SCREENING OF CABLES IN THE MHZ TO GHZ FREQUENCY RANGE EXTENDED APPLICATION OF A SIMPLE MEASURING METHOD

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Summary:

Although cables are passive components, which can't disturb the circumference by its own, their electromagnetic compatibility (EMC) is as important as the EMC of active devices, which are connected with those cables.

The increasing use of cables, especially in the field of information technologies and cable distributed television systems up to very high frequencies requires new measuring methods for determination of the screening effectiveness of those cables. This methods shall be cheap, easy to use, applicable up to the GHz range and with good reproducibility.

An investigation shows, that the well known triaxial measuring method with a one-sided short circuit commonly used for lower frequencies retains its merits also in the GHz range. The extension of this triaxial method up to higher frequencies was first investigated in 1993 in Germany [1]. After two years of discussion and improvement in the international working group IEC TC46/WG5 it became a Committee Draft, IEC 46A/320/CDV, „shielded screening attenuation test method" [2] [3].

1. Introduction

In many cases, above all in the lower frequency range, the screening effectiveness of cables is described by the transfer impedance Z_T. It is, for an electrically short peace of cable, defined as the quotient of the longitudinal voltage measured on the secondary side of the screen to the current in the screen, caused by a primary inducing circuit, related to unit length [4]. Although the transfer impedance Z_T covers only the galvanic and magnetic couplings it is common practice to use it also as a quantity which includes the effect of the coupling capacitance C_T through the cable screen [5]. In this case it is named equivalent transfer impedance Z_{TE} which includes the effects of galvanic, magnetic and capacitive coupling.

For the determination of the proper coupling capacitance there is, as standardised quantity, the capacitance coupling admittance Y_T. The coupling admittance, for an electrically short peace of cable, is defined as the quotient of the current in the screen caused by the capacitive coupling in the secondary circuit to the voltage in the primary circuit related to unit length [4].

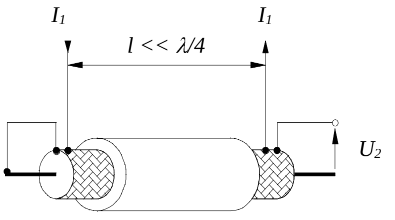


Fig. 1: Definition of transfer impedance

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

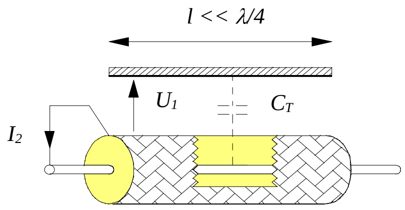


Fig. 2: Definition of coupling admittance

$$Y_T = \frac{I_2}{U_1 \cdot l} = j\omega \cdot C_T \qquad (1\text{-}2)$$

With electrically short cables, where wave propagation can be neglected, the screening quantities related to unit length can directly be used to calculate an induced disturbing voltage. In the higher frequency range the implications get similar complicated as the transmission characteristics of a simple line, dependent on the impedance and admittance per unit length as well as on the terminating resistors.
For the measurement of the transfer impedance two methods are standardised, the well known triaxial measuring method according to IEC 61196-1, clause 12.2 and the line injection method according to IEC 61196-1 clause 12.1. The triaxial method is limited to a frequency range in which the coupling length is short (less than approximately one sixth of the wave length). For example the upper frequency limit is appr. 30 MHz for a coupling length of 1m. The line injection method is applicable in a frequency range up to 3 GHz [6]. This procedure requires that the characteristic impedance and propagation velocity of the induced loop are matched to the cable under test. This complicates its use up to the GHz range considerably.
The measuring methods for the capacitance coupling admittance according to IEC 61196-1, clause 12.3 are limited to frequencies below 1 MHz. According to experience the coupling capacitance C_T is independent of frequency, at least up to 1000 MHz [5]. Therefore it is common practice to convert the measured results of Y_T proportionally to high frequencies.

2. Objectives

It is desirable to measure and evaluate the screening efficiency of cable screens also in the wave propagation frequency range such that its characteristics can be directly applied. This requires to look closer at the conditions of such applications.
In general, a system of electromagnetic induction consists of a transmission circuit in the cable, which is assumed to be fully defined, and of a surrounding transmission system, which is assumed to be universal with respect to the definition of cable screening. The screening effectiveness may be universally described by the maximum power output into the surroundings of the cable related to the power propagating in the cable. The power ratio is best expressed logarithmically as screening attenuation.
An often used procedure to determine the screening attenuation is the well known „absorbing clamp method“ according IEC 61196-1 clause 12.4. The drawback of this method is, that its set-up requires relatively much space, does not exclude environmental effects - unless the measuring area is enclosed in a shielded cabin -, and that the available absorbing clamp transformers considerably limit the measurement sensitivity.
It suggests itself to limit the free space such that the said problems don’t occur but wave propagation near the cable surface is not significantly changed. A triaxial measuring set-up is the solution. It has a one-sided short circuit between the metal tube and the cable screen. Power is fed into the terminated inner circuit of the cable and the disturbing power is measured at the opposite end of the outer circuit.

3. Theory of the triaxial measuring method

On the basis of the known reversibility of primary and secondary measuring circuits, the proposed measuring set-up, presented in Fig. 3, actually meets the IEC standard despite of the interchange of generator and receiver. The benefits of feeding the inner system, which is terminated by its characteristic impedance, are the matching of the generator and reflection free wave propagation over the cable length.
The characteristic impedance of the outer circuit depends on the diameter of the measuring tube and the cable design. The effect of the mismatch in the outer circuit is discussed later on.
The equivalent circuit using concentrated elements (shown in Fig. 4) facilitates the understanding of the theoretical relationships.

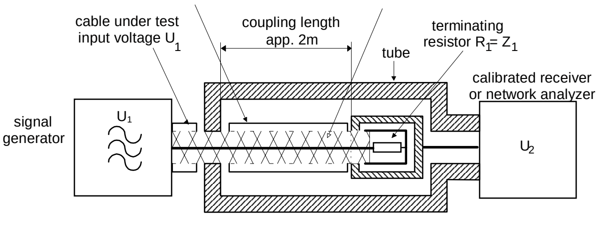


Fig. 3: Triaxial measuring set-up for screening attenuation

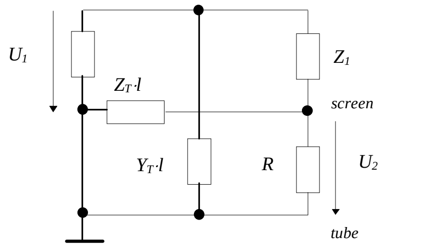


Fig. 4: Equivalent circuit of the triaxial measuring set-up

The following symbols are used:

- a_s screening attenuation
- a_{sn} normalised screening attenuation (Z_s=150Ω and $|\Delta v/v_1|$=10% or $\varepsilon_{r1}/\varepsilon_{r2n}$=1,21)
- α_1, α_2 attenuation constants of primary and secondary circuit
- β_1, β_2 phase constants of primary and secondary circuit
- $c_0 = 3\cdot 10^8 \ \frac{\mathrm{m}}{\mathrm{s}}$ propagation velocity in free space
- C_T through capacitance per unit length
- D_a outer diameter of cable screen
- D_m inner diameter of measuring tube
- ε_{r1}, ε_{r2} rel. dielectric permittivities of primary and secondary circuit
- $\varepsilon_{r2,n}$ normalised value of the rel. dielectric permittivity of the environment of the cable
- f measuring frequency
- l effective coupling length
- λ_0, λ_1, λ_2 wave length in free space, primary and secondary circuit
- M_T effective mutual inductance per unit length for braided screens $M_T = M'_{12} - M''_{12}$ where M'_{12} relates to the direct leakage of the magnetic flux and M''_{12} relates to the magnetic flux in the braid [5]
- P_1 feeding power of the primary circuit (cable under test)
- P_2 measured power received on the input resistance R of the receiver in the secondary circuit
- P_r radiated power in the environment of the cable, which is comparable to $P_{2,n}$+$P_{2,f}$ of the absorbing clamp method of IEC 61196-1, 12.4
- P_s radiated power in the normalised environment of the cable under test (Z_s=150Ω and $|\Delta v/v_1|$=10% or $\varepsilon_{r1}/\varepsilon_{r2n}$=1,21)
- R load resistance of secondary circuit (input resistance of receiver)
- R_T screen resistance per unit length
- U_1, U_2 input and output voltage of the primary and secondary circuit
- v_1, v_2 propagation velocity in primary and secondary circuit
- ω radian frequency
- $Y_T = j\,2\pi\ f\,C_T$ capacitance coupling admittance per unit length
- $Z_F = Z_1 Z_2 Y_C$ capacitive coupling impedance per unit length
- $Z_T = R_T + j\,2\pi\, f M_T$ transfer impedance per unit length
- Z_1, Z_2 characteristic impedance of primary and secondary circuit
- Z_s =150 Ω normalised value of the characteristic impedance of the environment of the cable under test

Based on the conditions of the objects to be measured it is assumed that the transfer impedance Z_T is low and the reciprocal quantity of the coupling admittance Y_T is high in comparison with the characteristic impedances Z_1 and Z_2 and the load resistance R. Therefore the feedback of the secondary circuit on primary circuit can be neglected.

When the frequency is low one may consider the primary circuit shown in Fig. 4 as voltage divider and read the disturbing voltage ratio directly. The one-sided short circuit in the measuring circuit prevents the efficiency of the capacitance coupling admittance Y_T.

$$\frac{U_2}{U_1} \approx \frac{Z_T \cdot l}{Z_1} \qquad (3\text{-}1)$$

In the high frequency range, where wave propagation has to be considered, one may expect the transfer impedance to be proportional to the frequency in most cases. Therefore it is expedient to use the following equation:

$$Z_T = R_T + j\omega\, M_T \approx j\omega\, M_T \qquad (3\text{-}2)$$

and consider the effective mutual inductance per unit length M_T at high frequencies as an approximated constant quantity as it is usually done with the through capacitance C_T.
It is common practice to describe the capacitive coupling in the form of the capacitive coupling impedance Z_F., which is nearly invariant with respect to the geometry of the outer circuit (tube). [5, 9]

$$Z_F = Z_1 Z_2 Y_T = Z_1 Z_2\, j\omega C_T \qquad (3\text{-}3)$$

Furthermore, the attenuation constants α_1 and α_2 of the circuits may generally be neglected as, for example, the value of nearly 1 dB/m of the common cable type RG 58 at 3 GHz is relatively small compared to the usual measuring uncertainty.
In the relevant literature it is common practice to describe wave propagation in the form of phase constant [5, 7]. If the ratio between effective length and wave length is used instead of the phase constant, the periodic phenomena become clearer. With wave length λ_0 in free space or λ_1, λ_2 in the circuits 1 and 2, the following relation exists:

$$\beta_{1,2} \cdot l = 2\pi \cdot \sqrt{\varepsilon_{r1,2}} \cdot \frac{l}{\lambda_0} = 2\pi \frac{l}{\lambda_{1,2}} \qquad (3\text{-}4)$$

According to the theory of wave propagation [7] and line crosstalk [8], a wave propagates in the matched inner circuit towards the matched end. In the outer circuit a part of the induced wave propagates forwards to the measuring receiver and the other part is moving backwards to the short circuit. The total reflection at the short circuit reverses this backward wave and superposes it to the original forward wave, i.e. the sum can be obtained as measured value.
If the second circuit is matched at both ends the backward wave would be measured at the generator end (near end) and the forward wave at the opposite end (far end) separately.
Hence for the near end derives from [5]

$$\frac{U_{2n}}{U_1} = \frac{Z_T + Z_F}{2Z_1} \frac{c_0}{j\omega\left(\sqrt{\varepsilon_{r1}} + \sqrt{\varepsilon_{r2}}\right)} \left\{1 - e^{-j2\pi\left(\sqrt{\varepsilon_{r1}} + \sqrt{\varepsilon_{r2}}\right)\frac{l}{\lambda_0}}\right\} \qquad (3\text{-}5)$$

and for the far end

$$\frac{U_{2f}}{U_1} = \frac{Z_F - Z_T}{2Z_1} \frac{c_0}{j\omega\left(\sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2}}\right)} \cdot \left\{1 - e^{-j2\pi\left(\sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2}}\right)\frac{l}{\lambda_0}}\right\} \cdot e^{-j2\pi\frac{l}{\lambda_2}} \qquad (3\text{-}6)$$

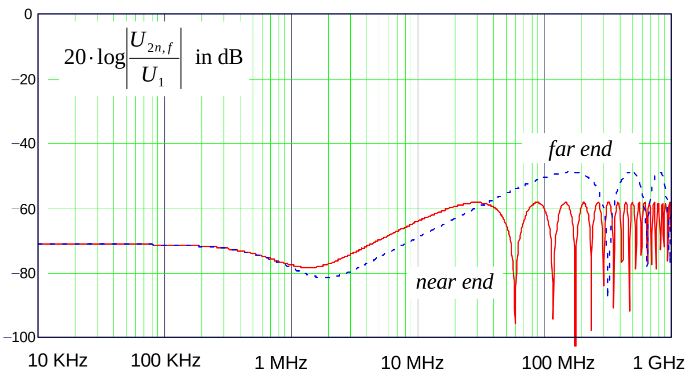


Fig 5.: calculated voltage ratio for a typical braided cable screen, quantities used:

C_T	=	0,02	pF/m			
M_T	=	0,4	nH/m			
R	=	50	Ω	l	=	2 m
Z_1	=	50	Ω	ε_{r1}	=	2,3
Z_2	=	120	Ω	ε_{r2}	=	1,1

With a short circuit and an unmatched measuring receiver these original voltage waves cause additional voltage portions. The sum of all voltage portions is zero at the shorted end (near end) and U_2 at the receiver end (far end). By use of the wave parameter and reflection factors or terminating resistors it is possible to calculate all voltage portions and the voltage U_2 from the primary induced voltage waves (3-5) and (3-6) as follows:

$$\left|\frac{U_2}{U_1}\right| \approx \left|\frac{Z_T - Z_F}{\sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2}}} \cdot \left[1 - e^{-j\varphi_1}\right] + \frac{Z_T + Z_F}{\sqrt{\varepsilon_{r1}} + \sqrt{\varepsilon_{r2}}} \cdot \left[1 - e^{-j\varphi_2}\right]\right| \cdot \left|\frac{1}{\omega \cdot Z_1}\right| \cdot \left|\frac{c_0}{2 + \left(Z_2 / R - 1\right) \cdot \left(1 - e^{-j\varphi_3}\right)}\right| \qquad (3\text{-}7)$$

or in consideration of equation (3-2), (3-3)

$$\left|\frac{U_2}{U_1}\right| \approx \left|\frac{M_T/Z_1 - C_T Z_2}{\sqrt{\varepsilon_{r1}}-\sqrt{\varepsilon_{r2}}}\left[1-e^{-j\varphi_1}\right] + \frac{M_T/Z_1 + C_T Z_2}{\sqrt{\varepsilon_{r1}}+\sqrt{\varepsilon_{r2}}}\left[1-e^{-j\varphi_2}\right]\right| \cdot \left|\frac{c_0}{2+\left(Z_2/R-1\right)\cdot\left(1-e^{-j\varphi_3}\right)}\right| \qquad (3\text{-}8)$$

where

$$\varphi_1 = 2\pi\left(\sqrt{\varepsilon_{r1}}-\sqrt{\varepsilon_{r2}}\right)\frac{l}{\lambda_0} \qquad \varphi_2 = 2\pi\left(\sqrt{\varepsilon_{r1}}+\sqrt{\varepsilon_{r2}}\right)\frac{l}{\lambda_0} \qquad \varphi_3 = \varphi_2-\varphi_1 = 4\pi\sqrt{\varepsilon_{r2}}\frac{l}{\lambda_0}$$

Another way to obtain the related induced voltage shows [10].
The functional equation (Fig. 6)

$$\left|1-e^{-j\varphi}\right| = \left|2\sin\varphi/2\right| \qquad \text{with } \varphi = \varphi_1, \varphi_2, \varphi_3 \qquad (3\text{-}9)$$

shows, that the equation of the voltage ratio contains three periodic partial functions of the ratio effective length l to wave length λ_0:

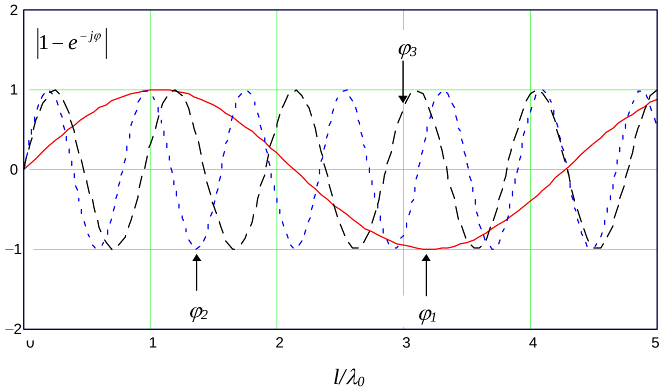


Fig. 6.: calculated periodic functions for ε_{r1} = 2,3 and ε_{r2} = 1,1

For low frequencies, when l<<λ_0 and, consequently, sinφ ≈ φ, equation (3-7) changes into (3-1), the result of the common measuring method for the transfer impedance.
An example of the theoretical curve of the voltage ratio is shown in fig. 7 in two diagrams. The left one with a logarithmic scale to extend the lower frequency range and the right one with a linear scale up to very high frequencies.

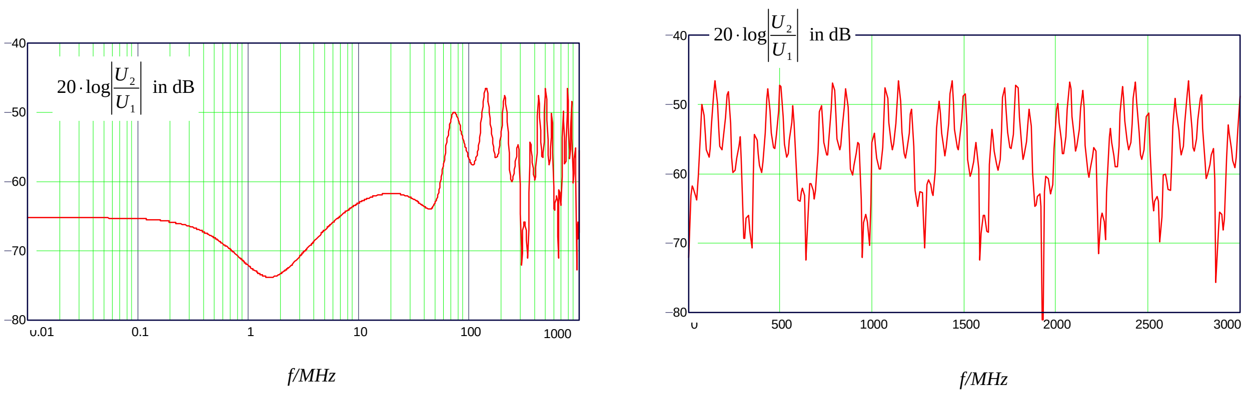


Fig 7.: calculated voltage ratio, quantities used:

C_T	=	0,02	pF/m	M_T	=	0,4	nH/m
R	=	50	Ω	l	=	2	m
Z_1	=	50	Ω	ε_{r1}	=	2,3	
Z_2	=	120	Ω	ε_{r2}	=	1,1	

It is not expedient to value the induced power for an exact length of cable at a single frequency, anywhere between a minimum and maximum of the function. Only the periodic maximum voltage is important for the evaluation of the screening effectiveness. In the outer circuit the wave propagation shall be nearly the same as in free space. Therefore the characteristic impedance Z_2 is higher than the common input resistance R of the measuring receiver, i.e. 50 Ω or sometimes 75 Ω.
Consequently periodic maximum values of the voltage ratio are obtained from (3-7), (3-8) which are independent of the input resistance of the receiver R and of effective cable length l:

$$\left|\frac{U_2}{U_1}\right|_{\max} \approx \frac{c_0}{\omega Z_1} \cdot \left| \frac{Z_\mathrm{T}-Z_\mathrm{F}}{\sqrt{\varepsilon_\mathrm{r1}}-\sqrt{\varepsilon_\mathrm{r2}}} + \frac{Z_\mathrm{T}+Z_\mathrm{F}}{\sqrt{\varepsilon_\mathrm{r1}}+\sqrt{\varepsilon_\mathrm{r2}}} \right| \tag{3-10}$$

or in consideration of equation (3-2), (3-3)

$$\left|\frac{U_2}{U_1}\right|_{\max} \approx \left| \frac{M_T/Z_1 - C_T Z_2}{\sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2}}} + \frac{M_T/Z_1 + C_T Z_2}{\sqrt{\varepsilon_{r1}} + \sqrt{\varepsilon_{r2}}} \right| \cdot c_0 \tag{3-11}$$

At first sight C_T, Z_2, ε_{r2} and Z_F appear as random quantities which depend on freely chosen dimensions of the measuring tube. In reality, however, the voltage ratio is independent of the characteristic impedance of the outer circuit since $C_T \cdot Z_2$ and Z_F are practically invariant with respect to the dimensions of the measuring tube [5,9]. Furthermore, the influence of the cable sheath on the resulting relative permittivity ε_{r2} is negligible if the design of the measuring tube takes into account the requirement for a wave propagation which is approximately the same as in the free space, in consequence $\varepsilon_{r2} \approx 1{,}0$.
The periodic maximum value is independent of the effective length l and frequency f or wave length λ. A measured frequency response would hint at a frequency related quantity rather than the pure mutual inductance M_T.
As it is seen from fig. 6 and 7 the envelope rise is reached with the first maximum of the wide period at:

$$\lambda_o / l \leq 2 \cdot \left| \sqrt{\varepsilon_\mathrm{r1}} - \sqrt{\varepsilon_\mathrm{r2}} \right| \quad \text{or} \quad f > \frac{c_0}{2 \cdot l \cdot \left| \sqrt{\varepsilon_\mathrm{r1}} - \sqrt{\varepsilon_\mathrm{r2}} \right|} \tag{3-12}$$

In this frequency range Z_T can be calculated if Z_F is negligible:

$$|Z_T| \approx \frac{\omega \cdot Z_1 \cdot |\varepsilon_{r1} - \varepsilon_{r2}|}{2 \cdot c_0 \cdot \sqrt{\varepsilon_{r1}}} \cdot \left|\frac{U_2}{U_1}\right|_{\max} \tag{3-13}$$

4. Screening Attenuation

The screening attenuation is defined as the logarithmical ratio of the maximum power in the secondary (outer) circuit to the power propagating in the primary (inner) circuit.

$$a_s = -10 \cdot \log_{10}\left(Env \left| \frac{P_{r,\max}}{P_1} \right| \right) \tag{4-1}$$

The power coupled into the outer circuit depends on Z_2 although the peak voltage is independent of it. Thus a normalised value of the characteristic impedance of the outer circuit Z_s has to be defined. It is common practice to define Z_s = 150 Ω [5].
In the standardised „absorbing clamp method“ (IEC 61196-1 clause 12.4) the outer circuit is matched with Z_2, and the radiated power is the sum of the near end and far end crosstalk. From the comparison of that measuring circuit with the measuring circuit of the triaxial method result the relation of the measured power to the radiated power.
The equivalent circuit for an electrical short part of the length *Δl* and for a negligible capacitive coupling illustrates the circumstances.

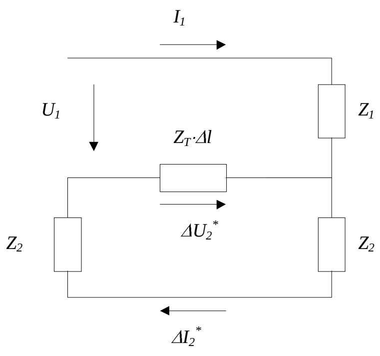


Fig. 8 Equivalent circuit for an electrical short part of the length Δl and negligible capacitive coupling

The power in the primary circuit is:

$$P_1 = U_1 \cdot I_1 = \frac{U_1^{\ 2}}{Z_1} = I_1^{\ 2} \cdot Z_1 \qquad \text{(4-2)}$$

The power in the secondary circuit, which is coupled by the transfer impedance Z_T is

$$P_2^* = \Delta U_2^* \cdot \Delta I_2^* \qquad \Delta U_2^* = I_1 \cdot Z_T \cdot \Delta l \qquad \Delta I_2^* = \frac{\Delta U_2^*}{2 \cdot Z_2} \qquad \text{(4-3)}$$

thus

$$\frac{P_2^*}{P_1} = \frac{\left(\Delta U_2^*\right)^2}{2 \cdot Z_2} \cdot \frac{1}{I_1^2 \cdot Z_1} = \frac{\left(Z_T \cdot \Delta l\right)^2}{2 \cdot Z_1 \cdot Z_2} \qquad \text{(4-4)}$$

If the secondary circuit is short circuited at one end and terminated by R at the other end the power measured at R is

$$\frac{P_2}{P_1} = \frac{(Z_T \cdot \Delta l)^2}{Z_1 \cdot R} \qquad \text{(4-5)}$$

thus

$$\frac{P_2^{\ *}}{P_2} = \frac{R}{2Z_2} \qquad \text{(4-6)}$$

or in the case of radiation due to the normalised characteristic impedance of the environment

$$\frac{P_r}{P_2} = \frac{P_{r,\max}}{P_{2,\max}} = \frac{R}{2Z_s} \qquad \text{(4-7)}$$

Thus the screening attenuation has to be calculated by:

$$\begin{aligned} a_s &= 10 \cdot \log_{10}\left|\frac{P_1}{P_{r,\max}}\right| = 10 \cdot \log_{10}\left|\frac{P_1}{P_{2,\max}} \cdot \frac{2 \cdot Z_s}{R}\right| \\ &= 10 \cdot \log_{10}\left|\left(\frac{U_1}{U_{2,\max}}\right)^2 \cdot \frac{2 \cdot Z_s}{Z_1}\right| \qquad \text{(4-8)} \\ &= 20 \cdot \log_{10}\left|\frac{U_1}{U_{2,\max}}\right| + 10 \cdot \log_{10}\left|\frac{300\Omega}{Z_1}\right| \end{aligned}$$

5. Normalised screening attenuation

From eq. (3-10) one may recognise, that the maximum voltage ratio and therefore the screening attenuation is depending on the velocity difference between the primary and secondary circuit. Therefore the test results may also be presented in normalised conditions where Z_s = 150 Ω and the velocity difference $|\Delta v/v_1|$ = 10 % or $\varepsilon_{r1}/\varepsilon_{r2,n}$ = 1,21.
The normalised screening attenuation is calculated by:

$$a_{s,n} = 20 \cdot \log_{10} \left| \frac{\omega \cdot \sqrt{Z_1 \cdot Z_s} \cdot \left| \sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2,n}} \right|}{Z_T \cdot c_0} \right| \qquad (5\text{-}1)$$

With respect to eq. (3-10), (4-8) and (5-1) and with neglect of Z_F, the difference Δa of the normalised and the measured screening attenuation is given by:

$$\Delta a = a_{sn} - a_s = 20 \cdot \log_{10} \left(\sqrt{2} \cdot \frac{\left| 1 - \sqrt{\dfrac{\varepsilon_{r2,n}}{\varepsilon_{r1}}} \right|}{\left| 1 - \dfrac{\varepsilon_{r2,t}}{\varepsilon_{r1}} \right|} \right) \qquad (5\text{-}2)$$

where $\varepsilon_{r2,t} \approx 1{,}1$ is the relative dielectric permittivity of the outer circuit (tube) during measurement.
The following table shows the difference Δa for typical cable dielectric:

ε_{r1}	2,3	2,1	1,6	1,3
$\varepsilon_{r2,n}$	1,9	1,7	1,3	1,1
Δa in dB	-12	-11	-8	-2

6. Measuring results

The measuring curve of common type of cables prove the validity of the theoretical basis. The voltage ratio U_2 / U_1 is measured by means of a network analyser having an internal resistance of 50 Ω. The screening attenuation a_s is presented in Figures 9 to 13 for three types of cables as a function of frequency.

- RG 58 according to MIL-C-17 with single copper braid
- HF 75 0,7/4,8 2YCY with a dielectric of solid PE and a single copper braid
- HF 75 1,0/4,8 02YCY with a dielectric of foamed PE and a single copper braid
- RG 223 according to MIL-C-17 with double copper braid

The theoretical relations of the transitions from low to medium and high frequencies - appearing in the calculated curve in Fig. 5 - become most evident with the single copper braid (see Fig. 9). Here the voltage ratio is independent of the frequency up to approx. 0,4 MHz but proportional to the effective length of the measuring tube like the transfer impedance. At high frequencies, higher than approx. 100 MHz, superpositioned periodic functions occur showing maximum values of approximately equal magnitude independent of frequency and effective length. The frequency, at which the superposition appears is reciprocal to the effective length just as the frequency spacing of the peak values (see Fig. 9,10). In contrast to the effective length of 2 m, the effective length of 0,5 m does not allow to plot the screening envelope curve with sufficient accuracy any more, due to the wide spacing of the long period maximum values.

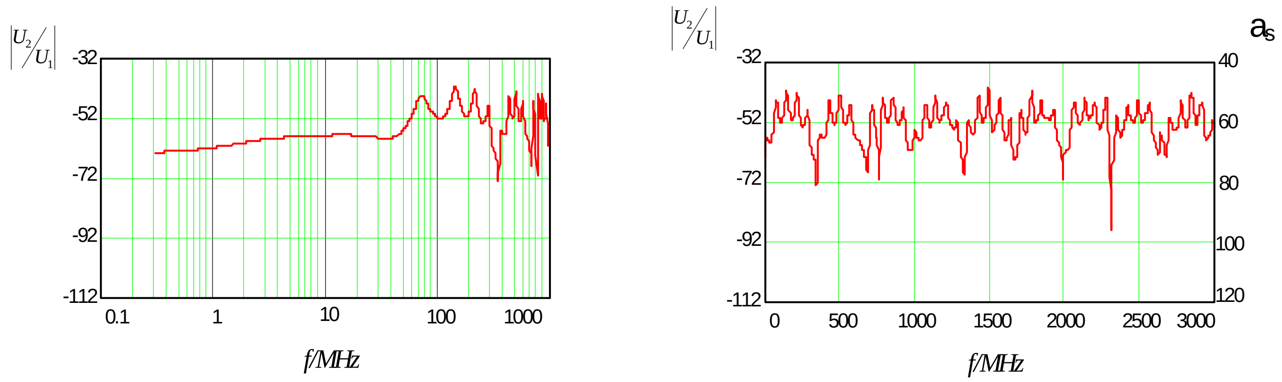


Fig. 9: Frequency response of logarithmic voltage ratio $|U_2/U_1|$ in dB (left hand scale) and screening attenuation a_s (right hand scale) of single braid screen, Cable Type RG 58, coupling length l = 2 m

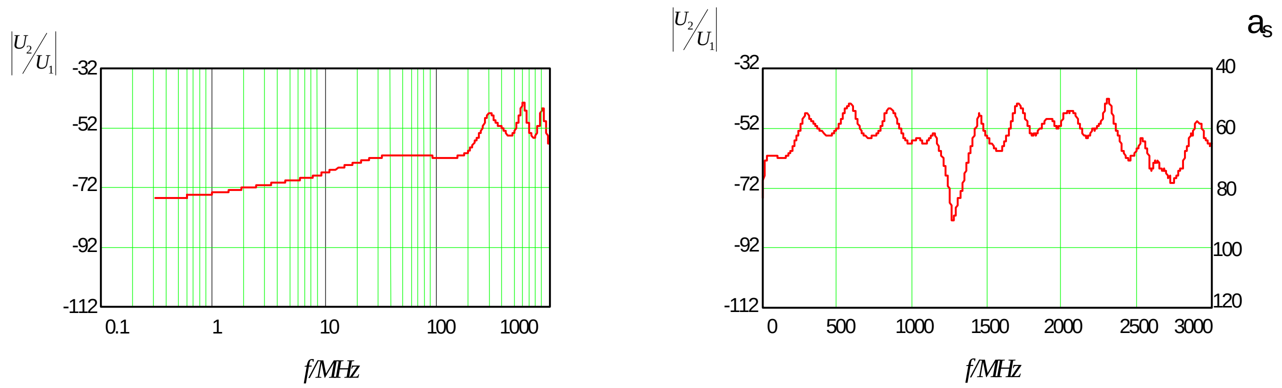


Fig. 10: Frequency response of logarithmic voltage ratio $|U_2/U_1|$ in dB (left hand scale) and screening attenuation a_s (right hand scale) of single braid screen, Cable Type RG 58, coupling length l = 0,5 m

The periodic frequency spacing in the measured curve and the screening attenuation are dependent on the velocity difference between primary and secondary circuit (eq. 3-7, 3-10). This theoretical relation become most evident in Fig 11 and 12. Where the cable screen of both cables are equal, but the relative permittivities of the cable dielectric ε_{r1} and thus the velocity difference in the test set-up differ. In Fig 11 we have ε_{r1}=2,3 and a velocity difference $|\Delta v/v_1| \approx 45\%$ whereas in Fig 12 ε_{r2}=1,7 and $|\Delta v/v_1| \approx 24\%$. Thus in Fig 12 we have a larger frequency spacing of the wide period and also a lower screening attenuation. But the normalised screening attenuation of both cable screens are equal, $a_s \approx 43$ dB.

For the cable with double copper braid (Fig. 13) the theoretical relations become apparent only if the measurement is very accurate and the receiver is sensitive enough for low induced voltage. Apart from its level and distinct function of frequency, the screening attenuation of the double copper braid is obviously similar to that of the single copper braid.

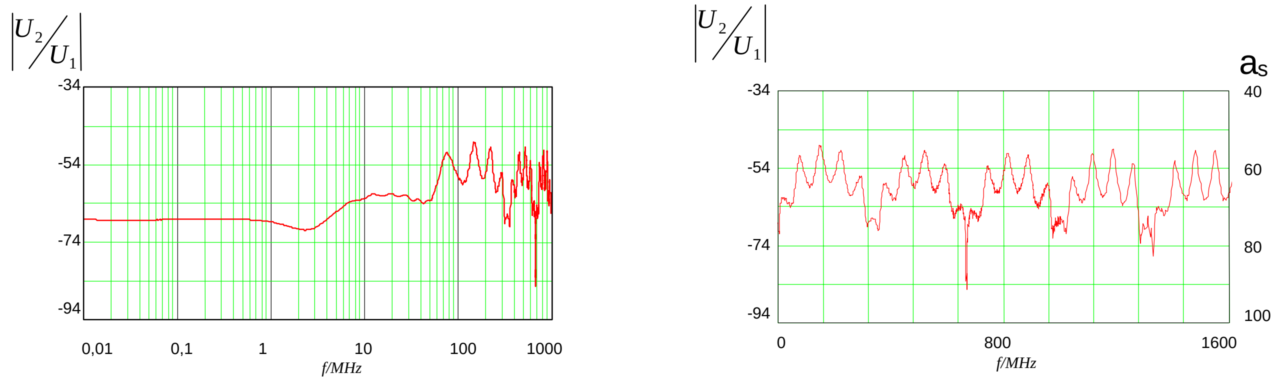


Fig. 11: Frequency response of logarithmic voltage ratio $|U_2/U_1|$ in dB (left hand scale) and screening attenuation a_s (right hand scale) of Cable Type HF 75 0,7/4,8 2YCY, e_{r1}=2,3 , $|\Delta v/v_1|$=45% , coupling length l = 2 m

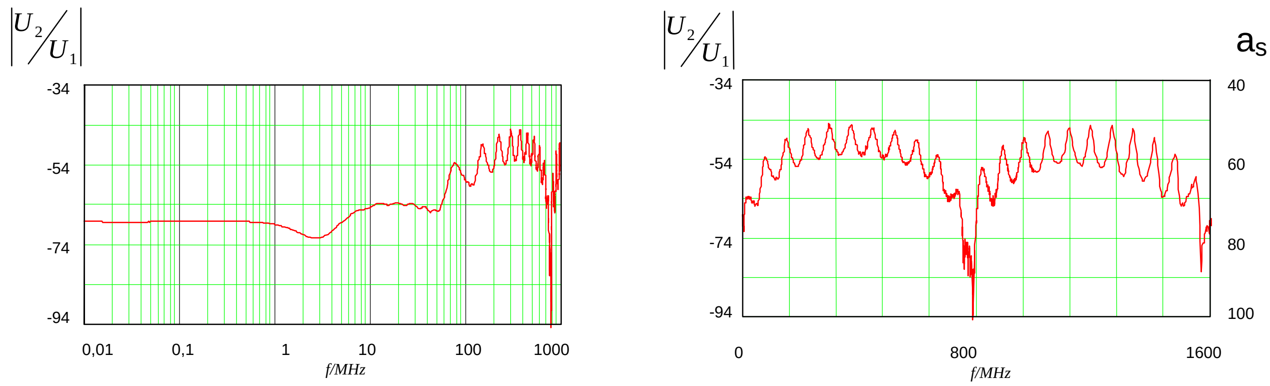


Fig. 12: Frequency response of logarithmic voltage ratio $|U_2/U_1|$ in dB (left hand scale) and screening attenuation a_s (right hand scale) of Cable Type HF 75 1,0/4,8 02YCY, e_{r1}=1,7 , $|\Delta v/v_1|$=24% , coupling length l = 2 m

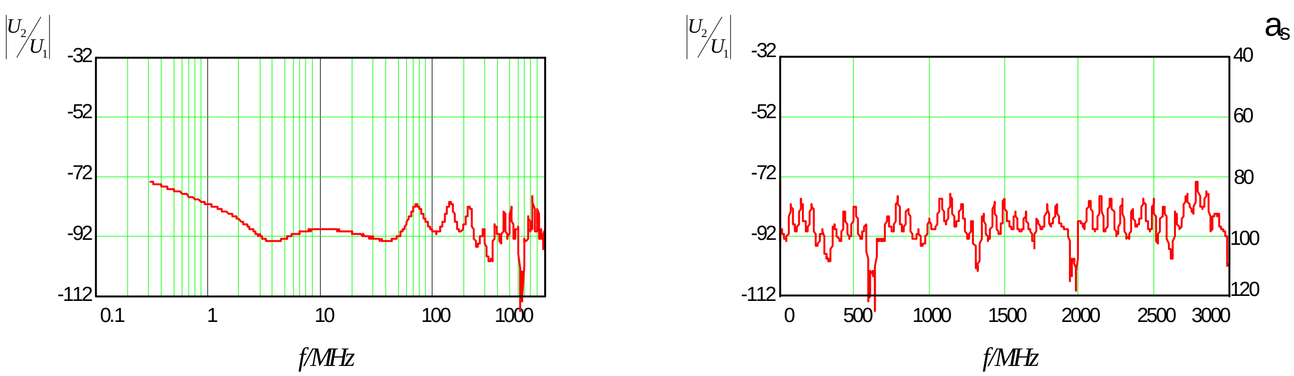


Fig. 13: Frequency response of logarithmic voltage ratio $|U_2/U_1|$ in dB (left hand scale) and screening attenuation a_s (right hand scale) of double braid screen, Cable RG 223, coupling length l = 2 m

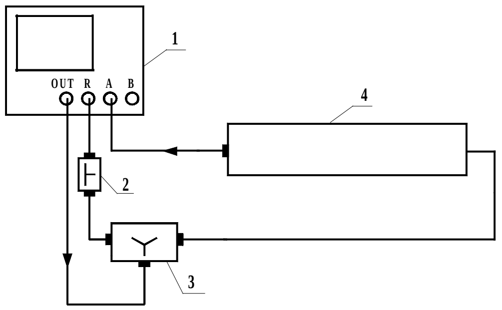
7. Comparison with absorbing clamp method

In the absorbing clamp method according to IEC 61196-1 clause 12.4 , in principle, the current on the outside of the cable under test is measured. The matched outer circuit is directly induced by the inner circuit. The power in the outer circuit is related to the current by calibration.
Comparison results are available of some coaxial cables of different screen designs. They show a difference of max. 3dB

	screening attenuation a_s in dB		
cable type, screen	frequency GHz	absorbing clamp method	triaxial method
RG 58, single braid	0,2	51	48
	0,8	52	50
	3,0	-	50
RG 214, single braided	0,2	51	50
	0,8	54	51
	3,0	-	53
RG 214, double braid	0,2	79	79
	0,8	82	81
	3,0	-	83
RG 223, double braid	0,2	86	88
	0,8	90	90
	3,0	-	83

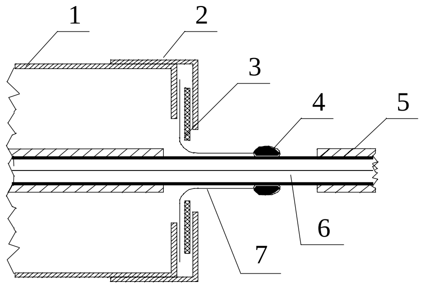
8. Practical design of the test set-up

The set-up to measure the screening attenuation a_S is in principle the set-up to measure the attenuation of RF devices, where the voltage ratio U_2/U_1 is measured. The cable under test is connected to the output of a RF-generator, the output of the coupling tube is connected to the measuring input of a RF-receiver. Generator and receiver may be included in a sensitive network analyser (see Fig. 3 and 14).
The measuring tube shall be of a material, which is not ferromagnetic and good conductive (for example brass), with an inner diameter of about 40 mm to 50 mm and a length of 2 m to 4 m or more, where the total length of 2 m or more may be achieved by screwing together single parts of tubes (RF-tight).
One way to realise the short circuit at the near end of the CUT is to solder a braid of silvered copper wires to a punched disk of copper. This "contacting braid" is fixed on the outer conductor of the cable sample where the sheath is removed, e.g. with cable clamps. The electrical contact between this contacting braid and the measuring tube may then be achieved by a jam-disk, which is fixed by the clasp cap, which is screwed to the tube (see Fig. 15).
The contacting braid, which is prepared once may be used several times. Soldering of the screen of the cable sample to the tube - as usual at the classic triaxial - set-up is no longer required and the time to prepare the CUT is minimised.
The termination at the far end of the CUT is achieved by a resistor of the same value as the characteristic impedance of the CUT. Experience has shown, that best results are obtained with SMD resistors respectively so called "Mini-Melf-Resistors" with low mechanical dimensions and good RF-characteristics, which are soldered directly between the inner and the outer conductor of the CUT. To avoid radiation and to contact the outer conductor of the CUT, this termination is shielded by a case, which is well conductive (see Fig. 3).



1 Network Analyser 3 Power Divider
2 Attenuator 20 dB 4 Measuring tube

Fig. 14: schematically arrangement for the measurement of the screening attenuation a_S



1 tube of brass 5 sheath of cable sample
2 clasp cap 6 cable under test (CUT)
3 jam disk 7 contacting braid
4 contact to the cable screen of CUT

Fig. 15: Short circuit between tube and cable screen of the CUT

Important for clear and reproducible results is the centring of the sample in the measuring tube. A slacked hanging cable under test in the measuring tube will led to deviations of the characteristic impedance Z_2 of the outer system over the coupling length and thus to additional reflections. Centring may be achieved by punched polyethylene disks which are placed in the measuring tube, or better by stretching the sample under test, e.g. with a desk vice. Also a vertical mounting of the measuring tube is useful.

9. Influence of mismatches

There may be mismatches in the inner or outer circuit of the test set-up which influence the results significant. Theoretical and practical investigations [11] show that a mismatch of the terminating resistor in the inner circuit is of low influence as long as:

$$\frac{\left|R_{termination} - Z_1\right|}{Z_1} \cdot 100\% \leq 10\% .$$

Additional mismatches in the outer circuit however result in significant errors. With the screening case of the terminating resistor a mismatch is inserted into the outer circuit, which affect the results significantly depending on the mechanical dimensions [11]. The mean characteristic impedance of the outer circuit, formed by the cable screen and the measuring tube, respectively in the outer circuit at the screening case is given by:

$$Z_2 \approx \frac{60\Omega}{\sqrt{\varepsilon_{r2}}} \cdot \ln\left(\frac{D_m}{D_a}\right) \qquad (9\text{-}1)$$

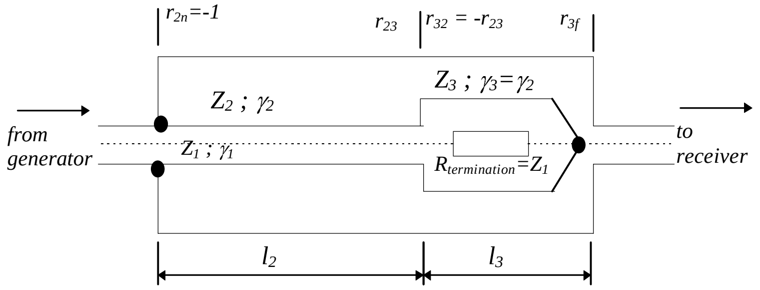
$$Z_3 \approx \frac{60\Omega}{\sqrt{\varepsilon_{r2}}} \cdot \ln\left(\frac{D_m}{D_{case}}\right) \qquad (9\text{-}2)$$

where

D_a outer diameter of cable screen
D_{case} outer diameter of screening case
D_m inner diameter of measuring tube

A deviation between D_{case} and D_a thus results in different impedance's and therefore in additional reflections in the outer circuit. For example a screening case with a outer diameter of $D_{case} = 1{,}2 \cdot D_a$ results in a impedance Z_3 which is 11 Ω less than Z_2 (ε_{r2}=1,0).

Fig. 16 facilitates the understanding of the theoretical relationships.



r_{2n} = -1(due to the short circuit)

$$r_{23} = \frac{Z_3 - Z_2}{Z_3 + Z_2}$$

$$r_{32} = -r_{23}$$

$$r_{3f} = \frac{R - Z_2}{R + Z_2}$$

Fig. 16: real triaxial set-up

The outer circuit thus consist of two lines with different characteristic impedance's. To calculate the voltage at the receiver some additional variables have to be defined.
U_h is the voltage, which is coupled from the cable under test into the outer circuit (Z_2, γ_2, l_2), propagation to the far end, including the total reflection at the near end.

$$\frac{U_h}{U_1} = \frac{U_{2f}}{U_1} + \frac{U_{2n}}{U_1} \cdot r_{2n} \cdot e^{-\gamma_2 \cdot l_2} \qquad (9\text{-}3)$$

Where U_{2f}, U_{2n} are the voltages in a matched outer circuit according eq. (3-5) and (3-7).
Multiple reflections of this wave between the short circuit at the near end of the outer circuit and the transition from Z_2 to Z_3 are described by T_{2f}.

$$T_{2f} = \frac{1 + r_{23}}{1 - r_{2n} \cdot r_{23} \cdot e^{-2 \cdot \gamma_2 \cdot l_2}} \qquad (9\text{-}4)$$

The superposition of the wave which is propagating from the line Z_2, γ_2, l_2 to the far end (receiver) of the line Z_3, γ_3, l_3 - including the multiple reflections between the transitions from Z_3 to Z_2 and Z_3 to R (receiver input) - is described by T_{3f}.

$$T_{3f} = \frac{1 + r_{3f}}{1 - r_{32} \cdot r_{3f} \cdot e^{-2 \cdot \gamma_3 \cdot l_3}} \cdot e^{-\gamma_3 \cdot l_3} \qquad (9\text{-}5)$$

The superposition of the wave which is propagating from line Z_3, γ_3, l_3 to line Z_2, γ_2, l_2 is described by T_{32}.

$$T_{32} = \frac{1 + r_{32}}{1 - r_{32} \cdot r_{3f} \cdot e^{-2 \cdot \gamma_3 \cdot l_3}} \cdot r_{3f} \cdot e^{-2 \cdot \gamma_3 \cdot l_3} \qquad (9\text{-}6)$$

The superposition of the wave which is propagating from line Z_2, γ_2, l_2 to line Z_3, γ_3, l_3 is described by T_{23}.

$$T_{23} = \frac{1 + r_{23}}{1 - r_{2n} \cdot r_{23} \cdot e^{-2 \cdot \gamma_2 \cdot l_2}} \cdot r_{2n} \cdot e^{-2 \cdot \gamma_2 \cdot l_2} \qquad (9\text{-}7)$$

In consideration of all these reflections the voltage at the receiver is calculated by:

$$\frac{U_{receiver}}{U_1} = \frac{U_h}{U_1} \cdot \frac{T_{2f} \cdot T_{3f}}{1 - T_{32} \cdot T_{23}} \qquad (9\text{-}8)$$

Fig. 17 and 18 show the calculated voltage ratio for a cable screen with the same characteristics as in Fig. 7 with different dimensions of the screening case.

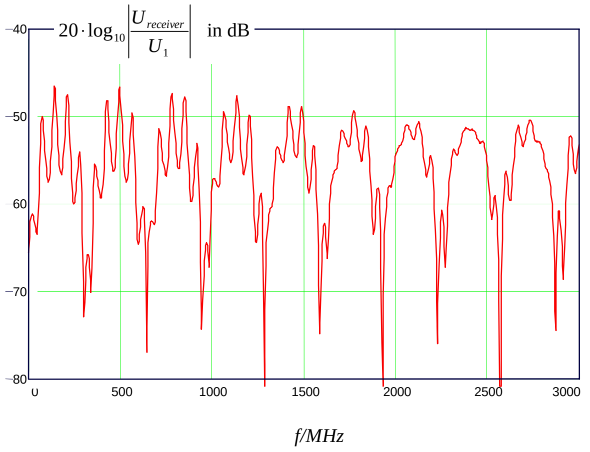


Fig. 17: calculated voltage ratio including multiple reflections caused by the screening case, quantities used:

$C_T =$ 0,02 pF/m

$M_T =$ 0,4 nH/m

R =	50	Ω	Z_1 =	50	Ω	ε_{r1} =	2,3	
Z_2 =	120	Ω	ε_{r2} =	1,1		l_2 =	2	m
Z_3 =	90	Ω	ε_{r2} =	1,1		l_3 =	0,03	m

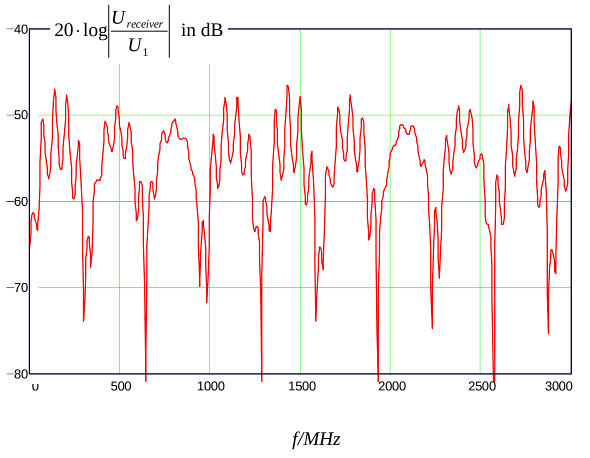


Fig. 18: calculated voltage ratio including multiple reflections caused by the screening case, quantities used:

$C_T =$ 0,02 pF/m

$M_T =$ 0,4 nH/m

R =	50	Ω	Z_1 =	50	Ω	ε_{r1} =	2,3	
Z_2 =	120	Ω	ε_{r2} =	1,1		l_2 =	2	m
Z_3 =	90	Ω	ε_{r2} =	1,1		l_3 =	0,1	m

To avoid the disturbing reflections at the screening case one have to minimise the reflection factor r_{23} or (and) r_{3f}. A worthwhile solution in praxis is to design the screening case in a way that the characteristic impedance Z_3 is approximately of the same value as the input resistance of the receiver. In this case the reflection factor $r_{3f} \approx 0$ and thus $T_{3f}=1$, $T_{32}=0$. That results in a voltage ratio which is equal to the ideal frequency response of eq. (3-9).

10. Preview

In the field of digital data transmission and telecommunication the use of symmetrical multi pair/quad cables increases. Therefore it is also necessary to determine their EMC behaviour.

The EMC behaviour of symmetrical cables depends on both, the unbalance attenuation a_U of the pairs and the screening attenuation a_S of their screen. The sum of the unbalance attenuation a_U of the pairs and the screening attenuation a_S of their screen is named coupling attenuation a_C.
With the new triaxial measuring procedure explained above, also the coupling attenuation a_C may be measured, when the cable is driven in the differential mode, see figure 19. This procedure is still under consideration [12].

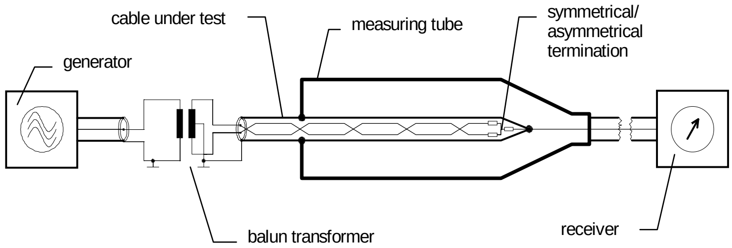


Fig. 19: test set-up for the determination of the coupled power attenuation of a shielded twisted pair.

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