elements of circuits (1) and (2)

Figure 12 – 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 12b. 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 [15 Ikrath], [2 Kaden], [14 Vance] 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 [1 Fowler], [3 Tyni] on M''_{12} but the matter has been studied by IEC SC 46A/WG 1 and its successor TC 46/WG 5. 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 decreasing 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 on the braid
  – in manufacturing,
  – of the jacket.

IEC TC 46/WG 5, Screening effectiveness, is studying the impact of these parameters on an optimized braid when preparing a guide for braided optimization based on theory and practice.

A guide on screening optimization of braids will help the IEC family to talk the same language when setting limits for electromagnetic screening parameters (Z_T; Z_F; a_s) of braided cables.

8 Test possibilities

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.

8.1 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, whilethose with 'o' under grouping the injection circuit is unscreened. The difference is the line interchange referred to in 5.4 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 3 and 8. The latter, with its wide frequency coverage is recommended for future testing.

8.2 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 (2 and 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.

8.3 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 address come from:

a) the fact that 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.

8.4 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 set-up or even to optimize the test set-up.

The following test equipment specifications are required for the calculation:

- minimum input (noise floor);
- maximum input;
- amplification/attenuation;
- maximum output.

Figure 13 gives a visualization of the different signal levels in a generic test set-up. 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.

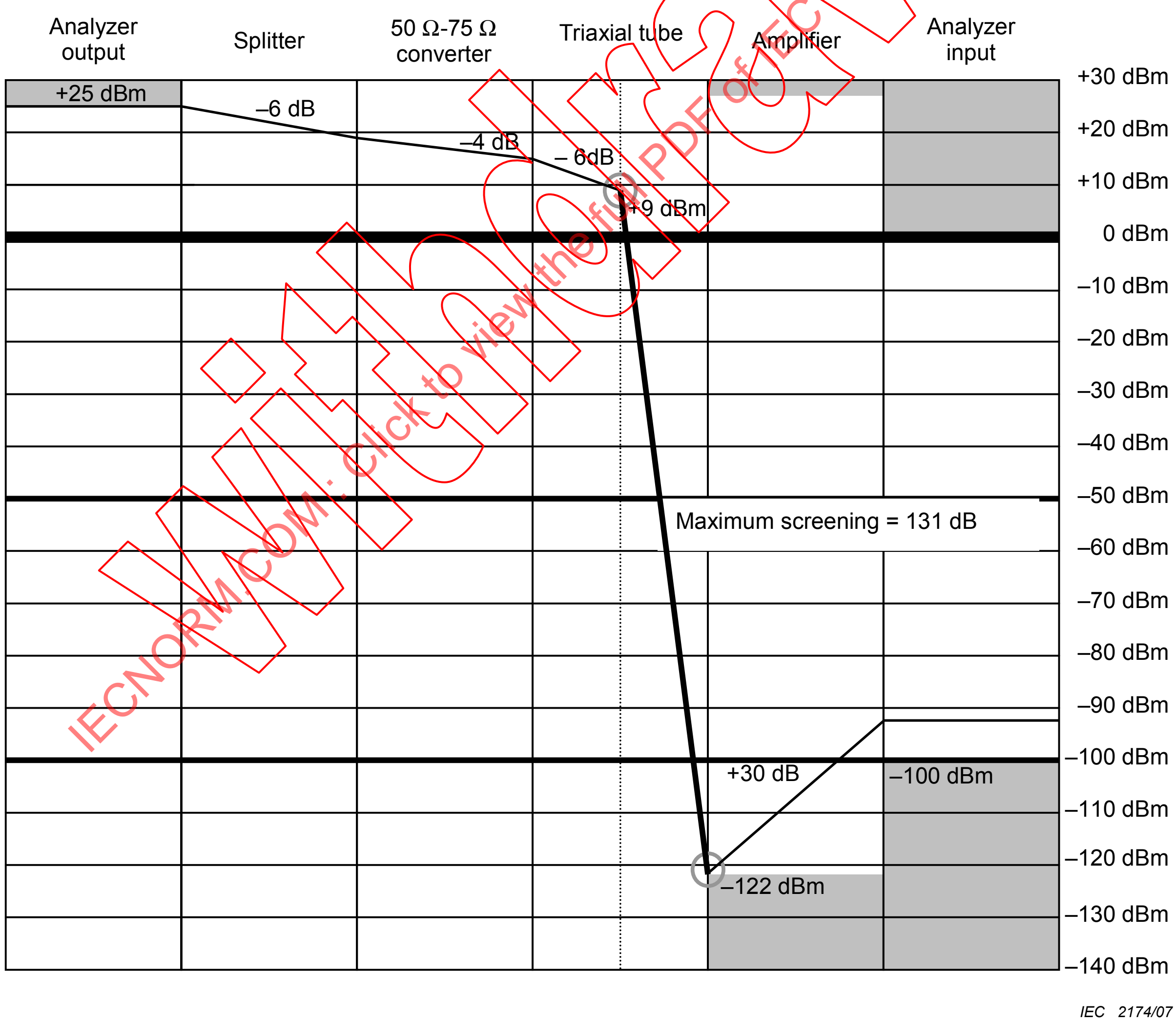


Figure 13 – Example of visualization of the maximum measurable screening level

Taking into consideration the noise level of 1 Hz bandwidth at room temperature being -173 dBm, (increase 10 x log (bandwidth) dB), and adding the noise figure of the amplifier, the theoretical noise level of the test set-up is obtained. Assuming that the amplifier in the Figure 13 example has a noise figure of 11 dB, the bandwidth (Δf) of the network analyser can be calculated to be smaller than 10 kHz.

This can be expressed as a general formula:

$$NL = (-173 + F + 10 \log \Delta f) \tag{25}$$

where

NL is the noise floor level of the 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;

dBm is dB(mW).

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

Short title	Reference	Grouping (note 1)	Frequency range		Injection N or F (note 2)	Advantages or shortcomings
			Possible	Actually used		
1 IEC triaxial	Fig. A9 of IEC 60096-1 Fig. 44 of IEC 61196-1	kf s	d.c. – 50 MHz	10 kHz – 30 MHz	F	Rigid test rig
2 Terminated triaxial (Simons)	Fig. A5 of IEC 60096-1	m s	10 kHz – 1 GHz	100 kHz – 500 MHz	N F	Flexible test jig relies on ferrites
3 Braid injection (Fowler)	AESS(TRG)71181 [6]	m o	d.c. – 500 MHz	10 kHz – 500 MHz	N F	Flexible test needs good screening on measuring system
4 Quadraxial	[7]	m s	100 kHz – 50 MHz	100 kHz – 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 [8] [9]	m s	1 kHz – 12 GHz	100 MHz – 10 GHz 10 kHz – 100 MHz	N F	Rigid test jig needs good screening
6 ERA triaxial (Smithers)	[10]	kf s	d.c. – 400 MHz	10 kHz – 300 MHz	F	Very short CUT requires amplifier or phase locked loop
7 Line injection (time domain)	IEC 60096-4-1 [11]	m o	d.c. – 100 MHz	1 kHz – 80 MHz (note 3)	N F	Very easy to use. Needs good screening in measuring amplifier
8 Line injection (frequency domain)	Figures 34 and 35 of IEC 61196-1 [4] [12]	m o	d.c. – 20 GHz	10 kHz – 3 GHz	N F	Flexible and cheap measuring set-up, equipment needs to be well shielded
9 Open screening attenuation test method (absorbing clamp)	Figures 50 to 52 of IEC 61196-1	m o	30 MHz – 2,5 GHz	30 MHz – 1 GHz 300 MHz – 2,5 GHz	N F	Poor sensitivity. Measuring of a_s is dependent on the surroundings
10 Reverberation chamber method	IEC 61726 [13]	kn kf	0,1 GHz →	0,3 GHz – 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 62153-4-3 IEC 62153-4-5 [16] [17] (note 4)	m s	d.c. – 20 GHz	10 kHz – 3 GHz	F	High-sensitivity measurements can be made without a screened room, Transfer impedance and Screening attenuation with one test set-up
12 Open multipin connector screening test method	[18] [19]	o	d.c. – 1 GHz	10 kHz – 700 MHz	N	Low cost and flexible

Table 2 *(continued)*

Short title	Reference	Grouping (note 1)	Frequency range		Injection N or F (note 2)	Advantages or shortcomings
			Possible	Actually used		
13 Coupling attenuation measurements of balance cables and cable-assemblies						Under study
13.1 Current clamp injection method	IEC 62153-4-2 [21]		50 MHz - 1 GHz	50 MHz - 1 GHz		Under study
13.2 Shielded triaxial test method	IEC 62153-4-9 [22]		d.c. - 1 GHz	d.c. - 1 GHz		Under study
13.3 Absorbing clamp method	IEC 62153-4-5		50 MHz - 2.5 GHz	50 MHz - 2.5 GHz		Under study
14 Shielded screening attenuation, test method for measuring the transfer impedance Z_T and the screening attenuation a_S of RF connectors up to and above 3 GHz; tube in tube method	IEC 62153-4-7 [23] [24]	m s	d.c. – 20 GHz	d.c. – 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	IEC 62153-4-10	m s	d.c. – 4 GHz	d.c. – 3 GHz		Under study

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 µs.

NOTE 4 Secondary circuit near end short-circuited.

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

9.1 Introductory remark

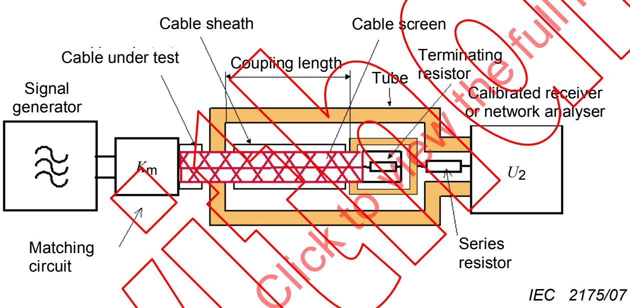
Different triaxial test set-ups for the measurement of the transfer impedance exist. One set-up is according to EN 50289-1-6 another according to IEC 61196-1. All of them are based on the same principle but use different load conditions. In one method, for example, the cable under test is matched, while 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 can be measured.

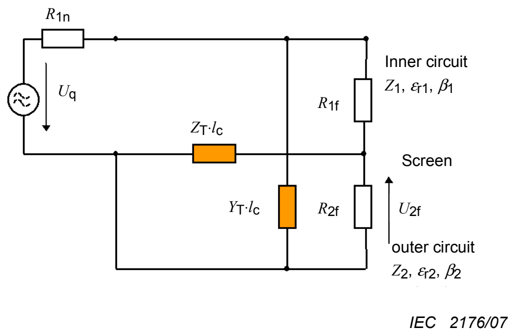
9.2 Physical basics

9.2.1 Triaxial set-up

9.2.1.1 General

The triaxial set-up is of the "triple coaxial" form, see Figure 14 and Figure 15. 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.





where

$Z_{1,2}$ is the characteristic impedance of the inner circuit (cable) respectively outer circuit (tube);
$A_{1,2}$ 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);
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;
$R_{1,f}$ is the load resistance at the far end of the inner circuit (cable). Depending on the method used, either equal to the characteristic impedance of the cable or a short-circuit;
$R_{2,f}$ is the load resistance at the far end of the outer circuit (cable). Equal to the output impedance of the generator respectively input impedance of the receiver, including an eventually used feeding resistor;
u_q is the EMK of the generator;
$u_{2,f}$ is the voltage at the far end of the outer circuit.

Figure 14 – Triaxial set-up for the measurement of the transfer impedance Z_T

Figure 15 – Equivalent circuit of the triaxial set-up

9.2.1.2 Load conditions of the different set-ups

EN 50289-1-6 uses 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.

The first edition of IEC 61196-1(1995) described 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, one has introduced IEC 62153-4-3 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 at the near end of the outer circuit (between the cable screen and the tube), the other at the far end of the cable. The advantage of that method is the simplification of the sample preparation. A short-circuit is easier to make than to solder a resistance, 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 one puts an additional braid, the measuring braid, over the cable sheath instead of using the measuring tube. The advantage is that the sample can be bend 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 below. The impedance of the outer circuit Z_2 varies 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 lower, than the input impedance of the receiver.

Table 3 – Load conditions of the different set-ups

	Generator	Receiver	$R_{1,n}/Z_1$	$R_{1,f}/Z_1$	$Z_2/R_{2,f}$
EN 50289-1-6					
Standard	IC[a]	OC[b]	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[c]	0	1…5 depending on tube diameter
With milked on braid	IC	OC	1[c]	0	0,1…0,4 depending on screen and sheath diameter of the cable

[a] IC inner circuit (cable under test).

[b] OC outer circuit (tube).

[c] 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 and outer circuit for any load conditions are described in 0 and 0. By taking into account the short-ircuit 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 one gets following equations.

$$\frac{u_{2\mathrm{f}}}{u_\mathrm{q}} = \frac{L}{R_{1\mathrm{f}} + R_{1\mathrm{n}}}\left[Z_\mathrm{T} \cdot g + Z_\mathrm{F} \cdot h\right] \tag{26}$$

$$g = \frac{1}{N} \cdot \frac{1}{1-n^2} \cdot \frac{j}{x}\left\{r \cdot [\cos x - \cos nx] - j \cdot n \cdot \sin nx + j \cdot \sin x\right\} \tag{27}$$

$$h = \frac{1}{N} \cdot \frac{1}{1-n^2} \cdot \frac{j}{x}\left\{nr \cdot [\cos x - \cos nx] - j \cdot \sin nx + j \cdot n \cdot \sin x\right\} \tag{28}$$

$$N = \left\{\cos x + \frac{j \cdot \sin x}{r + w}[1 + r \cdot w]\right\} \cdot \left\{\cos nx + j \cdot v \cdot \sin nx\right\} \tag{29}$$

$$x = \beta_1 \cdot L = 2\pi \cdot \frac{L}{\lambda_1} \qquad n = \frac{\beta_2}{\beta_1} = \frac{\lambda_2}{\lambda_1} = \sqrt{\frac{\varepsilon_{r2}}{\varepsilon_{r1}}} \tag{30}$$

$$r = \frac{R_{1f}}{Z_1} \qquad v = \frac{Z_2}{R_{2f}} \qquad w = \frac{R_{1n}}{Z_1} \tag{31}$$

where

$Z_{1,2}$ is the characteristic impedance of the inner circuit (cable) respectively outer circuit (tube);
$\varepsilon_{1,2}$ 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 wavelength 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;
$R_{1,f}$ is the load resistance at the far end of the inner circuit (cable). Depending on the method used, either equal to the characteristic impedance of the cable or a short-circuit;
$R_{2,f}$ is the load resistance at the far end of the outer circuit (cable). Equal to the output impedance of the generator respectively input impedance of the receiver, including an eventually used feeding resistor;
u_q is the EMK of the generator;
$u_{2,f}$ is the voltage at the far end of the outer circuit.

Factors g and h describe the frequency response of the test set-up. At low frequencies when $\lambda >> L$, factors g and h are equal to 1. However, with increasing frequency factors g and h start to oscillate and thus also the measurement results. The maximum frequency to which the transfer impedance can 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, if factor g, respectively h become $>\sqrt{2}$ respectively $<1/\sqrt{2}$.

9.3 Simulations

9.3.1 General

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

Due to the reciprocity of the materials one can 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 give 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=$R_{1,f}/Z_1$	v=$Z_2/R_{2,f}$	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, 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" 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 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 multipair/conductor cables. For the "milked on braid" method ,one has only taken into account typical combinations of insulation and sheath materials (PE/PVC, PE/LSZH, PTFE/FEP…) 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 to 1/√2. The simulations below show that this factor is a good compromise with respect to the maximum frequency to which the transfer impedance can be measured.

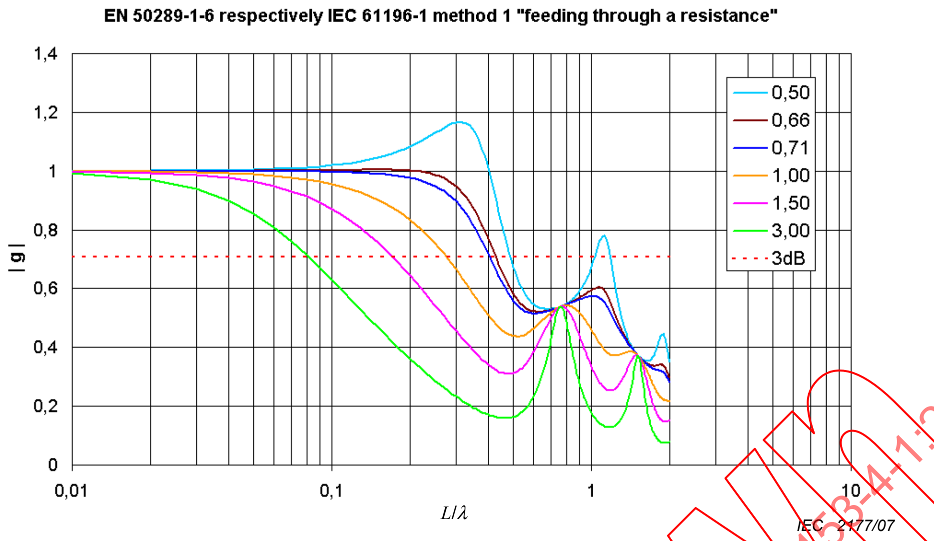


Figure 16 – Simulation of frequency response for different factors of v = $Z_2/R_{2,f}$ with ε_{r1} = 2,3 (solid PE), ε_{r2} = 1,0, n = 0,659

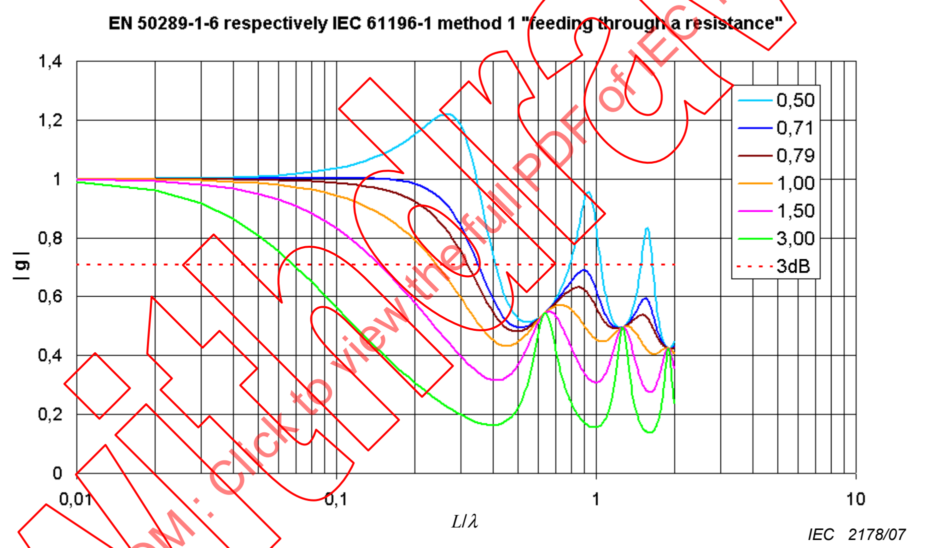


Figure 17 – Simulation of the frequency response for different factors of v = Z_2/R_{2f} with ε_{r1} = 1,6 (foam PE), ε_{r2} = 1,0, n = 0,791

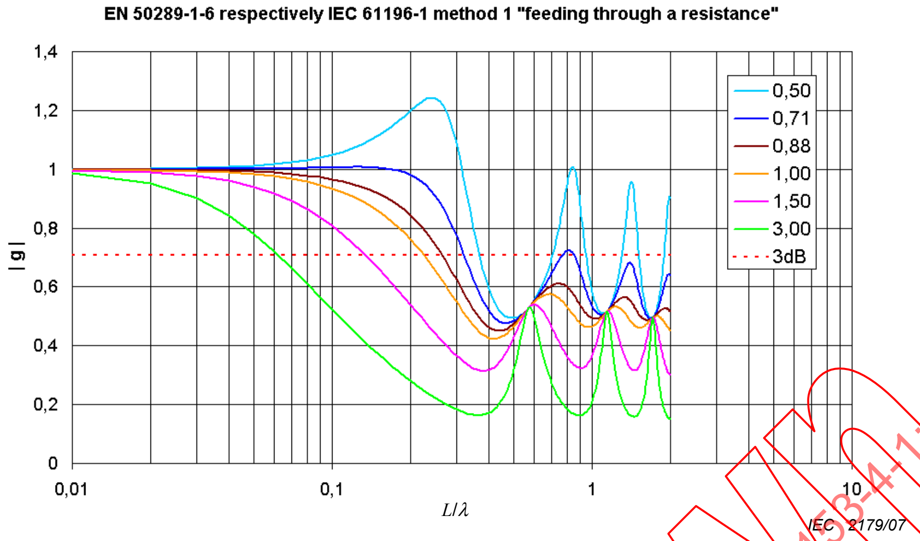


Figure 18 – Simulation of frequency response for different factors of v = Z_2/R_{2f} with ε_{r1} = 1,3 (foam PE), ε_{r2} = 1,0, n = 0,877

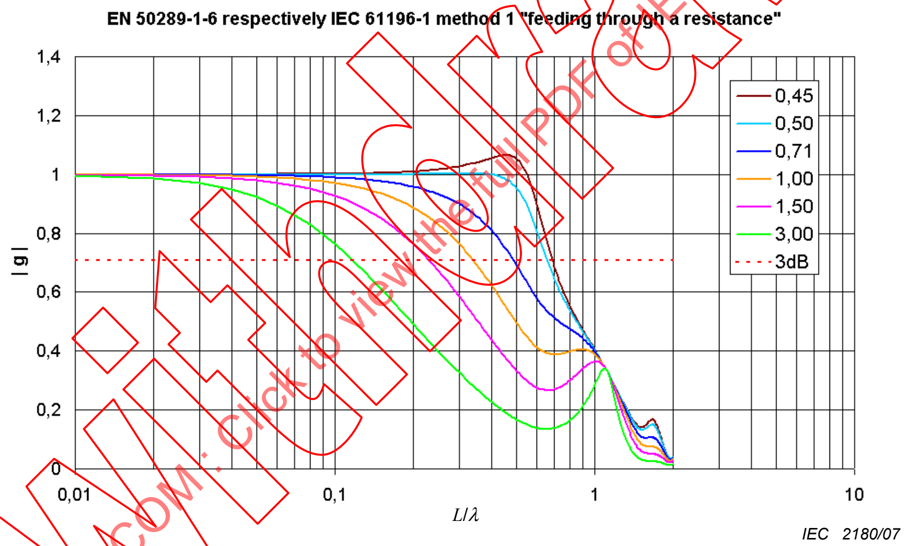


Figure 19 – Simulation of frequency response for different factors of v = Z_2/R_{2f} with ε_{r1} = 5 (PVC), ε_{r2} = 1,0, n = 0,447

The highest frequencies (respectively shortest wavelengths) are obtained if the factor v = 1/√2 respectively v = n, whichever is smaller. In Figure 16 and Figure 19 the highest frequency is obtained for v = n (= 0,659 respectively 0,447). But in Figure 17 and Figure 18 the highest frequency is obtained for v = 1/√2 = 0,71. If one falls below that value, than factor g overshoots, i.e. becomes higher than one. If one oversteps that value, than the cut-off frequency decreases.

In Figure 20 the 3 dB cut-off wavelength (L/λ_1) is calculated by iteration at which factor |g| becomes 1/√2. The graph is given as a function of the factor n = $\sqrt{\varepsilon_{r2}}/\sqrt{\varepsilon_{r1}}$ and for different factors v = Z_2/R_{2f}. The curves show a linear behaviour and can be interpolated by a straight line.

This has been done in Figure 21 for v = 1/√2, v = 1, v = 1,8 and v = 3,6. The factor v = 1/√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 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) one obtains an impedance in the outer circuit of 90 Ω to 180 Ω, thus for direct feeding resulting in v = 1,8…3,6.

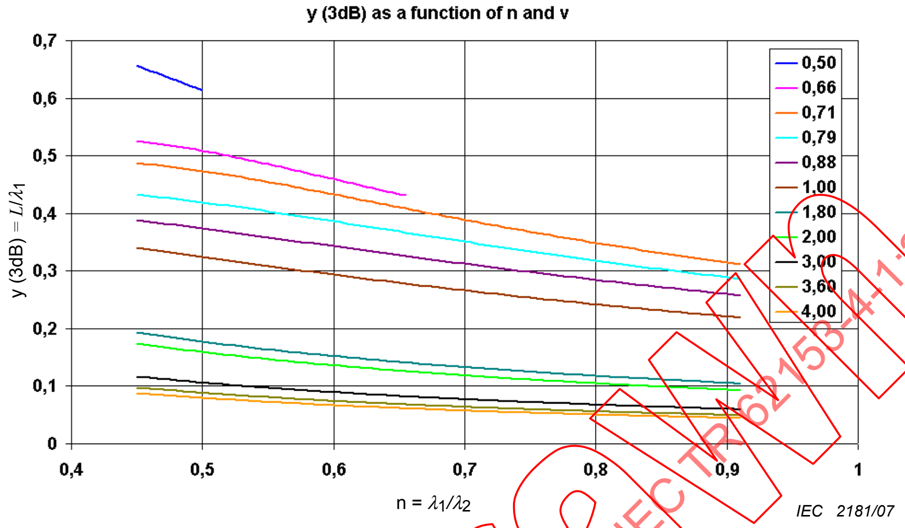


Figure 20 – Simulation of the 3 dB cut off wavelength (L/λ_1) as a function of factor n = $\sqrt{\varepsilon_{r,2}} / \sqrt{\varepsilon_{r,1}}$ given for different factors v = Z_2/R_{2f}

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

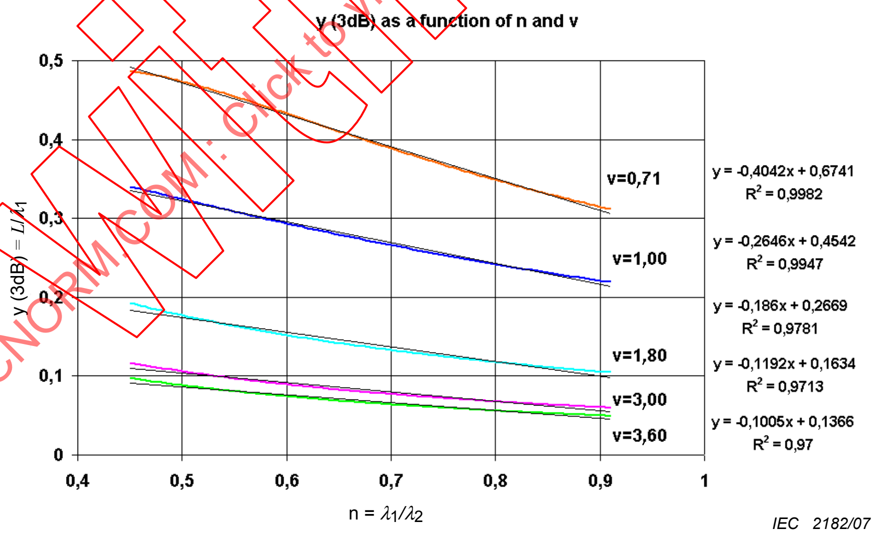


Figure 21 – Interpolation of the simulated 3 dB cut off wavelength (L/λ_1) as a function of factor n = $\sqrt{\varepsilon_{r,2}} / \sqrt{\varepsilon_{r,1}}$ given for different factors v = Z_2/R_{2f}

From the found linear interpolation, one can derive an equation to calculate the cut-off frequency length product from which the transfer impedance can be measured in a given triaxial test set-up. The equations are given in Table 5.

Table 5 – Cut-off frequency length product

EN 50289-1-6 **IEC 61196-1 method 1 "feeding through a resistance** **IEC 62153-4-3 method A "matched-short"**	**v = 1/√2**	$(f \cdot L)_{3\text{dB}} \approx \left[\frac{200}{\sqrt{\varepsilon_{r1}}} - \frac{120}{\varepsilon_{r1}}\right] \text{MHz} \cdot \text{m}$	(32)
Simplified EN 50289-1-6 **IEC 61196-1 method 2 "direct feeding"**	**v = 1**	$(f \cdot L)_{3\text{dB}} \approx \left[\frac{135}{\sqrt{\varepsilon_{r1}}} - \frac{80}{\varepsilon_{r1}}\right] \text{MHz} \cdot \text{m}$	(33)
	V = 1,8	$(f \cdot L)_{3\text{dB}} \approx \left[\frac{80}{\sqrt{\varepsilon_{r1}}} - \frac{55}{\varepsilon_{r1}}\right] \text{MHz} \cdot \text{m}$	(34)
	v = 3	$(f \cdot L)_{3\text{dB}} \approx \left[\frac{50}{\sqrt{\varepsilon_{r1}}} - \frac{35}{\varepsilon_{r1}}\right] \text{MHz} \cdot \text{m}$	(35)
	v = 3,6	$(f \cdot L)_{3\text{dB}} \approx \left[\frac{40}{\sqrt{\varepsilon_{r1}}} - \frac{30}{\varepsilon_{r1}}\right] \text{MHz} \cdot \text{m}$	(36)

The given equations are drawn in the graphs of Figure 22. For example, if a cable with a PE insulation – dielectric permittivity of ε_{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 v = 0,7, 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 can 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 can be measured is around 36 MHz.

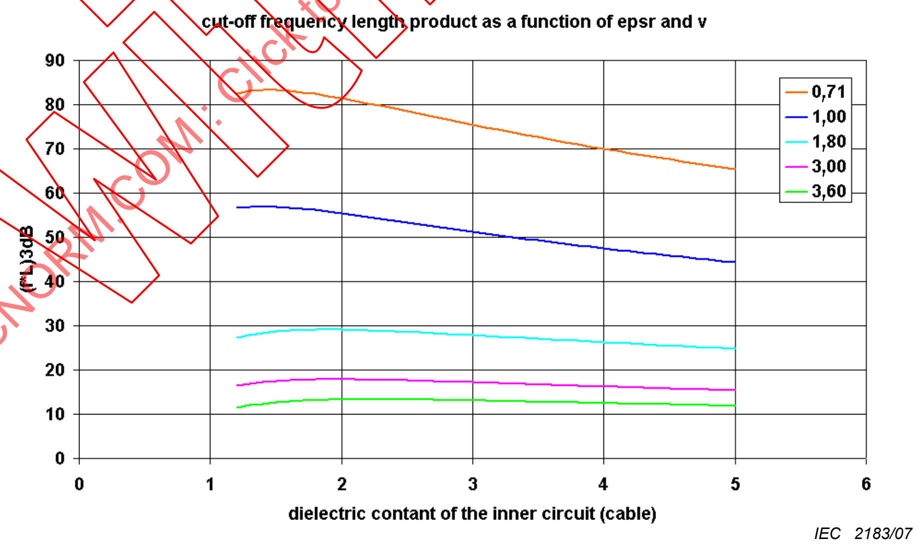


Figure 22 – 3 dB cut-off frequency length product as a function of dielectric permittivity of the inner circuit (cable) given for different factors v = $Z_2/R_{2,f}$

Figure 23 and Figure 24 show the measurement results of the normalized 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, once with a solid PE insulation (ε_{r1} = 2,3), the other with a foam PE insulation (ε_{r1} = 1,6). The measurement results confirm the simulations. From the equations given above one gets 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.

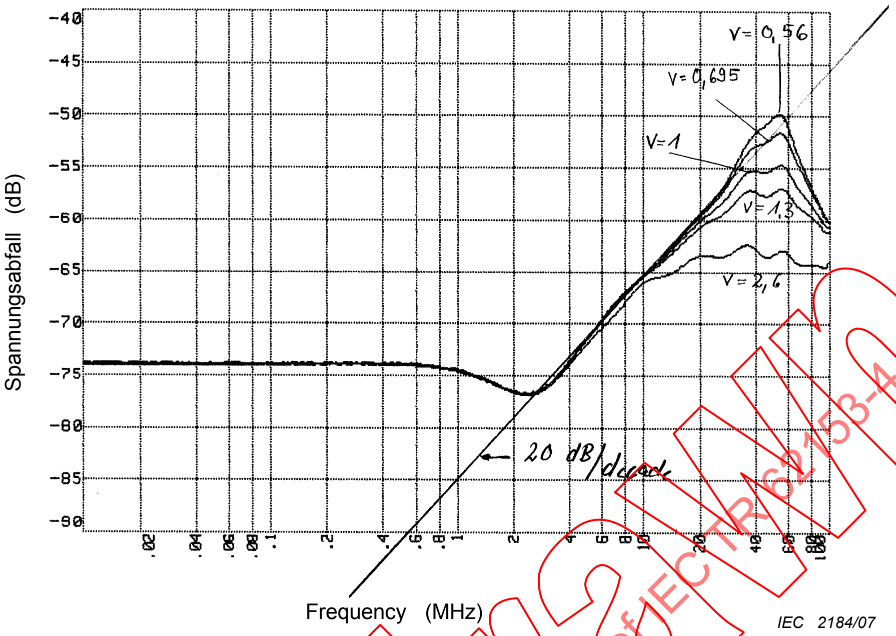


Figure 23 – Measurement result of the normalized voltage drop of a single braid screen in the triaxial set-up for different factors of v = $Z_2/R_{2,f}$ with ε_{r1} = 2,3 (PE), ε_{r2} = 1,0, n = 0,659, Z_2 = 130 Ω, L = 1 m

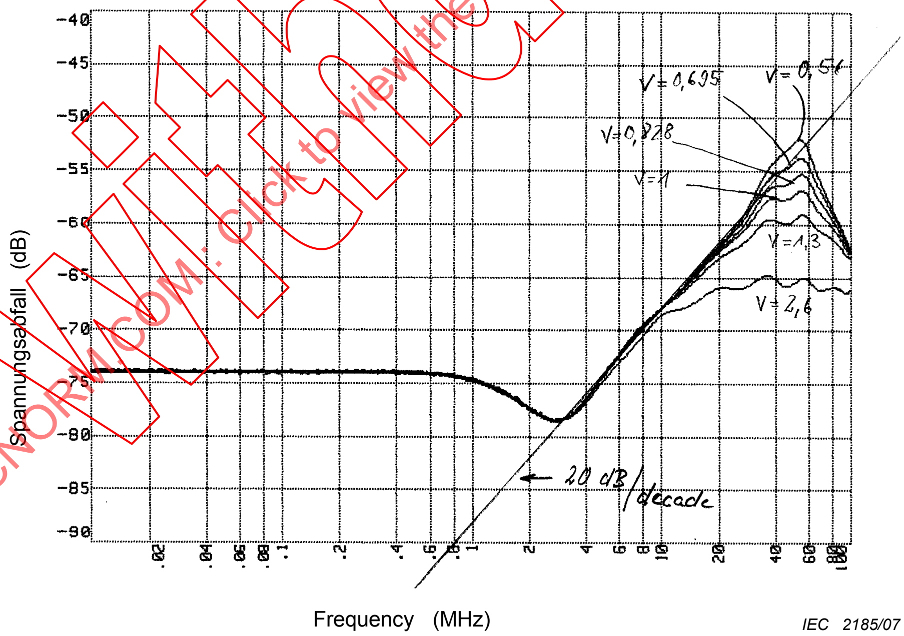


Figure 24 – Measurement result of the normalized voltage drop of a single braid screen in the triaxial set-up for different factors of v = Z_2/R_{2f} with ε_{r1} = 1,6 (foam PE), ε_{r2} = 1,0, n = 0,791, Z_2 = 130 Ω, L = 1 m

9.3.3 Simulation of the double short-circuited methods

9.3.3.1 General

For the double short-circuited methods, one has either a measuring tube or a “milked on braid”. When using a measuring tube, the dielectric permittivity of the outer circuit (tube) is nearly independent on the sheath material and could be assumed to be 1. However, in the “milked on braid” method, the dielectric permittivity is given by the sheath material. Thus, factor n is different for both methods. Also the impedance of the outer circuit is different for both methods; first, due to the different dimensions and, second, due to the different permittivities.

9.3.3.2 Simulation of the double short-circuited method using a measuring tube

The double short-circuited method, using a measuring tube, is shown in Figure 25. The outer circuit is fed over a fixed i.e. the same value for all cable types – feeding resistor, the value of which is equal to the output impedance of the generator (e.g. 50 Ω). Thus the load impedance of the outer circuit at the far end is equal to 2 times the output impedance of the generator. Factor v is then only dependent on the diameters of the screen and of the measuring tube.

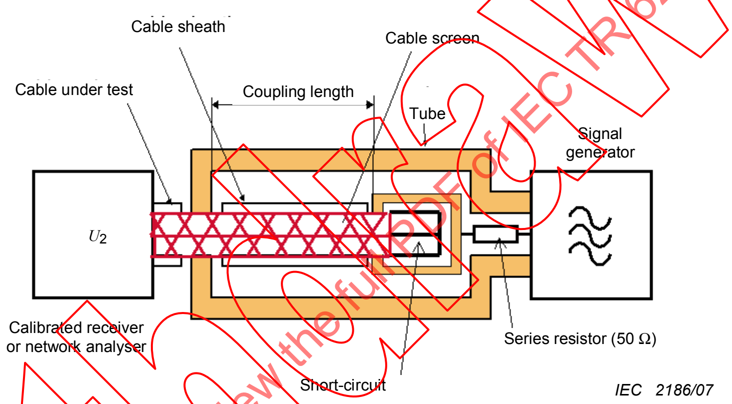


Figure 25 – Triaxial set-up (measuring tube), double short-circuited method

Table 6 – Typical values for factor v, for an inner tube diameter of 40 mm and a generator output impedance of 50 Ω

Ø screen mm	Z_2 Ω	v= $Z_2/R_{2,f}$
9	89	0,89
8	97	0,97
5	125	1,25
3.5	146	1,46
2	180	1,80

These values have been used in the following simulations. The graphs in Figure 26 to Figure 29 show the simulated frequency response for different dielectric permittivities of the cable and for the different factors of v given in Table 6 above.

In Figure 30 the 3 dB cut-off wavelength (L/λ_1) has been calculated by iteration at which factor |g| becomes $1/\sqrt{2}$. The curves have then been interpolated by straight lines.