The background for electromagnetic screening measurements of cylindrical screens

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The background for electromagnetic screening measurements of cylindrical screens*

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Summary: A brief scetch of the underlying physical phenomena is followed by the definition of transfer impedance Z_T, through capacitance C_T, normalized through elastance K_T and capacitive coupling impedance Z_F. The coupling equations through a screen as function of these parameters are given and discussed in detail, introducing the concepts of equivalent transfer impedance Z_{TE}, cut-off frequency f_c and cut-off cable length l_c. The definition of screening attenuation a_s is critically examined. Z_T, Z_F and Z_{TE} are considered as primary quantities, defining clearly the quality of a screen, while a_s should be given as additional information in rated conditions. The correct interpretation of cable screen parameters and of screening test data is the main concern of the authors.*

1 Introduction

Since the invention of the coaxial cable in the nineteenhundred and thirties it has been known that its immunity against electromagnetic disturbance may be poor, especially at low frequencies. The basic concept of low frequency measurement technics as well was developed long ago.

Yet, it has taken many years to establish a well constructed logical approach to the testing and standardization of RF coaxial screens. The difficulties were mainly related to the dependency of the dominating coupling mechanism on the cable construction, on the environment (geometrical and electromagnetic properties of the near surrounding) and on the frequency. The large frequency range in which coaxial cables are used increased these problems. Part of the development is done by the IEC Working Group on Screening Effectiveness of RF-cables. The results may be found in the IEC Publication 96-1 and in new IEC SC 46A documents (see Glossary).

This paper focuses on screening definitions and the relation between screen parameters and the coupling through a screen.

The screening effectiveness is measured by applying a well defined current and voltage to the screen in a test set-up. Usually, the set-up consists of a transmission line, the screen under test being one conductor of the transmission line, Fig. 4. Depending on frequency range, cable type, required accuracy and accepted complexity of the test, various set-ups are available for testing (see IEC Publ. 96-1).

The coupling at high frequencies through a general screen is a complicated function of the screen and set-up parameters (distributed coupling of two waves). The aim of the following sections is, to assist the correct use of set-ups and the interpretation of test data.

The resulting understandings are valid not only for test set-ups but also for the 'real world'. We stress that in the real world — in contrast to fully matched set-ups — near and far end coupling phenomena are present at the same time due to standing waves (forward and backward waves) on the outside part of the screen (see chapter 3.3. and Eq. 3.3-6).

Generally deductions will not be given except the coupling equation (see Annex 2). Detailed mathematical and physical treatment of the coupling equations and the screening parameters as function of the physical screen data and features of the set-ups are extensively given by [1] and in broad outline by [2]. The classical reference on electric and magnetic screening is [3].

Parts of the following text were prepared by the authors for the draft introduction on screening standard of the IEC SC46A, document 46A(Sec)138, March, 1987.

2 Screening basics

Screening effectiveness of cables is the ability of the outer conductor or screen, to protect the transmission line from being interfered by outside electromagnetic fields, or reciprocally, to protect the electromagnetic environment from disturbing signals emanating from the cable.

The outside electromagnetic fields induce currents and charges in the screen. These currents and charges are the main sources of the coupling into cables. The direct coupling into cables — not related to charges and currents in the screen — is usually negligible at radio frequencies above 10 kHz.

The screening effectiveness against currents is expressed in terms of the surface transfer impedance Z_T (from here on called transfer impedance) and against charges (i.e. against electric fields and voltages) in terms of the capacitive coupling impedance Z_F (or by associated parameters as feed through capacitance C_T, through admittance Y_T or through elastance K_T), see [5]. Through holes in the screen both the magnetic H and electric E fields are penetrating into the cable, *Fig. 1* [3].

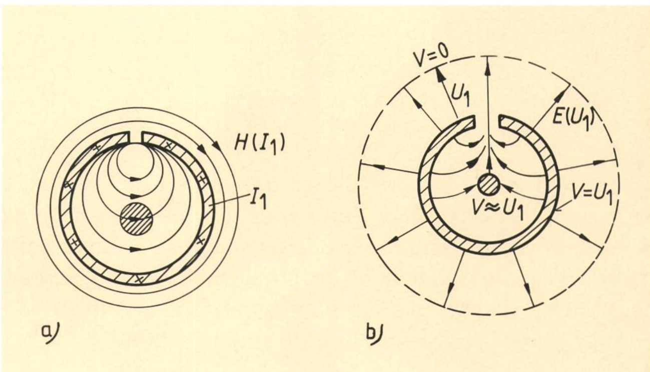

Fig. 1
Penetration of the magnetic a) and electric b) field through a hole in a screen

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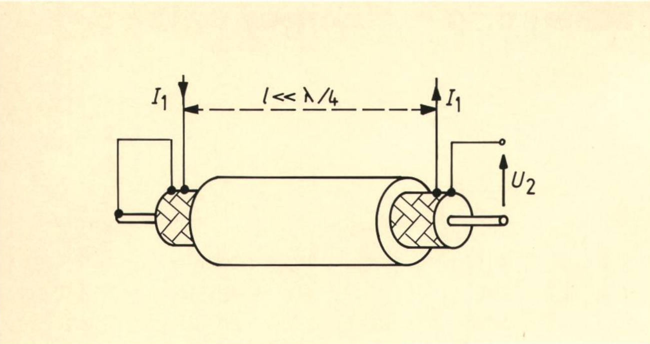

Fig. 2
Definition of the transfer impedance

In the transfer impedance definition only the current in the screen is considered. It induces a surrounding and penetrating magnetic field and a longitudinal ohmic voltage.

The transfer impedance is defined according to *Fig. 2* for an electrically short piece of cable:

$$Z_T = \frac{U_2}{I_1\, l} \qquad (2\text{-}1)$$

The transfer impedance of a homogeneous screen (i.e. of a tube without holes) is decreasing as a function of frequency depending on skin effect.

Typical transfer impedances of cables with braided-wire screens are shown in *Fig. 3*. 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 current or skin effect of the braid; see Annex 1 for further details. 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 we are always on the conservative side. This extrapolation can be used up to several GHz.

In the through capacitance definition only the voltage between the screen and the surrounding world is considered. It induces charges and an electric field that couples through holes. This capacitive coupling from the outer world can be described by a capacitance that is called the feed through capacitance or shortly the through capacitance C_T, and is defined by Fig. 6 when the voltage U_1 is applied as

$$Y_C = j\,\omega\, C_T = \frac{I_F}{U_1 l} \qquad (2\text{-}2)$$

According to experience C_T is independent of frequency at least up to 1000 MHz. The capacitive coupling impedance Z_F is proportional to ωC_T (see sec. 3.2.2). Z_F is negligible for cables having no or small holes in the screen (e.g. screens with overlapping spiral tapes, with braids of high optical coverage or with multiple braids).

For electrically long cables the capacitive and inductive coupling act on the cable at the same time, therefore a separation of Z_T and Z_F is not necessary from the view-

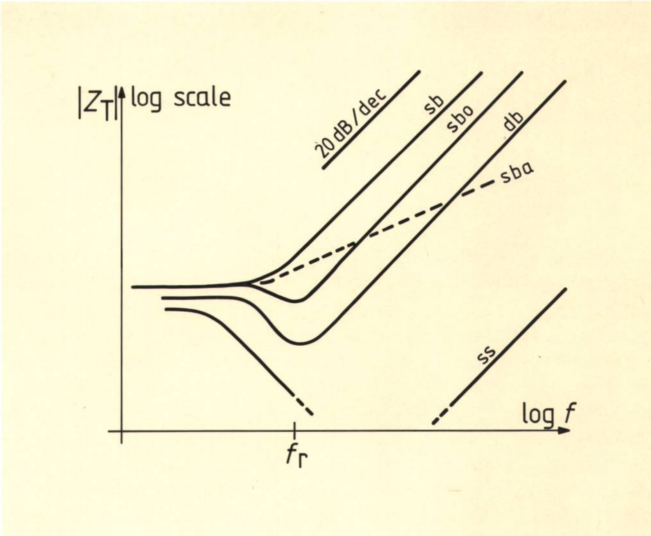

Fig. 3
Transfer impedances of typical cables

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

point of the user: either the equivalent transfer impedance Z_{TE} (that includes both effects) or the screening attenuation a_s may be an adequate measure of screening at high frequencies.

Note: A good criterion for an electrically long cable or for low frequencies may be taken as $\lambda > l/20$, where l is the length of the cable and λ is the wavelength, $\lambda \approx 300/(\sqrt{\varepsilon_r}\, f)$, where f is in MHz and λ is in metres.

3 Definitions and frequency-domain coupling-equations

31 General

Without loss of generality, the triaxial (i.e. triple coaxial) set-up with 'outer circuit' feeding is taken as model for the following sections. In this case the outer circuit is the 'primary circuit' (1) and the 'inner circuit' is the 'secondary circuit' (2), *Fig. 4*.

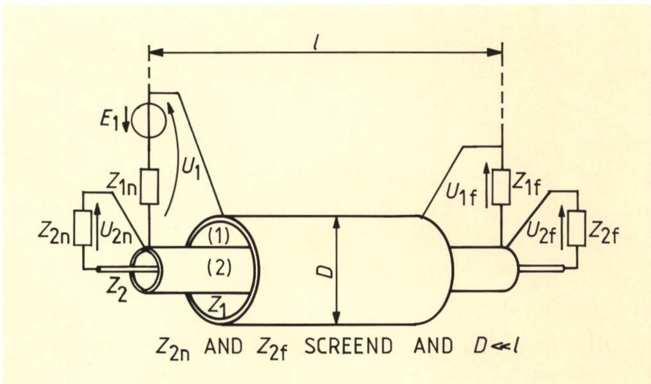

Fig. 4
Concept of the tricoaxial test set-up

Z_1, Z_2 are the characteristic impedances of the lines. In a matched set-up: $Z_{1f} = Z_1$, $Z_{2n} = Z_{2f} = Z_2$.

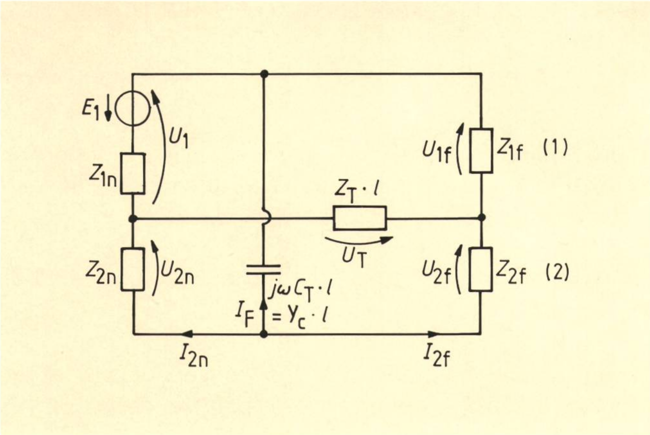

Fig. 5
Equivalent circuit of the set-up for low frequencies, i.e. for an electrically short cable ($l \ll \lambda$), with general load conditions

Note: The reference direction of voltage is chosen according to the 'normal phase' in coaxial lines. Z_T: transfer impedance per unit length (Ohm/m), C_T: through capacitance per unit length (Farad/m).

The definitions and the form of the coupling equations do not depend on the set-up (triaxial, wire injection etc.). Within the limits of the accuracy, even the measured screen parameters are independent of the set-up – except in a few differences [8].

It is not important whether the inner or the outer circuit is the primary circuit, as long as linear materials are considered. Due to reciprocity, the screen parameters and the transfer function T will be the same in both directions (outer to inner and inner to outer circuit). The coupling equations remain unchanged also for nonlinear materials, but the screen parameters Z_T, Z_F and consequently T may have different values depending on the choice of the primary circuit.

32 Definition of the transfer impedance Z_T, capacitive coupling impedance Z_F and associated parameters

321 Transfer impedance Z_T, through capacitance C_T and capacitive coupling admittance Y_C

Z_T and C_T are defined in an elementary length of cable (Fig. 2 and 6). I.e. a longitudinally uniform cable, that is electrically short, is assumed. Now we recall these definitions in terms of *Fig.* 4 and *5*:

The resistive and magnetic coupling through the screen, due to the currents, is described in terms of the transfer impedance. The transfer impedance Z_T is defined as the ratio of the longitudinal voltage per unit length in the secondary circuit, induced by the current in the primary circuit, to this current. It may be measured according to Fig. 2, using appropriate loads in the circuits of Fig. 4 and 5, i.e.:

$Z_{1f} = Z_{2f} =$ short circuit, $Z_{2n} =$ open circuit, thus

$$Z_T = (U_T/l)/I_1 = -(U_{2n}/l)/I_1 \text{ (ohm/metre).} \qquad (3.1\text{-}1)$$

Z_T is complex, see Fig. 13a!

The coupling of the electric field through holes in the screen is described in terms of the through capacitance C_T and capacitive coupling admittance Y_C. Y_C is defined as the ratio of the transversal current per unit length in the secondary system, induced by the voltage in the primary system, to this voltage. It may be measured according to *Fig. 6*, using appropriate loads in the circuits of Fig. 4 and 5, i.e.:

$Z_{1f} = Z_{2f} =$ open circuit, $Z_{2n} =$ short circuit, thus

$$Y_C = (I_F/l)/U_1 = -(I_{2n}/l)/U_1 = j\omega C_T \text{ (siemens/metre), imaginary} \qquad (3.1\text{-}2)$$

In the above equation we have implicitely defined the through capacitance $C_T = Y_c/j\omega$.

322 Invariant screen parameters, normalized through elastance K_T and capacitive coupling impedance Z_F

The statements of this section are very important, they may be deduced from textbooks like [1], [2], [3].

From the description of the coupling through a screen (sec. 2 and Annex 1) it is rather obvious that the transfer impedance Z_T is an invariant quantity with respect to the outer circuit of least for circularly symmetrical outer circuits like the tricoaxial circuit; see [8] for the discussion of exeptions. This is why Z_T shall be regarded as a pure cable screening parameter which is unaffected by the outer world. And it should be considered as the primary parameter in specifying the cable screening properties.

Unfortunately, C_T is dependent on the permittivity and geometry of the outer circuit. By introducing the new quantities K_T and Z_F, we arrive to a large degree of invariance of the capacitive coupling. The normalized through elastance K_T is defined as

$$K_T = C_T/(C_1\, C_2) \quad \text{(metre/farad)} \qquad (3.2\text{-}1a)$$

C_1 and C_2 are the capacitances per unit length of primary and secondary circuits. K_T is — to the same extent as Z_T — *invariant with respect to geometry.* Keeping the geometry constant, K_T depends on the relative permittivities of the outer and inner circuits as

$$K_T \sim 1/(\varepsilon_{r1} + \varepsilon_{r2}) \qquad (3.2\text{-}1b)$$

In order to obtain a high degree of symmetry and of transparency and to assist in the interpretation of the

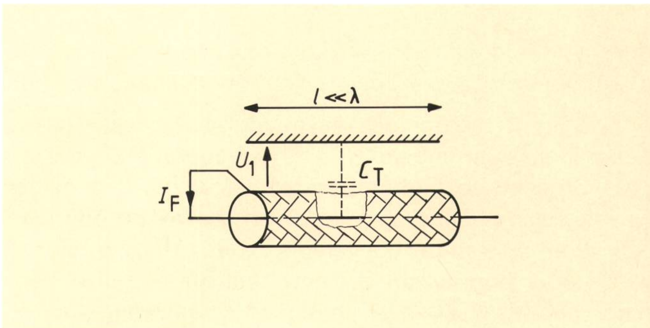

Fig. 6
Definition of the through capacitance C_T and the capacitive coupling admittance Y_C

coupling equations in sec. 3.3, the capacitive coupling impedance Z_F is introduced:

$$Z_F = Z_1 Z_2 Y_C = Z_1 Z_2 j \omega C_T = j \omega K_T/(v_1 v_2) \quad (3.2\text{-}2)$$

whereby v_1 and v_2 are the phase velocities in the primary and secondary circuit respectively.

Assuming that the phase velocity only depends on the permittivity and observing the invariance of K_T, we may conclude that also Z_F *is invariant with respect to geometry* and, for usual values of ε_r ($\varepsilon_r = 1....4$), Z_F *is nearly invariant with respect to permittivity:*

$$Z_F \sim \sqrt{\varepsilon_{r1} \varepsilon_{r2}}/(\varepsilon_{r1} + \varepsilon_{r2}) \quad (3.2\text{-}3)$$

The advantages of using C_T, K_T or Z_F are complementary: C_T is closest to the physical phenomena and is independent on frequency. K_T is a relative quantity, it is still independent on frequency and it is additionally invariant with respect to the geometry of the outer circuit. Z_F, a frequency dependent quantity, is nearly invariant to any change of the outer circuit. It shall permit a direct comparison of the capacitive coupling to the transfer impedance Z_T.

33 General coupling equations

The short or open circuit at the end of the set-up is no longer a short or open circuit along the test section if the frequency is high (i.e. the line is long). Therefore Z_T and Z_F may not be measured separately. As a first step towards high frequency definitions we replace the short and open circuits by matched loads in order to have well defined and frequency independent load conditions. In our Fig. 4, 5 and 15 we have then:

$$Z_{1n} = Z_{1f} = Z_1 \text{ and } Z_{2n} = Z_{2f} = Z_2$$

Next we still consider an electrically short set-up ($l \ll \lambda/4$) and observe in Fig. 5 the bridge-structure, the symmetry and the reasonable condition:

$$Z_T\, l \ll Z_2 \text{ and } Y_C\, l \ll 1/Z_2$$

by which we obtain for near (n) and far (f) end coupling respectively the 'low frequency' (i.e. 'short line') coupling equation of a matched circuit:

$$U_{2\substack{n\\f}}/U_1 = -(l/2)\,[(Z_F \pm Z_T)/Z_1] \quad (3.3\text{-}1)$$

Hereby and in the following, the stacked subscripts like $\substack{n\\f}$ (and later $\pm$) are associated to the stacked operation symbols $\pm$ in the obvious way: upper subscript → upper operation, lower subscript → lower operation. Now we shall proceed to higher frequencies and we consider waves. For *matched* circuits or for waves (in any load condition) it is more convenient to use the normalized voltages or normalized currents that are equal to each other and to the wave amplitude ('scattering matrix' representation). Note that the square of these quantities is equal to the power in the matched load or in the wave [4].

The normalized voltages and currents are:

$$U_1/\sqrt{Z_1} = I_1 \sqrt{Z_1}, \quad U_2/\sqrt{Z_2} = I_2 \sqrt{Z_2} \quad (3.3\text{-}2)$$

The coupling transfer function (or coupling function) is defined in terms of the normalized quantities. For *'low frequencies'*:

$$T_{\substack{n\\f}} = (U_{2\substack{n\\f}}/\sqrt{Z_2})/(U_1/\sqrt{Z_1}) = -(l/2)[(Z_F \pm Z_T)/Z_{12}] \quad (3.3\text{-}3)$$

Whereby $Z_{12} = \sqrt{Z_1 Z_2}$, and the conversions to other ratios, e.g. to U_{2n}/I_1 is made by using the definitions in Eq. 3.3-2.

We cannot separately measure Z_T and Z_F, but we could still separate them mathematically from near and far end measurements. This separation is usually not necessary from the point of view of the user. It is more convenient to introduce a new quantity called equivalent transfer impedance:

$$Z_{TE} = \max |\, Z_F \pm Z_T \,| \quad (3.3\text{-}4)$$

According to the reciprocity law, not the quantities U_2/U_1 or I_2/I_1 but the coupling transfer function T is the invariant quantitiy with respect to feeding direction, i.e., for linear materials:

$$T_{\substack{n\\f}} \text{ (outer to inner circuit)} = T_{\substack{n\\f}} \text{ (inner to outer circuit)}$$

For short cables there is no change of phase along the cable. If long cables i.e. high frequencies are considered, the coupling contributions along the line must be summed up in the correct phase. This summation (integration) leads to the following result in a matched system (see Annex 2: Coupling equations):

$$T_{\substack{n\\f}} = \frac{-(Z_F \pm Z_T)}{Z_{12}}\, \frac{l}{2}\, S_{\substack{n\\f}}(lf) \quad (3.3\text{-}5)$$

Whereby:

$$T_n = \frac{\underline{U}_{2n}/\sqrt{Z_2}}{\underline{U}_1/\sqrt{Z_1}} = \frac{\underline{I}_{2n}\sqrt{Z_2}}{\underline{I}_1\sqrt{Z_1}}$$

$$T_f = \frac{\underline{U}_{2f}/\sqrt{Z_2}}{\underline{U}_1/\sqrt{Z_1}} = \frac{\underline{I}_{2f}\sqrt{Z_2}}{\underline{I}_1\sqrt{Z_1}}$$

T: (coupling) transfer function of the normalized wave amplitudes for near and far end coupling respectively, with following notation:

$\underline{U}_1$: voltage wave in the primary circuit at the near end, travelling to the far end. (Due to matched circuits, and to the assumption that the total recoupled wave from the secondary to primary circuit is negligible, there is no wave in the primary circuit travelling to the near end. This assumption of $T_{\substack{n\\f}} \ll 1$ is true for practical RF-cables.)

$\underline{U}_{2n}$: voltage wave in the secondary circuit at the near end, travelling to the near end

$\underline{U}_{2f}$: voltage wave in the seconadry circuit at the far end, travelling to the far end.

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

$$S_{\substack{n\\f}}(lf) = \frac{\sinh \frac{\Gamma\pm}{2}}{\frac{\Gamma\pm}{2}} \exp - \frac{\Gamma+}{2} \qquad (3.3\text{-}6a)$$

= summing function for the general lossy network with

$$\Gamma\pm = (\gamma_2 \pm \gamma_1)\, l = [(\alpha_2 \pm \alpha_1) + j\,(\beta_2 \pm \beta_1)]\, l$$
$$= [(\alpha_2 \pm \alpha_1) + 2\,\pi\, f\,(1/\nu_2 \pm 1/\nu_1)]\, l$$

subscript $\pm$ refers to near/far end respectively
$+$ refers to both near/far ends

Γ: image transmission factor or propagation factor $\Gamma = \alpha\, l + j\, \beta\, l$
β_1, β_2: phase constants (rad/metre)
ν_1, ν_2: phase velocities (metre/second)
α_1, α_2: attenuation constants (Neper/metre).

In most practical situations the losses are negligible: $\exp(\alpha_1 l) \approx \exp(\alpha_2 l) \approx 1$. Eq. (3.3-6a) is then reduced to

$$S_{\substack{n\\f}}(lf) = \frac{\sin \frac{\beta\, l_\pm}{2}}{\frac{\beta\, l_\pm}{2}} \exp - j\, \frac{\beta\, l_+}{2} \qquad (3.3\text{-}6b)$$

with $\beta\, l_\pm = (\beta_2 \pm \beta_1)\, l = 2\,\pi\, fl\,(1/\nu_2 \pm 1/\nu_1)$.

The results in Eq. (3.3-5) are for matched conditions. If e.g. there is a mismatch at the far end of circuit (1) and the reflection factor between Z_1 and Z_{1f} is ρ we get the following coupling equations T_{np} for near end and T_{fp} for far end

$$T_{np} = T_n + \rho T_f \exp(-\gamma_1 l)$$
$$T_{fp} = T_f + \rho T_n \exp(-\gamma_1 l) \qquad (3.3\text{-}7)$$

From these equations we see the large effect the mismatch can have on the measuring results:

If the near and far end couplings (T_n and T_f) have significantly different magnitudes, then the test results may become very erroneous, due to the undesired superposition of ρT_n or ρT_f to the couplings T_f ot T_n.

34 Interpretation of the coupling equations

We have arrived to the crux of our investigation: We have obtained the coupling transfer function T. Now we want to see the effect of the various cable and set-up parameters on the transfer function we measure or we even must accept in real cable installations.

In the following we shall always neglect losses, i.e. Eq. (3.3-6b) is used.

341 Matched circuits

A directional effect i.e. a difference in forward and backward coupling may be caused at any frequency due to the fact that $Z_T + Z_F$ and $Z_T - Z_F$ appear in Eq. (3.3-5) for near and far end couplings respectively. Further we may see in the Annex 1 'Transfer impedance of a braided-wire outer conductor', that Z_T may have any arbitrary phase in the reactive coupling region ($f > f_r$ in Fig. 13), thus either the near end ot the far end coupling may be larger. But note that on cables with vanishing capacitive coupling no directive effects are present from DC up to the cut-off frequency f_{cn} (see below).

342 The summing function $S(lf)$

The only difference between low and high frequency coupling is represented by the summing function $S(lf)$. Note that its amplitude is a $(\sin\varphi)/\varphi$ function. *Fig. 7* illustrates the low pass behaviour of S having periodic extreme for $f > f_c$. S being a function of (lf), there is a cut-off value $(lf)_c$, corresponding to $\varphi = 1$ rad. From $(lf)_c$ either a cut-off frequency f_c is derived for a fixed length l or a cut-off length l_c is derived for a fixed frequency f:

$$f_{\substack{cn\\f}} = \nu_\pm/(\pi l),\quad l_{\substack{cn\\f}} = \nu_\pm/(\pi f) \qquad (3.4\text{-}1)$$

$$\nu_\pm = 1/\,|\,1/\nu_2 \pm 1/\nu_1\,| = c/\,|\,1/\nu_{r2} \pm 1/\nu_{r1}\,|$$

where $\nu_r = \nu/c$, c: velocity in vacuum.

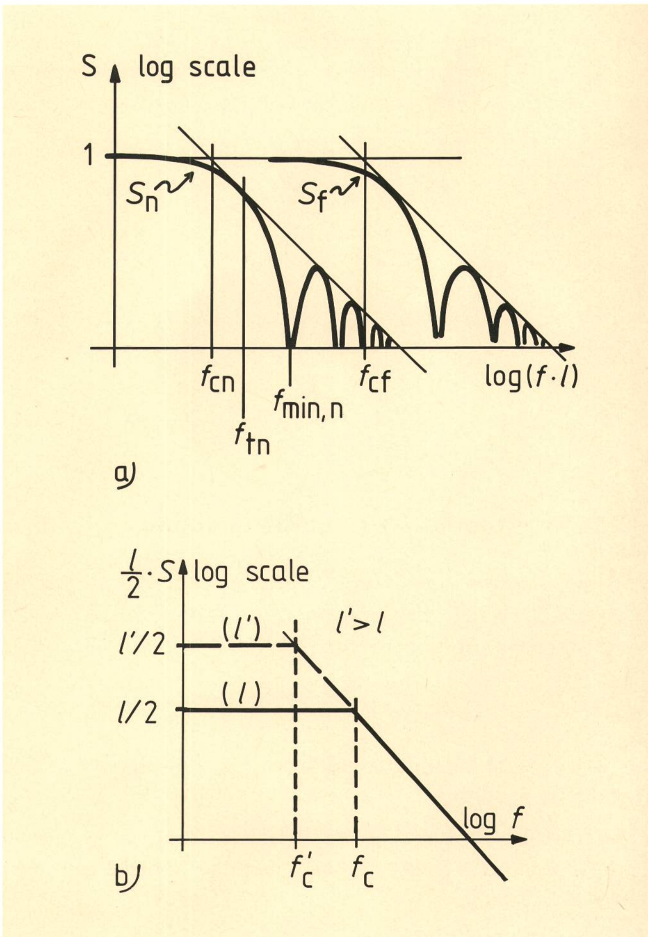

Fig. 7
a) The summing function $S(lf)$ for near and far end couplings
b) The envelope of the product $(l/2)\, S$ as function of frequency, with l as parameter. f_c: cut-off frequency

$f_c = \nu_\pm/(\pi l)$ $\quad$ $\nu_\pm = c/\,|\,1/\nu_{r2} \pm 1/\nu_{r1}\,|$
$f_t = \nu_\pm/(2l)$ $\quad$ $\pm$ *sign refer to n/f*
$f_{min} = \nu_\pm/l$

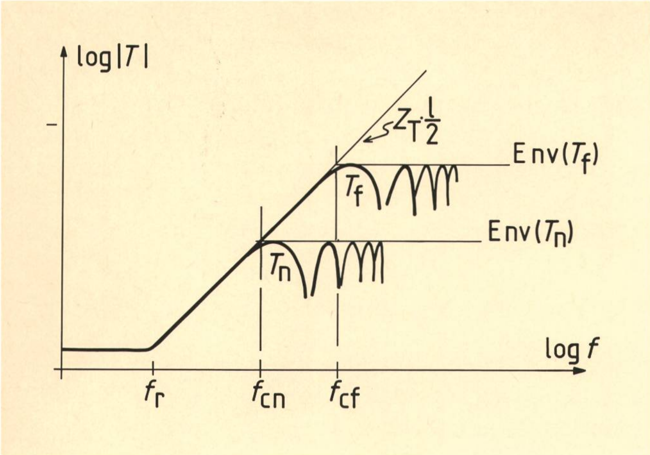

Fig. 8
Coupling transfer function for a typical cable without capacitive coupling ($Z_F = 0$), Z_T according to cable sb in Fig. 3. The minima and maxima are periodic on a linear scale
Example: $v_{r1} = 1$ (set-up), $v_{r2} = 1/1.5$ (cable), $l = 1$ m,
$\rightarrow f_{cn} = 40$ MHz, $f_{cf} = 200$ MHz.

If it is assumed $v = c/\sqrt{\varepsilon_r}$:

$$f_{\substack{cn\\f}} = (c/\mid \sqrt{\varepsilon_{r2}} \pm \sqrt{\varepsilon_{r1}} \mid)/(\pi l) \qquad (3.4\text{-}2)$$

with according expression for $l_{\substack{cn\\f}}$. Note that near and far end couplings have different cut-off values. Denoting by Env ( ) the envelope of a function, the mathematical expression for the envelope of the summing function with f as variable is:

$$\begin{aligned} f<f_c&: \text{Env}(\mid S \mid) = 1 \\ f>f_c&: \text{Env}(\mid S \mid) = f_c/f \end{aligned} \qquad (3.4\text{-}3)$$

Admitting an error less than factor 2, the envelope is approximated by

$$\text{Env}_{\substack{n\\f}}(\mid S \mid) = \frac{1}{1+f/f_c} = \frac{1}{1+\pi f l/v_\pm} \qquad (3.4\text{-}4)$$

343 The effect of $S\,(lf)$ on the coupling

The most important effect of the summing function is that for $(lf) > (lf)_c$ the screen parameters *must not be calculated* by the simple formula:

$$\mid Z_F \pm Z_T \mid = \mid T_{\substack{n\\f}} \mid Z_{12}/(l/2)$$

but they must be extracted from Eq. (3.3-5). There are 3 ways to proceed:

1) Exact formula for S is used: The problem is the large error when $f \gg f_c$ and v_1 and v_2 are known with limited accuracy only.
2) Only the envelope Env (| T |), i.e. Env (| S |), is used and measurement is made up to $f \gg f_c$: This gives appropriate information, because $Z_F \pm Z_T$ is a smooth function of frequency. We obtain:

for $f<f_c$: $$\mid T_{\substack{n\\f}} \mid = (\mid Z_F \pm Z_T \mid /Z_{12})\,(l/2) \qquad (3.4\text{-}5a)$$

$f>f_c$:
$$\text{Env}(\mid T_{\substack{n\\f}} \mid) = (\mid Z_F \pm Z_T \mid /Z_{12})\,(l/2)\,(f_c/f) \qquad (3.4\text{-}5b)$$

or with the approximation (3.4-4):

$$\text{Env}(\mid T_{\substack{n\\f}} \mid) \approx \frac{\mid Z_F \pm Z_T \mid (l/2)}{Z_{12}\,(1+f/f_{\substack{cn\\f}})} = \frac{\mid Z_F \pm Z_T \mid (l/2)}{Z_{12}\,(1+\pi f l/v_\pm)} \qquad (3.4\text{-}6a)$$

specially for $(lf) > (lf)_c$

$$\text{Env}(\mid T_{\substack{n\\f}} \mid) = \frac{\mid Z_F \pm Z_T \mid}{Z_{12}} \frac{v_\pm}{\omega} \qquad (3.4\text{-}6b)$$

3) Phase equalization: by special measures v_1 is made $v_1 \approx v_2$, i.e. $\sqrt{\varepsilon_{r1}} \approx \sqrt{\varepsilon_{r2}}$, by which: f_{cf} is shifted to much higher frequencies. Note that f_{cn} is not significantly changed. This procedure is only useful if $\mid Z_F \ll Z_T \mid$ or $\mid Z_F + Z_T \mid \ll \mid Z_F - Z_T \mid$, i.e. if T_n is not of interest.

In order to simplify the discussion, we consider in the rest of this section 3.4.3 only cables with negligible through capacitance (see Sec. 3.4.4 if C_T is not negligible). Three effects should be mentioned: directivity, length-effect and permittivity effect.

If $C_T = 0$, then $Z_F = 0$ and the coupling transfer function is

$$T_{\substack{n\\f}} = \mp (l/2)\,(Z_T/Z_{12})\,S\,(lf)$$

Although the directive effect of Z_F is not present (see Sec. 3.4.1), we note a directive effect at high frequencies, when $f > f_{cn}$ i.e. $T_n \neq T_f$, see *Fig. 8*. This directive effect is due to the different cut-off frequencies $f_{cn} < f_{cf}$.

The effect of cable length is illustrated in *Fig.* 7b and *9*. Fig. 9 is obtained by the addition of the log $|\underline{Z}_T|$ and log $|(l/2)S|$ functions of Figs. 3 and 7. For a fixed frequency there is a cut-off length l_c. If $l > l_c$, the coupling is inde-

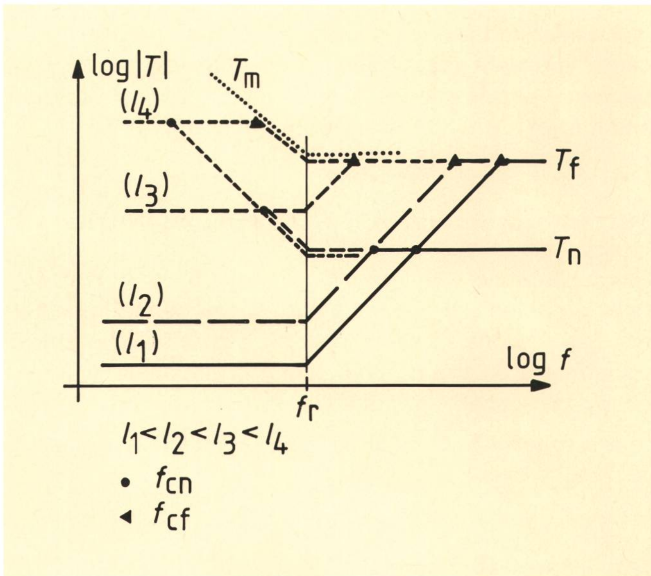

Fig. 9
Effect of the cable length on coupling. Same cable as in Fig. 8
$T(f)$ for different cable lengths. Only the envelope of the coupling is drawn. T_m: maximum attainable coupling ($l \geq l_c$).

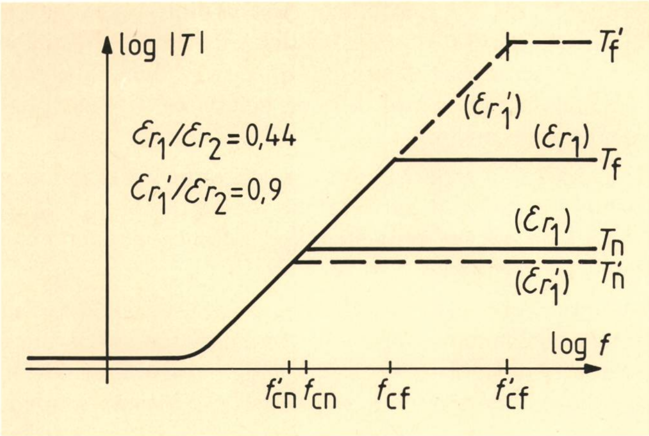

Fig. 10
Effect of the permittivity in the set-up on the coupling. Only the envelope is drawn. Reference cable: same parameters as in Fig. 8
Example: $\varepsilon'_{r1} = 0.9 \cdot \varepsilon_{r2} \rightarrow v_{r1}/v_{r2} = 1.05 \rightarrow f'_{cf} = 6.7 \cdot f_{cf}$,
for $f > f'_{cf}$: Env (T')/Env (T) = 6.7 = 16 dB.

pendent of the length. I.e. there is a maximum coupling T_m that can not be exceeded by increasing the cable length.

For short cables i.e. below cut-off length (or frequency) the coupling transfer function is proportional to the cable length and follows the frequency response of Z_T.

The effect of the phase velocity may be usually reduced to the effect of the permittivity, because most often $v = c/\sqrt{\varepsilon_r}$. The permittivity effect is illustrated in *Fig. 10.* The shift in the cut-off frequency (for a fixed l) and consequently in the height of the envelope at frequencies $f > f_{cf}$ is related to the change of the permittivity ε_{r1} in the outer circuit as follows [by prime (') we denote the new quantities]:

$$f'_{cf} = f_{cf} \mid 1-\sqrt{\varepsilon_{rl}/\varepsilon_{r2}} \mid / \mid 1-\sqrt{\varepsilon'_{rl}/\varepsilon_{r2}} \mid \qquad (3.4\text{-}7)$$

From the equation it is obvious that as ε_{r1} tends toward ε_{r2}, the far end cut-off length l_{cf} (for a fixed frequency) tends toward infinity and the coupling increases without limits proportionally to the cable length. However, $|T| \leq 1$ for physical reasons, see Annex 2.

344 Special effects on cables with nonvanishing Z_F ($Z_F \ll Z_T$ is not fullfilled)

The effect of the summing function S on the coupling to cables that have a through capacitance is the same as for cables without this capacitance, i.e. $Z_F = 0$, as discussed above. But for cables with nonvanishing Z_F, due to the expression $(Z_F \pm Z_T)$ in Eq. (3.3-5), additional directive effects are present at any frequency. The resulting coupling functions are more complex. The interpretation of test data (coupling transfer function) or the application of cable data hides more pitfalls.

Following features are important:

- Change of the permittivity of the set-up: Additionally to the directive effect of Z_F and to the aspects in Fig. 10, also C_T, K_T and very slightly Z_F will be changed, when ε_{r1} is changed! (see sec. 3.2.2).
- Variable dominance of T_f over T_n: Whether $T_f > T_n$, may depend on the permittivity (ε_{r1}, see above), on the frequency band and even on the cable length. The last fact is shown in *Fig. 11*. It was obtained from careful inspection of the coupling equations.

345 A special problem: Cables with multiple isolated screens

The above coupling equations are not directly applicable to multilayer screens that have low loss isolation between the different low loss metallic layers. Each pair of metallic layers with isolation represents a transmission line. The resulting coupling is the chain of the couplings from all these transmission lines. Different permittivities in the layers, short circuit at the end of the layers yield a summing function that is a more complicated function of l than Eq. (3.3-5 and -6). Simple extrapolation of the test data to other cable lengths than the test length is not possible. When testing such cables, it is recommended that at least the secondary circuit has the same length as used in practice.

35 Definition of the screening attenuation a_s

According to the present definitions [IEC document SC 46A(C.O.)113, March 86; intended for IEC publication 96-1] the screening attenuation a_s is measured with the long wire antenna set-up (absorbing clamp) and is specified for 30 MHz < f < 1000MHz.

The value of a_s is defined in above mentioned IEC document as the ratio of power in a matched primary circuit to the maximum ot total power in the secondary circuit, measured or extrapolated beyond the cut-off frequency ($f > f_c$) or cut-off length ($l > l_c$):

$$a_s = -10 \lg \left[(P_{n,max}/P_1 + P_{f,max}/P_1) \right] \qquad (3.5\text{-}1)$$

whereby:

$$P_{\substack{n,\\ f}\,max}/P_1 = [\text{Env}\,(T_{\substack{n\\ f}})]^2 = \left(\frac{\mid Z_F \pm Z_T \mid v_\pm}{\mid Z_{12} \mid \omega} \right)^2$$

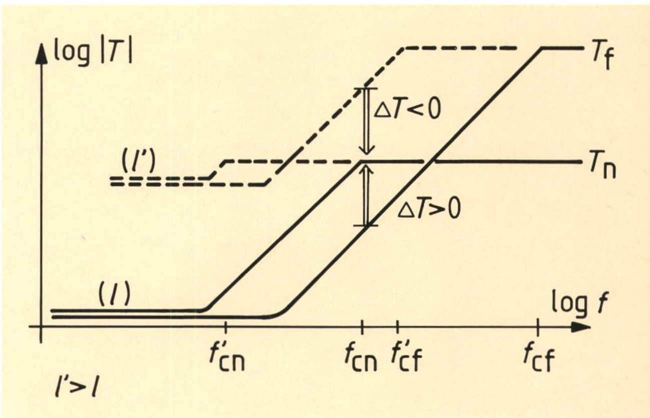

Fig. 11
Dependence of the couplings T_n and T_f on the length for a cable with following properties:
$\mid Z_T + Z_F \mid > \mid Z_T - Z_F \mid$, for $f < f_{cn}$: $T_n > T_f$
$f > f_{cf}$: $T_n < T_f$

The conversion $a_s \rightarrow |Z_T \pm Z_F|$ is ambiguous if only a_s is known but $P_{n,max}/P_1$ and $P_{f,max}/P_1$ are not known. But for negligible Z_F and $v_1 \neq v_2$ is valid:

$$| Z_T | = Z_{12}\, \omega \left| \frac{1}{v_2} - \frac{1}{v_1} \right| 10^{-(a_s/20)} \qquad (3.5\text{-}2)$$

where Z_1 in $Z_{12} = \sqrt{Z_1 Z_2}$ is taken from (3.5-3) and $v_1 \approx 3 \cdot 10^8$ m/s.

Although the test set-up and the 'true world' are governed by the same coupling equations, the maximum coupling in true world may be significantly different from the reciprocal value of a_s. The reason for it is clear: T_f is proportional to $1/|\sqrt{\varepsilon_{r2}} - \sqrt{\varepsilon_{r1}}|$ which means that for $\varepsilon_{r2} \approx \varepsilon_{r1}$ T_f is large and varying largely when the permittivity of the outer world is changing, see Fig. 11. For these reasons Z_T, Z_F and Z_{TE} are preferred to be regarded as primary set-up independent screening parameters of cables.

There is a further problem: the above definition may be in conflict with definitions proposed for connectors in the IEC (and also being reasonable for cables), because:

- the above definition of 'a_s' uses $(P_{n,max} + P_{f,max})$, while the 'connector-definition' uses max $(P_{n,max}, P_{f,max})$
- Z_1 of the long wire antenna set-up is different from Z_1 of connector set-ups.

The characteristic impedance of the long wire outer circuit in the a_s measurement of cables is:

$$Z_1 \approx 60\, \{\ln[\lambda_0/(\pi\, d)] + 0.12\}\ \text{Ohm} \qquad (3.5\text{-}3)$$

d: outer diameter of the screen

All these problems with a_s might be avoided by introducing a new definition of a_s:

$$a_s = -20\ \lg\, \{\max\, [\text{Env}\, (T'_n),\ \text{Env}\, (T'_f)]\} \qquad (3.5\text{-}4)$$

whereby T'_n and T'_f are referred to standardized values by calculation. E.g.
$Z'_1 = 150\ \Omega$, $\Delta v'_r = |v_{r2} - v'_{r1}|/v_{r2} = 0.1$. The condition $l > l_c$ shall be valid.

The screening attenuation has following merits as screening quantity:

- it directly indicates the power transmission through a long screen by a single dB value
- this dB value is practically constant over the specified range (30...1000 MHz) because of the form of Eq. (3.4-6b) and because of the usual behaviour
  $|Z_T \pm Z_F| \sim \omega$, for $f > 30$ MHz. Note that Z_1 in Eq. (3.5-3) is only weakly dependent on frequency.

4 Conclusions

The intrinsic screening properties of cylindrical screens can be described by the transfer impedance Z_T and by the capacitive coupling impedance Z_F. By intrinsic we mean that these properties are not or only slightly dependent on the environment. For practical reasons instead of Z_F often the associated parameters are given like through capacitance C_T, capacitive coupling admittance Y_C or normalized through elastance K_T, which depend on the environment.

The most appropriate quantity to describe the capacitive leakage is Z_F because it has the same dimension as Z_T and the amount of capacitive and inductive coupling can easily be compared.

At high frequencies or on long cables ($l > \lambda/20$), Z_T and Z_F act always simultaneously on the cable, their separation is not reasonable. Therefore it is more expedient to specify one quantity, i.e. the equivalent transfer impedance $Z_{TE} = \max |Z_F \pm Z_T|$.

The coupling through cylindrical screens is further characterized by the screen-independent cut-off parameter $(lf)_c$ = constant. Below cut-off the coupling is proportional to $l|Z_F \pm Z_T|$. Beyond cut-off the maximum coupling is independent of cable length and is proportional to $|Z_F \pm Z_T|/f$.

Above the cut-off frequency f_c and cut-off length l_c and in practice above 30 MHz, screening attenuation a_s can be used. It must be remembered that a_s is very much affected by the permittivity of the outer system when the permittivities of the two systems are nearly equal. For this reason Z_T, Z_F and/or Z_{TE} should be considered as primary quantities in specifying the screening effectiveness of coaxial cables, while the screening attenuation a_s — a 'user friendly' quantity — should be given in rated conditions and mainly for guidance.

Annex 1: Transfer impedance of a braided-wire outer conductor

Let us study more closely the transfer impedance of a braided-wire screen construction. An electrically short piece of braided coaxial cable is considered to be placed in a triaxial arrangement like in Fig. 4. 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 according to Fig. 1a.

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, according to Fig. 1a and induces 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, Fig. 12a.

Therefore it is necessary and unavoidable that Φ''_{12} is partly also in the inner circuit, Fig. 12b. 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 $\Phi''_{12}/2$. The voltages are in parallel:

$$U_{rh} = U_{lh} = U''_2 = -\frac{1}{2}\, j\, \omega\, \Phi''_{12} \qquad (A1\text{-}1)$$

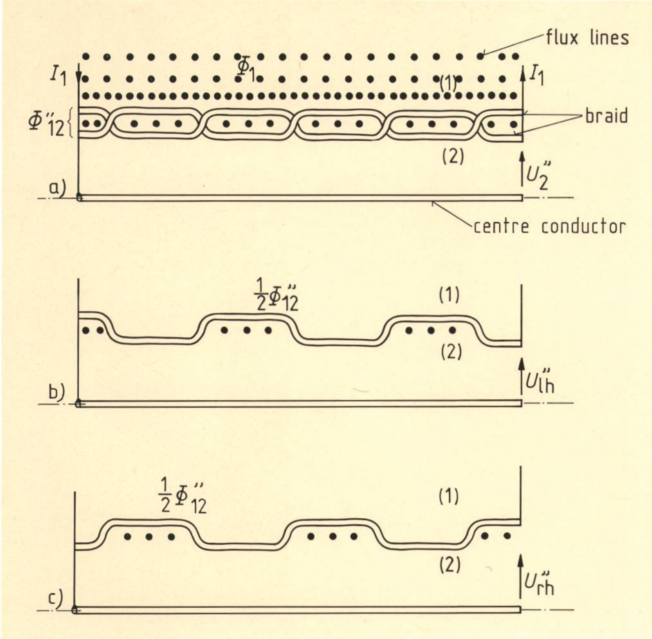

Fig. 12
Magnetic coupling in the braid (one half of the longitudinal section of a coaxial line is shown), a) complete flux, b) left hand lay contribution, c) right hand lay contribution

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

The two induced disturbing voltages oppose each other. This can also be roughly concluded from the directions of the leaking magnetic fluxes Φ'_{12} and Φ''_{12} and magnetic field-lines in *Fig.* 1a and *12.*

On this important physical fact is based the braid optimization. The both leakage phenomena can be described by mutual inductances:

$$M'_{12} = \frac{\Phi'_{12}}{I_1} \tag{A1-2}$$

$$M''_{12} = \frac{1}{2}\frac{\Phi''_{12}}{I_1} \tag{A1-3}$$

It is obviously possible to make braided-wire screens where either M'_{12} or M''_{12} are dominant or where they are cancelling each other. We are correspondingly speaking of underbraided, overbraided or optimized braids. In *Fig. 13a* are shown measured transfer impedances in the complex plane of such screens. 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 roughly like a homogeneous tubular screen. The same can be concluded from *Fig. 13b* where the transfer impedance amplitudes are shown as a function of frequency. But from *Fig. 13c* cannot be directly said if the screen is underbraided or overbraided.

The transfer impedance of a braided-wire screen consists of three above mentioned main components:

(i) At low and medium frequencies the tubular screen coupling behaviour (Z_{Th}) with eddy currents and decreasing Z_T. In [2] it is stated that a good approximation for Z_{Th} is a tubular homogeneous screen with the thickness of one wire diameter and the same DC resistance as the braid.

(ii) The mutual inductance M'_{12} related to direct leakage of the magnetic flux Φ'_{12}.

(iii) The mutual inductance M''_{12} (opposed polarity) related to the magnetic flux Φ''_{12} in the braid.

By adding these components we get a good approximation for the transfer impedance Z_T of a braided-wire screen

$$Z_T \approx Z_{Th} + j\,\omega\,(M'_{12} - M''_{12}) \tag{A1-4}$$

and the first approximation of the equivalent circuit is shown in *Fig. 14a.*

A more complete equivalent circuit where also the through capacitance C_T and surface impedances Z_a of the braided cable are incorporated is shown in *Fig. 14b.* L_1 and L_2 are the (external) inductances of the outer and inner circuit, whereas Z_a contain the self resistances and inductances of the conductors.

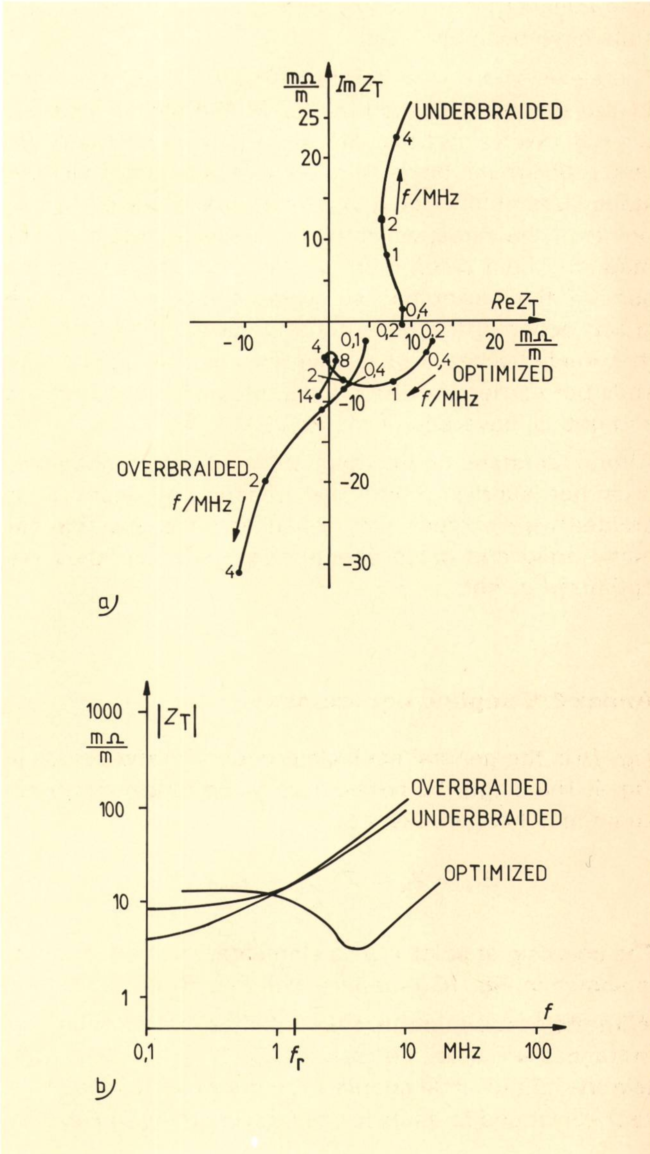

Fig. 13
Measured transfer impedances, a) complex plane, b) amplitude

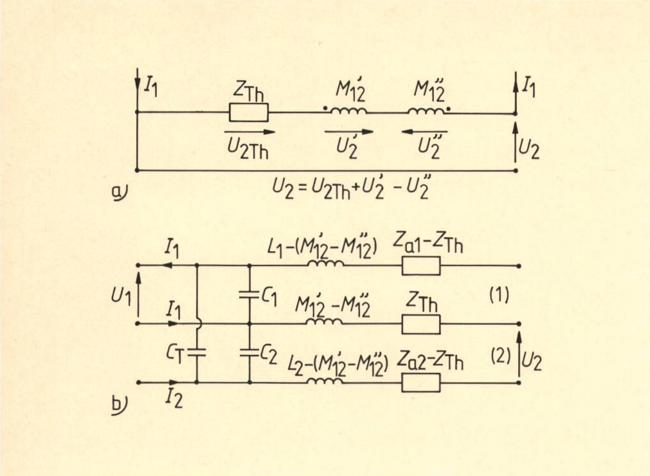

Fig. 14
Equivalent circuits of a braided-wire screen, a) contributions to the transfer impedance, b) significant elements of circuits (1) and (2)

Many attempts have been made to calculate the transfer impedance of a braided coaxial cable. Most of the literature { e.g. [7], [3], [2] } 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 { [5], [6] } on M''_{12} but the matter has been studied by IEC SC46A/WG1. Specially the calculation and stability of M''_{12} have shown to be very problematic because of so many uncertain and unstable parameters as e.g. the resistance of the crossover points of the wires, which certainly has an effect on the magnetic field distribution in the braid. Also the pressure of the jacket has surely 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 thightness and optical coverage of the braid.

After understanding the magnetic coupling mechanisms it is not suprising that the transfer impedances of braided-wire screens vary much and are unstable for many braid and cable constructions whether they are optimized or not.

Annex 2: Coupling equations

Fig. 15 is the general equivalent circuit for the set-up in Fig. 4. The length *l* is not limited. In the following we assume matched circuits, i.e.

$$Z_{1n} = Z_{1f} = Z_1, \; Z_{2n} = Z_{2f} = Z_2$$

The coupling at point *x* in an elementary short section d*x* is shown in *Fig. 16* (compare with Fig. 5).

With the assumption that the infinitesimal coupling immittances are small i.e. $Z_T\,dx \ll 2Z_2$, $Y_C\,dx \ll 2/Z_2$, we obtain the infinitesimal coupled voltage waves which travel to the near and far ends in line (2), according to *Fig. 16b*

$$dU_2\,(x) = -\frac{1}{2}\,(dU_F + dU_T) \qquad \text{(A2-1)}$$

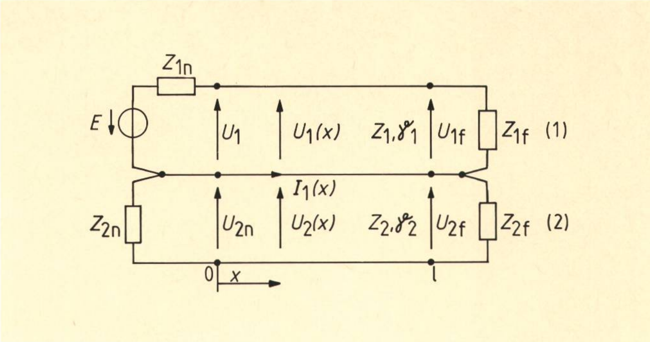

Fig. 15
Voltage and current definitions

$$dU_2\,(x+dx) = -\frac{1}{2}\,(dU_F - dU_T) \qquad \text{(A2-2)}$$

Inserting Z_T and $Z_F = Z_1\,Z_2\,Y_c$ and observing that $I_1\,(x) = U_1\,(x)/Z_1$:

$$dU_2\,(x) = -\frac{U_1\,(x)}{Z_1}\,\frac{1}{2}\,(Z_F + Z_T)\,dx \qquad \text{(A2-3)}$$

$$dU_2\,(x+dx) = -\frac{U_1\,(x)}{Z_1}\,\frac{1}{2}\,(Z_F - Z_T)\,dx \qquad \text{(A2-4)}$$

Taking into account the transmission parameters of line (1) and (2) we get at the near end and at the far end of the line (2) the infinitesimal coupling contributions from the line element d*x* that is located at *x*:

$$dU_{2n} = -\frac{U_1}{Z_1}\,\frac{1}{2}\,(Z_F + Z_T)\,dx\,\exp\,[-(\gamma_1 + \gamma_2)x] \qquad \text{(A2-5)}$$

$$dU_{2f} = -\frac{U_1}{Z_1}\,\frac{1}{2}\,(Z_F - Z_T)\,dx\,\exp\,(-\gamma_1\,x)\,\exp\,[-\gamma_2\,(l-x)] \qquad \text{(A2-6)}$$

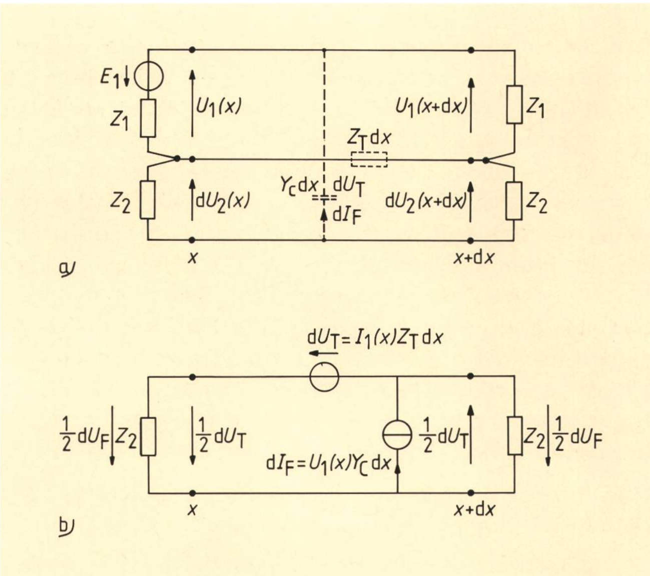

Fig. 16
Coupling of an elementary short section dx, a) total circuit, b) circuit (2).

For practical RF cables we may neglect the recoupling from circuit (2) to circuit (1) and the coupling losses in circuit 1. I.e. $U_1(x)$ is not influenced by the coupling (Note: The screening attenuation a_s attains many tens of dB). Thus, by straight forward integration and use of the definition in Eq. (3.3-3) we get the coupling transfer functions (3.3-5). The above approximations are valid for $|T| \ll 1$, else the rigorous three conductor line analysis must be used.

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Glossary

IEC	International Electrotechnical Commission
IEC SC46A	TC46 Cables, Wires and Waveguides for Telecommunication Equipment Sub-Committee SC46A Radio Frequency Cables
SC 46A/WG1	WG on Screening Effectiveness
TC	Technical Committee
WG	Working Group
RF	Radio Frequency

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