INTERNATIONAL ATOMIC ENERGY AGENCY
UNITED NATIONS EDUCATIONAL, SCIENTIFIC AND CULTURAL ORGANIZATION
INTERNATIONAL CENTRE FOR THEORETICAL PHYSICS
I.C.T.P., P.O. BOX 586, 34100 TRIESTE, ITALY, CABLE: CENTRATOM TRIESTE

H4.SMR/841-10

FOURTH ICTP-URSI-ITU(BDT) COLLEGE ON RADIOPROPAGATION: Propagation, Informatics and Radiocommunication System Planning

30 January - 3 March 1995

Miramare - Trieste, Italy

Noise, Interference and EMC

P. Degauque
Laboratoire de Radiopropagation et Electronique, CNRS
Villeneuve D'Ascq, France

MAIN BUILDING STRADA COSTIERA, 11 TEL. 2240111 TELEFAX 224163 TELEX 460392 ADRIATICO GUEST HOUSE VIA GRIGNANO, 9 TEL. 224241 TELEFAX 224531 TELEX 460449
MICROPROCESSOR LAB. VIA BEIRUT, 31 TEL. 224471 TELEFAX 224600 TELEX 460392 GALILEO GUEST HOUSE VIA BEIRUT, 7 TEL. 22401 TELEFAX 2240310 TELEX 460392

Outline

1) • Coupling to a network of transmission lines
   • Non uniform lines
   • Topological approach

2) Coaxial cable and multiwire shielded cables
   Transfer impedance

3) Penetration through apertures and into shielded boxes

4) Susceptibility of electronic equipment and measurement procedures

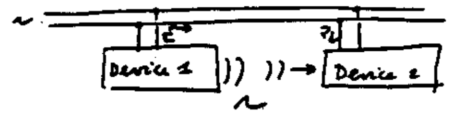
Annex · Source of interference and coupling to cables

COMPATIBILITY

E.M.C

Presentation

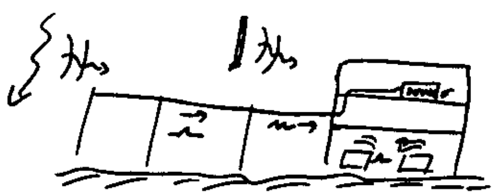
"Compatibility" of 2 electrical or electronic devices



Device 1 emits disturbing signal (radiated or conducted) → interference with device 2

• Compatibility if the E.M. noise generated by each of these devices does not "interfere" with their normal performances

More generally: Compatibility of a device (or system) in its environment

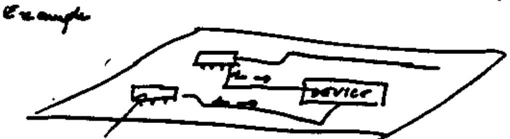


EMI exists when undesirable voltages or currents are present to influence adversely the performance of a device

Where are situated the cause of E.M.I

• Within the system one is dealing with → INTRA SYSTEM PROBLEM

Example



AC→DC converters

• E.M.I comes from the outside → INTER SYSTEM

Approach → Experimental and Theoretical one (a)

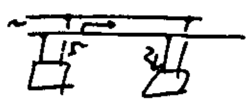
1st step: Sources: Identify the possible sources of EMI and have an idea of the disturbing signals which will be incident on the system

• theory
• measurement
• statistical approach (noise)

2nd step: Understand the coupling mechanism

- Conducted interference

Easy!



- Radiated interference $\vec{E},\vec{H} \underset{\text{determine}}{\longrightarrow} I, V$

Simplest case: lines // ground plane

Influence of the shielding (first for a cable)

→ coaxial cable

- multiwire shielded cable

Influence of the ground connections

Shielded box

• Knowledge of the E.M.I produced by any electrical system → Norms to avoid "E.I pollution" → Measurements

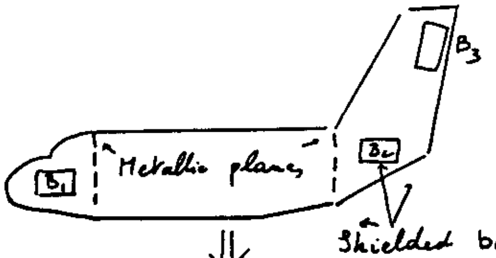
• levels of "susceptibility" of the equipment

Max level of the interference (either E, H or I, V) so that the performance of the system remains unchanged below this value → Norms: Any system must work in a standard or usual environment

• Biological effect

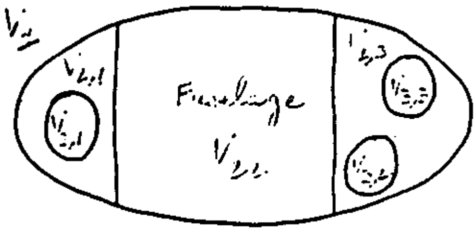
- Coupling to a network of transmission lines.
- Non uniform lines.
- Topological approach.

3



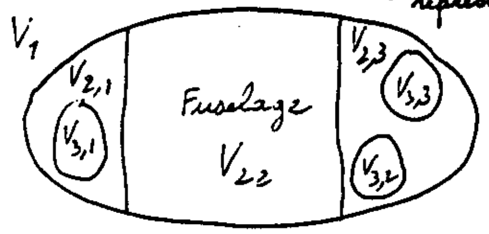
⇓ Topological representation

External volume

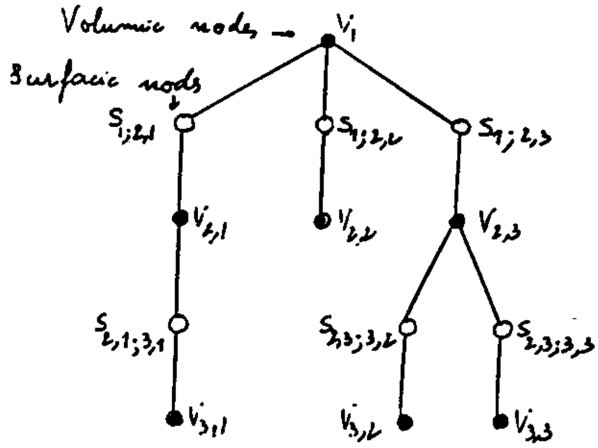


4

External volume — Topological representat.



⇓ interaction diagram

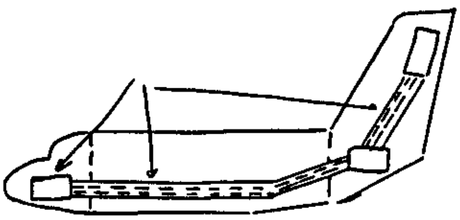


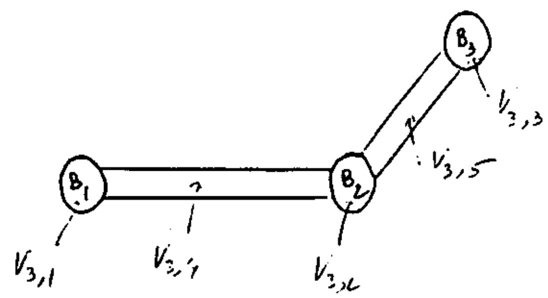
5

Example: multiwire shielded cable

⇓

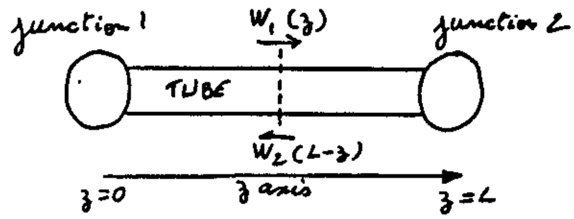
V_3





⇒ In this presentation → consider only the coupling to this structure

B_1, B_2, B_3 behave as "junctions" connecting "tubes"



W: "wave matrix" characterizing the propagation of I, V along the tube.

Transmission line theory ("uniform" tube)

$$\frac{d}{dz}[V] = -[Z][I] + [V^s]$$

$$\frac{d}{dz}[I] = -[Y][V] + [I^s]$$

$\}$ Source terms

$$[\gamma_c] = \{[Z][Y]\}^{1/2} \; ; \; [Z_c] = [\gamma_c][Y]^{-1}$$

Choice of W ? → Propagation along $z > 0$

$$[W] = [V] + [Z_c][I]$$

Z_c → Characteristic impedance matrix of the tube (Well defined for a uniform tube).

6

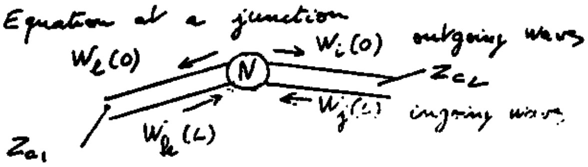
• Propagation along 1 tube (length L)

$$[\Gamma] = \exp(-[\gamma]L)$$

$$[W(L)] = [\Gamma][W(0)] + [W_s]$$

Prop. matrix — integration of the source terms along the line

• Equation at a junction



$$W_i(0) = S_{ij}\, W_j(L) + S_{ik}\, W_k(L)$$

$$W_\ell(0) = S_{\ell j}\, W_j(L) + S_{\ell k}\, W_k(L)$$

At the junction N

$$[W_N(0)] = [S_N][W_N(L)] \qquad \forall N \Rightarrow$$

$$([1] - [S][\Gamma])[W(0)] = [S][W_s]$$

B.L.T. Eq.

Remark: At the junction N → Supermatrix Z_c

$$[Z_c] = \begin{bmatrix} [Z_{c_1}] & 0 \\ 0 & [Z_{c_2}] \end{bmatrix}$$

$[S]$ depends on $[Z_c]$ → "topological" scattering matrix

Determination of $[S]$?

→ analytically

→ Measurements ? → Reference: scalar charact. impedance = 50 Ω (S_{50})

$$[S_{50}] \rightarrow [Y] = \frac{1}{50}\{[1] + [S_{50}]\}^{-1}\{[1] - [S_{50}]\}$$

$$\downarrow$$

$$[S_{Topo}] = \{[1] - [Z_c][Y]\}\{[1] + [Z_c][Y]\}^{-1}$$

ELECTROMAGNETIC TOPOLOGY

—> TRANSMISSION LINE NETWORK

B.L.T. Equation

$$\{[1]-[S][\Gamma]\}[W(0)]=[S][Ws]$$

$$\Updownarrow$$

Transmission line theory

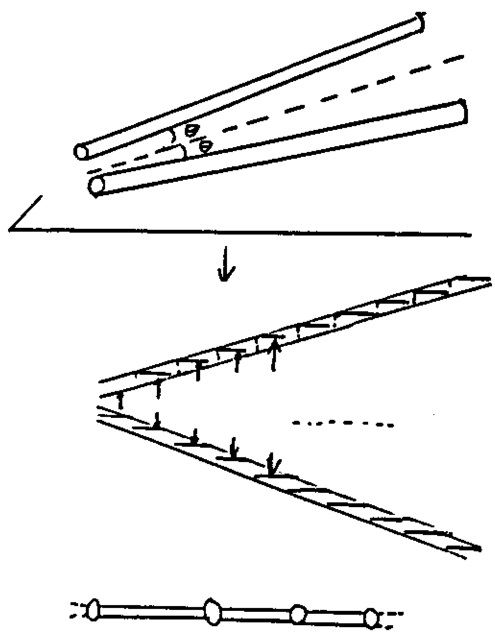
$$-\frac{dV}{dz}=(Z)(I)+(\varepsilon)$$

$$-\frac{dI}{dz}=(Y)(V)+(\zeta)$$

$$\boxed{\text{NONUNIFORM TRANSMISSION LINES } (Z(z)) \text{ and } (Y(z))}$$

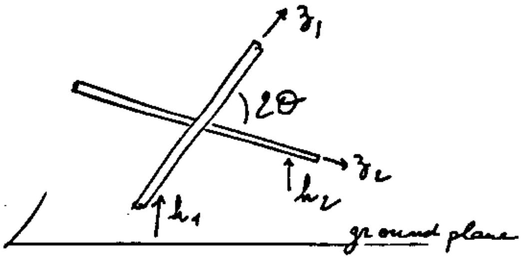
~~B.L.T. Equation~~

Non uniform line → series of uniform T.L.



Series of tubes + junctions

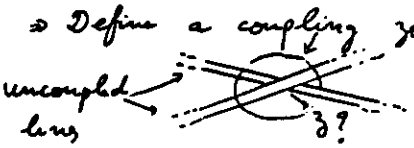
Other example: crossing of 2 conductors



~~Succession~~

- Series of tubes. The distance between each elementary tube → constant (but by modifying L_{ij} to take θ into account)
- What part of the lines must be discretized? Along what distance the lines 1 and 2 are "coupled"

⇒ Define a coupling zone



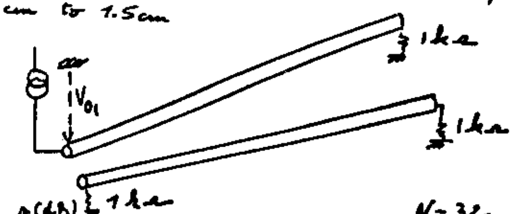
⇒ Introduce a "coupling ratio" R_c

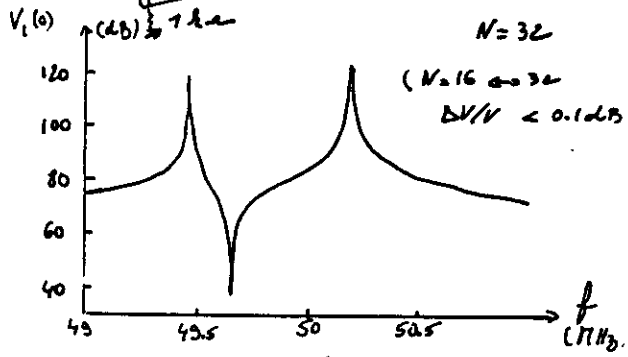
$R_c = \frac{L_{12}(z)}{L_{12}}$ ⇒ z_{max} such that $R_c <$ ref. level

(-20 dB, -40 dB, -60 dB... ??)

1st example: comparison with analytical models (perturbation method) → small angle between the conductors

Example: ∅ 8mm - L = 3m, 3 cm above a ground plane - Distance between the conductors, from 0.5 cm to 1.5 cm

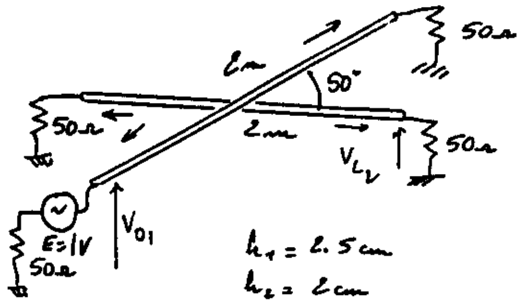




2 resonance frequencies $\begin{cases} 49.46 \text{ MHz} \\ 50.19 \text{ MHz} \end{cases}$

good agreement with analytical models: 49.5 MHz and 50.18 MHz.

(5)



$h_1 = 2.5$ cm

$h_2 = 2$ cm

$d_1 = d_2 = 0.18$ cm.

Transfer function V_{L2}/V_{01} at 1 MHz

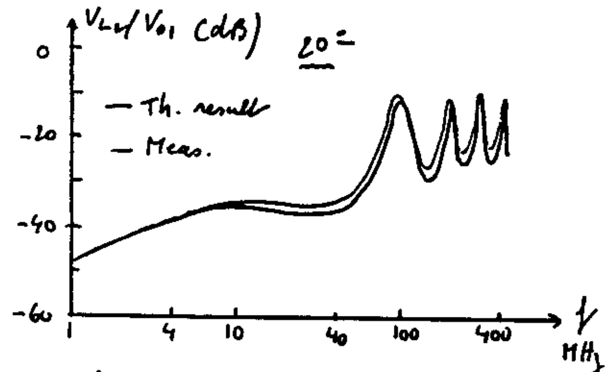
	R_c (dB)	Nb of elements	discretized length	V_{L2}/V_{01} (dB)
Reference	—	128	2 m	-59.6
	-60	128	1.6 m	-59.6
	-40	32	0.5 m	-60.1
	-20	16	0.14 m	-61.8

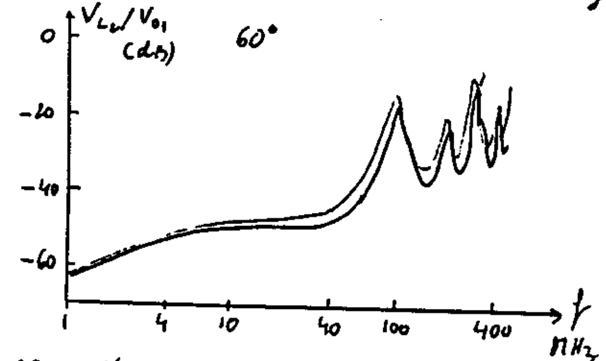
Same kind of result up to 20 MHz

Parametric study

Satisfactory result $\begin{cases} -40 \text{ dB} \to 32 \text{ elements} \\ -20 \text{ dB} \to 16 \text{ elements} \end{cases}$

Comparison between theoretical and exp. result





if $2\theta \to \pi/2$: capacitive coupling

→ lumped capacitance between the two lines

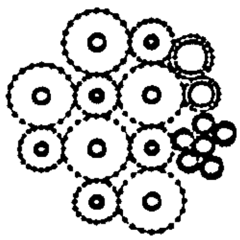
(6)

Characterization of multiconductor transmission line

- Experimental charact.
  - → Direct measurement of R, L, C coef.
  - → "Indirect" meas.

    Meas [S] par. → $[\Gamma], [Z_c] \rightarrow R, L, C$
- "Numerical" method

  [C] and [L] matrices from a code

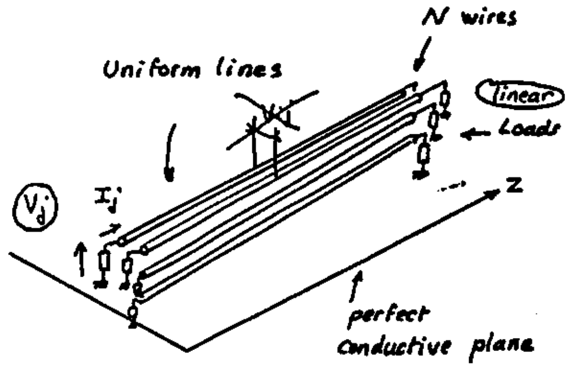


Pb: if $N \nearrow\nearrow$

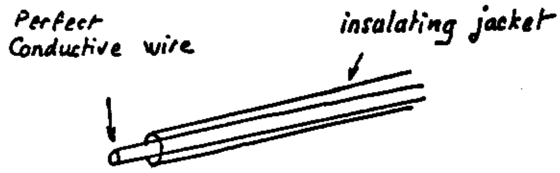
"statistical" approach (Usually number of different wires (ϕ, ε_r ...) is limited)

Meas. on a small number of cables ...

STRUCTURE OF THE CABLE BUNDLE

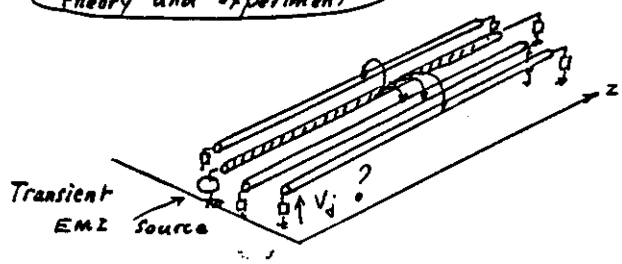


Structure of the wires

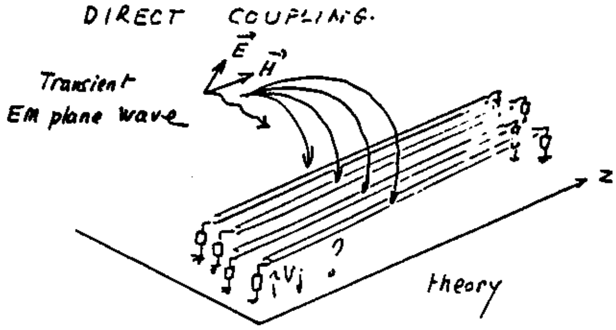


INDIRECT COUPLING

(theory and experiment)



DIRECT COUPLING



RIGOROUS MODAL THEORY

The transmission line equation may be solved in using

(L) and (C) matrices.

The coefficients of these matrices may be deduced either from measurements or computation by a finite elements code.

⇓

Structure of the modal propagation matrix

$$\left(e^{-\Gamma L}\right) = \begin{pmatrix} e^{-\Gamma_1 L} & & & & 0 \\ & e^{-\Gamma_2 L} & & & \\ & & \ddots & & \\ & & & e^{-\Gamma_k L} & \\ 0 & & & & \ddots \; e^{-\Gamma_N L} \end{pmatrix}$$

$\Gamma_1^2, \Gamma_2^2 \ldots \Gamma_k^2 \ldots \Gamma_N^2 \Longrightarrow$ eigen values

⇓

$\omega^2 (L)(C)$

SIMPLE PROCEDURE

(Approximation of the degenerate modes)

The coefficients of the (L) matrix are given by the usual analytical formulas.

$$L_{jj} = \frac{\mu_0}{2\pi} \operatorname{Log}\left(\frac{4h_j}{d_j}\right) \qquad L_{ji} = \frac{\mu_0}{4\pi} \operatorname{Log}\left(1 + \frac{4\, h_i h_j}{d_{ij}^2}\right)$$

d_i : wire diameters
h_i, h_j : height of the wires above the ground plan
d_{ij} : distance between the wires i and j

The coefficients of the (C) matrix are deduced in inverting (L) such :

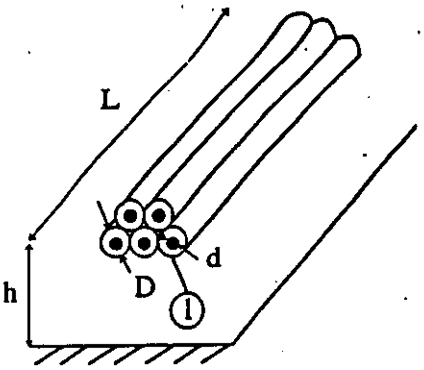
$$(C) = \frac{1}{v_c^2} (L)^{-1}$$

v_c : Corresponds to the velocity of the common mode propagated along the cable bundle.

$$v_c = \frac{3.10^8}{\sqrt{\varepsilon_{req}}}$$

ε_{req} : equivalent electric permitivity of an homogeneous dielectric medium around the wires.

GEOMETRICAL CONFIGURATION USED IN THIS STUDY



$L = 2.66$ m.
$D = 12.2$ mm → Diameter of the insulating jacket
$d = 7$mm → Diameter of the wire
h : height of the bundle above the ground plane

9

(L) and (C) given by measurements

$$(L) = \begin{pmatrix} 810 & 550 & 448 & 553 & 479 \\ 550 & 793 & 546 & 553 & 553 \\ 448 & 546 & 809 & 480 & 555 \\ 553 & 553 & 480 & 819 & 565 \\ 479 & 553 & 555 & 565 & 819 \end{pmatrix} \quad h = 100 \text{ mm.}$$

nH/m.

$$(C) = \begin{pmatrix} 52.1 & -21.8 & -2.2 & -21.8 & -1.2 \\ -21.8 & 79.2 & -21.6 & -16.9 & -15.9 \\ -2.2 & -21.6 & 54.7 & -1.3 & -22.9 \\ -21.8 & -16.9 & -1.3 & 66.1 & -21.9 \\ -1.2 & -15.9 & -22.9 & -21.9 & 66.4 \end{pmatrix}$$

pF/m.

Velocities of the eigen modes

$V_1 = 2.71\ 10^8$ m/s → Common mode
$V_2 = 2.35\ 10^8$ m/s
$V_3 = 2.22\ 10^8$ m/s
$V_4 = 2.16\ 10^8$ m/s.
$V_5 = 2.12\ 10^8$ m/s.

(L) given by the formula.

$$L_{jj} = \frac{\mu_0}{2\pi} \operatorname{Log}\left(\frac{4h_j}{d_j}\right) \qquad L_{ij} = \frac{\mu_0}{4\pi} \operatorname{Log}\left(1 + \frac{4 h_i h_j}{d_{ij}^2}\right)$$

$$(L) = \begin{pmatrix} 812 & 562 & 425 & 572 & 462 \\ 562 & 812 & 562 & 572 & 572 \\ 425 & 562 & 812 & 462 & 572 \\ 572 & 572 & 462 & 831 & 582 \\ 462 & 572 & 572 & 582 & 831 \end{pmatrix} \quad \frac{\Delta L}{L} < 5$$

(C) given by the relation ship.

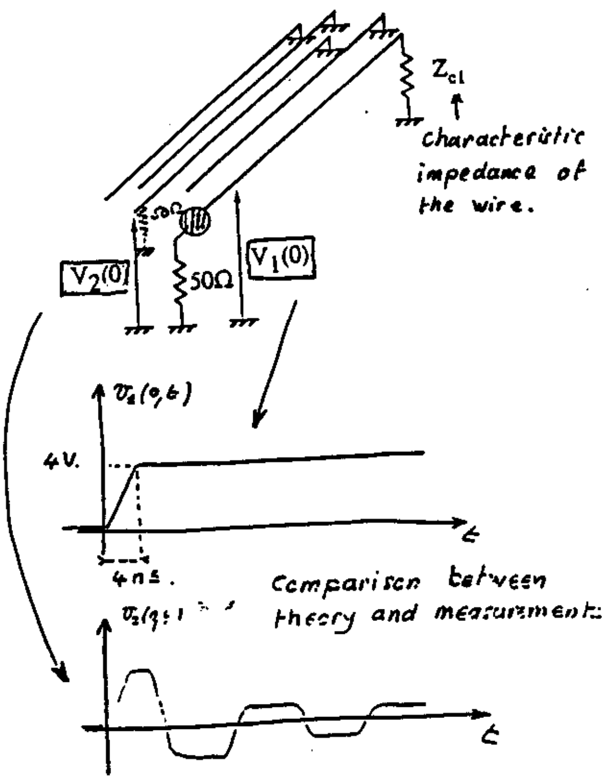
$$(C) = \frac{1}{v_c^2} (L)^{-1}$$

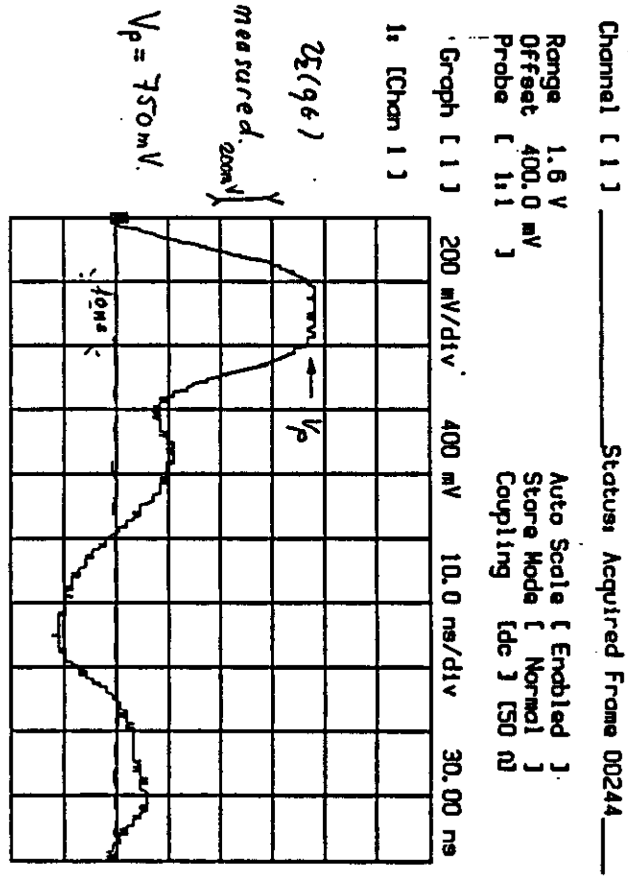
v_c Velocity of the common mode deduced from measurements.

$$(C) = \begin{pmatrix} 40.4 & -16.1 & -1.3 & -17 & \boxed{1.4} \\ -16.1 & 53.8 & -16.1 & -10 & -10 \\ -1.3 & -16.1 & 40.4 & \boxed{1.4} & -17 \\ -17 & -10 & \boxed{1.4} & 47.1 & -17.6 \\ \boxed{1.4} & -10. & -17. & -17.6 & 47.1 \end{pmatrix} \quad \frac{\Delta C}{C} <$$

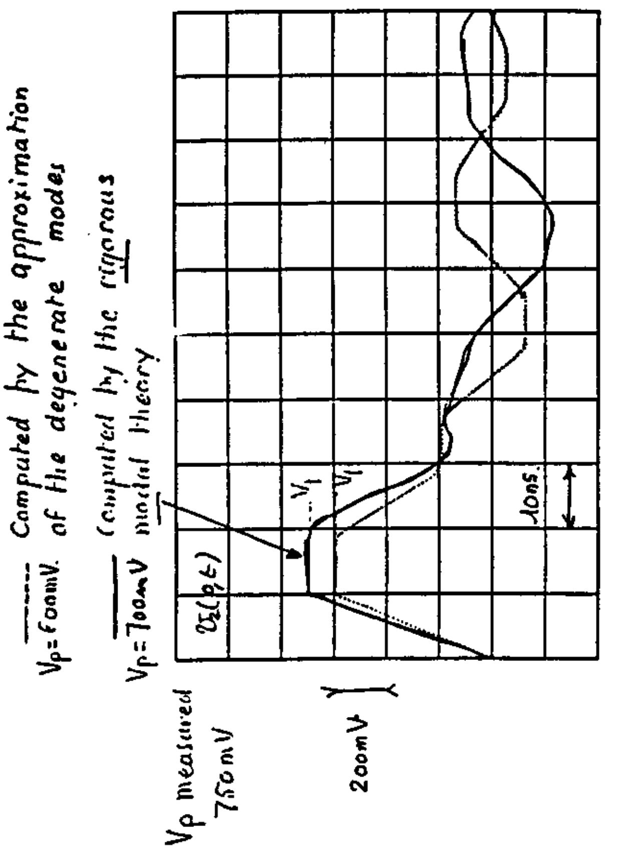
$\frac{\Delta C}{C} =$

TEST IN THE TIME-DOMAIN





(11)



RIGOROUS PROCEDURE

↓

OTHER SIMPLE PROCEDURE

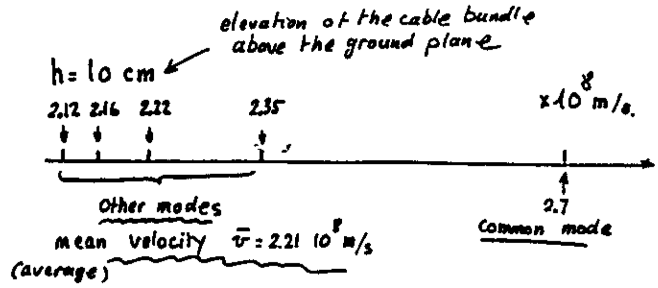
Quasi degenerate modes → Property of the modal velocities

↑

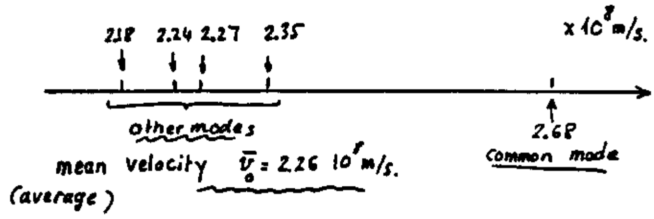
SIMPLE PROCEDURE using the approximation of the degenerate modes.

(16)

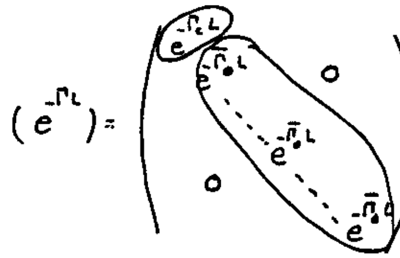
DISTRIBUTION OF THE MODE VELOCITIES



Cable bundle near the ground plane.



APPROXIMATION OF THE QUASI DEGENERATE MODES



$\bar{\Gamma}_c = j \dfrac{\omega}{v_c}$ v_c: velocity of the common mode

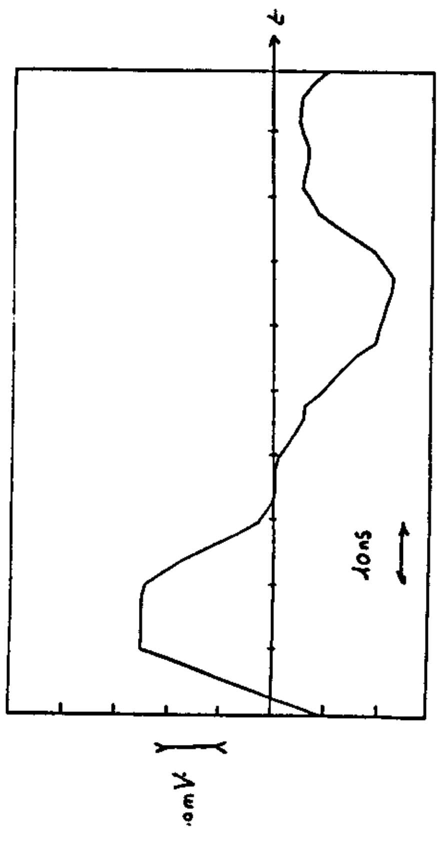
$\bar{\Gamma}_o = j \dfrac{\omega}{\bar{v}_o}$ $\bar{v}_o$: average velocity of the other modes

(L) ⇒ deduced from the usual analytical formulas

(C) ⇒ deduced from (L) and $(e^{-\Gamma L})$

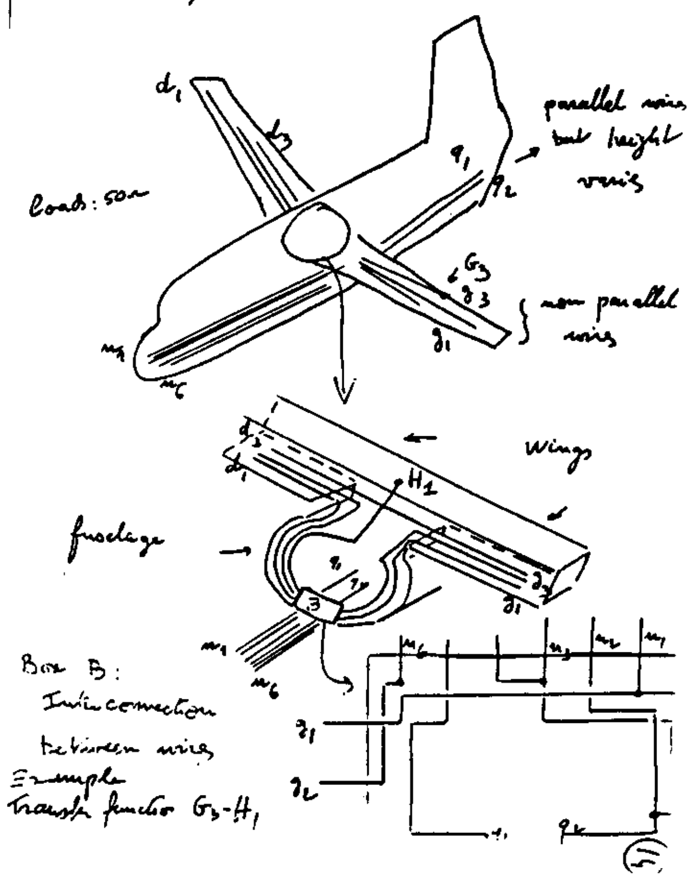
(5)

APPLICATION OF THE QUASI DEGENERATE MODES :
APPROXIMATION FOR THE COMPUTATION OF THE CROSSTALK VOLTAGES



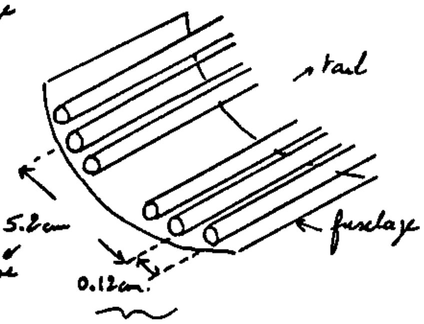
Experiment on a "Transall" airplane mock-up (1/10)

Diameter of the wires : 1.2mm



(5)

- Wings + tails → distance between the wires are >> diameter ⇒ Analytical expression for [L], [C]
- Fuselage

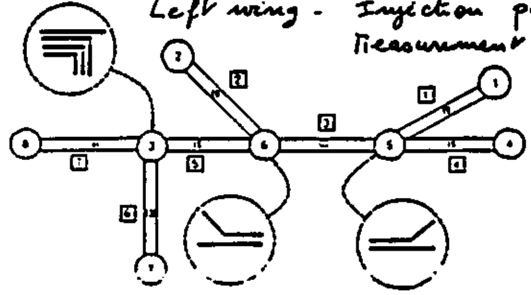


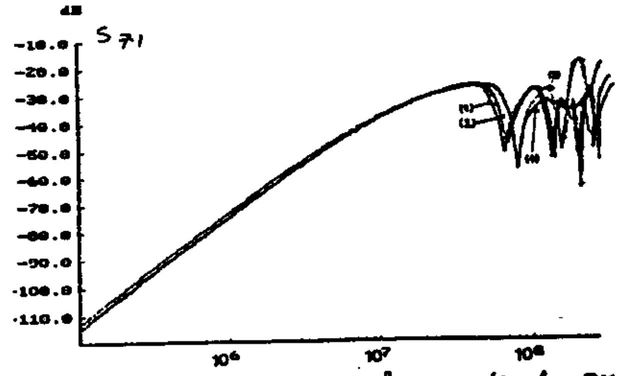
measurement of (L), (C) : line (3 wires) above a ground plane

- Each part has been tested separately.
- global test, 70 junctions
  "Wave" matrix $[W_0] \rightarrow 410$ elements.
- Box ? → "Black box" → Measurement of [S] (or) + analytical expression by neglecting the propagation effect inside the box

Influence of the meas. tech. of L - C

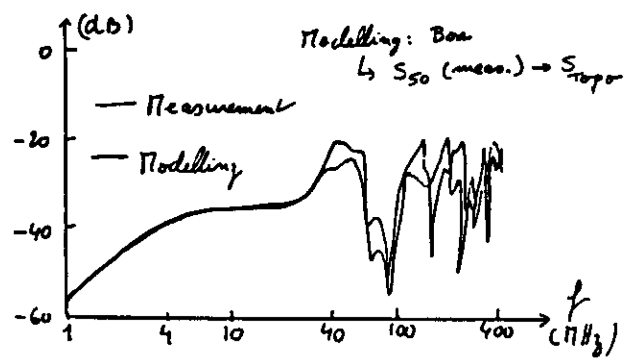
"Left wing" - Injection port 1
Measurement port 7



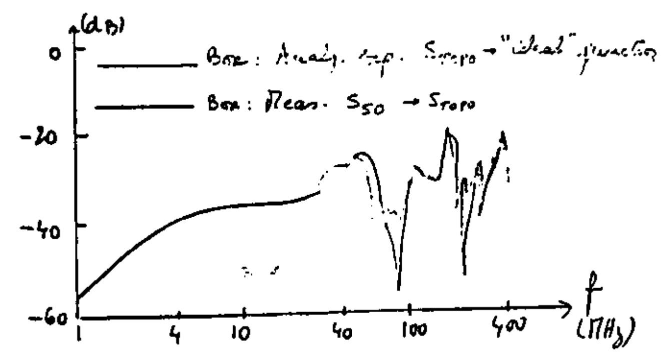


— Direct meas [L], [C] - Same network part above a ground plane
— Direct meas [L], [C] "in-situ"
— [L] and [C] deduced from [S] meas. "in-..."
— S_{71} measured

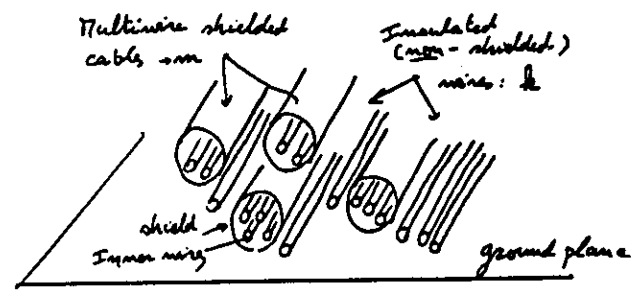
Transfer function G_3/H_1



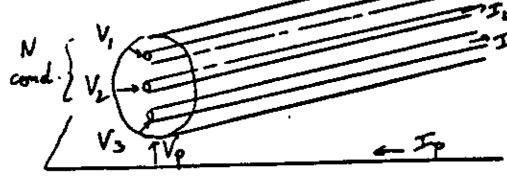
Modelling using analytical expression for S_{TOPO} (Box)
↑
S_{TOPO} deduced from S_{50} (meas.).



The general case:



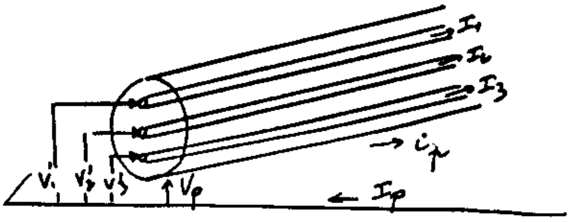
All the wires must be considered as belonging to the same tube - Multiwire shielded line ?



$$\begin{cases} -\dfrac{\partial (V)}{\partial z} = (Z)(I) - (Z_t) I_p \\ -\dfrac{\partial V_p}{\partial z} = -(Z_t)^t (I) + Z_{pp} I_p \end{cases}$$

$$(Z_t) = \begin{pmatrix} Z_{t1} \\ Z_{t2} \\ \vdots \\ Z_{tN} \end{pmatrix}$$ → "Common mode" transf. imped. - Measurement (usually $Z_{ti} \neq Z_{tj}$)

⇒ Not easy to introduce ...



Shield → additional wire ⇒ N+1 wires above ground plane

V's reference: ground plane ; $i_p = I_p - \sum_{k=1}^{N} I_k$

$$-\frac{\partial (V')}{\partial z'} = (Z')(I')$$

$$(V') = \begin{pmatrix} V'_1 \\ \vdots \\ V'_N \\ V_p \end{pmatrix} \text{ and } (I') = \begin{pmatrix} I_1 \\ \vdots \\ I_N \\ i_p \end{pmatrix}$$

$$(Z') = \begin{bmatrix} Z'_{11} & \cdots & Z'_{1N} & Z'_{1p} \\ Z'_{N1} & \cdots & Z'_{NN} & Z'_{Np} \\ Z'_{p1} & \cdots & Z'_{pN} & Z'_{pp} \end{bmatrix}$$

By identifying the two matrix relations

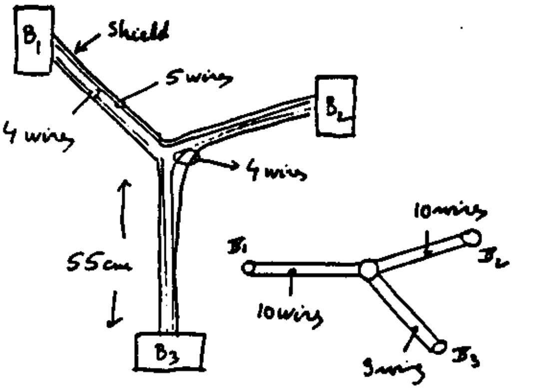
$$Z'_{jj} = Z_{jj} + Z_{pp} - 2Z_{tj} \qquad j = 1,N$$

$$Z'_{ij} = Z_{ij} + Z_{pp} - Z_{ti} - Z_{tj} \qquad \begin{cases} i = 1,N \\ j = 1,N \end{cases} \quad i \neq j$$

$$Z'_{jp} = Z_{pp} - Z_{tj} \qquad j = 1,N$$

$$Z'_{pp} = Z_{pp}$$

Other relations $Y' \leftrightarrow Y$

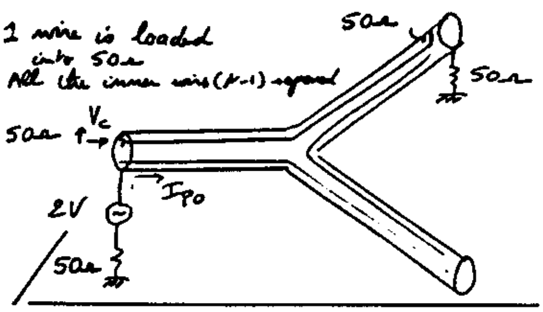


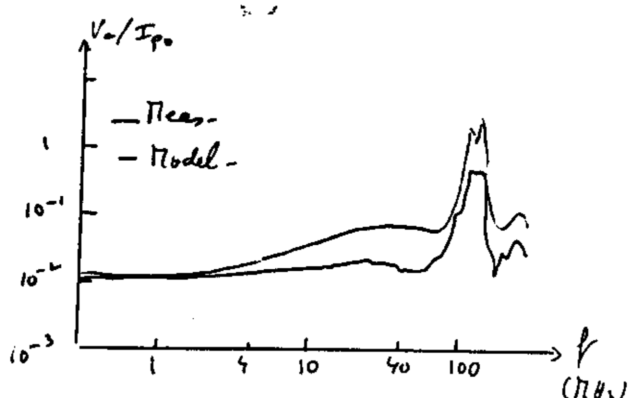
• 1st difficulty → "direct" measurement of Z_t
Assuming $Z_{ti} = R_{ti} + jL_{ti}\omega$ → impulse response on the total structure → R_t # 11 mΩ/m, L_t # 0.7 nH/m

• 2nd difficulty → (L) and (C) matrices → measurement.

7

• 1 wire is loaded into 50Ω
• All the inner wires (N-1) open





Difference → R_t, L_t

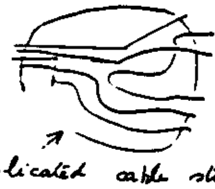
⇒ Much better agreement by measuring $Z_t(f)$ on a cable without branches.

Conclusion

• Topological approach (B.L.T eq) → interesting tool to predict the coupling to complicated wire structures

• Shielded + non shielded wires can be introduced

• Part of the structure can be replaced by its [S] matrix



Complicated cable structure → "Black box" N_p (Valid if there is no external source term inside the box).

8

Coaxial cable and

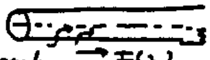
Multiwire shielded lines

—

Transfer impedance

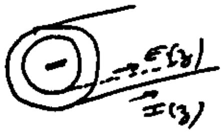
COAXIAL CABLE

27



Transfer $\rightarrow I(z)$ ⇒ disturbing voltage

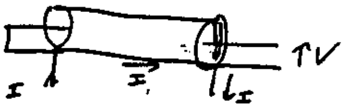
Case of a "concentric tube"



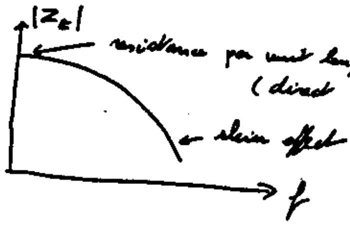
$$Z_t = E(z) / I(z)$$

Transfer impedance

other possible definition



$$Z_t = \frac{V}{I} \cdot \frac{1}{L}$$



Analytical formulas (radii, conductivity ...)

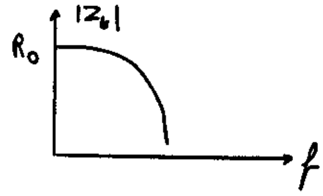
14

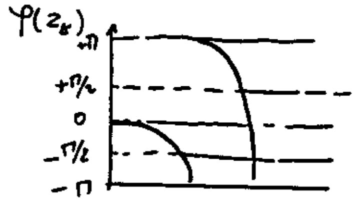
Thickness of the shield e

Diameter : D

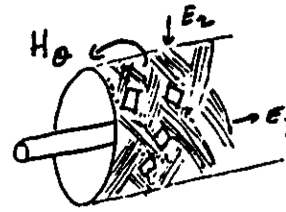
$$e \ll D \Rightarrow Z_t = R_0 \frac{j k_2 e}{sh(j k_2 e)}$$

$$k_2 = \sqrt{j \omega \mu_2 \sigma_2}$$





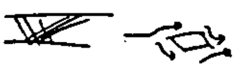
SHIELD → BRAIDS



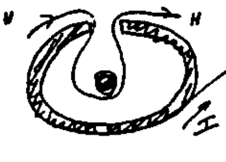
E_z, H_θ ⇒ DIFFRACTION

braided cables, helices ...

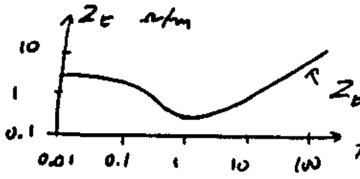
⇒ Propagation of the external current no longer along the z axis



⇒ Diffraction by the small distributed apertures

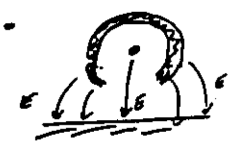


⇒ Magnetic coupling



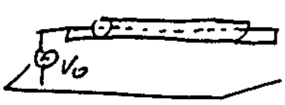
transfer inductance

$Z_t = L_t \omega$



Direct coupling of the electric field through the apertures

⇒ Transfer admittance



$$Y_t = \frac{1}{V_0} \frac{dI}{dz} = j\omega C_t$$

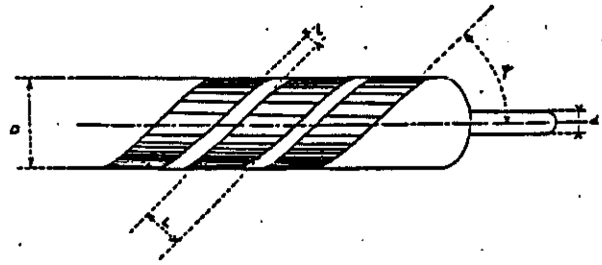
II-6

HELICOIDAL TAPES

RUBANS HELICOIDAUX

N RUBANS AVEC LA MÊME ORIENTATION

N TAPES WITH THE SAME WEAVE ANGLE ψ



$$Z_t = Z_{td} + j L_t \omega$$

DIFFUSION — INDUCTION COMPOSANTE MAGNÉTIQUE AXIALE — Induction due to the axial mg. component

$$Z_{td} = R_0 \frac{(1+j)\frac{1}{\delta}}{sh\left[(1+j)\frac{1}{\delta}\right]}\left[1 + tg^2\,\psi\; ch\left[(1+j)\frac{1}{\delta}\right]\right]$$

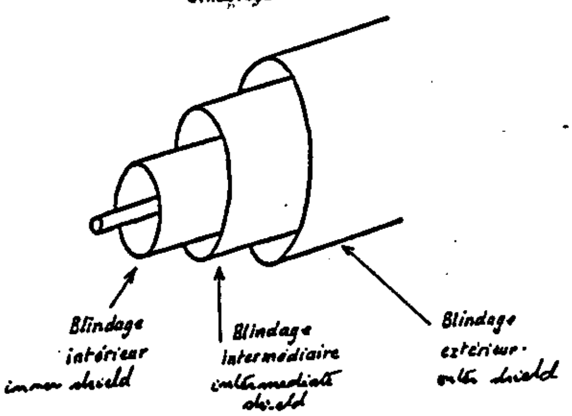
R_0 RÉSISTANCE LINÉIQUE DU RUBAN ; Resistance per unit length of each tape

$$L_t = \frac{\mu_0}{4\pi}\left[\left(1 - \frac{d^2}{D^2}\right) tg^2\,\psi + \frac{1}{N}\left(\frac{1}{2d}\right)^2 \frac{1}{\cos\psi}\right]$$

45

HIGH IMMUNITY SHIELDED CABLES
CABLES BLINDES A HAUTE IMMUNITE

MULTI-SHIELDS STRUCTURES
STRUCTURE MULTICOUCHES



MECANISMES DE COUPLAGE
Coupling mechanism

INFLUENCE DES CONNEXIONS ENTRE BLINDAGES
Influence of the connection between the shields

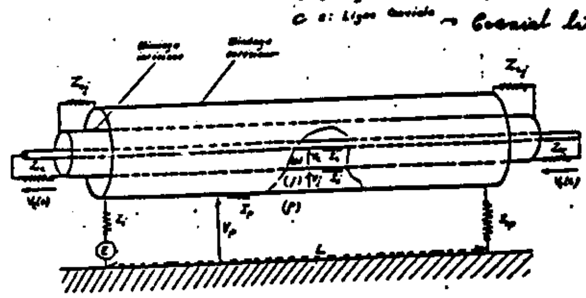
STRUCTURE À DEUX BLINDAGES
two shields

ACTION DES FERROMAGNETIQUES
ACTION DES COMPOSITES
Composite + Ferromag materials

STRUCTURE À TROIS BLINDAGES
three shields

CABLE A 2 BLINDAGES
2 SHIELDS

P : Ligne perturbatrice → Disturbing line
J : Ligne intermédiaire → "Intermediate"
C : Ligne coaxiale → Coaxial line



LES PARAMETRES LES PLUS INFLUENTS SONT :
- The most important parameters are :

L'IMPEDANCE DE TRANSFERT DU BLINDAGE EXTERIEUR Z_{te}
- The transfer impedance of the outer shield

L'IMPEDANCE DE TRANSFERT DU BLINDAGE INTERIEUR Z_{ti}
- Transfer impedance of the inner shield

LE MODE DE CONNEXION DES BLINDAGES Z_{0j} ET Z_{Lj}
- Impedances connected between the shield.

LES CARACTERISTIQUES DE LA LIGNE INTERMEDIAIRE.
- Characteristics of the intermediate line

47

EXPRESSIONS APPROCHEES DE $V_c(0)$ ET $V_c(L)$
Approximate expression of $V_c(0)$ and $V_c(L)$
VALABLES AUX GRANDES LONGUEURS D'ONDE
valid at low frequency
$\lambda >> L$

- BLINDAGES SUR COURTS-CIRCUITS $Z_{0j} = Z_{Lj} = 0$
  Shields short-circuited at both ends

$$V_c(0) = V_c(L) = \frac{1}{2}\frac{Z_{te}\,Z_{ti}}{jL_j\,\omega}\,I_p(0)\,L$$

L_j INDUCTANCE LINEIQUE DE LA LIGNE INTERMEDIAIRE
series inductance per unit length of the intermediate line

- BLINDAGES ADAPTES $Z_{0j} = Z_{Lj} = Z_{cj}$
  Shields terminated on matched loads

$$V_c(0) = V_c(L) = \frac{1}{4}\frac{Z_{te}\,Z_{ti}}{Z_{cj}}\,I_p(0)\,L^2$$

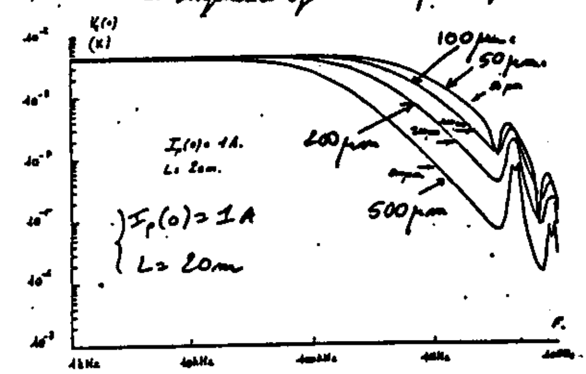
Z_{cj} IMPEDANCE CARACTERISTIQUE DE LA LIGNE INTERMEDIAIRE

- BLINDAGES OUVERTS $Z_{0j} = Z_{Lj} = \infty$
  open-ended shields

$$V_c(0) = V_c(L) = \frac{Z_{te}\,Z_{ti}}{12}\,j\,C_j\,\omega\,L^3$$

C_j CAPACITE LINEIQUE DE LA LIGNE INTERMEDIAIRE

INFLUENCE DE LA PROPAGATION ET
DE L'ESPACEMENT DES BLINDAGES
Influence of the propagation between the shields
⇒ Influence of their spacing



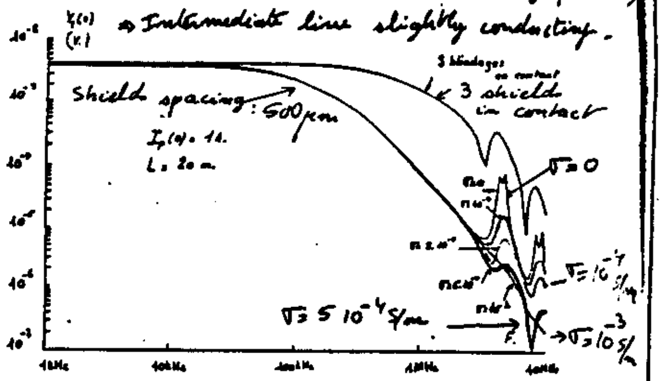
→ Theoretical curves with a 3 shields cable (ohic copper)
MODELISATION OBTENUE AVEC UN CABLE À 3 ECRANS HOMOGENES EN CUIVRE ESPACES D'UN DIELECTRIQUE.
→ between each shield: dielectric material.

LES BLINDAGES SONT COURT-CIRCUITES
→ Shields short-circuited at both ends → resonances

LES RESONANCES SONT DUES AUX REFLEXIONS SUR LES COURTS-CIRCUITS

ACTION D'UN DIELECTRIQUE CHARGE
INTERPOSE ENTRE LES BLINDAGES

Influence of a "loaded" material
⇒ Dielectric medium + small conducting particles
⇒ Intermediate line slightly conducting.



CÂBLE À 3 ÉCRANS HOMOGÈNES - 3 shields.

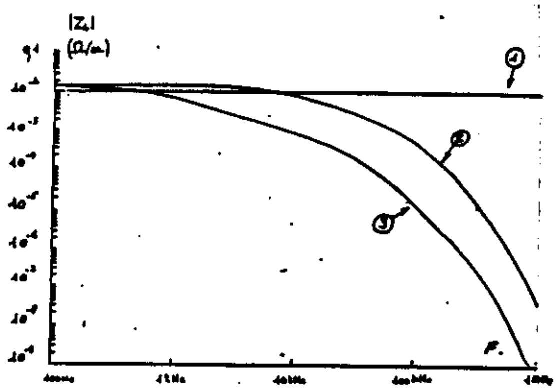
L'ESPACEMENT DES BLINDAGES EST DE 500 µm

LA CONDUCTIVITÉ DU COMPOSITE VARIE ENTRE 10^{-4} mho/m ET 10^{-3} mho/m

À 10^{-3} mho/m LES RÉSONANCES SONT BIEN ATTENUÉES

ACTION DES ECRANS FERROMAGNETIQUES
A HAUTE PERMEABILITE MAGNETIQUE

Ferromagnetic screens
-high magnetic permeability-



1 { 3 ÉCRANS EN CUIVRE ET EN CONTACT
3 screens (copper) in contact

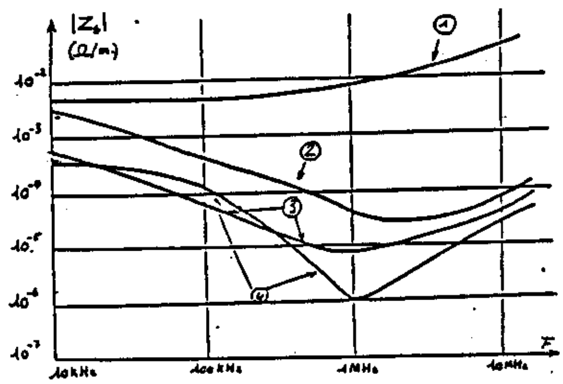
2 { ÉCRANS EXTÉRIEUR ET INTÉRIEUR EN MUMÉTAL $\mu_r = 20\ 000$
outer and inner screen µmetal

3 { ÉCRAN INTERMÉDIAIRE EN MUMÉTAL $\mu_r = 20\ 000$
intermediate screen: mumetal

NB: Variation $\mu_r(f)$

EFFICACITES COMPAREES DE COAXIAUX
A HAUTE IMMUNITE

Experimental results on high immunity cables



MESURES FAITES SUR ÉCHANTILLONS L = 2 m.

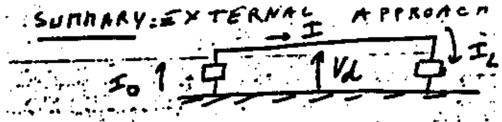
1 - TRESSE CUIVRE Copper braid

2 - TRESSE CUIVRE - DIÉLECTRIQUE - TRESSE CUIVRE
Copper braid - Dielectric mat. - Copper braid

3 - TRESSE CUIVRE - COMPOSITE FERROMAGNÉTIQUE - TRESSE CUIVRE
Ferromag mat.

4 - QUATRE TRESSES CUIVRE EN CONTACT.
4 copper braids in contact

SUMMARY: EXTERNAL APPROACH



$$V^d = \int_h^0 E_x^d \, dx \quad (\text{due to } I)$$

$$\begin{cases} \dfrac{\partial V^d}{\partial z} + Z I = \underbrace{E_z^c(h)} \\ \dfrac{\partial I}{\partial z} + Y V^d = 0 \end{cases}$$

Appr boundary conditions

$$-Z_0 I_0 + \underbrace{\int_0^h E_x^c(0)\, dx} = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=0}$$

$$Z_L I_L + \underbrace{\int_0^h E_x^c(L)\, dx} = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=L}$$

If resonance → the amplitudes of the current will be appr

Special case: infinitely long wire

→ Incident plane wave

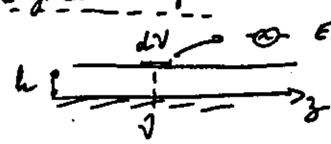
→ No boundary conditions

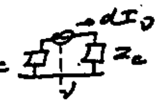
$$E_z^c(h) = E_{z0}^c(h)\, e^{j\omega t - k_z z}$$

Prop. constant of the incident wave along the z axis

$$\Rightarrow I(z) = \int_{-\infty}^{+\infty} \frac{E_{z0}^c(h)\, e^{-k_z \nu}}{2 Z_c} e^{-\gamma_p |z-\nu|} \, d\nu$$

Physical interpretation



Infinite line ⇒ Z_c  Z_c

$dI(\nu) = E_z^c(h)\, d\nu / 2 Z_c$

↳ at a point z → $dI_\nu\, e^{-\gamma_p (z-\nu)}$

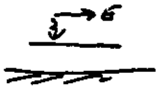
$\int_{-\infty}^{+\infty}$ ⇒ Σ of all the eq. generators

Plane wave: $E_{z0}^c(h) = $ cte

(4)

$$I(z) = \frac{E_{z0}^c(h)\, e^{-k_z z}}{Z_c} \; \frac{\gamma_p}{\gamma_p^2 - k_z^2}$$

Simultaneous illumination

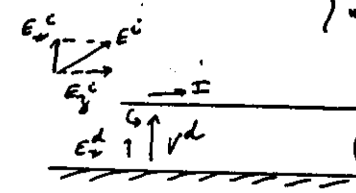


$\Rightarrow k_z = 0$

$$\boxed{I(z) = \frac{E_{z0}^c(h)}{\gamma_p Z_c} = \frac{E_{z0}^c(h)}{Z}}$$

→ order of magnitude for finite lengths

own positive approach: "magnetic coupling" / "electric coupling"

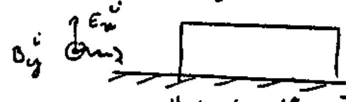


$E_x^c + E_x^d = E_x^t \Rightarrow V^t = \int_h E_x^t$

$$\begin{cases} \dfrac{\partial I}{\partial z} + Y V^t = -Y \int_0^h E_x^c\, dx & \to \text{current source} \\ \dfrac{\partial V^t}{\partial z} + Z I = j\omega \int_0^h B_y^c\, dx & \to \text{voltage source} \end{cases}$$

Source $E_x^c + B_y^c$ instead of E_z^c !!

Magnetic coupling — Electric coupling



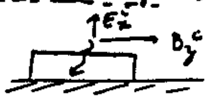
"hybrid" coupling

Instead of 1 coupling to the vertical part or (and) to the horizontal part of the line

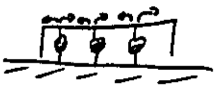
(∞)

Different results depending on the approach?? (2)

Electric coupling



$$\frac{\partial I}{\partial z} + Y V^t = -Y \int_0^h E_x^i \, dx$$



→ Boundary condition + solution

only $E_x^i(0)$ and $E_x^c(L)$

coupling to the vertical part of the line!

⇒ Whatever the configuration – same results (Maxwell eq! Relation between B_y and E_x, E_z!

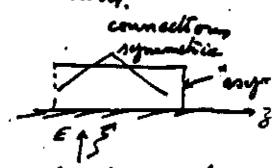
Interest of this approach → B_y^i is easily measurable at low frequency – It is not the case for E_x^i

but the previous approach can be easily extended to other shapes (antenna theory: source terms

Simple analytical formulas (short cables $L \ll \lambda$)

Example: simultaneous illumination.

Coupling to the horizontal part



connections symmetric

"asym"

Coupling to the ver[tical part]

	Coupling to the horizontal part	Coupling to the vertical part
Sym. conn. (2 pts)	$\frac{E_z^i(h)}{2}$ ①	$-j\frac{E_x^i h}{Z_c}\frac{2\pi}{\lambda}\left(\frac{L}{2}-z\right)$
Asym. 1 pt	$\frac{E_z^i(h)}{2}\left(\frac{2\pi}{\lambda}\right)^2\left(\frac{z}{2}-L\right)z$	$-j\frac{E_x^i h}{Z_c}\frac{2\pi}{\lambda}z$ ③

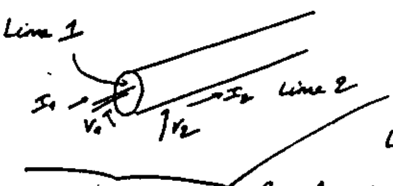
① eq. gen. $E_z^i(h) \cdot L$ load → Z_L → independent of L

② 2 eq. gene. on both sides of the line → $I = 0$ at the center

③ 1 eq. voltage generator connected to an open end transmission line —

(6)

Illuminated coaxial cable → system of 2 coupled lines (7)



Line 1 { series imp. Z_{11}, Y_{11}

Line 2 { Z_{22}, Y_{22}

Coupling terms Z_t, Y_t

$$\begin{cases} \frac{dV_1}{dz} + Z_{11} I_1 = Z_t I_2 \\ \frac{\partial I_1}{\partial z} + Y_{11} V_1 = -Y_t V_2 - Y_t \int_0^h E_x^{it}(z)\, dx \end{cases}$$

$$\begin{cases} \frac{dV_2}{dz} + Z_{22} I_2 = Z_t I_1 + j\omega \int_0^h B_y^{it}(z)\, dx \\ \frac{dI_2}{dz} + Y_{22} V_2 = -Y_t V_1 - Y_{22} \int_0^h E_x^{it}(z)\, dz \end{cases}$$

Usually Y_t negligible + well shielded appr. → simple formulas. If the coaxial cable is connected onto its charact. impedance

$$\begin{cases} V_1(0) = \frac{1}{2} Z_t \int_0^L I_2(z)\, e^{-\gamma_1 z}\, dz \\ V_1(L) = \frac{1}{2} Z_t\, e^{-\gamma_1 L} \int_0^L I_2(z)\, e^{\gamma_1 z}\, dz \end{cases}$$

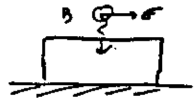
$\{ V \propto Z_t$

→ ... → ... Z_t = ...

Connection	Coupling to horizontal part	Vertical part
2 pts (sym.)	$-\frac{E_z^{ip} h}{Z_{cp}} Z_t L$	$\frac{j E_x^{ip} h}{24 Z_{cp}}\left(\frac{2\pi L}{\lambda}\right)^2 \sqrt{E_{z^i}}\, Z_t$
1 pt asym.	$\frac{E_z^{ip} h}{3 Z_{cp}}\left(\frac{2\pi L}{\lambda}\right)^2 Z_t L$	$-\frac{j E_x^{ip} h}{4 Z_{cp}}\frac{2\pi L}{\lambda} Z_t L$

$V(0)$ or $V(L) \propto L$ only for 1 config. —

At V.L.F → most worst case



→ "Loop"

Long cables → resonance + beating

(8)

Current distribution

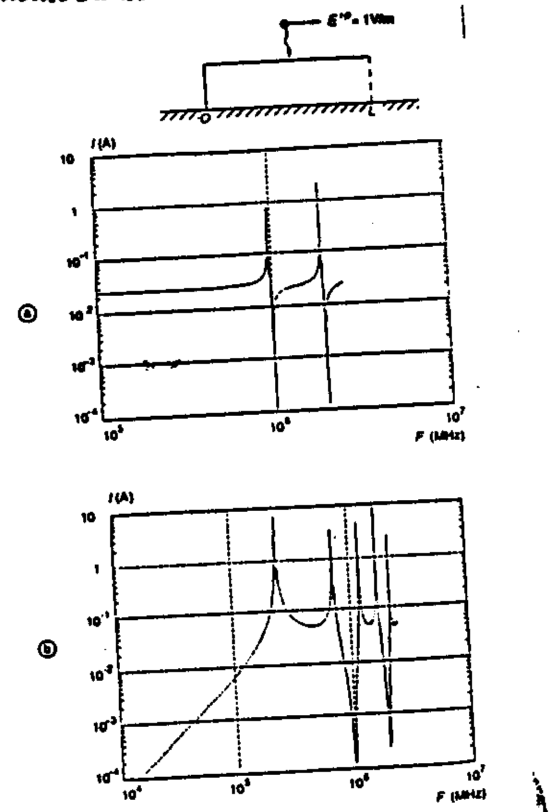


Fig. 7.14 — Variation d'un courant à l'extrémité d'une ligne de 300 m située à 6 m au-dessus d'un plan de masse. Couplage à la partie horizontale.
a) Ligne court-circuitée aux deux extrémités.
b) Cas d'une seule connexion de masse en z = 0.

Disturbing voltage at the end of the coaxial cable

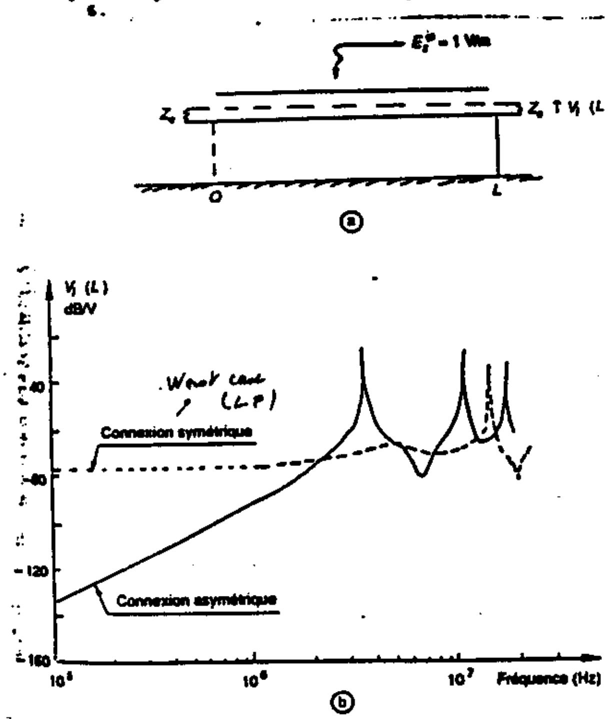


Fig. 7.17 — a) Illumination du câble coaxial : Couplage au fil horizontal:
b) Variation de $V_i(L)$ en fonction de la fréquence pour deux types de connexion.

L = 20 m, h = 40 cm

$Z_c = R_0 + j L_t \omega$ $\quad R_0 = 5$ mΩ/m, $L_t = 0.5$ nH/m

Disturbing voltage

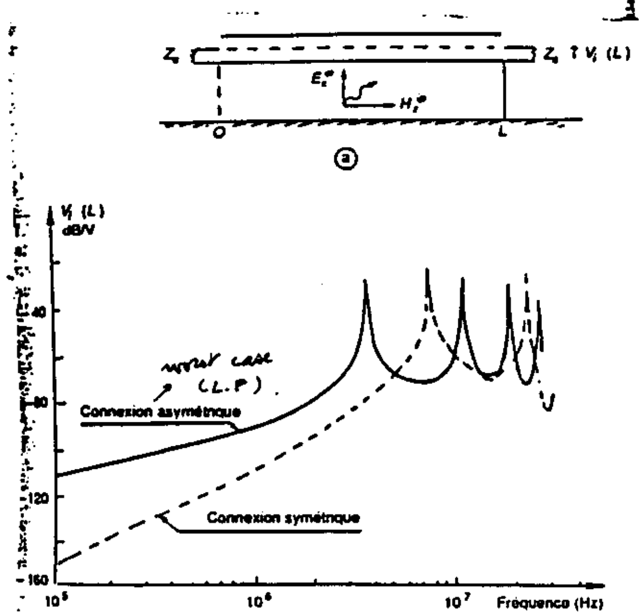
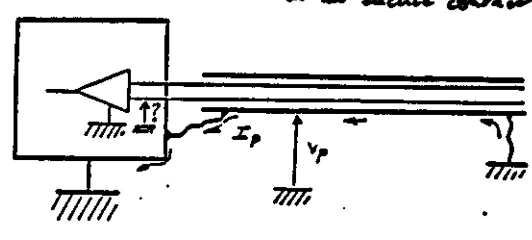


Fig. 7.18 — a) Illumination du câble coaxial : Couplage aux fils verticaux:
b) Variation de $V_i(L)$ en fonction de la fréquence.

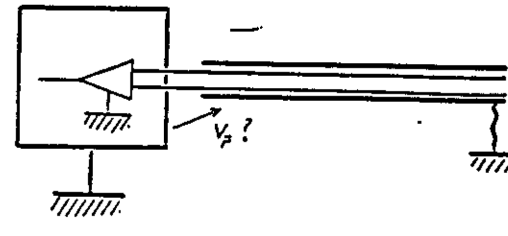
⇒ Usually connection at BOTH ends.
(except in some special cases when the low frequency source has been identified and is the only one)

LES DISCONTINUITES DE BLINDAGE
SHIELD DISCONTINUITY

DISCONTINUITE AVEC CONTACT ELECTRIQUE
with an electric contact



DISCONTINUITE SANS CONTACT ELECTRIQUE
without

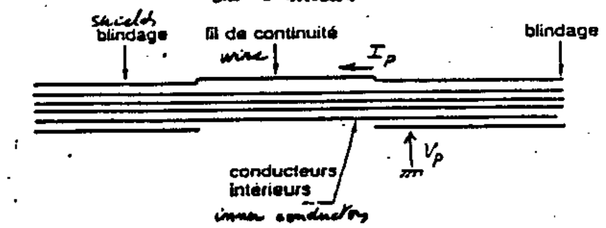


Modelization of a "discontinuity"

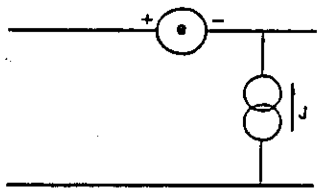
MODELISATION D'UNE DISCONTINUITE AVEC CONTACT ELECTRIQUE

with an electric contact between the 2 shields

48 (53)



- LA DISCONTINUITÉ AGIT COMME UN GÉNÉRATEUR DE F.E.M. ET UN GÉNÉRATEUR DE COURANT
- The discontinuity acts as a voltage and a current generator.



$e = (R_0 + j L_t \omega)\, I_p$ — R_0 RESISTANCE DE LA DISCONTINUITE

$J = j C_t \omega V_p$ — L_t INDUCTANCE DE LA DISCONTINUITE

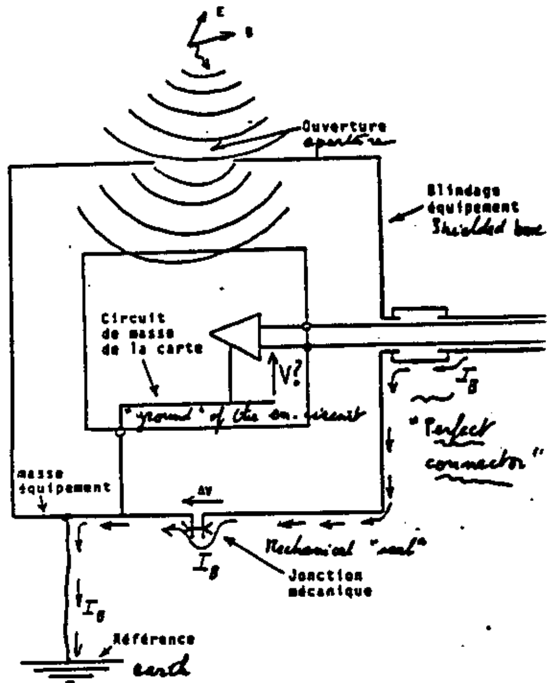
generally, the influence of C_t is negligible

EN GÉNÉRAL LA CAPACITÉ DE TRANSFERT C_t DE LA DISCONTINUITÉ N'INTERVIENT PAS (COUPLAGE PAR LE COURANT DOMINANT)

ENCEINTE BLINDEE RELIEE A LA REFERENCE DE MASSE

Shielded box with a ground connection

40 (5)



(3)

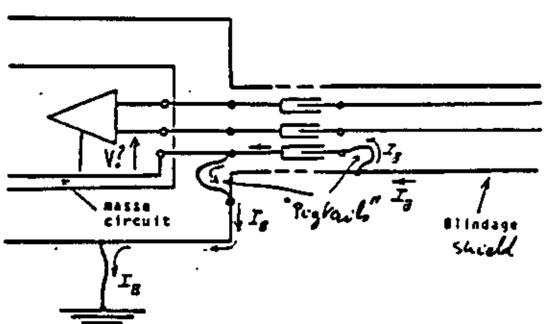
BLINDAGE DES CONNECTEURS

SHIELDING OF A CONNECTOR

49 (55)

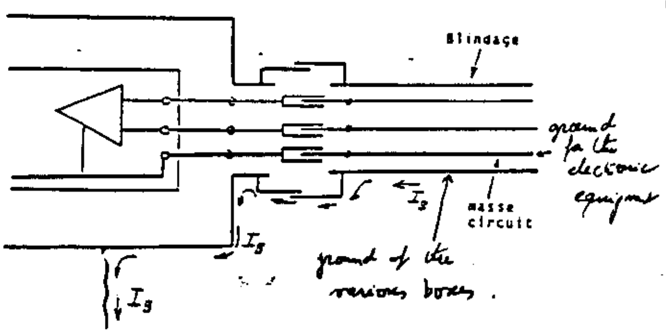
CONNECTEUR USUEL

USUAL CONNECTOR



CONNECTEUR BLINDE

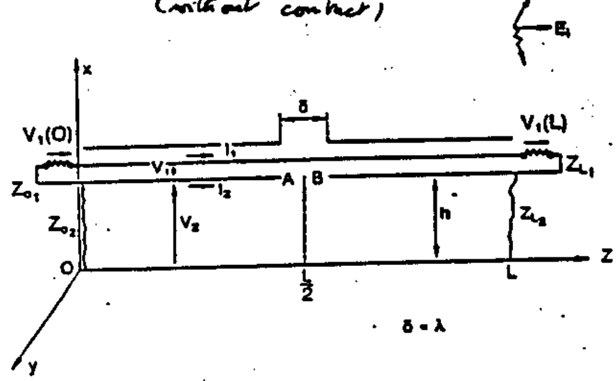
SHIELDED CONNECTOR



MODELISATION THEORIQUE D'UNE DISCONTINUITE DE BLINDAGE SANS CONTACT

(4)

Theoretical model of a shield discontinuity (without contact)



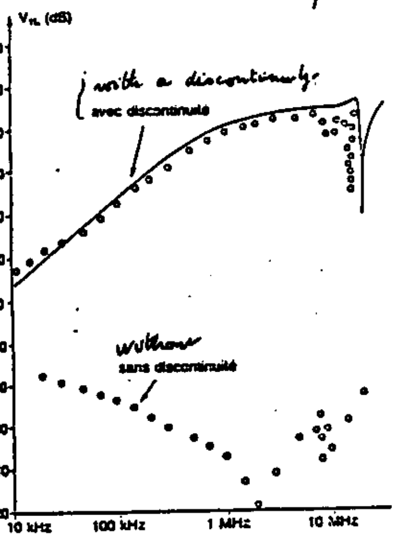
- IL FAUT TOUT D'ABORD ÉVALUER LA TENSION INDUITE PAR L'ONDE INCIDENTE SUR LA DISCONTINUITÉ V_{AB}
- 1st step: determine the induced voltage V_{AB}
- ENSUITE TRANSFÉRER CETTE TENSION SUR LES CHARGES (ÉQUIPEMENTS SOIT Z_{01} ET Z_{L1} PUIS ÉVALUER $V_1(0)$ ET $V_1(L)$
- 2nd step, Determine the current inside the cable, then $V_1(0)$ and $V_1(L)$
- L'AMPLITUDE DE $V_1(0)$ ET $V_1(L)$ DÉPEND DE LA NATURE DE Z_{01} ET Z_{L1}
- ($V_1(0)$ and $V_1(L)$ will depend on the loads Z_{01} and Z_{L1}

(5)

COMPARAISON THEORIE EXPERIENCE

(DISCONTINUITE SANS CONTACT)

Theor. and experimental results



Travelling disturbing wave (along the cable axis)
ONDE PROGRESSIVE SE PROPAGEANT PARALLELEMENT AU CABLE

$E_0 = 1$ V/m

DISTANCE DES EQUIPEMENTS

L = 15 m
Cable length

DISCONTINUITE d = 5 cm

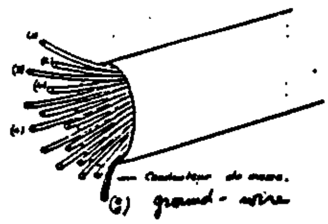
$Z_{01} = Z_{L1} = 50\ \Omega$

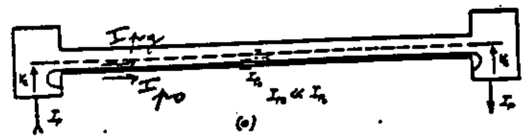
COAXIAL BLINDE

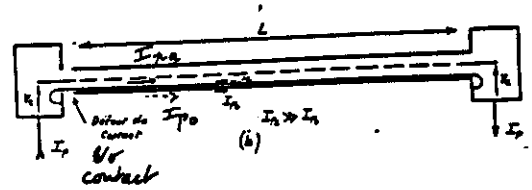
- $V_1(0)$ ET $V_1(L)$ SONT PRATIQUEMENT INDEPENDANTES DE d
  $V_1(0)$ and $V_1(L)$ are nearly independant on δ
- Z_{01} ET Z_{L1} CORRESPONDENT AUX IMPEDANCES DE CHARGE DU COAXIAL
- EN B.F. LES TENSIONS SONT PROPORTIONNELLES A f ENSUITE ELLES SONT INVARIANTES AVEC f LORSQUE LA REACTANCE DE LA BOUCLE DE MASSE EST SUPERIEURE A 50 Ω
  At Low Freq, $V \propto f$ and then tends to a limit value when the impedance of the ground loop is greater than 50 Ω

ROLE JOUE PAR LE FIL DE CONTINUITE DE MASSE

Influence of the ground-wire





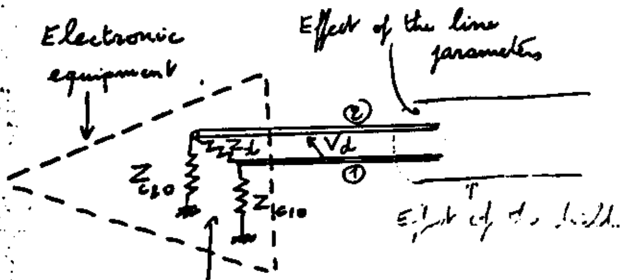


A) LE BLINDAGE DU CABLE EST CONTINU LE COURANT PERTURBATEUR EST ECOULE PAR LE BLINDAGE
Continuous shield → I_{pg} ≪

B) LE BLINDAGE EST DISCONTINU LE COURANT PERTURBATEUR EST ECOULE PAR LE FIL DE MASSE. IL PEUT EN RESULTER DES TENSIONS PARASITES DE GRANDE AMPLITUDE.
DISCONTINUITY → all the disturbing current flows locally on the ground wire → Induced current on the other conductors

Multifilar → Bifilar

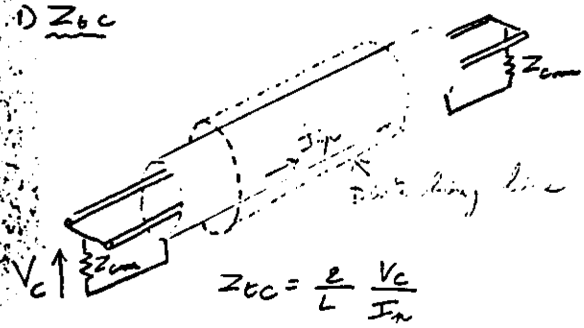
TWO-WIRE SHIELDED LINE



Effect of the input impedance of the equipment

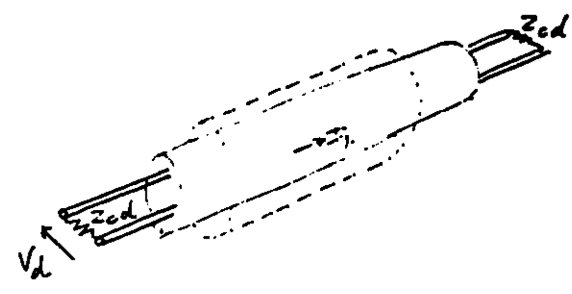
Z_t → Z_{tc} → common mode
Z_t → Z_{td} → differential mode ⇒ Measurement

1) Z_{tc}



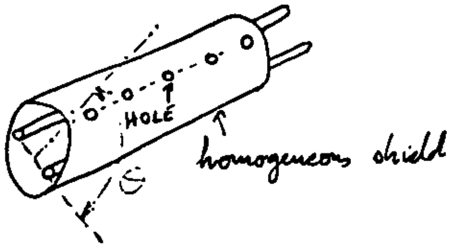
$$Z_{tc} = \frac{2}{L}\,\frac{V_c}{I_p}$$

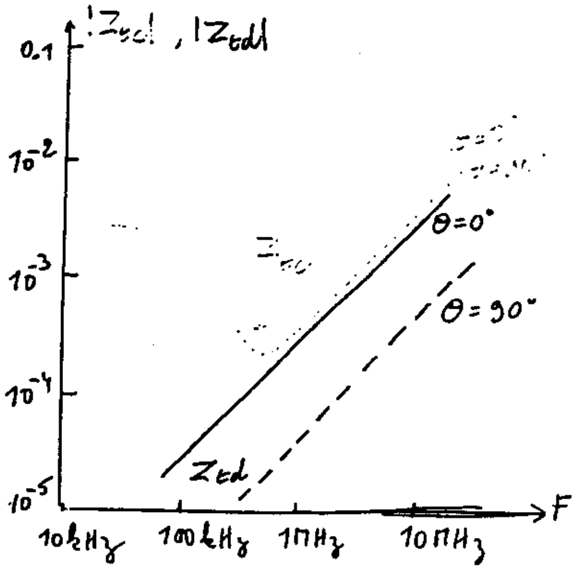
Measurement Z_{td}



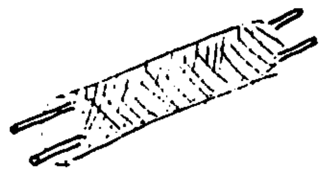
$$Z_{td} = \frac{2}{L}\,\frac{V_d}{I_p}$$

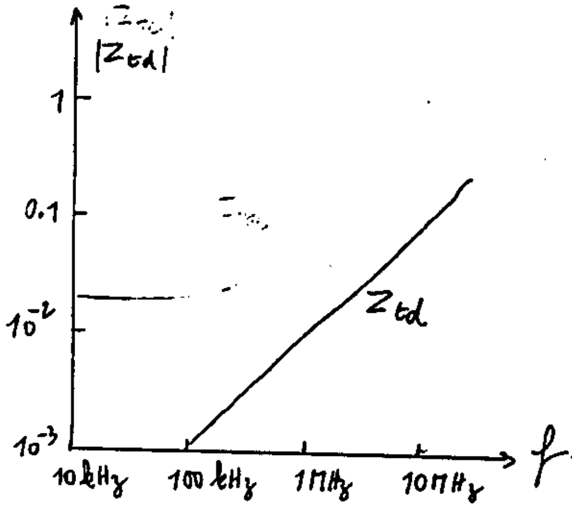
Effect of the shield dissymetry





Measurement on a braided shield





(1)

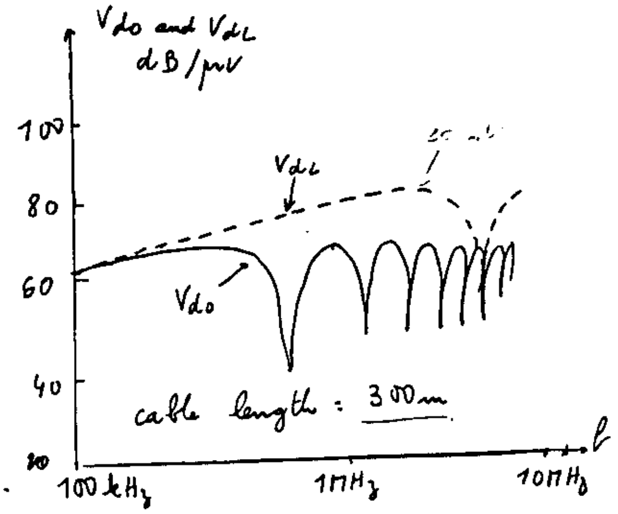
What is (are) the most important parameter(s):
Dissymetry of the shield, of the load impedance, or of the inner lines parameters?

⇒ Theoretical approach!

1) Simulation of the dissymetrical shield

$Z_t = j L_t \omega \rightarrow L_{t1} = 1.3\,nH/m$, $L_{t2} = 1.2\,nH/m$

- bifilar line - symmetric
- common mode input impedance - equal ...
- bifilar mode: matched



2) Dissymetrical load impedances

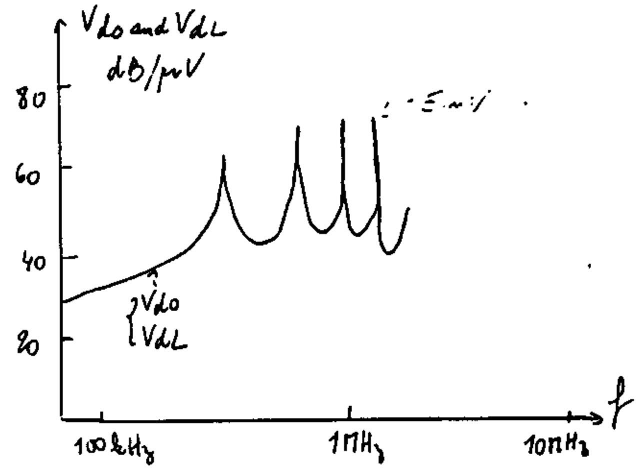
$Z_{c1L} = 10\,k\Omega$, $Z_{c2L} = 11\,k\Omega$

$Z_{c10} = 10\,k\Omega$, $Z_{c20} = 10\,k\Omega$

Symmetrical shield: $L_{t1} = L_{t2} = 1.2\,nH/m$

$(L_{td} = L_{t1} - L_{t2} = 0)$

Symmetrical bifilar line $\left(\begin{matrix} L_{11} = L_{22} \\ C_{11} = C_{22} \end{matrix}\right)$



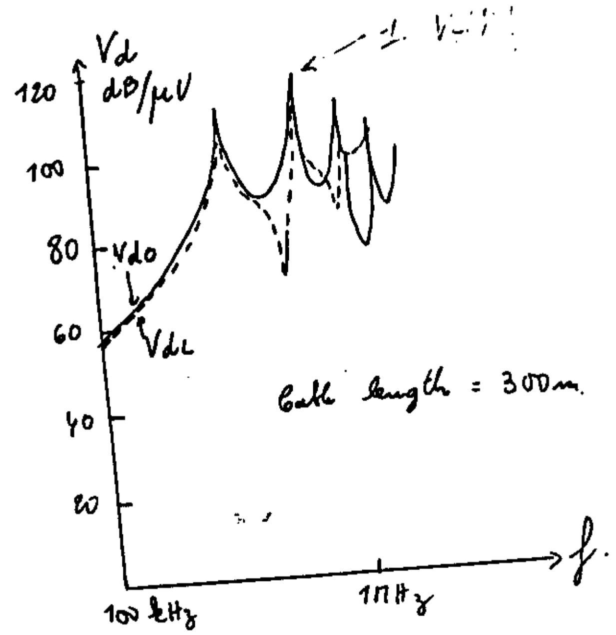
(2)

L_{2b}

a) Dissymmetrical line parameters

$\begin{cases} L_{11} = 330\ nH/m & L_{22} = 300\ nH/m \\ C_{11} = 65\ pF/m & C_{22} = 71\ pF/m \end{cases}$

$\begin{cases} \text{Sym. shield} \quad L_{t1} = L_{t2} = 1.2\ nH/m \\ \text{Sym. load impedance} \end{cases}$



Conclusion (bifilar)

Short cable or low frequency ⇒ Differential mode voltage

↓

shield dissymmetry

⇓

Differential transfer impedance Z_{td}

long cable or high frequency → Differential mode voltage

↓

dissymmetry of the inner lines parameters

⇓

Numerical simulation (?) or measurement on long cable

⇒ Now "generalization" to 3 or 4 inner lines both in $\begin{cases}\text{theory} \\ \text{exp.}\end{cases}$

(5)

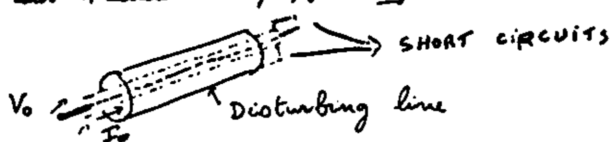
Z_{t4}

PRINCIPLE OF MEASUREMENT

2 parameters : Z_t and Y_t → amplitude, → phase angle (useful?)

$(j L_t \omega \ , \ j C_t \omega)$

1st possibility . Measurement of Z_t (only) and Measurement of Y_t (only)



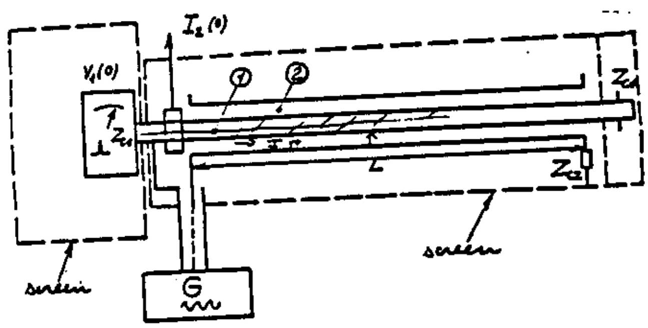
• $Z_t = \frac{1}{L} \frac{V_0}{I_p}$ (but $L << \lambda$)

$Y_t = \frac{V_0}{V_p} Y_a$; Y_a → capacitance of the coaxial line

Problems $\underline{L << \lambda}$, $\underline{\text{open}}$ circuit ? $\underline{\text{short}}$ circuit ?

Z_t

MATCHED IMPEDANCES



V(0) depend on Z_t and Y_t

advantage → $L \leq \lambda$

$$V_1(0,\omega) = -\frac{I_2(0,\omega)}{2(\gamma_1+\gamma_2)}\left(Z_{t0} + j L'_t \omega\right)\left[1 - e^{-(\gamma_1+\gamma_2)}\right.$$

$$V_1(L,\omega) = -\frac{I_2(0,\omega)}{2(\gamma_2-\gamma_1)}\left(Z_{t0} + j L''_t \omega\right)\left[1 - e^{-(\gamma_2-\gamma_1)L}\right]$$

$$\begin{cases} L'_t = L_t + Z_{c1} Z_{c2} C_t \\ L''_t = L_t - Z_{c1} Z_{c2} C_t \end{cases}$$

sum or difference ⇒ L_t, C_t.

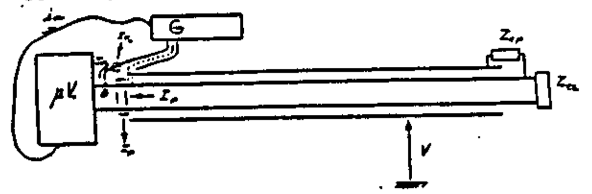
(8)

MODES D'INJECTION DU COURANT PERTURBATEUR

INJECTION OF THE DISTURBING CURRENT

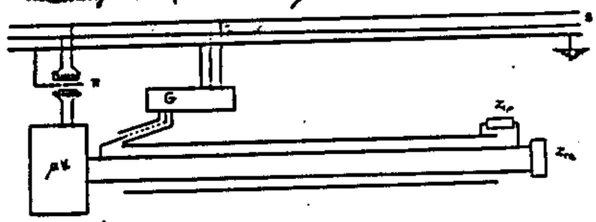
(51)

1) PAR CONTACT DIRECT AVEC BLINDAGE À L'EQUIPOTENTIELLE DE MASSE
Direct contact to the shield



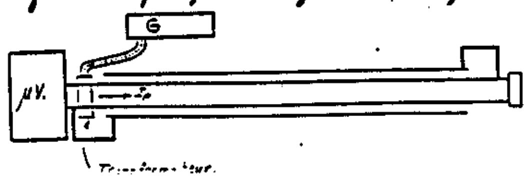
PAR CONTACT DIRECT AVEC BLINDAGE AU POTENTIEL. IL FAUT ISOLER LE RÉCEPTEUR DE LA MASSE SECTEUR.

To avoid that part of the disturbing current flows through the ground of the receiver → It is necessary to put a transformer



2) PAR COUPLAGE MAGNETIQUE AU MOYEN DE TRANSFORMATEURS.
LA LIGNE PERTURBATRICE EST CONNECTÉE SUR DES COURTS-CIRCUITS

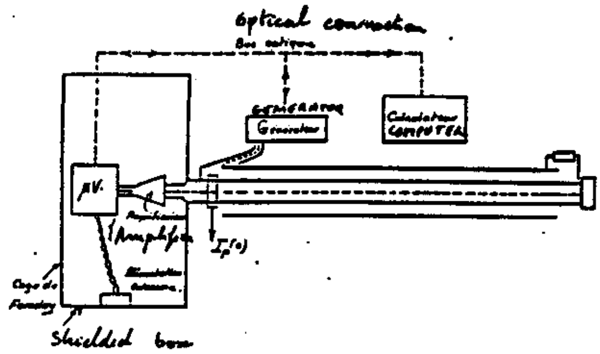
"Magnetic coupling" → Injection through a transformer



PROTECTION DE LA CHAINE DE RECEPTION.

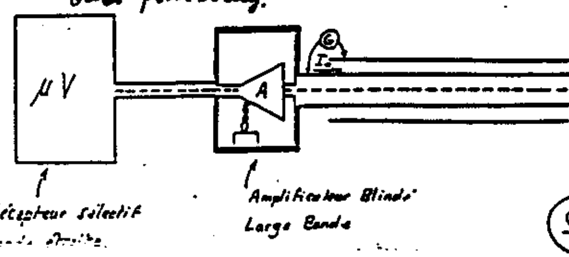
Measurement on well-shielded cable → Necessity to shield the receiver

PROTECTION SIMULTANÉE DE L'AMPLI DE TENSION ET DU RÉCEPTEUR

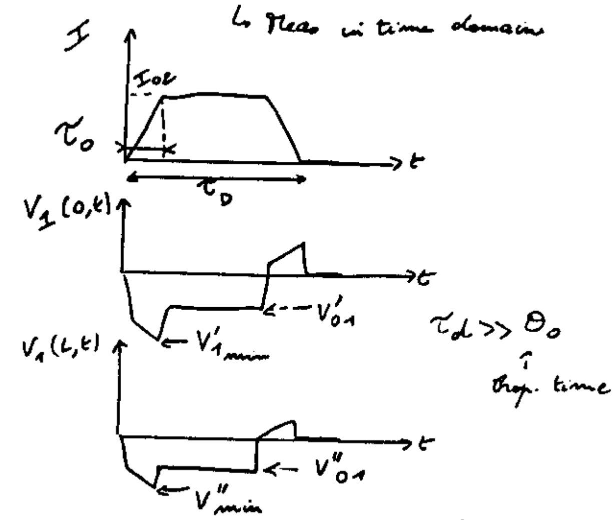


PROTECTION DE L'AMPLI DE TENSION DONT LE NIVEAU DE SORTIE SOIT SUFFISANT POUR ÉVITER TOUT RISQUE D'INTERFERENCE

Other possibility.



Measurement of amplitude and phase angle
- → Vector analyzers → Meas. in frequency dom.
- → Time domain + FFT
  ↳ Meas. in time domain



$\tau_d >> \theta_0$
Prop. time

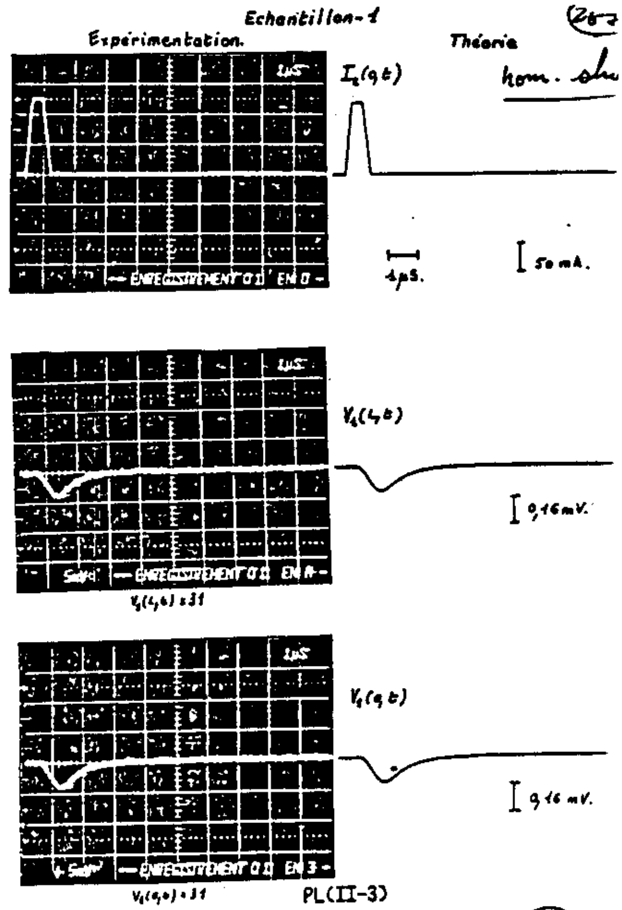
If resistance of the braid indep. of frequency → R_0

$$V'_{1\,min} \simeq -\frac{I_{0L}}{2}\left(R_0 + \frac{L'_t}{\tau_d}\right) L$$

$$V'_{01} \simeq -\frac{I_{0L}}{2} R_{0L}$$

$$V''_{1\,min} \simeq -\frac{I_{0L}}{2}\left(R_0 + \frac{L''_t}{\tau_d}\right) L$$

Echantillon-1

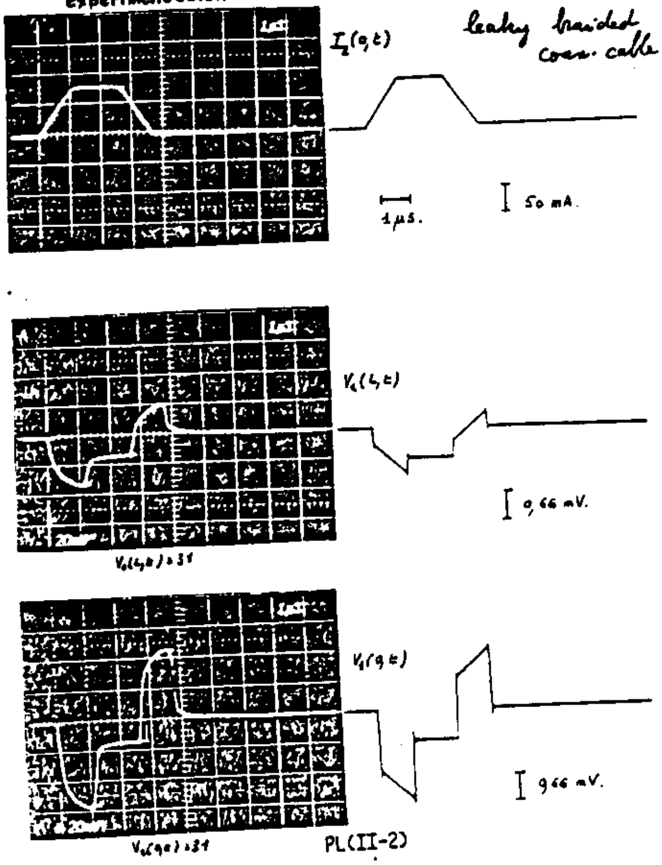


PL(II-3)

Echantillon - 3 (2t8)

Expérimentation — Théorie

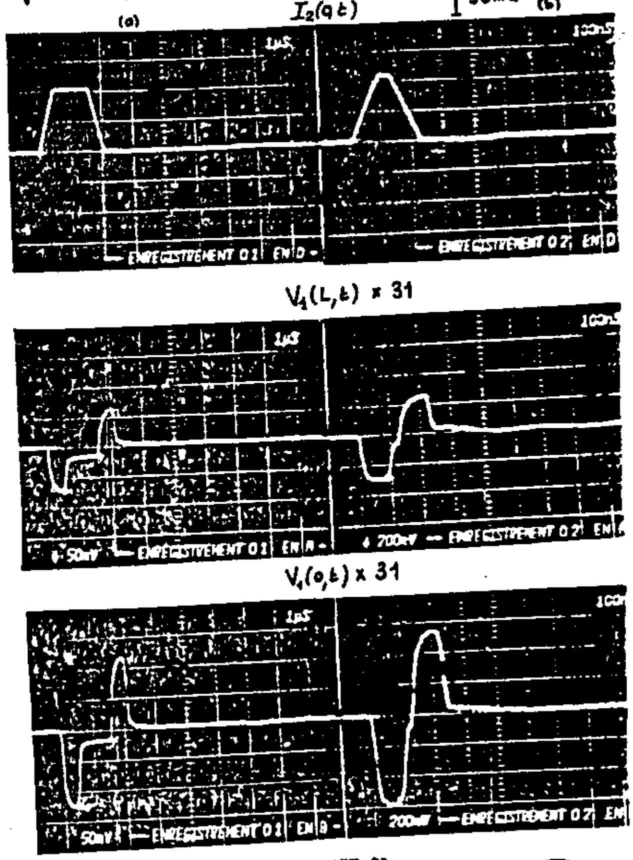
leaky braided coax. cable



PL(II-2)

L stops ⇒ τ_d.

Echantillon - 3 (2t9)

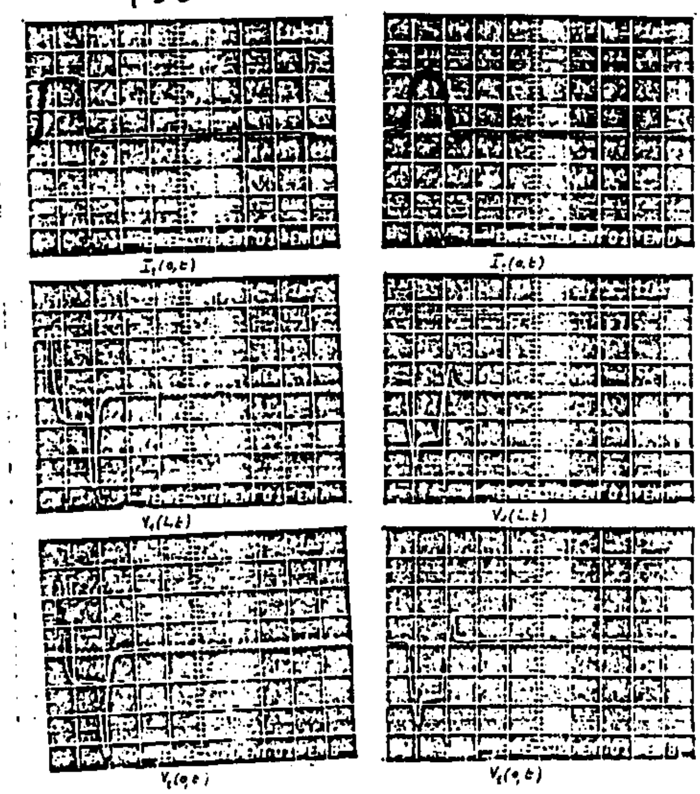


PL(II-6)

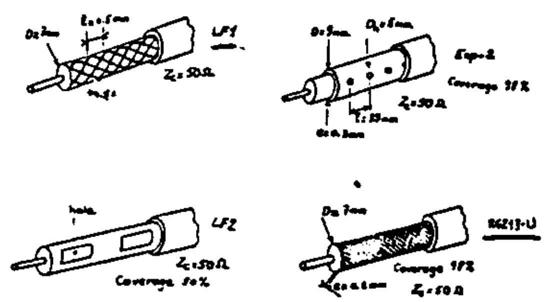
Braids with high optical coverage ratio (2t10)

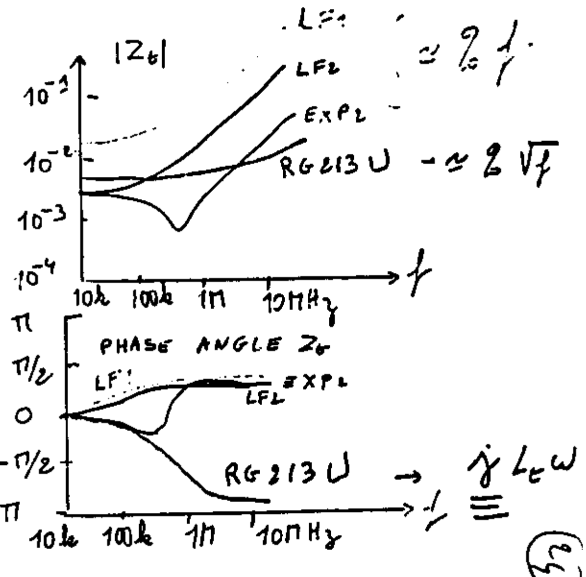
Weave angle ψ

$\psi = 20°$ — $\psi = 60°$



(2t)





Comparison with the theoretical results (Z_{t13})

→ based on electric and magnetic polarizat of the apertures (diffraction by the apertures)

→ good agreement BUT differ [illegible] !!

$\vec{m}$: magnetic dipole
$\vec{p}$: electric dipole

$\vec{m} = \underset{m}{d} : \vec{H}_0$ mg. polar.

$\vec{p} = \mathcal{E}_0 d_e \vec{n} \cdot E_{0n}$ elec. polar

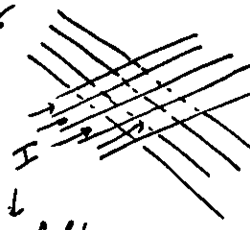
$\Rightarrow L_t \simeq \dfrac{\nu\, d_m \mu_0}{\pi^2 D^2}$ and $C_t = g\, \dfrac{\nu\, d_e C_1 C_2}{\pi^2 D^2 \mathcal{E}_1}$

intrinsic char. of the shield; depend on the environment.

(but d_e and $d_m \not\approx (L \text{ or } \ell)^3 \Rightarrow$ errors.

But high coverage and $\Psi < 45°$?

Theoretical explanation based on induction effects (Z_{t14})



↓
mg. field
↓
eddy currents on the lower belt

Σ effects ⇒ "equivalent" transfer imped.

$Z_t = Z_{td} + Z_{tc}$

diffusion; additional term

$Z_{tc} \simeq -k \sin\left(\frac{\pi}{2} - 2\Psi\right) \sqrt{\frac{\omega\mu}{\sigma}}\; e^{j\pi/4}$

$\varphi(Z_{tc}) \rightarrow -3\pi/4$, $|Z_{tc}| \nearrow \sqrt{f}$
if $\Psi < 45°$

All the coupling mechanisms present → depend on their relative amplitude

⇒ [illegible]

predict Time domain response of a system...

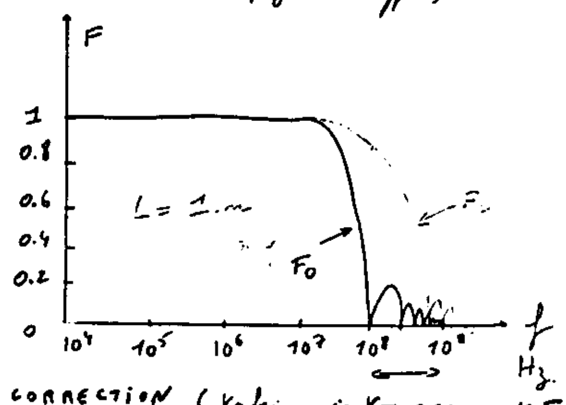
(25)

Limitation of the usual experimental set-up at high frequency (Z_{t14bis})

$V_0(0) = -\frac{1}{2}(Z_t + Z_{c1} Z_{c2} Y_t)\, I \cdot L \cdot \underline{F_0}(\gamma, L)$

$V_0(L) = \frac{1}{2}(Z_t - Z_{c1} Z_{c2} Y_t)\, I \cdot L \cdot \underline{F_L}(\gamma, L)$

Propagation effects



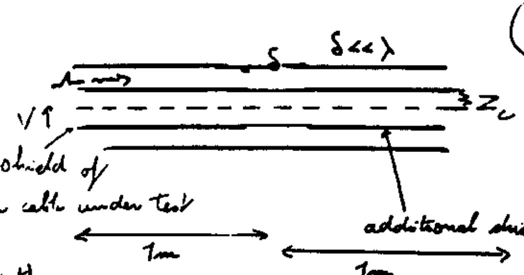
CORRECTION (taking into account F_0 and F_L) not obvious above 100 MHz !

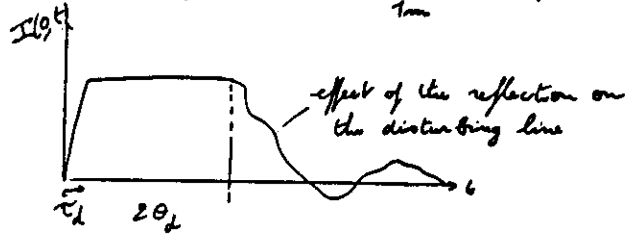
→ lack of precision

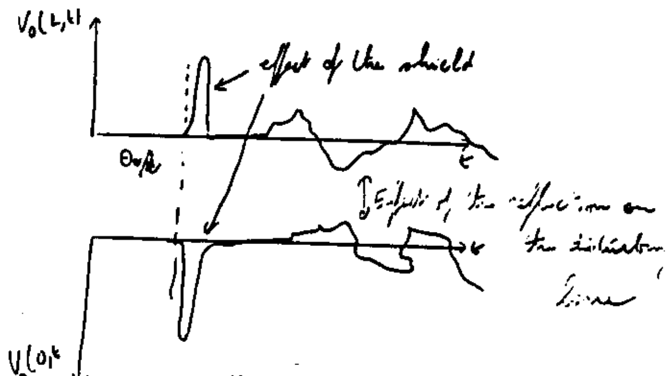
Solution $L \searrow$ ⇒ [illegible] (connection)

Idea → test only a small part of the cable

(Z_{t14b})



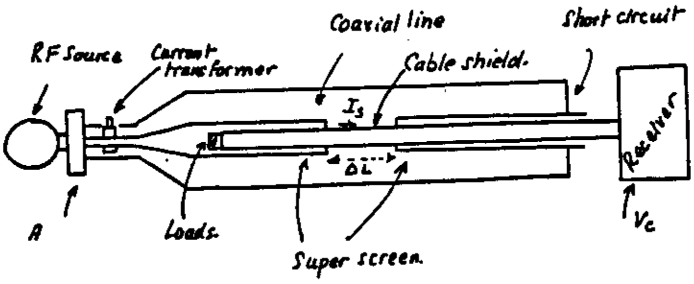




$\tau_d = 400$ ps

(26)

MODIFIED TRIAXIAL TECHNIQUE
(Shield discontinuity method)



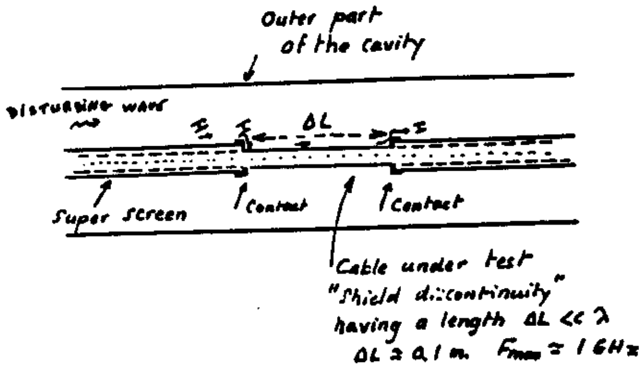
$\Delta L = 0.1\ m < \lambda = 0.3\ m \Rightarrow f = 1\ GHz$

The current flowing on the cable shield I_s is deduced from the measurement of the current transformer.

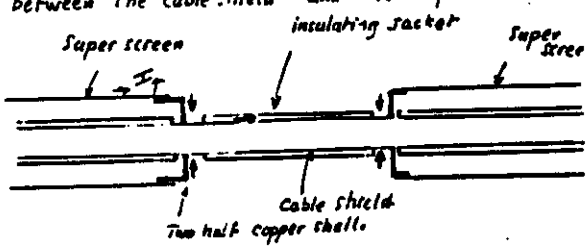
At low frequencies Z_t is given by the ratio $\frac{V_c}{I_s}$ for all frequencres

At high frequencies Z_t is determinated for each resonance of the outer coaxial line which produces a maximum value of the voltage V_c.

SHIELD DISCONTINUITY

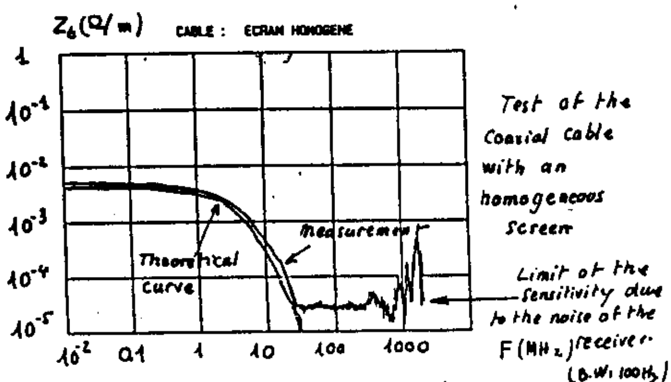


Detail of the electrical contact between the "cable shield" and the super screen.



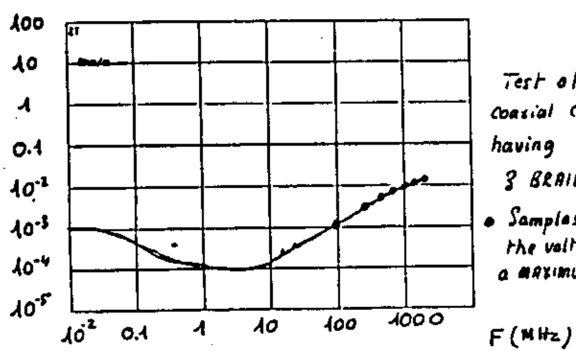
The contact between the "super screen" and the "cable shield" is made by the "pressure" of two half copper shell

(47)



Test of the coaxial cable with an homogeneous screen

The shield discontinuity method.



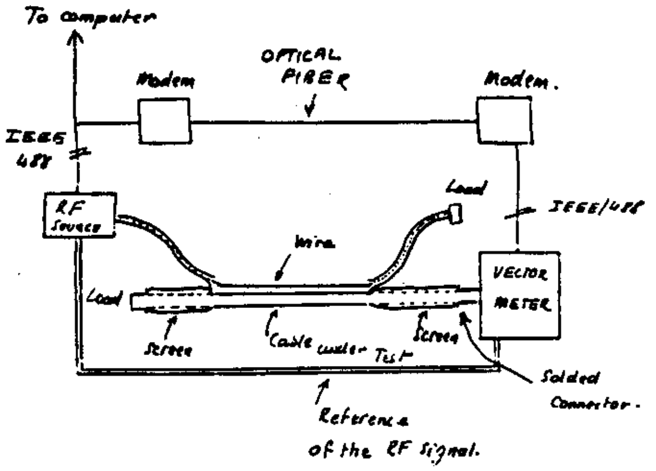
Test of a coaxial cable having 3 BRAIDS

• Samples where the voltage takes a maximum value

THE SHIELD DISCONTINUITY

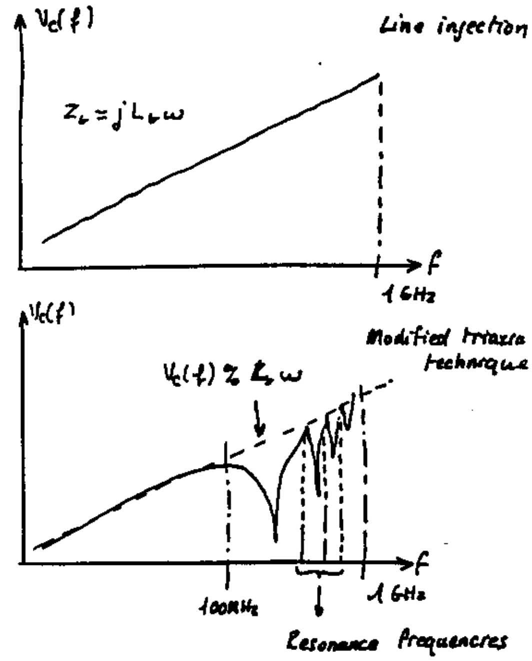
- Large frequency range (1 kHz - 2000 MHz)
- Good sensitivity (high immunity cable)
- May be extended to the measurement of the transfer impedance of the connectors
- The characteristic Zt (f) is sampled above the first resonance frequency of the cavity

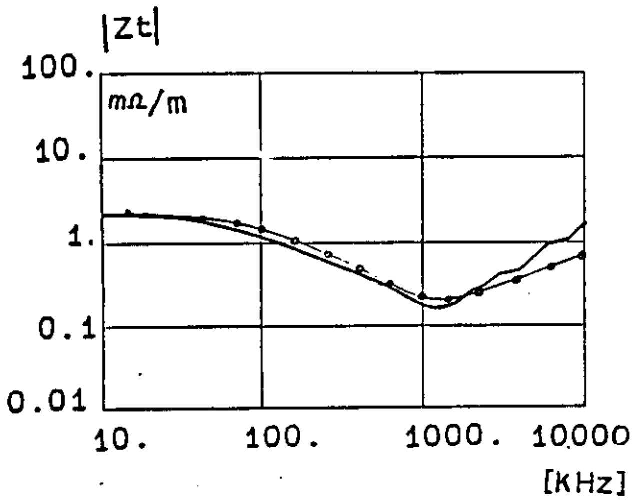
(48)



LINE INJECTION METHOD

VARIATIONS OF THE VOLTAGE "V_c" versus the frequency





Double Braid. 10kHz < f < 10 MHz

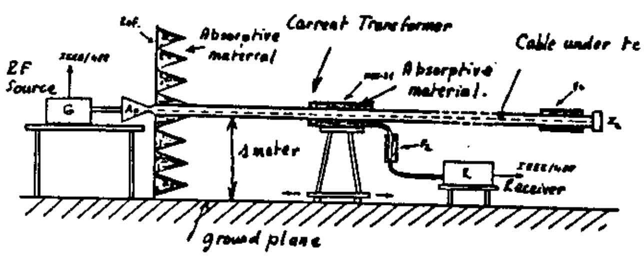
Line injection method.

Shield discontinuity method.

THE ABSORBING CLAMP METHOD

which is described in the I.E.C. Standard

The cable under test having a length very much greater than the wavelength is put above a ground plane at a distance equal to 1 meter



The RF power is injected in the cable and the measurement consists in determining the current due to the surface wave induced around the cable shield

The surface wave is damped by an absorptive material put near the C.T. and the RF source

We determine the ratio S between the power P injected in the cable and the average power P measured by the receiver connected on the C.T. which is moved along the cable during the measurement

The next formula is used to make the conversion between S and Zt
The coefficient A introduced in this formula is given by a calibration procedure

$$Z_t = \frac{\omega(\sqrt{\varepsilon_r} - 1)}{c}\sqrt{Z_{cr} Z_{cc}}\; 10^{-\frac{a_s}{20}}$$

$$a_s = 10 \log\left(\frac{P_o}{P_r}\right) - A$$

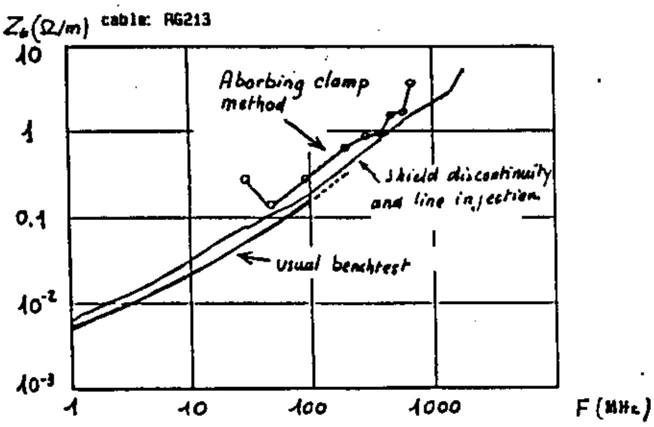
$$Z_{cr} = 60\left[\mathrm{Log}\left(\frac{\lambda}{\pi D}\right) - 0.116\right]$$

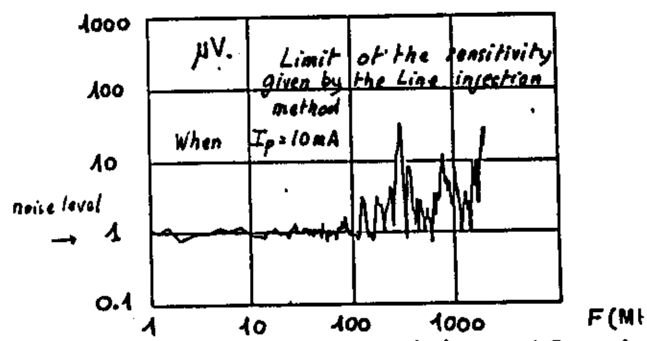
Z_{cc} : characteristic impedance of the cable under test

THE ABSORBING CLAMP METHOD

- Simple and low cost bench-test

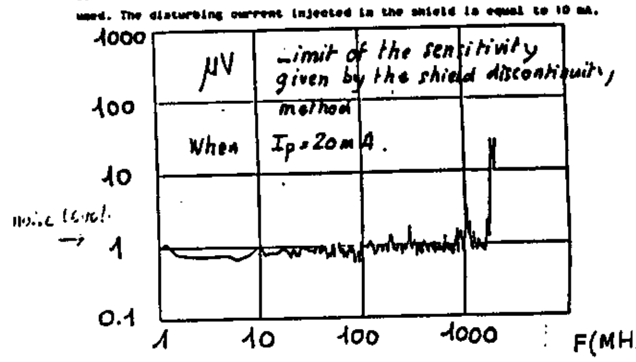
- Lack of sensitivity
- Calibration procedure
- Frequency range limited to 30 - 700 MHz
- Lack of accuracy S to Z_t conversion





Measurement on a copper tube (thickness: 0.3mm)

Variation of the voltage obtained on the receiver connected into the cable under test when the line injection method is used. The disturbing current injected in the shield is equal to 10 mA.



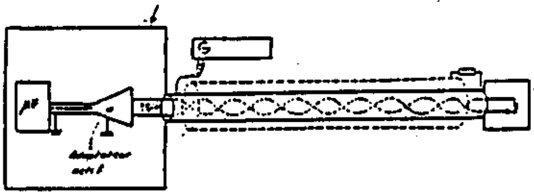
Variation of the voltage obtained on the receiver connected into the cable under test when the triaxial method is used. The disturbing current injected in the shield is equal to 20 mA.

MESURE DE L'IMPEDANCE DE TRANSFERT DIFFERENTIELLE Z_{td}

MEASUREMENT OF THE DIFFERENTIAL TRANSFER IMP.

PAR SYMÉTRISEUR ACTIF : AMPLIFICATEUR OU CHAINE ÉLECTRO OPTIQUE

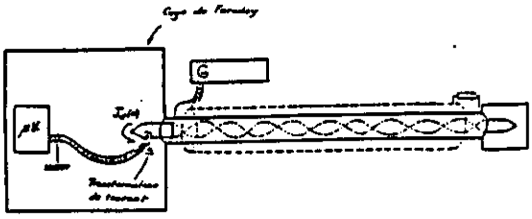
Measurement with a differential amplifier.



PAR SYMÉTRISEUR PASSIF : TRANSFORMATEUR DE COURANT.

IL FAUT ALORS FAIRE LA CONVERSION COURANT TENSION

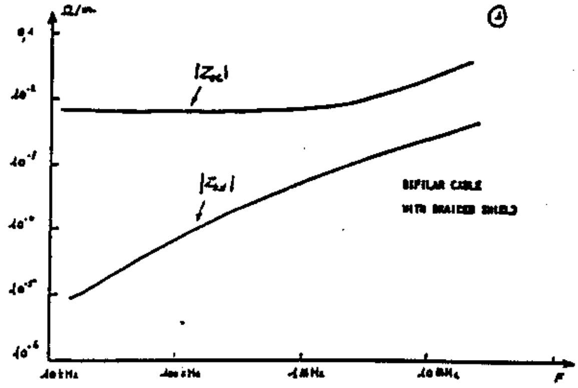
Use of a transformer



EXEMPLE

Z_{tc} ET Z_{td} MESUREES SUR UN BIFILAIRE PROTEGE PAR UNE TRESSE

1 - BIFILAR CABLE WITH BRAIDED SHIELD AT LOW FREQUENCIES Z_{TC} IS DUE TO THE DIFFUSION COUPLING BUT Z_{TD} IS DUE TO THE MAGNETIC COUPLING.

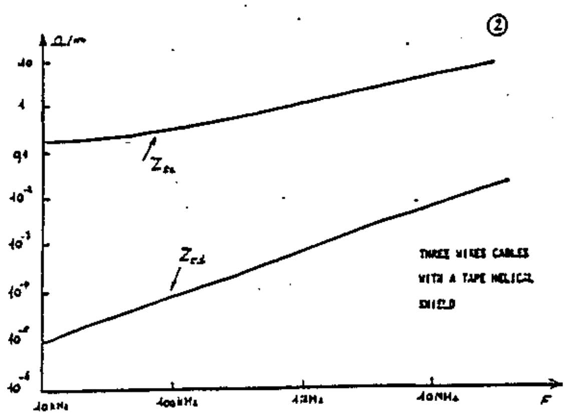


33

EXEMPLE

Z_{tc} ET Z_{td} MESUREES SUR UN TRIFILAIRE PROTEGE PAR UN RUBAN HELICOIDAL

2 - THREE WIRES CABLE WITH A TAPE HELICAL SHIELD (POWER CABLE)
THE HIGH SYMETRY OF THE INNER CONDUCTOR LEADS TO A VERY LOW Z_{TD}.



34

Penetration through apertures and into shielded boxes

PENETRATION TROUGH APERTURES

→ Perfectly conducting plane

- Polarizabilities

→ "Loaded aperture"

- Surface impedance
- Contact resistance

PENETRATION INTO NON PERFECTLY CONDUCTING SHIELDS

→ Attenuation by a conducting plane

→ Effect of the "volume"

Concept of "magnetic current"

Duality : Maxwell equations

$\mathrm{rot}\,\vec{H} = \vec{J}_e + \frac{\partial \vec{D}}{\partial t}$ $\quad\vec{J}_e$: current density or $\frac{\partial \vec{D}}{\partial t}$, same unit

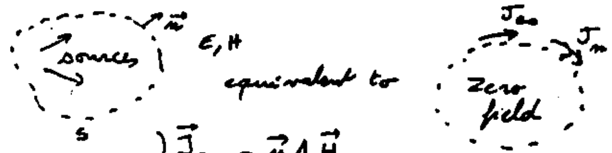
$\mathrm{rot}\,\vec{E} = -\frac{\partial \vec{B}}{\partial t} + ?$

→ Introduce a magnetic current density $\vec{J}_m$

$\Rightarrow \mathrm{rot}\,\vec{E} = -\frac{\partial \vec{B}}{\partial t} - \vec{J}_m$ fictive source

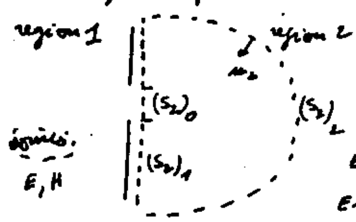
Advantage : Change a problem of aperture rad. into a problem of magnetic current source radiation

Equivalence theorem



$$\left\{\begin{array}{l}\vec{J}_{eo} = \vec{n} \wedge \vec{H} \\ \vec{J}_{mo} = \vec{E} \wedge \vec{n}\end{array}\right.$$

Application : penetration through an aperture (infinite plane)



$S_2 = (S_2)_0 + (S_2)_1 + (S_2)_2$

Equivalence :
Exact fields in region 2
zero fields in region 1

On S_2 → $\left\{\begin{array}{l}\vec{J}_{e2} = \vec{n}_2 \wedge (\vec{H}_2)_{S_2} \\ \vec{J}_{m2} = (\vec{E}_2)_{S_2} \wedge \vec{n}_2\end{array}\right.$

The field radiated by $\vec{J}_{e2}$ and $\vec{J}_{m2}$ can be expressed in terms of the potential vector $\vec{A}$

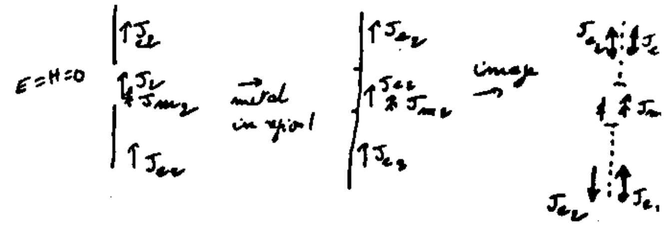
$$\vec{A}_{e2} = \frac{\mu}{4\pi}\int_{S_2} \vec{J}_2(r')\,\psi(r,r')\,dS \qquad \psi = \frac{e^{-jk|\vec{r}-\vec{r}'|}}{\vec{r}-\vec{r}'}$$

$$\vec{A}_{m2} = \frac{\varepsilon}{4\pi}\int_{S_2} \vec{M}_2(r')\,\psi(r,r')\,dS \qquad k = 2\pi/\lambda$$

$\vec{r}$: point of calculation ; $\vec{r}'$ any point on S_2

$$\vec{E}_2 = \frac{1}{j\omega\varepsilon\mu}\left[\overrightarrow{\mathrm{grad}}\,\mathrm{div}\,\vec{A}_{e2} + k^2 \vec{A}_{e2}\right] - \frac{1}{\varepsilon}\,\mathrm{rot}\,\vec{A}_{m2}$$

Simplification (Note on the metal $\vec{J}_{m2} = 0$ since $E_{tg} = 0$)



⇒ The aperture is equivalent to the radiation of magnetic current on its surface, radiating in free space. $\quad J_{m2} = 2\vec{E} \wedge \vec{n}$

Boundary condition ?

Express the fields and equivalence on both sides + continuity of the tg field on the aperture
⇒ Integral eq.

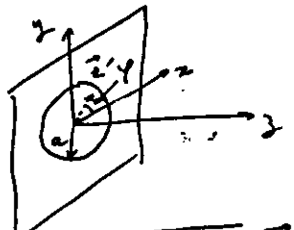
$$\left.\frac{H_{cc}}{2}\right|_{tg} = \frac{1}{j\omega\varepsilon\mu}\left[\overrightarrow{\mathrm{grad}}\,\mathrm{div}\,\vec{A}_m + k^2 \vec{A}_m\right]$$

where $\vec{A}_m = \frac{\varepsilon}{4\pi}\int_S \vec{J}_m(r')\,\psi(\vec{r},\vec{r}')\,dS$

M.o.M. Then, knowing J_m → calculate the field at any point.

(4) Simplification: small apertures ($d \ll \lambda$)
<u>incident uniform</u> field (important (of experiments)

Example: circular aperture

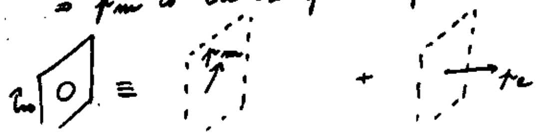


$$\vec{J}_m \sim -j\omega\mu \frac{4}{\pi}\sqrt{a^2 - z'^2}\,\vec{H}_{cc} + \frac{2}{\pi\sqrt{a^2-z'^2}}\,\vec{z}' \wedge \vec{E}_{cc}$$

Radiation at great distance of the aperture → asymptotic expansion. Radiation as an electric and a magnetic dipole with a moment

$$\begin{cases} \vec{p}_e = -\dfrac{\varepsilon_0}{2}\displaystyle\int_S (\vec{z}' \wedge \vec{J}_m)\, dS & = j\omega I\, dl_{eq} \\ \vec{p}_m = \dfrac{1}{j\omega\mu_0}\displaystyle\int_S \vec{J}_m\, dS & = I\, S_{eq} \end{cases}$$

Since $\vec{J}_m$ and $\vec{z}'$ are in the aperture plane
→ $\vec{p}_m$ is in the aperture plane; while $\vec{p}_e \perp$ --



It can be shown that $\vec{p}_e$ and $\vec{p}_m$ can be put in the form

$$\underline{\vec{p}_e = \varepsilon_0\, \alpha_e\, \vec{E}_{cc}} \qquad \underline{\vec{p}_m = -\overline{\overline{\alpha}}_m\, \vec{H}_{cc}}$$

α_e and α_m: electric and mag POLARIZABILITIES of the aperture

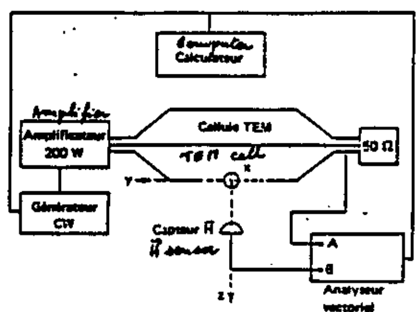
$\overline{\overline{\alpha}}_m$ → dyadic term (diagonal if the aperture has two axes of symmetry)

For a circular aperture (radius a)

$$\alpha_e = \frac{4a^3}{3} \quad ; \quad \alpha_{mxx} = \alpha_{myy} = \frac{8a^3}{3}$$

Note $\alpha \to p \to$ radiated field $\nearrow a^3$

(5) 17)



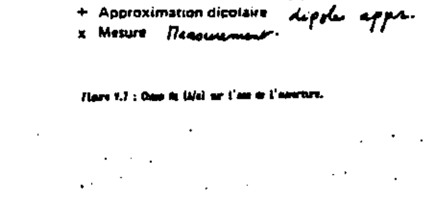
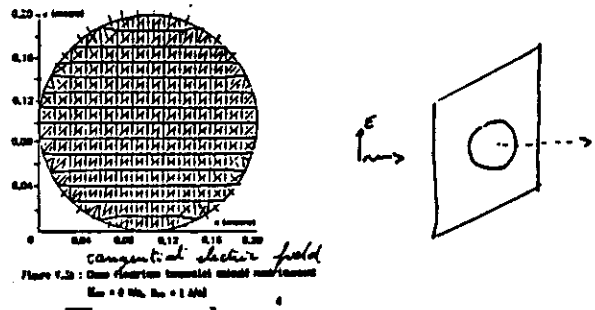
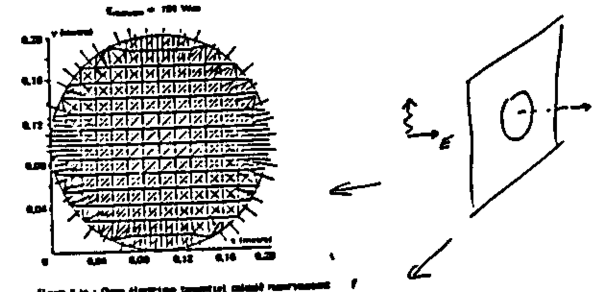


Figure V.7 : Onde H (A/m) sur l'axe de l'ouverture.

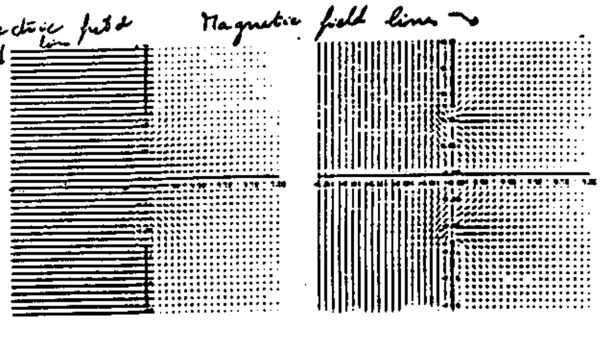
(6) 19)



tangential electric field



electric field lines ↓ Magnetic field lines →



Magnetic field lines

Influence of the orientation of the transmitting loop.

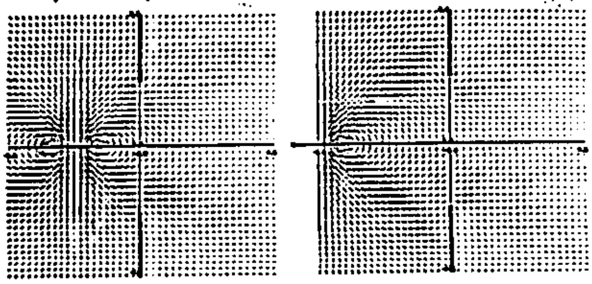


Figure [illegible] : [illegible] de l'ouverture.

Figure [illegible] : [illegible] de l'ouverture.

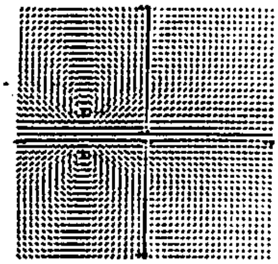


Figure [illegible] : [illegible] de l'ouverture.

→ important remarks

non uniform field

polari~~zabilités~~

Influence of the thickness of the material.

→ small if thickness d << transverse dimens.

→ on the contrary → wave guide

$\lambda > \lambda_c$

Circular aperture $\lambda_c = \frac{\pi D}{1.84}$

$\updownarrow D$

$\overleftrightarrow{d}$

Attenuation during the propagat.

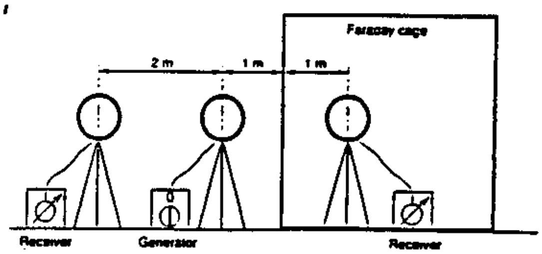
$S_A = e^{2\pi d\sqrt{1-\lambda_c^2/\lambda^2}} \simeq e^{2\pi d/\lambda_c} \simeq 54.6\, d/\lambda_c$ dB

Useful → Shielded enclosure.

→ Meas. of the shielding eff.

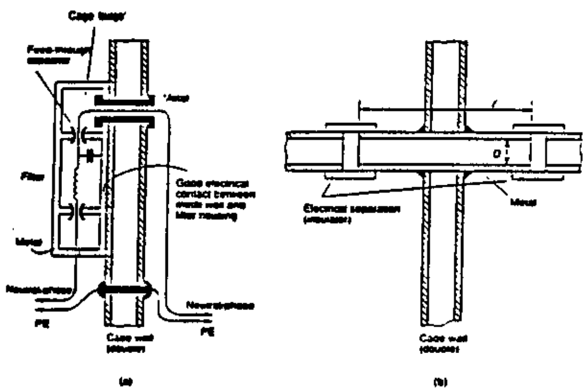
↓

Normalized.



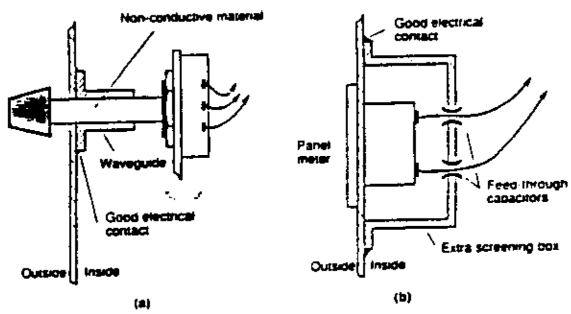
(Difficult if small box!)

Example de précaution - traversée d'enceinte blindée



(a) (b)

Figure 7.8 Bringing conduits through the wall of a Faraday cage: (a) mains cable; (b) gas/water (etc.) service lines.



(a) (b)

Figure 7.10 (a) Passage of a non-conductive control spindle through a screen with EM sealing by the waveguide principle; (b) panel meter with extra screening box and feed-through capacitors.

Influence of a "load" on the aperture

→ loaded aperture

→ Define a surface impedance Z_s for a material

$\boxed{J, E}$ $\updownarrow d$

In the medium $\text{rot}\,\vec{H} = \vec{J} + j\omega\varepsilon_0\vec{E}$

where $\vec{J} = (\sigma + j\omega(\varepsilon - \varepsilon_0))\vec{E}$

→ Surface current density J_s : integral of $\vec{J}$ along a "vertical" line

$|\vec{J_s}| = \int_{\mathcal{C}_s} |\vec{J}|\, dl$

If the skin effect is negligible (low conduct. material and (or) low frequency

$J_s = \int_{\mathcal{C}_s} |\vec{J}|\, dl = |\vec{J}| \cdot d = [\sigma + j\omega(\varepsilon - \varepsilon_0)]|\vec{E}|\, d.$

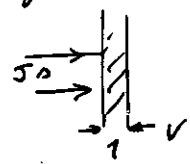
Usually σ remains $\gg \omega\varepsilon$

$|\vec{J_s}| \simeq \sigma|\vec{E}|\, d$. Definition: $\vec{E_{tg}} = Z_s\, \vec{J_s}$

⇒ homogeneous medium $\boxed{Z_s = \frac{1}{\sigma d}}$

Contact resistance

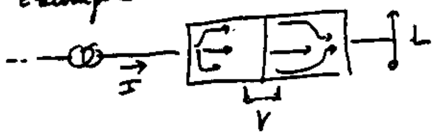
Usually a contact admittance / unit length is defined G_c



$G_c = J_0 / V$

Current density / length $\perp$ to the joint

Example



Total admittance $G = \frac{I}{V} = \frac{JL}{V} = G_c L$

$\Rightarrow$ Contact resistance per unit length

$R_c = 1/G_c$

→ Change the boundary conditions

Simplest case: circular aperture illuminated by a uniform field

$$\vec{p}_m = \vec{p}_{mo}\left(1 + j\frac{f}{f_m}\right)^{-1} ; \; p_e = p_{eo}\, j\frac{f}{f_c}$$

where

$$f_m = \frac{3 Z_0}{8 \mu_0 a}\left(1 + \frac{R_c}{a Z_0}\right) ; \; f_c = \frac{3}{16 \varepsilon_0 Z_0 a \left(1 + \frac{R_c}{a Z_0}\right)}$$

R_c: contact resistance

loading material → low pass filter for $\vec{p}_m$
→ important attenuation of $\vec{p}_e$ at low frequency

Note: the additional attenuation depends on the radius a.

To shield will be { easy on $\vec{E}$ (charge)
{ difficult on $\vec{H}$ (current)

$a = 10\text{ cm}, \; Z_0 = 1\,\Omega$

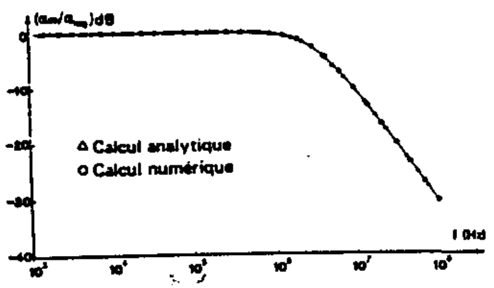


Figure V.10a : Variation de α_m/α_{mo} en fonction de la fréquence.

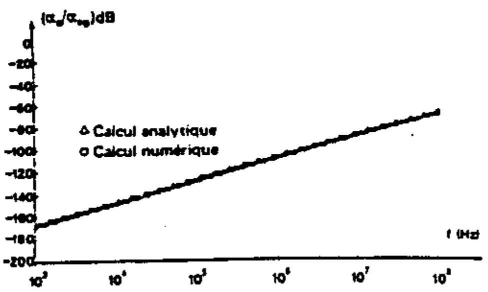


Figure V.10b : Variation de α_e/α_{eo} en fonction de la fréquence.

Influence of the contact resistance R_c

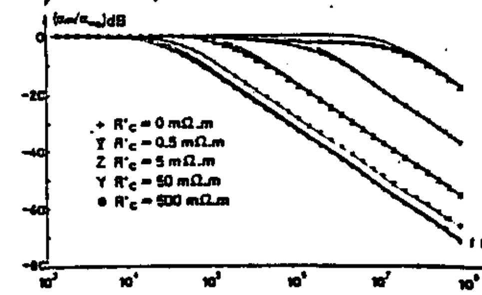


Figure V.11 : Influence de l'impédance de contact : résultats théoriques.

Influence de l'orientation de la boucle

{ $Z_0 = 4\,\text{m}\Omega$ – ouverture carrée de 60 cm de côté
rayon de la boucle d'émission 10 cm. – Distance à l'ouverture 20 cm

→ ouverture "libre"

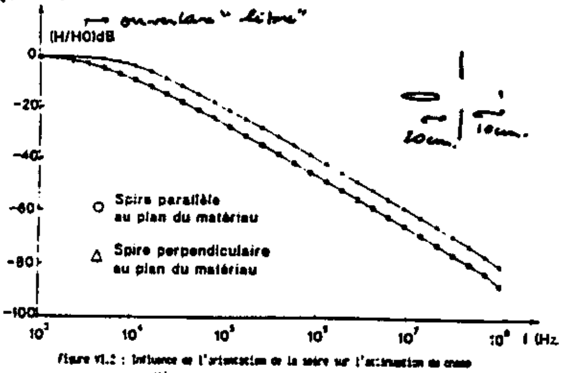


Figure VI.2 : Influence de l'orientation de la spire sur l'atténuation du champ magnétique.

Influence of the contact resistance R_c

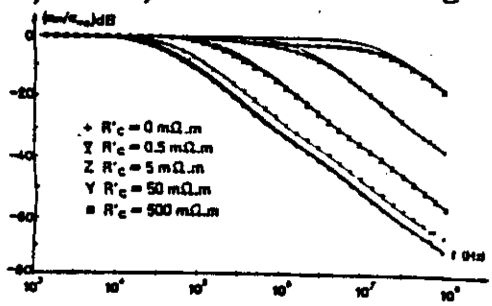


Figure V.11 : Influence de l'impédance de contact : résultats théoriques.

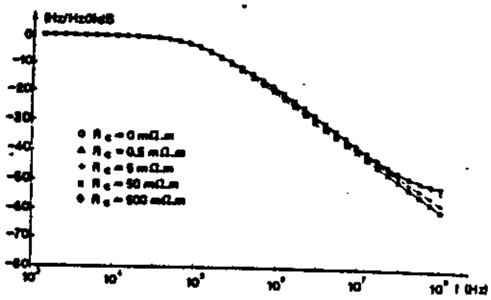


Figure VI.3 : Mise en évidence de la faible influence des joints dans le cas où la paroi est considérée ... extérieur.

(13b) Repartition des lignes de courant sur un matériau homogène transversale

Distribution of the current lines in a transverse plane

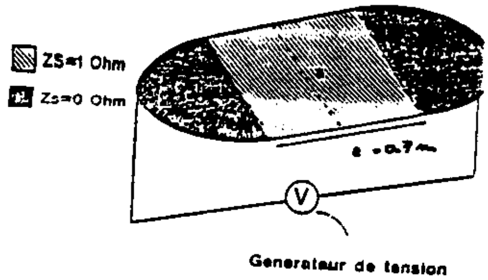


Figure III.4 : Exemple de circuit résistif.

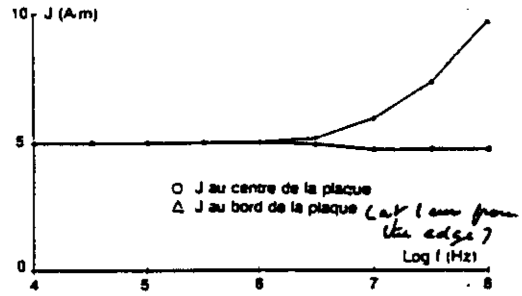
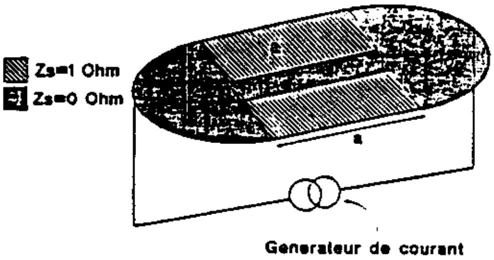


Figure III.7 : Densité de courant au centre et au bord de la plaque.



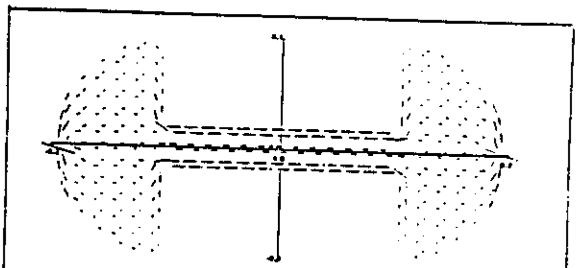


Figure III.11 : Répartition du courant à 10 kHz.

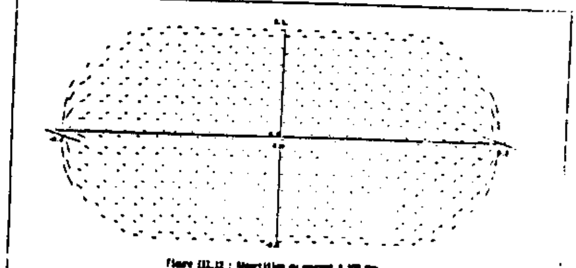
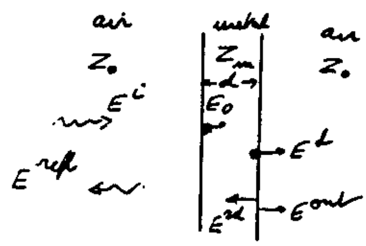


Figure III.12 : Répartition du courant à 100 MHz.

Attenuation by a plane (∞^2 long)



Shielding efficiency (Comparison with/without plane

$S(dB) = S_A(dB) + S_R(dB) + S_{MR}(dB)$

S_A : Attenuation due to the path d inside the m

S_R : Losses due to the reflection.

S_{MR} : contribution of the multiple reflections

↳ play a leading part if S_A is small < 10 dB,

Inside the metal $\sigma \gg \omega\varepsilon \Rightarrow Z_m = \sqrt{\frac{\omega\mu}{\sigma}} = \frac{\sqrt{2}}{\sigma\delta}$ ∀ sou

Absorption loss $S_A = \frac{E_0}{E_d} = \frac{H_0}{H_d} = e^{d/\delta} \simeq 8.7 d/\delta$ ca

Reflection loss : impedance mismatch at the interfa

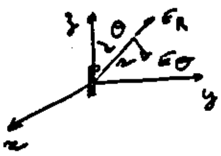
$S_R = \frac{(Z_0 + Z_m)^2}{4 Z_0 Z_m}$ but Z_0? Plane Wave → No P ... BUT source?

Multiple reflection

$S_{MR} = 1 - e^{-2d/\delta} \simeq 2d/\delta$

Source: $\ddot{V}EO$

→ Preliminary approach → radiation of the VEO



$$E_R = \frac{Z_0}{2\pi}\frac{I l\cos\theta}{r^2}\left(1+\frac{1}{jkr}\right)e^{-jkr}$$

$$E_\theta = j\frac{kZ_0}{4\pi}\frac{I l\sin\theta}{r}\left(1+\frac{1}{jkr}-\frac{1}{k^2r^2}\right)e^{-jkr}$$

$$H_\varphi = \frac{jk}{4\pi}\frac{I l\sin\theta}{r}\left(1+\frac{1}{jkr}\right)e^{-jkr}$$

$Z_0 = \sqrt{\mu/\varepsilon}$ Near zone $E \nearrow 1/r^3$; $H \nearrow 1/r^2$

Far zone (asymptotic range) $E, H \nearrow 1/r$
↳ wave: locally plane

Wave impedance

. Far zone $Z_W = \left|\frac{E_\theta}{H_\varphi}\right| = Z_0$

. Near zone $Z_W = \left|\frac{E_\theta}{H_\varphi}\right| = \frac{Z_0}{kr} = Z_E$ ($kr \ll 1$)

same def (direction $\theta = \frac{\pi}{2}$) ⇓

$Z_W \gg Z_0$

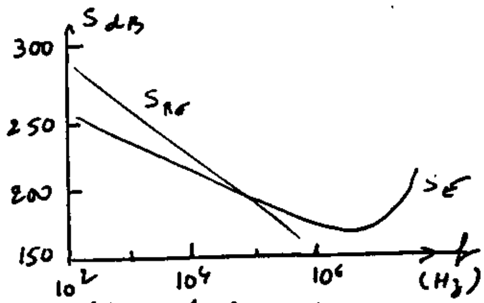
Electric field → high impedance

Attenuation near an electric antenna

$$S_{RE} = \left|\frac{(Z_E+Z_m)^2}{4Z_E Z_m}\right| \simeq \left|\frac{Z_E}{4Z_m}\right| = \frac{1}{4\sqrt{2}}\frac{\sigma\delta}{\omega\varepsilon_0 r}$$

Example . Cu , $d = 0.1$ mm , $r = 10$ cm.

$f \nearrow$ |S|



Difference between the 2 curves $\begin{cases} LF \to \text{Multiple reflect} \\ HF \to \text{skin effect.} \end{cases}$

. Attenuation ↘ if f ↗ (then skin effect)

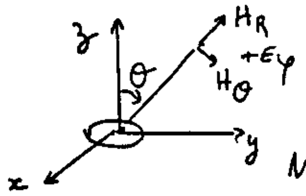
. Attenuation ↘ if distance source → plane ↗ (see formula) ... ⚠ Measurement !!

. Shielding → easy !
"high impedance" near the electric ant.

---

(12)

Magnetic dipole → duality



Near zone $\begin{cases} E \nearrow 1/r^2 \\ H \nearrow 1/r^3 \end{cases}$

In the direction $\theta = \pi/2$

$$|Z_W| = Z_H = \left|\frac{E_\varphi}{H_\theta}\right| = Z_0 kr$$

$Z_H \ll Z_0$

⇒ "low impedance

As previously S will depend on r

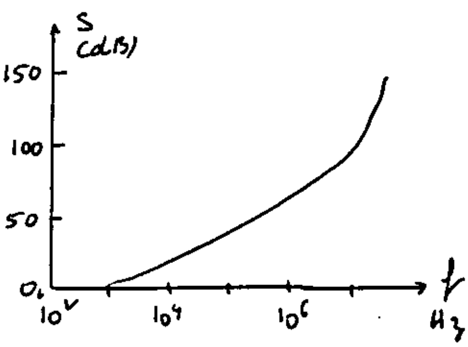
⇒ Shielding char. depend on the distance

... except for the plane wave ($r \to \infty$)

Attenuation near the magnetic dipole

$$S_{RH} = \left|\frac{(Z_H+Z_m)^2}{4Z_H Z_m}\right| \simeq \left|\frac{Z_H}{4Z_m}\right| = \frac{\omega\mu_0 r\sigma}{4\sqrt{2}}$$

Same example



⇒ Shielding ? → difficult at L.F.

. Attenuation ↗ if f ↗

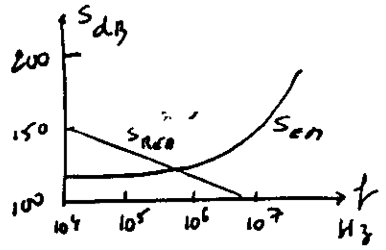
. Attenuation ↗ if r ↗

Variation → opposite (electric dipole)

Plane wave attenuation

If $z \gg \lambda \Rightarrow S_{RE} = S_{RH} = S_{Rem}$

$$S_{Rem} = \left| \frac{4Z_0}{4Z_m} \right| = \frac{377\, \sigma\, \delta}{4\sqrt{2}}$$



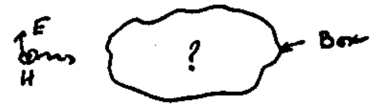
At low frequency ($d/\delta \ll 1$), the total att. tends to a limit value $\forall f \rightarrow 133\, \sigma\, \delta$

$\Rightarrow$ This att. remains cte in a wide frequency range

Important value.

But, assumption: ∞ plane

– but volume ?



Numerical methods $\begin{cases} \text{Integral eq} \\ \text{Finite diff. or Finite element} \end{cases}$

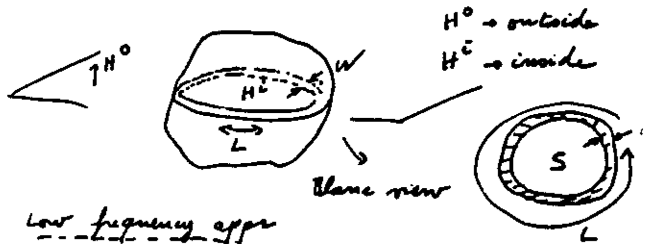
$\rightarrow$ qualitative + approximate approach

Dimensions of the cavity $\ll \lambda$, Field nearly unif inside the cavity.

Influence of the incident magnetic field

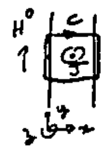
$H^o \rightarrow$ outside

$H^i \rightarrow$ inside



Low frequency appr

$H_0 \rightarrow$ uniform current density in the walls



$$\int_c \vec{H}\, \vec{dl} = \iint \vec{J}\, ds$$

$$H_0 L' - H_i L' = L' \int_0^W J(x)\, dx = L' J$$

$$\Rightarrow H_0 - H_i = J W$$

Current $\Rightarrow$ electric field inside the cavity tg to the material : $E^i = J/\sigma$

(14)

Now let us consider the contour defined by the intersection of the plane and the box (see p14)

$$\int \vec{E^i}\, \vec{dl} = \iint_S j\omega\mu\, \vec{H^i}\, \vec{ds}$$

$E^i L = j\omega\mu H^i S$ (Homog. field in the cavity)

$$H^i = \frac{E^i L}{j\omega\mu S} = \frac{J}{\sigma} \frac{L}{j\omega\mu S} \quad \text{But} \quad H_0 - H_i = JW$$

$$\Rightarrow \left| \; J = \frac{H_0}{W} \frac{1}{1 + \dfrac{1}{j\omega\mu\sigma W \frac{S}{L}}} \right.$$

Very low frequency $\omega\mu\sigma W \frac{S}{L} \ll 1$

$$E^i \sim j\omega\mu \frac{S}{L} H_0 \quad ; \quad H^i \simeq H^o$$

since $\dfrac{E^o}{H^o} = 120\pi$

$$\begin{cases} \dfrac{E^i}{E^o} \sim \dfrac{\omega\mu}{120\pi} \dfrac{S}{L} & \rightarrow \text{important attenuation} \quad E^i \nearrow \text{ if } f \nearrow \\ \dfrac{H^i}{H^o} \sim 1 & \rightarrow \text{No attenuation} \end{cases}$$

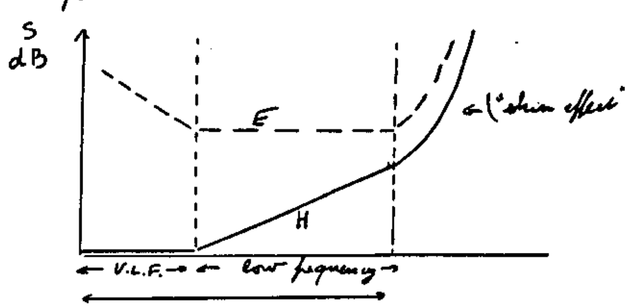
Low frequency $\omega\mu\sigma W S/L \gg 1$

$$J \sim H^o / W$$

$$\begin{cases} \dfrac{E^i}{E^o} = \dfrac{1}{120\pi\sigma W} & \rightarrow \text{cte} \\ \dfrac{H^i}{H^o} = \dfrac{1}{\omega\mu\sigma W \frac{S}{L}} & \rightarrow H^i \searrow \text{ if } f \nearrow \end{cases}$$

High frequency $\rightarrow$ skin depth

The current density is no more uniform in the Exponential attenuation

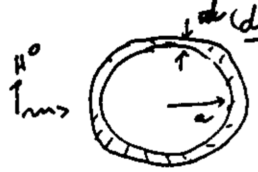


low frequency: Effect of the skin depth is negligible

Effect of the incident electric field $\rightarrow$ negligible

Remark: Importance of the ratio S/L

Example: penetration in a sphere



→ Analytical formulas

At the centre of the sphere

$$\frac{H^i}{H^0} = \left[ch\,\gamma d + \frac{1}{3}\left(\frac{j\omega\mu_0 a}{\eta} + \frac{2\eta}{j\omega\mu_0 a} \right) sh\,\gamma d \right]^{-1}$$

where $\eta = \sqrt{j\omega\mu_0/\sigma}$; $\gamma = \sqrt{j\omega\mu_0\sigma}$

At low frequency $\gamma d \to 0 \Rightarrow ch \approx 1$; $sh\,\gamma d \approx \gamma d$

$$\frac{H^i}{H^0} = \left[1 + \frac{1}{3}\left(j\omega\mu_0 a \sigma d + \frac{2d}{a} \right) \right]^{-1}$$

if $d \ll a$ and $Z_s = 1/\sigma d$

$$\left\{ \frac{H^i}{H^0} = \left[1 + j\frac{\omega\mu_0 a}{3 Z_s} \right]^{-1} \right.$$

Spherical shield behaves as a low-pass filter

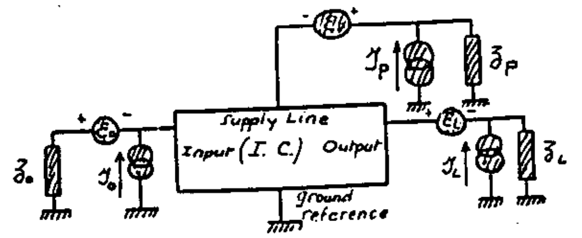
cut-off frequency $\left\{ f_c = \dfrac{3 Z_s}{2\pi\mu_0 a} \right.$

$f_c \searrow$ if $a \nearrow$ ⇒ For a given value of Z_s, a large sphere present a best shielding effect than a small one ! (See previously influence of S!)

Susceptibility of electronic equipment and Measurement procedures.

3

The induction mechanisms on the printed lines. produce VOLTAGE sources and CURRENT sources on the various ports of the component.

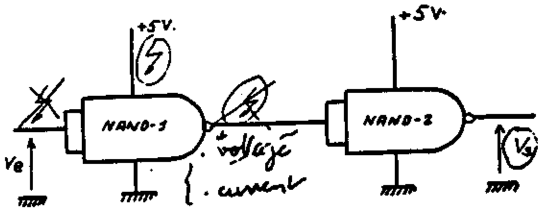


VOLTAGE sources ⟶ MAGNETIC Coupling

CURRENT sources ⟶ ELECTRIC Coupling

The loads Z_o, Z_P, Z_L correspond to the inner impedance of the sources

METHODOLOGY



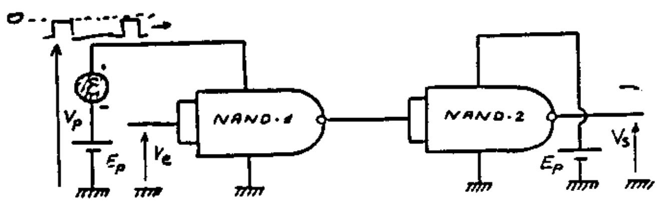
The characterization of the sensitivity of a digital integrated circuit to an electromagnetic disturbance is made by considering

TWO IDENTICAL CIRCUITS CONNECTED TOGET

This example shows two NAND-functions working as two inverters

5

VOLTAGE INJECTION APPLIED ON THE SUPPLY LINES OF THE FIRST NAND CIRCUIT



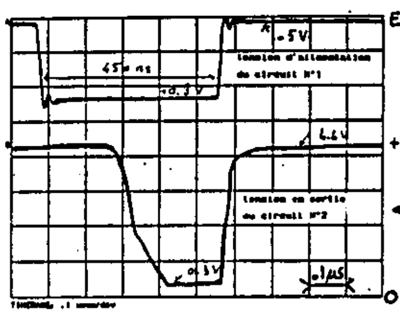
$E_p = +5V$.

The voltage source is characterized by a NEGATIVE PULSE.

The pulse width may be modified during the test monitored with either a high or a low level applied on the input port

Behavior of the OUTPUT VOLTAGE of the second circuit when a TRANSIENT VOLTAGE disturbs the supply of the first circuit

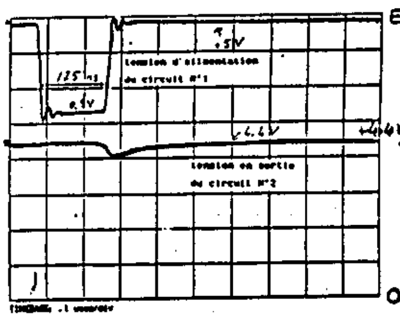
The duration of the disturbance is respectively 450 ns and 175 ns, the input voltage applied on the first circuit corresponding to a high level 5



Voltage on the supply line pulse width = 450 nS

Voltage on the output port

TTL circuit



pulse width = 175 ns

+4.4V: output voltage

Summary of the results given by the tests carried out on TTL and CMOS technologies

ΔV characterizes the peak to peak amplitude of the pulse which, applied on the supply line, produces a disturbing effect on the output voltage of the second circuit

Δt is the minimum value of the pulse duration producing a disturbance

Ve is the input voltage applied on the first circuit

TTL Technology NAND 74LS00

$V_e = 0$	$\Delta V < -4.3\,V.$	$\Delta t = 800\,ns$
$V_e = 5\,V$	$-4.2\,V < \Delta V < -2.5\,V.$	$\Delta t = 200\,ns.$

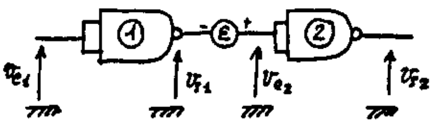
CMOS Technology NAND 74HC00

$V_e = 0\,V.$	$\Delta V < -2\,V$	$\Delta t = 15\,ns$
$V_e = 5\,V.$	No disturb.	—

CMOS Technology NAND 74C00

$V_e = 0\,V.$	$\Delta V < -2\,V$	$\Delta t = 60\,ns$
$V_e = 5\,V.$	No disturb	—

VOLTAGE PULSE "INJECTED" ON THE OUTPUT PORT OF THE COMPONENT



The voltage source "E" is added to the output voltage of the first component

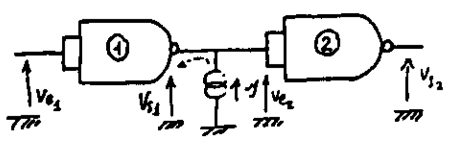
$$v_{e2} = v_{s1} + E.$$

When V_{e2} reachs the threshold voltages of the input port of the second component the logical state V_{s2} changes

The sensitivity depends of alone parameter

The threshold voltage of the second component

CURRENT PULSE "INJECTED" ON THE OUTPUT PORT OF THE COMPONENT



The current pulse J injected in the output port of the first component changes its output voltage V_{s1}

$$V_{s1} = V_{e2}$$

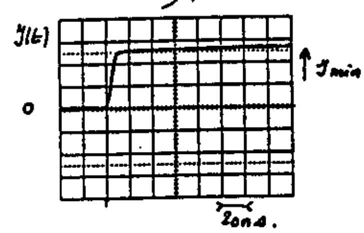
When V_{e2} reachs the threshold voltages of the input port of the second component the logical state V_{s2} changes.

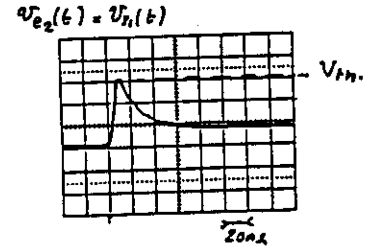
The sensitivity depends of two parameters.

- The output impedance of the first component
- The threshold voltages of the second component

RESULT OBTAINED WITH TTL INVERTOR

J_{min} corresponds to the smallest current amplitude which apply on the output port produces the threshold voltage

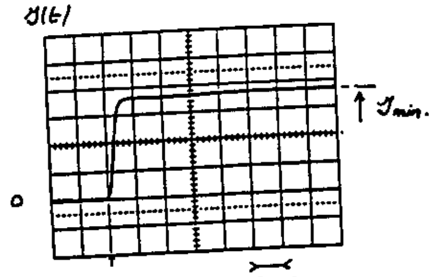


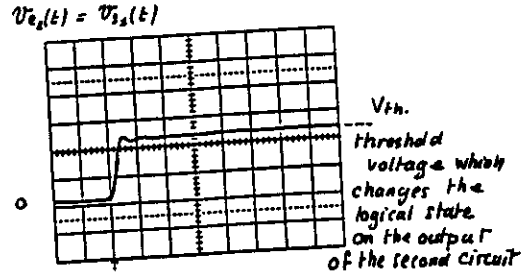


The peak amplitude reached by $v_{s1}(t)$ depends of the PN junction of the bipolar transitor disturbed by the transient phenomenon.

RESULTS OBTAINED WITH CMOS INVERTORS

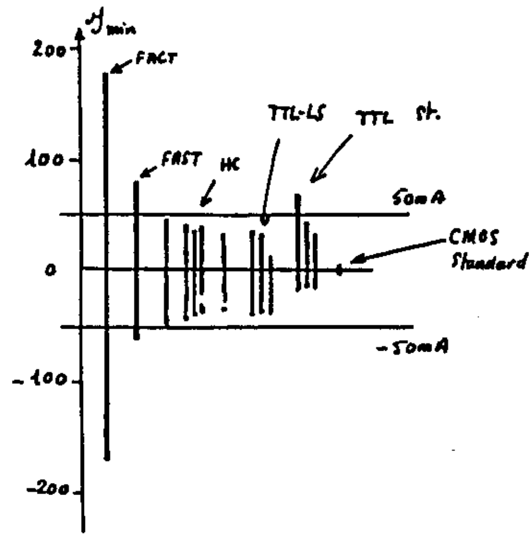
$\mathcal{I}_{min}$ corresponds to the current amplitude which produces the threshold voltage V_{th}



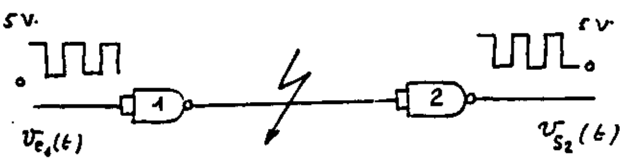


SUMMARY OF THE SENSITIVITIES MEASURED ON VARIOUS COMPONENTS TTL and CMOS technologies

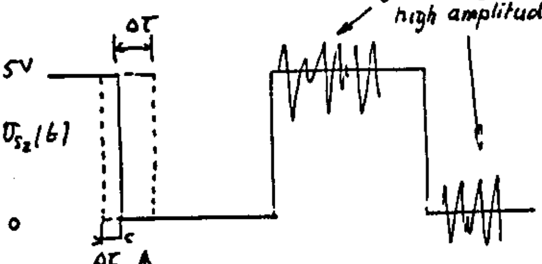
— CMOS — TTL Standard
— TTL LS — HC
— High speed technologies



EFFECTS OF THE DISTURBANCES DURING THE DIGITAL SWITCH OF THE COMPONENT

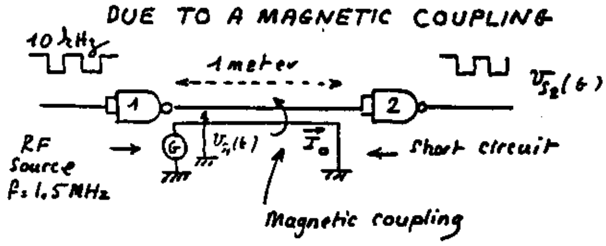


The disturbance appears on the stationary level of the digital signal only for high amplitude of the E.M.I.

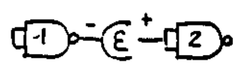


For low level of the E.M.I. the disturbance is characterized by a shift "Δτ" of the output signal $v_{s_2}(t)$

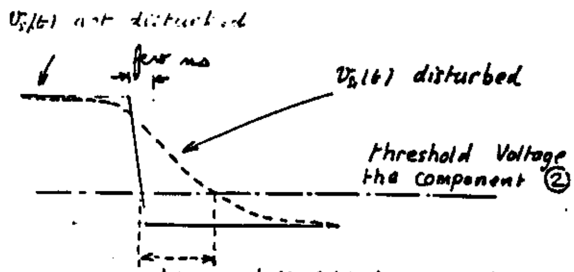
EFFECT OF THE VOLTAGE SOURCE DUE TO A MAGNETIC COUPLING



Equivalent circuit

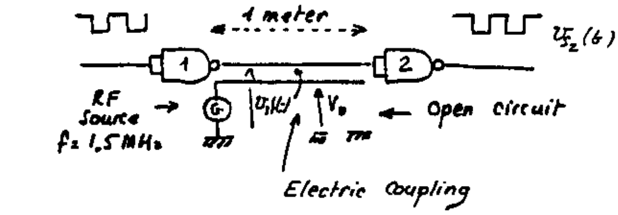


$\varepsilon = j L_{12} \omega I_0$

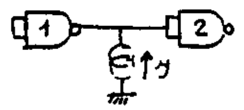


Δτ shift (delay) produced on $v_{s_2}(t)$

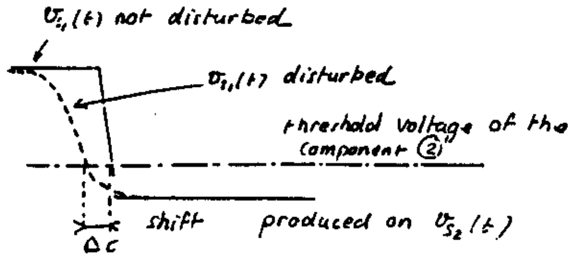
EFFECT OF THE CURRENT SOURCE DUE TO AN ELECTRIC COUPLING



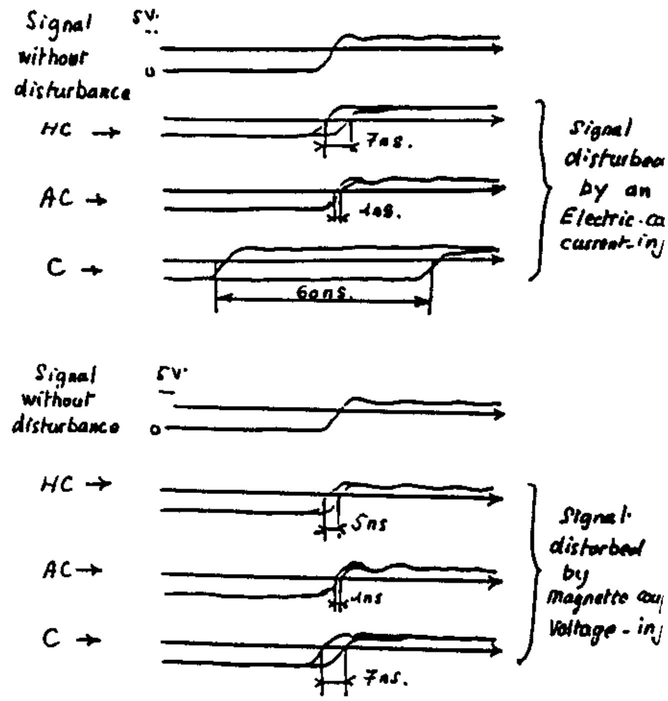
Equivalent circuit



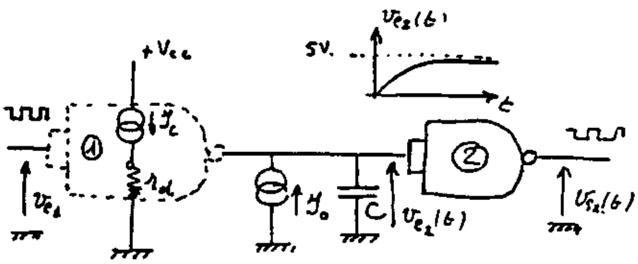
$$y = j C_{1,2}\,\omega\, V_c$$



MAXIMUM TIME SHIFT OBTAINED WITH VARIOUS "MOS" TECHNOLOGIES DISTURBED BY MAGNETIC COUPLING OR ELECTRIC COUPLING



PHYSICAL INTERPRETATION OF THE TIME SHIFT

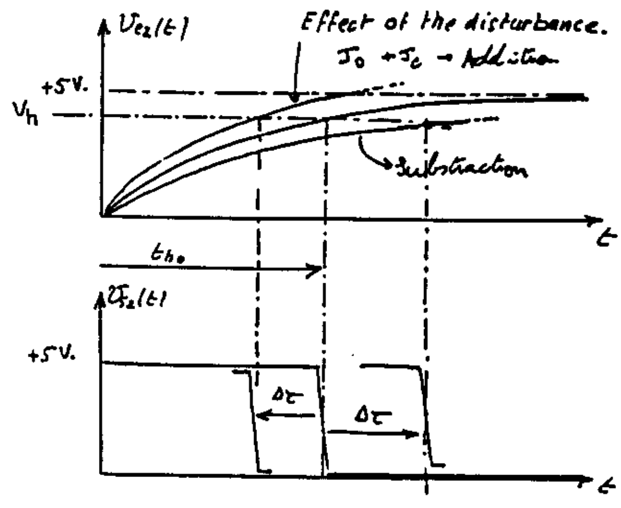


- J_0 current source due to the disturbance
- C Capacitance of the driving line between circuit (1) and circuit (2)
- r_d dynamic output resistance given by circuit (1)
- J_c transient current injected by circuit (1) during the digital switch.

$$J_c = g_0 V_{e_1}$$

g_0 equivalent transconductance

J_0 and J_c → addition or substraction depending on the alternance of J_0 at the time of commutation



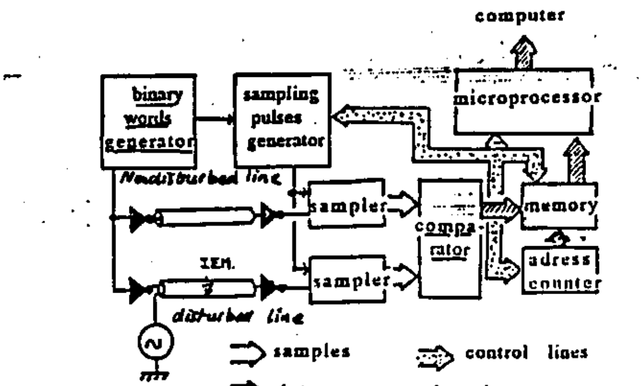
$$\Delta\tau_{maxi} = -\tau_0 \,\mathrm{Log}\left|\frac{1-\dfrac{V_h}{J_c' r_d}}{1-\dfrac{V_h}{J_c r_d}}\right|$$

$$J_c' = J_c - J_0$$

$$\tau_0 = C r_d$$

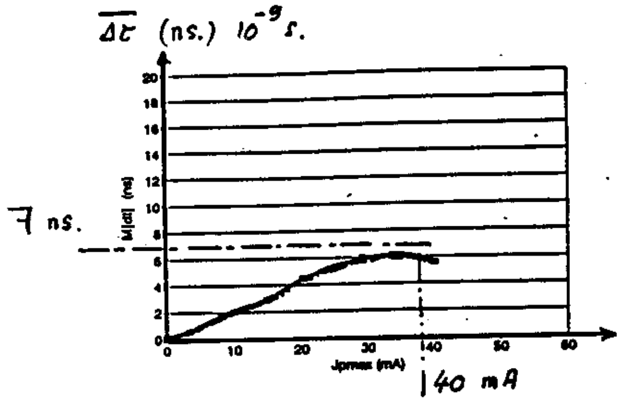
AC Tech.

MEASUREMENT OF THE TIME SHIFT BY MEANS OF A STATISCAL APPROACH



We compare the rise time and decay time of the disturbed and non disturbed BINARY WORDS.

TIME SHIFT (expressed in terms of mean value VERSUS THE CURRENT SOURCE AMPLITUDE

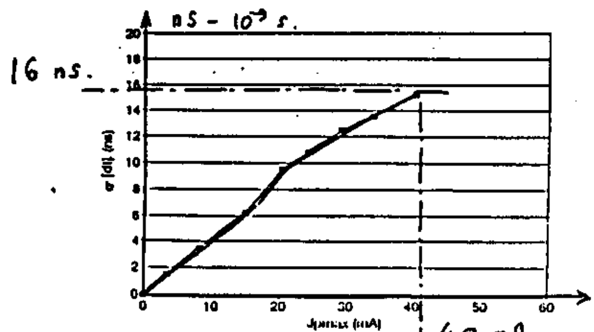


Equivalent current source due to the electric coupling

(peak value)

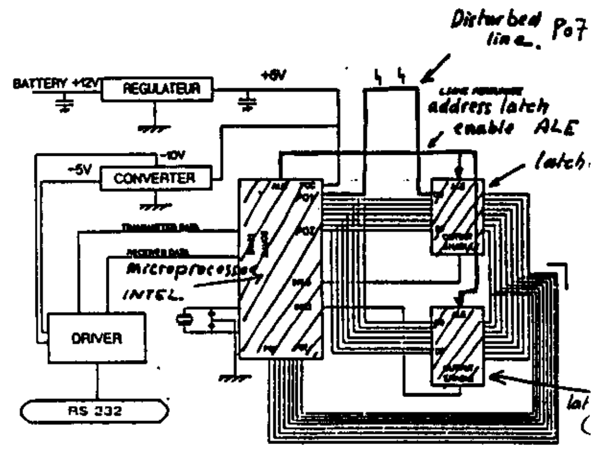
TIME SHIFT (expressed in terms of Standard deviation) VERSUS THE CURRENT SOURCE AMPLITUDE

$$\sigma_{\Delta t} = \sqrt{\overline{\Delta t^2} - (\overline{\Delta t})^2}$$



Equivalent current Source due to the electric coupling

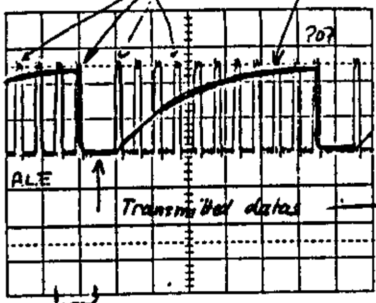
TEST MADE ON DATA LINES BETWEEN A MICROPROCESSOR AND A LATCH CIRCUIT



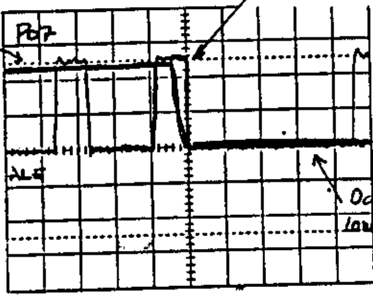
Errors on the datas transmitted by one line "Po7" disturbed by an electric coupling may be detected.

10

ALE Signal and Voltage on line Po7 without disturbance

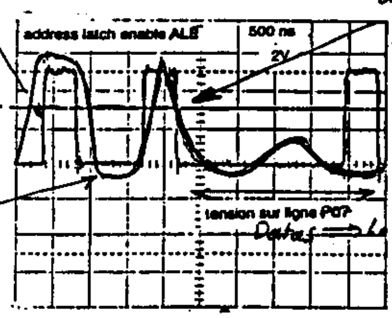


zoom ↓



Effect of the EMI coupled on the data lines!

Corresponding to a disturbing current source equal to 30 mA



During this time → high input impedance important voltage on line Po7

At the instant of validation of the first data (adress) → transition high input imp → low input imp. but due to the capacitance of the line ⇒ level on line Po7 can be still higher than the threshold of the high ... on the second data

MEASUREMENTS ②

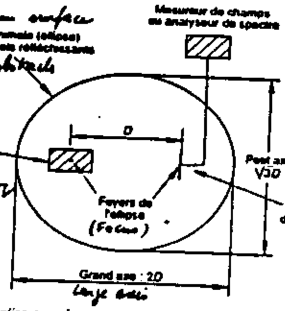
1) Emitted disturbing signals

"Free space" → avoid the presence of obstacles between the object under test and the receiving antenna

D = 100 m	f ≤ 30 MHz
D = 30 m	30 MHz ≤ f ≤ 100 MHz
D = 10 m	100 MHz ≤ f ≤ 300 MHz
D = 3 m	300 MHz ≤ f

("Measurement of the radiated field")

Normalization CCISPR, FCC, VDE



— Site normalisé pour la mesure en espace libre de perturbations rayonnées.

But reflections on the ground surface

Change the height of the receiving antenna → Max. signal

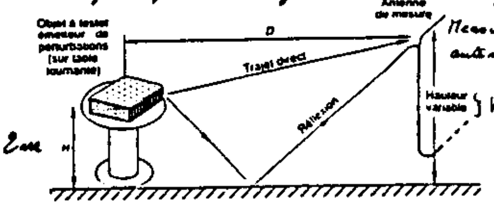


Fig. 10.2 — Réflexion sur le sol dans une installation en espace libre.

Other problem : Measurement of all the E.M. environment! → Choose a very quiet site!

Ambiant noise

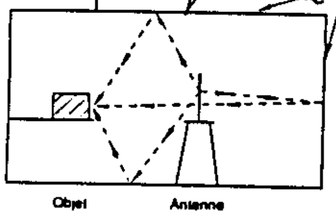
object but receiving antenna

Avoid ambiant noise ? → Make measurement in a shielded room

Drawbacks? wall → metal

⇒ internal reflections

⇒ resonance at high frequencies and the receiving sig. depends on the position of the antenna



— Réflexions multiples dans une cage de Faraday.

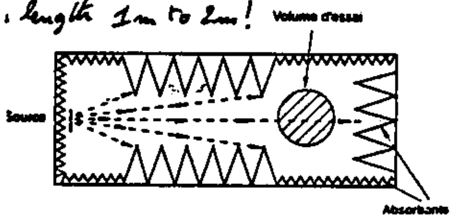
Used either for the illumination of the object or to measure the radiated field (by this object) (Compatibility + Susceptibility)

⇒ Put absorbing materials on the walls

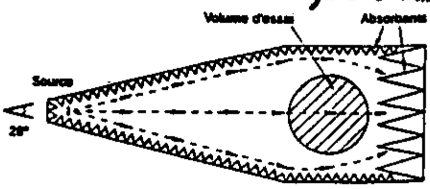
Example at μ. Waves

Performance ↗ if f ↗
(few GHz)

At 50 MHz - 100 MHz, to get acceptable performance cons. length 1m to 2m!



– Utilisation d'absorbants dans une chambre rectangulaire.
Rectangular chamber.



— Chambre anechoïde adaptée (Tapered anechoic chamber).

Another solution: Do not use the "free-space" radiation of a source but a guided TEM wave

→ { TEM line
T.E.M cell

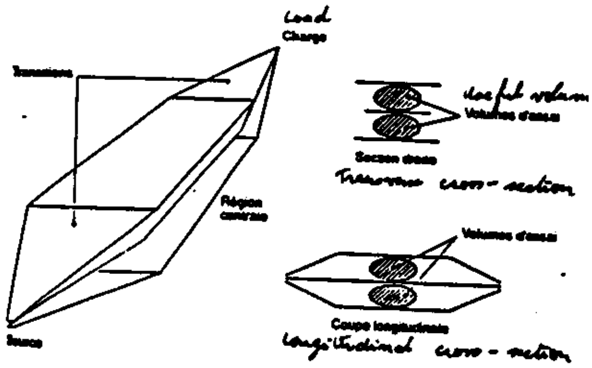


Fig. 10.7 — Ligne triplaque. . TEM line

Drawbacks. At high frequency, the transverse dimen of the cell ~ same order of magnitude as λ ⇒ high order modes → E^c ?

Usual cells h ~ 1m ⇒ f_{max} ~ 100 MHz

13

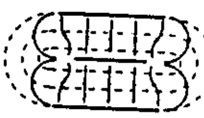
To avoid the coupling between the inside and the outside regions of the T.E.M. line → complete shield

⇓

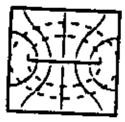
vertical metallic walls

⇓

TEM cell

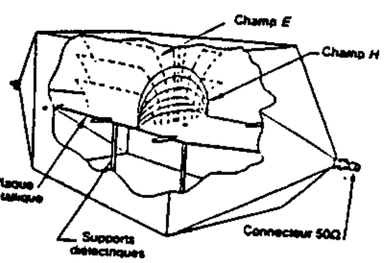


Ligne triplaque TEM line



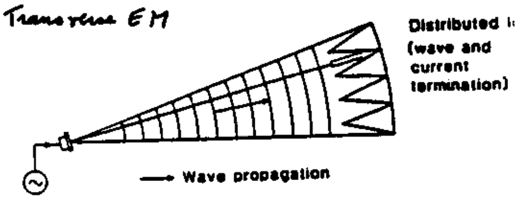
E field lines are less uniform

Cellule TEM TEM cell (compared to TEM line)

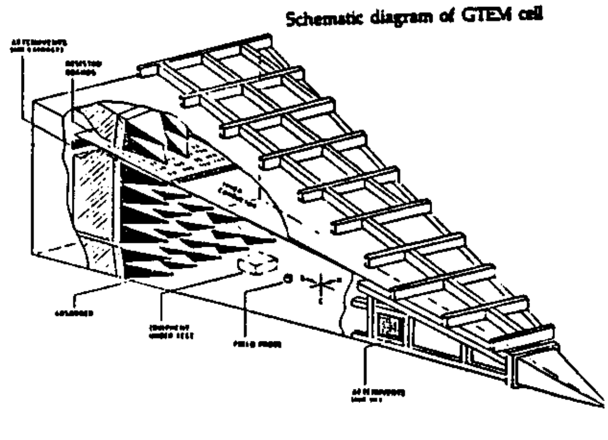


Cellules TEM.

GTEM cell
↑
gigahertz Transverse EM



Schematic diagram of GTEM cell



Cutaway drawing of a GTEM cell. (Reproduced by permission of Ray Pond and ABB Research Centre, Sw

Guide rectangulaire - ouverture faible (15°) - terminé → absorbant : front d'onde légèrement sph

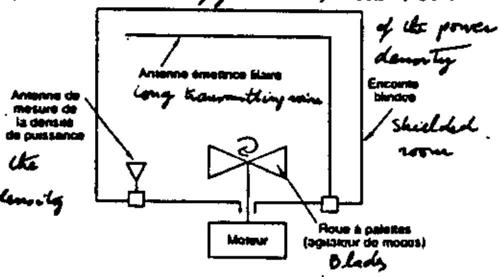
Avantage principal: champ pratiquement uniforme ds V.L.F → 1GHz. Possibilité de gde taille

Ds 1/3 du volume total (⇒ volume utile) uniforme du champ ± 1dB, variation de ± 3dB en fct de fréquence

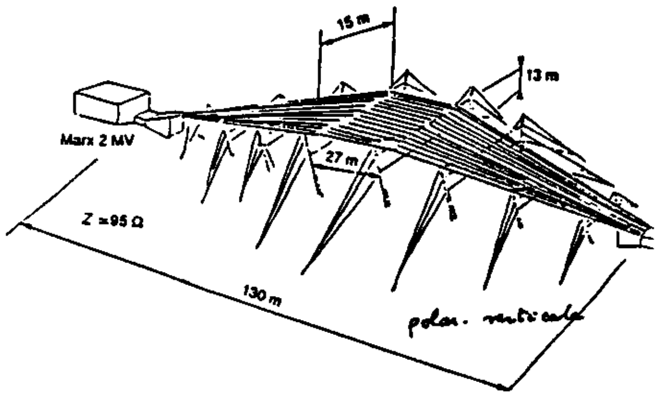
(23)

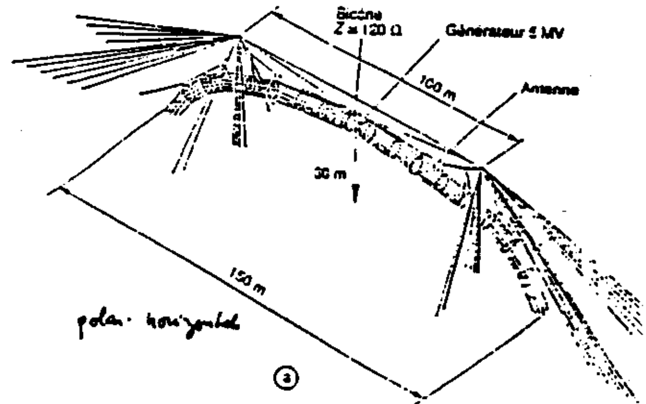
Mode stirred chamber

If $f \gg 100\,MHz$ (typically above few GHz)
How to define the incide
→ "Statistical approach"
Transverse dimensions of the room (reflecting walls) $\gg \lambda$ ⇒ Many modes ⇒ Moving (rotating) blades ⇒ change the mode configuration → mean value of the power density



" grands simulateurs d'impulsion. (onde plane)





③

(51)

(24)

Measurement of the conducting disturbances emitted by an object

Measurement on the power line or on the lines connecting the object and its peripherals for example
But a RSIL (Reseau stabilisateur d'impedance) "artificial mains network" is inserted to get
- A well defined impedance
- Filter the disturbance on the power line

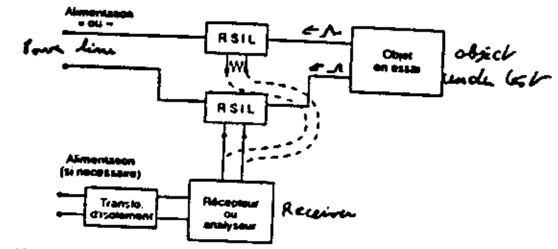
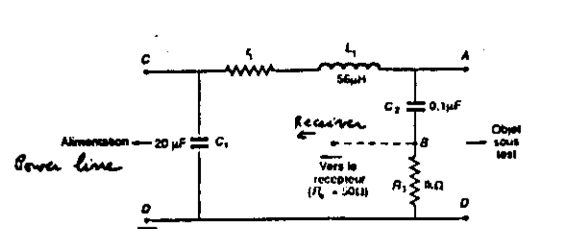


Fig. 10.12 — Mesure des perturbations conduites émises sur les conducteurs d'alimentation à l'aide de RSIL.



— Schéma simplifié du RSIL de la norme MIL-Std-462.
used if
$10\,kHz < f < 50\,MHz$

(25)

other possibility: use a current transformer

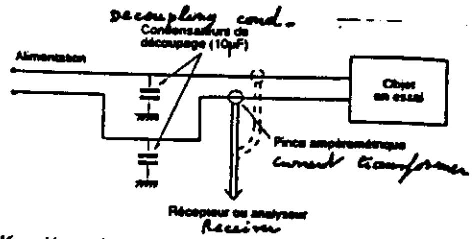


Fig. 10.16 — Mesure des perturbations conduites émises sur des fils d'alimentation à l'aide d'une pince ampèremétrique.

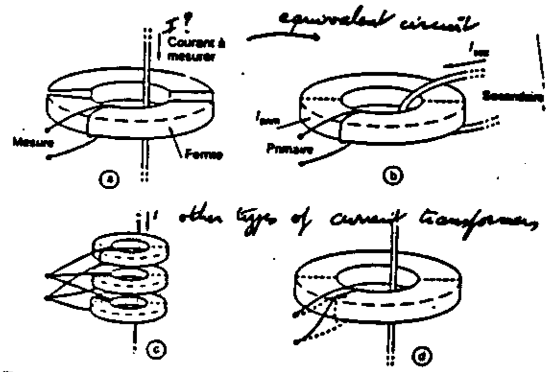


Fig. 10.17 — Pince ampèremétrique. a) pince; b) transformateur équivalent; c) et d) : variantes.

(51)

SUSCEPTIBILITY (26)

Determine the susceptibility level ⇒ inject disturbance on the object under test and verify its working

1) FREQUENCY DOMAIN

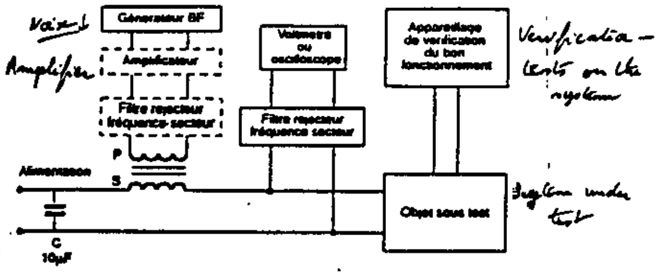


Fig. 10.19 — Mesure de la susceptibilité par conduction sur des fils d'alimentation à l'aide d'un transformateur de couplage («basses fréquences» : 30 Hz ≤ f ≤ 50 à 100 kHz). low frequency

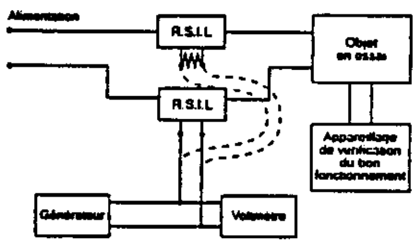


Fig. 10.20 — Mesure de la susceptibilité par conduction sur des fils d'alimentation à l'aide de RSIL («hautes fréquences» : 10 à 20 kHz ≤ f ≤ 300 à 400 MHz).

TIME DOMAIN (27)

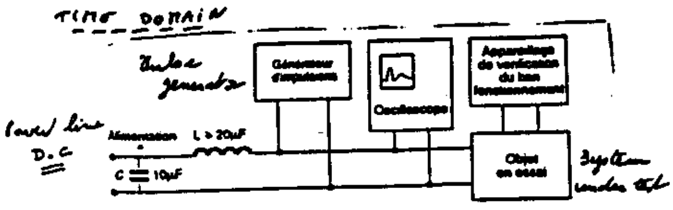


Fig. 10.22 — Susceptibilité aux signaux impulsionnels injectés sur les alimentations : montage série.

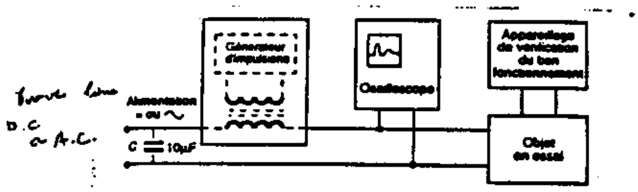


Fig. 10.21 — Susceptibilité aux signaux impulsionnels injectés sur les alimentations : montage parallèle.

7

(28)

Example of typical pulse injected in the power line (GAM T-13) + MIL 461

E : 100V up to 5kV

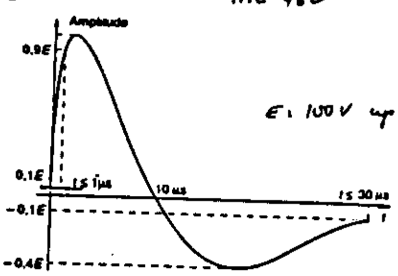


Fig. 10.23 — Allure des impulsions pouvant être injectées sur les alimentations pour une mesure de susceptibilité [13].

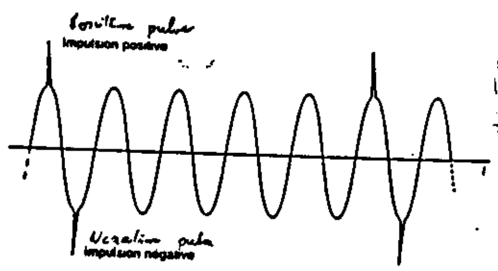


Fig. 10.24 — Synchronisation d'impulsions perturbatrices de fréquence de répétition 10 Hz sur secteur 50 Hz.

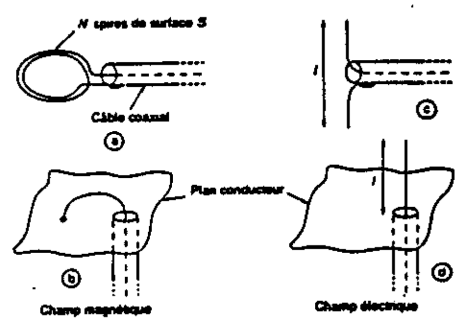
Synchronization of the disturbing signal with the AC of the power line

Measurement of radiated signals

2 types → Near field measurement
Far field

At low freq. (30Hz – 30MHz)

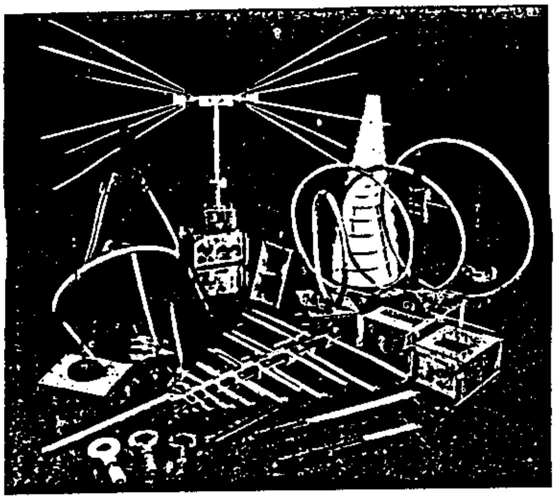
small loops small dipoles (l << λ)

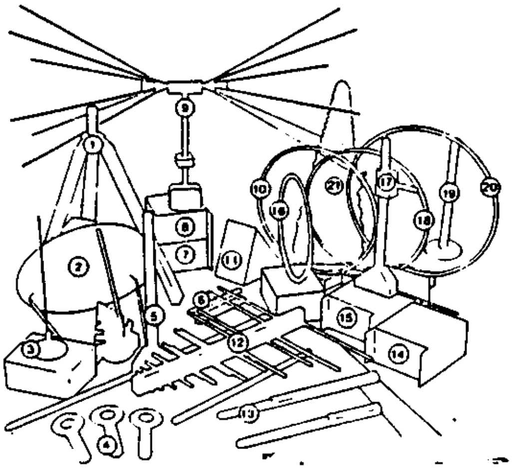


— Antennes pour mesures en champ proche.

At high freq or in the far field : usual receiving antennas = loops, biconical ant., log-periodic ---

8





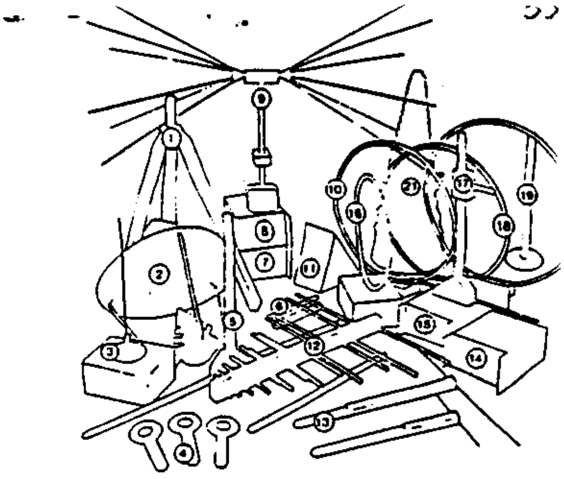


Table 5.3. Antennas and associated equipment shown in Figure 5.20, supplie[d by]
Electro-Metrics Inc., New York

Item	Name	Antenna type	Frequency ran[ge]
2	MTA-60	Parabolic	1–10 GHz
3	RVR-30M	Manual vertical	9 kHz–30 MH[z]
4	MFA/B/C	Magnetic field probes	9 kHz–230 M[Hz]
5	BDA-30	Broadband dipole	20–220 MHz
6	EFP-30	Electro-static probes	9 kHz–1 GHz
9	BIA-30C	Collapsible bi-cone	30–200 MHz
10	ALP-51H	Loop	100 kHz–30 M[Hz]
11	RGA-60	Double ridged guide	1–18 GHz
12	LPA-30	Log periodic	200 MHz-1 G[Hz]
13	TDA-30/1-4	Tunable dipoles	28 MHz–1 G[Hz]
16	ALR-30M	Manual loop	9 kHz–30 MH[z]
17	BDA-30S	Broadband dipole	20 MHz–200 [MHz]
18	ALP-51L	Loop	30 Hz–30 MH[z]
19	ELS-11	Sensor (magnetic field)	20 Hz–50 kHz
20	ALP-70	Loop	10 kHz–30 M[Hz]
21	LCA-30	Conical log spiral	200 MHz–1 G[Hz]
		Associated equipment	
1	TRP-163	Support tripod	
7	BPA-1000	Broadband preamp	10 kHz–1 GH[z]
8	CIG-25	Impulse generator	
14	PA-1000	Power amplifier	10 MHz–1 G[Hz]
15	FE-60	Frequency extender	

[illegible] probes cover this frequency range

(35)

Measurement in "free space"
(inside a shielded room

(36)

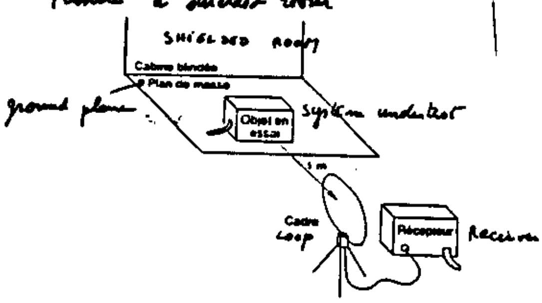


Fig. 10.29 — Mesure des perturbations rayonnées. Champ magnétique.
Magnetic field

SUSCEPTIBILITY MEASUREMENT

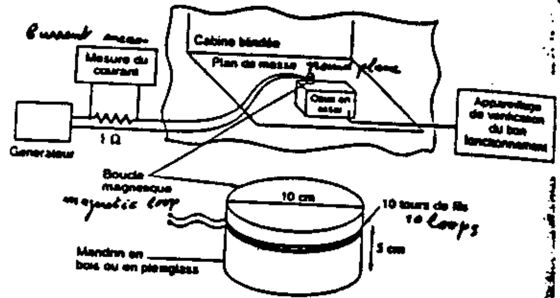


Fig. 10.31 — Mesure de la susceptibilité par rayonnement à l'induction magnétique (30 Hz ≤ f ≤ 50 KHz).
Susceptibility to the mg field induc.

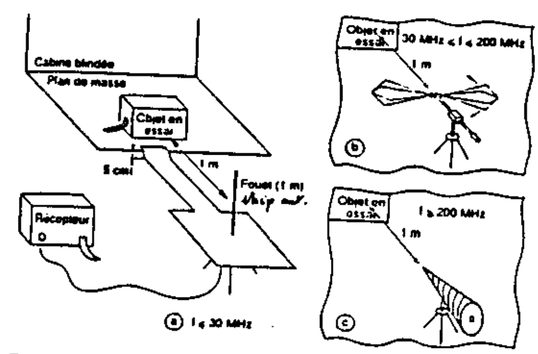


Fig. 10.30 — Mesure des perturbations rayonnées. Champ électrique.
Electric field.

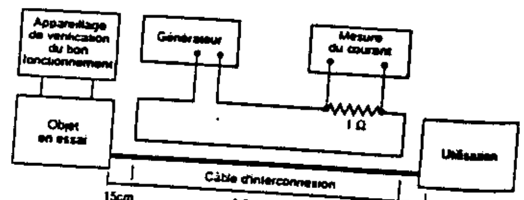
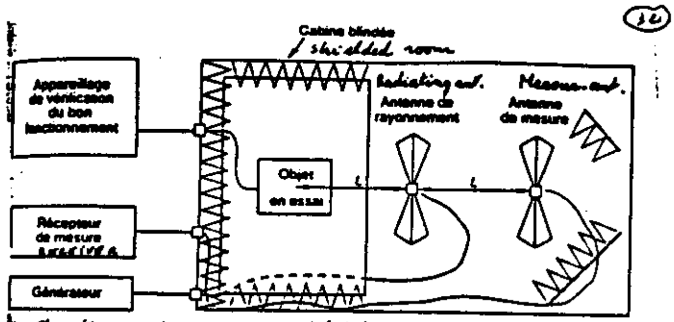


Fig. 10.32 — Mesure de la susceptibilité par rayonnement à l'induction magnétique appliquée aux câbles d'interconnexion (30 Hz ≤ f ≤ 50 kHz).

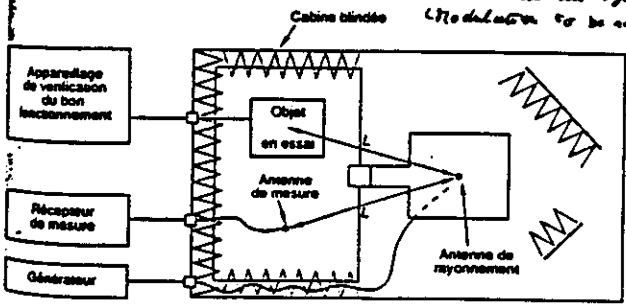
Same but to test interconnecting cables.

(8)



The transmitter can be modulated to produce the most critical disturbance on the object (Modulation to be adjusted)

(a) Utilisation d'antennes biconiques (30 MHz ≤ f ≤ 200 MHz)



(b) Utilisation d'antennes fouet de 1m (10 kHz ≤ f ≤ 30 MHz) Whip antenna.

or TEM cell

Fig. 10.33 — Exemples de disposition pour mesures de susceptibilité au champ électrique en cabine blindée (vues de dessus). En général, la distance L vaut de 1 à 3 m selon le type d'essai.

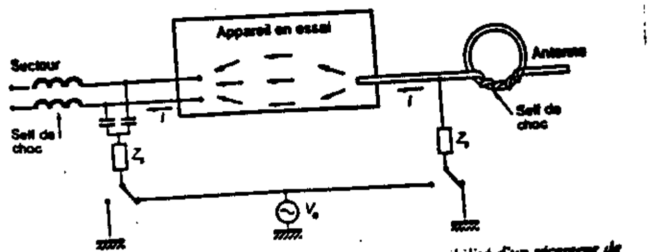
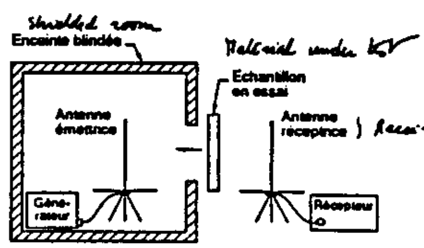


Fig. 10.35 — Méthode d'injection pour mesure de la susceptibilité d'un récepteur de télévision dans la bande 3-200 MHz [37].

If the main coupling mechanism = coupling E^i, H^i to cables connected to the equipment ⇒ it is possible to avoid the illumination of the equipment but simply use a current injection.

Current generator has the same characteristics as the equivalent Norton generator (equivalent for the current induced on cables)

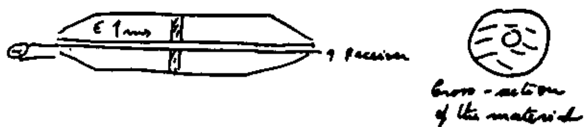
example: Susceptibility of T.V. set.

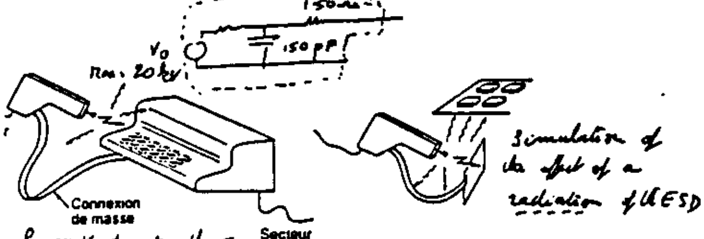


- *Mesure d'efficacité de blindage à l'aide d'une enceinte blindée.*

Measurement of the shielding efficiency of materials,

or conical cell.





Current due to the ESD injected on the object.

(a) (b)

– *Exemples d'emploi d'un simulateur de décharges électrostatiques.*

Susceptibility of equipment to electrostatic discharge.

Annex

- Source of interferences
- Coupling to cables

NATURAL NOISE - (1)

→ "Lightning phenomena"

1) Magnetic field on the ground surface. Magnetic storms

Earth → Magnetic dipole N-S. ($\vec{H}$, horizontal line) depends on the position of the observation point

$\vec{H}$ { H_m → due to the earth nucleus
{ H_a → due to the "superficial" layer

and far from the earth — R, d ~ 10R ← solar particles — pressure → distortion of the magnetic field lines

Variation of H → slow: { thermal mvt of the atmosphere, tide ...

→ rapid: interaction between the magneto- and the solar particles ⇒ magnetic storms. Amplitude of few tens of γ (10^{-9} T or $7.96\,10^{-4}$ A/m) ⇒ "Micro"pulses: duration few minutes to few hours.

Variation of B ⇒ Variation of E ($E/H = 120\pi$)

1 γ ⇒ 0.3 V/m → (Figure)

2) Global electric circuit

Lower atmosphere → slightly conduct. Interaction between solar X, UV rays, cosmic rays, α and β particles (radioactive component of the earth crust) ②

- Between 0 and 100 m → main ionizing cause: RADON (from U_{239}) ⇒ α
- Cosmic particles ↗ if altitude ↑. Max at 15 km
- Altitude of 60-70 km → mainly X, UV solar rays.

Fine weather conditions ⇒ σ_{air} (ground level) = 1.33 10^{-14} S/m

Integration (R) ionosphere ⇒ R_{eq} ≈ 230 Ω

Assuming lower layer of the iono. electrically charged

⇒ iono / J ↓↓ E_y earth — $|E_y|$ = 100 V/m

$\vec{J} = \sigma \vec{E}$ ⇒ $|\vec{J}| \approx 2\,10^{-12}$ A/m²

Integration I_{eq} ≈ 1 kA ; Voltage $V_a - V_o = R_{eq} \times I$ = 230 kV

Discharge? ⇒ other phenomena: lightning "activities"

About 2000 storms → simultaneously ⇒ global atmospheric elec. circuit.

3) Atmospheric noise

How to characterize the lightning "activities" on a site?

⇒ keraunic level.

keraunic level → days per year when the thunder has been heard. Why this criteria (vibration - sound and not EM?) → limitation of the area (sound → range of ≠ 10 km). Example - see p5

Not a quantitative criteria. 1 lightning stroke → same "weight" as a succession of lightning discharges a whole day. (But practical way - still used)

Empirical formulas → ground strike point / km²

Electric discharge → E.M. radiation in a wide freq. range.

Correlation between keraunic level and E.M. radiation —

"CIGRE" antennas: vertical antennas (1 m or 2 m high), tuned at few kHz, adjustable sensitivity → detection at a distance of 10 km - 20 km

- optical data from satellites → world-wide scale.
- Knowledge of the E.M. signature ⇒ goniometric systems (p9(?)) → determination of the ground strike point

- Statistical analysis of $\frac{\text{Intracloud}}{\text{Cloud to ground}}$ discharge (Important to define the mean disturbing field in a given region)

This ratio strongly depends on the latitude (p.7)

Example of H variation due to a mag. storm

Enregistrement d'un signal H dû à un orage magnétique (1986)

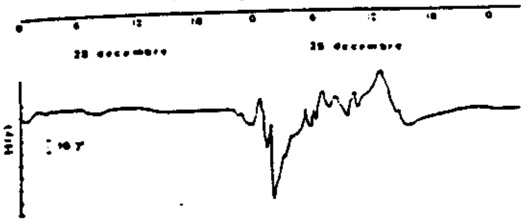
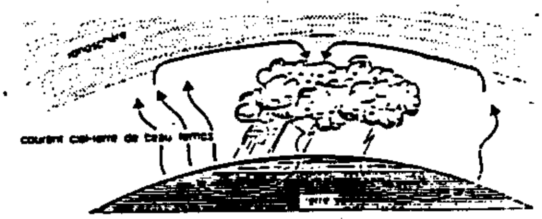


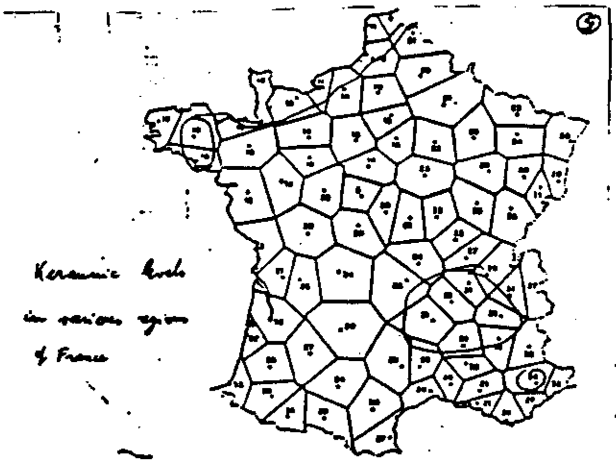
Schéma de circuit électrique atmosphérique global 1



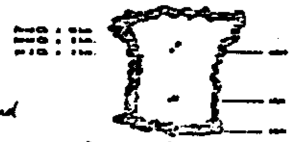
Mouvement des porteurs de charges négatives autour de l'orage

Movement of the negative charges

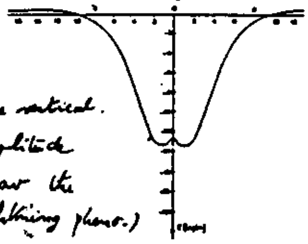
②



Keraunic levels in various regions of France



Charge distribution inside the cloud



Variation of the vertical electric field amplitude (ground level) below the cloud. (before lightning phase.)

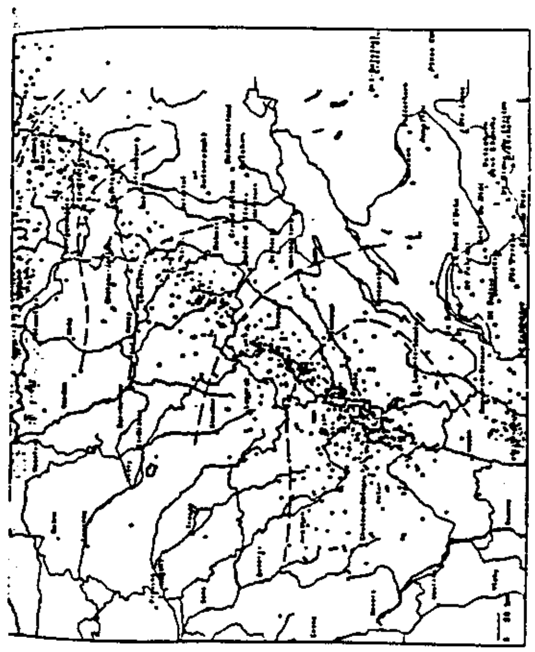
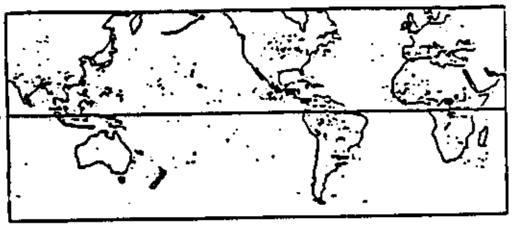
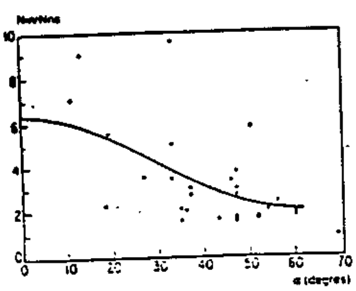


Fig. 2.74 — Activité orageuse enregistrée par le réseau Météorage entre 4 et 10 h TU le 27 septembre 1987 sur l'est de la France [105] (l'original de ce document est en couleurs [105]). La progression de l'orage heure par heure est indiquée approximativement par les lignes pointillées. Nombre d'impacts : négatifs – 982, positifs + 70, total 1052. Par tranche horaire 4-5 h : 392 ; 5-6 h : 227 ; 6-7 h : 144 ; 7-8 h : 106 ; 8-9 h : 136 ; 9-10 h : 45.





Effects of the lightning discharge

→ local : few E.M. pulses of large amplitude (distance of tens kms)

→ global effect due to lightning "activity" on the earth surface

earth

iono

→ Propagation in the earth - ionosphere waveguide

"average noise" - "atmospheric noise" (whistlers)

Important in the low frequency range (< 100 kHz)

(- resonance
- attenuation of the waveguide)

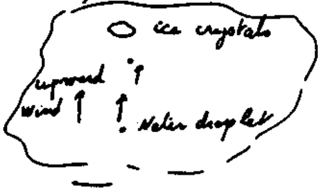
→ lightning discharge mechanism

→ local effect (E.M. field - currents ..)

→ Noise

1st step: Separation of the charge inside the cloud

Appears in cloud (cumulus) due to the thermal instability (hot and humid air in the lower layer) → cumulo-nimbus height ~ 10 km. Electrization within few hours. Many theories

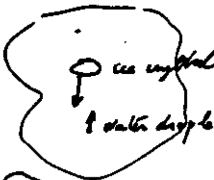


→ high altitude

→ ice crystals

crystal ↗↗ → weight ↑↑

Then



→ collision → charge transfer

crystal → $\bar{e}$ → −

droplet → − $\bar{e}$ → +

} statistically.

→ other charge pockets

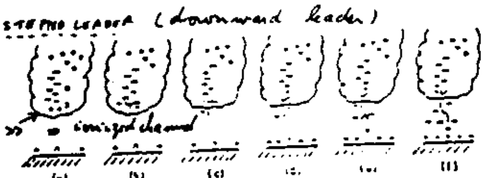
Complex phenomena (neutralization, breakdown...)

2nd step: Cloud to ground discharge

Total duration of the phenomena ~ 0.5 s

Successive stages.

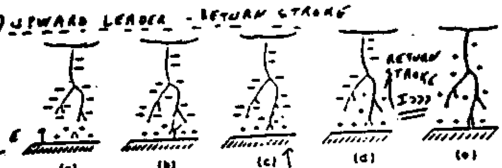
1) STEPPED LEADER (downward leader)



Each step → length of ~ 50 m, duration 1 μs. Between 2 steps: 50 μs. Velocity 1.5 10^5 m/s

Figure III.2: a) démarrage et propagation du traceur par pas; b) répartition des charges avant l'éclair; c) décharge préliminaire; d) à f) progression du traceur par pas vers le sol

2) UPWARD LEADER - RETURN STROKE



E ~ for 100 kV/m - Same - JUNCTION

Figure III.3: a) démarrage et propagation du pré-ionisateur; b) étape finale dans la descente du traceur par pas; c) démarrage du traceur ascendant; d) à e) propagation du pré-ionisateur du sol vers le nuage

3) SUBSEQUENT LEADER (Dart leader)

Subsequent return stroke

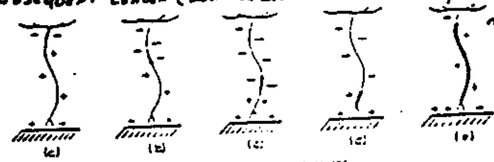


Figure III.4: traceur continu et arc-en-retour subséquent; a) à c) un traceur continu se propage dans le canal thermalisé par l'arc précédent; d) à e) propagation de l'arc-en-retour subséquent vers le nuage

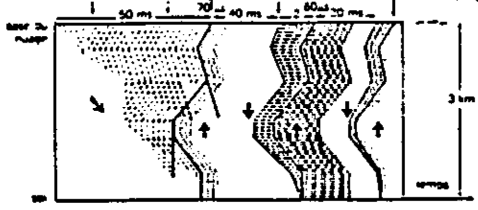
1st RETURN STROKE

Stepped leader / précurseur par pas

Dart leader / précurseur continu

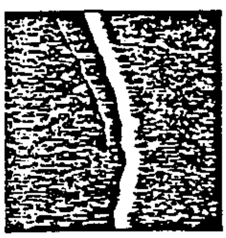
2nd RETURN STROKE / arc en retour

11



La première prédécharge (précurseur par pas) avance du nuage vers le sol par bonds successifs. Ce précurseur crée un canal fortement ionisé ; à l'issue de cette jonction, une décharge (arc en retour) va progresser du sol vers le nuage. L'arc en retour est caractérisé par l'accroissement brutal du courant, donc de la luminosité ; sa vitesse voisine de c/3 (c : vitesse de la lumière) est environ mille fois supérieure à celle du précurseur. Quelque temps plus tard (30 à 40 ms), un autre couple précurseur-arc en retour peut succéder au précédent, et ainsi de suite une dizaine de fois.

Photo : Groupe de Saint-Privat-d'Allier - Dimension horizontale : 1 ms, dimension verticale : 2 km



Cette photo illustre la figure ci-contre. A l'aide d'une caméra à défilement continu (caméra « streak »), nous pouvons observer le précurseur suivi par l'arc en retour, celui-ci étant caractérisé par une plus forte intensité lumineuse.
La décharge est particulièrement lumineuse dans la partie basse qui correspond à la présence du fil mais également ...

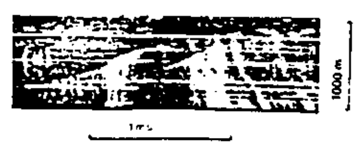
12

Negative downward leader

Negative upward leader

Positive downward leader

Positive upward leader



Différents types de décharges de foudre (d'après K. Berger [21]).
a) précurseur négatif descendant
b) ...

Study the phenomenal
↳ simulation (long sparks) (13)

. Streamers : energetic electrons moving in a cold plasma

. Streamers → leader ("Thermalization" of the medium)

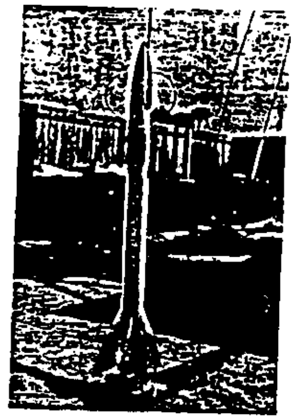
characterization of the discharge depend on
- the gap length
- the waveshape and the amplitude of the voltage

3 Développement d'une grande étincelle positive (50 µs entre vues successives).

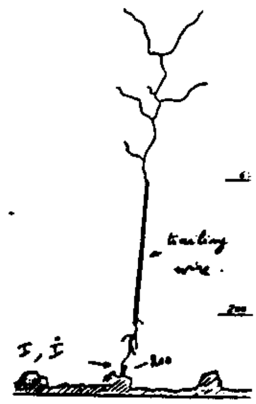
Diametre thermique du *leader* : 1 à 10 mm.
Champ électrique le long du *leader* : 500 à 1 000 V/cm.
Extension de la couronne (*streamers*) : 0,5 à 3 m.
Courant : 0,5 à 1 A.
Vitesse du *leader* : 1,2 à 2 cm/µs.
Champ électrique le long des *streamers* : 4 à 5 kV/cm.

Autre moyen d'étude : déclenchement (14)
→ lieu d'impact . mesure de I et E

other possibility : artificially triggered lightning
Rocket + trailing wire connected to a rod.



1 Système LRS.



2 Déclenchement LRSA (...)

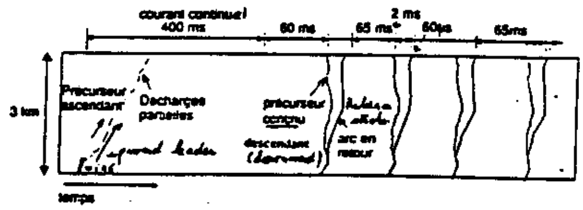
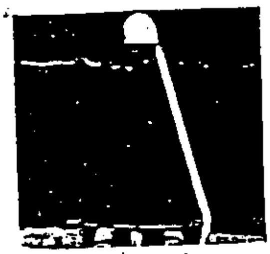


Fig. I-3 Evolution caractéristique d'une décharge déclenchée

Example of configuration (15)



2 Capteur de champ électrostatique.

Measurement of the electrostatic field ("Field mill")

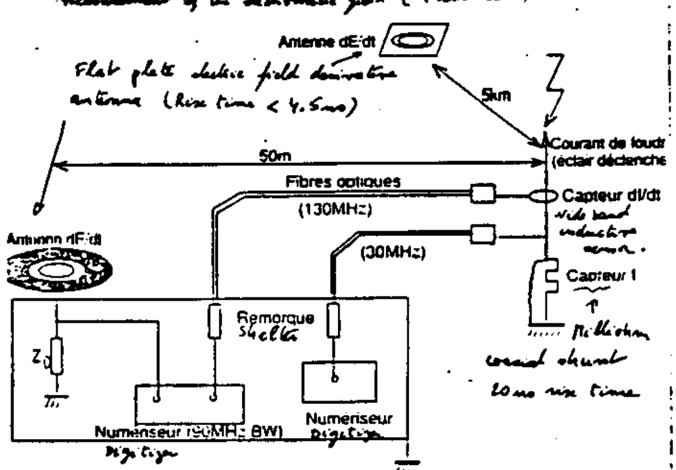


Fig. II-8 Synoptique des moyens de mesures en 1985 à Kennedy Space Center

CNET Lannion MAY 8...

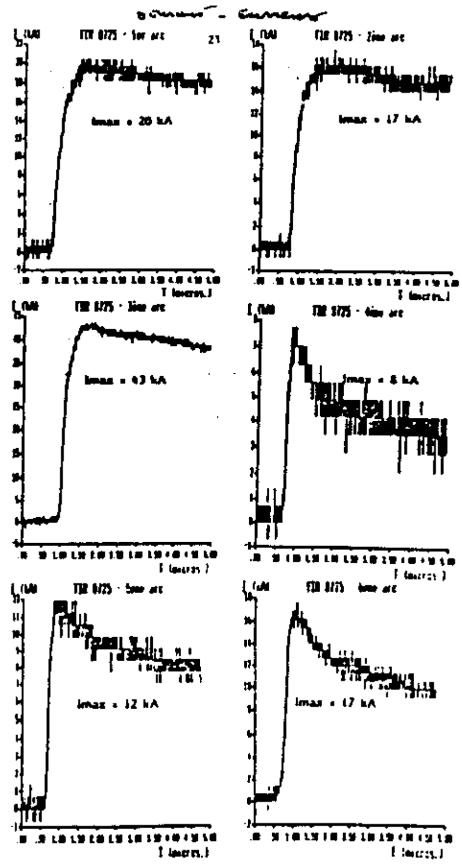


fig. I-5 éclair déclenché 8725 -
Impulsions de courant de chacun des 6 arcs en retour pendant 5 ...

Measured dI/dt (...) 55
$\dot{I}$

(17)

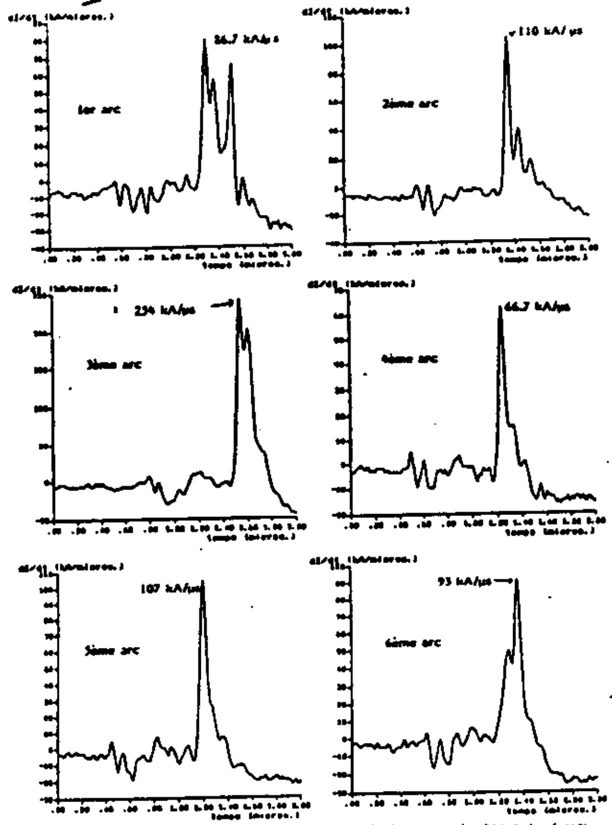


Fig. II-16 Eclair déclenché 8725 - Derivée temporelle du courant de chacun des 6 arcs-en-retour pendant 2 microsecondes. Les amplitudes crête varient de 65 à 255 kA/micros.

(18)

I crête

a)

dI/dt crête

b)

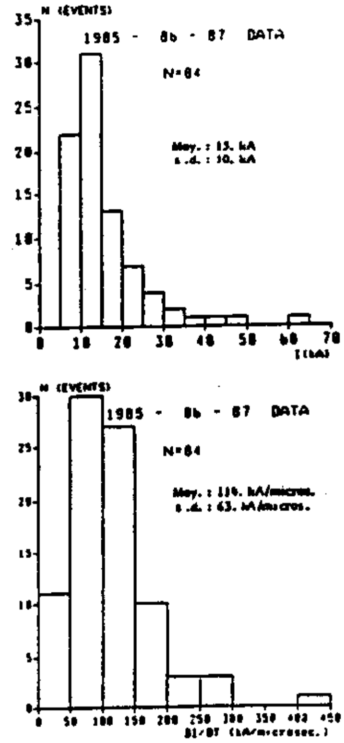


Fig. II-18 Histogrammes des valeurs crête des courants I mesurés et de la dérivée temporelle des courants dI/dt mesurées.

56

Mesure of E and $\dot{E}$ (19)

Measurement made at 50 m from the lightning.

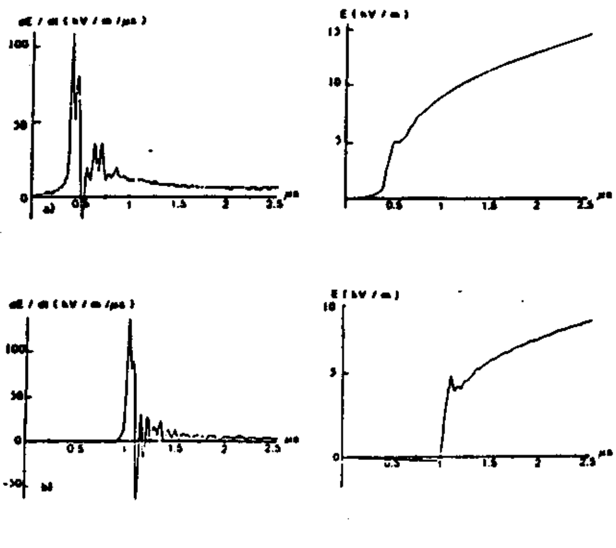
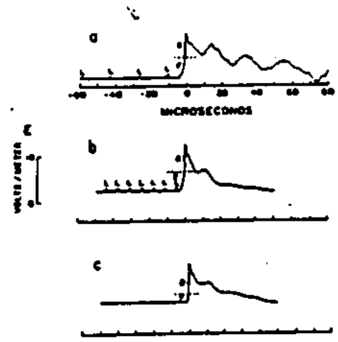


Fig. II-12 Signaux dE/dt enregistrés à 30 m de l'éclair déclenché et signaux $E = \int dE/dt$

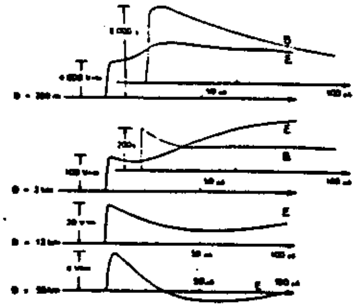
a) Floride - éclair 5304 - 4ème arc-en-retour; I crête = 11 kA (Peak current)
b) Floride - éclair 8533 - 3ème arc-en-retour; I crête = 10 kA

(20)



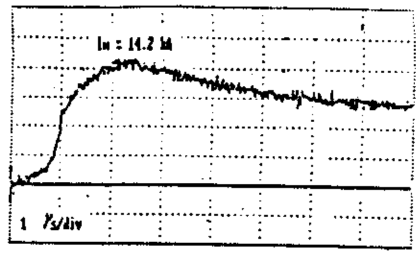
Typical waveshapes of the E field of
a) stepped leader (impulses) followed by the return stroke
b) Same as a) but for a subsequent return stroke
c) "continuous" leader followed by the subsequent return stroke.

21



Typical wave-shapes of the electric and magnetic fields.
Normalized to $I_{peak} = 10$ kA.

Distortion of the waveshape due to the decrease of the L.F. component ($1/R^3$, $1/R^2$) while the H.F. component $\propto 1/R$.



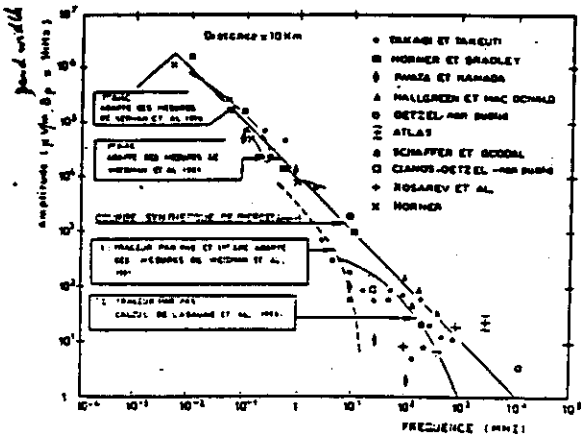


Figure III.7 : Domaine spectral de l'éclair

23

Summary

Stepped leader → "steps" of few tens of meters, 1 μs
velocity ~ $2 \cdot 10^5$ m/s. Between each step ~ 500 μs
Total duration ~ 50 μs. Total charge involved 5 C.
Peak current during the step: 1 kA, rise time < 0.1 μs
⇒ VHF-UHF radiation

Return stroke "Usually", current propagating along the ionized channel and neutralization of the charges in the channel. Upward velocity 0.8 c along the first 100 m, then $v \neq c/3$.
Total duration ~ 100 μs. Sudden thermalization of the channel ($t = 30.000$°K) ⇒ brightness of the channel + acoustic shock.

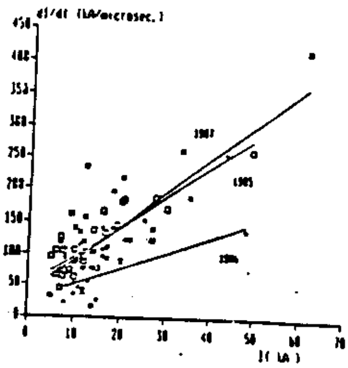
I_{peak} ~ 20 kA - 40 kA ; rise time < 1 μs
⇒ E, H (E ~ few 10 kV/m at ~ 100 m)

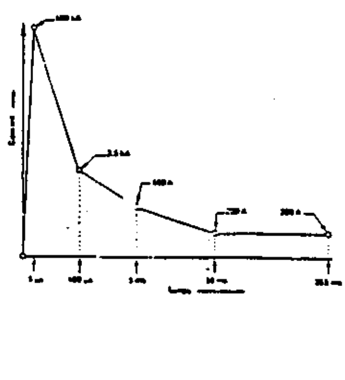
Continuous leader same way in the low resistive channel.
$v \sim 3 \cdot 10^6$ m/s ; $\bar{I} = 500$ A ; charge 1 C

Subsequent return stroke --

Number of return strokes varies but ≈ 10
During the 1st μs - E.M. power 10^9 W

24

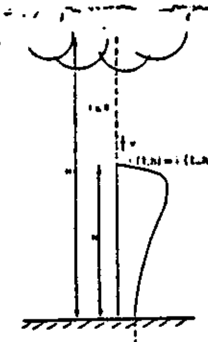




(7)

25

Modelization ?
coupling
to systems



" Transmission line model "

Figure III.9 : Modèle à onde progressive

Travelling current generator

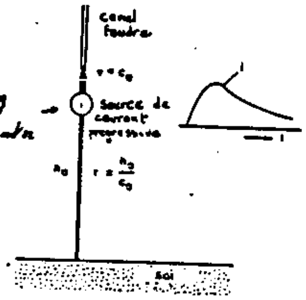
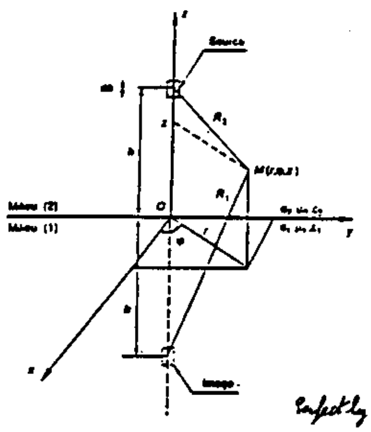


Figure III.12 : Modèle de la source de courant progressive proposé par Heidler

Dipole - Antenna - Sommerfeld - Appr. - -

26



Perfectly conducting ground.

$$E_z(t) = \frac{dh}{2\pi\varepsilon_0}\left[\frac{2h^2 - r^2}{R^5}\int_0^t i\left(\tau - \frac{R}{c}\right)d\tau - \frac{r^2 - 2h^2}{cR^4}\, i\left(t - \frac{R}{c}\right) - \frac{r^2}{c^2R^3}\frac{di}{dt}\left(t - \frac{R}{c}\right)\right]$$

$$H_\varphi(t) = \frac{dh}{2\pi}\left[\frac{r}{R^3}\, i\left(t - \frac{R}{c}\right) - \frac{r}{cR^2}\frac{di}{dt}\left(t - \frac{R}{c}\right)\right]$$

$$\frac{E_z}{E_r} = \sqrt{\varepsilon_r + \frac{\sigma}{j\omega\varepsilon_0}}$$

$$\begin{cases} E_z(H, t) = \dfrac{v}{2\pi\varepsilon_0 c^2 r}\left[i(t - r/c - H/v) - i(t - r/c)\right] \\ H_\varphi(H, t) = \dfrac{1}{2\pi c r}\displaystyle\int_0^H \frac{di}{dt}(t - r/c - h/v)\,dh \end{cases}$$

Influence of the ground conductivity ?
(far field)

Relative amplitude of E_z and E_r

27

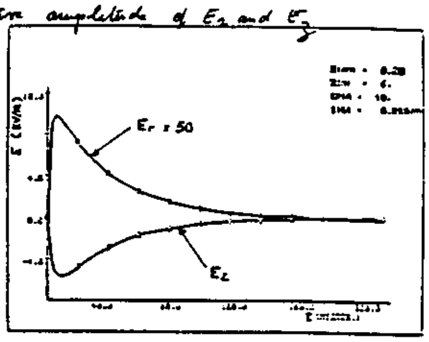


Figure III.31 : Composantes du champ électrique à 200 m

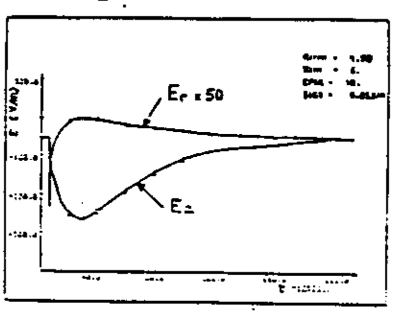
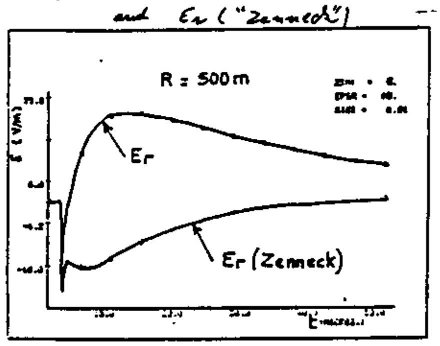
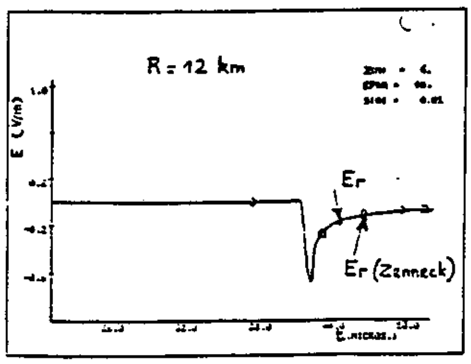


Figure III.33 : Composantes du champ electrique à 1,5 km

Comparison between E_r ("exact") and E_r ("Zenneck")





(29)

"Atmospheric noise"

→ Due to various lightnings occurring simultaneously around the earth + industrial activity (global) ⇒ average continuous noise

⇒ equivalent noise factor F_a (dB)

$F_a = 10 \log (f_a)$ when $f_a = P_m / k T_0 b$

P_m : average power (W) received by omnidirect. antenna

k : Boltzman constant , T_0 : temperature

b : receiver bandwidth.

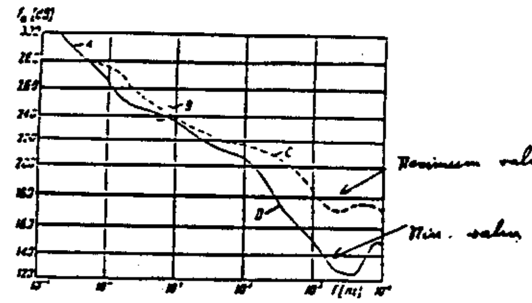
From F_a → deduce E_z

$$E_z = F_a + 20 \log f + 10 \log b - 95.5$$

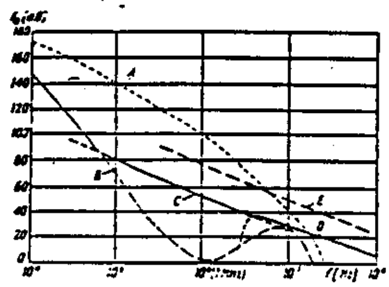
frequency expressed in MHz

⇒ statistical measurement

(10)



– on a given site –



A → Atmospheric noise (value exceeded during 0.5% of the time)

B → same but ⇒ during 99.5%

C → Man made noise (Residential area)

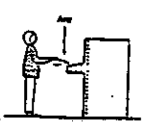
D → galactic noise

E → Man made noise (Industrial area)

(31)

ELECTROSTATIC DISCHARGE

Well-known effect on human body!



Due to "friction" between 2 different type of materials

1 | 2 Friction → increase of the contact surface between the media

⇒ charge on "conductors" 1 and 2

Discharge (breakdown) → important effect on the ground but also on satellite.

Discharge (Effect) ⇒ 1) pulse of current propagating on the structure

2) Radiation of this pulse

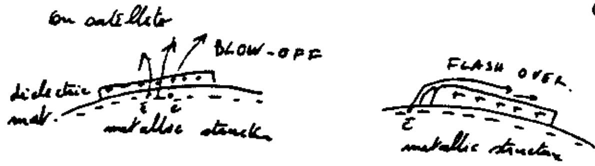
Length of the spark is very small → Electric moment of the antenna ≪ : Radiation only in the vicinity of the spark and not too critical

Failures due to the discharge current.

On the structure, pbs due to the insulating materials or lack of contact between conducting panels

(32)

On satellite



⇒ Not a direct breakdown on the component but electric pulse. At 30 cm

$|\vec{E}| = 1200$ V/m ; rise time ≠ 15 ns ; duration 40 ns

ENERGY

Composants COMPONENTS	Énergie [illegible] Perturbations (J) LATCH UP	Destructions (J) LASTING EFFECTS
CMOS, Circuits intégrés	10^{-7}	10^{-6} (Joules)
Transistors	10^{-4}	10^{-3}
Diodes	10^{-5}	10^{-4}
Redresseurs	10^{-4}	10^{-3}
Diodes Zener	10^{-1}	10^{-2}
Transistors de moyenne puissance	10^{-3}	10^{-4}
Relais Rectifying device	-	10^{-4}
Transistors de puissance Power Tr.	10^{0}	10^{1}
Diodes de puissance et SCR	10^{0}	10^{2}

(33)

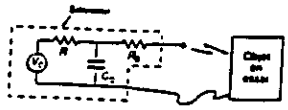
Tab. 5.1.4
Perturbations de composants électroniques soumis à une impulsion de choc rapide
Effect of a short pulse of current

Tab. 5.1.5 Seuils de destruction des circuits intégrés CMOS |5.1.7
Lasting effects on CMOS Integrated Circuits.

Circuit	Valeur minimum Minimum Value	Largeur d'impulsion Pulse width 10 ns	100 ns	1 µs
CD4001A Entrée	Tension (V)	350	150	60 Voltage
	Courant (A)	10	7	1.2 Current
	Puissance (W)	3500	1050	72 Power
	Energie (µJ)	87.5	105	72 Energy
CD4016 Entrée	Tension (V)	150	120	20
	Courant (A)	2.0	4.0	2.0
	Puissance (W)	300	480	40
	Energie (µJ)	7.5	48	40
CD4049 Sortie	Tension (V)	150	25	12
	Courant (A)	15	6.0	3.0
	Puissance (W)	2250	150	36
	Energie (µJ)	56.2	15	36
CD4050 Sortie	Tension (V)	170	60	20
	Courant (A)	13	7.5	3.0
	Puissance (W)	2210	450	60
	Energie (µJ)	55.2	45	60

(35)

Discharge due to a human body
Equivalent circuit $R_0 - C_0$



. Dry conditions $R_0 = 50\,k\Omega$

$V_{max} = 8.8\,kV$

$I_{max} = 0.18\,A$

Rise time of few ns

. Humid — $R_0 = 5\,k\Omega$

$\Rightarrow I_{peak} = 1.8\,A$

Important parameters:

- I_{peak}
- dI/dt
- Energy
- Repetition --

(34)

Man-made noise

⇒ characterization of the { "disturbing source" / "Coupling mechanism" }

Radiation ↙ ↘ Conduction.

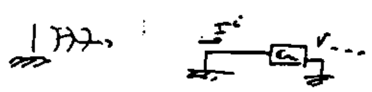
Radiated interference

A radiated interference is defined as any interference transferred through a medium by an electromagnetic field.

The spectrum of sources of radiated disturbance can be very extensive, covering frequencies going from a few kilohertz to a few gigahertz. Of course the amplitude of the incident electromagnetic field depends on the power at the source, but also on the distance separating the source from the disturbed circuits ; if the distance is large compared to the wavelength, we have a distant source and the dispersion of the radiated field is then inversely proportional to the distance from the emitter to the receiver. If the distant source condition no longer holds, the law of variation of the field with this distance is a much more complex affair and depends very much on the geometry of the source.

A few figures will help give an idea of the great diversity of amplitudes and frequencies that sources of disturbance can radiate.

An intentional radiation (e.g. broadcast--) can be a source of interference on sensitive electronic systems



"Authorized" emissions

(36)

Broadcast, T.V, Radio transmitter

- High power (few MW) ⇒ $E > 1V/m$ distance < 7.

Statistically, it appears that if $E > 1V/m$ on sensitive electronic equipment ⇒ latch up (misfunction)

F.M. radio $P > 10\,kW$ $f \approx 80\,MHz$

⇒ $E > 1V/m$ at a distance < 100 m.

Radars

1.2 GHz - 1.4 GHz, P_{peak} - 1.2 MW + antenna gain (30dB)

Pulses of few µs

At 1km - 2km → 3 V/m up to 120 V/m ! (peak value)

Disturbances of µP, satellite, ...

Mobile telephone

150 MHz, 200 MHz, 450 MHz, 900 MHz

P - 5W up to 20W ⇒ $E > 1V/m$ if $d < 10\,m$

EMC pb near the vehicle or inside the vehicle

→ Radiation out of the authorized B.W.

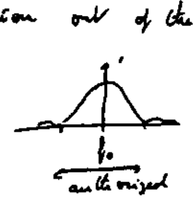


Table 1 Restricted Radiation Devices

Intentional Radiators
Campus radio stations
Wireless monitoring systems
Walkie-talkies
Low-power telemetering transmitters
Wireless microphones
Security transmitters
Intrusion detectors
Field disturbance sensors
Hearing or auditory transmitters
Door opener transmitters
Licensed transmitters
Phono oscillators

Unintentional Radiators
Cable television systems
Closed-circuit television systems
Microwave ovens
Industrial heaters
Radio receivers
Master antenna systems
TV cameras to monitor circuits
TV cameras to TV receiver circuits
TV/electronic games
Electronic equipment:
Computers
Digital weight scales
Data processing equipment
Tape recorders
Electronic watches
Switching power supplies
Digital displays
Devices using digital techniques

Table 2 Incidental Radiation Devices

Electric motors and pumps
Home appliances
(e.g., electric shavers, hair dryers, food mixers, etc.)
Ignition systems (and other automotive sources)
Light dimmers
Electric power lines
Fluorescent Lights
Defective insulators

Table 3 Sources of Radiated Interference

Source	Spectrum	Magnitude
Bistable circuits	15 kHz - 400 MHz	
Harmonic generator	30 MHz - 1000 MHz	
Heater thermoswitch		
Contact arc	30 kHz - 300 kHz 20 MHz - 200 MHz	
Actuator		
Motor	10 kHz - 400 kHz	
Switch arcs	30 MHz - 200 MHz	
Cams	10 MHz - 20 MHz	200 µV/m/kHz
Teleprinter		
Magnet armatures	1.8 MHz - 2.6 MHz	
Print magnets	1.0 MHz - 3.0 MHz	200 µV/m/kHz
Transfer switch		
Coil decay	15 kHz - 150 kHz	
Contact arc	20 MHz - 400 MHz	
DC power supply		
Switching circuit	100 kHz - 30 MHz	
Equipment case		
Unscreened access covers	10 kHz - 10 MHz	1000 µV/m/kHz
Fluorescent lamp		
Arc	100 kHz - 3.0 MHz	
Multiplexer		
Solid state switching	300 kHz - 600 kHz	
Power console		
Circuit breaker cam contacts	10 MHz - 20 MHz	
Power switching devices	100 kHz - 300 MHz	
Power controller		
Power wires	50 kHz - 4.0 MHz	
Chopper, relay, bistable circuits	10 kHz - 200 MHz	

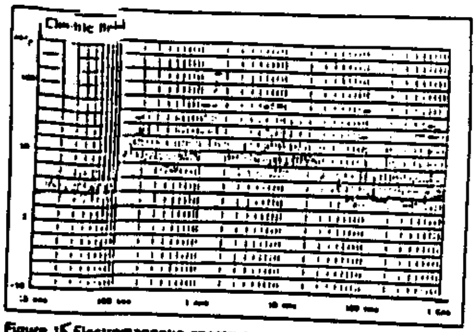


Figure 15 Electromagnetic spectrum recorded seven meters from the metro tracks, with no train passing.

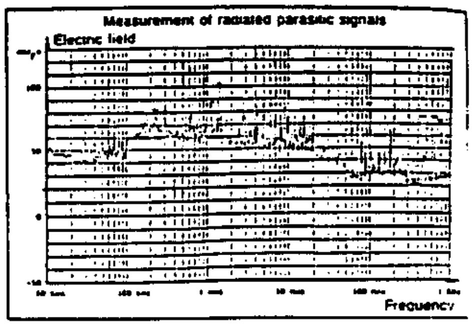


Figure 16 Electromagnetic spectrum recorded seven meters from the tracks with a train passing by.

Measurement of E radiated by a car

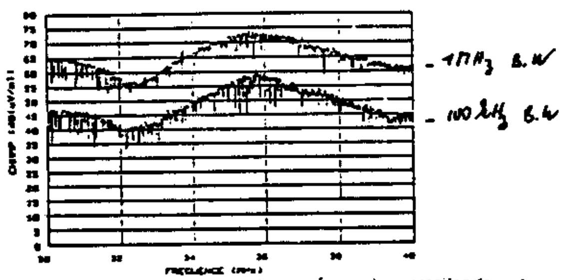


Figure 14 : Comparaison des mesures de champ crée par les parasites de voiture dans 1MHz et dans 100kHz de bande passante d'analyse

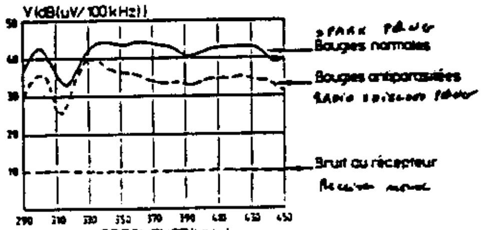


Figure 15 Diminution du bruit d'allumage à l'aide de bougies antiparasitées

Standards and specification

- Spectrum of the authorized emission

  Rules - U.I.T (Union Int. Telecom)

  IFRB (International Frequency Registration Board

- Committee:

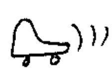
  CCIR → Radiocommunication
  Broadcast, mobile tel., Hertzian links
  satellite...
  → Divide the spectrum between the users!

  CCITT → Telegraph - Telephon
  for Telecommunication - Norm (Standard ISDN...)

- Limitation of the radiated levels

  CISPR - Comité Int. Special pour les perturbations radioélectriques

  Example



- MIL - STD

Conducted interference

In this case, the noise or the disturbing signal created by a device or an electric equipment propagates towards the "victim" through a metallic conductor such as wiring or any metallic structure. It is obvious that a H.F. or an impulse current propagating along a cable also radiates in the environment. Thus conducting and radiating processes will occur together. However, the classification between conducted and radiated interference comes from the fact that in many practical cases one of the two coupling mechanisms will be dominant.

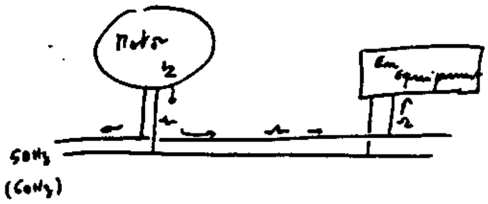


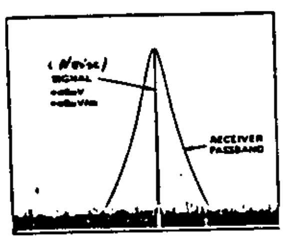
Table 4. Sources of Conducted Interference

Source	Spectrum	Magnitude
Heater Circuits (Contact Cycling)	50 kHz to 25 MHz	
Fluorescent Lamps	0.1 to 3 MHz (peak at 1 MHz)	20 to 300 μV/kHz
Mercury Arc Lamps	0.1 to 1.0 MHz	8000 μV/kHz
Computer Logic Box	50 kHz to 20 MHz	
Command Programmer		
signal lines	0.1 to 25 MHz	
power lines	1 to 25 MHz	
Multiplexer	1 to 10 MHz	
Latching Contactor		
coil pulses	1 to 25 MHz	
contact cycling	50 kHz to 25 MHz	
Transfer Switch	0.1 to 25 MHz	
Power Supply Switching Circuit	0.5 to 25 MHz	
Power Controller	2 to 15 kHz	
Power Transfer Controller		
constant noise	10 to 25 MHz	
transients	50 kHz to 25 MHz	
Magnet Armatures	2 to 4 MHz	20.000 μV/kHz (250 V transient spike)
Circuit Breaker Cam Contacts	10 to 20 MHz	
Corona	0.1 to 10 MHz	100 μV/kHz
Vacuum Cleaner	0.1 to 1.0 MHz	3000 μV/kHz

WIDE BAND - NARROW BAND INTERF.

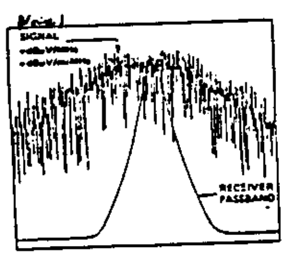
NOT an "absolute" cl. (intrinsic) charac. of the disturbance BUT also depends on the receiver B.W.

2.2. Classement par le spectre



→ the receiver can clearly identify various rays.

Bande étroite



Bande large

Wide band - example

$f(t) = c_0/2 + \sum_n c_n \cos(n\omega_0 t + \phi_n)$

$c_n = 2A \frac{\tau}{T} \frac{\sin(n\pi\tau/T)}{n\pi\tau/T}$

let $dF = 1/T$. If $T \to \infty \Rightarrow c_n$ and $dF \to 0$

but $Q_n = c_n/dF$ tends to a limit value

$\Rightarrow f(t) = c_0/2 + \sum Q_n \cos(n\omega_0 t + \phi_n)\, dF$

$\int Q_n \cos(n\omega_0 t + \phi_n)\, dF$

Q_n - spectrum density of Fourier coefficient

$Q = \lim_{T\to\infty} Q_n = \lim_{T\to\infty} c_n T = \lim 2A\tau \frac{\sin n\pi\tau/T}{n\pi\tau/T}$

$\Rightarrow Q = 2A\tau$

→ Instantaneous peak value measured with a B.W = B

$V_c = 2A\tau B$

• $V_c \sim B$ → only for wide band impulse signal

• White noise : often extreme case $P_c \sim B \Rightarrow V_c \sim \sqrt{B}$

general → $V_c = a B^{\alpha}$ $0.5 < \alpha < 1$

(35)

Table 5 Radiated disturbing sources.

	Electric field E	Magnetic field H	Power surface density dP
Narrowband	V/m	A/m	W/m²
Wideband	V/(m.Hz) or V/(m.√Hz)	A/(m.Hz) A/(m.√Hz)	W/(m².Hz²) or W/(m².Hz)

Table 6 Conducted disturbing interferences.

	Voltage V	Current I	Power P
Narrowband	V	A	W
Wideband	V/Hz or V/√Hz	A/Hz A/√Hz	W/Hz² or W/Hz

Narrow band or Wide band interference on a signal receiver

Narrow band → Measuring instrument → output that rises sharply as the instrument is tuned

Wide band → low (small) variation.

criteria: Change the bandwidth B of the receiver

IF NOT POSSIBLE
Change the tuned frequency

Centre Frequency $\{ F_0$ Bandwidth $B_0 \to U_0$ (measure.

⇓

Bandwidth $B_1 \to U_1$

If $\Delta U > 10 \log \frac{B_1}{B_0}$ → Wide band

$\Delta U < 3dB$ → Narrow band.

$F_0 \to F_1 = F_0 + B \quad (B = cte)$

$\Delta U > 3dB$ → Narrow band

$\Delta U < 3dB$ → Wide band

(15

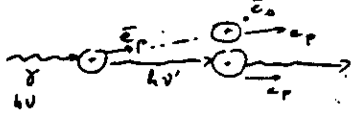
III.4) Nuclear E.M. pulse

The nuclear electromagnetic pulse is generated by the interaction between almost the gamma rays and the atmosphere. Indeed when a gamma ray generated by a nuclear bomb interacts with an atom, there is a creation of a secondary gamma but the energy is important enough to also remove one electron of the atom, as shown in Figure 17 |2|. This Compton

Figure 17

effect occurs since the average energy of the incident gammas are greater than 1 MeV. Then there is a cumulative effect giving rise to an important current density in air. The phenomen is very complex since both the amplitude and the direction of the current density vector will depend on the characteristics of the air, on the presence of the soil, on the interaction with the terrestrial magnetic field...

Then



⇒ Compton electron → current J_2
(radial current near the source.

⇒ Secondary electron → current J_s

$\overleftarrow{J_s}$ $\overrightarrow{J_2}$ $\}$ → total current

For a high altitude explosion, the terrestrial magnetic f. plays a leading part in the distortion of the current lines. (Figure

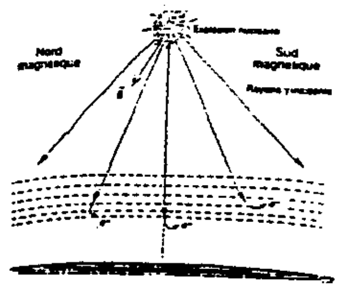


Figure 18

As a result, the electric field at the ground level radiated by th currents has no more a revolution symmetry around the vertical line the explosion point. (Figure 19)

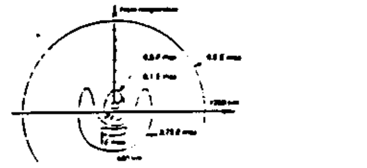


Figure 19

Theoretical calculation of the electric field leads to an electr pulse having a rise time of few nanoseconds and a decay time of a 300 ns. The peak value of this short pulse depends on the type of weapon but a value as high as 50 kV/m can be assumed.

This pulse is generated over a wide area (see Figure 19) couples particularly to long wires, such as the e... telecommunications cables, producing a short peak of current, but high amplitude, on these cables.

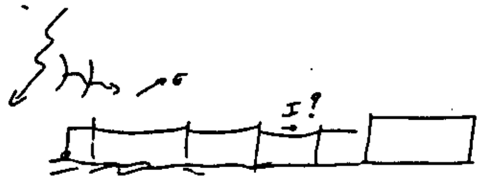
(3)

COUPLING
TO STRUCTURES

→ Disturbing current on the equipment
(Conducted interference)

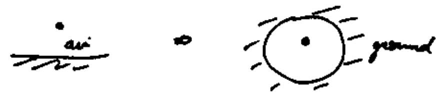
⇒ the disturbing signal is known

→ Radiated interference



Coupling field - structure.
Simplest case: cable parallel to a ground plane (infinite conductivity) or to the earth.

(1). 1) Understand the propagation



Characteristic impedance, surface impedance, external inductance ..

2) Coupling to a line // to ∞^{ly} conducting plane
- differential eq (volt. + currents)
- approximate formulas (short cables)

3) Cable (line) → current but if it is a coaxial cable or a multiwire shielded line.

Transfer impedance - Transfer admittance
Influence of the ground connection.

4) general case → Numerical solution.

Propagation in a homogeneous medium

(3)

$\text{rot}\,\vec{E} = -\frac{\partial \vec{B}}{\partial t}$ $\quad$ $\text{rot}\,\vec{H} = \vec{J} + \frac{\partial \vec{D}}{\partial t}$

$\text{div}\,\vec{B} = 0$ $\quad$ $\text{div}\,\vec{D} = \rho$ $\quad (0)$ hom. medium

$\vec{B} = \text{rot}\,\vec{A}$ $\quad$ $\vec{A}$: Vector potential

$\text{rot}(\vec{E} + \frac{\partial \vec{A}}{\partial t}) = 0 \Rightarrow \vec{E} + \frac{\partial \vec{A}}{\partial t} = -\text{grad}\,\phi$

ϕ : scalar potential

In the region free of electric charges.

$(\Delta - \Gamma^2)\vec{A} = 0$ $\quad$ $(\Delta - \Gamma^2)\phi = 0$

$\Gamma^2 = \mu \frac{d}{dt}(\sigma + \varepsilon \frac{d}{dt})$

The 2 terms $\vec{A}$ and ϕ can be related ⇒ expressed in terms of a new potential vector $\vec{\Pi}$ (HERTZ POT.)

$\vec{A} = \varepsilon \mu \frac{\partial}{\partial t}\vec{\Pi}$ and $\phi = -\text{div}\,\vec{\Pi}$.

$(\Delta - \Gamma^2)\vec{\Pi} = 0$ $\quad$ Frequency domain

$\Gamma \to \gamma$ $\quad$ $(\partial/\partial t = j\omega)$

$\gamma^2 = j\omega\mu(\sigma + j\omega\varepsilon)$: non constant

(4)

Plane wave propagating towards $z > 0$

E, H, z ⇒ $|E| = E_0 e^{j\omega t - \gamma z}$

$\gamma = \alpha + j\beta$ — α: Attenuation constant; β: Phase angle.

① High conducting medium $\sigma \gg \omega\varepsilon$

$\gamma^2 \simeq j\omega\mu\sigma \Rightarrow \gamma = \frac{1+j}{\delta}$ → skin depth

$\delta = \sqrt{\frac{2}{\omega\mu\sigma}} = \frac{500}{\sqrt{\sigma f}}$ $\quad (\mu = \mu_0)$

attenuation as $e^{-z/\delta}$

sea water $\sigma = 4$ S/m

ground 10^{-4} S/m $\leq \sigma \leq 0.1$ S/m
granite $\quad$ clay
+ depends on the water content

Typical value: $\sigma = 10^{-2}$ S/m

Copper $\sigma_{cu} = 6\,10^{7}$ S/m

ε_r? $\begin{cases} \text{ground} & 2 \leq \varepsilon_r \leq 10 \\ \text{water} & \varepsilon_r = 81 \end{cases}$

Example $f = 1\,MHz$

Cu $\delta = 64\,\mu m$, sea water $\delta = 25\,cm$,

ground ($\sigma = 10^{-2}\,S/m$) $\delta = 5\,m$

ground $\begin{cases} \sigma = 10^{-2}\,S/m \\ \varepsilon_r = 5 \end{cases}$ $\sigma = \omega\varepsilon$ if $f = 36\,MHz$

③ High frequency - low conducting material ($\omega\varepsilon \gg \sigma$)

$\gamma = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} \Rightarrow \gamma = \alpha + j\beta$

si $f \gg$: $\alpha \to \frac{1}{L_c}$ L_c: charact. length of the medium

(same influence as δ)

$L_c = \frac{\varepsilon}{\sigma}\sqrt{\frac{\varepsilon}{\mu}}$

Example $\sigma = 10^{-2} \Rightarrow L_c \sim 1\,m$

Boundary conditions on the interface

$\begin{cases} E_1 \\ B_1 \end{cases}$ ① | ② $\vec{n}$ $\begin{cases} E_2 \\ B_2 \end{cases}$

$\vec{n} \cdot (\vec{B}_2 - \vec{B}_1) = 0$; $\vec{n} \wedge (\vec{E}_2 - \vec{E}_1) = 0$

$\mu_1 \gamma_2^2 E_{n2} = \mu_2 \gamma_1^2 E_{n1}$

example air - ground $\frac{\uparrow E_z}{\uparrow E_z}$ air

$\left(\frac{E_{z\,sol}}{E_{z\,air}}\right) = \frac{\omega\varepsilon_0}{\sigma}(E_z)_{air}$ At L.F. $\sigma \gg \omega\varepsilon$

$(E_z)_{ground} \ll (E_z)$

Application to coaxial structures

① + ③ → good conductors σ

② → dielectric medium ε_0 (or $\varepsilon_0\varepsilon_r$)

$\vec{E}(r,\theta,z,t) = \vec{E}(r,\theta)\,e^{j\omega t - \gamma_p z}$

$\vec{H}(r,\theta,z,t) = \vec{H}(r,\theta)\,e^{j\omega t - \gamma_p z}$

γ_p: propagation constant of the wave along the z-axis

Wave equation $(\Delta - \Gamma_n^2)\vec{E}_n = 0$

where Γ_n^2: propagation constant in the medium "n"

$\Gamma_n^2 = j\omega\mu_n(\sigma_n + j\omega\varepsilon_n)$

$\Rightarrow \left[\Delta - (\Gamma_n^2 - \gamma_p^2)\right]\vec{E}_n(r,\theta) = 0$

general case E_r, E_θ, E_z, H_r, H_θ, H_z

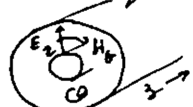
divided into 2 waves

TE → H_r, H_θ, E_θ TM → E_r, E_z, H_θ

Boundary condition on the interfaces

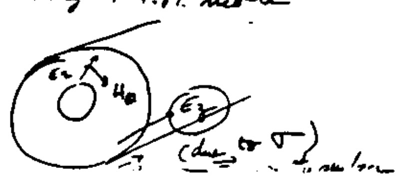
- ① + ③ ∞^{t} cond $\Rightarrow \Gamma_2^2 - \gamma_p^2 = 0 \Rightarrow \gamma_p = \Gamma_2$

$\begin{cases} \text{prop. constant} = \text{prop. constant of the dielect. mat. } \gamma \\ \text{TEM prop.} \end{cases}$



$\gamma_p = j\omega\sqrt{\varepsilon_2\mu_2}$

- ① + ② : finite conductivity → T.M. mode



Boundary condition E → Bessel, Hankel function

Characteristic impedance

$V = \int_{a_1}^{a_2} E_{r_2}\,dr$ $H_\theta = \frac{I}{2\pi r}$

(at low frequency)

$\Rightarrow \frac{V}{I} = \boxed{Z_c = \frac{j\gamma_p}{2\pi\omega\varepsilon_2}\log\frac{a_2}{a_1}}$

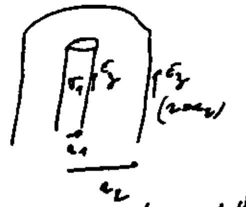
γ_p = complex number → Z_c complex + depends on f.

If conductivity → ∞ $\Rightarrow \gamma_p = \Gamma_2 = j\omega\sqrt{\varepsilon_2\mu_0}$

$Z_c = 60\log\frac{a_2}{a_1}$ (si $\varepsilon_2 = \varepsilon_0$)

Surface impedance

- for the internal conductor
- " " external conductor



Surface impedance of the internal conductor

$$Z_{si} = \left(\frac{E_z}{I}\right)_{r=a_1} = \frac{j}{2\pi a}\sqrt{\frac{j\omega\mu}{\sigma_1}}\;\frac{J_0(j\Gamma_1 a_1)}{J_1(j\Gamma_1 a_1)}$$

where $\Gamma_1 = \sqrt{j\omega\mu\sigma_1}$

At high freq.; $a_1 \gg \delta_1 \Rightarrow Z_{si} \neq \frac{1+j}{2\pi a_1 \sigma_1 \delta_1}$

Surface impedance of the external conductor

$$Z_{se} = \left(\frac{E_z}{I}\right)_{r=a_2} = \frac{-j}{2\pi a}\sqrt{\frac{j\omega\mu}{\sigma_3}}\;\frac{H_0^{(1)}(j\Gamma_3 a_2)}{H_1^{(1)}(j\Gamma_3 a_2)}$$

where $\Gamma_3 = \sqrt{j\omega\mu\sigma_3}$

(a) Impedance and admittance per unit length of a coaxial line

. $Z = jL\omega + Z_{si} + Z_{se}$

L: inductance $= \frac{\mu}{2\pi}\log a_2/a_1$

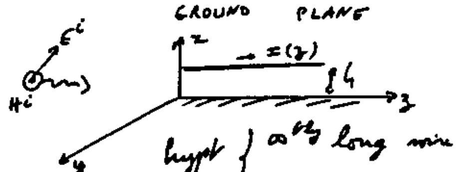
. $Y = j\omega C = j\omega \frac{2\pi\varepsilon_0}{\log a_2/a_1}$

Braided cable

Same general expression but Z_{se} has a more complicated form

$Z_{se} = f(a_2, a_3)$

CABLE PARALLEL TO A GROUND PLANE



hyp. $\begin{cases} \infty\text{ly long wire} \\ h < \lambda \Rightarrow \text{T.M. mode} \end{cases}$

$I = I_0 e^{j\omega t - \gamma z}$; $\gamma = \alpha + j\beta$

Antenna theory $\to \int eq \Rightarrow$ M.o.M.

$\searrow$ Transmission line theory

Short cable $\to$ MoM ; long cable $\to$ T.L.

$E^i, H^i \to$ "incident" field on the line, i.e., in the absence of the line but in the presence of the ground plane

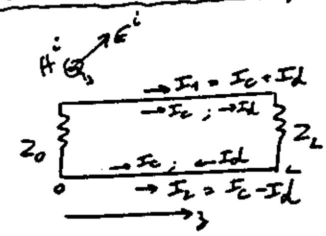
1st step $L \to \infty$

then influence of the boundary conditions

Image theory $\to$ two-wire line

("bifilar" line)

(c) Two-wire line illuminated by a plane wave



I_c vanishes at $z=0$ and $z=L$

I_d flows through the load impedances

$I_c \to$ dominant on the main part of the line (resonance ($\sim \lambda/2$))

$I_d \to$ due to E^i on wire 1 $\neq E^i$ on wire 2

Line above a ground plane

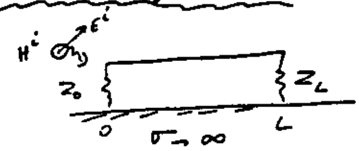
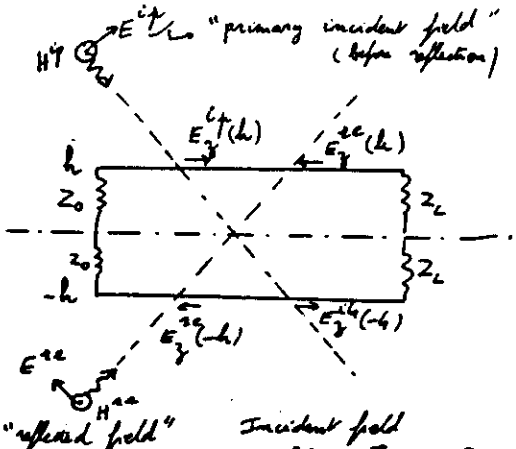


Image theory (for the line + for the source

(16)



"reflected field"   Incident field

$$\vec{E}^{i} = \vec{E}^{ip} + \vec{E}^{re}$$

tangential component on $\vec{E}^{i}$ on the wires (at the same abscissa) are the same (but opposite signs)

$\Downarrow$

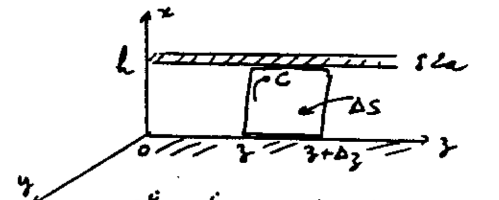
~~common mode~~   differential mode

T. L. theory well suited

Differential equations for V, I

{ Conductivity of the line: σ_f
{ $h \ll \lambda$ → T.M (quasi T.E.M. approx)

$I = I_0\, e^{j\omega t - \gamma_f z}$



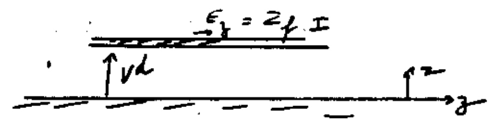
E^{i}, H^{i} → incident field ($\vec{E}^{ip} + \vec{E}$

⇒ Induced current on the wire ⇒ "scattered" (or "diffracted" or "radiated") field

Total field $\vec{E} = \vec{E}^{i} + \vec{E}^{d}$ and $\vec{H} = \vec{H}^{i} + \vec{H}^{d}$

$$\int_{C} \vec{E}\, d\vec{l} = -j\omega \iint_{\Delta S} \vec{B}\cdot d\vec{S}$$

To introduce the transverse voltage   To introduce the self inductance of the cable



V^d: transverse voltage (from the definition of the T.L. theory ⇒ due to the current I on the wire)

$$V^d = \int_h^0 E_x^d(x,0,z)\, dx$$

Stokes theorem

$$\frac{\partial V^d}{\partial z} - \int_0^h \frac{\partial E_x^i}{\partial z}\, dx + Z_f\, I(z) = j\omega \int_0^h B_y^i\, dx + j\omega \int_0^h B_y^d\, dx$$

Maxwell $\quad j\omega B_y^i = \dfrac{\partial E_z^i}{\partial x} - \dfrac{\partial E_x^i}{\partial z}$

$$\Rightarrow \frac{\partial V^d}{\partial z} + Z_f\, I(z) - \underbrace{j\omega \int_0^h B_y^d\, dx}_{jL\omega I} = E_z^i(h)$$

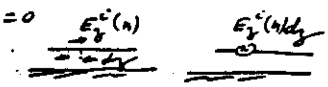
Series impedance $Z = Z_f + jL\omega$

$$\boxed{\frac{\partial V^d}{\partial z} + Z\, I(z) = E_z^i(h)}$$

2nd diff. eq → $\dfrac{\partial I}{\partial z} + Y V^d = 0$

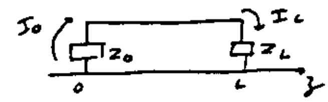
⇒ System of differential equation

$$\begin{cases} \dfrac{\partial V^d}{\partial z} + Z I = \underbrace{E_z^i(h)}_{\text{source}} \\[2mm] \dfrac{\partial I}{\partial z} + Y V^d = 0 \end{cases}$$



$ZY = \gamma_f^2$

BOUNDARY CONDITIONS



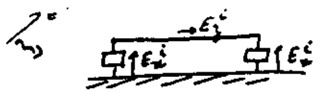
$$V^d(0) = -Z_0 I_0 = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=0}$$

$$V^d(L) = Z_L I_L = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=L}$$

Valid if E_z^i

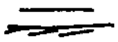
BUT if E^i has a vertical component → coupling to the "risers" (vertical parts) !!

(17)

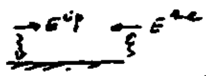


NOT possible to take rigorously E_x^i into account

T.L. →



BUT it is IMPORTANT to introduce E_x^i, even if $h \ll L$. Indeed



$E^i = E^{ip} + E^{re}$

nearly same ampl. but opposite signs!

⇒ vertical component can play a leading part (depending on the angle of incidence)

!

Since $h \ll \lambda$ → riser ~ antenna

e.m.f. $V^i(0) = \int_0^h E_x^i(x,0,0)\,dx$

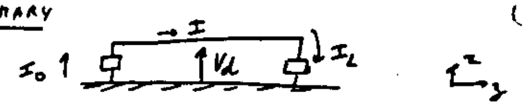
$V^i(L) = \int_0^h E_x^i(x,0,L)\,dx$

⇒ Appr. boundary conditions

$$-Z_0 I_0 + V^i(0) = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=0}$$

$$Z_L I_L + V^i(L) = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=L}$$

SUMMARY



$V^d = \int_h^0 E_x^d\,dx$ (due to I)

$$\begin{cases} \dfrac{\partial V^d}{\partial z} + ZI = E_z^i(h) \\ \dfrac{\partial I}{\partial z} + YV^d = 0 \end{cases}$$

Appr. boundary conditions

$$-Z_0 I_0 + \int_0^h E_x^i(0)\,dx = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=0}$$

$$Z_L I_L + \int_0^h E_x^i(L)\,dx = -\frac{1}{Y}\left(\frac{\partial I}{\partial z}\right)_{z=L}$$

If resonances → the amplitudes of the current will be appr.

Special case: infinitely long wire

→ Incident plane wave

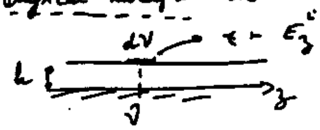
→ No boundary conditions

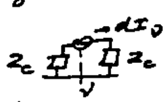
$E_z^i(h) = E_{z0}^i(h)\, e^{j\omega t - k_z z}$

prop. constant of the incident wave along the z axis

$$\Rightarrow I(z) = \int_{-\infty}^{+\infty} \frac{E_{z0}^i(h)\, e^{-k_z v}}{2Z_c}\, e^{-\gamma_p|z-v|}\, dv$$

Physical interpretation



infinite line ⇒ 

$dI(v) = E_z^i(h)\,dv/2Z_c$

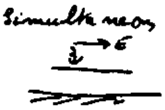
at a point z → $dI_v\, e^{-\gamma_p|z-v|}$

$\int_{-\infty}^{+\infty}$ ⇒ Σ of all the eq. generators

Plane wave: $E_{z0}^i(h) = $ cte

$$I(z) = \frac{E_{z0}^i(h)\, e^{-k_z z}}{Z_c}\,\frac{\gamma_p}{\gamma_p^2 - k_z^2}$$

Simultaneous illumination

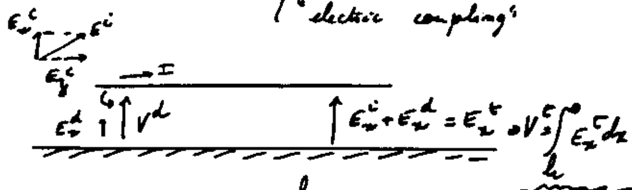


⇒ $k_z = 0$

$$\boxed{I(z) = \frac{E_{z0}^i(h)}{\gamma_p Z_c} = \frac{E_{z0}^i(h)}{Z}}$$

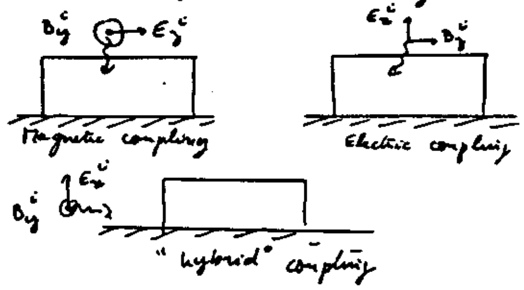
→ order of magnitude for finite lengths

Other possible approach: { "magnetic coupling" / "electric coupling"



$$\begin{cases} \dfrac{\partial I}{\partial z} + YV^t = -Y\displaystyle\int_0^h E_x^i\,dx & \rightarrow \text{current source} \\ \dfrac{\partial V^t}{\partial z} + ZI = j\omega \displaystyle\int_0^h B_y^i\,dx & \rightarrow \text{voltage source} \end{cases}$$

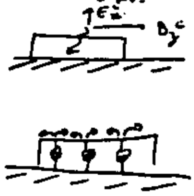
Source, $E_x^i + B_y^i$ instead of E_z^i !!



Instead of 1 coupling to the vertical part or (and) to the horizontal part of the line

Different results depending on the approach ??

Electric coupling



$$\frac{\partial I}{\partial z} + YV^t = -Y\int_0^h E_x^i\,dx$$

→ Boundary conditions + solution

only $E_z^i(0)$ and $E_z^i(L)$

coupling to the vertical part of the line!

⇒ Whatever the configuration → same results (Maxwell eq! Relation between B_y and E_x, E_z!)

Interest of this approach → B_y^i is easily measurable at low frequency. It is not the case for E_x^i

but the previous approach can be easily extended to other shapes (antenna theory: source term is the tg. electric field on the structure)

Simple analytical formulas (short cables $L \ll \lambda$)

Example: simultaneous illumination.

	Coupling to the horizontal part	Coupling to the risers
Sym. con. (2 pts)	$\dfrac{E_z^i(h)}{2}$ ①	$-j\dfrac{E_x^i h}{Z_c}\dfrac{2\pi}{\lambda}\left(\dfrac{L}{2}-z\right)$ ②
Asym. 1 pt	$\dfrac{E_z^i(h)}{2}\left(\dfrac{2\pi}{\lambda}\right)^2\left(\dfrac{z}{2}-L\right)z$	$-j\dfrac{E_x^i h}{Z_c}\dfrac{2\pi}{\lambda}z$ ③

① eq. gen. $E_z^i(h)\times L$ → independent on L; load → ZL

② 2 eq. gene. on both side of the line → I = 0 at the center

③ 1 eq. voltage generator connected to an open end transmission line