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SECOND COLLEGE ON THEORETICAL AND EXPERIMENTAL RADIOPROPAGATION PHYSICS
(7 January - 1 February 1991)

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Electromagnetic Compatibility

and

The Electromagnetic Environment

[illegible]
[illegible] of Lille
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ELECTROMAGNETIC COMPATIBILITY

AND

THE ELECTROMAGNETIC ENVIRONMENT

by

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Lecture notes for the "Second College on Theoretical and Experimental Radiopropagation Physics"

**International Centre for Theoretical Physics
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UNESCO**

Part of the content of these lectures notes comes from the following books or journals.

1) B. KEISER, "Principles of Electromagnetic Compatibility", Artech House Ed., 1983.

2) P. DEGAUQUE et J. HAMELIN, "Compatibilité Electromagnétique", Dunod Ed., Paris, 1990.

3) M. IANOVICI et J.J. MORF, "Compatibilité Electromagnétique", Presses Polytechniques Romandes", Lausanne 1983.

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5) E.F. VANCE, "Coupling to shielded cables", J. Wiley Ed., 1978.

6) K.S.H. LEE, "EMP Interaction, Principles, Techniques and Reference Data", Hemisphere Pub., New York, 1986.

7) P.E. LAW, "Shipboard Electromagnetics", Artech House Ed., Boston, 1987.

8) I.E.E.E. Transactions on Electromagnetic Compatibility, Journal edited by I.E.E.E.

9) B. DEMOULIN, P. DEGAUQUE and M. HEDDEBAUT, "Approaches to electromagnetic compatibility in transport systems", Recherche - Transport - Sécurité Rev., n° 4, pp. 1-14, 1990.

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INTRODUCTION

As defined in reference |1|, electrical and electronic devices are said to be electromagnetically compatible when the electrical noise generated by each does not interfere with the normal performance of any of the others. Electromagnetic compatibility (EMC) is that happy situation in which systems works as intended, both within themselves and in their environment.

Electromagnetic interference (EMI) is said to exist when undesirable voltages or currents are present to influence adversely the performance of a device. Furthermore, these voltages or currents may reach the victim device either by conduction or by electromagnetic field radiation.

The cause of an EMI problem may be either within the system one is dealing with, in which case the problem is labelled an "intrasystem" problem, or the EMI may come from the outside, in which case the problem is given the "intersystem" designation.

The objective of this lecture is to make a comprehensive approach of the methodology which has to be used in order to prevent or at least to interpret bad performances of systems due to EMC problems in order to find adequate solutions. EMC is a very wide subject and this lecture must be thus considered as a general presentation rather than a succession of "tricks of the trade". First, the possible sources of interferences will be presented by considering the natural noise and the man made noise successively. Then, is a second part the coupling mechanism will be described and some practical aspects, mainly on the measurement techniques will be detailed in the third part. Lastly informations on biological effects of electromagnetics waves are given.

A) ELECTROMAGNETIC ENVIRONMENT CHARACTERISTICS

I) INTRODUCTION

The source of interferences are usually divided into two broad categories : the natural noise and the man-made noise. As it is obvious from the designation of these two types of noise, the so-called natural noise is due to natural phenomena such as the lightning discharge, the global atmospheric noise and also the electrostatic discharge. The man-made noise comes from the human activity and to give some examples correspond to broadcast emissions, industrial, scientific and medical equipment (ISM), automotive sources (ignition systen, voltage regulator)... The characteristics of such electromagnetic sources will be described successively.

II) NATURAL NOISE

II.1) Cosmic noise [1]

Cosmic noise, or noise from outer space, is of three types : galaxy noise, thermal noise and anomalous star noise. Galaxy noise peaks in the 150-200 MHz band. Thermal noise is radiated from celestial bodies at a frequency range of 3 to 30 GHz. Anomalies solar noise is associated with sunspots and other unexplained solar phenomenas. The 20 to 35 MHz band is the range where such interference is most often found, although the frequencies vary with the type of anomaly. As it will be outlined in the next paragraphs, the level of the cosmic noises below 10 MHz are lower than those of atmospheric and man-made interference. Therefore cosmic noises are critical primarily above 50 MHz at locations remote from man-made noise environments.

II.2) Atmospheric noise [2]

The mean reason of the atmospheric noise is the global lightning activity. On a given site, a keraunic level is first defined. It corresponds to the number of days per year during which the thunder has been heard. This "thunder" criteria has been adopted since it allows to limit the geographical area around the point of observation since the sound has a range of about 10 km. One of the main criticism on this criteria is the fact that it is not a quantitative one since only one lightning stroke has the same "weight" as a succession of lightning discharges a whole day. However this keraunic level has been kept and empirical formulas different from one region to another ore have been proposed to deduce the density of ground strike point per km^2. Usually this correllation between these two characteristics has been obtained with the help of CIGRE antennas. These are vertical antennas, about 1 m or 2 m high tuned at 500 Hz or 10 kHz with an adjusted sensibility such that only a lightning discharge occuring at a distance smaller than 10 to 20 km can be detected.

It must be outlined that this situation is changing rapidly. Indeed, optical data from satellites allows to determine and to locate the number of lightning discharges at a world-wide scale, including the oceans. Observations made from satellite DMSP are shown in Figure 1.

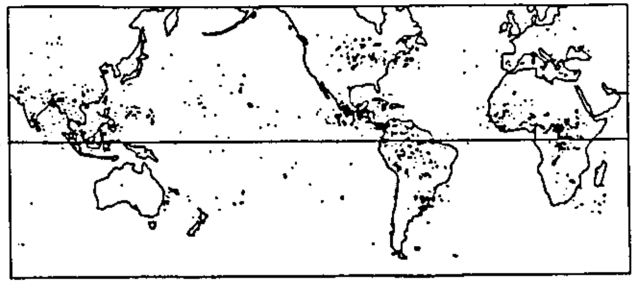

Figure 1

Furthermore from the knowledge of the electromagnetic signature of the discharge, goniometric systems are now working in the U.S., France, Japon, Sweden... It becomes possible to locate in a real time all the strike points in a given region. As an example if the distance between the stations is 300 km, the precision in the location of the discharge is about 3 km.

Another important point to define the electromagnetic aggression is to know if most of the lightning discharges are cloud-to-cloud or ground-to-cloud flashes. A statistical analysis shows that the ratio between the number of intracloud discharge N_{in} to the number of cloud to ground discharge depends on the latitude α as it appears from Figure 2. Indeed in equatorial regions most of the events are intra-cloud discharges, the ratio previously defined being equal to 6.

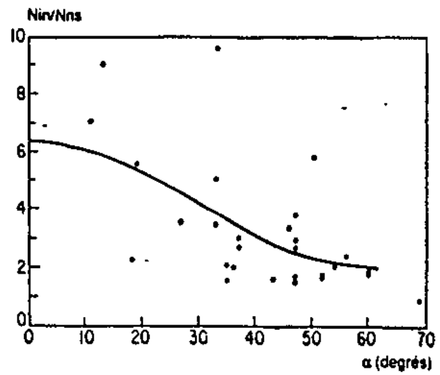

Figure 2

Many lightning discharges occur simultaneously at various points of the world. The radiated electromagnetic field can propagate at very large distances, at least in the low frequency range, in the earth-ionosphere waveguide. It results an average continuous noise called atmospheric noise or whistlers. To have a more quantitative approach, an equivalent noise factor F_a expressed in dB is introduced

$$F_a = 10 \log (f_a) \quad \text{where } f_a = P_n / kT_o\, b \qquad (1)$$

P_n is the average power, expressed in Watt, received by an omnidirectional antenna, k is the Boltzmann constant, T_o is a temperature of reference (i.e. $T_o = 288°K$), b is the receiver bandwith (Hz). From F_a, it is possible to deduce the amplitude of the vertical electric field component E_z :

$$E_z = F_a + 20 \log f + 10 \log b - 95.5 \qquad (2)$$

when f is expressed in MHz.

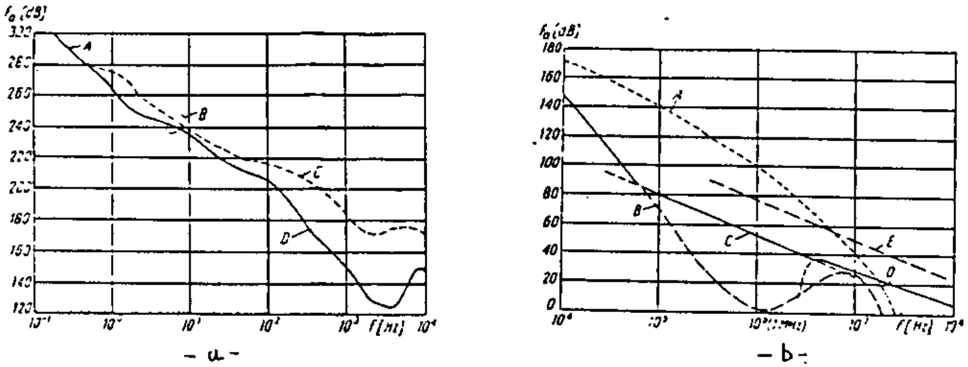

Figure 3

Figures 3a and 3b present the contribution of the various noise sources to the equivalent noise factor F_a. In Figure 3a, the curves A,B,C and D, plotted for frequencies ranging from 0.1 Hz to 10 kHz, correspond to micropulses, atmospherics, maximum and minimum values respectively. In the range 10 kHz - 10^8 Hz, the curves A,B,C, D and E in Figure 3b are : A- the atmospheric noise (value exceeded during 0.5 % of the time, B- atmospheric noise which is exceeded during 99.5 % of the time, C- noise due to the human activity in a residential area, D- galactic noise, E- noise in an industrial area. Lastly, it must be noted that the impulsive noise due to a lightning discharge at few hundred kilometers from the observation point has a typical waveshape because the earth - ionosphere waveguide behaves as a band pass filter for few frequencies. At great distances, all the frequencies greater than 300 kHz are strongly attenuated.

II.3) Characteristics of a lightning discharge

II.3.1) Physical mechanism

The conductivity of the low atmosphere is due to the β and γ particles generated by the cosmic rays, the UV and X solar rays, the natural radioactivity of the ground.. In "fine weather" conditions, this conductivity at sea level is $\sigma = 1.33\ 10^{-14}$ S/m. Since the lowest layers are electrically charged, there is a continuous current from the ground to the ionosphere with a current density J_z of about 2.10^{-12} A/m^2. By integrating on the earth surface, this leads to a total current of 1 kA. If this phenomena is the only one, the ionosphere would be rapidly discharged. Thus there is another phenomena which is the lightning activity near the earth surface such that one can define a global equivalent electric circuit as shown in Figure 4.

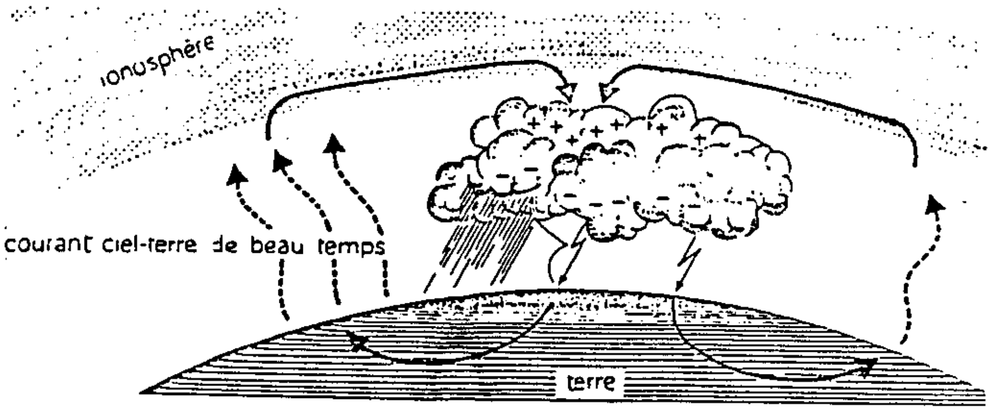

Figure 4

The lightning discharge is due to the mechanism of separation of charges inside the clouds. This electrisation process is not yet well understood but comes from the convection mechanism occuring inside the clouds. In most cases, as shown in Figure 5a, the bottom of the cloud is usually charged negatively while the upper part of the cloud is positively charged. If the electric field is sufficiently high, an ionized channel appears at the bottom of the cloud and the so-called "stepped leader" drains the negative charges and moves towards the earth surface with a stepped process and following the electric field line existing at each step. Each step has a length of about 50 m, a duration of 1 µs a direction between two steps of 50 µs and moves at a velocity of 1.5 10^5 m/s. Then, as it appears from Figure 5b and when the stepped leader is at a height of 100 m from the ground surface, the electric field intensity on the ground reaches 500 kV/m. Usually it is enough to produce an upward leader, starting from the ground and moving towards the downwards leader. Ater the junction, a continuous ionized channel exists between the cloud and the ground and a "return stroke" occurs. It is characterized by a sudden increase of the current amplitude and of

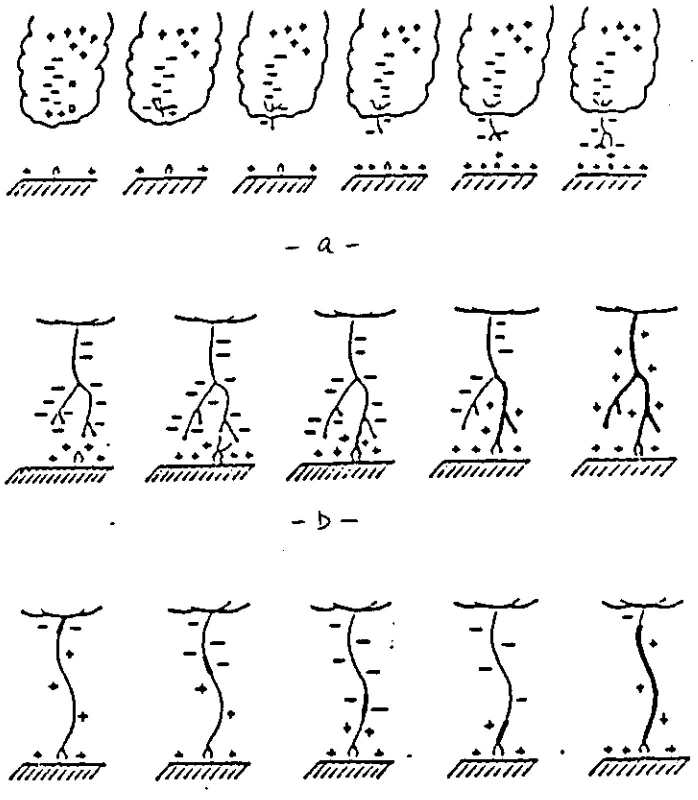

Figure 5

the brightness of the channel. After this discharge, a subsequent leader can occur but following the existing one still an ionized channel remains. As shown in Figure 5c, few subsequent return strokes usually occur in a lightning discharge. The duration of the various events is given in Figure 6.

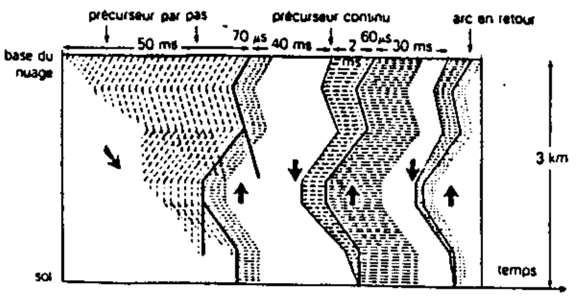

Figure 6

An intracloud discharge may also appear between two volumes of opposite charges, either in the cloud itself or between two clouds.

II.3.2) Return stroke current

From an electromagnetic compatibility point of view, the most critical part corresponds of course to the return stroke phenomena. Although many experimental works are now undertaken to characterize the intra-cloud discharges, most of the quantitative data are related to the cloud to ground discharge. Curve in Figure 7 gives the average value of the return stroke current versus time.

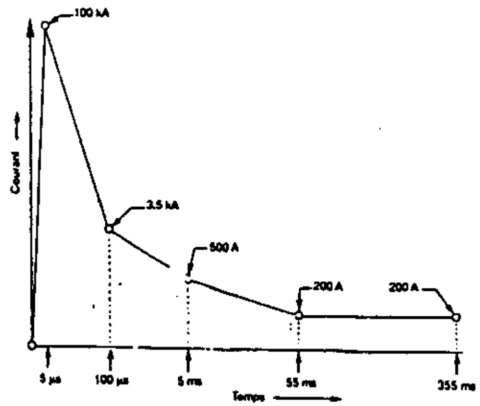

Figure 7

Many observations have been made either on triggered lightning or on instrumental towers. Typical shapes of measured current and current derivative are given in Figure 8a and 8b respectively. It must be

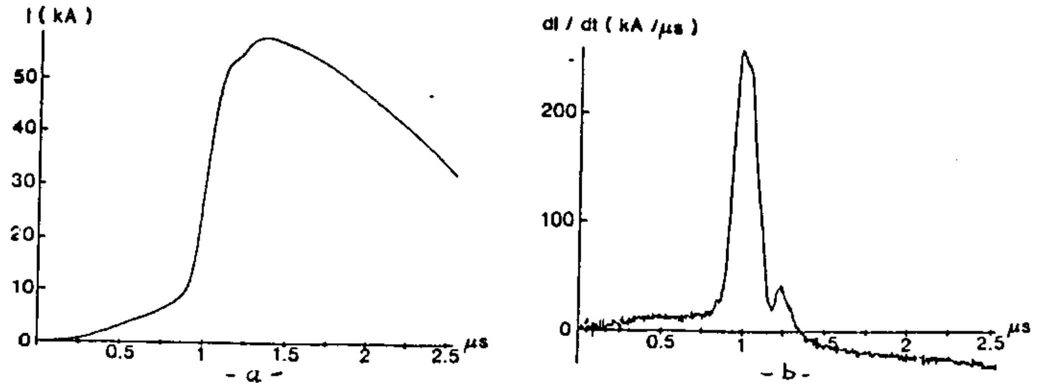

Figure 8

outlined that the time current derivative is also an important parameter since in the case of a direct stroke to electric cables, the protecting devices must work under the two extreme values prescribed by I_{max} and $(dI/dt)_{max}$. Figure 9 shows the possible correlation between the peak values of I and dI/dt. All the points, crosses... correspond to different measurements made on triggered lightning either in France or in the U.S. We note that peak values of dI/dt can reach values as high as 450 kA/µs.

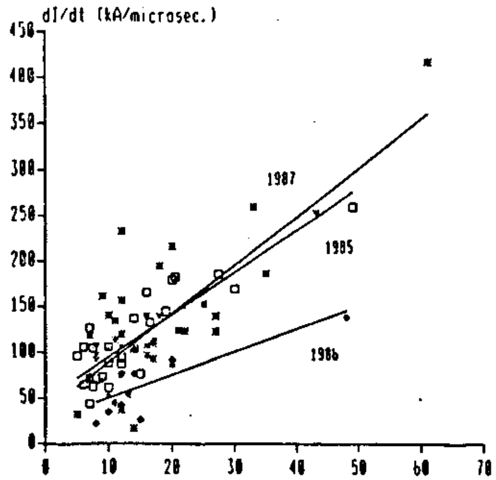

Figure 9

II.3.3) E.M. field radiated by a lightning stroke

During the initial phases i.e. when predischarges occur, radiofrequency signals are emitted, almost in the 60 MHz - 175 MHz band, and correspond to short pulses, each burst having a duration of 10 ms to 100 ms. The step-leader radiates in a wide frequency range but almost in the 60 MHz - 500 MHz range. This VHF-UHF radiation can be used in interferometric systems to localize, with a very good precision, the discharge either in the cloud or between the cloud and the ground.

The most disturbing source is of course the one due to the return stroke process. Indeed at distances as great as 200 km, the peak amplitude of the vertical electric field reaches 1 V/m while at distances smaller than 100 m, these peak values exceed 10 kV/m. As an example if we consider a region where the keraunic level is 30 days per year, thus 2 ground strike points per km^2 and per year, it has been determined that any point of this area is illuminated by few tens thousands impulses per year, with a peak amplitude greater than 20 V/m.

The electromagnetic field is due to the propagation of the lightning current along the nearby vertical channel, few kilometers long. This equivalent long antenna can be considered as a continuous series of elementary electric dipoles. Therefore, the expression of the radiated field can be put in the form of the sum of three terms. The

first one decrases as $1/D^3$ (where D is the distance between the lightning channel and the observation point) and is the "quasi-static component", the second term decreases as $1/D^2$ (induced component) and the third one is the "radiated component", behaving as 1/D. Usually all the measurements and thus all the characteristics of the electric field are normalized to 10 km or to 100 km, assuming an 1/D variation. As an example, the well-known curve of Pierce presented in Figure 10 gives the spectrum of the peak amplitude of the vertical radiated field normalized to 10 km and for a bandwidth of 1 kHz.

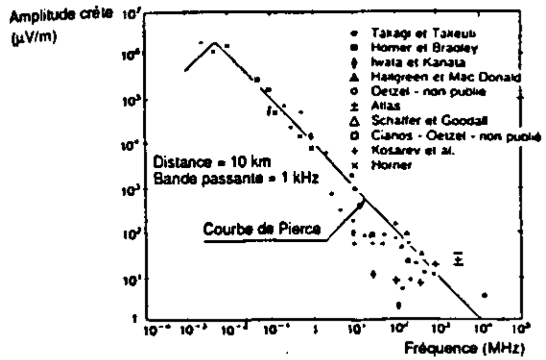

Figure 10

The magnetic field component can also be divided into two terms, the first one varying as $1/D^2$ and the second one as 1/D.

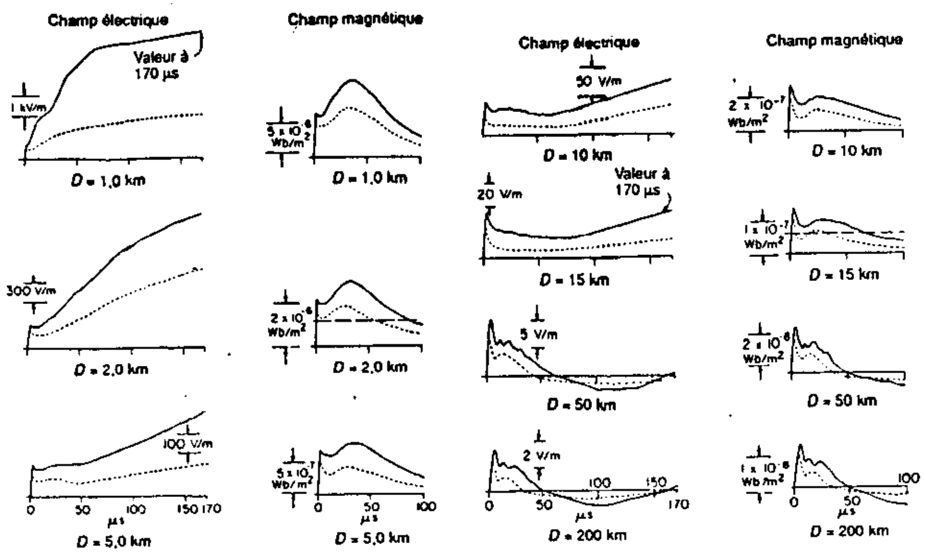

Figure 11

Figure 11 shows typical shapes of the electric and magnetic field both for the first return stroke (continuous curves) or for the subsequent return stroke (dotted lines) and for two distances : 1 km and 10 km.

This electromagnetic field is distorded when it propagates along the ground surface and as it can be seen from the measurements or from the experimental approach, the rise time is a rapidly decreasing function of the distance.

II.3.4) Numerical modelling

Numerous models have been proposed to calculate the electromagnetic field radiated by the return stroke. One of the oldiest and a quite popular one is called the "transmission line model". As indicated in Figure 12, it is assumed that the return stroke current propagates along a vertical channel with a constant velocity v. If the ground (medium 1) is assumed to be a perfect conductor, the vertical

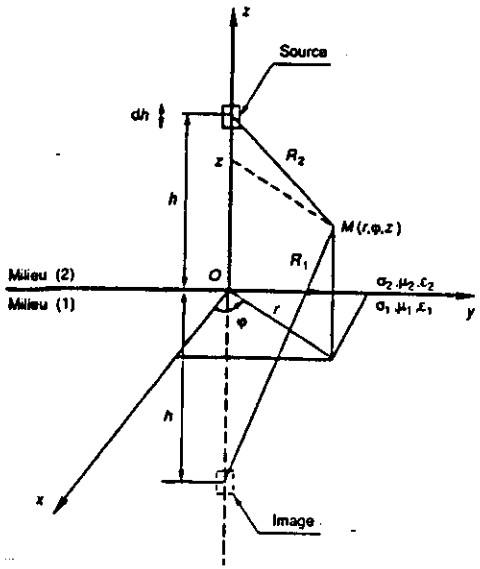

Figure 12

electric field E_z and the azimuthal magnetic field H_y radiated by each elementary part dh of the channel are given by

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

$$H_\varphi(t) = \frac{dh}{2\pi}\left[\frac{r}{R^3}\, i\left(t-\frac{R}{c}\right) + \frac{r}{cR^2}\frac{di}{dt}\left(t-\frac{R}{c}\right)\right] \quad (4)$$

If the observation point is situated at a horizontal distance r much greater than the height H of the channel and if the velocity v of the return stroke current is a constant, one can be easily shown that expressions (3) and (4) can be put on the form

$$E_z(H,t) = \frac{v}{2\pi\varepsilon_0 c^2 r}\left[i(t - r/c - H/v) - i(t - r/c)\right] \quad (5)$$

$$H_\varphi(H,t) = \frac{1}{2\pi c r}\int_0^H \frac{di}{dt}(t - r/c - h/v)\,dh \quad (6)$$

These simple formulas give at least an order of magnitude of the disturbing field. The horizontal component E_r may also play a leading part to the coupling to long horizontal cables. It must be calculated either from an exact numerical approach taking the ground conductivity into account, or, at great distance, from the Zenneck formula, relating the horizontal to the vertical electric field components

$$\frac{E_z}{E_r} = \sqrt{\varepsilon_r + \frac{\sigma}{j\omega\varepsilon_0}} \quad (7)$$

where ε_r and σ are the electrical characteristics of the ground.

II.4) Electrostatic discharge

The electrostatic discharge is a frequently entcountered problem and the well-known geometrical configuration is given in Figure 13, a discharge occuring when the voltage difference between the human body and the apparatus becomes greater than few kV.

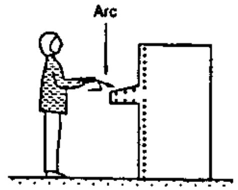

Figure 13

This discharge produces a current which may reach a peak value of few Amps and with a rise time of few nanoseconds. This current is of course a disturbing current injected in the electronic equipment but also radiates a high frequency electromagnetic field in its nearby environment. A numerical model describing the phenomena is rather difficult and thus an experimental simulation is rather used to test the susceptibility of an equipment. As shown in Figure 14 a capacitance C_0

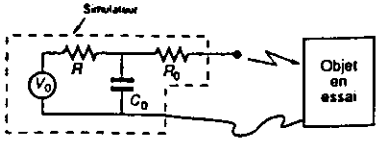

Figure 14

is charged at a voltage V_0 and thus discharged through a resistance R_0. The two parameters R_0 and C_0 must simulated the equivalent electric parameters of a human body for example. The usual values are

$R_0 = 150\ \Omega$, $C_0 = 150$ pF and $V_{omax} = 20$ kV

III) MAN-MADE NOISE

III.1) 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 course.

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

We have seen that the spectrum radiated by a lightning strike covers practically all of the frequencies from continuous, direct current up to several tens of megahertz. The maximum electric field recorded at one kilometer from the point where the lightning strikes may be as much as 100 V/m. On an entirely different order of magnitude, the electric field radiated by a 100 W local FM broadcasting station is the order of one millivolt per meter a few hundred meters from the antenna. On the other hand, the electric field radiated by a one-watt walkie-talkie at 150 MHz is a few volts per meter when measured one meter away. The electrical field radiated by powerful radars emitting at frequencies of a few gigahertz is still a few volts per meter several tens of kilometers away from the transmitter facility (ref. 26).

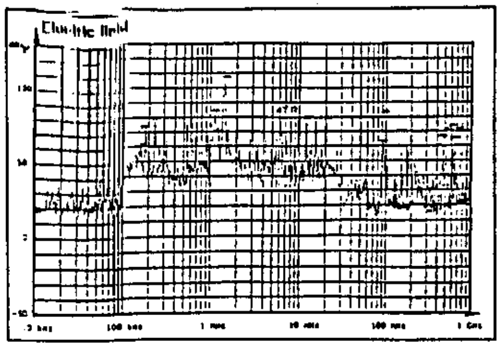

Figure 15 Electromagnetic spectrum recorded seven meters from the metro tracks, with no train passing.

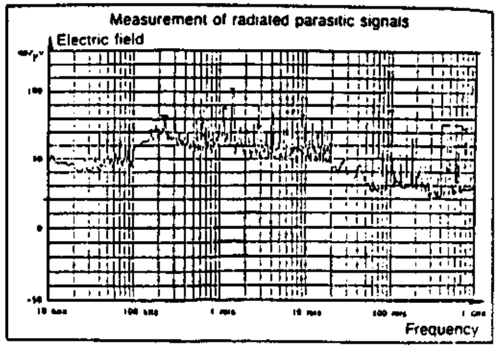

Figure 16 Electromagnetic spectrum recorded seven meters from the tracks with a train passing by.

As an example, Figures 15 and 16 show the electromagnetic spectrum of a disturbance, as read near a location where metrotrains were running on 750 V direct current (ref. 19). The spectrum corresponds to the variation of the vertical component of the electrical field, between 10 kHz and 1,000 kHz. The electric antenna was set up seven meters from the rails supplying the trains. Figure 15 shows the spectrum when there were no trains present and Figure 16 when a train is passing near the antenna. It can be seen that the perceptible field between 10 kHz and 100 kHz increases by 30 dB when the train passes by. These are the RF components radiated by the highorder harmonics of the currents absorbed by the choppers in the drive equipment. Tables 1 to 3|1| give few examples of radiation devices and sources of radiated interference.

III.2) 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 mechanism will be dominant. Example of possible sources are given in Table 4|1|.

III.3) Narrowband and Wideband interferences

The terms "narrowband" and "wideband" are generally used. It must be emphasized that the adjective "narrow" or "wide" cannot be a characteristic of the source alone but mainly depends on the bandwidth of the sensitive receiving equipment. Indeed, a signal or noise occupying a narrow bandwidth relative to that of a given measuring instrument will produce an output that rises sharply as the instrument is tuned across the band occupied|1|. It is thus possible to identify the disturbance by changing either the bandwidth B of the receiver or, if it is not possible, by changing the tuned frequency|2|.

Table 1 Restricted Radiation Devices

Intentional Radiators
Campus radio stations
Wireless announcing 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 camera to monitor circuits
TV camera 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 - 3.6 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		
Untreated access covers	10 kHz - 10 MHz	1000 µV/m/kHz
Fluorescent lamp		
Arc	100 kHz - 3.0 MHz	
Multiplexer		
Solid state switching	300 kHz - 500 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	

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

In the first case, let us assume that the receiver is tuned at a frequency F_o and its bandwidth is changed from B_o to B_1. If the signal level, expressed in dB has a variation greater than 10 log (B_1/B_o), the disturbance is "wideband". If the variation of the signal level is smaller than 3 dB, this corresponds to a "narrowband" disturbance. This method also allows to determine the coherence of the wideband interference.

In the second case, it is not possible to change the bandwidth of the receiver. Thus in a first step, if the filter is tuned to a frequency F_o, a level U_o (dB/µV) is obtained while, by changing F_o to $F_1 = F_o + B$, a level U_1 is measured.

If $DU = |U_1 - U_o| \leq 3$ dB ==> wideband interference

If $DU = |U_1 - U_o| > 3$ db ==> narrowband interference.

Tables 5 and 6 recall the usual units characterizing the signals.

Table 5 Radiated disturbing sources.

	Electric field E	Magnetic field H	Power surface density dP
Narrowband	V/m	A/m	W/m^2
Wideband	V/(m.Hz) or V/(m.$\sqrt{Hz}$)	A/(m.Hz) A/(m.$\sqrt{Hz}$)	W/(m^2.Hz^2) or W/(m^2.Hz)

Table 6 Conducted disturbing interferences.

	Voltage V	Current I	Power P
Narrowband	V	A	W
Wideband	V/Hz or V/$\sqrt{Hz}$	A/Hz A/$\sqrt{Hz}$	W/Hz^2 or W/Hz

III.4) <u>Nuclear E.M. pulse</u>

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

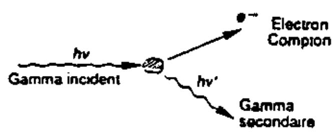

Figure 17

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...

For a high altitude explosion, the terrestrial magnetic field plays a leading part in the distortion of the current lines. (Figure 18)

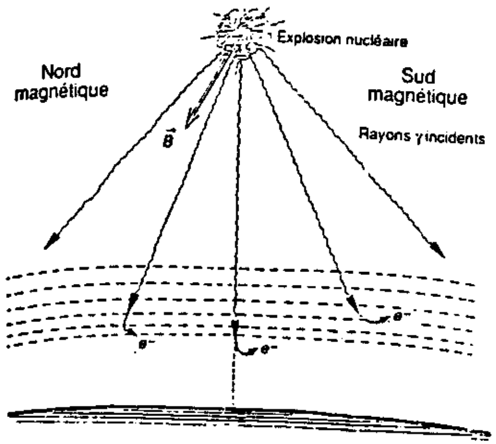

Figure 18

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

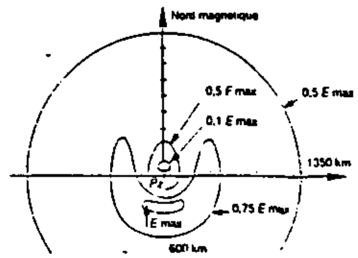

Figure 19

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

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

B) SUSCEPTIBILITY AND COMPATIBILITY |9|

I) INTRODUCTION

Having established that EMI sources are characterized by their frequency spectrum or pulsed character, we will now address the question of how they couple with the systems.

The idea of a "system" is taken in a very general sense, considering that the configurations and arrangements encountered in practice are in the process of diversifying without limit. It is this very large complexity that makes it difficult to predict the voltage levels that might reach a given point in an electronic circuit.

So the goal is not to look for a precise method of calculating these levels, nor to find a universal formula for predicting EMI effects, but simply to understand the physical phenomena governing the coupling of an EMI with a system, which we will attempt to simplify.

The most elementary system will include a wire pair line connecting two electronic circuits. This line is not, in any case, the only element common to the two circuits : often the ground circuits are also in common, and are intentionally laid out at the time the system is designed, or perhaps are introduced accidentally by the environment external to the system. The spurious voltages that may appear in the equipement are strongly influenced by the impedance networks that load the end of the two-wire line. We will admit that these impedances are simple in structure and that they are equivalent to an impedance Z_d connected to the ends of the two conductors, and impedances Z_{c1} and Z_{c2} connected between conductor 1 and conductor 2 and the ground reference of each piece of equipment. The ground circuit in the diagram of Figure 20 symbolizes the link between the reference of the equipment installed at z = 0 and the reference of the equipment installed at z = L

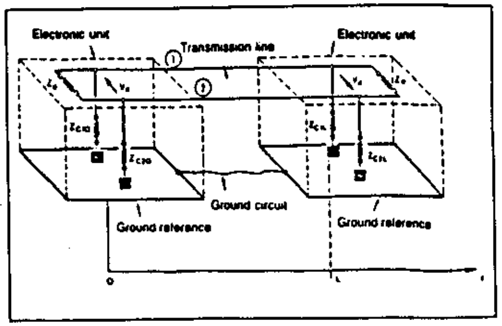

Figure 20 Example of a schematic system reduced to a two-wire transmission line connecting two pieces of electronic equipment. The potential references of each are interconnected through the ground network.

There are three possible modes of electromagnetic coupling to the system, i.e. by :

- influence or mutual coupling
- conduction
- radiation.

In reality, coupling by mutual coupling and by radiation have the same physical origin since they are the effects of the electric and magnetic field components on the system. The distinction, as we will hereafter show, has to do only with the relative arrangement of the sources and disturbed systems, and the mathematical formulas used to predict the effects of these couplings.

We can say that influence coupling occurs when lines designed to carry heavy currents follow a path parallel to lines transmitting low-level messages. The best known example is railroads equipped for electric drive. An influence exists between the catenary lines, the rails and the signaling lines running along in a trench parallel to the tracks.

Conductive coupling involves an electrical contact between the source of disturbance and the sensitive circuits. If we exclude those disturbances transmitted through the circuit power supplies, the most common example in a train is the flow of residual RF currents emitted by the ground line on a self-propelled train (Figure 21). Two electronic

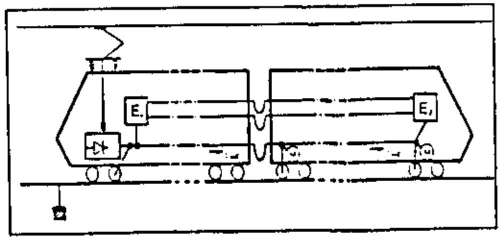

Figure 21 Conductive coupling through the ground wiring in a self-propelled train.

units E_1 and E_2 located at the two ends of the train, interconnected by a transmission line and in contact with the ground circuit, will find parasitic voltages due to the differences of potential produced by the flow of RF components in the ground circuit. This is because, for the fundamental frequency of the high currents, the impedance of the ground circuit is small and consequently has no effect on the circuits ; on the other hand, for the higher-order harmonics, this impedance can increase and create high voltages.

Radiative coupling quite often comes from a point source very far from the equipment. The equipment wiring behaves like a receiving antenna with respect to this source, and can generate spurious voltages that may modify the operation of the circuits. The most typical example is the effect of a radio transmitter on the electronic equipment in an automobile.

The phenomenological analysis is extremely complex here, since it depends on the effect of the shielding offered by the metal bodywork, and on the presence of other elements that make it easier for the field to penetrate (such as an antenna, openings in the bodywork, the use of composite materials, etc.). Any attempt to model these phenomena exactly is of course difficult, so we will attempt to describe hereafter the general principles and the simplest equations that can offer a few solutions to these problems of electromagnetic compatibility.

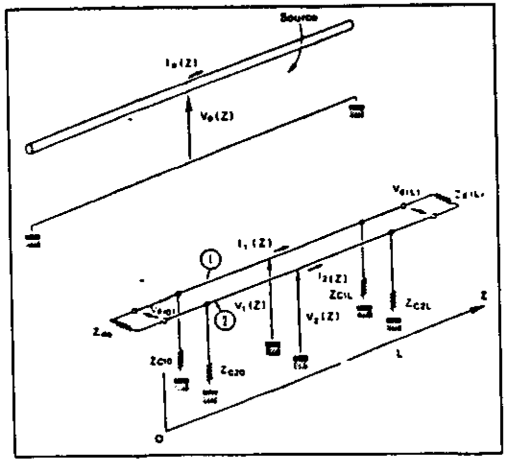

Figure 22

II) MUTUAL COUPLING BETWEEN CABLES

Let us consider a two-wire line parallel to another line carrying the electromagnetic disturbance characterized by the current I_p (Z, ω). This will induce currents I_1 (Z, ω) and I_2 (Z, ω) in the pair, along with the voltages V_1 (Z, ω) and V_2 (Z, ω). These spurious voltages are the differences of potential induced between each conductor and the ground reference, common to the source of disturbance and to the system. This ground reference is assumed to equipotential (figure 22)

This is a system of coupled lines for which we can write a set of differential equations relating the current I_1 and I_2 and the voltages V_1 and V_2 to the source of disturbance:

$$-\frac{dV_1}{dz} = Z_{11}I_1 + Z_{12}I_2 + Z_{p1}I_p$$

$$-\frac{dV_2}{dz} = Z_{12}I_1 + Z_{22}I_2 + Z_{p2}I_p \qquad (1)$$

$$-\frac{dI_1}{dz} = Y_{11}V_1 + Y_{12}(V_2 - V_1) + Y_{p1}(V_p - V_1)$$

$$-\frac{dI_2}{dz} = Y_{12}(V_1 - V_2) + Y_{22}V_2 + Y_{p2}(V_p - V_2) \qquad (2)$$

Z_{11} and Z_{22} represent here the impedances per unit length of conductors 1 and 2 with respect to the reference plane, and Y_{11} and Y_{22} the corresponding admittances per unit length. Z_{12} and Y_{12} are the impedances and admittances per unit length of electromagnetic coupling between conductors 1 and 2. Z_{p1} and Z_{p2} as well as Y_{p1} and Y_{p2} express the effect of the electromagnetic disturbance on the two-wire system, and Z_p the magnetic influence between the disturbing line and conductor 1 or conductor 2. An inductance per unit length of coupling Lp can be associated with Z_p such that.

$$Z_p = j\,L_p\omega \text{ in which } j = \sqrt{-1} \qquad (3)$$

Y_p expresses the electrical influence that appears between the source of disturbance and each wire element of the system. A capacitance per unit length of coupling C_p can also be associated with Y_p such that.

$$Y_p = j\,C_p\omega \text{ in which } j = \sqrt{-1} \qquad (4)$$

Though the voltage and current with respect to the reference match nicely in equations (1) and (2), the equations themselves are of little practical interest. Often, what is preferred is the common-mode voltage and current I_c (Z, ω) and V_c (Z, ω) and the differential-mode current and voltage I_d (Z, ω) and V_d (Z, ω), expressed by:

$$V_c(z,\omega) = \frac{V_1(z,\omega) + V_2(z,\omega)}{2} \qquad (5)$$

$$I_c(z,\omega) = \frac{I_1(z,\omega) + I_2(z,\omega)}{2} \qquad (5)$$

$$V_d(z,\omega) = |V_1(z,\omega) - V_2(z,\omega)| \qquad (6)$$

$$I_d(z,\omega) = |I_1(z,\omega) - I_2(z,\omega)|$$

The differential mode voltage is superimposed on the data transmitted through the pair and is therefore of great interest if we

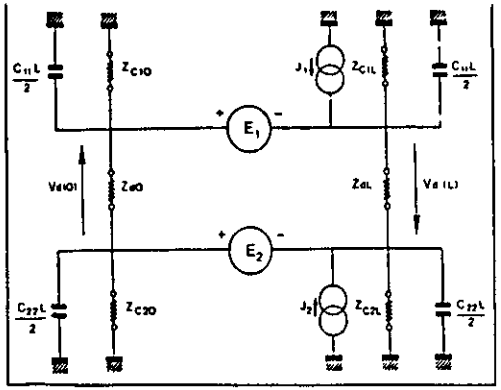

Figure 23 Equivalent diagram of the disturber-disturbed line system in the long wavelengths (λ ≫ L). The impedances Z_{c10}, Z_{c20}, Z_{c1L} and Z_{c2L} are large and the voltage and current generators represent the action of the disturber.

want to assess the system EMC. In reality, it is mainly the differential voltages at the ends of the line that will be considered hereafter, i.e. V_d (o, ω) and V_d (L, ω).

The way these voltages are generated is rather complex. We will speak first of the low frequencies, where the hypothesis holds that the disturbance wavelength is much greater than the length of the lines, or λ ≫ L.

The differential voltage may come from a direct action due to the electromagnetic coupling between the disturbing current and voltage and the wire pair itself. Or it may be an indirect effect due to the transfer of common-mode voltages into a differential voltage. It should be known in fact that the differential voltage is quite often small compared with the common mode voltage, so that any common-mode transfer will be large and even in certain cases preponderant compared to the direct differential voltage.

To clarify this point of obvious practical interest, let us refer to the equivalent diagram of the disturbed system in figure 23

In this figure, we see voltage and current generators, the former being equivalent to the effect of the magnetic coupling on pair conductors 1 and 2, while the latter symbolize the effect of the electrical coupling. These generators are expressed simply, as a function of the source terms in equations (1) and (2), since:

$$E_1 = jL_{p1}\omega\, I_pL$$
$$E_2 = jL_{p2}\omega\, I_pL \qquad (7)$$

$$J_1 = jC_{p1}\omega\, V_pL$$
$$J_2 = jC_{p2}\omega\, V_pL \qquad (8)$$

in which $j = \sqrt{-1}$

The direct action of the disturber on the differential mode in the diagram will correspond to voltage generators of slightly differing amplitudes, and to current generators of different amplitudes. $V_{d(0)}$ and V_d (L) are then expressed

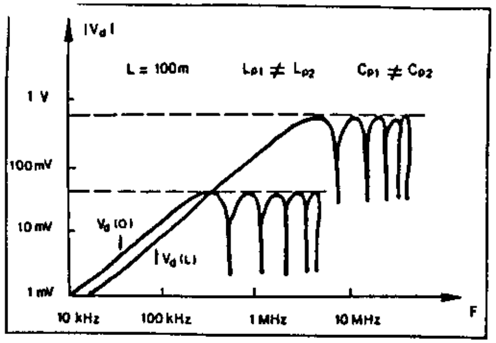

Figure 24 Curves showing the variation of the differential voltages V_d (O) and V_d (L) as a function of the frequency in the presence of direct action of the disturber.

$$V_{d0} = \frac{1}{2} \left| (E_1 - E_2) + \frac{Z_d}{2} (J_1 - J_2) \right|$$

$$V_{dL} = \frac{1}{2} \left| (E_1 - E_2) - \frac{Z_d}{2} (J_1 - J_2) \right| \qquad (9)$$

This relation, established under the long wavelength hypothesis ($\lambda \gg L$), assumes that the disturbance propagates as a standing wave, or:

$$I_p (Z, \omega) = I_p (0, \omega) = I_p e^{-\gamma_p z}$$

$$V_p (Z, \omega) = Z_{pc} \, I_p (Z, \omega) \qquad (10)$$

in which γ_p is the propagation factor per unit length of the disturber line and Z_{pc} its characteristic impedance. Relations (9) are also established under the hypothesis that Z_{c0} and Z_{cL}, connected to the ends of the wire pair, are identical with each other and with the differential mode characteristic impedance Z_{cd}. In the presence of a very weak electric coupling, the differential voltages at the cable ends are identical in absolute value, but this is not a general rule.

Admitting now that the magnetic and electrical couplings acting on conductors 1 and 2 are identical, then $E_1 = E_2$ and $J_1 = J_2$ and the differential parasitic voltages are nul. This situation occurs when the geometry of the wire pair obeys certain strict symmetry criteria. A twisted wire pair may fulfill these conditions, so other causes must be found for the differential parasitic voltages. They may, as was mentioned above, stem from a common-mode transfer, as would occur for instance when the impedances connected between the conductors and ground references are not strictly identical. Generally, these impedances are very large and the effect of this asymmetry is masked by the linear capacitances of the conductors represented in figure 23

As has already been emphasized, the differential voltages are generally very small compared with the common mode voltages. Even when masked by the conductor capacitances, this phenomenon may be perceptible. To calculate these differential voltages, we simply solve for the circuit in figure They can be approximated by the relations:

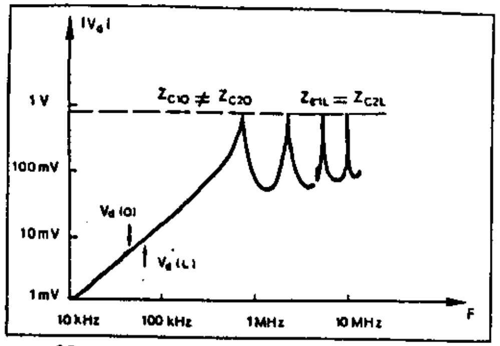

Figure 25 Variation of the differential voltages with the frequency by indirect action of the disturber due to transfer from the common mode to the differential mode caused by a lack in which : of balance of Z_{c10}, Z_{c20}.

in which:

$$Z_1 = \frac{Z'_{01} \cdot Z'_{L1}}{Z'_{01} + Z'_{L1}}$$

$$Z_2 = \frac{Z'_{02} \cdot Z'_{L2}}{Z'_{02} + Z'_{L2}} \qquad (12)$$

$$Z' = \frac{Z_c}{1 + jZ_c \frac{L}{2} C\omega} \quad \text{où } j = \sqrt{-1}$$

in which $j = \sqrt{-1}$

Another source of transfer may be a small unbalance in the line capacitances C_{11} and C_{12}. The differential voltages induced by these phenomena are often of the second order, but may have an effect when the wires are very long

The expressions (9) and (11) become highly erroneous, however, when the propagation phenomena come into play. That is, when the condition $\lambda \gg L$ no longer holds, neither does the schemes in figure 23 and systems (1) and (2) must be solved strictly.

The propagation phenomena then bring out two new facts. Firstly, contrary to relations (9) and (11), the numerical results show that above a certain frequency limit that depends on the length of the cable, the amplitude of the induced voltages becomes insensitive to this parameter. Secondly, the variations of these voltages with the frequency are accompanied by fluctuations that can be attributed either to a beat frequency between waves propagating at different velocities, or to standing wave phenomena, which we will demonstrate with a few examples.

Figures 24 and 25 illustrate how V_d (O) and V_d (L) vary, in numerical simulations of the systems (1) and (2).

Figure 24 corresponds to a system in which only the direct action of the disturber is at play, in differential mode:

$$E_1 \neq E_2, \; J_1 \neq J_2, \; Z_{c10} = Z_{c20} = Z_{c1L} = Z_{c2L}$$

Figure 25 on the other hand, corresponds to a system with indirect action by common mode transfer:

$$E_1 = E_2 \quad J_1 = J_2 \; Z_{c10} \neq Z_{c20} \; Z_{c1L} = Z_{c2L}$$

$$V_d (0) = -V_d (L) \cong \frac{1}{2} Z_d \; \frac{(E_1 + E_2)(z_1 - z_2) - (J_1 + J_2)(z'_{01} z_1 - z'_{02} z_2)}{z'_{01} + z'_{02} + z'_{L1} + z'_{L2}}$$

Considering the imposed matching conditions in two-wire mode, the propagation phenomena look like a beat between two standing waves propagating at different velocities, as is quite clear in figure 24 The propagation phenomena also contribute to modifying the relative maximum amplitude the parasitic differential voltages at the cable ends.

As concerns the reaction of the differential voltages to the common mode transfer, the propagation produce resonances that originate from common mode reflections at the ends of the line. We may point out that the spurious signals are then always the same, regardless of the disturber frequency.

Of course in reality things are more complex. The disturber is not necessarily a standing wave, and direct and indirect couplings may be present everywhere. But the theoretical model can still be a precious tool, insofar as it provides a way of evaluationg the amplitude of the coupling and the role played by the input impedance of the equipment connected to the system.

III) COUPLING OF CONDUCTED INTERFERENCES

As indicated above, we will bring the ground reference plane into play here. That is, we will attribute a conductivity to it, and this non-infinite conductivity means it has a surface impedance Z_s. If it is a metal plane of good conductor material, the surface impedance can be introduced in the solution of the coupled-line equations (1) and (2) by simply adding the plane impedance to the coupling impedance Z_p by:

$$Z_p = Z_s + jL_p \omega \qquad (13)$$

Metal ground planes are rare, however, and what we usually have is conductors that abet the flow of "ground currents". These conductors have a specific impedance which contributes to the conductive coupling. We distinguish here between the case where this impedance is uniformly distributed and the case where it is localized at certain points in the ground circuit.

In the first case, the impedance is linear in nature and generally reactive. To the coupling impedance L_p we add a surface inductance per unit lenght L_s:

$$Z_p = jL_s\omega + jL_p \omega \qquad (14)$$

The discontinuities will be treated as voltage sources in the coupled line system of equations (1). These impedance discontinuities may result from loops formed in the ground circuit, which is a situation very frequently encountered on railroad vehicles.

When the systems in question are physically large, the ground reference may be the earth or any network of conductors in contact with the system. The structure of the ground circuit then becomes very complex and is often difficult to define, compounded by the fact that the impedance of the ground connectors becomes very important when the frequency of the disturber increases.

In contrast to influence coupling, conductive coupling calls for an empirical model, which can be tidily summed up as a precise analysis of the site topography, along by a few preliminary measurements to get an idea of the coupling "level".

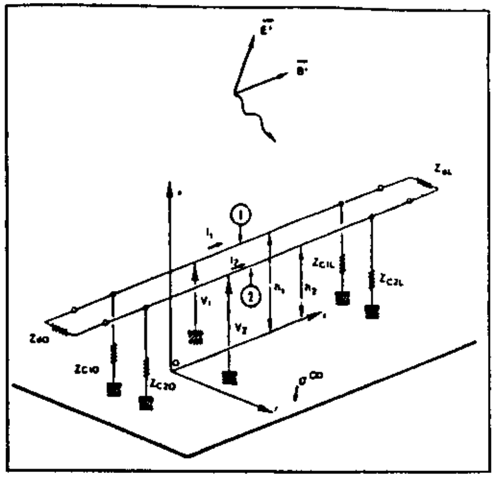

Figure 26: Simplified model for simulating the radiative coupling on a two-wire line parallel to an infinite conductive plane.

IV) COUPLING OF RADIATED INTERFERENCES

If we refer to the system with the wire pair running parallel to the ground plane, radiative coupling can be analyzed by the coupled line theory (figure 26. The coupled transmission line equations can be generalized to the problem of wave-system interaction. To do this, equations (1) and (2) from the previous section will be expressed:

$$-\frac{dV_1}{dz} = Z_{11} I_1 + Z_{12} I_2 + E_1$$

$$-\frac{dV_2}{dz} = Z_{12} I_1 + Z_{22} I_2 + E_2 \qquad (15)$$

$$-\frac{dI_1}{dz} = Y_{11} V_1 + Y_{12} (V_2 - V_1) - J_1$$

$$-\frac{dI_2}{dz} = Y_{12} (V_2 - V_1) + Y_{22} V_2 - J_2 \qquad (16)$$

The source terms E_1, E_2, J_1 and J_2 refer to the action of the disturber. If it is a plane wave, the induced emf's E and currents J are expressed by the integrals:

$$E_{1,2} = j\omega \int_0^{h_{1,2}} B^i_{y1,2} \, dx \qquad (17)$$

$$J_{1,2} = Y^{1}_{2\,2} \int_0^{h_{1,2}} E^i_{x1,2} \, dx \qquad (18)$$

In which $E^i_{x1,2}$ and $B^i_{y1,2}$ are the electromagnetic field components of the wave incident on conductors 1 and 2, and h_1 and h_2 are the heights of conductors 1 and 2, respectively, above the plane.

At large wavelengths $\lambda \gg L$, it is entirely possible to transpose the scheme of figure 8. On the other hand, the propagation phenomena come into play at high frequencies and then (15) and (16) must be solved strictly.

... coupled lines equations, if applied to the problem of the interaction of a wave with a wire system, can only solve systems consisting of wires parallel to a plane of good conductor, and at a distance from the plane that is small compared with the wavelength. When these conditions are not met, the system must be treated as a receiver antenna for which we want to find the induced current distribution.

The problem is even more complex when the system is installed in a partially shielded three-dimensional enclosure. We must then proceed in two phases, first evaluating by diffraction theory the field that enters the enclosure and then, once the field distribution is known, evaluating the currents and voltages induced in the wiring.

Except for a few very simple situations, the interaction of a wave with a system requires an attentive study of the phenomena accompanying the propagation of the electromagnetic field. The solution will often consist of breaking the system down into a set of subsystems whose electromagnetic properties are easier to analyze.

V) Shielding

CABLE SHIELDING

To modify the consequences of the electromagnetic coupling caused by a disturber, it is worthwhile protecting the systems with shielding. The effectiveness of the shielding used depends on its intrinsic properties, but also on a few rules of rational use. If we refer to the system comprising two wire conductors, installing shielding adds the following new components to the electromagnetic chain (figure
1. wire pair cable shielding;
2. equipment shielding;
3. Shielded connectors;
4. Ground reference links;
5. Ground reference.

Impinged by an electromagnetic interference, the shielded system can be broken down into two interdependent subsystems. The outer one, directly exposed to the disturber, includes the shielding and ground connection. The inner one is reduced to the wire pair and the equipment input impedances, and receives electromagnetic energy that is theoretically attenuated by the outer system. If the shielding is of sufficient quality and is correctly used, the reaction of the inner system on the outer one can be neglected. This considerably simplifies the approach to the problem since, to begin with, we then have only to determine the amplitude of the currents I_B (Z, ω) and voltages V_B (Z, ω) induced on the components of the outer system. Knowing the intrinsic properties and parameters of the shielding, we then evaluate the parasitic signals transferred from the outer to the inner system.

One essential task will therefore be to characterize the four elements of the chain diagrammed in figure 27

The electromagnetic efficiency of the cable shielding can be characterized by a reduction factor and a transfer impedance. The reduction factor, used mostly at low frequencies (50 Hz and harmonics, is the ratio of the spurious signals induced between a wire and a ground reference when the cable is impinged without shielding, with respect to when it is shielded.

The transfer impedance is a parameter per unit length that can be expressed by the ratio

$$Z_t(\omega) = \frac{1}{L} \frac{V_c(\omega)}{I_B(\omega)} \qquad (19)$$

$$L \leq \lambda$$

V_C (ω) is the parasitic voltage that appears at the end of a coaxial when the shielding is carrying the current $I_B(\omega)$ (figure 28)

The excerimental determination of the ratio (19) nonetheless requires the condition of the electrically small line. The transfer impedance can be considered as a coupling parameter just like the coupling impedance Z_o in the coupled line equations (1). The variation of the transfer impedance with frequency will therefore tell us something about the quality of the shielding

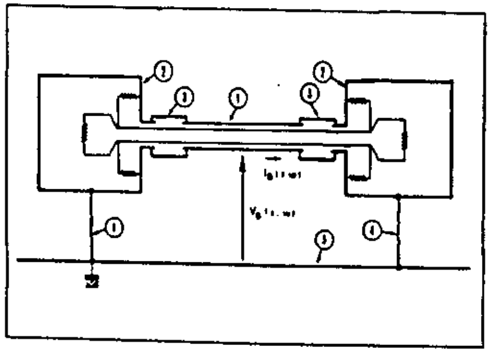

Figure 27 Model of two-wire transmission line protected by electromagnetic shielding.

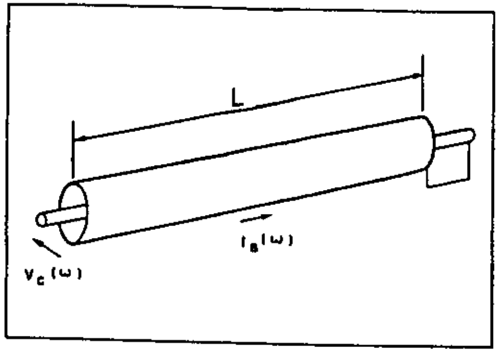

Figure 28 Principle of defining and measuring a coaxial cable transfer impedance.

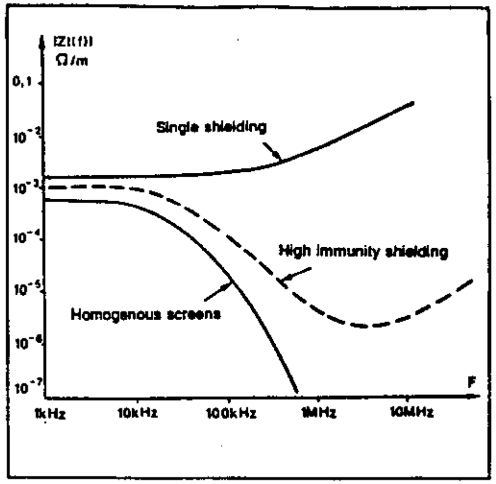

Figure 29 Transfer impedance characteristics with coaxial cables.

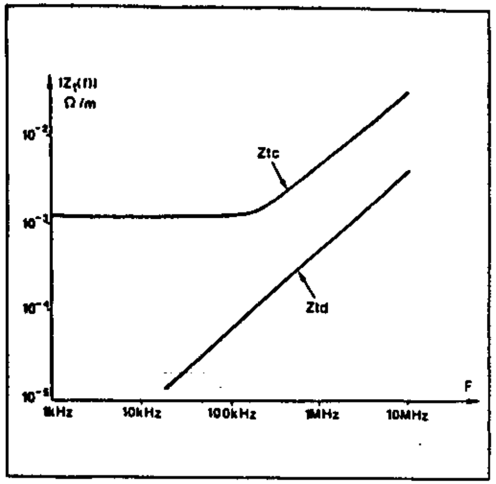

Figure 30 Transfer impedance characteristics for the common mode Z_{tc} and transfer impedance for the differential mode Z_{td} measured on a sample of a braid-shielded cable pair.

and provide us with a quantitative parameter for predicting the amplitude of the residual parasitic signals reaching the equipment.

Concretely, the transfer impedance characteristics lie in an envelope between two behavior limits (figure 29 cables for which the transfer impedance will increase beyond a certain transition frequency, and those for which Z_t is a monotonically decreasing function as the frequency increases. Ordinary cables with a single shielding system of braid or wound tape are of the first kind. Cables protected by a homogeneous screen of good conductor are of the second kind. There are other technologies combining the two kinds of shielding (braid, tape and homogeneous screen) and with intermediate properties, and these are cables of high electromagnetic immunity.

An ordinary coaxial cable of 7 mm diameter shielded by a copper braid with good cover has a transfer impedance of a few milliohms per meter at about 1 kHz. Over 1 MHz, this impedance will begin to increase and will reach 10 mΩ/m at about 10 MHz. The same cable, when equipped with a homogeneous copper screen a few tenths of a millimeter thick, has the same transfer impedance at 1 kHz and below, while as soon as the frequency goes beyond 10 kHz it will decrease, down to as low as a few microohms per meter at about 1 MHz. This is an effect of the electromagnetic field diffusion phenomena in the conductor materials. This diffusion can be further increased and made more sensitive to the lower frequencies by combining screens of good conductor material with screens of high magnetic permeability. To reduce or eliminate the rise in the transfer impedance, we can also combine braids with thin homogeneous screens, which reconciles the requirements for mechanical flexibility and electromagnetic compatibility.

Transfer impedance as we have just described it has no reality except for coaxial cables. If we are dealing with shielded pairs, or even moreso with shielded multiconductors, we must again distinguish between common-mode and differential-mode disturbance.

For the common mode, the transfer impedance Z_{tc} is practically the same as what we find on the equivalent coaxial structure. But the differential mode induced on the shielded pair may be produced by juxtaposing two effects: a common-mode transfer by a process analogous to the one described for influence coupling, and a mutual induction caused by direct coupling through the shielding. The differential transfer impedance Z_{td} for the direct coupling can be measured in the laboratory on test specimens of reasonable sizes.

For ordinary cables including a braid or wound tape, Z_{td} increases monotonically with the frequency of the disturber. The differential transfer impedance is often very much less than that of the common-mode (figure 30. The effectiveness of a shielded pair or multiconductor will therefore depend on how much common-mode transfer there is, compared with the direct differential mode induction.

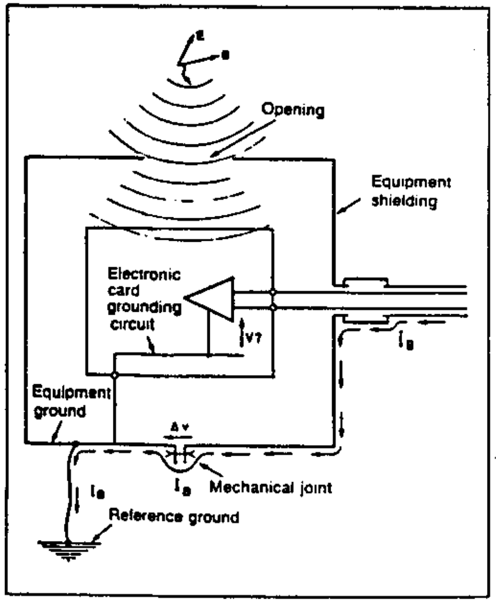

Figure 31 Diagram showing the main elements that might reduce the effectiveness of a shielded enclosure intended to protect electronic equipment from the effects of electromagnetic radiati...

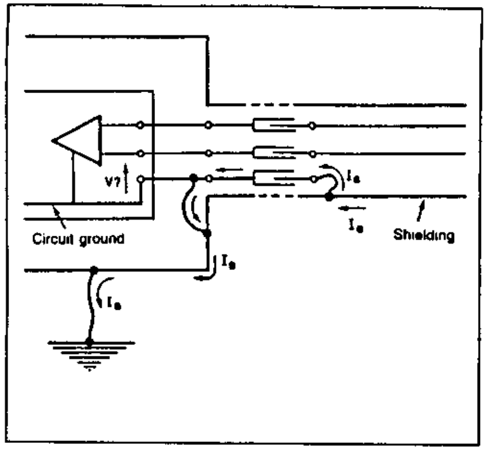

Figure 32 Shielded cable connection to shielded equipment, using an ordinary connector.

Equipment shielding

Considering the three-dimensional geometry of the enclosures protecting electronic equipment, it is nowhere near as easy applying the methods laid out for shielded cables. We will prefer a more qualitative approach, attempting to list the most vulnerable points of the enclosure and indicate the openings and risks of electrical discontinuity in the shielding.

The metal enclosure that protects the equipment is generally not hermetically sealed and there remain openings that make it easier for electric and magnetic field components from a disturber to enter the enclosure. This will result in direct induction of spurious voltages on the wiring inside. Also, the diagram in figure 27 shows that the metal enclosure is used at the same time as a ground reference for the equipment circuits and as a surface of flow for the current the disturber induces on the shielding. Moreover, any electrical continuity defect at the mechanical joints between the metal parts may set up a local parasitic emf that will couple with the equipment ground circuits by conduction (figure 31).

It is above all the search for adequate technological solutions that will guarantee the quality of the enclosure shielding, while reducing the sizes of the openings and bypassing suspect metal joints with braids.

Connector shielding

Connectors provide threefold continuity: of the signals transmitted by the wire pair, of the ground circuits between the equipment units, and the electrical continuity of the shielding. Two technologies are commonly used, i.e. ordinary three-pole connectors and shielded three-pole connectors. In the former (figure 32 the neutral contact provides the equipment ground continuity at the same time as the electrical continuity of the shielding. This often breaks the symmetry of revolution of the shielding over a space of a few centimeters.

The second solution (figure 33 differentiates the ground line from the electrical continuity of the shielding. The neutral contact connects the equipment grounding circuit electrically by means of a third conductor inside the shielded cable, and the symmetry of revolution of the shielding is guaranteed by the metal cylinder protecting the connector.

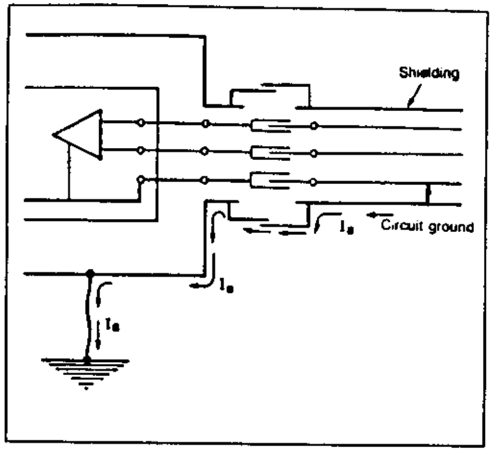

Figure 33 Cable connection to equipment by shielded connector.

The usual connector is of course not as good from an EMC point of view, since the disturber current flows in the neutral contact and especially through the shielding connections These shielding connections may exhibit an inductance of a few nanohenrys. This inductance is sufficient to produce spurious voltages that are equivalent to those obtained over several meters of ordinary shielded cable.

The shielded connector therefore seems to be the most appropriate solution, since it also separates the equipment ground from the shielding. The shielding quality of a connector can be measured in the laboratory. It is also similar to a transfer impedance that may vary widely depending on the level of uniformity of the contact between the fixed and mobile ring of the connector.

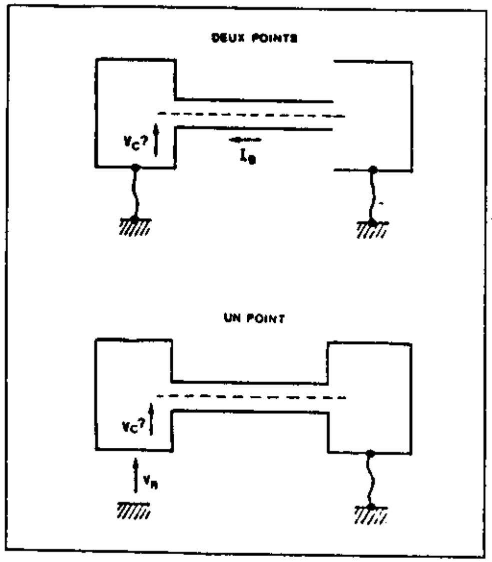

Figure 34 Very simplified diagram illustrating the ground connection at two points (top) and at a single point (bottom).

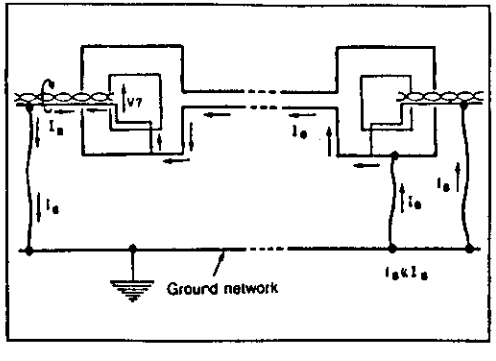

Figure 35 Illustration of the risk of branching of induced currents by the ground wire in the cable supplying line power to a piece of equipment.

Ground connections

Ground connections provide the electrical links between the shielding and the ground reference, consisting of networks connected to the ground connectors. In appearance, they play two contradictory roles. By holding the shielding at the equipotential of the reference, they provide a path of flow for the currents induced by the disturber. On the other hand, the flow of current in the shielding may cause a parasitic voltage in the equipment. From a practical point of view, the ground link is indispensable. So there is a choice to be made between linking the shielding at one or at more than one point to the reference ground. This is a hotly debated subject. The example taken from the simplified model of figure 27 will shed some light on the difficulty of the choice. We will distinguish here between the case where the two units are in contact with the reference ground, and the case where only one is (figure 34).

We analyze first what happens at very low frequencies, i.e. at wavelengths much greater than the space separating the units. The system can then be equated to a very simple element.

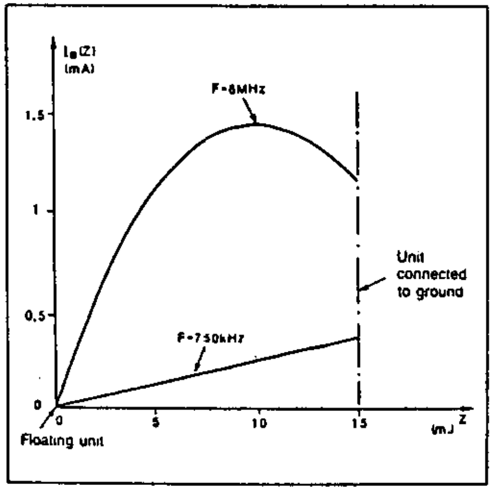

Figure 36 Distribution of the current induced in the shielding of a coaxial cable interconnecting one unit that is connected to the ground, to another unit that is not. The units are 15 m apart and the measurement is taken with an electromagnetic disturbance of the plane wave type, at frequencies of 750 kHz and 8 MHz.

Impinged by a plane electromagnetic wave, the shielding, ground connections and reference behave like a loop that is the seat of an induced current I_B independent of the frequency. The risk for the electronic circuits in this case is a spurious signal due to the imperfection of the shielding. If a single piece of equipment is connected to the reference, the disturber sets up a voltage V_B between the floating equipment and the ground reference. This difference of potential, proportional to the frequency of the disturber, has no direct incidence on the circuits. On the other hand, it may have an indirect effect when the floating equipment is supplied from line power (figure 35).

In this configuration, the floating equipment is in contact with the reference anyway, through the ground circuit of the line power supply. There results a loop current I_B capable of producing a magnetic coupling on the line cable and inducing voltages V_C in it. If the amplitude of these voltages is high enough, they can move toward the electronic circuits and disturb the most vulnerable components.

This risk is not so great for the enclosure connected to the reference, insofar as the conductor linking it to the reference has a low impedance. The loop through the line power ground circuit can be eliminated, and the advantages of the floating equipment can be maintained, by installing filters on the line power supply. This solution is generally good if the disturbing spectrum does not exceed a few kilohertz.

If we now consider disturbances at very high frequencies, the condition of the electrically small circuit no longer holds and the connection to the ground at a single point of contact no longer offers any great advantages. The propagation phenomena come into play and, even with a floating unit, a current I_B is induced in the shielding elements, with an amplitude that is greater as the disturbing frequency increases.

The curves in figure 36 illustrate the variation of the current distribution I_B (z) on the shielding of a cable connecting two electronics units 15 m apart. One is connected to the ground reference and the other is floating (z = 0).

The system is illuminated by a plane electromagnetic wave at a frequency of 750 kHz and then 8 MHz. At 750 kHz, the hypothesis of the electrically small circuit is well verified, as the current measured in the cable shielding is small. This current is maximum at the level of the unit that is connected to ground, illustrating the effect of the capacitance that exists between the system and the reference plane. At 8 MHz, there is a major increase in the current and a shift of the maxima to a point between the two units. This behavior is closely related to the

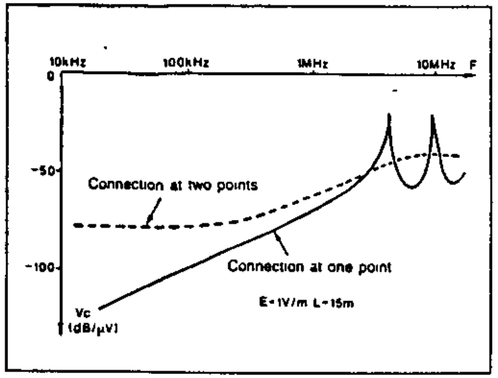

Figure 37 Variation, as a function of frequency, of the parasitic voltage measured at the end of the coaxial, with both units connected to ground and with only one. It can be seen that the second solution offers satisfactory protection at low frequencies, while propagation phenomena create resonances of an appreciable level at the higher frequencies.

propagation phenomena that arise on the transmission line comprising the shielding and the reference. These phenomena cause the residual parasitic voltage resonances at the equipment inputs.

The curves in figure 37 referring to the conditions of the electromagnetic disturbance described above, show the variation of the parasitic voltages measured during the experiment, when the shielding of the coaxial cable is connected to the reference at two points and then at a single point. Though the second solution is advantageous at low frequencies, the situation changes radically when the resonances appear.

This example shows how primordial the connection mode of a system to its ground references can be for its EMC. It is sometimes preferred to get around this phenomenological approach by using overdimensioned shielding or filters. This (economically) not very attractive solution is inciting more and more designers to concern themselves with the ground connection organization. When confronting a complex system, though, the optimum connection mode is not the immediate concern of research. It is after an attentive study of the equipment, of its power supply and of the spectrum of disturbances to which it will be exposed, that is will then be possible to establish the architecture of a spider, lattice or hybrid ground network.

Effect of EMI on electronic components

We will speak here only of integrated logic circuits subject to nondestructive EMI. Contrary to the electromagnetic coupling phenomena on wiring, or to the transfer of the field through the shielding, EMI acts in a highly nonlinear fashion on electronic components . Any method designed to characterize the sensitivity of the component to EMI must account for this nonlinear character.

The usual transfer functions do not apply to quadripoles or "n" linear poles, so we must analyze the reaction of the component to the various time-dependent signatures to which it is exposed. That is, an integrated circuit will react differently depending on whether it is subjected to a narrow, large-amplitude pulse, a broad pulse of moderate amplitude, a damped wave or to a sustained harmonic wave.

Disturbances also have very different consequences, depending on whether they superimpose on the logic signals at the circuit input or output or rather act on the bias currents and voltages. Considering the complexity of the problem, we often have to settle for analyzing very simple situations such as the example of two NAND logic gates in figure 38

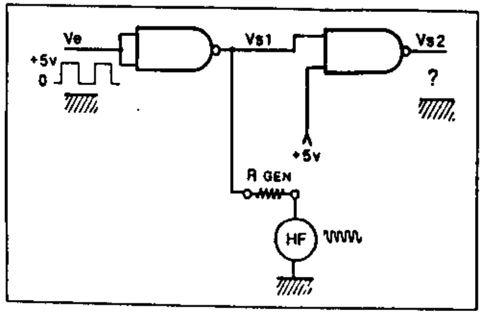

Figure 38 Experimental setup for studying the sensitivity of logic circuits to EMI.

A logic signal consisting of recurrent pulses passing back and forth between low and high states is applied to the input of circuit 1. Experience shows that the component is mainly disturbed when the disturbance is applied to the output of the first flip-flop. It is under these conditions that the threshold will be found beyond which the RF level will cause a change of logic state at the second flip-flop output, as a function of the frequency of a harmonic signal. The curves in figure 39 show that the RF threshold varies as a function of the EMI frequency and depending on the component technology.

These experiments also confirm the conclusions of numerical simulations of circuits as published in the literature. Both the numerical and experimental results are, however, too small in number to come to any conclusions in this very particular field of EMC. However, this line of research merits further investigation insofar as the security of the systems will often depend on the reaction of the components under exceptional circumstances.

Conclusion

No general statements can be made about the electromagnetic compatibility of systems as complex as the electronic equipment carried on board vehicles. Applying the few principles and properties that have been discussed in this paper assumes that an attentive study is first made of the whole system, to break it down into elementary subsystems. This is the only possible way of discovering the electromagnetic couplings that will provide a basis for deciding on the level of shielding protection needed. The technologies that have been developed for onboard equipment generally lend themselves well to this type of approach insofar as the architecture of the systems responds to a series of clearly perceivable hierarchical levels. As the common point for any equipment is the ground reference, a significant reduction in the risk of parasitic signals sensed by the electronic circuitry can be expected if this ground network is organized rationally. This is particularly true if the shielding is connected electrically to the reference ground.

There is no hurry to choose between the spider connection and the lattice, and the decision will often depend on parameters that are considered to be secondary, such as the effect of the equipment power supply cables or of the propagation phenomena on the outer part of the shielded cables.

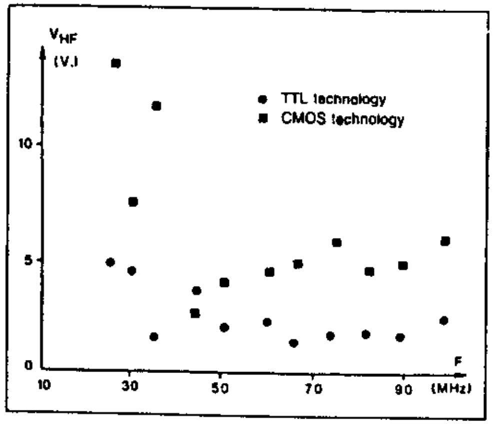

Figure 39 Variation of the peak-to-peak threshold of the RF signal triggering a logic state change in NAND flip-flops, versus the frequency of the signal. Above 30 MHz, TTL and CMOS components behave differently.

The effectiveness of the protection provided for an installation also depends for a large part on the quality of the shielded cables, along with an excellent regularity of the electrical continuity of the shielding through the transition of the connectors.

If it is correctly protected against the usual electromagnetic disturbances, a system may nonetheless be found to be sensitive to disturbances of exceptional amplitude. This can result in parasitic signals large enough to change the function of an active electronic component. The characterization of the components under nondestructive impingement will have an impact on the level of safety or availability of a system.

This inhabitual approach to component characterization is certainly indispensable as a help to designing future automatic systems integrated in vehicles. One possible perspective of the EMC studies might in fact concern the search for component tolerance thresholds as a function of their technology, switching speed, frequency and the type of disturbing signals.

C) BIOLOGICAL EFFECT OF E.M. WAVES

I) INTRODUCTION

The electromagnetic radiation can be divided into two broad categories : ionizing and non-ionizing radiation. If the frequency is high enough, a photon can interact with the medium giving rise to an electron pulled out of an atom. If this atom belongs to a molecule, the resulting ionization can destroy the chemical bond. Furthermore if the molecule itself belongs to a gene inside the nucleus of a cell, its genetic code may be modified. If the cell survives, it may either proliferate (cancerigenic effect) or simply keep a change producing a permanent mutation.

Of course the probability for such effects occurs and the risk is directly proportional to the total quantity of photons received by the human body. Norms have been established specifying the maximum density of ionizing photons which can be received by a human body for a given duration. The ionizing radiation extends from U.V. to X and gamma rays. (Figure 40)

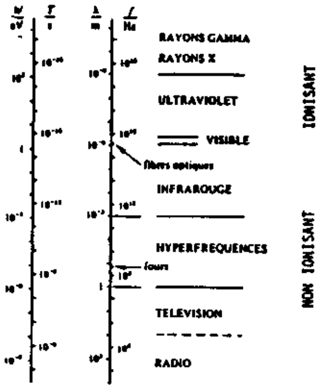

Figure 40

At smaller frequencies : visible, I.R and all the electromagnetic radiation used for telecommunication purposes for example, a single photon is not enough energetic to ionize an atom or to modify a chemical bond. This radiation is called "non-ionizing".

The interaction between the incident wave and the biological medium is due in this case to the combined effect of an important quantity of low energetic photons. The electromagnetic field may produce either a heating or an ion migration. Destructive effects only appear if the incident power density raises to a certain level. Below this level, the effect is usually non-cumulative. In the following only the effects of non-ionizing radiations will be briefly presented.

II) HEAT EFFECT

When a material is illuminated by an electromagnetic wave, a heat effect appears. Indeed, depending on the frequency and on the conductivity of the material, the electromagnetic power penetrates more or less in this material due to the skin depth effect. For the human body, the heat effect is mainly due to the rotation of the water dipolar molecule, at a frequency identical to the one of the disturbing field, and to the presence of equivalent frictional force. This fact is characterized by an electric parameter ε_r'' which is the imaginary part of the relative dielectric permittivity.

The power absorbed per m^3 is given by

$$P_{abs} = 2\pi\varepsilon_0 \, f \, |E|^2 \, \varepsilon_r''$$

In the different parts of the human body, the average value of ε_r'' varies from 1.5 for the lung, 8 for the crystalline lens of the eye and to 14 for a muscle. Favorable effects of a heating of the human body are well-known : increase of the flow of blood, vasodilatation, muscle relaxation... The maximum temperature which can be reached before cells destruction is about 43° C. Various frequency bands have been allocated for biomedical applications : 13, 27, 40, 900 et 2400 MHz.

An excessive heating in a special place may also be used to kill cancerous cells.

In conclusion burns may occur if the electromagnetic field power density is too intense but can be easily avoided if the norms are followed.

III) NON HEATING EFFECT

Athermal effect has been and remains a controversial subject. Indeed experiences must be made on a great number of animals, divided into two parts. The first one is exposed to the EM field while the other is the reference group. As usual is medical experiments, many parameters may influence the results and it is sometimes difficult to get precise answers to the possible non heating effects of E.M. waves.

Most of the "positive" experiences have been performed using a very low frequency E.M. field. Indeed, it seems that extremely low frequencies may interact with the nervous system.

The influence of the 50 Hz and 60 Hz electric field has been extensively studied since near high voltage lines, the vertical electric field reaches few kV/m. However one cannot conclude that 400 kV lines are detrimental to health, even in the vicinity of these lines.

Others athermal effects may occur if the human body is illuminated by signals with a very high peak values but having a too small average value to give rise to a heating (radar signals for example). An interaction between the E.M. waves and the internal ear has been reported.

However any of the numerous experiences has clearly shows that a harmful athermic effect appears if the E.M. field intensity remains below the usual norms.

IV) NORMS

Figure 41 shows the limits of the radiation level for various norms or countries. It must be noted the decrease of the authorized

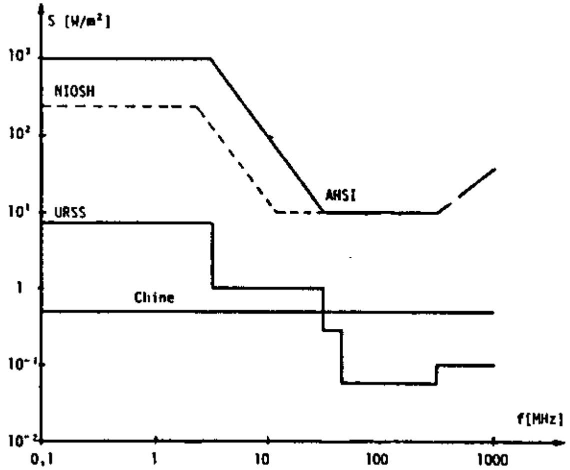

Figure 41

maximum level in the 30 MHz - 300 MHz frequency range. This band corresponds to the maximum absorption of the human body since resonances may occur. From each side of this band, the absorption varies as $1/f^2$ below 30 MHz and as $1/f$ above 300 MHz. Furthermore, for the most critical case corresponding to an incident electric field parallel to the human body, the specific absorption rate (SAR) should not rise above 0.4 W/kg. The levels may exceed the limits defined by the norms if the duration of exposure is less than 6 minutes as long as the mean value of the received level does not exceed the norm.

Lastly, the U.S. norms specify maximum values while USSR norms apply to a mean value on the whole body. Thus, although the norms apparently differ from a factor 100, the time difference is much smaller.