AGARD-LS-177

AGARD

ADVISORY GROUP FOR AEROSPACE RESEARCH & DEVELOPMENT
7 RUE ANCELLE 92200 NEUILLY SUR SEINE FRANCE

AGARD LECTURE SERIES 177

Electromagnetic Interference and Electromagnetic Compatibility

(Interférences Electromagnétiques et Compatibilité Electromagnétique)

NORTH ATLANTIC TREATY ORGANIZATION

Distribution and Availability on Back Cover

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This material in this publication was assembled to support a Lecture Series under the sponsorship of the Electromagnetic Wave Propagation Panel of AGARD and the Consultant and Exchange Programme of AGARD presented on 10 to 11 June 1991 in Kjeller, Norway, 13 to 14 June 1991 in Königswinter (near Bonn), Germany, and 17 to 18 June 1991 in Lisbon, Portugal.

North Atlantic Treaty Organization

Organisation du Traité de l'Atlantique Nord

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— Providing scientific and technical advice and assistance to the Military Committee in the field of aerospace research and development (with particular regard to its military application);

— Continuously stimulating advances in the aerospace sciences relevant to strengthening the common defence posture;

— Improving the co-operation among member nations in aerospace research and development;

— Exchange of scientific and technical information;

— Providing assistance to member nations for the purpose of increasing their scientific and technical potential;

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Abstract

Two aspects of the current electromagnetic (EM) environment having great significance for NATO systems are:

(a) electromagnetic interference (EMI) arising from both natural and man-made sources;

(b) electromagnetic compatibility (EMC), i.e. the ability of an EM system to function as specified without being susceptible to EMI, and without itself generating excessive EMI which would cause other systems to malfunction.

The Lecture Series will first set EMC in a NATO operational context. Major EMI generation mechanisms will then be reviewed and their characteristics outlined. The manner in which EMI energy couples into EM systems will be discussed, together with the analysis and modelling tools available to assist in computing such interactions.

Modern EMC testing methods and environments will then be examined and their limitations indicated. The relationship between NATO and civilian EMC requirements will be examined in the light of the European Community EMC Directive. Consideration will also be given to the problems of spectrum management and conservation as they affect systems intentionally radiating EM energy, e.g. radio, radar, etc.

Finally, design principles and techniques for EM systems with effective EMC characteristics will be presented. A Round Table Discussion will enable attendees to interact in some detail with the lecturing team.

This Lecture Series, sponsored by the Electromagnetic Wave Propagation Panel of AGARD, has been implemented by the Consultant and Exchange Programme.

Abrégé

Les deux aspects du milieu électromagnétique (EM) actuel qui revêtent une importance particulière pour les systèmes mis en oeuvre par les pays membres de l'OTAN sont:

(a) Les interférences électromagnétiques (EMI) créées aussi bien par des éléments naturels qu'artificiels.

(b) La compatibilité électromagnétique (EMC), qui est l'aptitude d'un système EM à tenir les spécifications sans être sensible l'EMI et sans générer lui-même des perturbations EMI excessives qui perturberaient fortement d'autres systèmes.

Le cycle de conférences situera d'abord l'EMC dans son contexte opérationnel au sein de l'OTAN. Les principaux mécanismes générateurs d'EMI seront ensuite étudiés et leurs caractéristiques indiquées. La manière dont l'énergie EMI est couplée aux systèmes EM sera discutée, ainsi que les moyens d'analyse et de modélisation proposés pour le calcul de telles interactions.

Les méthodes et les environnements d'essai EMC modernes seront passés en revue, avec indication de leurs limitations. Le rapport qui existe entre les besoins de l'OTAN et ceux de l'industrie civile en matière d'EMC sera examiné à la lumière de la directive de la communauté européenne sur l'EMC. Il sera tenu compte également des problèmes de la gestion et de la conservation du spectre dans la mesure où ils ont une incidence sur les systèmes qui ont pour fonction l'émission de rayonnements EM, à savoir les systèmes radar, radio etc...

Enfin, les principes et les techniques qui permettent la réalisation de systèmes EM dotés de caractéristiques EMC effectives seront présentées. Une table ronde organisée en fin de séance permettra aux participants de discuter des points particuliers avec les conférenciers.

Ce cycle de conférences est présenté dans le cadre du programme des Consultants et des Echanges, sous l'égide du Panel AGARD sur la Propagation des Ondes Electromagnétiques.

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INTRODUCTION & OVERVIEW: ELECTROMAGNETIC INTERFERENCE & ELECTROMAGNETIC COMPATIBILITY

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Electromagnetic interference (EMI) and electromagnetic compatibility (EMC) are topics which have been of concern to the military community for many years, and especially during the past half century when electrical and electronic equipments have proliferated. The ability of NATO forces to operate as planned over the complete range of operational scenarios is critically dependent on the correct functioning of a wide variety of electromagnetic (EM) systems, both individually and collectively; therefore, any foreseeable interactions and performance degradation of such EM systems must be minimised - preferably by effective design.

Excessive generation of, or susceptibility to, EM noise and interference, resulting in inadequate EMC, is one potential source of interaction and performance degradation which may result in reduced operational effectiveness. For this reason, the importance of achieving EMC has been stressed within NATO for at least the last two decades; bodies such as the EMC Section within NATO and the NATO EMC Advisory Group have been responsible for a continuing appraisal and analysis of actual and predicted EMC problems, leading to recommendations for action and specific projects. Within AGARD, the Electromagnetic wave Propagation Panel (EPP) and Avionics Panel (AVP) have also continued to promote scientific analysis and discussion of EMI and EMC problems, eg (AGARD, 1974) and (AGARD, 1987).

In general, it can be stated that the military EM environment is reasonably well defined in terms, say, of maximum threat levels; hence, it has also been possible to quantify EMC requirements with a degree of precision. In the civilian domain, however, an informed appreciation of the significance of EMI and EMC has been slower in developing, probably because, until relatively recently, that environment had a much lower average density and variety of EM systems than the typical military situation. Over the past decade, the civilian environment has become increasingly "rich" in sources of EMI and hence achieving EMC has now become a significant concern. Increased awareness of the likely implications of the installation of a vast range of EM equipments and systems, unspecified from an EMC viewpoint, has led to an increasingly comprehensive range of specifications and standards, culminating in the formulation and ratification of the EMC Directive by the European Community (EC Council Directive, 1989), by which virtually all EM equipments and systems manufactured and sold in the EC will be subject to legal EMC requirements with effect from 1992.

An increasing number of EM systems continue to be deployed for military purposes; these now have to take account not only of other military systems, but also of the parallel increase in civilian systems. In addition, it is becoming more common for sub-systems, developed primarily for civilian purposes, to be incorporated into military installations - a trend which is likely to escalate as the specifications of civilian systems improve. Also, a prolonged period of reduced international tension in the NATO theatre of operations is likely to lead to proportionately lower military budgets and expenditure on military EM systems. Thus, there will be an increased economic impetus in favour of civilian programmes. Again, it becomes probable, therefore, that military system designers will have to make greater use of, and adapt, primarily civilian EM equipments and systems to meet their requirements. Thus, military EMC planners and designers must be aware of related civilian developments.

One specific area in which the requirements of military and civilian systems are, to some extent, in conflict is in radio spectrum allocation. Increased numbers of civilian radio systems and services are creating a situation in which civilian bands are becoming progessively more congested and there is pressure to release bands in the spectrum which are currently largely assigned to military users. Also, the general level of manmade radio noise in all bands, due to a proliferation of EM systems, can be expected to become greater with time. A combination of reduced spectrum availability and higher noise levels inevitably puts an increased emphasis on the EMC aspects of radio-based systems, in both military and civilian contexts.

EMC is a term which can mean "all things to all men", and so it is appropriate here to provide a formal definition which will apply in this Lecture Series. There are two aspects to the definition (Darnell, 1988), ie EMC describes

(a) the extent to which a given EM system can function according to its design specification without generating EMI at a level which would cause other specified EM systems to malfunction - the "emission" aspect;

(b) the extent to which a given EM system can function according to its design specification in a defined EMI environment without itself malfunctioning - the "immunity" aspect.

It is seen, therefore, that EMC is "bi-directional" in that it is concerned both with EM energy emitted by a system and energy absorbed by a system. Essentially, therefore, EMC specifications and standards are concerned with the control of the "interface" between an equipment/system and its EM environment; Fig. 1 shows this interface diagramatically. On either side of the interface illustrated in Fig. 1, are listed a number of factors and trends which are tending to make the attainment of EMC a more difficult task as time progresses.

As an introduction to the detailed specialist lectures that follow, a brief overview of the objectives and content of Lecture Series No. 177 will now be given.

The first two lectures are of a fundamental nature and will provide essential introductory background material. Initially, the lecture on "Natural and man-made noise and interference: mechanisms and characteristics" will survey naturally-occurring and man-made noise and interference sources which have the potential to degrade the performance of NATO EM equipments and systems of various types. The source mechanisms will be described, together with their spectral "signatures" in the various bands of the radio spectrum. Analysis models for the effects of internal and external sources of EMI on EM system performance will also be presented. This will then be followed by a lecture on "Propagation and coupling mechanisms" which will provide a description of the physical basis of EM theory required to appreciate the coupling phenomena which can give rise to a lack of EMC.

In order to be able to treat EMI and EMC problems in a quantitative manner, it is necessary to be able to apply systematic analysis and modelling procedures. The lecture entitled "Numerical analysis and modelling techniques" will describe a range of such techniques which can potentially be applied to the EMC analysis of various types of coupling configurations. A companion lecture entitled "Cables and crosstalk" will discuss the analysis models applicable to the prediction of coupling effects between circuits in close proximity, and different conductors in cables, via inadvertent energy transfer mechanisms.

The first day of the Lecture Series will conclude with a lecture on "Shielding, grounding and bonding". Here, the principles, practice and implications of these three basic topics, of vital importance in achieving EMC, will be outlined.

Day 2 of the Lecture Series will commence with a lecture entitled "A review of the NATO EMC analysis programme and related civilian developments". This will describe in some detail specific activities associated with the NATO EMC analysis programme, particularly in relation to frequency assignment and planning; in this context, the interaction of military and civilian interests will be discussed.

The following lecture on "Spectrum management and conservation" will review the propagation mechanisms available to radio systems in the various bands of the radio frequency (RF) spectrum. The way in which frequency planning procedures are influenced by these mechanisms and equipment characteristics and imperfections will be considered, together with spectrum control procedures.

Attention will then be turned to common methods of assessing EMC characteristics of EM systems in a lecture on "Measurement environments and testing". The various test environments available will be described; their rationale, limitations and precision will also be examined.

The final lecture on "Design principles for effective EMC" will take a systems level approach to achieving adequate levels of EMC; the basic characteristics of systems affecting their EMC performance will be reviewed, and a number of design rules postulated. A general approach to tackling EMC problems will be suggested.

It is obvious that in two days only broad principles and essential concepts can be presented by the lecturers. A concluding Round Table session will provide an opportunity for participants to discuss with the lecturing team any, possibly more detailed, points arising from the individual lectures, and to examine general principles in greater depth.

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EQUIPMENT/SYSTEM	**EM ENVIRONMENT**
MORE EM SYSTEMS CAPABLE OF BEING AFFECTED BY EMI	GREATER NUMBERS AND DIVERSITY OF EMI SOURCES
LOWER-POWER LOGIC *FAMILIES, MORE SUSCEPTIBLE* TO EMI	FASTER CLOCK RATES SPREADING EMI ENERGY TO HIGHER FREQUENCIES
NON-METALLIC PACKAGING GIVING POORER ISOLATION	MORE MOBILE SOURCES OF EMI
INCREASED NUMBERS OF LARGE, DISTRIBUTED EM INSTALLATIONS	MORE ENVIRONMENTAL COMPLEXITY, EG ELECTRICAL NON-LINEARITIES
NON-EMC-ORIENTED DESIGN PROCEDURES USED FOR THE MOST PART	INCREASE IN EM EMISSIONS WITH TIME DUE TO AGING OF EQUIPMENT

INTERFACE
(CONTROLLED BY EMI/EMC
SPECIFICATIONS & STANDARDS)

Fig. 1: EMC as an interface control problem

NATURAL AND MAN-MADE NOISE AND INTERFERENCE: MECHANISMS AND CHARACTERISTICS
(Outline Only)

by

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1. Overview Lecture
2. Definitions of Noise and Interference
3. Sources of Radio Noise
   - 3.1 Categories of Sources
   - 3.2 Examples of Source Characteristics
4. System Noise Figure and Noise Factor
5. Worldwide Minimum Noise Levels
6. Empirical Noise Models
   - 6.1 Natural Noise from Lightning
   - 6.2 Man-Made Noise
   - 6.3 Analytical Noise Models
7. Effects of Noise on Systems Performance
8. Channel Occupancy and Congestion
   - 8.1 Definitions of Occupancy and Congestion
   - 8.2 Examples of Occupancy and Congestion
   - 8.3 Modeling of Congestion
9. Noise Measurements and Standards
   - 9.1 Field Strength Measurements
   - 9.2 Noise Detectors
   - 9.3 Standards (IEEE, SAE, CISPR)
10. Future Noise Trends

NATURAL AND MAN-MADE NOISE AND INTERFERENCE: MECHANISMS AND CHARACTERISTICS

by

G.H. Hagn

ABSTRACT

The objective of the lecture is to define noise (a cause) and interference (an effect) and describe some of the sources of both natural and man-made radio noise and interference which have the potential to degrade the performance of radio and other electromagnetic systems of interest to NATO. The following types of sources will be included: natural noise (e.g., from lightning, the sun and the cosmos), and man-made noise (e.g., from external sources such as powerlines and ignition systems and from sources internal to a receiving system associated with electronic devices and components). The source mechanisms will be discussed as well as the characteristics of the sources and their "signatures" in the bands from ELF through SHF. CCIR estimates of worldwide minimum effective antenna noise figures versus frequency and empirical noise models (for noise of natural and man-made origin) will be presented and discussed. Modeling the composite noise environment generated by multiple types of sources will be discussed. The analytical noise models of Hall and Middleton will be described. The proper combining (in a model of overall system noise figure) of the predictions of models for individual external and internal noise sources will be discussed. The effect of noise on analog voice and on digital communications systems will be discussed. HF channel occupancy and band congestion will be defined, and the measurement and modeling of congestion will be discussed. Noise measurements and standards will be reviewed, and future noise trends will be discussed.

Propagation and Coupling Mechanisms.

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Summary.

This lecture aims to cover the background electromagnetic theory needed to understand the phenomena that lead to EMC problems, the techniques used to alleviate the problems and the measurement techniques used to assess the problems.

The lecture starts with a discussion of the concepts associated with vector and scalar fields, and then review Maxwell's Equations and their solution in the form of electromagnetic waves. The structure and properties of electromagnetic waves are reviewed.

Concepts associated with inductive and radiative coupling between circuits will be discussed, and the boundary between the two coupling types is explored in terms of the separation of the coupled circuits as a fraction of the wavelength.

In view of the limited time available to cover this revision material, the approach during the lecture will be intuitive. The lecture notes contain a more rigourous mathematical treatment.

Introduction.

In order to understand the causes and solutions of Electromagnetic Compatibility problems it is necessary for the engineer to have a working knowledge of electromagnetics. This subject is traditionally unpopular with engineering students as many of the concepts involved seem to be abstract, and the mathematical formulations required to make the analysis tractable are often unfamiliar, not being seen in other branches of electrical engineering. Unfortunately, there are no short cuts to a proper understanding of the subject, however, much can be gained by an appreciation and application of the important results of electromagnetic theory. In this lecture I have attempted to distil the important results and concepts concerning electromagnetic waves and to use them to describe the propagation and coupling of waves in and around a system. The lecture attempts to lay a foundation for others on more specific topics of concern in the study of Electromagnetic Compatibility. For a more detailed treatment the reader should consult one of the many books on the subject. Two I have found to be readable and useful, and these I recommend to students and those who need a refresher course. They are the books by Christopoulos [1] and Liao [2].

1) Field Concepts and Basic Electromagnetics.

In dealing with electromagnetic interference and electromagnetic compatibility we are concerned with the generation, propagation and measurement of electric and magnetic fields. It is useful, therefore, at the outset to review the concepts associated with the physical phenomena described as fields.

Definition; If a physical phenomenon exists and can be measured over a volume of space, then the phenomenon can be described as a *field*. Graphical or other visual representations of the magnitude of the field are described as field plots.

One of the most common field plots in everyday use is a weather map. The isobars on the map are a contour plot of the atmospheric pressure at sea-level over part of the earth's surface - a two dimensional pressure field plot. Atmospheric pressure has a magnitude but no direction associated with it. The air can exert a pressure in any direction. The pressure field is thus a *scalar field.*

The isobars on the weather map are also an indication of the atmospheric wind direction at heights around 500m above the ground. As such, they are also a plot of the wind velocity field. As this field has both magnitude and direction, it is a *vector field.*

In electromagnetics we are concerned with both scalar and vector fields.

If a scalar field is differentiated with respect to position, the result is a vector field. For example, if the height information of land on a map is regarded as a plot of a scalar field, then the result of differentiation is the gradient of the land which has a direction associated with it - e.g. the land slopes to the north.

Electric Fields.

The concept of an electric field arises out of Coulomb's Law describing the observed vector force **F*** experienced by two point charges Q_1 Q_2 separated by a distance r as shown in Fig 1.1.

$$\mathbf{F} = (Q_1Q_2/4\pi\varepsilon r^2).\mathbf{a}_{12} \text{ Newtons}$$

The term ε is the permittivity of the medium surrounding the charges and is one of the electrical parameters of the medium. The term $\mathbf{a}_{12}$ is a unit vector in the direction of a straight line joining the charges.

(*Bold characters are used to denote vector quantities is used throughout this paper)

If one of the charges, say Q_2 is replaced by a vanishingly small test charge δq as shown in Fig 1.2, which does not perturb the charge Q_1 then an *electric force field* **E** due to the charge Q_1 can be defined as;

$$\mathbf{F} = \delta q\mathbf{E} \text{ Newtons}$$

and therefore

$$\mathbf{E} = (Q_1/4\pi\varepsilon r^2).\mathbf{a}_r \text{ Newtons/Coulomb}$$

The unit vector is in a radial direction from the charge Q_1 towards the test charge.
The electric force field can be measured at any point around Q_1 by measuring the force exerted on the test charge. It is a vector field, having both magnitude and direction and has the units Newtons/Coulomb. As the electric force field **E** is proportional to the magnitude of its source Q_1, the system is linear and the principle of superposition applies. The total electric force field at any point can be obtained by vector summation of all the contributions from all the charges in a system.

A further vector field quantity **D** the *electric flux density* is defined as;

$$\mathbf{D} = \varepsilon\mathbf{E} \text{ Coulombs/m}^2$$

or alternatively

$$\mathbf{D} = Q_1/4\pi r^2.\mathbf{a}_r \text{ Coulombs/m}^2$$

The electric flux density is independent of the medium surrounding the source charge.

Electric Potential.

In moving the test charge around in an electric force field an amount of work is performed. The amount of work required to move a test charge from some reference point to any point in the space surrounding the charge system is independent of the route taken to that point, and can be evaluated as the integral of the force experienced along the route by the test charge. Work is a scalar quantity, and hence a scalar work field can be plotted in the space around the charge system. This scalar field is termed the *electric potential field* and is the amount of work required to move one Coulomb of charge from the reference point to a given position in space. The natural units of electric potential are Joules/Coulomb. By definition;

$$1 \text{ Joule/Coulomb} = 1 \text{ Volt}$$

In elementary text books the reference point used to define the potential of a point charge is at infinity where the electric force field is zero. This gives a potential value V at a distance r from a point charge Q of;

$$V = Q/4\pi\varepsilon r \text{ Volts}$$

As the potential V is proportional to the charge Q, the principal of superposition applies, and the potential at a point due to a charge distribution can be evaluated by scalar summation of the potentials of the point charges comprising the distribution.

As the scalar electric potential field was evaluated by integration along a path through the vector electric force field, differentiation of the electric potential field V with respect to position yields the electric force field.

$$\mathbf{E} = -\text{grad } V$$

The units of the electric force field can thus be re-defined in terms of those of potential.

$$1 \text{ Newton/Coulomb} = 1 \text{ Volt/m}$$

where the gradient operator is used to differentiate the potential field with respect to position.

Conduction and Displacement Current.

A material with mobile charge is called a conductor. If an electric field is applied to such a material the charge moves under the influence of the field and a conduction current flows. In most materials which are homogeneous and linear, the relationship between the applied electric field **E**, the conductivity σ (Siemens/m) and the resultant conduction current density $\mathbf{J}_C$ is given by Ohm's Law.

$$\mathbf{J}_C = \sigma\mathbf{E} \text{ Amps/m}^2$$

The time derivative of the electric flux density $d\mathbf{D}/dt$ also has the units Amps/m^2. This is called the displacement current density $\mathbf{J}_D$.

In a general material with an applied electric field E the total current density **J** is.

$J = J_C + J_D = \sigma.E + \varepsilon.dE/dt$

For sinusoidal time variation at angular frequency ω this becomes
$J = (\sigma + j\omega\varepsilon)E$

Note that the displacement current is in time phase quadrature with the conduction current. The classification of materials into dielectrics or conductors depends on the ratio of the conduction current to the displacement current.

Magnetic Fields.

A parallel set of arguments can be used to *show the existence of magnetic fields* associated with the forces observed between conductors carrying electric currents. The *magnetic field* **H** has units **Amps/m**. The force field associated with magnetic phenomena is the *magnetic flux density* **B**;

$B = \mu H$ Tesla

Fig 1.3 shows the two short current carrying conductors or current elements each of length dl. The force F_2 experienced by conductor 2 is given by the vector product;

$F_2 = I_2 dl \times B_1$

where for direct currents B_1, the magnetic flux density set up by conductor 1, is given by the Biot-Savart Law. The term μ is the permeability of the surrounding medium the material constant associated with magnetic phenomena.

$B_1 = \mu.I_1 dl \times r_{12}/4\pi r^2$

The vector $I_1 dl$ represents the magnitude and direction of the short current element, the convention being that the vector direction is that of the current flow.

Faraday's Law of Electromagnetic Induction.

If a loop of wire of area A has a uniform magnetic flux density *B passing through it, the* total *magnetic flux* cutting the loop is Φ where;

$\Phi = |B|.A.\cos\theta$ Webers.

θ is the angle between a line along the flux vector direction and the axis of the loop. If the magnetic flux changes, a voltage V is induced around the loop;

$V = -d\Phi/dt$ Volts.

This is illustrated in Fig 1.4. Faraday's law is the basis of the operation of transformers and many other devices. It is also the cause of many spurious coupling problems in and around electronic circuits.

Statement of the EMI/EMC Problem.

The concepts described above should be familiar to any electrical/electronic engineering student. For the purposes of this lecture series we can state them as follows:

A system of moving charges gives rise to a set of time varying electric and magnetic fields. These fields will interact with other distant mobile charge giving rise to electric currents. If the interaction is intentional we call it a radio system, if not it's EMI.

2) Maxwell's Equations and Electromagnetic Waves.

The concepts described in section 1 are described fully by Maxwell's equations listed in vector differential form below.

$\mathrm{curl}E = -dB/dt$

$\mathrm{curl}H = J_c + dD/dt$

$\mathrm{div}E = \rho/\varepsilon$

$\mathrm{div}H = 0$

Here J_c is the conduction current density (Amps/m^2), and ρ is the charge density (Coulombs/m^3). These two terms jointly represent the moving charges referred to in section 1 as the sources of the electric and magnetic fields. The four differential equations can be solved to yield coupled wave equations for the electric and magnetic fields - i.e. an electromagnetic wave. Examination of Maxwell's equations shows that both field components must be present in the time varying case. If electric fields originate from electric charges, then a change in an electric field requires charge to move - an electric current, the source of a magnetic field, must therefore occur.

The conventional form of the solution assumes that the fields occupy a region of space well away from the field sources where both ρ and J_c are zero. The solution can then be configured in the form of a spherical wave propagating away from a source region of negligible dimensions as shown in Fig 2.1. In spherical co-ordinates, with the source at the origin and sinusoidal excitation, the field equations are;

$$\mathbf{E}_\theta = E_0 \exp(j\omega t).\exp(-j\beta r)/r.\ \mathbf{a}_\theta \quad V/m$$

$$\mathbf{H}_\phi = H_0 \exp(j\omega t).\exp(-j\beta r)/r.\ \mathbf{a}_\phi \quad A/m$$

Here ω is the angular frequency and β is the phase constant defined as;

$$\beta = 2\pi/\lambda$$

where λ is the wavelength and r is the distance from the source. The two exponential terms describe the temporal and spatial oscillatory behaviour associated with a sinusoidal wave. The fields decay at a rate inversely proportional to the distance from the source. The power flux density of the wave is given by the Poynting Vector **P**

$$\mathbf{P} = 1/2(\mathbf{E} \times \mathbf{H}^*) \text{ Watts/m}^2$$

where the asterisk * denotes the complex conjugate. The power flux density decays at a rate inversely proportional to the square of the distance from the source as the energy propagated by the wave spreads out. As the fields are orthogonal to the radial direction and to each other, the vector product of the fields gives a power flow in the radial direction.

The ratio of the electric field to the magnetic field is determined by the electrical parameters of the medium in which the wave is propagating. For a dielectric medium these are the permittivity ε and the permeability μ, and

$$\mathbf{E}/\mathbf{H} = Z = (\mu/\varepsilon)^{1/2}\ \Omega$$

Z is the intrinsic impedance of the medium. For free space (vacuum, ε_o, μ_o) the intrinsic impedance is 377Ω. The wave impedance can be used to give other expressions for the power flux density;

$$\mathbf{P} = E^2/Z\ \mathbf{a}_r = H^2.Z\ \mathbf{a}_r \quad W/m^2$$

The velocity v of the wave is also determined by the parameters of the medium, and solution of Maxwell's equations shows it to be given by;

$$v = 1/(\mu\ \varepsilon)^{1/2} \text{ m/sec.}$$

For free space the velocity is 3 x 10^8 m/sec. For most dielectric media, the permeability is that of free space and the intrinsic impedance and velocity are reduced by a factor equal to the square root of the relative permittivity (dielectric constant) of the medium ($\varepsilon = \varepsilon_o\varepsilon_r$).

If the wave propagates into a conducting medium, conduction currents are present as well as displacement currents. The behaviour of the wave in the medium depends on the ratio of conduction current to displacement current. In media that are poor conductors, for example most practical dielectrics, the conduction current is small ($\sigma << \omega\varepsilon$), and the properties of the wave are those of a perfect dielectric of the same permittivity with the addition of a small attenuation term. The wave equations become;

$$\mathbf{E}_\theta = E_0 \exp(j\omega t).\exp(-j\beta r).\exp(-\alpha r)/r.\ \mathbf{a}_\theta \quad V/m$$

$$\mathbf{H}_\phi = H_0 \exp(j\omega t).\exp(-j\beta r).\exp(-\alpha r)/r.\ \mathbf{a}_\phi \quad A/m$$

where α is the attenuation due to power dissipation given by;

$$\alpha = (\sigma/2)(\mu\varepsilon)^{1/2}$$

In the case of a good conductor such as a metal ($\sigma >> \omega\varepsilon$) the solution of Maxwell's equations results in a wave that is heavily attenuated with the wave impedance and velocity both substantially reduced. The phase constant and the attenuation factor are numerically equal and given by;

$$\alpha = \beta = (\pi f \mu \sigma)^{1/2}$$

where f is the frequency. The situation is illustrated in Fig 2.2 where it can be seen that the wave fields are attenuated to insignificant values in less than a wavelength. As the wave is propagating in a good conductor with predominantly conduction current, the field amplitude is proportional to the conduction current in the conductor. As the current decays rapidly with depth of penetration, a wave entering a conductor from another medium is confined to a surface layer of current. The depth of penetration of the surface layer is determined by the attenuation, and the skin depth δ is defined as that depth by which the wave has decayed to 1/e (0.37) of its surface value.

$$\delta = 1/\beta$$

The velocity is given by $\omega\delta$, and the wave impedance is complex and equal to $(1 + j)/\sigma\delta$.

For example if copper ($\sigma = 5.8 \times 10^7$S/m, $\mu = 4\pi \times 10^{-7}$ H/m) is considered at a frequency of 1MHz, the following results are obtained;

skin depth = 66μm

wave velocity = 414m/sec

wave impedance = $(1 + j) \times 2.6 \times 10^{-4}\Omega$

Compare these with a velocity of 3×10^{8}m/sec and impedance of 377Ω for free space.

3) The Radiation Mechanism.

The preceding section did not specify the source of the electromagnetic wave other than to state that a set of time varying charges are required. The wave description is also for a region well away from the wave source. In the region close to the wave source, the wave properties are modified by the proximity of the source, and the nature of coupling between circuits is determined by the distance between the circuits. Clearly, if the coupling is to be understood, the structure of the fields close to a radiation source must be studied.

In section 1 it was stated that the sources of electric and magnetic fields are electric charge ρ and conduction current density $\mathbf{J}_c$ (moving charge). These two sources are related by the continuity of current relationship.

$$\mathrm{div}\mathbf{J}_c = -d\rho/dt$$

As the divergence of the magnetic flux density is zero, a standard vector identity (div.(curlA) = 0) can be used to define the vector magnetic potential **A** such that:

$$\mathbf{B} = \mathrm{curl}\mathbf{A}$$

For direct currents the vector magnetic potential of a current element **Idl** can be derived from the Biot-Savart Law

$$\mathbf{A} = \mu.\mathbf{Idl}/4\pi r$$

where r is the distance from the current element to the observer.

For oscillating fields at some distance from the source retarded potentials must be used to account for the finite propagation time of field variations. These are of the form;

$$V = (Q/4\pi\varepsilon r).\exp(j\omega t - \beta r)$$

$$\mathbf{A} = (\mu.\mathbf{Idl}/4\pi r).\exp(jwt - \beta r)$$

The vector magnetic potential can be substituted into the Maxwell equation for the curl of E to give;

$$\mathrm{curl}(\mathbf{E} + d\mathbf{A}/dt) = 0$$

Again a standard vector identity

$$\mathrm{curl}(\mathrm{grad}\ V) = 0$$

can be used to set the Lorentz gauge condition for electromagnetic radiation;

$$\mathbf{E} + d\mathbf{A}/dt = -\,\mathrm{grad}\ V$$

In the static case (d/dt = 0) we get the relationship used to define electric potential in section 1. In the time varying case where the charge distribution is changing by means of conduction currents the electric field is sourced by both the scalar electric potential V (derived from the charge distribution) and the vector magnetic potential A (derived from the conduction current distribution). Thus from any current distribution, representing any electronic equipment, the electric and magnetic fields can be evaluated. It is often computationaly more convenient to evaluate the vector magnetic potential for a current distribution, thereby enabling the magnetic field to be evaluated. The electric field can then be found by application of the appropriate Maxwell equation.

In order to illustrate the complexity of the field structure close to a radiating source, the two elemental source types will be considered. The first is the elemental electric dipole or current element as shown in Fig 3.1. This comprises an isolated current filament of negligible assumed diameter and length dl which is small compared to the wavelength of the radiation and also to the distance from the dipole to the observer at P. The current is I and is assumed to be sinusoidally varying with angular frequency ω. Evaluation of the vector magnetic potential at P and thus the fields gives the following result. In spherical co-ordinates, the magnetic field has only a φ component while the electric field has both radial and θ directed components. Z is the intrinsic impedance of the medium surrounding the dipole.

$$H\phi = (I\,dl / 4\pi) \,.\, (1/r^2 - j\beta/r)\ \exp(j\omega t - j\beta r) \,.\, \mathrm{Sin}\theta \,.\, a\phi$$

$$E\theta = (I\,dl / 4\pi) \,.\, (j\omega\mu/r + Z/r^2 + 1/j\omega\varepsilon r^3)\ \exp(j\omega t - j\beta r) \,.\, \mathrm{Sin}\theta \,.\, a\theta$$

$$Er = (I\,dl / 4\pi) \,.\, (2Z/r^2 + 2/j\omega r^3)\ \exp(j\omega t - j\beta r) \,.\, \mathrm{Cos}\theta \,.\, ar$$

The second elemental source type is the elemental magnetic dipole or current loop as shown in Fig 3.2. In this case the loop of area A has dimensions small compared to the distance to the observer and the wavelength. The current loop has a magnetic field with radial and θ components and an φ directed electric field.

$$E\phi = (I\,A / 4\pi) \,.\, (jZ\beta/r^2 - Z\beta^2/r)\ \exp(j\omega t - j\beta r) \,.\, \mathrm{Sin}\theta \,.\, a\phi$$

$$H_\theta = -(I\ A / 4\pi) . (\beta^2/r - j\beta/r^2 - 1/r^3) \exp(j\omega t - j\beta r) . \sin\theta . a_\theta$$

$$H_r = (I\ A / 4\pi) . (j\beta/r^2 - j/r^3) \exp(j\omega t - j\beta r) . \cos\theta . a_r$$

For either source, the field equations can be divided into four sections. The first is a source term indicating that the radiated fields are proportional to the dipole size and current. This term is often refered to as the dipole moment. The second is a term describing the amplitude decay of the fields as a function of distance r. It is this term that is of most interest. The third is a general wave description of the fields, and the fourth is the directional properties of the fields.

Returning to the second term in the equations, it can be seen that the fields decay as inverse functions of the first three powers or the distance. At large distances from the source, the only significant term is the inverse distance term resulting in a wave of the type described in section 2. For either type of source, the inverse distance term is only present in the θ and ϕ directed components both of which are orthogonal to the radial direction of propagation. These field components transverse to the radial direction of propagation are refered to as the radiation fields of the sources. The radial field components have only second and third order inverse distance terms and are thus insignificant away from the source. Note also the $\cos\theta$ directional properties of the radial fields as opposed to the $\sin\theta$ directional properties of the transverse fields. Very close to the sources the inverse cube and inverse square terms are dominant. These are known as the induction fields. The boundary between the dominant radiation fields at large distances from the sources and the dominant induction fields close to the sources is not distinct. A guide can be obtained by equating the magnitudes of the radiation field and induction field terms. If the magnetic field of the electric dipole is used then;

$$1/r^2 = \beta/r$$

and

$$r = 1/\beta = \lambda/2\pi$$

Thus at distances less than about one sixth of a wavelength the induction field terms are dominant, and at greater distances the radiation field terms dominate. At distances greater than about one wavelength, the only significant terms are the radiation fields. Examination of the field equations in the radiation field region shows that they are identical to the spherical wave solutions to Maxwell's equations given in section 2 with the addition of a term expressing the directional properties of the dipole.

In the radiation field region the two source types are indistinguishable. There are no radial fields, and the ratio of the electric to magnetic fields is the intrinsic impedance of the surrounding medium. The two source types differ significantly in the induction field region where $1/r << 1/r^2, 1/r^3$. The differences can be best illustrated by examining the transverse wave impedances Z_T, defined as the ratio of the θ and ϕ directed field components for θ equal to $\pi/2$, for both source types. For the electric and magnetic dipoles respectively we have;

$$Z_{Te} = 1/j\omega\varepsilon r$$

$$Z_{Tm} = j\omega\mu r$$

In each case the impedance is reactive and dependent on both distance and frequency. Manipulation of the relationship between ω and λ, the wavelength, gives the transverse wave impedances in normalised wavelength terms;

$$Z_{Te} = -j(Z/2\pi)/(r/\lambda)$$

$$Z_{Tm} = j(2\pi Z).(r/\lambda)$$

The magnitudes of these impedances are plotted in Fig 3.3 for all values of (r/λ). Very close to the source the electric dipole has a relatively high transverse wave impedance indicating that the electric field is dominant. The magnetic dipole has a very low transverse wave impedance indicating that the magnetic field is dominant.

4) Radiation from Circuits and Equipment.

An electronic device comprises a large number of circuits, each of which may be carrying a time varying current and thus emitting radiation. At typical clock frequencies of 10MHz to 50MHz significant harmonics are present to at least 1GHz, a wavelength range of 30m to 300mm. As individual tracks and loops on circuit cards have dimensions in the range 10mm to 100mm, they can be considered as elemental radiators over most of that frequency range. Examination of the equations for the radiation from electric and magnetic dipoles indicates that the magnitudes of the radiation fields are proportional to the frequency and the square of the frequency respectively. Thus a circuit carrying a digital signal with the time waveform and frequency power spectrum shown in Fig 4.1 will radiate a spectrum modified by the frequency response of the radiation mechanism as shown in Fig 4.2. The highest radiated fields are associated with spectral energy at frequencies beyond those normally considered important for the operation of the circuit. With clock frequencies in the 10MHz to 50MHz region, the radiated energy is predominantly in the VHF/UHF frequency range.

A typical circuit which is small compared to the wavelength under consideration may comprise both types of elemental source. Consider the circuit shown in Fig4.3. If it is regarded as an isolated circuit on a circuit card, the circuit diagram is representative of both the electrical equivalent circuit and its physical layout. Electric currents that circulate around the circuit loop clearly give rise to magnetic dipole radiation. The presence of the circuit tracks on the card means that parasitic capacitance exists between the tracks. The finite load and source impedances Z_S, Z_L give rise to a potential between the tracks, and thus an oscillating current component flows in the circuit as well as a circulating one as shown in Fig 4.3. This oscillating current component gives rise to electric dipole radiation. The axis of the electric dipole is orthogonal to that of the magnetic dipole as defined in Figs 3.1 and 3.2 and lies in the plane of the loop. If impedance loading is present in other arms of the circuit loop, a second electric dipole, orthogonal to the first, with its axis in the plane of the loop is present. The total radiation from the circuit is the sum of that from the three elemental dipoles.

The relative magnitudes of the individual dipole moments depend on the loop area, its linear dimensions and the impedance loading around the loop. In general high impedance circuits have predominantly electric dipole behaviour while low impedance circuits behave predominantly as magnetic dipoles.

As a circuit card has many loops, the overall dipole moments of a card, small compared to a wavelength, could be derived by summation of all the individual dipole moments at any particular frequency. In a general electronic device, that has cards in more than one orientation but is still small compared to a wavelength, a set of three orthogonal electric dipoles and three orthogonal magnetic dipoles for each frequency can be used as a complete description of the device as a radiation source. Clearly such a description is too unwieldy for general use. It does, however, serve to illustrate the process of radiation from devices.

In the case of larger equipment, the radiation process is more complex. If a standard equipment enclosure size is considered, say 19in (0.5m) along with a power cable of say 2m length, then the equipment can no longer be considered to be electrically small. In the frequency range up to 1GHz, the equipment and its power cable with a maximum dimension of 2.5m is up to eight wavelengths in extent.

Clearly, equipment of this size cannot be considered to be a set of co-located elemental dipoles, and must be considered as a distributed source. The principle of superposition applies, and the structure of the equipment can be divided into a set of elemental dipoles, each with its own current. As the current on one element radiates a set of fields that impinge on all the other elements, this mutual coupling between the elements must be accounted for when evaluating the currents on the equipment. There exist a number of numerical methods which enable such solutions. The most useful one is the method of moments developed by Harrington , the computer code Numerical Electromagnetics Code (NEC) being the most common implementation of this method. These are addressed in the next lecture.

Such equipment is also likely to be electrically screened and so any radiation from the equipment will arise from currents existing on the outside of the screened equipment enclosure and on the power cable, excited by fields existing across apertures in the screen. The accurate prediction of the radiated fields under these circumstances is not usually possible as such predictions rely on a complete knowledge of the current distribution on all parts of the equipment at each frequency. As these currents are not generally part of the design requirement of the equipment, and steps may have been taken to prevent them by screening and filtering, such detailed knowledge is improbable. The values of these currents and hence the radiated fields also relies critically on the position and orientation of the equipment and its cable. The ramifications of this are discussed in the paper on measurement environments and testing.

5. Reception of Electromagnetic Energy and Reciprocity.

So far I have concentrated on the generation and propagation of electromagnetic waves. In this section the reception of waves is briefly considered. If any source victim pair is considered as a pair of antennas, one transmitting and one receiving, then the operation of the receiving antenna would be related to its operation as a transmitting antenna by the principle of reciprocity. This is a convenient ploy for antenna engineers because it is generally easier to describe the operation of an antenna in transmission rather than in reception. In the reception case the currents induced on an antenna by an incident electromagnetic wave must be deduced. Reciprocity avoids this difficult task. From the viewpoint of the interference source victim pair, the *electromagnetic* reception properties of a victim can be deduced from the same properties of the victim considered as a source. Thus if an equipment has low emissions as a result of good electromagnetic design, i.e. adequate screening and filtering and good circuit layout avoiding large current loops with proper decoupling, then the immunity of the equipment is likely to be high as the inefficient transmitting antenna is also an inefficient receiver. This only applies to the electromagnetic aspects of the design. If the low emissions are a result of a low energy logic family being used in the equipment, for example low voltage CMOS, without thoughtful electromagnetic design, then the immunity may be poor. The amount of energy required to disrupt a low energy logic family is much less than that required to disrupt a high energy logic family such as TTL.

6. Capacitive and Inductive Coupling.

The sections above have concentrated on the generation of electromagnetic radiation from equipment. The radiation process has been described in terms of elemental sources. Examination of the field equations for the elemental sources reveals that the radiation fields are dominant at distances from the source greater than $\lambda/2\pi$, and that they are the only significant fields at distances in excess of a wavelength. In many cases the potential for interference is between equipments that are adjacent to each other, with a typical separation in the order of 1m. If the $\lambda/2\pi$ criterion is applied for a 1m separation then coupling at frequencies below approximately 50MHz is by the induction fields. These fields decay with distance as the second or third inverse power, and can be seen to be equivalent to the quasi-static fields associated with low frequency analysis of inductance and capacitance.

The lower the frequency, the larger the equipment separation that can be analysed by the quasi-static approximation. At such small separations in wavelength terms, the interference source equipment and the victim equipment cannot be regarded as separate electrically. The coupling of interference between them is by mutual capacitance and inductance, and the presence of say the victim equipment alters the parasitic capacitances and inductances of the source equipment and vice versa. Under these circumstances there is also likely to be some direct electrical connection between the two equipments. This may take the form of a connection through power cables sharing a common phase in the electrical mains, or through data/signal inter-connections. The situation is illustrated in Fig 6.1 At these low frequencies the distinction between conducted interference, cross-talk and inductive/capacitive coupling becomes difficult to draw, as indicated in Fig 6.1 where an interference current path relies partly on capacitive coupling and partly on direct conduction. Aspects of shielding grounding and crosstalk are discussed in more detail in the following lectures.

The distinction between the low frequency quasi-static coupling and the high frequency radiative coupling is clear. An appreciation of the mechanisms operating in the intermediate frequency range is more difficult to come by. The quasi-static analysis works well for equipment separations up to around $\lambda/10$ and adequately for separations up to $\lambda/2\pi$. For separations of λ or more the radiation model is appropriate. The problem is also illustrated by the wave impedance relationship shown in Fig3.3. The quasi-static analysis corresponds to the purely reactive impedance region, whereas the radiation region corresponds to the 377Ω region. In the intermediate region the wave impedance is complex, and the field decay function is depends on more than one inverse power of distance. If inter-connecting cables exist between the interference source and victim. The energy coupling may be best understood by considering propagation along the non-uniform transmission line that exists between the equipments.

Conclusions.

In this lecture I have attempted to highlight the more important aspects of the generation, propagation and coupling of electromagnetic energy. This is far too large a topic to cover in a one hour lecture, and only the surface can be skimmed. Some appreciation of the basics of this subject is essential for any engineer designing or operating equipment today's complex electromagnetic environment. The lecture has been written with the aim of providing the tools necessary to understand the lectures that follow on numerical techniques, cables and crosstalk, shielding, design and measurements.

References.

[1] C. Christopoulos "An Introduction to Applied Electromagnetism" Wiley Student series in Electronic and Electrical Engineering. John Wiley1990. ISBN 0 471 92761 9.

[2] S.Y. Liao "Microwave Devices and Circuits" Prentice-Hall 1980. ISBN 0 13 581207 0.

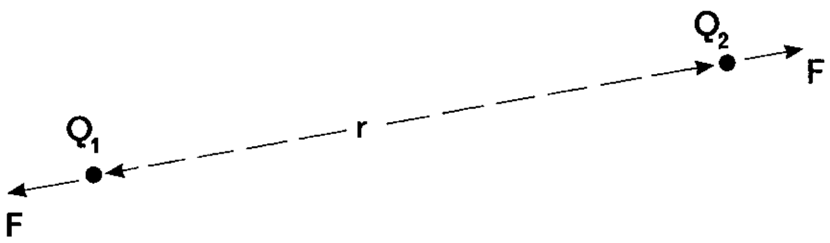

Figure 1.1
Illustration of the Forces between Two Point Charges.

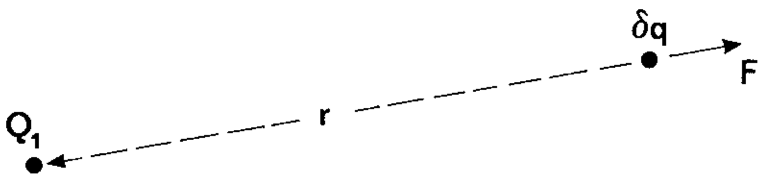

Figure 1.2
Derivation of the Electric Field of a Point Charge.

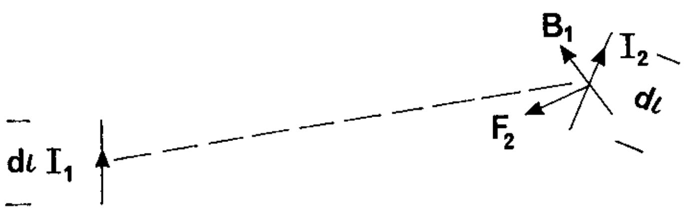

Figure 1.3
Illustration of the Magnetic Force between Two Current Elements.

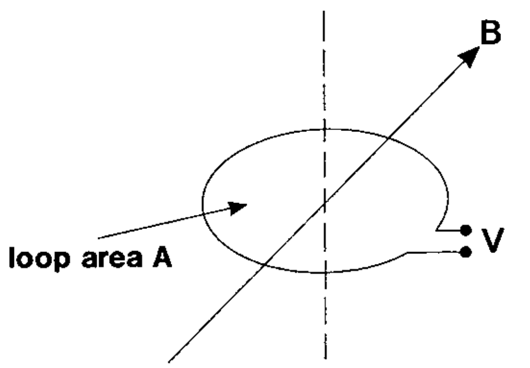

Figure 1.4
Illustration of Faraday's Law of Induction.

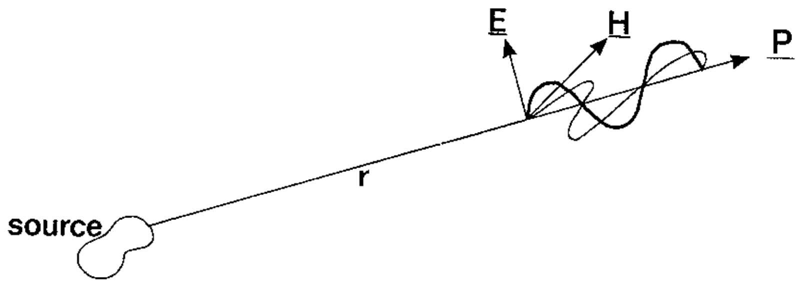

Figure 2.1
A Spherical Electromagnetic Wavefront.

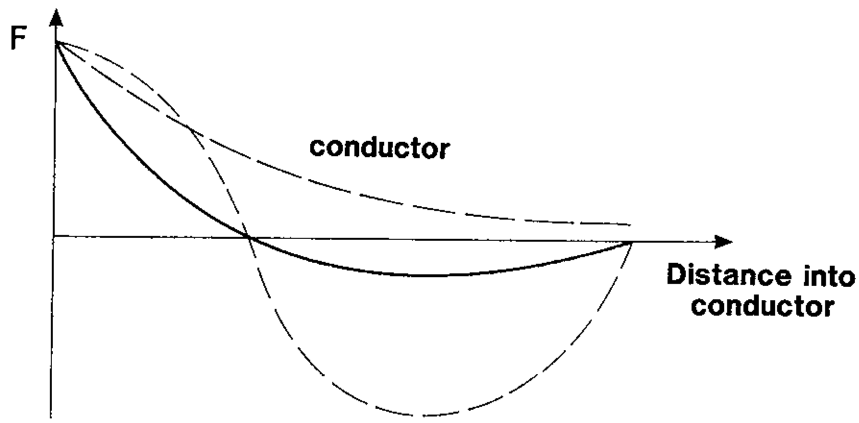

Figure 2.2
Wave Penetration into a Conductor.

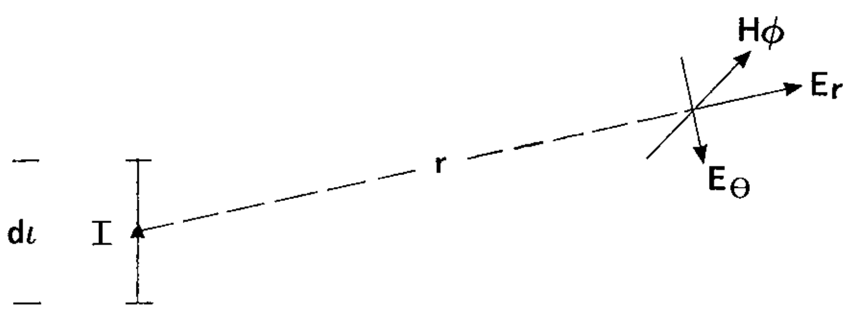

Figure 3.1
Field Components of an Electric Dipole.

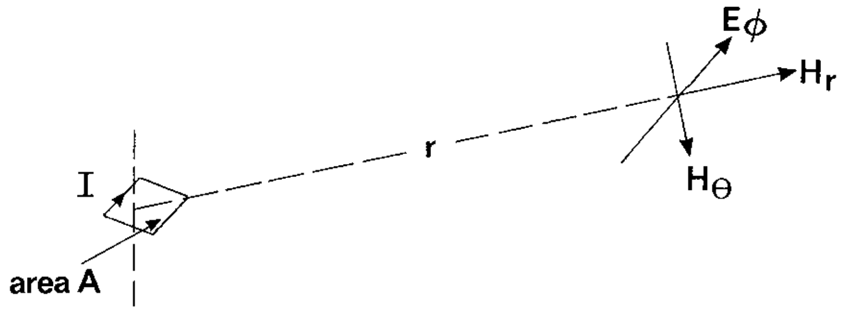

Figure 3.2
Field Components of a Magnetic Dipole.

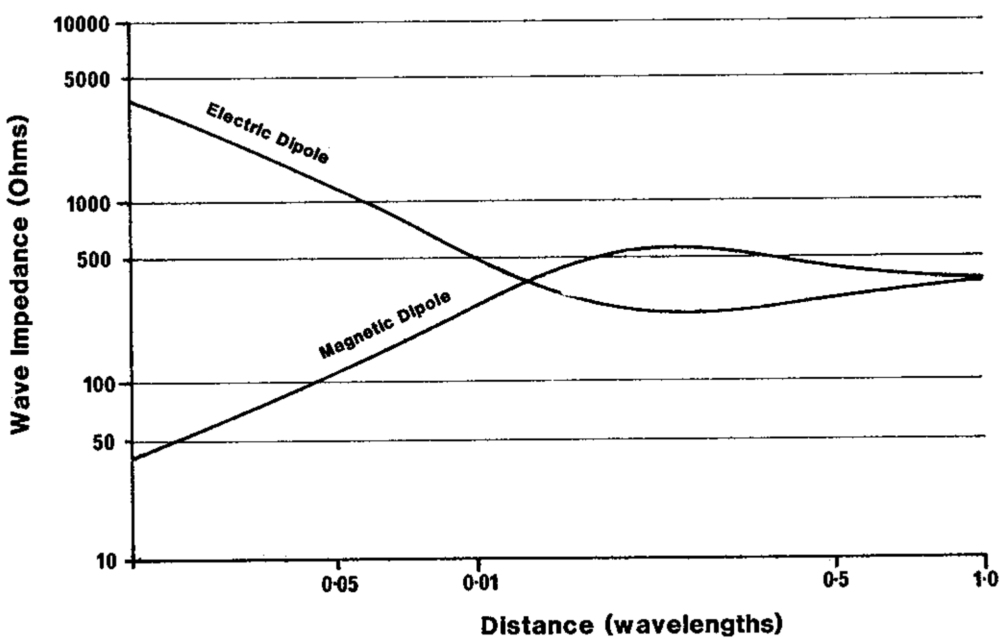

Figure 3.3
Plot of the Transverse Wave Impedances of Electric and Magnetic Dipoles.

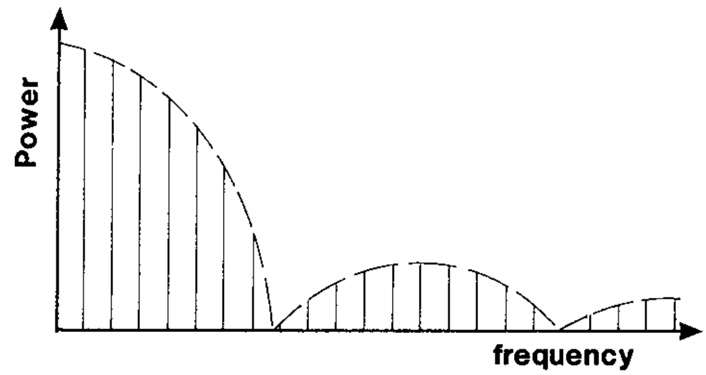

Figure 4.1
Power Spectrum of a Digital Pulse Train.

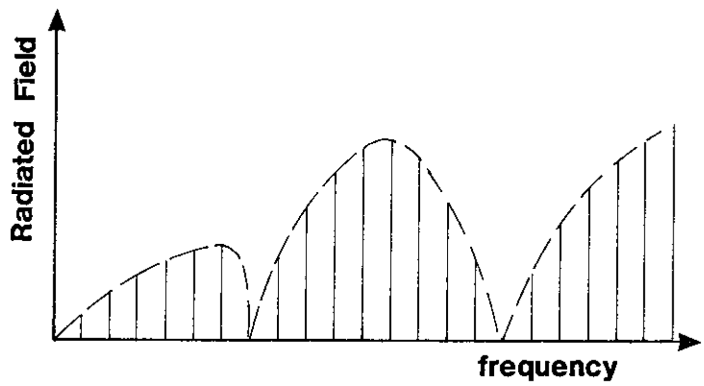

Figure 4.2
Radiated Spectrum of a Digital Pulse Train.

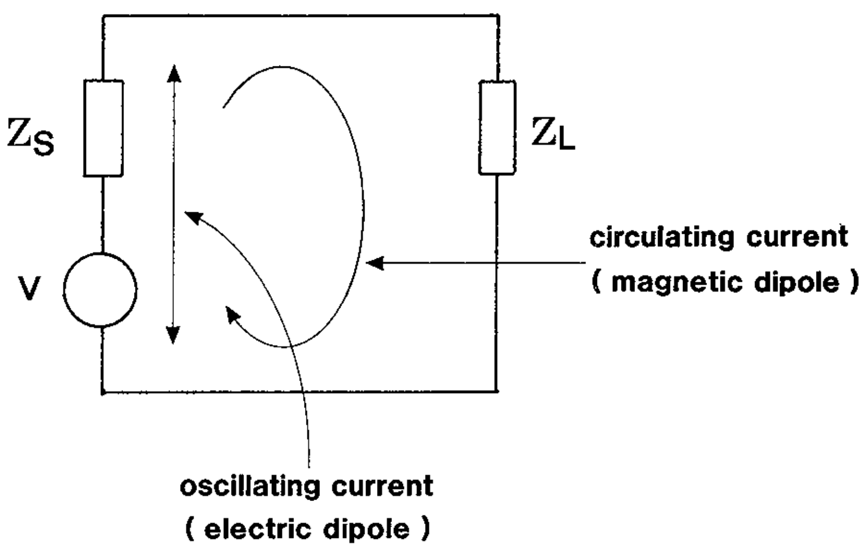

Figure 4.3
Radiating Currents in an Electric Circuit.

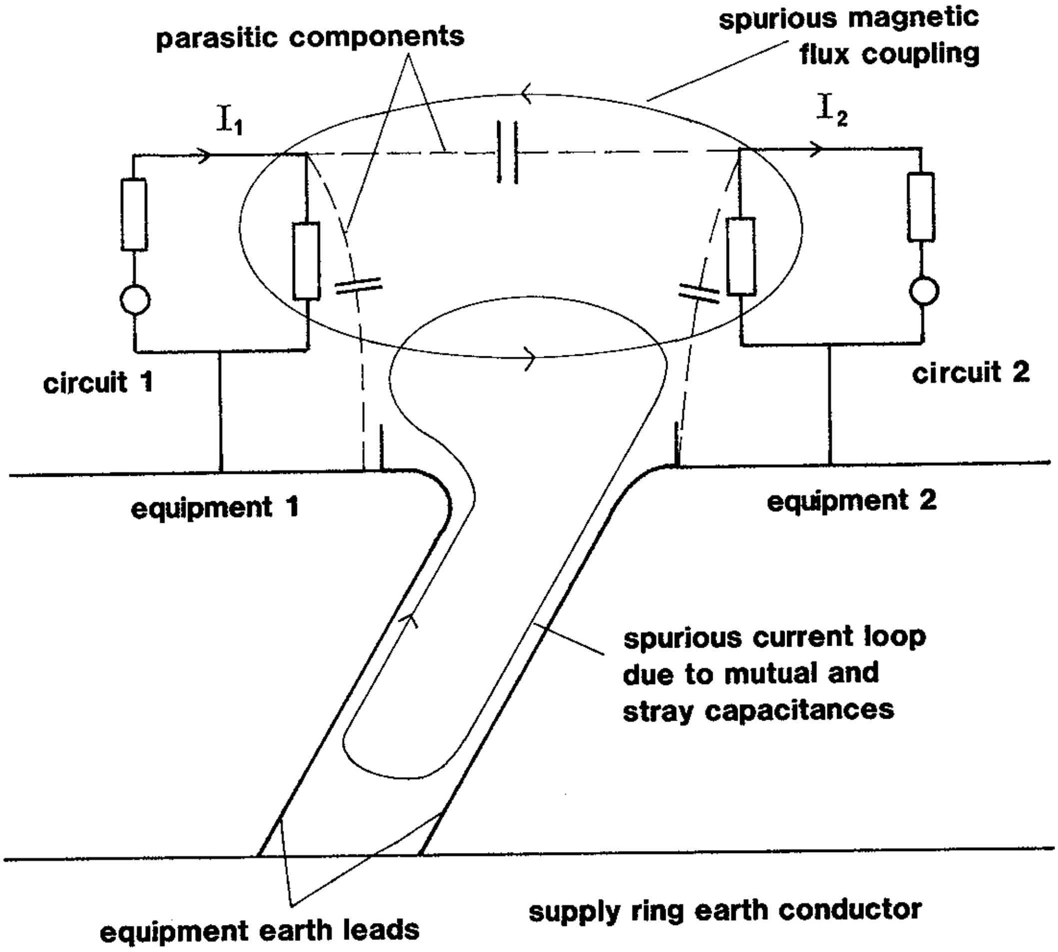

Figure 6.1
Schematic Illustration of Capacitive and Inductive Coupling between Circuits due to Parasitic Components.

Numerical Analysis and Modelling Techniques

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1. INTRODUCTION

This paper deals with the modern computational tools which can be exploited in the complex task of assuring the electromagnetic compatibility of modern avionic/weapon systems. The life-cycle of aerospace systems ranges from the conceptual stage, to initial design, to prototype test and development, to production design and test, field operation, major up-date or mid-life improvement, interspersed with retrofit installation of new systems. In many countries off-the-shelf systems are purchased and require integration into existing airframes.

It has often been stated that electromagnetic compatibility or EMC must be designed into systems. This is a truism because invariably corrective action is costly. Often deficiencies in EMC can be associated with serious operational limitations and loss of life. However the attempt to design for EMC involves a requirement for the awareness of the total set of possible undesired interactions, and their consequences as intersystem or intrasystem events in an operational electromagnetic environment. The development of this awareness is an awesome task. It requires knowledge of the electromagnetic environment, the characteristics of antennas in their sited locations, the EMC characteristics of avionic, electrical and instrumentation and control systems and a knowledge of the coupling modes which can be relevant.

For military systems, a coherent methodology is implied by the requirements of specifications such as MIL-E-6051 [1] and MIL-HDBK-335 [2]. These in turn call up specifications and requirements governing equipment and components. The former requires the implementation of an overall **EMC CONTROL PLAN** which eventually would culminate in the development and execution of an **EMC TEST PLAN**. The intention is that the test plan when properly executed, would define all likely operational limitations and help define and verify corrective action. It is also appropriate to reflect on what information the EMC engineer might have about the environment, equipment, antennas, aircraft layout and equiment interconnection during the life-cycle of an aerospace system. Understandably, the extent of the information available would govern the type of analysis which would be suitable to develop the required awareness of likely interactions. Thus the progressive development of a data base of relevant information plays an important part in the planning for EMC.

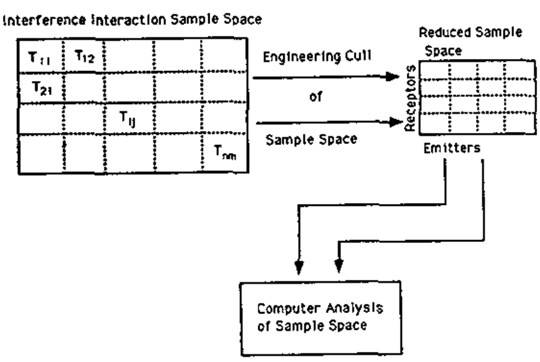

Fig. 1 Interference Interaction Matrix

In AGARD Lecture Series 116, Spina[3] proposed a visualization of all potential interference interactions in terms of an interaction sample space between emitters and receptors in terms of a global transfer function matrix $\mathbf{T}_{ij}$ as shown in Figure 1. The transfer function elements arise from all the possible signal coupling paths which can occur between systems as represented in Figure 2. The representation of these possible coupling paths is the subject of some of the other papers in this Lecture Series. The objective of an EMC program must be, first to apply engineering judgment in order to reduce this matrix to a manageable size, incorporating those interactions whose severity must be monitored and then address these by computer analysis techniques for further evaluation and reduction.

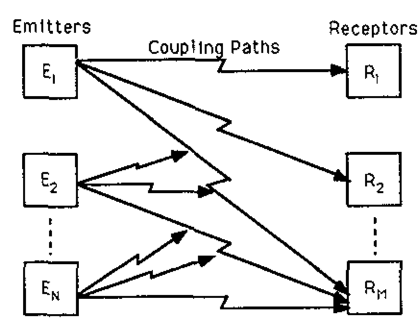

Fig. 2 Intrasystem EMC Setting

Two system-level analysis programs are discussed in this paper. The Intrasystem Electromagnetic Compatibility Analysis Program(IEMCAP)[4] is a comprehensive system-level computer analysis program designed to assist in the quantification process. A general description is presented to make readers aware of its existence and some of its features. A smaller interactive program called AAPG: Antenna-to-Antenna Propagation with Graphics is described in more detail. Both of these foster the progressive development of the data base which is essential to any comprehensive approach.

In addition to the modelling required for system analysis purposes, it is necessary to be able to determine radiated field levels or the coupling between antennas more precisely. For these purposes computational techniques based on the electric field integral equation or Transmission Line Modelling(TLM), as well as ray-optical methods based on the Geometric Theory of Diffraction, can be used. Salient features of the former two methods are presented below.

2. INTRASYSTEM ANALYSIS METHODS

About two decades ago, the United States Air Force began a co-ordinated effort to address the problem of intrasystem electromagnetic compatibility. This development effort was entitled the Intrasystem Analysis Program(IAP). This development effort and the programs, such as IEMCAP, which were developed under its sponsorship are described in Spina's paper[3]. Not all users are able to take advantage of the size and scope of IEMCAP, and subsequent to its promulgation, smaller specialized computer codes, such as AAPG were developed. The AAPG code is a computer program originally developed in 1978 at Concordia University in Montreal under the sponsorship of the Canadian Department of Defence. Since 1980, the Defence Research Establishment Ottawa (DREO), USAF Rome Air Development Centre (RADC), the US Department of Defence Electromagnetic Compatibility Analysis Centre (ECAC) and Concordia University have pursued AAPG developments on a joint basis.

2.1 Summary of IEMCAP Features

IEMCAP was designed to satisfy the primary requirements for EMC analysis. The first requirement of an EMC analysis program is that it should bring all relevant EMC factors into focus. Thus as indicated in Figure 2, all potential emitters, receptors and coupling paths must be identified and characterized to a minimum level of detail. Geometry considerations and operational scenarios need to be specified.

Another important feature deals with an interactive capability to allow various designs to be compared as to their EMC impact. This allows options to be examined before design decisions are firm. Examples of this are antenna placement, cable and wire bundling and routing, and frequency assignments.

The program should include the capability to vary equipment EMC specifications according to the specific electromagnetic environment in which it is to be used. This type of analysis requires considerable technical maturity and experience on the part of the analyst to keep within safety margins when compromises are made.

Requests for waivers from a system EMC specification are common in a new weapon system design. Having a waiver analysis capability provides the system planner with a useful tool because it removes uncertainties from ad hoc waiver decisions by providing some ability to assess its impact on the rest of the system.

An EMC analysis program should have the ability to predict EMC problems to the extent that a comprehensive EMC profile is produced which can be used to design effective test programs. Then when the transfer functions involve similar mechanisms, the testing of a dominant few makes for an efficient test program which validates the overall prediction process.

The final essential feature stressed in IEMCAP is the intrasystem data base. This contains electrical data which describes equipment EMC characteristics, wire routing information, the system geometry, electromagnetic apertures, and equipment and antenna placement. This information forms the baseline from which future EMC analyses can be made in a perceptive and effective manner as changes occur or the information is refined.

2.1.1 IEMCAP Capabilities

A complete description of the capabilities, models and operation of IEMCAP is provided by Capraro[4]. The program provides an EMC analysis methodology for a system, whether it be ground based, airborne, or a space/missile system. The basic medium for modelling signals is the frequency domain. To predict interference for a set of receptors due to a set of emitters in the system, each emitter's characteristics are modelled by its power output, tuned frequency, emission spectrum in the vicinity of the tuned frequency, and spurious-emission levels and frequencies. The model assumes that harmonic spurious output levels can be approximated by one or more straight-line segments. Spurious output frequencies are determined by the user or as harmonics of the tuned frequency, or generated by the computer code. The illustrations presented below for AAPG also apply to IEMCAP modelling.

The receptor is characterized by its sensitivity, tuned frequency, selectivity curve, spurious response levels, and spurious frequencies. It is assumed that the spurious response levels can be approximated by one or more straight lines. Spurious response frequencies are generated by the code or the user must determine these frequencies, external to the program, by using available techniques, such as those applicable to the superheterodyne conversion process.

Antenna gains are determined by preprogrammed equations for low-gain types and medium and high gain are represented by multi-level patterns, in which each level is specified by a gain and associated azimuth and elevation beamwidths. Provision is made for three discrete gain levels.

Various models of coupling or transfer functions are included in the program. Filter models used are single tuned, transformer coupled, Butterworth tuned, low and high pass, bandpass, and band reject. The filter transfer models calculate the "insertion loss" in dB provided by a filter at a given frequency, i.e., the reduction in delivered power due to insertion of a filter.

There are two antenna-to-antenna coupling models available. For ground systems, the propagation model is a simplified theoretical ground-wave model which assumes a smooth-earth surface with a 4/3 earth radius accounting for atmospheric refraction. An intravehicular propagation model calculates the propagation loss associated with an electromagnetic coupling path when both emitter and receptor are located on the same aircraft or spacecraft. The power received is related to the power transmitted, free-space transmission(Friis equation[8]), and a shading factor due to the presence of the vehicle whose bulk may be interposed in the region between emitter and receptor.

Environmental electromagnetic field interaction with the system wiring is determined. External fields enter a vehicle through dielectric apertures in the system's skin and couple onto wires immediately adjacent. The coupled RF energy is a function of the aperture size and location. A transmission-line model is then used to compute the currents induced in the wire loads. These models are described in this Lecture Series by Prof. Clayton Paul. For ground systems, artificial apertures are required for determining certain field-to-wire conditions.

Wire coupling between wires in a common bundle considers capacitive coupling due to the interwire capacitance and inductive coupling due to the mutual inductances between the wires. Relatively complex configurations, e.g. shielded(single or double), unshielded, twisted pair, balanced or unbalanced, can be handled.

The equipment case model treats each case as though it were a dipole. The source model assumes a fall off of $(1/R)^3$, where R is the distance between cases, for both the electric and magnetic fields.

Interference is determined by IEMCAP as computed point and integrated EMI margins. These represent the ratios of coupled power to receptor susceptibility at an individual frequency and acrosss a broad frequency range, respectively.

The data size limits of IEMCAP are shown in Table 1.

Table 1
IEMCAP Size Limits

Equipments	40
Ports per Equipment	15
Total Ports(40x15)	600
Apertures	10
Antennas	50
Filters	20
Wire Bundles	140
Segments/Bundle	10
Wire/Bundle	280

Thus, using IEMCAP, the matrix of EMI margins can be analyzed to obtain the desired reduced sample space of interactions for further monitoring.

2.1.2 Present Status of IEMCAP

The computer program is now distributed by the Data & Analysis Center for Software, operated for the Rome Air Development Center by Kaman Sciences Corporation. A graphics mode has been added to display emitter and receptor spectra, antenna locations on the simplified aircraft model and to produce frequency plots of a received signal for an emitter/receptor pair.

Work on improvement of the models continues. A recent change to the field-to-wire coupling model has been described by Brock et al. [5]. A new receptor model development is described by Capraro et al. [6]. Current release is under Version 6.0. The program consists of some 16,000 lines of ANSI FORTRAN code and has been installed on a variety of mainframe systems. Kaman Sciences have also developed a PC version which requires a 32-bit co-processor and 2 Mbytes of RAM.

2.2 The AAPG Computer Code

AAPG has its roots in the IEMCAP analysis formalism. It uses a similar representation for emitters and receptors, a more complex aircraft model representation and similar ray-optical coupling paths between antennas. It also uses Geometric Theory of Diffraction formulations to calculate the propagation loss along these paths. For each of the interactions between a transmitter and a receiver via their respective antennas, AAPG computes the EMI point margin. Its considerable advantage lies in its interactive graphics capability which is used to present the results of all its operations in formats which make visible the constituent components of the EMI margin. A description of AAPG was presented[7] during AGARD Lecture Series 116.

The version in wide use at present is Version 07. The description below relates primarily to this release. Later, the special features of Version 09 are also described. This version is presently undergoing evaluation. Both versions are designed to operate with a high resolution graphics terminal such as the Tektronix 4014 and a hardcopy unit. The code is written in FORTRAN 77 and Version 07 occupies approximately 200k words of memory. With overlay techniques it has been installed on small microcomputers. It has been installed also on IBM personal computers driving displays with TEK4014 emulation. PC versions for VGA displays are expected to be available during 1991.

2.2.1 The Structure of AAPG

AAPG fulfils two fundamental objectives: to accurately compute EMI margins and to present the results of these computations in a concise, visual manner. It consists of two principal software modules:

a) The Electromagnetic Compatibility Computation System (EMCCS); and

b) The Graphical Data Management System (GDMS).

The EMCCS, which uses analytic geometry and the Geometrical Theory of Diffraction (GTD), determines geodesic coupling paths and their losses and fills a mass-storage data file (MSDF) with the results of its calculations.

Subsequently, the GDMS accesses portions of the MSDF to display those particular aspects of the analysis with which the user may be concerned. The GDMS has four distinct graphical modules: Frequency Coincidence, Antenna Location Display, Propagation Path or EMI Margin Display, and Antenna Position Input. These are illustrated below.

It will be seen that from a minimal data base, the T_{ij} interactions between emitters and receptors can be examined in turn. The more meaningful interactions are viewed and kept to form a reduced set from which the more critical ones can be identified for corrective action or evaluation by ground and flight tests.

The ease of relocation of antenna systems and the re-evaluation of the interference margin on a real-time basis provides the user with a dynamic approach to EMC analysis that is useful for tolerance studies. The execution of the program requires the preparation of an input data file as a series of ordered card images.

2.2.2 Input Data File

The Input Data File contains a geometrical description of the aircraft and the specifications of the on-board transmitting and receiving equipment.

The aircraft model as shown in Figure 5, consists of a cylinder representing the fuselage with a cone at one end. The conical portion may be truncated with a flat bottom below. Planar segments are utilized to model the wings and stabilizers. The user must have access to dimensioned three-view drawings in order to create such a model.

Immediately following the aircraft geometrical data in the input file, AAPG expects to read the electronic equipment data. The user must group the electronic equipment in transmitting/receiving subsystems called "TRS".

Each TRS is assigned a name, a security classification and will contain technical information on each subsystem such as:

a) Power output for fundamental and up to five higher order harmonics and receiver sensitivity threshold,

b) Tuning range,

c) Upper/lower frequency "roll-off" rates,

d) Identification of the antenna(s) which are used,

e) Antenna gain and pattern.

The antenna pattern that is used is a two-level "keyhole" pattern with main beam and sidelobe levels specified.

To facilitate the preparation of the input data set, a program called AAPG.DI is available. It prompts the insertion of the correct information in fields shown on a PC screen in inverse video form.

2.2.3 The Computation Parameters

The main objective of the software algorithms is to accurately evaluate the magnitude of potential antenna-to-antenna coupled interference and compute the EMI margin.

The numerical value (in dB) of the EMI Margin is calculated from the following equation.

$$M = [P_t - L_{ct} + G_t] - TL + [G_R - L_{CR} - S_R]$$

where M = EMI Margin, in dB

P_t = Transmitter power, in dBm

L_{ct} = Transmitter-to-antenna cabling loss, in dB

G_t = Transmitter antenna gain in the coupling path direction, in dBi

TL = Transmission loss, in dB

G_R = Receiver antenna gain in the coupling path direction, in dBi

L_{CR} = Receiver-to-antenna cabling loss, in dB

S_R = Receiver sensitivity, in dBm

These EMI margin components, computed at the frequency of greatest interference (FGI), can be viewed as three distinct blocks. The first and third blocks of data concern the emitter and receptor information which is entered as part of the TRS data in the input file.

The second block or Transmission Loss information, represents the data computed by the AAPG code. The TL factor contains the loss incurred along the coupling path between the two antennas and may be composed of any one or a combination of the following:

a) Free space loss,

b) Surface Shading loss, and

c) Edge shading loss.

The free space loss (L_{FS}) is calculated using the Friis [8] formula for point to point spreading loss:

$$L_{FS} = -20 \text{ Log}_{10} (\lambda/4\pi D)$$

where λ = wavelength of the interference frequency in meters

D = distance along the path in meters.

The surface shading loss (L_{CS}), representing the loss incurred in wave propagation over a curved-surface or a "creeping wave" loss, is evaluated using an approximation based on the work of Hasserjian and Ishimaru [9]:

$$L_{CS} = A/(\eta A + \varepsilon)$$

and $A = \rho_s . \theta_s / 2\pi/\lambda D$

ρ_s = geometric mean of the radii of the spiral end points in meters

θ_s = angle spanned by the coupling path spiral

D = spiral distance, in meters

and where

η = .005478 for $A < 26$
.003340 for $A \geq 26$

ε = .5083 for $A < 26$
.5621 for $A \geq 26$.

Finally the edge shading loss (L_{KE}) is evaluated using the formulation for edge diffraction presented by Kouyoumjian and Pathak [10]:

$$L_{KE} = 20 \log_{10}((D_1+D_2)\pm(D_3+D_4))/(r+s)/rs\varepsilon^{(-j2\pi(r+s))}$$

where r and s are distances to/from the source/receiver to the diffraction point, and the D_i's are the complex diffraction coefficients, described at length in reference [10].

It can be appreciated that the error-free incorporation of these path loss computations into the geodesic algorithms

It can be appreciated that the error-free incorporation of these path loss computations into the geodesic algorithms is always problematic. It will be seen that the visualization provided by the graphics displays helps in the verification of the values and the appreciation of their magnitude.

2.2.4 Use of AAPG

Once the EMCCS has read the input file and completed its calculations and stored the pertinent information in the MSDF, control is given to the GDMS. From this point on, the user is guided by a series of menu-driven displays. The GDMS is controlled by a Display Manager with a menu showing the four available options to be selected for display. Each of these has its own menu list of options which are called up by simple mnemonic commands[7].

The Frequency Coincidence Display Package is invariably the first that is entered. It can be used to generate five types of information, all of which pertain to specific frequency-coincident receiver/transmitter pairs. The first of these is a summary of coincidence data (Figure 3) listing the code numbers of all transmitters which overlap in frequency with each receiver. This code serves as an index to the other selected displays within AAPG. A line frequency-coincidence plot is available to show the frequency overlap of a receiver with all coincident transmitters at their operating frequency bands and harmonics. A power-versus-frequency plot is also available (Figure 4), which overlays the transmitter emission spectrum and the receiver sensitivity threshold for a particular receiver/transmitter pair. This shows at a glance, the power level differences which exist and which must be attenuated in any coupling paths between the systems. It is common practice to represent receiver spurious responses as separate receivers and thus make the data base complete. Cursor readout of power levels and frequencies are available for finer analysis or for flight test planning purposes.

The Antenna Location Display package provides a means to view the location and pattern for each antenna relative to the aircraft model. This data (Figure 5) comprise a useful tool to validate the antenna information as to the antenna location as butt line, water line, and fuselage station on the model and the modelled antenna patterns.

The antenna pattern can also be verified by accessing the antenna pattern diagram, Figure 6, where all pertinent information on each antenna may be examined. AAPG also provides the capability to view the platform using full graphic options i.e., side, front, top views individually or on a single display, which permits viewing the aircraft from any elevation or azimuth angle. A close-up view option is also available for coupling path displays.

```
UNCLASSIFIED        FREQUENCY COINCIDENCE DATA SUMMARY        TRACKER
                    ----------------------------------        1/19/91

RCVR        XMTRS  RCVR        XMTRS  RCVR        XMTRS  RCVR        XMTRS

      ARA25              ARC27A             ARN6/44             APX77

 1 DFRX             5 UHFRX            9 ADF6R            13 IDENRX
   1.01 HFSST         5.01 UHFTX         9.02 HFAMT          13.01 UHFTX
   1.02 HFAMT         5.02 VHFAMT                            13.02 TCNTX
   1.03 UHFTX         5.03 VHFFMT          APS504            13.03 IDENTX
   1.04 VHFAMT
   1.05 VHFFMT      6 UHFIM           10 SCHRDR                ARN508
                      6.01 UHFTX        10.01 SCHRDT
      APN 503         6.02 VHFAMT       10.02 RDALTT      14 VORLOC
                      6.03 VHFFMT                            14.01 HFSST
 2 DOPLR                                   APN22             14.02 HFAMT
   2.01 DOPLT            ARN6/44                             14.03 VHFAMT
   2.02 RDALTT                        11 RDALTR
                    7 DF44R1            11.01 RDALTT      15 VORIM
      ARC505          7.01 HFSST        11.02 TCNTX          15.01 HFSST
                      7.02 HFAMT                             15.02 HFAMT
 3 HFAMR                                   ARN504
   3.01 HFSST       8 DF44R2                              16 VORSP+
   3.02 HFAMT         8.01 HFSST      12 TCNRX               16.01 HFSST
                      8.02 HFAMT        12.01 UHFTX          16.02 HFAMT
 4 HFSSR                                12.02 TCNTX          16.03 UHFTX
   4.01 HFSST       9 ADF6R             12.03 IDENTX         16.04 VHFAMT
   4.02 HFAMT         9.01 HFSST

                          PAGE  1     (  .00)

                    TYPE >N< TO SEE NEXT PAGE            UNCLASSIFIED
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Fig. 3 Frequency Coincidence Data Summary

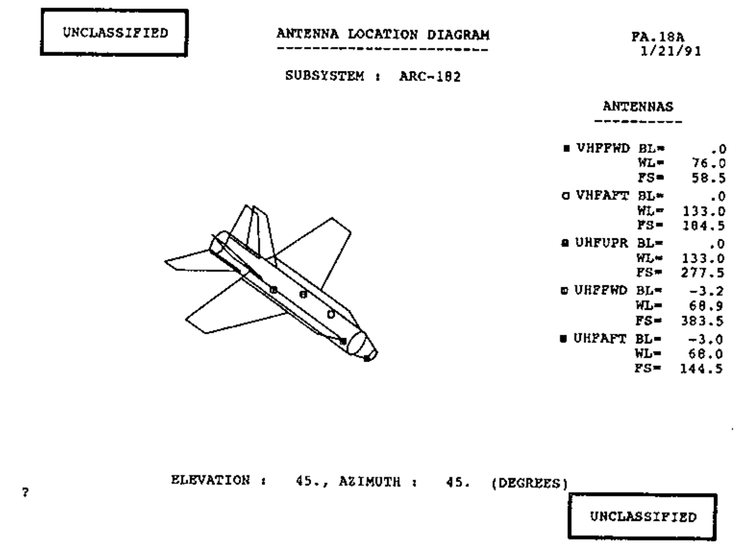


Fig. 4 Antenna Location Display

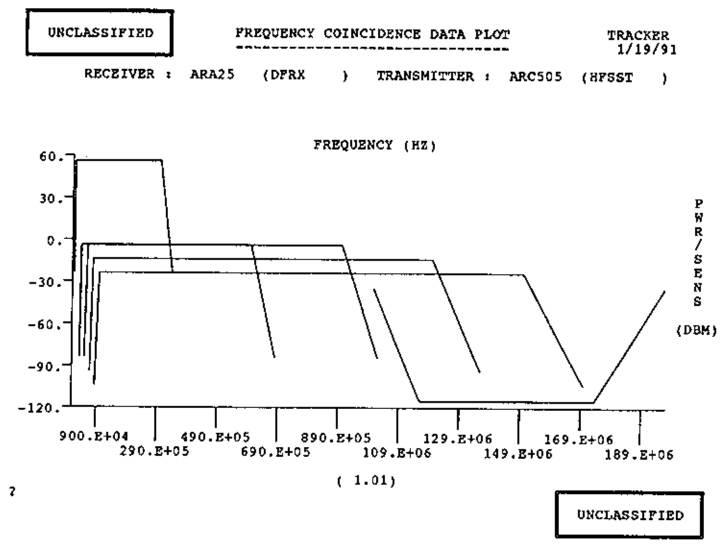


Fig. 5 Frequency Coincidence Plot

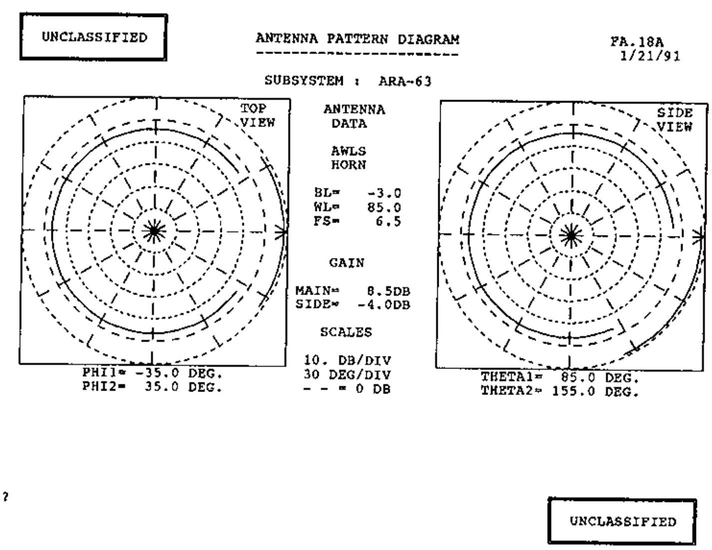


Fig. 6 Antenna Pattern Display

The Propagation Path Display Package provides a means to analyze each interaction, T_{ij}, calculated in the EMCCS. The user may select the antenna pair for which he/she wishes to perform a detailed analysis.

The selection of a particular T_{ij} produces a display as shown in Figure 7. This display contains the tabulation of all the component values used to derive the EMI margin value for the worst-case frequency of each transmitter harmonic. Also present in that display, is the aircraft model with the antenna locations and the actual coupling path between these antenna locations. The two antennas are well identified in this graph and if the users wish to have a closer look at the path, simple mnemonic commands can be entered to have a close up-view side view, top view or front view of the platform (Figure 8). The model can also be displayed from any elevation and/or azimuth. The set of displays and data available in this package provides a means to obtain a detailed analysis of each T_{ij} interaction value. The user obtains an appreciation of power levels, frequencies, and a view of the path of maximum coupling. Thus he validates the input information and appreciates each of the calculation elements.

The "Antenna Position Input" module permits the relocation of each antenna and also allows the user to alter its characteristics. By selecting the antenna of interest, a display similar to that of Figure 9 can be obtained. By utilization of electronic cursor or by typing the coordinates or a combination of both, the antenna can be relocated on the two views of Figure 9, where both the old and new positions are displayed along with their coordinates.

Once the antenna relocation is confirmed, the code produces a listing of the antenna characteristics which can be modified individually. Subsequently, the user executes a recompute command which has the effect of recycling the entire set of interactions T_{ij} and renders available the new set of data such that a complete new analysis can be done. A previous paper[7] shows example results of such relocations.

This process can be repeated as often as necessary to analyze or optimize the relocation of an antenna.

2.2.5 Evaluation Comments

The AAPG system, even with limited system data, provides the EMC test engineer with an appreciation of the overall interaction matrix so that he can anticipate difficulties and make the best choice of test frequency combinations in this total context. In the process he has contact with all the modelling elements - input data, aircraft model, antenna model and coupling paths. He becomes aware of the power level differences and the span of frequencies over which they are significant for EMC analysis purposes. The display of the coupling path in the simplified model leads naturally to the examination of the physical differences between the actual aircraft and the model. Coupled with the ability to relocate antennas, this feature provides a means to perform a tolerance analysis that produces a fuller appreciation of the numerical values in each interaction.

The interactive graphics package with its rapid response and comprehensive templates provides a dynamic process where the validity of the information is readily evident and the EMC evaluation has aspects of visibility that are difficult to achieve in any other way. When flight test results are correlated with results of such prediction, the combination forms a useful EMC core data base that can be exploited during the entire life cycle of an aircraft system. Corroborations with actual measurements have been presented by Hodes and Widmer[11].

The path algorithms within AAPG use creeping wave expressions which had been validated for a limited UHF and IFF data set. Recent measurements [12] at higher frequencies and for a variety of antenna positions suggest that an improved formulation based on the work of Pathak, Burnside, and Marhefka [13] should be sought. Important new capabilities have also been added in Version 09 of the code.

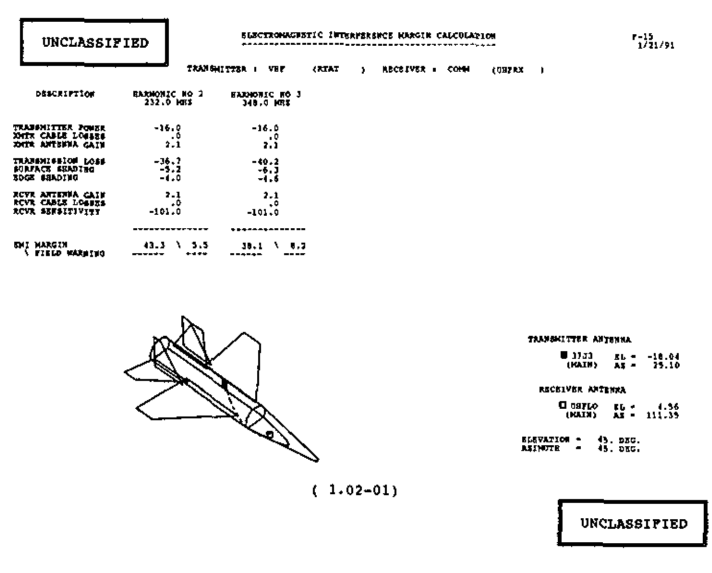


Fig. 7 EMI Margin Display

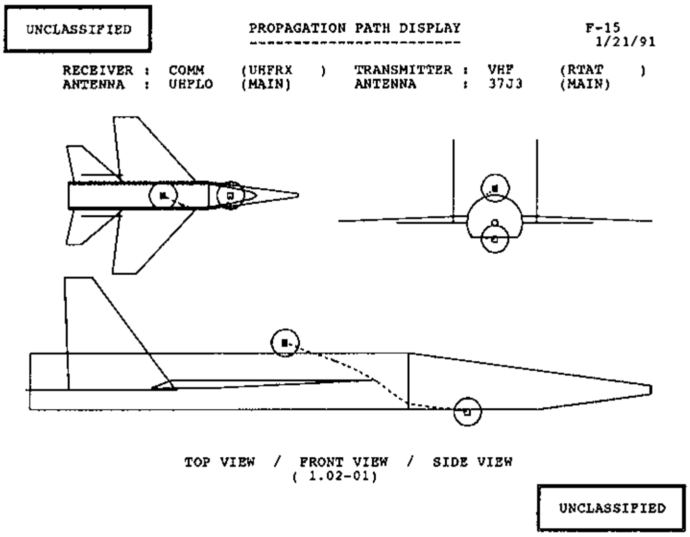


Fig. 8 Close-up Propagation Path

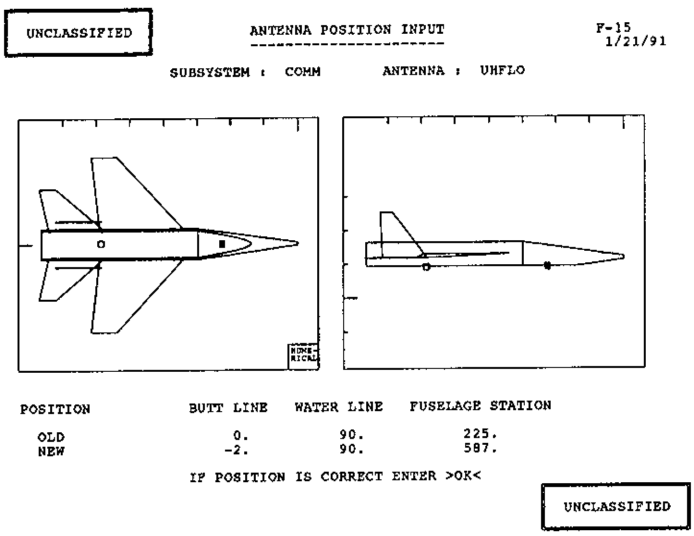


Fig. 9 Antenna Position Input

2.2.6 Features of AAPG Version 09

The three areas of additional capability that have been added in Version 09 of AAPG are: Gaussian approximation of the antenna pattern, representation of aircraft fuselage cross-sectional shapes by multiple points and the possibility to include forward and aft cones which may also be tilted.

For directional antennas, AAPG calculates the gain as a function of angle by assuming a Gaussian falloff in two orthogonal planes between the user input values for the main beam and sidelobe gains. The Gaussian function in each plane is parametrized to the -3dB beamwidth of the main beam in that plane. An example is shown in Figure 10.

In addition to the circular cylinder, with or without flat bottom, the cross-section of the fuselage cylinder and cones can be specified as a series of radius values from the centroid at specific angles. These points are joined within the code by spline fitting. This provides more flexibility in the representation of modern complex aircraft shapes. The forward and aft cones are assumed to have the same cross-sections as the fuselage. An example of an Version 09 display is shown in Figures 11(a) and 11(b).

A version of AAPG.DI has been adapted for the preparation of input data sets for AAPG Version 09.

3. FIELD COMPUTATIONAL MODELLING METHODS

In addition to the system level analysis discussed above, there are many occasions where separate and more detailed field or coupling computations must be undertaken. It is always useful to have techniques available for the calculation of radiation patterns of antennas for performance assessment purposes and for use in deriving their approximations as input for a system analysis.

Moment methods applied electric field integral equation formulations and GTD/UTD(Uniform Theory of Diffraction) computer codes are the most commonly used instruments in computational modelling for low and high frequency antennas respectively. At the same time the potential of the TLM method is being explored by several investigators. The latter method is particularly attractive because of its potential in the representation of composite surfaces.

For frequencies above VHF, GTD/UTD techniques are being applied. Molinet[14] describes the level of model complexity which can be achieved by UTD methods at present. Some of the codes described by Molinet are not widely available. The Ohio State codes[15] are in wider use. However, most of the examples reported with GTD/UTD codes show results in the roll plane, rather than in the pitch plane or volumetrically. Also, the modelling of complex antenna source patterns is not completely developed in these codes. Readers should consult the references for an appreciation of the potential and limitations of these methods.

3.1 Moment Methods

There are now a number of computer codes such as NEC[16], MININEC[17] and GEMACS[3] which are used for low frequency radiation analysis. They use moment methods to solve an electric field integral equation (EFIE) as applied to a wire-grid representation of complex surfaces. NEC also allows a surface patch representation, and other surface patch codes[18] are becoming available. NEC and MININEC are distributed by the Applied Computational Electromagnetics Society[19] to their members. The following discussion applies to the development and use of wire-grid models with the NEC code.

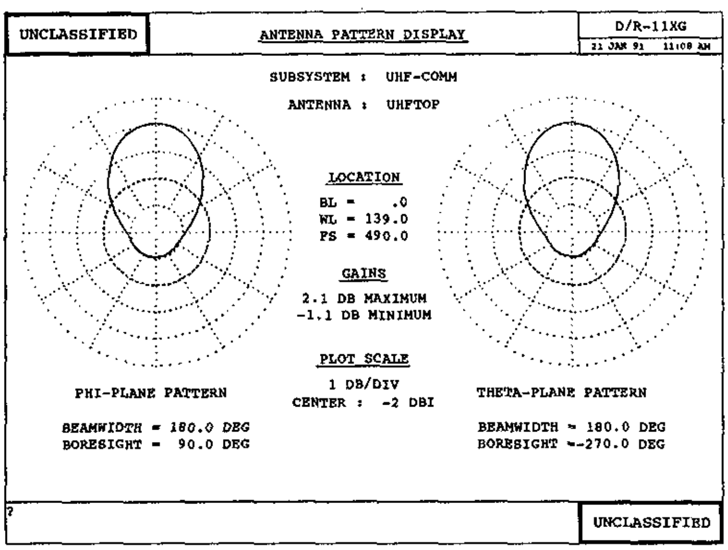


Fig. 10 Antenna Pattern Display V.09

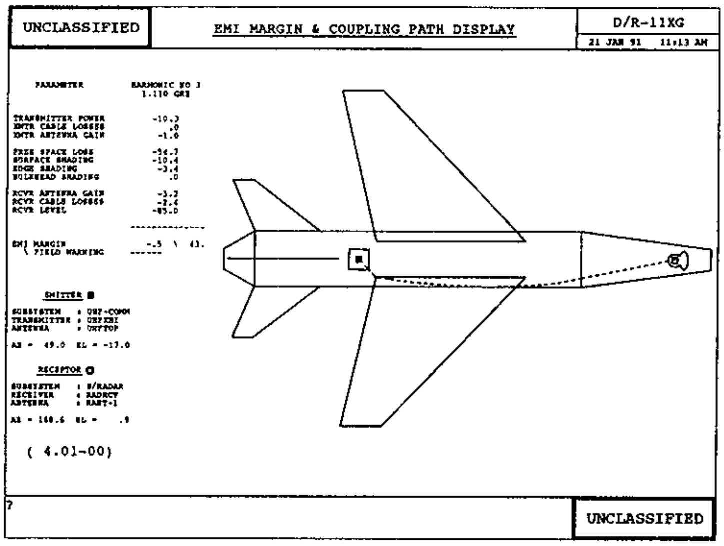


Fig. 11(a) EMI Margin Display V.09

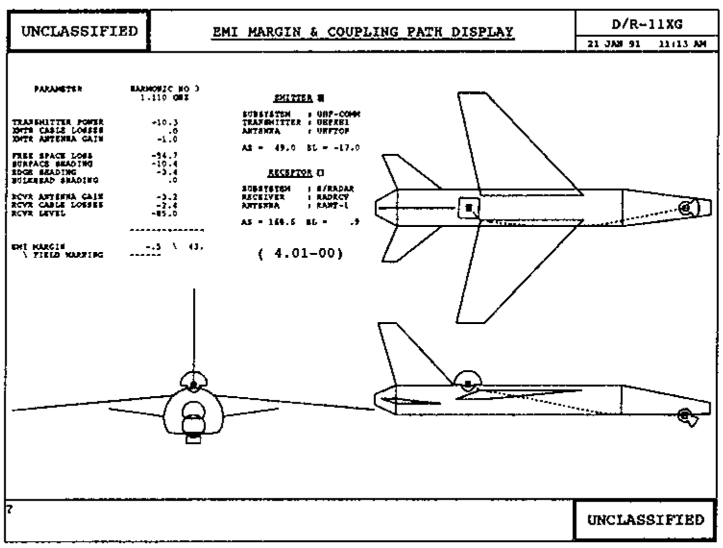


Fig. 11(b) EMI Margin and Three-Views V.09

3.1.1 The Electric Field Integral Equation

Consider an interconnection of highly-conducting wires, of radius much, much less than the wavelength. The current on each wire gives rise to an electric field. Thus in Fig. 12, the axial surface current density $\overline{J}_i$ is assumed to be uniformly distributed about the wire periphery, hence the current is $I_i = 2\pi a J_i$. This current gives rise to an electric field component at the observer in the direction $\hat{s}$ given by[16]

$$\overline{E}_i \cdot \hat{s} \;=\; \frac{-j\eta}{4\pi\beta}\int_{L_i} I_i(s_i)\left\{\beta^2 \hat{s}\cdot\hat{s}_i - \frac{\partial^2}{\partial s\,\partial s_i}\right\} G(s,s_i)\,ds_i$$

where η is the characteristic impedance of the medium, β is the wave number, and s is distance in direction $\hat{s}$ at the position of the observer. Kernel function $G(s,s_i)$ is given by

$$G(s_i,s) \;=\; \frac{1}{2\pi}\int_0^{2\pi} \frac{e^{-j\beta R}}{R}\,d\phi$$

where R is the distance from the "source point" on the surface of wire #i to the observer. If there are N wires in the grid, then the $\hat{s}$ component of the electric field at the observer due to all N wires is the "secondary field" and is given by

$$\overline{E}_{sec}\cdot\hat{s} \;=\; \sum_{i=1}^{N_w} \overline{E}_i\cdot\hat{s}$$

The "primary field" $\overline{E}_{primary}$ in a scattering problem is an incoming plane wave. The boundary condition for thin wires states that, at any point on any one of the wires, the axial component of the electric field must be zero. Thus, if s_k is a point on wire #k, then the boundary condition states that

$$\sum_{i=1}^{N_w} \frac{-j\eta}{4\pi\beta}\int_{L_i} I_i(s_i)\left\{\beta^2 \hat{s}_k\cdot\hat{s}_i - \frac{\partial^2}{\partial s_k\,\partial s_i}\right\} G(s_k,s_i)\,ds_i \;=\; -\overline{E}_{primary}(s_k)\cdot\hat{s}_k$$

This version of the EFIE is called Pocklington's Integral Equation.

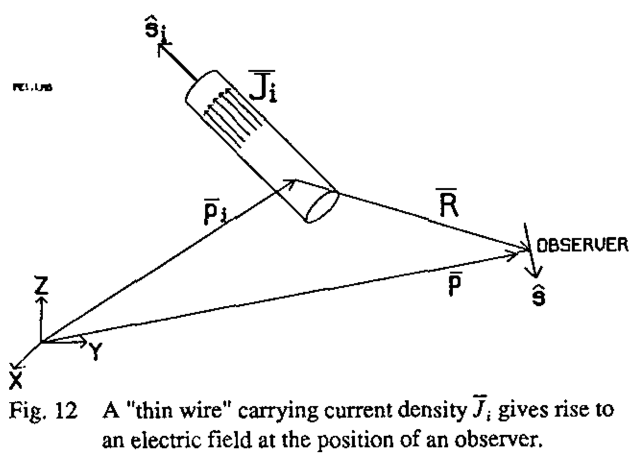


Fig. 12 A "thin wire" carrying current density $\overline{J}_i$ gives rise to an electric field at the position of an observer.

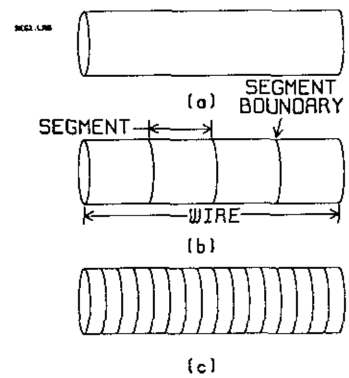


Fig. 13 A wire divided into segments, with various ratios of the segment length to the radius. (a) ratio 8; (b) ratio 2; (c) ratio 1/2.

3.1.2 Moment Method Solution of the EFIE

The numerical solution of the EFIE for an antenna or scatterer made up of an interconnection of thin wires is well-known in the literature[20]. In the Numerical Electromagnetics Code(NEC)[16,21,22], each wire of the model is subdivided into "segments", as illustrated in Fig. 13. The current on segment j of wire #i is of the form

$$I_{ij}(s_i) \;=\; A_{ij} + B_{ij}\sin(\beta(s_i - s_{ij})) + C_{ij}\cos(\beta(s_i - s_{ij}))$$

Thus there are three unknown complex-valued current amplitudes for each segment, and hence $3N_s$ unknowns in all where N_s is the total number of segments. Constraints for determining the values of these unknowns are obtained by requiring that the current and its derivative, the charge density, be continuous functions from segment to segment along each wire. At junctions of wires, Kirchoff's Current Law must be satisfied. Also, an excess charge is permitted to accumulate on the wires adjacent to a junction, and to be distributed among the wires according to the King-Wu junction constraint[16,23]. These charge and current constraints provide a set of $2N_s$ equations.

The remaining set of N_s equations are obtained by using the "moment method" to satisfy the boundary condition. Thus a "match point" is defined at the center of each segment of the antenna, and the EFIE is enforced at that point. Each match point provides one linear equation, that is, one row of a "moment method" matrix. The NEC code is formulated in such a way that the full set of $3N_s$ linear equations is never explicitly assembled. Rather, KCL and the charge density constraints are enforced as the $N_s x N_s$ "moment method" matrix is assembled.

The integration required by the kernel function is approximated in two different ways in the NEC code. These are obtained by expanding the kernel function as an infinite series, in powers of the wire radius. Retaining only the first term obtains the "thin wire kernel" or "normal kernel" approximation. This is equivalent to considering that the current flows along the centerline of the wire, and that the "match point" is located on the surface of the wire. The "extended kernel" approximation retains two terms of the power series, and is more accurate.

3.1.3 Modelling Guidelines

The assumption of "thin wires" made in deriving the EFIE, the approximations of the kernel, and the approximate solution of the EFIE by the "moment method" lead to restrictions on the geometry of the interconnection of "thin wires" that the NEC program can solve correctly. Table 2 gathers these restrictions as "modelling guidelines". The guidelines are based on extensive tests of the results obtained by solving simple structures with the NEC code. Those presented here are based both on the NEC User's Guide[21] and the authors' own experience.

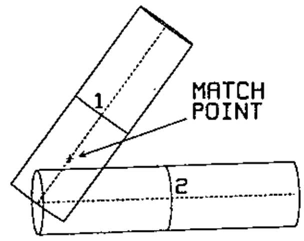
NEC requires that the segments be sufficiently short in terms of the wavelength. Segments comparable to $\lambda/10$ are usually adequate, but segments shorter than $\lambda/20$ are required in critical regions of the antenna. The wires must be "electrically thin": wires fatter than $\lambda/30$ are considered in error. The approximations of the kernel place restrictions on the ratio of the segment length to the radius. If the "normal kernel" is used, the segment length is best maintained at least 8 times the radius, but acceptable results are sometimes obtained for segments as short as twice the radius. Fig. 13 illustrates segments of length-to-radius ratio eight in part (a), and two in part (b). If the "extended kernel" is used, then the segment length is best kept twice the radius, but segments as short as half the radius are sometimes acceptable, illustrated in Fig. 13(c).

At wire junctions, there are restrictions aimed at making the joined segments not too dissimilar. The ratio of the longest segment to the shortest segment length at a junction must be less than five. The ratio of the fattest radius to the thinnest radius must be less than 10, with values less than five preferred. Because the NEC code does not fully use the "extended kernel" for segments which are part of wire junctions[16], the "normal kernel" rules for segment to radius ratio should be applied at junctions. Table 2 considers that, for a segment which is part of a wire junction, a segment length to radius ratio of six is fully acceptable.

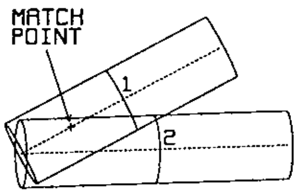
Another important restriction applies at a wire junction. The match point at the center of any segment at the junction must lie outside the volume of all the other wires at the junction, or else a "match point error" occurs. This is illustrated in Fig. 14. Part (a) shows a junction of two wires, each having two segments. The segments are relatively fat compared to the length, which is a common occurrence in a wire grid. The match point on wire #1 lies barely outside the volume of wire #2. This is undesirable. Part (b) shows the wires meeting at a shallower angle. Now the match point lies inside wire #2, a match point error. Part (c) shows the case of a short wire joining a longer wire, in which the match point on the first segment of the short wire lies inside the volume of the longer wire. It is an error to use two segments on the short wire, but it is one that is easily made. If NEC is run on a configuration of wires including some "match point errors", then the resulting currents may be incorrect and can be quite misleading. Clearly this guideline limits the angle at which wires can join, in terms of the segment length and the wire radius.

TABLE 2
SUMMARY OF THE MODELLING GUIDELINES
Δ = segment length
a = wire radius
λ = wavelength

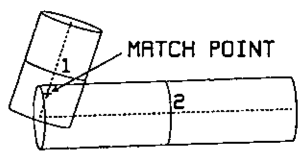
INDIVIDUAL SEGMENTS	Warning	Error
segment length	$\lambda/10 < \Delta < \lambda/5$	$\Delta > \lambda/5$
radius	$30 < \lambda/a < 100$	$\lambda/a < 30$
segment to radius ratio	$0.5 < \Delta/a < 2$	$\Delta/a < 0.5$
JUNCTIONS		
segment length ratio		$\Delta_{big}/\Delta_{small} > 5$
radius ratio	$5 < a_{big}/a_{small} < 10$	$a_{big}/a_{small} > 10$
segment to radius ratio	$2 < \Delta/a < 6$	$\Delta/a < 2$



(a) match point close to the other wire's surface.



(b) match point inside the other wire's volume.



(c) match point inside the other wire's volume.

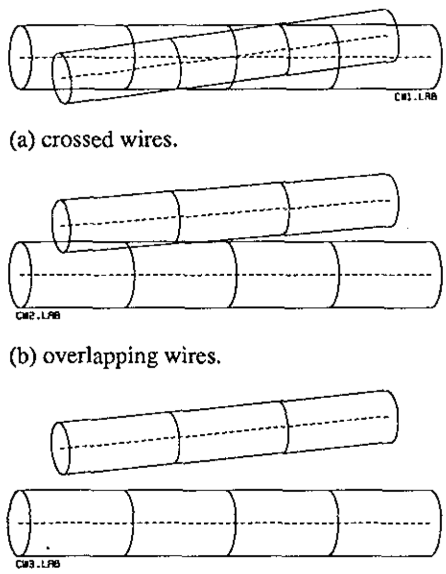
Fig. 14 The match point must lie outside the volume of the other wire at a wire junction.

Modelling guidelines for the spacing of wires concern both gross errors, and the "thin wire" assumption. Fig. 15(a) shows a pair of nearly-parallel wires which cross, that is, have a common point on their centerlines. In the NEC program, if the common point is not a segment boundary on both wires, then NEC will not form a junction between the two wires, and a "crossed wires" error occurs. The configuration of Fig. 15(a) also has "match point errors". A typical computer graphics display of a wire grid depicts wires by their centerlines, effectively hiding a crossed-wires error. It is essential that both wire radius and segment boundaries be shown on computer graphics, as in Fig. 15(a), if the user is to be able to see a crossed wires error.

If two wire centerlines pass closer than the sum of the wire radii, then the wires overlap and would have to be physically joined. Fig. 15(b) shows an "overlap error" for nearly-parallel wires. NEC does not form junctions between wires which overlap in this fashion: they are treated as unconnected wires.

The "thin wire" assumption requires that the geometry of the wires be such that the current flow on any wire is entirely axial, with no circumferential component. Further, the axial current density must be uniformly distributed about the wire periphery. This requires that wires be spaced sufficiently far apart. It is difficult to obtain guidance from the literature on just how far. Ludwig[24] recommends several diameters. Clearly the wires of Fig. 15(c) are too closely spaced to satisfy the "thin wire" assumption, even though they do not overlap. This geometry is a "near miss".

Wire grids are occasionally constructed of nearly-perpendicular wires, using NEC's feature of forming junctions at crossing-points if the point is a segment boundary on both wires. Such grids often contain errors such that crossing points are not segment boundaries, hence contain "crossed wires" errors. Then the wires are not joined by the program, and the grid does not express the model builder's intentions.



(a) crossed wires.

(b) overlapping wires.

(c) a near miss.

Fig. 15 Wire spacing errors for nearly parallel wires.

A systematic examination of a wire grid geometry file is required to ensure that all segments and junctions conform to Table 2, that there are no "match point errors", and that the spacing criteria are met. Visual inspection of graphics displays of various views of the wire grid have proven woefully inadequate for this purpose. Program CHECK[25] examines each wire, each junction and each wire pair to ensure compliance. A model verified by CHECK is one that is more likely to lead to correct results when analysed with the NEC program.

The "modelling guidelines" are local in nature. They concern individual segments, pairs of segments at wire junctions, and wire spacing on a scale of a few wire radii. To use wires to model solid surfaces, guidelines of a more global nature are required.

3.1.4 Wire Grid Guidelines

In replacing a highly-conducting, continuous surface with a grid of wires, fundamental questions arise. What is the best grid topology: triangular meshes or rectangular meshes? What is the bandwidth of a wire grid of a given mesh size? What is the best wire radius for wires representing a continuous surface? The wire grid guidelines of Table 3 provide principles upon which a model can be designed. These guidelines are oriented toward modelling complex structures such as ships or aircraft, often with associated wire antennas.

A wire grid using square mesh cells requires fewer segments per square wavelength to cover a surface than does a wire grid using equilateral-triangle mesh cells. Whether the triangular grid offers wider bandwidth in return for the larger number of segments has not been adequately explored[26]. Table 3 recommends an orthogonal grid with square mesh cells.

TABLE 3
WIRE GRID GUIDELINES

1. Use a rectangular grid of wires oriented parallel to edges, with wires along edges.
2. Choose the segment length equal to $\lambda/10$ at the highest frequency of the band. Keep the area of the grid cells comparable to $(\lambda/10)^2$.
3. Keep grid cells square. Keep the cell area constant, hence segment length constant, throughout the grid.
4. Provide elegant transitions.
5. Put wires where current is expected to flow.
6. Avoid wires meeting at very shallow angles.
7. Keep the center of any segment outside the volume of any other segment.
8. Use the "equal-area rule" radius.
9. All meshes in the grid must have peripheries longer than $\lambda/25$.

3.1.5 Model Development and Results

The application of the modelling guidelines to the creation of models for complex shapes such as aircraft is discussed in reference[25]. To assist in the error-free creation of a wire-grid or surface patch model, the computer-aided system called DIDEC[27] has been developed. This acronym means Digitize, Display, Edit and Convert. Vertices are created by the digitization of points on aircraft drawings or by numerical entry. The digitization process is accompanied by a multi-ported display of the data. Vertices are joined to form the wires and wire segments of the wire grid. They can be colour-coded as to radius or length. Editing steps can add or delete vertices etc.. The final model data set is converted to a NEC input data format by a simple transformation command.

Results of NEC wire-grid modelling for antennas on aircraft, helicopters and ship's topside structures have been presented in Lecture Series 165 [28]. One of the wire-grid models which had been used is shown in Fig.16. It consists of 371 wires and 400 segments and in the main, wire radii were selected based on the "equal area rule" with some variations based on trade-offs with other working guidelines. The segment length had been chosen to be 0.1λ at 15 MHz. This model was executed at 2 MHz increments over the HF band for radiation patterns to correspond to measurement frequencies. Impedance computations are also compared therein with measured data. The results show that carefully developed models can provide reliable radiation patterns results over the entire 2-30MHz band.

How are these computations related to EMC requirements? Although initially related to systems communications performance alone, the results of these computations serve to "validate" the computational model. As can be seen from the AAPG discussion, such patterns refine the approximations used in this code. The models can be used for computations of the near-field at other antenna and weapons locations. The computed current distributions on the wire-grid model also result in an appreciation of HF antenna coupling as in the case of the CP-140 aircraft model shown[28]. Near-field intensities are also important for estimates of RADHAZ conditions.

Whereas the 400 segment model shown in Fig. 16 would take approximately an hour on a CYBER 325 mainframe available during the development period, it now takes approx. 567 secs. on a MIPS desktop which is being evaluated. Such laboratory computing power allows a new outlook on model development, less constrained by the number of segments and execution times.

An extensive re-examination of wire-grid modelling guidelines (described above) had been undertaken for an RCS project[26]. Applied to the same aircraft, these guidelines have resulted in the model shown in Fig.17. This more "elegant" model contains 685 segments instead of 400 and takes 6128 secs. on the MIPS machine. The model more directly results in a finer structure at the feedpoints of antennas which had been found critical in previous work. Thus there is every indication that this would be a safer approach for "ab initio" computational model evaluation and use.

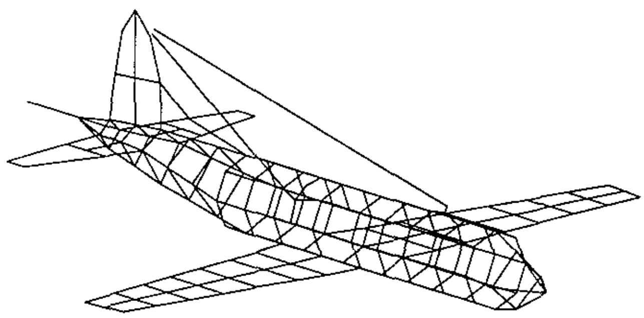


Fig. 16 Aircraft Model, 400 segments

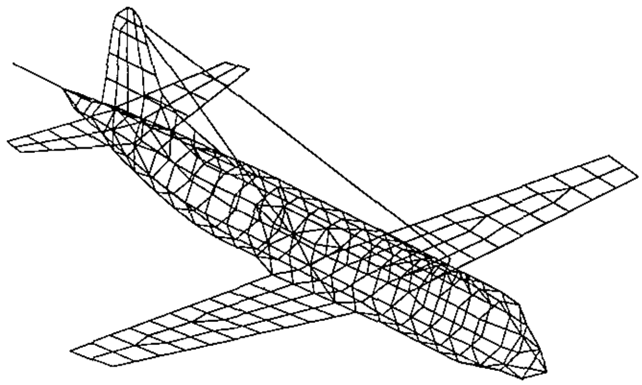


Fig. 17 "Elegant" Aircraft Model, 685 segments

3.2 Transmission Line Modelling

Transmission-line modelling (TLM) is a numerical electromagnetic technique developed mainly by Johns [29,-..,37] in the 70's and 80's. The basic solution provided by TLM is the time-domain variation of the fields in a defined region of space. Within the region to be modelled, metal and dielectric objects can be defined and the boundaries of the region can have various properties, such as perfectly conducting wall or a free space approximation. A very good introduction to the basic theory behind TLM is provided by Hoefer [38], but a brief outline will be presented here.

3.2.1 Outline of TLM

The basis of the technique is Huygens' Principle [39] which states that a wavefront can be made up of a number of secondary radiators, each producing spherical wavelets. These wavelets form a new wavefront which in turn can be broken down into a series of wavelets. In TLM, this principle is discretized in space and time to allow solution on a computer. Figure 18(a) shows the formation of a wavefront according to Huygens' and this is shown in Figure 18(b) superimposed upon a Cartesian mesh in a form suitable for TLM. The distance between nodes (dl) and the time taken for electromagnetic pulses to travel from one node to the next (dt) are related by the velocity of light.

The 2D Cartesian mesh is made up of transmission lines, with each node being the junction between four transmission lines. A typical node is shown in Figure 19 for the junction between parallel wire transmission-lines. The relationship between a voltage pulse incident upon the node and the resulting scattered voltages is defined by a scattering matrix. The scattering matrix is given below for the voltages scattered at all four ports of the 2D node. The scattered pulses immediately become incident voltages upon adjacent nodes. An analogy can be formed between voltages and currents on the transmission lines and electric and magnetic fields in the region being modelled. The permittivity and permeability of the region are defined in the model by the capacitance and inductance per unit length of the transmission lines.

$$\begin{pmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{pmatrix}_s = \frac{1}{2} \begin{pmatrix} -1 & 1 & 1 & 1 \\ 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \end{pmatrix} \times \begin{pmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{pmatrix}_i$$

At the boundaries of a region a number of options are available. In order to model a perfectly conducting wall, all transmission lines meeting the boundary are terminated in a short-circuit. Alternatively, it is possible to approximate a free space boundary by terminating the lines in their intrinsic impedance. The free space condition works perfectly for an incident wavefront perpendicular to the boundary, but boundary reflections occur if the wave is incident at other angles. It is possible to improve the free space boundary by assessing the wavefront angle of incidence at each time step and adjusting the terminating impedance accordingly. This computation has to be performed at every node along the boundary.

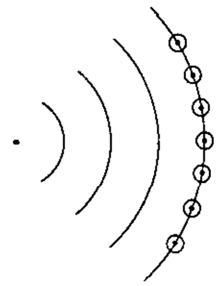


Fig. 18(a) Wavefront divided into wavelets according to Huygens

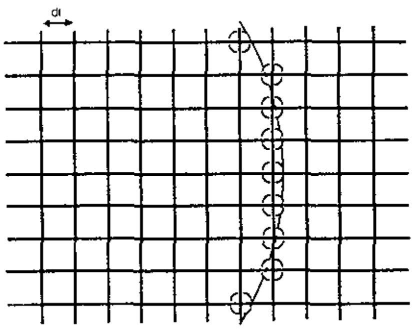


Fig. 18(b) Wavefront superimposed on TLM mesh showing approximate scattering centres.

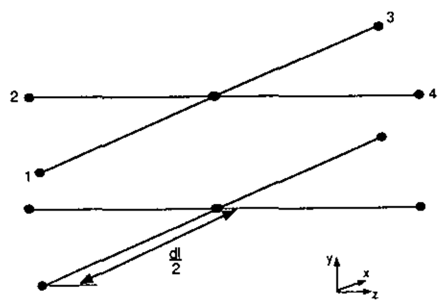


Fig. 19 TLM 2D node made up of parallel wire transmission line.

A metal object within the mesh is modelled in a similar way to a perfectly conducting boundary, i.e. by short-circuiting all the transmission lines at the boundary of the object. One of the advantages of the TLM technique is the ability to model dielectric and magnetic materials. This is achieved by adding an extra stub transmission line at each node within the object, where the admittance of the stub determines the relative properties of the object. TLM is even capable of modelling lossy dielectrics and this can be achieved in two ways. One possibility is to use lossy transmission lines between nodes, whilst the other used lossy stubs at each node to absorb energy from the mesh. The latter technique is preferred in situations where the region to be modelled is not homogeneous.

Once a mesh has been set up, the initial step is to set all the voltages at the nodes of the mesh to zero. Then an excitation is applied and the time domain response is calculated. Typically, excitations are produced by setting up pulses at a node at the initial time step. In this way it is possible to deduce the impulse response of a structure. Alternatively, it is possible to excite a line of nodes with pulses of identical magnitude which varies over time as $\sin(wt)$ to give a plane wave of angular frequency w.

TLM is equally applicable to two and three-dimensional problems. A schematic diagram of the full 3D node is given in Figure 20. This is the most efficient and accurate 3D node and was derived by Johns [40]. The basic 3D node has a 12x12 scattering matrix, but modelling a lossy dielectric requires an 18x18 matrix. Therefore, storage is required for 18 terms at each node in a 3D mesh. Even a small problem requires a relatively large amount of memory.

The transmission lines in the 3D node do not interact with each other and this allows them to have differing properties. In this way, TLM can even model substances that have anisotropic properties, such as carbon-fibre composites.

3.2.2 Mesh Construction

The mesh discretization is critical to an accurate TLM solution. Effectively, enough nodes must be used to model a wavefront accurately. Normally a $\lambda/10$ criterion is used as the maximum node spacing per wavelength. If fewer nodes are used, the wavefront velocity becomes dispersed until the point is reached (at $\lambda/4$ spacing) where no propagation occurs. This dispersion effect is very dependent upon the direction of propagation across the mesh, with maximum dispersion along the axes of the mesh and no dispersion at 45° to the axes.

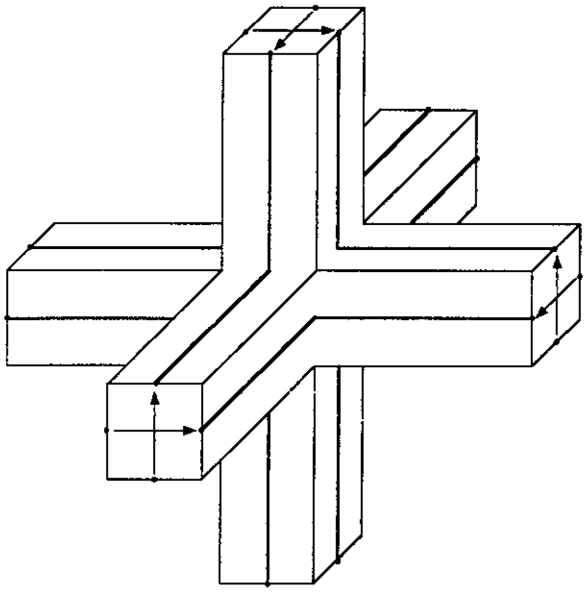


Fig. 20 TLM 3D Node showing each arm made up of two parallel wire transmission lines.

3.2.3 Potential and Limitations

As a numerical technique, TLM has a number of advantages over some computational solutions. It is unconditionally stable and computationally simple, the most complex operation being a matrix multiplication. Furthermore, because no matrix inversion is required, the condition of the matrix is unimportant. TLM can also solve certain types of problems relatively easily. Some of these, such as lossy dielectrics and anisotropic materials have already been mentioned. It is also ideally suited for solving problems with internal resonances.

Disadvantages with TLM include difficulty in representing wires of diameter less than the node spacing (which most wires are, without very high node density). It is also difficult to excite a problem in a practical way, such as voltage source on a structure. The large amount of information stored at each node usually limits the size of a problem due to the finite computer storage available.

TLM has many applications, both within electromagnetics and outside. For example, it has been used to study fluid flow and diffusion problems [41]. Typical electromagnetic problems solved by TLM include waveguide propagation [29,30,32] and screened room resonances. Waveguides have been modelled with dielectric loading [42] and bifurcations [29]. Other examples include direct coupling to a transmission line and coupling through an aperture [43]. The time domain nature of TLM makes it well suited to analysing lightning strike and EMP problems.

Gothard, German and Riggs [44] use TLM to assess the effect of RAM coating on a shipboard HF wire rope antenna. The aim in this case is to minimise the RCS of the antenna to radars also on-board, thus minimising false targets and pattern blockage for the radar. This uses near-field to far-field transformation techniques to reduce the size of the TLM mesh.

A number of techniques are available to reduce the storage required in a TLM problem. Grading the mesh is one possibility [45] whereby the $\lambda/10$ node spacing requirement can be relaxed for regions of the mesh where the fields are unimportant. The other alternative is to use a technique called diakoptics [46] in which two meshes can be solved separately and then joined together. This is applicable where only a small part of a large mesh is changed between solutions. The majority of the problem can be solved once and stored and further modifications require only a solution for the part of the mesh that has been modified.

4. SUMMARY AND COMMENTARY

The complex task of ensuring the electromagnetic compatibility of modern avionic/weapon systems can be made more manageable by the effective use of numerical analysis and modelling techniques. The acquisition and use of computer codes for system analysis and electromagnetic modelling is rewarded by a clearer appreciation of the individual interactions between systems and by the definition of coupling mechanisms. Such use also involves and requires the generation of a progressively improved data base of information on equipment characteristics and radiating elements.

The effective use of these modern computational tools does require the training of technical personnel in the their use. Technical judgment must be developed by analysis of progressively more complex cases. Awareness of the validity of the computational results at each step is most important. The validation of computational models is now the subject of sub-committees of the IEEE Antennas and Propagation Society and the Applied Computational Electromagnetics Society. A data base of canonical examples and real-world practical problems is being generated by ACES in order to assist the technical community in this task. It is important that users become aware of the results and in turn share their experiences so that duplication of effort and costly errors can be avoided.

5. ACKNOWLEDGEMENTS

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CABLES and CROSSTALK

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SUMMARY

Crosstalk is the unintentional electromagnetic coupling between circuits which are connected by parallel conductors that lie in close proximity to each other. Some examples are wires in cable harnesses or metallic lands on Printed–Circuit Boards (PCB's). This unintended interaction between two or more circuits via their electromagnetic fields can cause interference problems. Signals from one circuit that couple to another circuit will appear at the terminals of the devices that are interconnected by the wires. If these signals are of sufficient magnitude or spectral content, they may cause unintended operation of the device or a degradation in its performance. A summary of the standard models used for predicting crosstalk in various types of configurations is presented. The discussion will focus on the relative accuracies, regions of applicability and computational complexity of the models. A simple explanation of the ability (or inability) of shielded wires and twisted pairs of wires to reduce the crosstalk will also be given.

I. INTRODUCTION

Crosstalk is the unintentional coupling of electromagnetic fields between circuits that are connected by conductors. Fig. 1 illustrates this phenomenon. A source is connected to a load via one pair of wires. Another source is connected to its load via another pair of wires and the wires of each circuit are parallel and lie in close proximity. Currents flowing along these wires create magnetic fields that couple to the other circuit. Similarly, charges on the wires generate electric fields that couple to the other circuit. These electromagnetic fields induce signals in the two circuits. Portions of these signals, $\alpha_1 v_{S1}$ and $\alpha_2 v_{S2}$, appear at the inputs to each circuit. Prediction of these signals at the inputs to the circuits is the task in modeling crosstalk. The essential question is whether these unintended signals will cause the respective loads to malfunction.

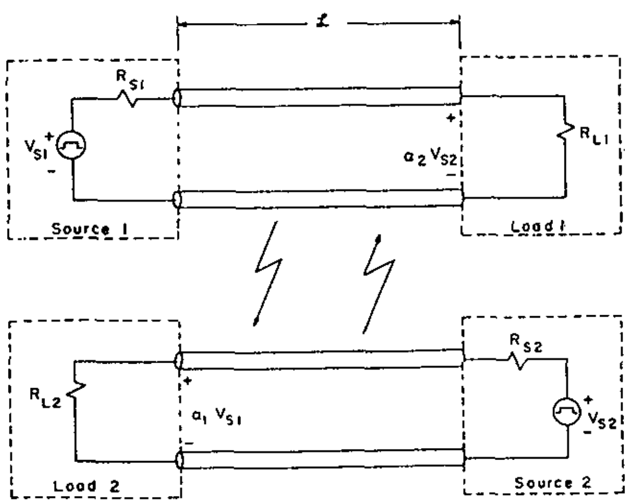

Fig. 1. Illustration of crosstalk

The objective of this paper is to review the standard mathematical models that are used to predict this crosstalk. The relative accuracies, computational complexities, and relative insights gained from each model will be discussed. If the crosstalk voltages are of sufficient magnitude and/or spectral content to cause unintended operation of the load, certain alternative wiring measures may be used to reduce this crosstalk to acceptable levels. The typical measures are to surround the emitter or the receptor wires with a shield or to replace either wire with a twisted pair. Although it is often assumed that these correction measures will reduce the crosstalk, this is not always the case. Simple explanations of how a shielded wire or a twisted pair of wires will or will not reduce crosstalk will be given. The key to understanding this is a low–frequency, inductive–capacitive coupling model. In addition to providing insight into the phenomenon, this simple model allows first–order predictions of the crosstalk that are obtainable with hand calculations. This simple low–frequency, inductive–capacitive coupling model will also be discussed.

This paper is intended to be a review of the various aspects of crosstalk. To assist the reader in reviewing the applicable literature, an extensive list of publications by the author on this subject is provided at the end of this paper. These publications are grouped in categories to further focus on the applicable reference. These references are numbered according to the category: A. Books, B. General, C. Per–Unit–Length Parameters, D. Cable Harnesses, E. Ribbon Cables, F. Shielded Wires, G. Twisted Pairs of Wires, H. Effects of Incident Fields, I. Digital Computer Programs, and J. Printed Circuit Boards.

II. The Multiconductor Transmission Line (MTL) Model

The fundamental assumption in crosstalk prediction models is that the electric and magnetic fields surrounding the conductors satisfy a *Transverse ElectroMagnetic* (*TEM*) field structure in that the electric and magnetic fields lie in a plane that is transverse (perpendicular) to the line axis [A.1,B.1]. This assumption considerably simplifies the modeling and is generally valid so long as the cross–sectional dimensions of the line (wire separation and cross–sectional size) are electrically small, i.e., much less than a wavelength, λ, at the frequency of interest. Crosstalk prediction models have been investigated by the author and are listed in the references. Experimental data that confirm the prediction accuracies of the models are documented in those publications. The purpose of this paper is to summarize those prediction models.

In order to illustrate the essential features of each model we will consider a three–conductor line illustrated in Fig. 2. In general, cable bundles and coupled lands on Printed–Circuit Boards (PCB's) consist of numerous

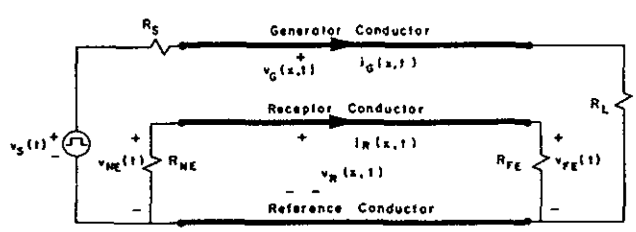

Fig. 2. A three–conductor transmission line

parallel conductors that interact. However, it has been shown that omitting consideration of the other *parasitic conductors and circuits* leaving a three–conductor line tends to give upper bounds or a worst case estimate of the crosstalk. References [D.1], [D.2], [D.3] and [D.4] contain those data. These data also indicate that the crosstalk in a random cable bundle (cable harness) wherein the relative positions of the wires are not known and vary along the bundle is quite sensitive to this relative wire position. Therefore it is appropriate to compute bounds on this crosstalk rather than attempting an exact prediction. Detailed FORTRAN programs that compute the crosstalk for a general multiconductor line consisting of (n+1) parallel conductors are contained in [I.1–I.4]. For a large number of coupled conductors, one has no recourse but to utilize these types of digital computer programs to calculate the crosstalk and little general insight is gained. Restricting our discussion to the three–conductor line of Fig. 2 will allow considerable insight to be obtained and will illustrate the essential points of the more general case of a large number of parallel conductors. The effects of incident fields that are coupled to the line can also be predicted with these digital computer programs, and these programs are described in [H.1–H.9].

The three–conductor line of Fig. 2 consists of three parallel conductors of uniform crossection along their lengths. This is referred to as a *uniform* transmission line in the sense that cross–sectional views of the line at two different positions along the line are identical. Therefore the conductor crossections as well as the crossections of any surrounding dielectric inhomogeneities such as wire insulations must not vary along the line. The conductors are parallel to the x axis in a rectangular coordinate system, and the y–z plane is perpendicular to this direction and contains the electromagnetic fields. According to the TEM field structure assumption, the electric and magnetic fields have no component in the x direction (along the line axis). As the frequency of excitation is increased such that the line cross–sectional dimensions become electrically large, higher–order, non–TEM field structures become important. For typical line crossectional dimensions of interest, the TEM field structure will be the primary field structure on the line up to frequencies on the order of a GHz so that the TEM model provides predictions over a substantial frequency range of interest.

The three–conductor line of Fig. 2 consists of a *generator conductor*, a *receptor conductor*, and a *reference conductor*. The generator conductor along with the reference conductor comprise the *generator circuit*. Similarly, the receptor conductor along with the reference conductor comprise the *receptor circuit*. The generator circuit is driven at the left end by a source represented by a voltage source, $v_S(t)$, and a source resistance R_S. Resistors will be used to represent sources and loads although the majority of the following results will apply to more complex impedances consisting of inductors, capacitors and controlled sources. The generator circuit is terminated in a load represented by the resistor R_L. Similarly, the receptor circuit is terminated in loads represented by the resistors R_{NE} and R_{FE}. The subscripts on these loads refer to the "near end" and "far end" with respect to the end of the receptor circuit nearest to the source of the generator circuit. The assumption of the TEM mode of propagation allows us to define, unambiguously, the voltages between the conductors and the currents along them. The voltage of the generator conductor, $v_G(x,t)$, is defined for the generator conductor with respect to the reference conductor. The current of the generator conductor, $i_G(x,t)$, is defined as flowing to the right along that conductor. Similarly, the voltage and current of the receptor conductor are defined as $v_R(x,t)$ and $i_R(x,t)$. These voltages and currents are dependent on position, x, along the line and time, t. The objective in a crosstalk analysis is, given the source voltage, $v_S(t)$, the termination resistances, R_S, R_L, R_{NE}, and R_{FE}, along with the cross–sectional dimensions of the line (conductor radii and separation) and the line length $\mathscr{L}$, predict the near–end and far–end crosstalk voltages, $v_{NE}(t)$ and $v_{FE}(t)$. There are two types of crosstalk predictions that may be desired; *time–domain crosstalk* and *frequency–domain crosstalk*. Time–domain crosstalk prediction refers to the prediction of the time–domain waveshapes of $v_{NE}(t)$ and $v_{FE}(t)$. Frequency–domain crosstalk prediction refers to the use of a sinusoidal source, $v_S(t)=V_S\sin(\omega t+\theta)$, and the prediction of the *magnitude* and *phase* of the steady–state voltages, $V_{NE}\angle\theta_{NE}$ and $V_{FE}\angle\theta_{FE}$. In this paper, we will be primarily interested in frequency–domain crosstalk since the various regulatory limits are in the frequency domain.

The general model of Fig. 2 is used to consider a number of different, practical configurations. Fig. 3 illustrates typical wire–type cross–sectional configurations. In all these configurations, the generator and receptor conductors are considered to be wires (conductors of circular cylindrical cross section). The reference conductor may be another wire (Fig. 3(a)), an infinite ground plane (Fig. 3(b)), or an overall cylindrical shield (Fig. 3(c)). Fig. 4 illustrates additional configurations that may be considered with our general model. The generator and receptor conductors are rectangular cross–section conductors typical of lands on PCB's. Fig. 4(a) illustrates a double–sided board wherein lands may be placed on both sides and connected through the board with *vias*. Fig. 4(b) illustrates a less–commonly used board, the single–sided board with a ground plane on one side that serves as the reference conductor. This is commonly referred to in the microwaves literature as the microstrip configuration. Fig. 4(c) illustrates the multilayer board wherein wiring channels are buried at various layers within the board and connected with vias. The general results that we will obtain apply to all these configurations.

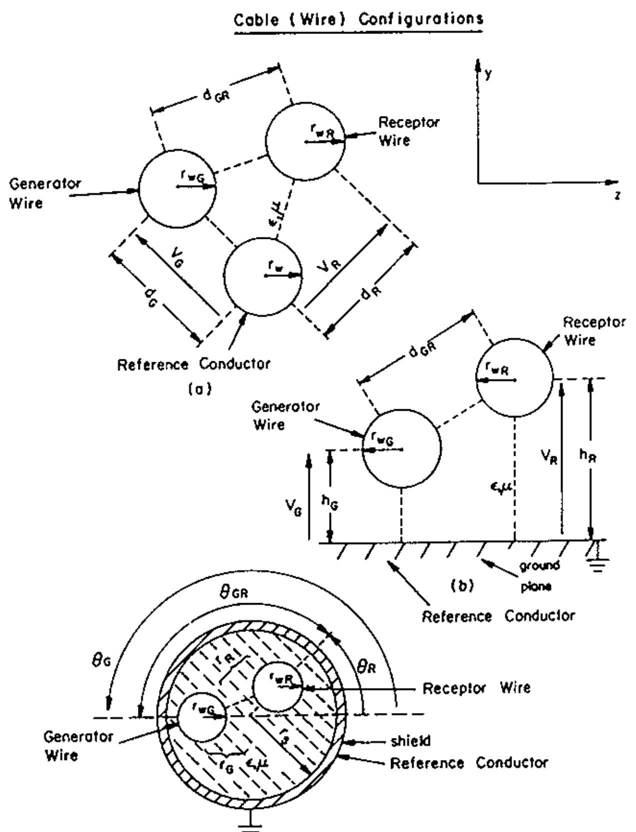

Fig. 3. Crossectional configurations of typical lines composed of wires in a homogeneous media

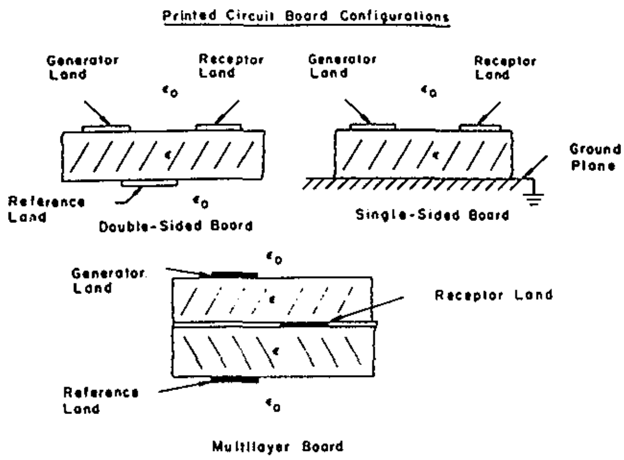

Fig. 4. Crossectional configurations of PCB lands

So long as the TEM mode of propagation is the only mode of propagation on the line, the *per–unit–length model* shown in Fig. 5 is a complete characterization of the coupling between the two circuits. For the moment, we will assume that all conductors are *perfect conductors* and that the surrounding medium is lossless. Therefore there are no losses in the line and it is said to be *lossless*. Later we will illustrate how to include losses. An electrically–small Δx section of the line is represented by the per–unit–length parameters of self inductance, l_G and l_R, and mutual inductance, l_m. The units of these are Henrys per meter and represent any section of the line since it is a uniform line. These parameters represent the effects of the transverse magnetic field. Similarly, the effects of the transverse electric field are represented by the per–unit–length parameters of self capacitance, c_G and c_R, and mutual capacitance, c_m. The units of these parameters are Farads per meter. These per–unit–length parameters are computed from static field considerations and apply for higher frequencies so long as the TEM mode of propagation is the only mode of propagation on the line [A.1,B.1]. The reader is referred to [B.1] and [C.1–C.7] for the computation of these parameters for the configurations of Fig. 3. Typically it is assumed that the wires are widely separated so that they can be replaced with filaments for this computation. For practical dimensions, this approximation is adequate and does not impose any significant restrictions on the applicability of the results. For the configurations shown in Fig. 4 wherein the conductors have rectangular cross sections, there are few closed–form results for the per–unit–length parameters and numerical methods must be employed to determine them. Although there are some closed–form equations for these self elements when the circuits are not coupled (l_G and c_G or l_R and c_R) these do not apply when the lines are in close proximity. Furthermore there are no similar closed–form results for the mutual elements (l_m and c_m). Essentially, all of the information about the

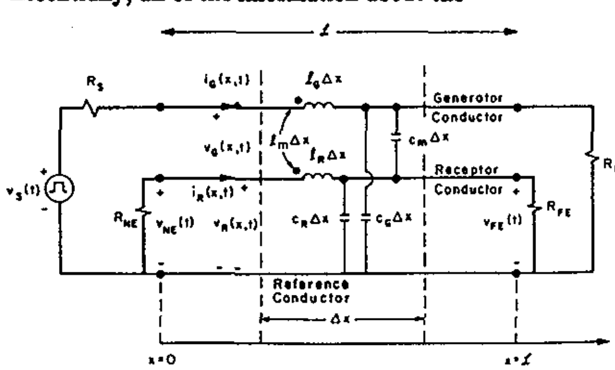

Fig. 5. The per–unit–length transmission–line model

cross–sectional dimensions that distinguishes one configuration from another is contained in these per–unit–length parameters. Therefore it is essential in a complete crosstalk prediction to be able to obtain these per–unit–length parameters for the specific configuration.

The *Multiconductor Transmission Line (MTL) Equations* that describe the coupling under the TEM mode assumption can be obtained from the per–unit–length equivalent circuit of Fig. 5 in the limit as $\Delta x \to 0$. These become a coupled set of first–order, linear, partial differential equations given by [A.1,B.1]:

$$\frac{\partial v_G(x,t)}{\partial x} = -l_G \frac{\partial i_G(x,t)}{\partial t} - l_m \frac{\partial i_R(x,t)}{\partial t} \tag{1a}$$

$$\frac{\partial v_R(x,t)}{\partial x} = -l_m \frac{\partial i_G(x,t)}{\partial t} - l_R \frac{\partial i_R(x,t)}{\partial t} \tag{1b}$$

$$\frac{\partial i_G(x,t)}{\partial x} = -(c_G+c_m) \frac{\partial v_G(x,t)}{\partial t} + c_m \frac{\partial v_R(x,t)}{\partial t} \tag{1c}$$

$$\frac{\partial i_R(x,t)}{\partial x} = c_m \frac{\partial v_G(x,t)}{\partial t} - (c_R+c_m) \frac{\partial v_R(x,t)}{\partial t} \tag{1d}$$

These can be written in a more compact form similar to the case of two–conductor lines as [A.1]

$$\frac{\partial}{\partial x}\mathbf{V}(x,t) = -\mathbf{L}\frac{\partial}{\partial t}\mathbf{I}(x,t) \tag{2a}$$

$$\frac{\partial}{\partial x}\mathbf{I}(x,t) = -\mathbf{C}\frac{\partial}{\partial t}\mathbf{V}(x,t) \tag{2b}$$

where the 2×1 vectors are

$$\mathbf{V}(x,t) = \begin{bmatrix} v_G(x,t) \\ v_R(x,t) \end{bmatrix} \tag{3a}$$

$$\mathbf{I}(x,t) = \begin{bmatrix} i_G(x,t) \\ i_R(x,t) \end{bmatrix} \tag{3b}$$

Matrices and vectors will be denoted by boldface. The 2×2 per–unit–length inductance matrix, **L**, and 2×2 per–unit–length capacitance matrix, **C**, are given by

$$\mathbf{L} = \begin{bmatrix} l_G & l_m \\ l_m & l_R \end{bmatrix} \tag{4a}$$

$$\mathbf{C} = \begin{bmatrix} (c_G+c_m) & -c_m \\ -c_m & (c_R+c_m) \end{bmatrix} \tag{4b}$$

The objective in crosstalk prediction is to solve these equations for the near–end and far–end voltages as $v_{NE}(t)=v_R(0,t)$ and $v_{FE}(t)=v_R(\mathscr{L},t)$ for a line whose total length is $\mathscr{L}$ and extends from $x=0$ to $x=\mathscr{L}$. This represents the *time–domain solution* since the complete waveforms of these voltages are desired.

The *frequency–domain solution* assumes the source voltage of the generator line is a sinusoid, i.e., $v_S(t)=V_S\sin(\omega t)$, and the line is in steady state. The above transmission line equations become for this case [A.1,B.1]

$$\frac{d}{dx}\hat{\mathbf{V}}(x) = -\hat{\mathbf{Z}}\,\hat{\mathbf{I}}(x) \tag{5a}$$

$$\frac{d}{dx}\hat{\mathbf{I}}(x) = -\hat{\mathbf{Y}}\,\hat{\mathbf{V}}(x) \tag{5b}$$

where the *phasor* voltage and current vectors are denoted as

$$\hat{\mathbf{V}}(x) = \begin{bmatrix} \hat{V}_G(x) \\ \hat{V}_R(x) \end{bmatrix} \tag{6a}$$

$$\hat{I}(x) = \begin{bmatrix} \hat{I}_G(x) \\ \hat{I}_R(x) \end{bmatrix} \tag{6b}$$

and the steady–state, time–domain voltages and currents are found as $v_G(x,t)=V_G(x)\sin(\omega t+\theta_{VG}(x))$, $v_R(x,t)=V_R(x)\sin(\omega t+\theta_{VR}(x))$, $i_G(x,t)=I_G(x)\sin(\omega t+\theta_{IG}(x))$, $i_R(x,t)=I_R(x)\sin(\omega t+\theta_{IR}(x))$ and the phasor solutions are $\hat{V}_G(x)=V_G(x)\underline{/\theta_{VG}(x)}$, $\hat{V}_R(x)=V_R(x)\underline{/\theta_{VR}(x)}$, $\hat{I}_G(x)=I_G(x)\underline{/\theta_{IG}(x)}$, $\hat{I}_R(x)=I_R(x)\underline{/\theta_{IR}(x)}$. The 2×2 per–unit–length impedance matrix, $\hat{Z}$, and 2×2 per–unit–length admittance matrix, $\hat{Y}$, are given by

$$\hat{Z} = j\omega L \tag{7a}$$

$$\hat{Y} = j\omega C \tag{7b}$$

where $\omega=2\pi f$ is the radian frequency of the source. The frequency–domain, phasor crosstalk voltages are $\hat{V}_{NE}=\hat{V}_R(0)$ and $\hat{V}_{FE}=\hat{V}_R(\mathscr{L})$.

Note that for frequency–domain responses, the partial differential equations reduce to ordinary differential equations. As such they are easier to solve using conventional state–variable techniques [B.1]. The partial differential equations characterizing the time–domain response are more difficult to solve. However, an exact time–domain model suitable for implementation in lumped–circuit analysis programs such as SPICE is described in [I.4] so that the time–domain solution is essentially solved for lossless lines. This exact SPICE model can also be used to determine the frequency–domain solution.

As an alternative, the solution to (6) for the frequency–domain solution may be obtained with direct FORTRAN codes in terms of the *chain parameter matrix* (*CPM*) as described in [B.1,I.1]. The result is

$$\begin{bmatrix} \hat{V}(\mathscr{L}) \\ \hat{I}(\mathscr{L}) \end{bmatrix} = \underbrace{\begin{bmatrix} \hat{\phi}_{11}(\mathscr{L}) & \hat{\phi}_{12}(\mathscr{L}) \\ \hat{\phi}_{21}(\mathscr{L}) & \hat{\phi}_{22}(\mathscr{L}) \end{bmatrix}}_{\hat{\phi}(\mathscr{L})} \begin{bmatrix} \hat{V}(0) \\ \hat{I}(0) \end{bmatrix} \tag{8}$$

The 4×4 chain parameter matrix, $\hat{\phi}(\mathscr{L})$, relates the voltages and currents at the near–end of the line, $\hat{V}(0)$ and $\hat{I}(0)$, to the voltages and currents at the far–end of the line, $\hat{V}(\mathscr{L})$ and $\hat{I}(\mathscr{L})$, and does not explicitly solve for these desired results. The terminal voltages and currents of the line can be determined once the terminal characterization of the line is specified, i.e., V_S, R_S, R_L, R_{NE}, and R_{FE} are specified. For the transmission line of Fig. 2, this can be accomplished by characterizing the left and right terminations as Generalized Thévenin Equivalents as [B.1]

$$\hat{V}(0) = \hat{V}_0 - \hat{Z}_0\,\hat{I}(0) \tag{9a}$$

$$\hat{V}(\mathscr{L}) = \hat{Z}_{\mathscr{L}}\,\hat{I}(\mathscr{L}) \tag{9b}$$

where

$$\hat{V}_0 = \begin{bmatrix} V_S \\ 0 \end{bmatrix} \tag{10a}$$

$$\hat{Z}_0 = \begin{bmatrix} R_S & 0 \\ 0 & R_{NE} \end{bmatrix} \tag{10b}$$

$$\hat{Z}_{\mathscr{L}} = \begin{bmatrix} R_L & 0 \\ 0 & R_{FE} \end{bmatrix} \tag{10c}$$

Substituting (9) into (8) gives [B.1]

$$[\hat{Z}_{\mathscr{L}}\hat{\phi}_{22}-\hat{Z}_{\mathscr{L}}\hat{\phi}_{21}\hat{Z}_0-\hat{\phi}_{12}+\hat{\phi}_{11}\hat{Z}_0]\hat{I}(0)=[\hat{\phi}_{11}-\hat{Z}_{\mathscr{L}}\hat{\phi}_{21}]\hat{V}_0 \tag{11a}$$

$$\hat{I}(\mathscr{L})=\hat{\phi}_{21}\hat{V}_0+[\hat{\phi}_{22}-\hat{\phi}_{21}\hat{Z}_0]\hat{I}(0) \tag{11b}$$

Once (11) is solved for the currents at x=0 (the near end) and at $x=\mathscr{L}$ (the far end), the voltages are obtained from (9) and the solution is complete.

It is worth noting that the above matrices and vectors are of dimension 2 for this three–conductor line. For a general multiconductor line consisting of (n+1) conductors, all of the above equations remain of the same form and only the dimensions of the matrices change (from 2×1 and 2×2 for the three–conductor line) to n×1 and n×n for the (n+1)–conductor line. These were solved in literal form for a three–conductor line in [B.5]. For n>2, computer solution methods must be employed.

It is important to emphasize, once again, that both the time–domain and the frequency–domain solutions for the general, lossless, (n+1) conductor line can be directly obtained using the SPICE circuit analysis program or other similar programs using the technique described in [I.4]. This approach has been used by the author and is highly recommended as an alternative to the above direct solution of the MTL equations.

III. The Per–Unit–Length Parameters

As indicated previously, the per–unit–length parameters are essential ingredients in the solution. For the wire–type configurations of Fig. 3 we may obtain so–called *wide–separation approximations* wherein the wires are assumed to be separated sufficiently from each other and the reference conductor such that the current and charge distributions around their peripheries are uniform; that is, *proximity effect* is not significant. Under this assumption, the per–unit–length inductances for the case of three wires in Fig. 3(a) are given by [B.1,C.4]

$$l_G = \frac{\mu_0}{2\pi}\ln\left[\frac{d_G^2}{r_{wG}\;r_w}\right] \tag{12a}$$

$$l_R = \frac{\mu_0}{2\pi}\ln\left[\frac{d_R^2}{r_{wR}\;r_w}\right] \tag{12b}$$

$$l_m = \frac{\mu_0}{2\pi}\ln\left[\frac{d_G d_R}{r_w d_{GR}}\right] \tag{12c}$$

where $\mu_0=4\pi\times10^{-7}$ H/m is the permeability of free space (assuming the medium surrounding the wires is not ferromagnetic). Similarly, for the case of two wires above an infinite ground plane in Fig. 3(b), the inductances are [B.1,C.4]

$$l_G = \frac{\mu_0}{2\pi}\ln\left[\frac{2h_G}{r_{wG}}\right] \tag{13a}$$

$$l_R = \frac{\mu_0}{2\pi}\ln\left[\frac{2h_R}{r_{wG}}\right] \tag{13b}$$

$$l_m = \frac{\mu_0}{4\pi}\ln\left[1+\frac{4h_G h_R}{d_{GR}^2}\right] \tag{13c}$$

For the case of two wires within an overall, cylindrical shield in Fig. 3(c), the inductances are given by [B.1,C.4]

$$l_G = \frac{\mu_0}{2\pi}\ln\left[\frac{r_S^2-r_G^2}{r_S r_{wG}}\right] \tag{14a}$$

$$l_R = \frac{\mu_0}{2\pi}\ln\left[\frac{r_S^2 - r_R^2}{r_S r_{wR}}\right] \tag{14b}$$

$$l_m = \frac{\mu_0}{2\pi}\ln\left[\left[\frac{r_R}{r_S}\right]\left[\frac{(r_G r_R)^2 + r_S^4 - 2r_G r_R r_S^2\cos\theta_{GR}}{(r_G r_R)^2 + r_R^4 - 2r_G r_R^3\cos\theta_{GR}}\right]^{1/2}\right] \tag{14c}$$

Very few similar closed–form approximations for the case of rectangular cross section conductors of Fig. 4 are available.

The per–unit–length capacitances for the wire–type configurations of Fig. 3 can be easily obtained if we assume that the medium surrounding the wires is *homogeneous.* Dielectric inhomogeneities such as wire insulation, strictly speaking, violate this condition but the configuration can be reasonably approximated as a homogeneous one by removing the insulations leaving free space having permittivity $\epsilon_0 = \frac{1}{36\pi}\times10^{-9}$ F/m. These are obtained using the reciprocal relation for a homogeneous medium [B.1]

$$LC = \mu\epsilon\, 1_n \tag{15a}$$

where the surrounding homogeneous medium is characterized by μ and ϵ and 1_n is the n×n identity matrix with ones on the main diagonal and zeros elsewhere. Therefore, the entries in the capacitance matrix can be found from the inductance matrix as

$$C = \mu\epsilon\, L^{-1} \tag{15b}$$

For the case of three–conductor lines of Fig. 3, n=2 and we have

$$\begin{bmatrix} (c_G + c_m) & -c_m \\ -c_m & (c_R + c_m) \end{bmatrix} = \mu\epsilon \begin{bmatrix} l_G & l_m \\ l_m & l_R \end{bmatrix}^{-1} \tag{16}$$

Defining

$$\Delta = v^2(l_G l_R - l_m^2) \tag{17a}$$

where the velocity of propagation of waves on the line is given by

$$v = \frac{1}{\sqrt{\mu\epsilon}} = \frac{3\times10^8}{\sqrt{\mu_r \epsilon_r}}\ \text{m/s} \tag{17b}$$

and μ_r and ϵ_r are the relative permeability and permittivity, respectively, of the surrounding homogeneous medium, we obtain from (16)

$$c_m = \frac{l_m}{\Delta} \tag{17c}$$

$$c_G = \frac{l_R}{\Delta} - c_m \tag{17d}$$

$$c_R = \frac{l_G}{\Delta} - c_m \tag{17e}$$

If the surrounding medium is not homogeneous, then the capacitance matrix with the medium present, C, and with it removed (replaced with free space), C_0, may be computed, and the inductance matrix is found from

$$L = \mu_0\epsilon_0\, C_0^{-1} \tag{18}$$

Typically, numerical methods must be used to compute C and C_0 [C.1–C.7]. This is typically the case for ribbon cables although ignoring the dielectric insulations gives reasonable results [E.1–E.3].

Again, these results for the per–unit–length parameters, although stated for the special case of a three–conductor line, can be generalized to the case of a (n+1)–conductor line in a homogeneous medium [B.1,C.4]. The forms of those results are very similar to the above results.

IV. Lumped–Circuit Iterative Techniques

In the past, lumped–circuit approximate models were employed to avoid the direct solution of the MTL equations. The advantage in doing so is that lumped–circuit analysis codes could be used to solve for the time– and frequency–domain responses directly. However these models are valid only for frequencies where the line is electrically short, e.g., $\mathscr{L} < \frac{1}{10}\lambda$ [A.1,B.1]. Typical such models are shown in Fig. 6. The *Lumped Pi* model of Fig. 6(a) has a structure that resembles the symbol π. Each of the per–unit–length inductances is multiplied by the total line length and lumped as single elements. Similarly the total capacitances of the line are split and placed at the beginning and end of the model. The *Lumped Tee* model of Fig. 6(b) is the dual and resembles the character T.

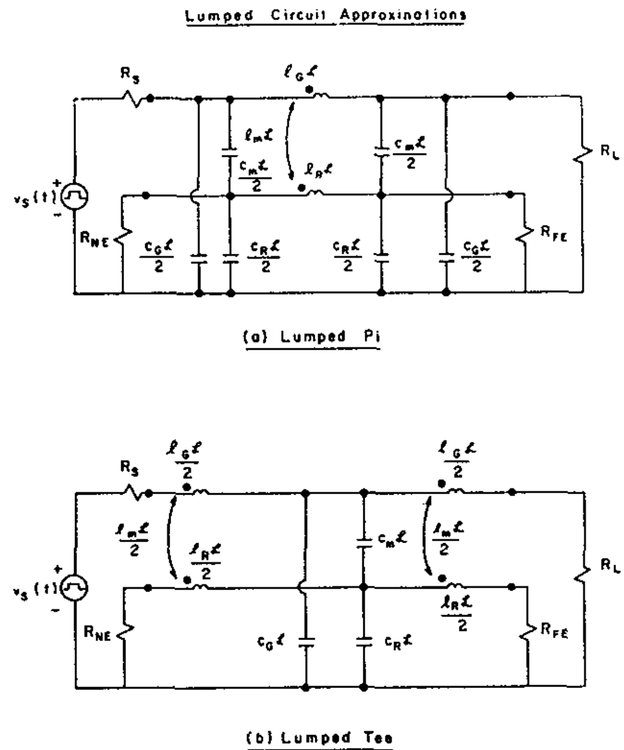

Fig. 6. Lumped–circuit approximate models

The advantage in using these models is that no knowledge of the solution of partial differential equations in (2) or ordinary differential equations in (5) is required; only computation of the per–unit–length parameters is required. However, the model is valid only for frequencies where the line is electrically short. For higher frequencies, the line is broken into sections each of which is electrically small, and each section is modeled with Lumped Pi or Lumped Tee models. This is the origin of the name lumped–circuit *iterative* models.

The natural question that arises is which of the two lumped models in Fig. 6 give better prediction accuracies. The answer is that the Lumped Pi model gives predictions that are valid over a larger frequency range than does the Lumped Tee if the termination impedances are "high impedance". Conversely, the Lumped Tee model gives predictions that are valid over a larger frequency range than does the Lumped Pi if the termination impedances are "low impedance". *The terms "low impedance" and "high impedance"* are with respect to the characteristic impedances of the circuits [B.2,B.13]. This is intuitively seen from the fact that the termination impedances are in parallel with the capacitances of the Lumped Pi model, whereas the termination impedances are in series with the inductances of the Lumped Tee model. Thus low impedance loads tend to render the capacitances of the Lumped Pi model ineffectual, whereas high impedance loads tend to render the inductances of the Lumped Tee model ineffectual.

V. Low–Frequency, Inductive–Capacitive Coupling Models

The MTL equations for a lossless, three–conductor line in a homogeneous medium and sinusoidal steady state given in (5) were solved in *literal form*; that is, in terms of symbols rather than numerical values of the elements, in [B.5]. This resulting solution showed that the solution could be reduced to a very simple result if the following three conditions were satisfied: (1) the lines are *weakly coupled*, (2) the frequency of interest is such that the line is very short, electrically, i.e., $\mathscr{L}<<0.1\lambda$, and (3) the termination impedances do not differ substantially from the line characteristic impedances. These criteria are not precise as shown in [B.7,B.10], and the frequency range of applicability of this simple model is a rather strong function of the termination impedance levels. Nevertheless, this very simple model yields crosstalk predictions that are sufficiently accurate over a wide range of useful frequencies for typical line dimensions and termination impedance levels.

The simple model of the receptor circuit is illustrated in Fig. 7. Fig. 7(b) illustrates the time–domain model and Fig. 7(a) illustrates the frequency–domain model. Both models depend on the DC or low frequency current and voltage of the generator circuit which can be approximated by

$$v_G(t) \cong \frac{R_L}{R_S+R_L} v_S(t) \tag{19a}$$

$$i_G(t) \cong \frac{1}{R_S+R_L} v_S(t) \tag{19b}$$

In other words, the frequency is sufficiently small that the effect of the generator line may be disregarded as can the effect of the receptor circuit on the generator circuit. For the frequency–domain response, the model of Fig. 7(a) gives

$$\hat{V}_{NE} = j\omega M_{NE} V_S \tag{20a}$$

$$\hat{V}_{FE} = j\omega M_{FE} V_S \tag{20b}$$

where the terms M_{NE} and M_{FE} are given by

$$M_{NE} = M_{NE}^{IND} + M_{NE}^{CAP} \tag{21a}$$

$$M_{FE} = M_{FE}^{IND} + M_{FE}^{CAP} \tag{21b}$$

and

$$M_{NE}^{IND} = \frac{R_{NE}}{R_{NE}+R_{FE}} l_m \mathscr{L} \frac{1}{R_S+R_L} \tag{22a}$$

$$M_{NE}^{CAP} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} c_m \mathscr{L} \frac{R_L}{R_S+R_L} \tag{22b}$$

$$M_{FE}^{IND} = -\frac{R_{FE}}{R_{NE}+R_{FE}} l_m \mathscr{L} \frac{1}{R_S+R_L} \tag{22c}$$

$$M_{FE}^{CAP} = \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} c_m \mathscr{L} \frac{R_L}{R_S+R_L} \tag{22d}$$

Therefore the frequency–domain crosstalk voltages increase linearly with increasing frequency (20 dB/decade). The time–domain solution is obtained from the equivalent circuit of Fig. 7(b) as

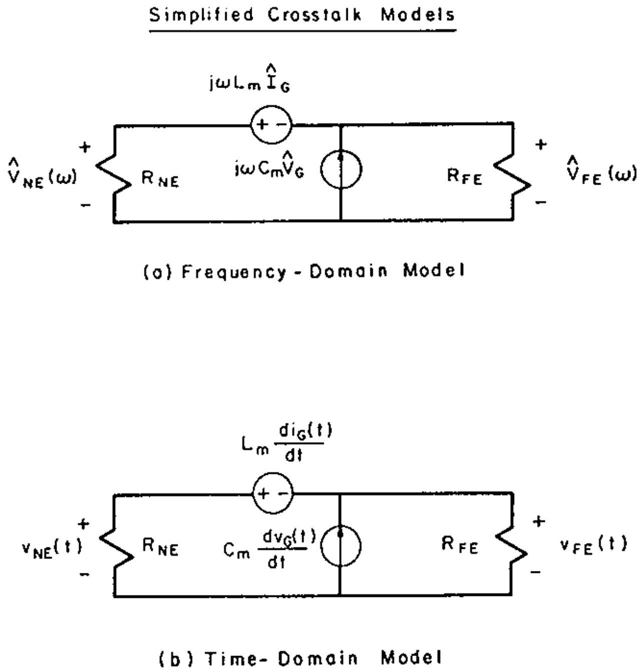

Fig. 7. Simplified inductive–capacitive coupling models

$$v_{NE}(t) = M_{NE} \frac{dv_S(t)}{dt} \tag{23a}$$

$$v_{FE}(t) = M_{FE} \frac{dv_S(t)}{dt} \tag{23b}$$

Therefore, the time–domain crosstalk voltages are direct functions of the instantaneous derivatives of the source voltage. Computed results will be given in the next section to illustrate the relative accuracy of this model.

Fig. 8 illustrates the frequency–domain variation of crosstalk voltage with frequency. The portions of the crosstalk coefficients in (22) which have the IND superscripts are referred to as the *inductive coupling contributions* whereas the portions which have the CAP superscripts are referred to as the *capacitive coupling contributions*. These are so named because they depend either on the *mutual inductance* or the *mutual capacitance* between the generator and receptor circuits. Observe in Fig. 8 that for "low–impedance" loads, the inductive coupling contribution dominates the capacitive coupling contribution. Conversely, for "high–impedance" loads, the capacitive coupling contribution dominates the inductive coupling contribution. In addition to being a simple crosstalk prediction model, this separation of the total crosstalk easily explains the effect of a shield or a twisted pair on the crosstalk as will be shown in later sections.

VI. Comparison of the Prediction Accuracies of the Three Models

We will now give an example comparing the relative prediction accuracies of the above three models for typical dimensions of a three–conductor line. The predictions of the three models are verified with experimental data in the publications listed in the references and will not be shown. For illustration we will consider the case of two wires above an infinite ground plane illustrated in Fig. 3(b). The wires are #22 gauge stranded ($r_{wG}=r_{wR}=12.65$ mils or .3213mm). Both wires are at a height above the ground plane of $h_G=h_R=1.5$cm. The wires are assumed to have insulation thicknesses of 15 mils and are touching so that the separation between the wires is $d_{GR}=55.3$ mils or 1.4mm. The total line length is $\mathscr{L}=5$m which is one wavelength at 60 MHz. The termination impedances are chosen to be equal to 50Ω: $R_S=R_L=R_{NE}=R_{FE}=50\Omega$. The predictions of all three

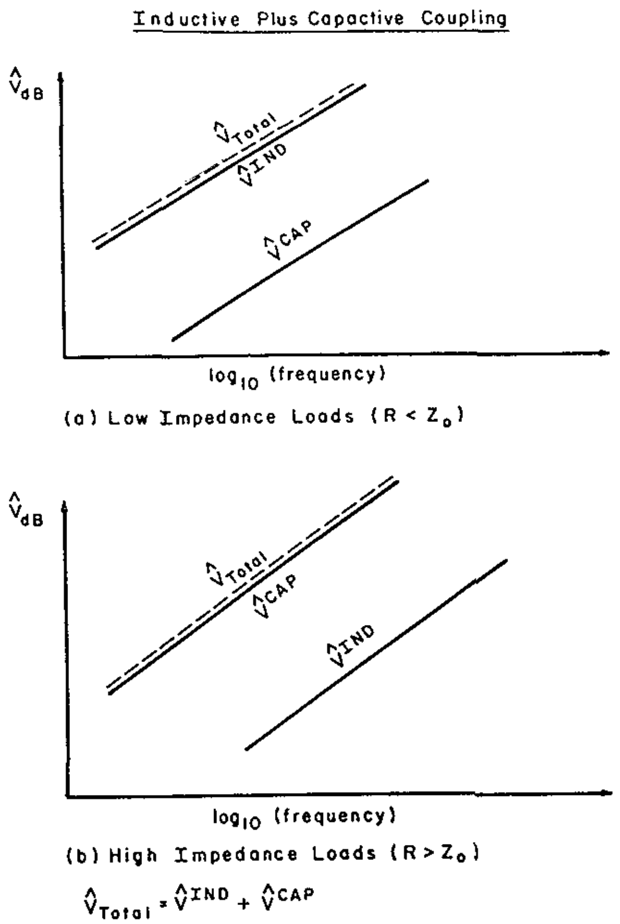

Fig. 8. Frequency response of inductive–capacitive coupling models

models (the MTL model, a Lumped Pi model, and the inductive–capacitive coupling model) are shown for the frequency domain in Fig. 9. The predictions of the transmission–line model [I.4] are denoted as VDB(NE1), the predictions of the Lumped Pi model are denoted as VDB(NE2), and the predictions of the inductive–capacitive coupling model are denoted as VDB(NE3). Observe that the MTL model and the Lumped Pi model predictions agree up to some 20 MHz where the line is $1/3\lambda$ long. The predictions of the MTL model and the inductive–capacitive coupling model agree up to some 4 MHz where the line is $1/15\lambda$ long.

Time–domain predictions are shown in Fig. 11. A trapezoidal pulse train typical of digital signals is used for $v_S(t)$ as illustrated in Fig. 10. The frequency of this pulse train is 500 kHz, the duty cycle is 50%, the transition voltage is 5V, and the rise/fall times of the pulse are $\tau_r=\tau_f=500$ns. The one–way time delay of the line is $T_D=16.7$ns. Therefore the pulse train has $\tau_r,\tau_f = 30\ T_D$. It is shown in [J.1,J.4] that in order for the inductive–capacitive coupling model to provide adequate predictions, $\tau_r,\tau_f>10\ T_D=\frac{10\,\ell}{3\times10^8}$. The crosstalk voltages appear during the transitions of the source voltage since they are related to the derivative of $v_S(t)$ according to (23). Therefore the larger the "slew rate" of the source, i.e., $\frac{dv_S}{dt}$, the larger the crosstalk pulse amplitude.

All of the predictions shown in Fig. 9 and Fig. 11 were computed using PSPICE, the personal computer version of SPICE. The SPICE MTL model is that described in [I.4]. The Lumped PI model is implemented using lumped inductor and capacitor elements of SPICE. The inductive–capacitive model used controlled sources to implement the models of Fig. 7. Programming all three models in SPICE in this fashion is quite simple.

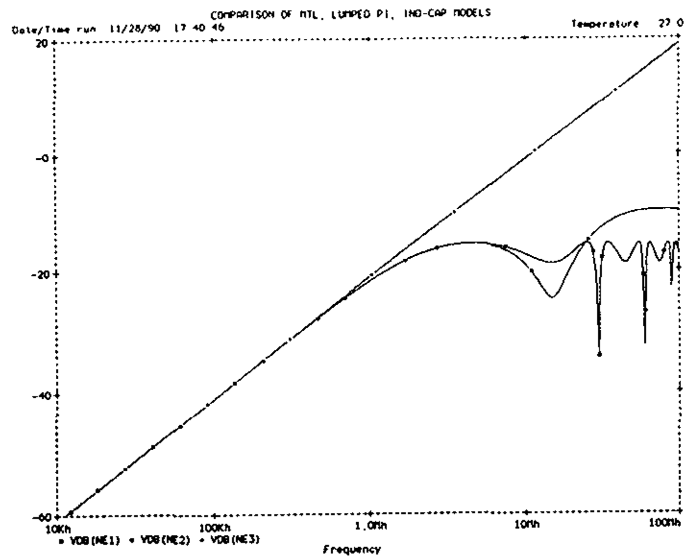

Fig. 9. Comparison of the frequency–domain predictions of the MTL, Lumped Pi, and inductive–capacitive coupling models

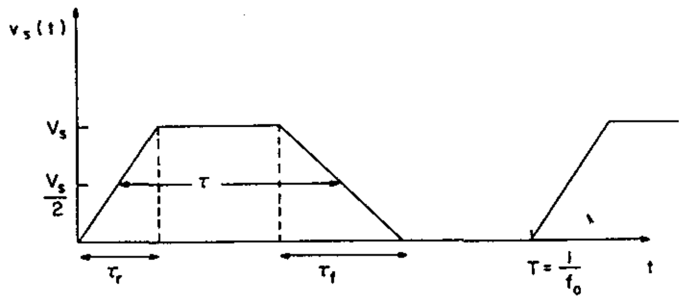

Fig. 10. A periodic, trapezoidal pulse train

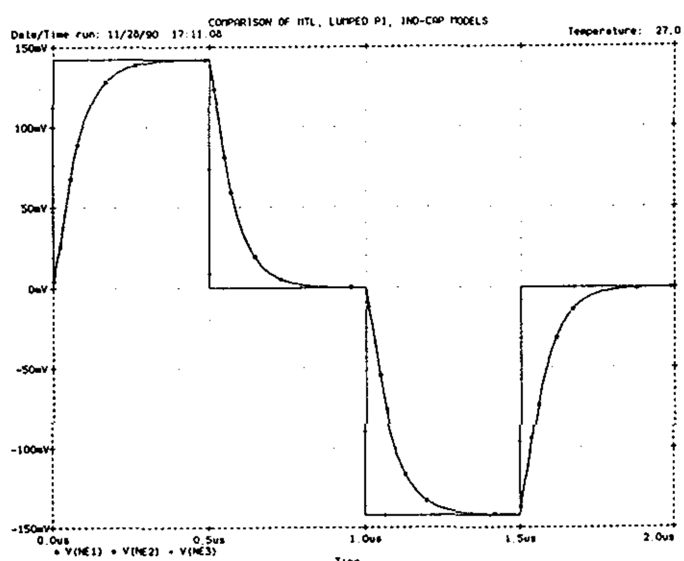

Fig. 11. Comparison of the time–domain predictions of the MTL, Lumped Pi, and inductive–capacitive coupling models for a trapezoidal pusle train

VII. Inclusion of Losses

The previous models assumed that the line was *lossless*; that is, the line conductors are perfect conductors and the surrounding medium is lossless. Assuming a lossless medium is typically a reasonable assumption for frequencies below the GHz range. However, there are cases where the assumption of perfect line conductors is not adequate. This generally occurs for lower frequencies and results in an additional contribution to the inductive and capacitive coupling contributions. This phenomenon is illustrated in Fig. 12. Suppose that we relax the requirement for perfect conductors in the case of the reference conductor, i.e., we assume that the reference conductor has a finite but nonzero conductivity giving it a resistance per unit length. Suppose we further lump that distributed resistance into one net resistance of the reference conductor, R_0. At low frequencies where the line is very short, electrically, the distributed effects of the line may be neglected and we may use lumped-circuit analysis. The phasor current of the generator circuit is approximately

$$\hat{I}_G \cong \frac{V_S}{R_S+R_L} \tag{24}$$

and approximately all of this flows through the reference conductor generating a voltage across it of

$$\hat{V}_0 \cong R_0 \hat{I}_G \tag{25}$$

This gives contributions to the near-end and far-end crosstalk voltages of

$$\hat{V}_{NE}^{CI} = \frac{R_{NE}}{R_{NE}+R_{FE}} R_0 \frac{V_S}{R_S+R_L} \tag{26a}$$

$$\hat{V}_{FE}^{CI} = -\frac{R_{FE}}{R_{NE}+R_{FE}} R_0 \frac{V_S}{R_S+R_L} \tag{26b}$$

These crosstalk contributions are referred to as *common-impedance coupling* and are denoted with the superscripts CI. These contributions are frequency independent and appear as a "floor" that limits the crosstalk as illustrated in Fig. 12. The total crosstalk can be approximated by the sum of the common-impedance coupling in (26) and the inductive-capacitive coupling in (20)-(22) for electrically-short lines.

This is, of course, an approximate way of including line losses, and in particular, imperfect line conductors. A more exact way of including the resistance of the conductors is to add a per-unit-length resistance matrix to the time-domain MTL equations in (2) or the frequency-domain MTL equations in (5) [B.1]:

$$R = \begin{bmatrix} (r_G+r_0) & r_0 \\ r_0 & (r_R+r_0) \end{bmatrix} \tag{27}$$

where r_G and r_R are the per-unit-length resistances of the generator and receptor conductors, respectively, and r_0 is the per-unit-length resistance of the reference conductor. The DC values of the resistances are usually sufficient since skin effect is not well developed over the low frequencies where these parameters substantially effect the total crosstalk.

VIII. Shielded Wires

Suppose the crosstalk for the above three-conductor line is of sufficient magnitude or spectral content to cause interference problems. There are typically two courses of action that are used to attempt to reduce this crosstalk to tolerable levels: (1) place a circular-cylindrical shield around either the generator wire, the receptor wire or both [F.1-F.7] or (2) replace the generator and/or the receptor conductors with a twisted pair of wires [G.1-G.10]. In the remaining sections of this paper we will discuss how and when the use of either of these methods will or will not reduce the crosstalk.

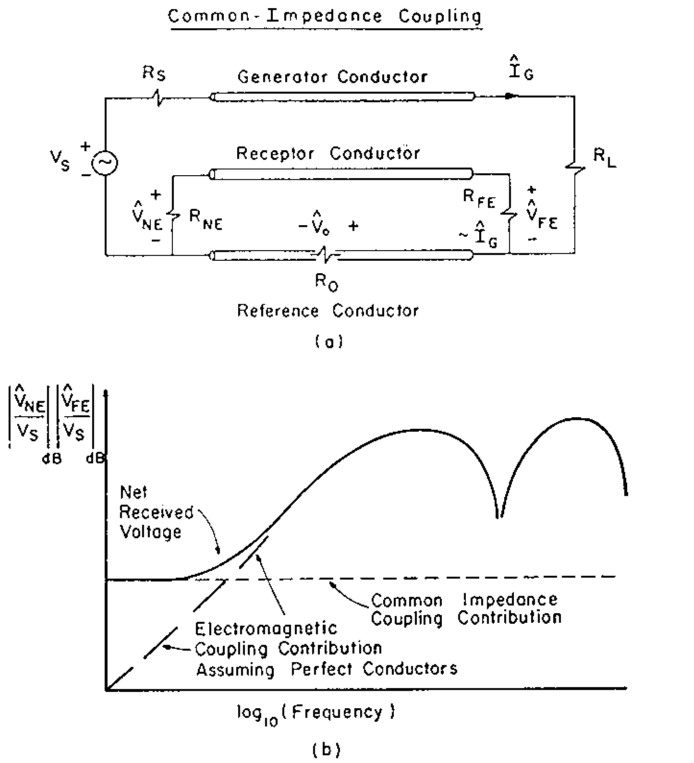

Fig. 12. Common-impedance coupling due to imperfect conductors

Consider placing a shield around the receptor wire of the two wires above a ground plane in Fig. 3(a). The shield could also be placed around the generator wire. We will examine the effect of the shield using the simple low-frequency, inductive-capacitive coupling model. The emphasis will be on the frequency-domain response. More exact MTL models can be formulated and solved as described in [F.1,F.2,F.4-F.7] and [I.2,I.3]. Also, the usual lumped-circuit iterative models such as the Lumped Pi and Lumped Tee models as in Fig. 6 can be easily adapted to handle this case. Although the inductive-capacitive coupling model covers a smaller frequency range than the MTL or lumped iterative models, it yields considerable insight and will be used for that purpose.

The capacitive coupling contribution is illustrated in Fig. 13. The self capacitances between the generator wire and the ground plane and between the shield and the ground plane are omitted since they provide second-order effects. The total mutual capacitance, C_m, is the per-unit-length mutual capacitance between the generator wire and the shield multiplied by the line length. This can be computed using the equations of section III, (13b), where we replace the radius of the receptor wire, r_{wR}, with the overall radius of the shield. The mutual capacitance between the shield interior and the receptor wire, C_S, is the per-unit-length capacitance of a coaxial cable multiplied by the line length [A.1]. This gives the equivalent circuit shown in Fig. 13. If the shield is not "grounded" (connected at either end to the ground plane), then the capacitive coupling becomes [F.1,F.3]:

$$\hat{V}_{NE}^{CAP} = \hat{V}_{FE}^{CAP} \cong \frac{R_{NE}R_{FE}}{R_{NE}+R_{FE}} j\omega C_m \frac{R_L}{R_S+R_L} V_S \tag{28}$$

which is essentially the same as with the shield removed since, for typical dimensions, $C_m << C_S$. Now suppose that the shield is grounded at either or both ends. If the line is electrically short, then the shield voltage, $\hat{V}_{SH}$, is essentially zero at all points along it and the capacitive coupling is reduced to zero. Thus, the shield reduces the capacitive coupling only if the shield is grounded at one or both ends.

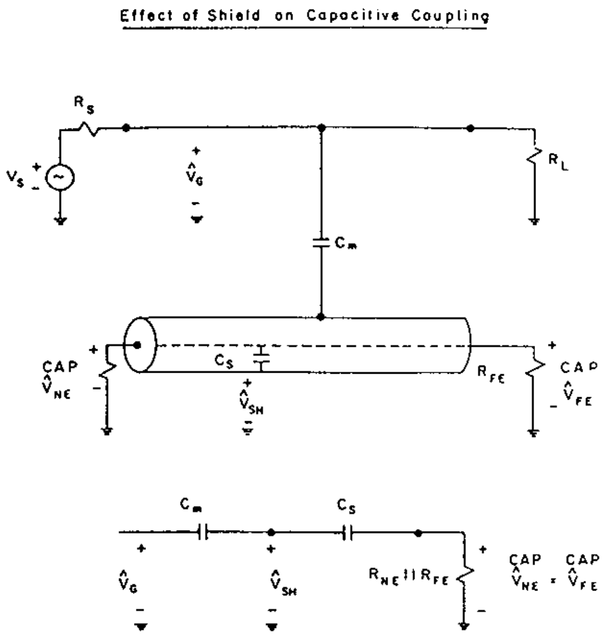

Fig. 13. Effect of a shield on capacitive coupling

Now let us consider the effect of the shield on inductive coupling. Consider Fig. 14. The generator circuit current produces a magnetic flux that threads both the receptor circuit and the circuit formed between the shield and the ground plane. This induces a current flowing back along the shield, $\hat{I}_{SH}$, which produces a counteracting flux that couples with and tends to reduce the net flux penetrating the receptor circuit. The equivalent circuit for the receptor and shield circuits is shown in Fig. 15. The generator current, $\hat{I}_G$, induces a source in the receptor circuit, $j\omega L_m \hat{I}_G$, and a source in the shield–ground plane circuit, $j\omega L_m \hat{I}_G$, and the two mutual inductances between the generator circuit and the receptor circuit and between the generator circuit and the shield circuit are equal because the receptor conductor is collocated with the shield. This mutual inductance can be computed from (13c). The self inductance of the shield–ground plane circuit, L_{SH}, can be computed from (13b) by replacing the receptor wire radius with the shield radius. The shield current induces a source, $j\omega L_{SH} \hat{I}_{SH}$, in the receptor circuit. It can be shown that the mutual inductance between the shield–ground plane circuit and the receptor–ground plane circuit are identical since the receptor wire is located on the axis of the shield [F.1–F.7]. The total resistance of the shield is denoted as R_{SH}. This circuit can be solved to yield

$$\hat{V}_{NE}^{IND} = \frac{R_{NE}}{R_{NE}+R_{FE}} j\omega L_m \frac{R_{SH}}{R_{SH}+j\omega L_{SH}} V_S \qquad (29a)$$

$$\hat{V}_{FE}^{IND} = -\frac{R_{FE}}{R_{NE}+R_{FE}} j\omega L_m \frac{R_{SH}}{R_{SH}+j\omega L_{SH}} V_S \qquad (29b)$$

For frequencies where $R_{SH} > \omega L_{SH}$ this expression reduces to that with the shield removed and the shield has no effect on inductive coupling for these frequencies. Above this frequency, $\omega L_{SH} > R_{SH}$, the inductive coupling becomes independent of frequency as illustrated in Fig. 16. All of this discussion of inductive coupling assumed that the shield was grounded at both ends to allow a shield current to flow back along the shield to produce this counteracting flux. If the shield is not grounded at both ends, no shield current can flow, and the shield has no effect on inductive coupling. Also, even though the shield is grounded at both ends, if the frequency of excitation is below $\omega_0 = R_{SH}/L_{SH}$, then the shield also has no effect

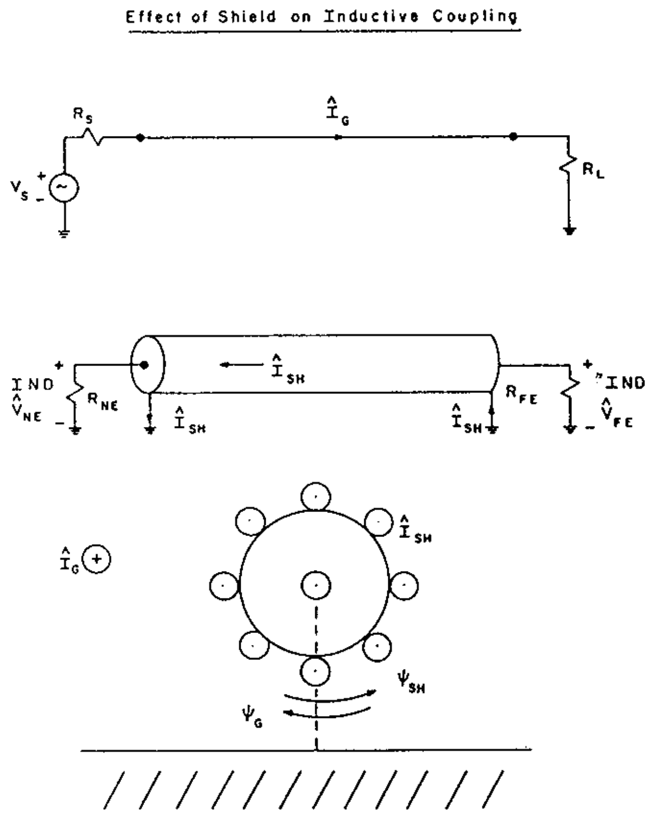

Fig. 14. Effect of a shield on inductive coupling

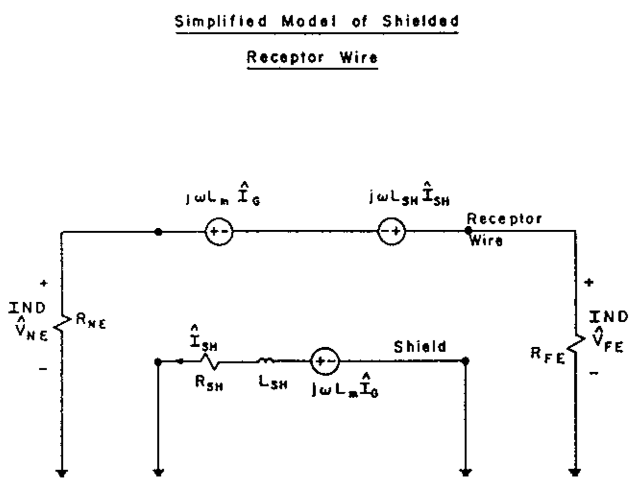

Fig. 15. Simplified model for inductive coupling to a shielded receptor wire

even if it is grounded at both ends.

This simple analysis explains when the addition of a shield will or will not reduce crosstalk. Consider Fig. 17. The notations in this figure denote the shield ungrounded at either end (OO), the shield grounded at the near end only (SO), the shield grounded at the far end only (OS), and the shield grounded at both ends (SS). The total crosstalk is again the sum of the inductive and capacitive coupling contributions. Suppose the termination impedances are "low impedances". Prior to the addition of the shield, the inductive coupling will dominate the capacitive coupling. If the shield is added (either around the generator wire or the receptor wire) and is grounded at only one end, only the capacitive coupling will be reduced, and since the total coupling is inductive, no reduction in the total coupling will be observed. If the shield is grounded at both ends, the capacitive coupling will again be removed but the inductive coupling will be reduced above $\omega_0 = R_{SH}/L_{SH}$. Thus for low–impedance loads a

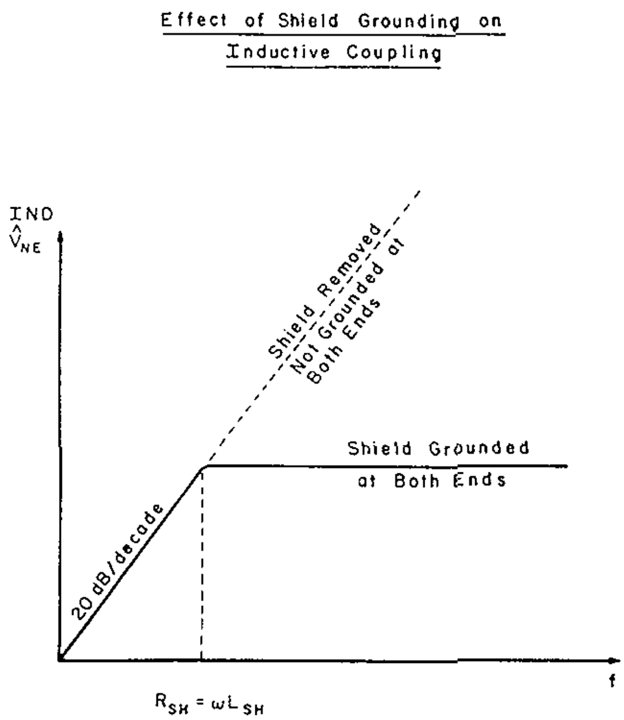

Fig. 16. Effect of shield grounding on inductive coupling for a shielded receptor wire

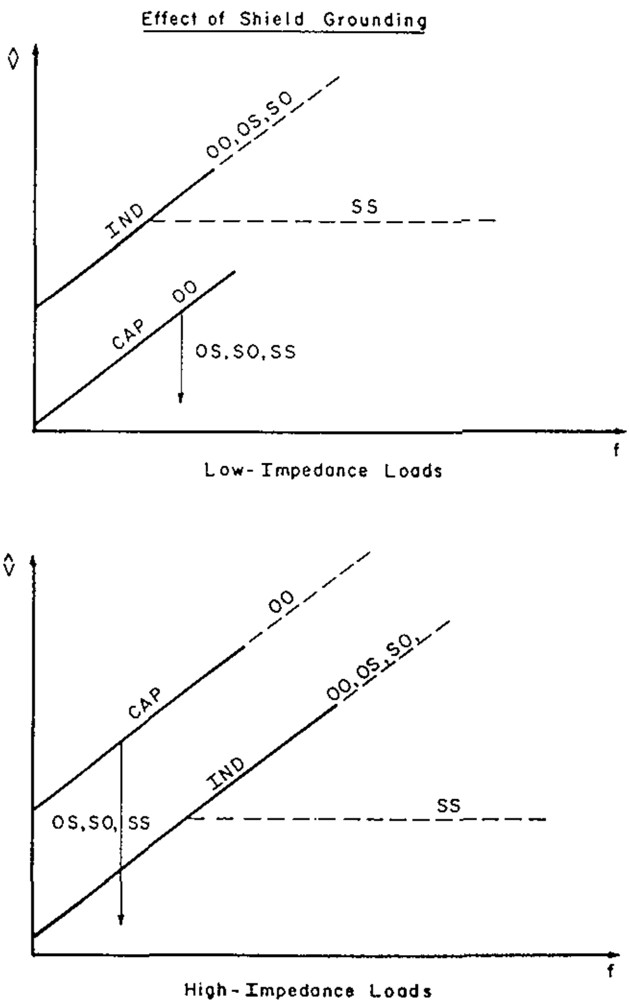

Fig. 17. Effect of shield grounding on crosstalk to a shielded wire

shield will reduce the total coupling only if the shield is grounded at both ends and only above ω_0. Conversely consider the case of "high–impedance" loads. Here the capacitive coupling dominates the inductive coupling before the shield is added. Suppose a shield is added to either the generator wire or the receptor wire and is grounded only at one end. The capacitive coupling is removed but since it was dominant the total coupling is reduced (to the inductive coupling level which is unchanged). Now if the shield is grounded at both ends, and the frequency is above ω_0, the total coupling will be further reduced above ω_0 since the inductive coupling is reduced. Thus for high–impedance loads, a shield will reduce the total coupling if the shield is grounded at at least one end and above ω_0 if it is grounded at both ends. Consequently, placing a shield around a wire may or may not reduce the total crosstalk and the above notions explain when they will or will not reduce crosstalk.

IX. Twisted Pairs

The second option is to replace either the generator wire or the receptor wire with a twisted pair. Consider replacing the receptor wire with a twisted pair. There are basically two configurations illustrated in Fig. 18. In the *unbalanced configuration*, the far end is ungrounded and one end of the near end is grounded. Both ends are usually not grounded in order to avoid ground loops. In the *balanced configuration*, each wire of the twisted pair sees the same impedance to ground. This is commonly implemented with center–tapped transformers (BALUNS) or with dual–input, dual–output operational amplifiers (line drivers and line receivers). We will concentrate on a simplified explanation of the coupling to twisted pairs using the inductive–capacitive coupling notions. The reader is referred to [G.1–G.10] for a more detailed discussion.

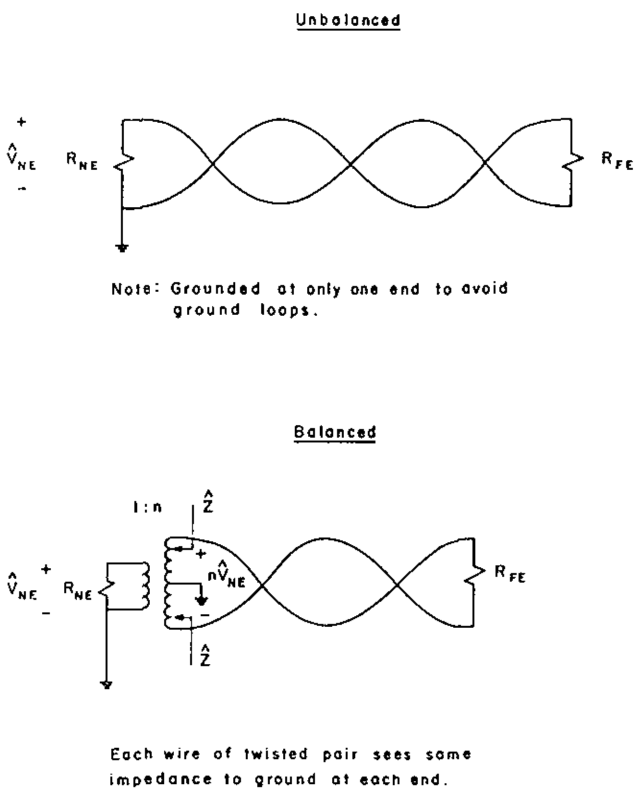

Fig. 18. Twisted pair configurations for unbalanced and balanced modes

The twisted pair inherently affects only the inductive coupling in the manner shown in Fig. 19. The current of the generator circuit produces a flux that threads the loops of the twisted pair of the receptor circuit. This induces opposing emf's in adjacent loops as illustrated in Fig. 20 where $\mathscr{L}_{HT}$ denotes the length of a "half twist". Similarly, the capacitive coupling is represented as current sources attached to each wire of the pair in the fashion of Fig. 7. If we "untwist" the twisted pair, we arrive at the model shown in Fig. 21. Observe that the emf's in adjacent loops tend to cancel leaving the net inductive coupling as that of one half twist. Thus the inductive coupling has been reduced by the twist. On the other hand, the current sources attached to the grounded wire are shorted out so that the total capacitive coupling is approximately unchanged. Thus the twist does not, in itself, reduce the capacitive coupling. In order to reduce the capacitive coupling, balanced loads such as in Fig. 18 must be used.

These notions are summarized in Fig. 22 for the case of an unbalanced twisted pair. This figure compares the difference in crosstalk between the receptor wiring being an "untwisted pair" and being a twisted pair. For the untwisted pair, either inductive or capacitive coupling will again be dominant. For "low–impedance" loads, assume that inductive coupling dominates capacitive coupling prior to twisting the pair of wires to produce the twisted pair. Twisting the wires reduces only the inductive coupling so that the total coupling drops to the capacitive coupling level of the original configuration (the untwisted pair). Conversely, for "high–impedance" loads, assume that capacitive coupling dominates inductive coupling prior to twisting the pair of wires to produce the twisted pair. Again twisting the wires reduces the inductive coupling component but does not substantially change the capacitive coupling component so that the reduction of the inductive coupling by the twist does not cause the total coupling to change. *Thus for the* **unbalanced case, use of a twisted pair will reduce the total crosstalk only for "low–impedance" loads.** Since balancing the loads will reduce capacitive coupling, for the balanced **case, use of a twisted pair will reduce the total crosstalk for both "low–impedance" and "high–impedance" loads.** References [G.1–G.10] contain experimental data that confirm these notions.

Effect of Twisted Pair on Inductive Coupling

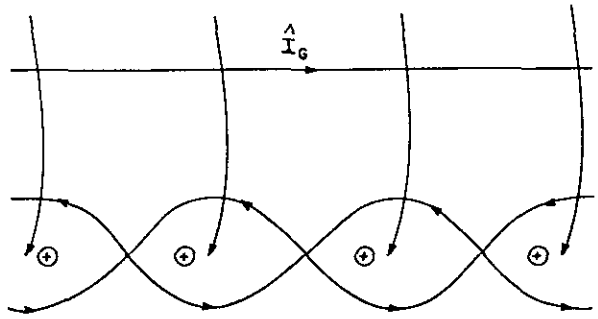

Fig. 19. Effect of a twisted pair on inductive coupling

Simplified Model of Twisted Pair

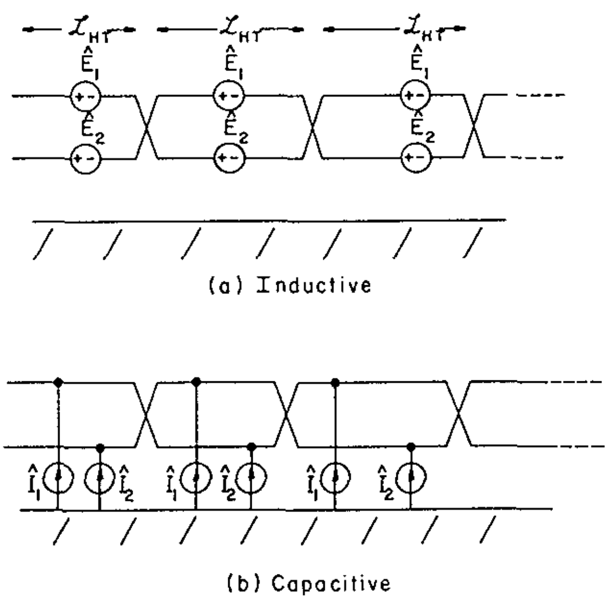

Fig. 20. Simplified model of coupling to a twisted pair

X. Summary and Conclusions

This paper has summarized the methods for predicting and reducing crosstalk. The fundamental prediction model is the distributed–parameter, MTL model whose fundamental assumption is that the TEM mode of propagation is the dominant mode of propagation on the line. Typically this is satisfied for lines whose crossectional dimensions are much less than a wavelength or for frequencies typically up to the GHz range. All other models are approximations to this model. The lumped–circuit iterative models attempt to approximate the MTL model for frequencies where the line length is electrically short. The applicable frequency range depends on the type of line. For cable bundles up to 5m in length, these models are valid up to some 10 MHz. For shorter cables they are valid to higher frequencies. For typical PCB's they are valid to hundreds of MHz [J.2].

Considerable insight is gained with the inductive–capacitive coupling model. This model is valid for lines that are very short, electrically, and are weakly coupled. These restrictions seem to be severe but for practical dimensions, they allow reasonable predictions over a significant frequency range and are amenable to hand calculations. The MTL and lumped–circuit iterative models generally must be solved with digital computer codes. The simple inductive–capacitive coupling models allow a simple explanation of when the addition of a shield or a twisted pair will or will not reduce crosstalk.

And finally, it is worth pointing out that most of these concerns with crosstalk would be eliminated if the metallic conductors were replaced with fiber–optic cables. There is certainly virtually no crosstalk between these types of cables. Any and all coupling will take place at the ends where the "photons are converted to electrons" to adapt to conventional signal processing. However, it is unrealistic to expect that fiber optic cables will solve the crosstalk problem in the near future since metallic–conductor cables are in considerable use, and a complete conversion to fiber–optic cables is not realistic to expect. Some use of fiber optic cables is being made in the newer weapon systems, but there is a large inventory that uses metallic conductor cables which are not likely to be replaced in any great volume. Therefore we will continue to have to deal with this problem of crosstalk between metallic–conductor cables for some time in the future.

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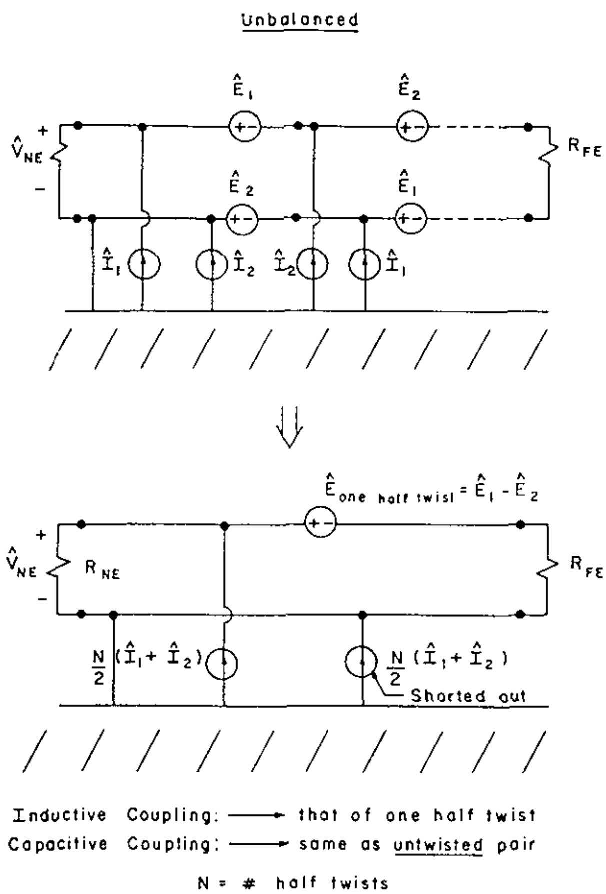
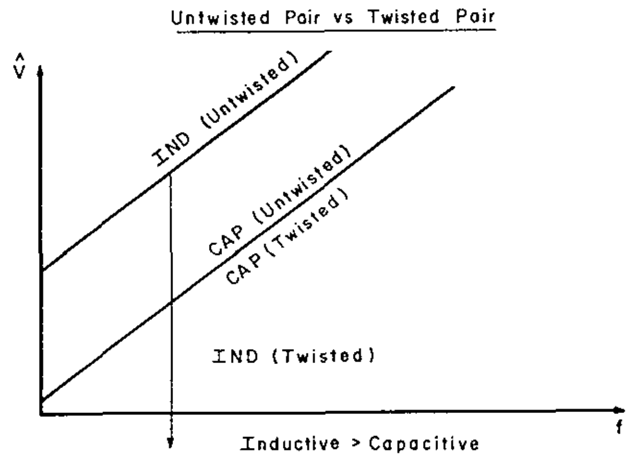


Fig. 21. Untwisting the twisted pair



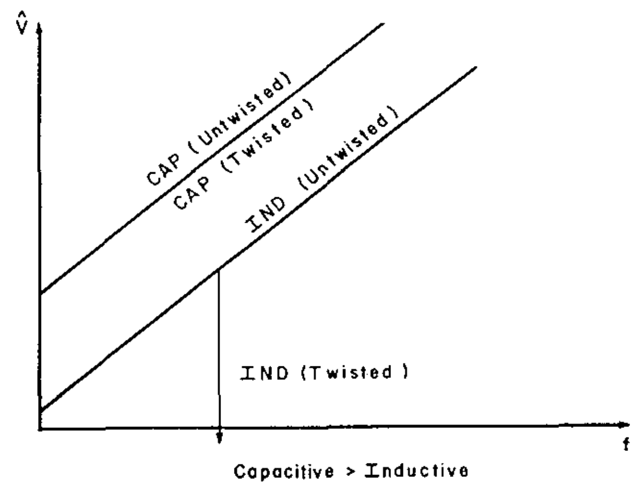


Fig. 22. Effect of twist on inductive and capacitive coupling for an untwisted pair compared to a twisted pair

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GROUNDING, SHIELDING AND BONDING

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In the EMC-design of systems and circuits, both grounding and shielding are related to the coupling mechanisms of the system with (radiated) Electromagnetic Fields. Grounding is more related to the source or victim circuit (or system) and is determining the characteristic of the coupling mechanism between fields and currents/voltages.
Shielding is a way of interacting in the radiation path of on electromagnetic field. In this text, basic principles and practical design rules will be discussed.

0. Introduction

Considering an electronic system, and looking for the coupling mechanisms with the "outside" world, two important coupling paths from circuit level into ambient fields (or vice-versa) may be distinguished. This is sketched in fig. 0.1.

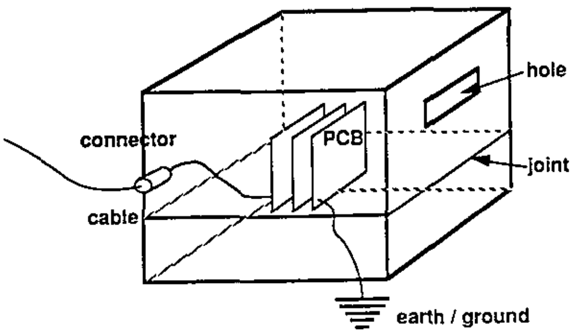


Fig. 0.1. : Coupling of systems to fields.

It should be mentionned that "system" means any electronic system or circuit. Referring to fig. 0.1., the sketched PCB's are again a system by thimselves. And the given system can be a subassembly in a larger context. This means that all characteristics and design rules and conclusions are valid at all levels of an electronic circuit. As an example should be mentionned that coupling characteristics of a couple of wires are exactly determined by the same physical properties as it is for tracks on a PCB. The same may be said for shielding subassemblies in a larges system compared with the shielding of the whole system. So, general conclusions have to be translated every time into the typical parameters at the apropiated level of the design.

1. Grounding

In electronics, grounding is a very general word and concept, and is used to describe a lot of techniques. The only common item in all applications is that a system is connected to the point of the reference potential used in the system. For a lot of applications, this means the "earth", but not any time.

1.1. What not ?

In electronic design, grounding is used for a lot of targets. First of all, it is used as the conductor for the return cirrent of data and signals or of the power system. When the same wire or track is used for the return-current of different subsystems, problems may occur because of these currents causing a voltage drop over the impedance of the wire (fig. 1.1). This effect is called common impedance and will be discussed in another paper.

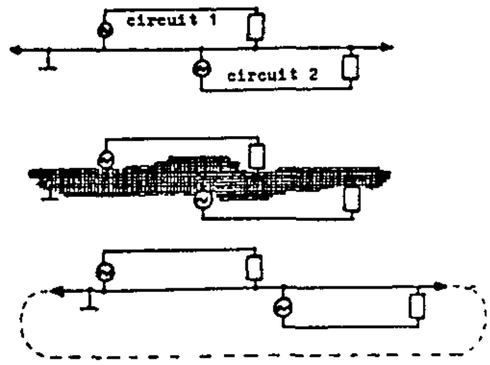


Fig. 1.1. : Common mode impedance coupling.

The presence of a ground wire or plane is also influencing the characteristic impedance of a couple of wires. The value of this characteristic impedance is important for matching sources and loads, or to a avid us wanted reflections (ex. in binary bus systems). This problem is discussed in another paper. See fig. 1.2.

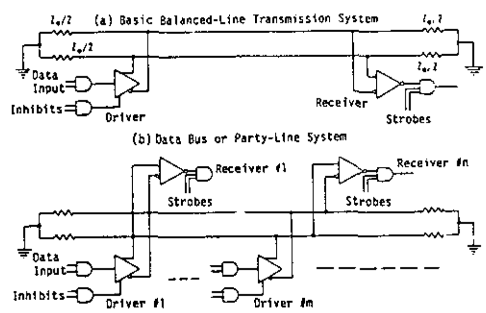


Fig. 1.2. : Characteristic impedance for matching bus systems.

And thirdly, the presence of a ground return wire is influencing the crosstalk between wires or tracks. Fig. 1.3. sketches crosstalk geometry. Also this effect is discussed in other papers.

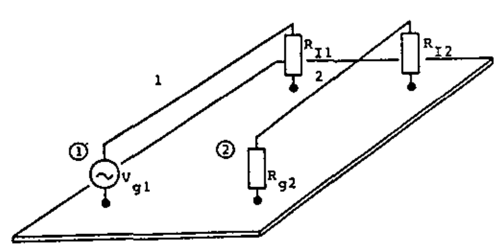


Fig. 1.3. : Xtalk geometry of wires or tracks.

Grounding is a very important tool to avoid problems from the mentionned effects. However, grounding is also very important in the coupling of systems to electromagnetic fields.

1.2. Ground loops.

As might be seen from fig. 1.4., connecting two systems together causes some current flowing from the first to the second one. Depending on the convection mode used, the return current is flowing over the earth/ ground (fig. 1.4.), or through a well defined return wire.

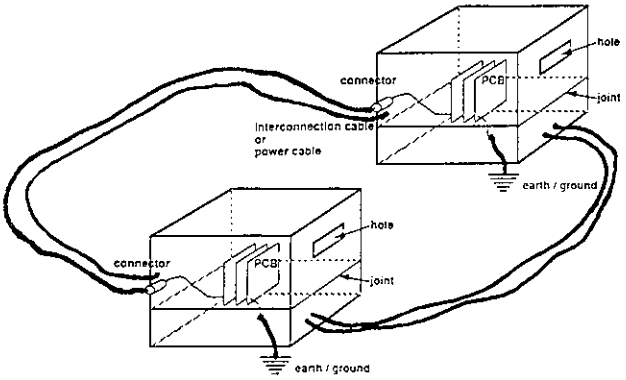


Fig. 1.4. : Return current flowing over earth.

In any case, a kind of loop is created where the same current is flowing through. This generates a magnetic field, which is radiated in the ambient. Otherwise, a noosy electromagnetic ambient induces in such a loop an induced voltage. For both cases (emission, reception), characteristics of the coupling mechanism may be derived from the theory of loop antenna's. From electromagnetic wave theory, it follows (see fig. 1.5) :

- emission : near field $H \doteq 1/\lambda^3 .A.I$

  far field $H \doteq f^2 . \frac{1}{d} A.I$

  $\doteq f \frac{1}{d} A.E$

- reception : $E_i \doteq \mu_o \; A \; dH/dt \doteq \mu_o \; A\omega E/120\pi$ (Plane wave)

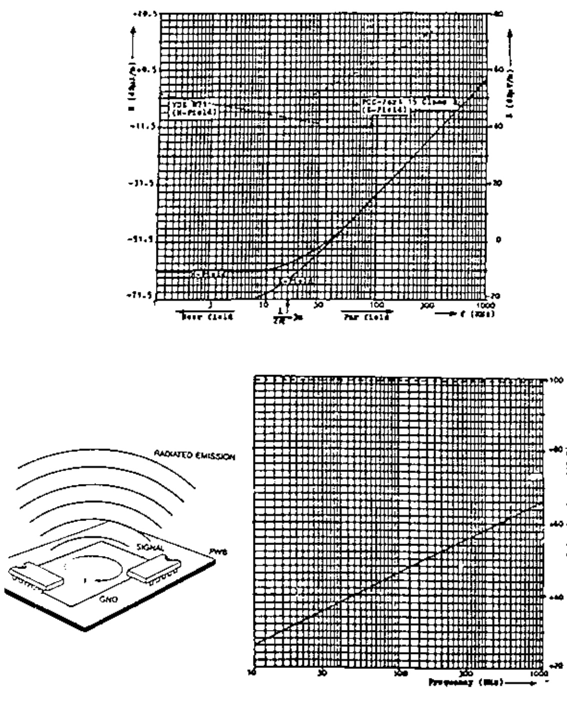


Fig. 1.5.: Radiation from a loop:
- H/E field radiation at distance of 3m from a 1 cm^2 loop/1 m A
- induced voltage in 1 cm^2 loop by plane wave 1 V/m

For the emission problems, it is easily seen that 3 important parameters are influencing the radiated level, frequency f, current I (or voltage E) and the loop area A. Both, frequency f and current-level I are directly related to the electronic design choice of clock , logic family, power consumption, disc of decoupling capacitors, ...
The loop area has only to do with the "mechanical" design, i.e. layout on PCB, wiring of subassemblies, cable connection between systems,... . The area to be considered is the area of the loop formed by the wire (on PCB-track) carrying the current and the ground wire (track, plane) carrying the return current. An example is given in fig.1.5.
Also for the susceptibility or receiving problem, nearly the same parameters are coming in : field level (E or H), frequency and loop area. The difference is however that field level and frequency are coming from the "unknown" ambient and are not directly related to the system under consideration. The only design parameter which may be taken into account, and is under the control of the system designer, is the loop area A. The fact that designers have to deal with unknown ambient levels makes the susceptibility problem harder than the emission one.
For both cases may be concluded that the problems are not only depending on the electronic design itself (ex. choice of components), but also on the mechanical design, normaly the generation of current loops. Because the ground is used to carry the return current (in asymmetrical system), the physical emplacement of the ground is the dominant factor in this loop coupling mechanism between fields (E/H) and the system (V/I).
However the effect is depending of the transmission mode used in a system (Common mode/ Differential mode). But as a general conclusion may be stated that loop area's should be minimised, and a good choice of the routing of the grounding wires or planes is crucial for the good working of a system.

1.3. Common mode/ Differential mode

Systems are reacting to the ambient in a different way, following a common or differential mode coupling. This will be discussed in detail in another paper. However, some characteristics are summarised. In fig. 1.6, basic concepts of CM/DM are sketched, and also the coupling area with the ambient. It is clear that for a normal concept of systems, the CM-loop area is dominant.
Minimising this loop area is one of the very important design parameters in circuits and systems by routing apropiately the ground-reference.

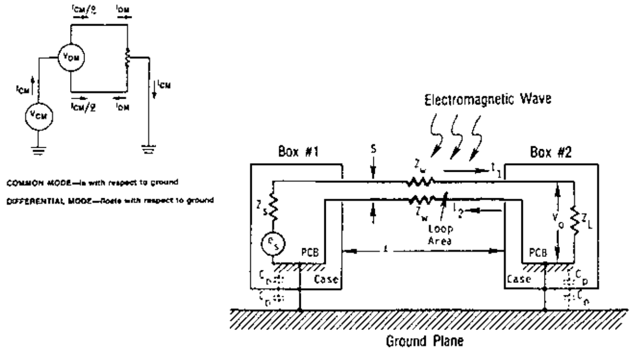


Fig. 1.6. : CM/DM coupling.

It should be noted that on a PCB, a CM system is used for routing, even bringing in a common impedance problem. For DM-like systems, an unwanted ground loop may induce an interfering signal, due to unbalances in the DM-system. Opening such "unwanted" loops may be done by "ungounding" systems (ex. not connecting PCB's to the shielding box). Minimising loop area's may be achieved by using good, large ground planes, with a well known position referred to the system. An example is given in fig. 1.7. for a small PCB and for a large computer system.

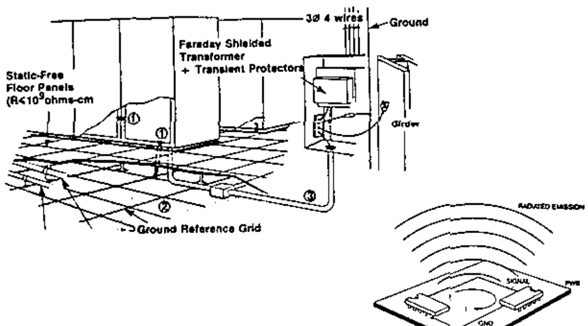


Fig. 1.7. : Ground planes.

1.4. Other grounding rules.

In the sections above, a lot of grounding rules were mentionned with respect to the electronic system itself. Other reasons for grounding (or referencing to earth) are related to safety and shielding. It should also be noted that grounding is not always connecting a system to the earth. Some reasons for grounding are :

- Shock and safety hazard control (earthing):
  - lightning
  - power system failures
- EMI-control :
  - control of a current discharge parth for ESD
  - faraday-shield return reference
  - common-mode loop control.

When looking for the last item, this may be in conflict with the other requirements for safety (ex. opening a ground-loop). Safety, shock and lightning require a low-impedance connection to the earth, and even a good distribution over the system of this earth-reference.
As a general rule, it should be stated that there should be NO mixing of grounding/earthing connections related to different topics. As a typical example can be mentionned the direct dischange current of an ESD-hazard. This current may flow through the reference ground of a.system, causing dammage or at least common impedance induced voltage (see fig. 1.8.).

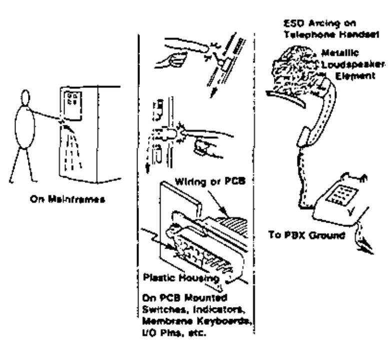


Fig. 1.8.: ESD discharge paths.

For the control of hightning and ESD-effects, and also for the faraday-shielding (of transformers, ...), it is essential for a good working that the connection between parts of the system and the ground/earth is done at a really low impedance level. This problem is known as banding. It is referred to section III for the discussion of banding techniques.

1.5. Conclusions.

Grounding is a very important action in electronic system design, because of the influence of the ground in different topics related to:

- safety and shock-protection
- lightning and ESD-control
- faraday-shielding
- common mode loop control
- Xtalk and transmission-line related characteristics

As a common requirement, for all applications, is the need of a low impedance ground reference. This means not only a low resistance, but also a low inductance for both the ground reference and the banding wires. This is a hard requirement for ambients and interference signals with a spectrum holding high frequencies.
And as a general rule, all grounding and earthing references and connections may not be mixed when they are related to different topics, requirements, systems or subassemblics.
In practica, this means a seperate ground reference for critical signals for analog and digital signals, for noisy systems, ...
All these seperated wires shouldn't only be connected at the REAL reference point of the system (Single point/Star grounding).
And as a last point, it should be mentionned that the mechanical design and layout is crucial for the good working of a system :

- ground-reference acting as the return current conductor, creating a ground loop (coupling to the ambient EM fields).
- crosstalk, shielding and characteristic impedance of wires /cables/tracks on PCB's, backplanes and subsystems.
- controlled routing of discharge currents of ESD/lightning and safety earth.

A typical example is the pigtail connection to cable shields, ESD-earths, safety earths of main filters, ... creating a loop-antenna inside the system to protect and causing in this way a new source of EMI problems instead of offering a solution, even using a shielded box.

2. Shielding.

Tackling interference problems may be done at 3 levels : by acting on the source level by reducing the emission levels and spectrum. By acting on the victim level by reducing the susceptibility (or increasing the immunity). And by acting on the transmission path. This latter action is depending on the coupling path. For conduction problems, the use of filtering techniques must be considered. For radiation problems, the use of shielding techniques is required.Because shielding is directly related to radiation, electromagnetic field theory is concerned with. In fig. 2.1., an example is given of a radiated spectrum, compared with a typical emission-level standard.

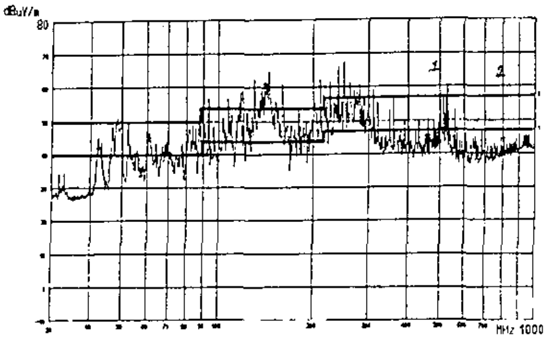


Fig. 2.1. : Emission spectrum from a PCB.

2.1. Basic Shielding Theory

Basic shielding theory is starting from maxwell equations. Depending on the model used for the shielding material or housing, three different techniques are used in practice :

- Schelkunoff, using a transmission-line like model and an infinite flot panel of shielding material.
- Koden, using a uniformous field distribution for closed boxes
- circuit theory, using induced currents in the materials and an equivalent circuit diagram.

Referring to practical box design, Koden and circuit theory are giving best theoretical results. But referring to the measuring techniques for evaluating the Shielding Effectiveness of materials, the Schelkunoff-theory is best fitting to these measurement methods.
When a wave is impinging on a barrier, a reflection is generated.
Part of the wave is transmitted through the barrier. During this transmission, the wave may be attenuated. Multiple reflections may be generated, due to the second transition of the barrier to the aim.
Global attenuation or shielding value will be a combination of these 3 effects :

- first reflection (R)
- multiple reflections (B)
- attenuations (A)

This is sketched in fig. 2.2.

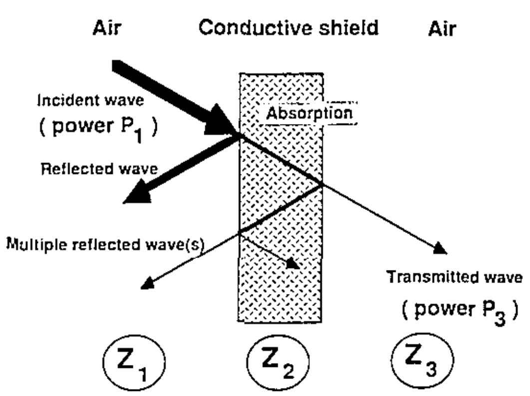


Fig. 2.2. : Shielding of materials.

Shielding effectiveness (SE) is defined as the ratio of the field strength in a point II without and with the shielding material. This refers to the definition of insertion loss (I.L.) in circuit theory.

Starting from Maxwell theory, it may be calculated using Schelkunoff theory that global shielding is given by :

$$SE_{(dB)} = R_{(dB)} + A_{(dB)} + B_{(dB)}$$

where : $R = -20 \log \| (1+\rho)(1-\rho) \|$

$A = 8.68\ t \sqrt{\mu\sigma\omega/2} = 8.68\ t/\delta$

$B = 20 \log \| 1 - \rho^2 \exp(-2t/\delta) \|$

δ = skin depth = $\sqrt{2/\mu\sigma\omega}$

$\omega = 2\pi$ frequency

σ = conductivity of the material

$\mu = \mu_o . \mu_r$ = permeability of material

ρ = reflection coefficient = $\dfrac{Z_{metal} - Z_{wave}}{Z_{metal} + Z_{wave}}$

For materials where $t \ll \delta$, $SE = R + B$
$t > \delta$, $SE = R + A$

Note that δ is frequency dependant.

As in classical transmission line theory, the reflection coefficient is function of the ratio of characteristic impedances in the material. Electromagnetic wave theory gives :

$$Z_{wave} = \sqrt{\frac{\mu_o}{\varepsilon_o}} = 120\pi = 377\,\Omega = Z_{air} \text{ for far field(plane wave)}$$

$$Z_{wave} = \frac{\lambda}{2\pi r} \sqrt{\frac{\mu_o}{\varepsilon_o}} \text{ for near field, E-field}$$

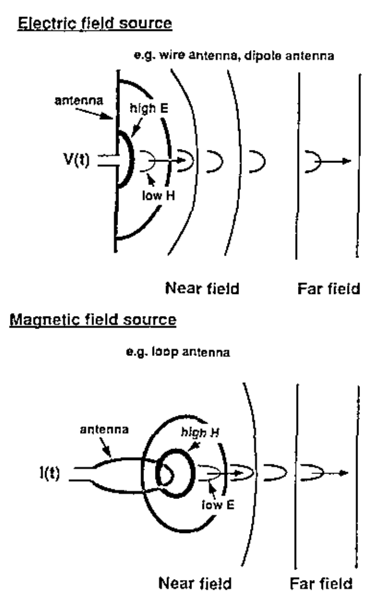
$$Z_{wave} = \frac{2\pi r}{\lambda} \sqrt{\frac{\mu_o}{\varepsilon_o}} \text{ for near field,H-field}$$

In general, Z_{wave} is noted as $Z_{wave} = k\sqrt{\dfrac{\mu_o}{\varepsilon_o}}$

$$Z_{metal} = \sqrt{\frac{\omega\mu}{2\sigma}}\,(1+j) \approx (1+j)\,\frac{1}{\sigma\delta}$$

The distinction between E/H field conditions are made for near field conditions, i.e. the distance r from source to barrier is smaller than $\lambda/2\pi$ (λ= wavelength). This is discussed in another paper.

The wave impedance is sketched in fig. 2.3 as a function of frequency, distance to the source and type of source.



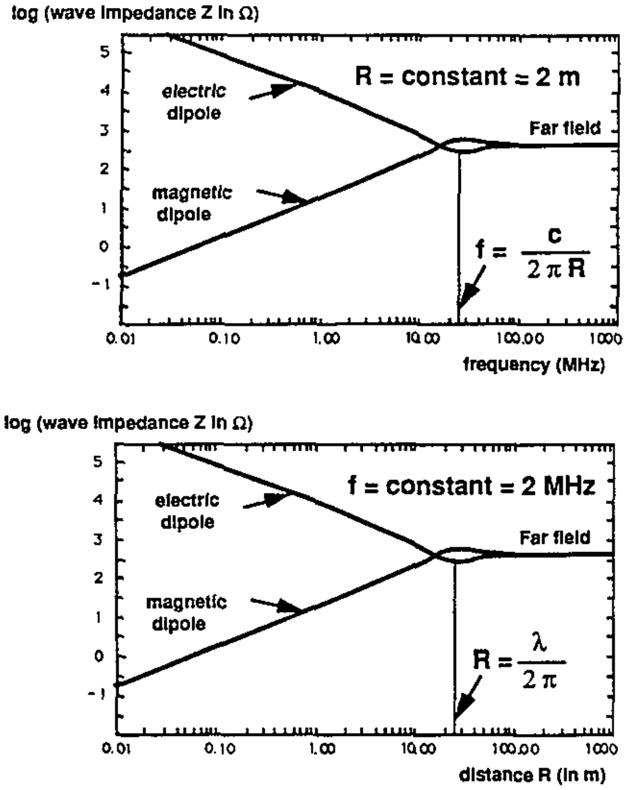


Fig.2.3.: Wave impedance.

Given the above expressions, it follows that for the SE-value SE = R + A + B, all three components *are depending on some parameters* :

* Reflection loss (R)= $-20 \log \|(1+\rho)(1-\rho)\| =$

$$\frac{4\, Z_{metal}\, Z_{wave}}{(Z_{metal} + Z_{wave})^2}$$

$$\rho = \frac{Z_{metal}/Z_{wave} - 1}{Z_{metal}/Z_{wave} + 1}$$

$$\frac{Z_{wave}}{Z_{metal}} = \frac{k\sqrt{\mu_o/\varepsilon_o}}{(1+j)\sqrt{\omega\mu/2\sigma}}$$

It follows that R is a function of

- distance source to barrier (near field/ far field)
- conductivity σ of the material
- permeability of the material
- frequency ω

R is not directly related to the thickness of the material.
An example of the *Reflection component* is given in fig. 2.4.

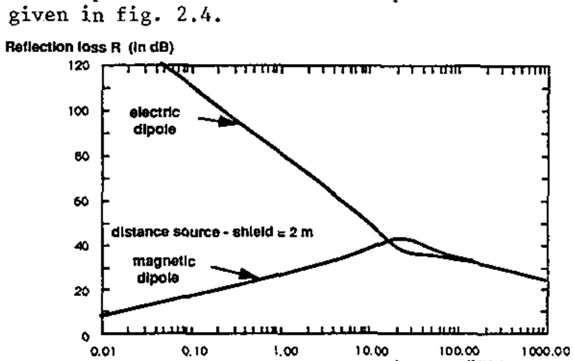


Fig. 2.4. : Reflection loss R as function of frequency

* Absorption loss (A) = 8.68 t/δ
t = thickness of the material

δ = *skin depth* =

$$= \sqrt{\frac{2}{\mu\sigma\omega}}$$

It follows that A is a function of

- conductivity σ of the material
- permeability μ of the material
- *frequency* ω
- thickness t of the material

A is not directly related to the near field/ far field conditions of the system.

* Re-reflection loss (B) = $20 \log \| 1 - \rho^2 \exp(-2t/\delta) \|$

This factor B is a function of :

- conductivity σ
- permeability μ
- *frequency* ω
- *thickness t*

An example of the Re-reflection component B is given in fig. 2.6.

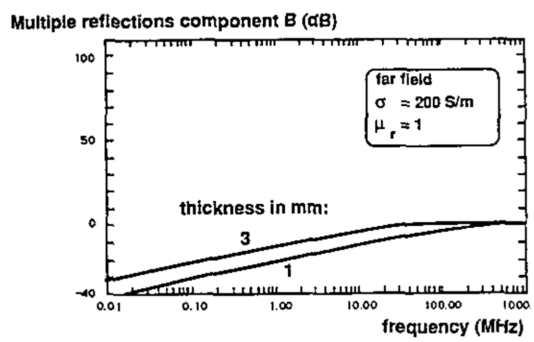


Fig. 2.6.: Re-reflection B as function of frequency

However, the three factors don't have a similar relationship to the mentionned parameters or variables. It means that different sets of values (R, A, B) may occur, giving the same amount of SE = R + A + B. This combination for SE is given in fig. 2.7 for a typical example, for far field conditions. It is seen that for these conditions SE is nearly flot up to higher frequencies, where the absorption component A becomes the dominant one.

The Re-reflection effect disappears also automatically when the thickness t of the conductive layer comes in the same order as the skin depth δ, because of the absorption (or attenuation) effect in the material. Approximate formulas are given in the next table.

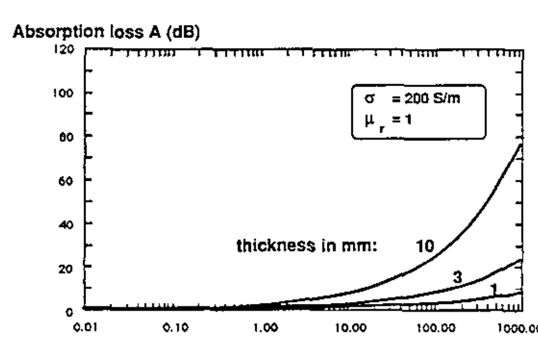


Fig. 2.5. : "A"

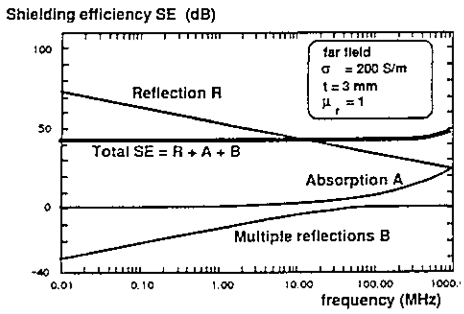


Fig. 2.7. : SE

Field condition	Shielding (dB) $t > \delta$ (A + R)
Far field ($r > \lambda/2\pi$)	$8.69\,\frac{t}{\delta} + 20\log\left[\frac{1}{4}\left(\frac{\sigma}{\omega\varepsilon_0\mu_r}\right)^{\frac{1}{2}}\right]$
Near electric dipole field ($r < \lambda/2\pi$)	$8.69\,\frac{t}{\delta} + 20\log\left[\frac{1}{4}\,\frac{c}{\omega r}\,\left(\frac{\sigma}{\omega\varepsilon_0\mu_r}\right)^{\frac{1}{2}}\right]$
Near magnetic dipole field ($r < \lambda/2\pi$)	$8.69\,\frac{t}{\delta} + 20\log\left[\frac{1}{4}\,\frac{\omega r}{c}\,\left(\frac{\sigma}{\omega\varepsilon_0\mu_r}\right)^{\frac{1}{2}}\right]$
Far field ($r > \lambda/2\pi$)	$20\log\left[1 + \frac{Z_0}{2R_s}\right]$
Near electric dipole field ($r < \lambda/2\pi$)	$10\log\left[1 + \left(\frac{c}{\omega r}\,\frac{Z_0}{2R_s}\right)^2\right]$
Near magnetic dipole field ($r < \lambda/2\pi$)	$10\log\left[1 + \left(\frac{\omega r}{c}\,\frac{Z_0}{2R_s}\right)^2\right]$

Because the square resistance $R_S = 1/\sigma\delta = \sqrt{\mu\omega/2\sigma}$, most of the expressions are given in literature as function of R_S. R_S is a well known material parameter in practice. The only problem is to measure R_S for filled conductive plastics, where the conductive material is not available at the surface.

2.2. Measurements.

Specifying shielding materials in modern electronic system design is a rather complicated decision. The influence of a lot of effects has to be taken into account : holes, openings, joints, etc. Therefore, the choice of a shielding material will depend on a series of requirements, also for mechanical design. The shielding effectiveness (or SE) of a material may be specified using different methods. These methods are used for the evaluation of materials in both near and far field. All of these methods have advantages and disadvantages, ranging from careful sample preparation up to very time consuming measurements in order to obtain exact SE-values, directly measured SE-values or calculation work.

Far Field (or plane wave)
All testing methods for far field conditions are based on the assumption that the ratio of E-field/H-field (impedance) is constant. Therefore, all test-cells are based on a coaxial transmission line, where the ratio of voltage and current (characteristic impedance) is constant.

2.2.1. ASTM-ES7-83

This obsolete cell (see figure 2.2.1.) is an expanded coaxial line, with a continuous inner conductor.

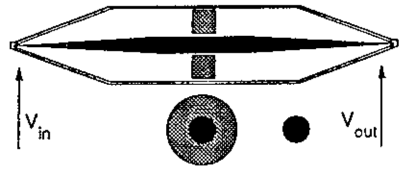


Fig. 2.2.1. : ASTM-ES7-83

It follows that samples must be prepared carefully, because good contact must be made with both the outer and inner conductor. Otherwise, the effect of these contact impedances is measured and may be the dominant factor. This may significantly interfere or hamper the determination of the SE-value of conductive plastics.

$$IL = \frac{V_{out(empty\ cell)}}{V_{out(loaded\ cell)}}$$

The empty-cell reference needs no special preparations.

2.2.2. ASTM D 4935

This new standard (see figure 2.2.2.) is based on a test-cell with an interrupted inner conductor, and a flanged outer conductor. For surface conductive materials (metal sheets, foils, painted plastics, etc.), the method seems similar to the ASTM-ES-7-method. For conductive plastics, the method is based on a capacitive coupling.

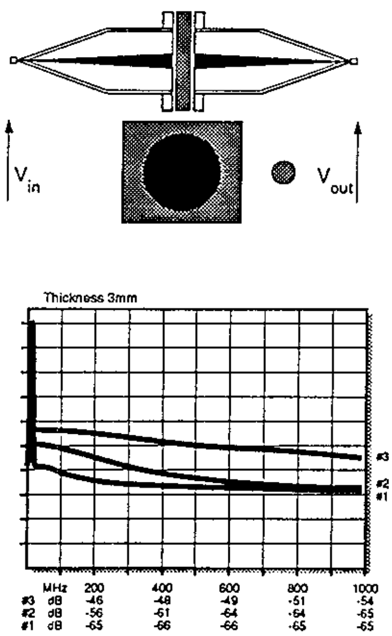
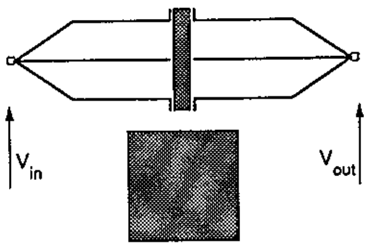


Fig. 2.2.2. : ASTM D 4935

Samples must completely cover the outer langes and no contact is allowed between the halves of the cell. (Plastic screws must be used).
The reference measurement for the empty cell is done with a reference sample in the cell, covering only the flanges and the inner conductor. This method is very complicated because for every measurement or material, the appropriated reference sample (of the same material) is needed. However, the accuracy is very high, although problems may occur for some surface conditions of the samples.

2.2.3. TEM-t cell

A very simple and easy new test-method has been developed. The test-cell (see figure 2.2.3.) based on a TEM-cell with an interrupted inner conductor and a rectangular cross-section.



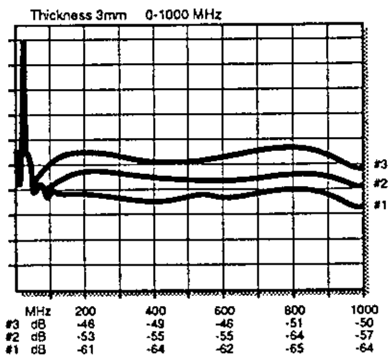


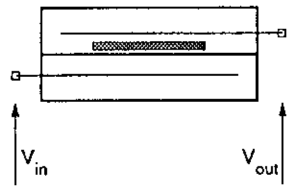
Fig. 2.2.3. : TEM-t cell

All measurements are done in a non-contacting, capacitive coupled manner for a variety of materials. Samples must completely cover the outer flanges, but smaller samples may be measured using special sample holders. Even for the reference measurement, no sample preparation is needed.

Near field
In the near-field, measurements may be done in E-field or H-field.For both field conditions, test methods have been developed. Referring to SE-theory, H-field measurements are more important.

2.2.4. E-field-ASTM-ES7-83

The E-field method is based in coupled wave guides (or transmission lines); it is called a double box-method (see figure 2.2.4).



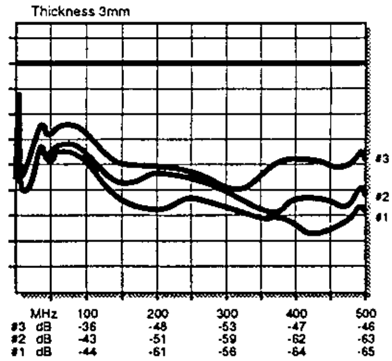


Fig. 2.2.4. : E-field-ASTM-ES7-83

This is a very simple method to use, keeping two caveats in mind :
First, the dynamic range is limited by the sealing between both parts of the cell.
And second, the field conditions are different from the far field cells, because the E-field is perpendicular to the sample. Therefore, it is not possible to match near field SE-values to farfield measurement results.

2.2.5. Takeda-Riken E/H

Using the Takeda-Riken method, two small wire antennas (for E-field) and small loop antennas (for H-field) are used at a distance of 1 cm to the sample (see figure 2.2.5).

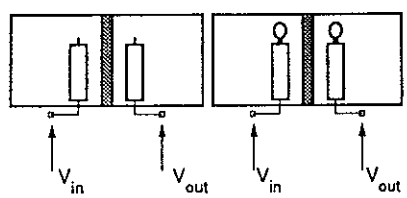


Fig. 2.2.5. : Takeda-Riken E/H

The system is an open system, so ther may be a coupling all over the system. Good sealing between both halves of the cell is necessary and very difficult. The sample must be grounded well, though the antennas are unsymmetrical (and ground-referenced).
This gives unbridgeable problems for conductive plastics. The effect is that measurements with conductive plastics are not repeatable.

2.2.6. Magnetic H-t cell

This system (see figure 2.2.6) uses two electrically shielded loop antennas, at a distance of 3 mm to the sample. The system is a closed system, and the loop-antennas are coplanar, so the measurements are done under the same conditions as the MIL-STD 285 specifications.

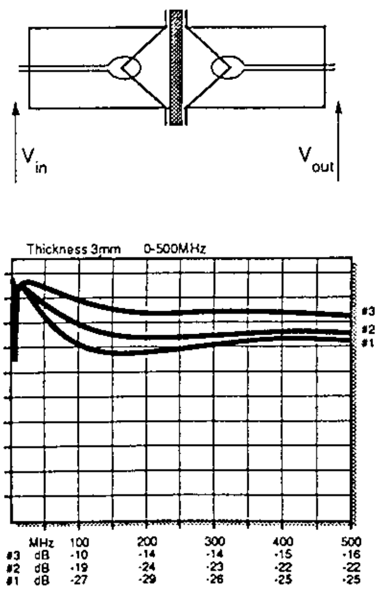


Fig. 2.2.6. Magnetic H-t cell

The system is made sample-compatible with the far-field TEM-t cell, and no sample preparation is needed.

2.2.7. Double TEM-cell

As an alternative method for the ASTM-ES7, a double TEM-cell method (see figure 2.2.7) may be used.

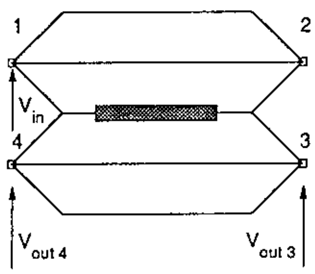


Fig. 2.2.7 : Double TEM-cell.

The advantage of this method is that it is a conditioned measurement - 50 Ω - system. By using port 1 as input and measuring at both ports 3 and 4 the E and H components of the near field, the SE values may be obtained.
Problems are that good contact of the sample with the cell is needed, and the E-field is perpendicular to the sample (thus different from far field measurement conditions).

For conductive plastics, problems may arise in trying to achieve good contact with the cell, further research is being conducted in this area. Another method may also be used : MIL-STD 285.

2.2.8. MIL-STD 285/MIL-G-83528A

The MIL-STD 285/MIL-G-83528A method is a modification of the MIL-STD 285 for testing shielded enclosures.
The method (see figure 2.2.8) uses a shielded box with an open window (± 1 x 1 m). SE is defined as the ratio of field strengths for the open window and the shielded window. The measuring distance d is 12" (± 30 cm). For lower frequencies, coplanar loop antennas are used.

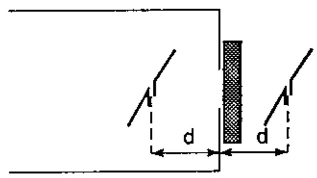


Fig. 2.2.8. : MIL-STD 285/MIL-G-83528A

The method needs rather large panels and measures a combination of a windowed box with the material, rather than the shielding material alone.

Conclusion
As a conclusion, it may be stated that all discussed test methods result in a good discrimination of the different samples and a correlation can be realized between the different methods.

For research purposes on the effects of holes and joining techniques, both TEM-t and H-t cells offer a very interesting opportunity for the far-field and near-field (H-field) characterization.

2.3. Boxes and housings

For the real world of boxes and housings, a lot of other effect are coming in, influencing the overall value of the SE of a shielded box.
The first problem deals with the need to open and close a box for servicing the electronic circuits. This means that a box is made of at least two parts, and a joint is existing between both parts.

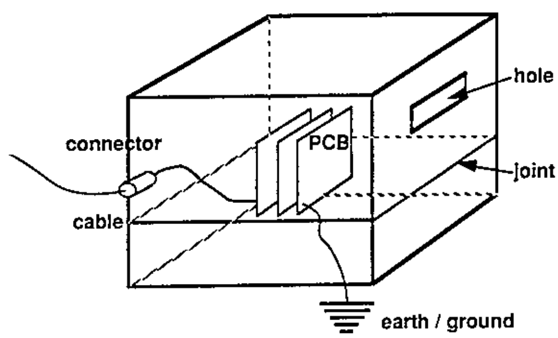


Fig. 2.3.1. : Real shielded boxes.

A second problem is related to incoming or outgoing cables, and connectors of the power supply, data & signal cables, remote control, ...

And a third one is related to openings and holes.
In the context of this paper, only a short discussion is made on all three problems.
Concerning the joints, the important point is to create a good conductive contact between both parts of the box. It should be mentionned that this should be regarded as a frequency-dependant characteristic.
This means that a capacitive coupling between both parts gives good SE results at higher frequencies.

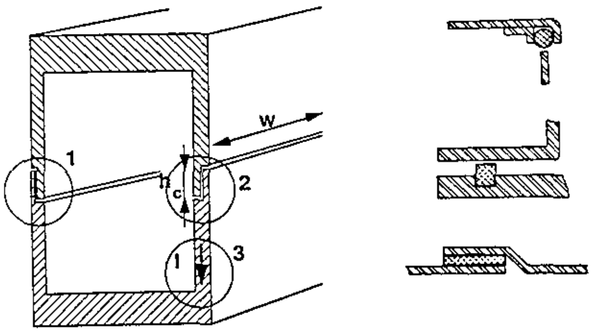


Fig. 2.3.2. : Joint techniques in box design (ref. Don White/ICT)

A hole is creating a leaky slot in a shielding wall. In some cases, one is placing a real radiating source in holes, such as displays. Therefor, holes should be *as small as possible (compared with the* wavelength) or should be shielded by other means. For displays is another requirement very important, namely of optical transparency. To give an idea about the effect of small holes or grids and meshes, the leakage of some holes are given in fig. 2.3.3.

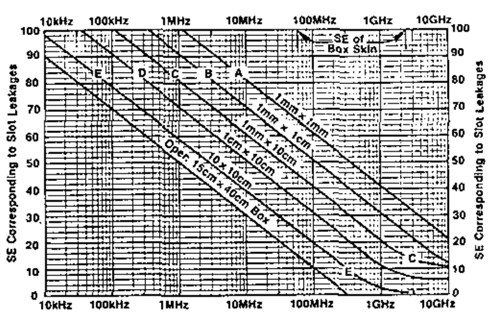


Fig. 2.3.3. : Leakage of slots (ref. Don White/ICT)

The third problem of cables and connections may be referred to both others, maintaining a good contact between all shielding materials of a construction a system, avoiding small openings.

Techniques used to maintain good electrical contact are the use of gaskets. They exist in a wide range of variaty, for all kind of applications. Also for the use of shielded windows and holes, it is very important to have a good conductive contact between all parts. This means also that - for long duration performance the effects of galvanic corrosion must be taken into account.

3. Banding.

As discussed above, making good low impedance connections to the ground reference, or between (shielding) parts of a system is determining the final EMC results of an electronic system.
Banding problems may be divided in two different aspects : the contacting problem itself and the banding wire or material (ex. gasket).
For grounding of systems, it is very important to have a low impedance connection. This may be obtained by :

- ° realising a low impedance contact between system, wires and ground
- ° using low impedance connection wires.

The first problem is related to have cleaned contact points or surfaces with a direct metal to metal contact. Corrosion is the problem on long term characteristics. This will be discussed furtheron.

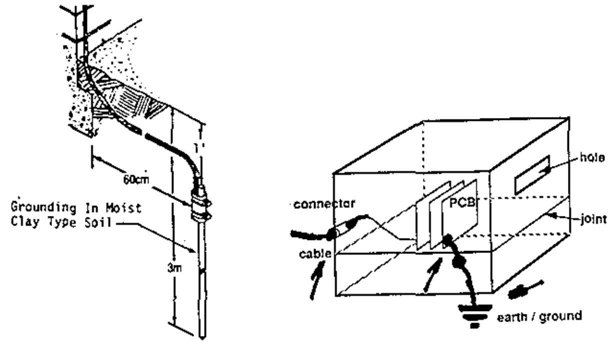


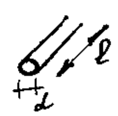
Fig. 3.1. : Wires for banding.

The impedance of wires is not only given by the resistance R, but also by an inductor $\mathcal{L}$. This influences the characteristics at higher frequencies:

$$R = \frac{\text{length } l}{\sigma . \text{ cross section } A}$$

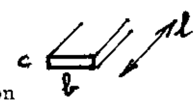
$$\mathcal{L} = 0.002\ l\ [\ \ln \frac{4l}{d} - 0,75]\ \mu H\ ;$$

round wire



$$\mathcal{L} = 0.002\ l\ [\ \ln \frac{2l}{b+c} + 0,5]$$

rectangular cross-section



It follows that low impedance may be achieved by using a multiwire, small diameter combination or a very large plate.
An example are the braided tresses used for grounding/earthing in large power systems.

Also for the application of gaskets, two factors are important :

- ° realising a low impedance contact between the housing and the gasket.
- ° the characteristics of the gasket itself.

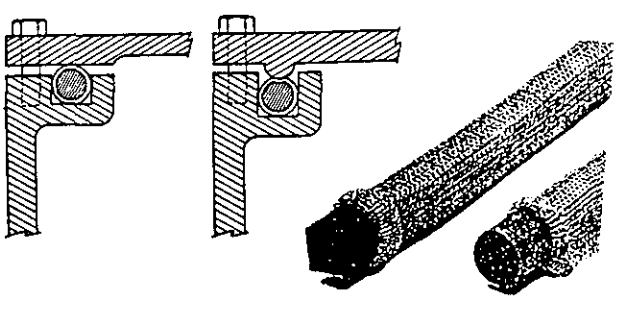


Fig.3.2.: Applying a gasket in practice

For the second item, a wide variety of materials and gaskettypes are available, depending on the application. For the first one, a good (metallic) conductive contact must be achieved. Parameters as compressibility and flexibility of the gasket are influencing the final result, but also the galvanic corrosion effect between different conductive materials.
And the same is true for the mating surface between both parts of a shielding box.
If two dissimilar metal surfaces are placed in a moisty ambient, galvanic effect is generated by a movement of electrons. As a result, the surfaces are eroded or corroded. The same effect occurs when two identical metallic surfaces are carrying a current in a moisty ambient. This is the case for shielding materials, due to induced currents in the shield.
Two techniques are used to avoid galvanic corrosion :

- avoiding the penetration of moisture by an appropiate sealing. If this is the case, there are no restriations in the choice of the metallic (or conductive) parts.
- choosing metals from adjacent galvanic groups. These groups are given in the next table. If metals from non-adjacent groups have to be used, an intermediate metal layer must be introduced (generating a higher contact impedance !). These conditions are very severe for the use of banding contacts in the presence of salt water.

GROUP					
I	II	III	IV	V	VI
Gold, Platinum Gold/Platinum alloys		Chromium plate, Molybdenum, Tungsten			
Graphite, Rhodium, Palladium, Titanium, Silver & Silver alloys, Silver-filled coatings		Copper & Copper Alloys		(Aluminum &) Aluminum Alloys, Cadmium, Zinc, Beryllium	
		Brass & Bronze			
	Titanium, Nickel, Cobalt, Nickel & Cobalt alloys, Nickel/Copper alloys (inc. Monel)		Tin, Lead, Tin/Lead solders Indium, Aluminum (2000 & 7000 Series)		Tin, Magnesium
	A286 Steel				
	AISI 300 Series Steel				
		431, 440, AM355, pH hardened Steels			
			AISI 410, 416, 420, Alloy & Carbon Steels		
Silver-filled Elastomers				Galvanized Steel	

Table : Galvanic groups of metals.

In this paper, a short overview of the influence of grounding, shielding, bonding and their related properties have been discussed. In the context of this course, only a tutorial discussion of the major problems was made.

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A REVIEW OF THE NATO EMC ANALYSIS PROGRAMME AND RELATED EUROPEAN COMMUNITY DEVELOPMENTS

by

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Summary

In support of the International Military Staff and the Allied Radio Frequency Agency, there has been a NATO EMC Analysis Programme since 1973. Early work involved compatibility of fixed frequency systems and the development of frequency assignment models. More recently the work has involved problems associated with the compatibility of frequency hopping systems and their management. Increasing competition for limited spectrum space by both military systems themselves and also by civil systems coupled with the need for more dynamic frequency management in the battlefield poses many compatibility interests for the future. Military and civil EMC interests are also becoming more inter-related because of European Community decisions concerning telecommunications and electronic devices.

1. Background to the NATO EMC Analysis Programme

1.1 The NATO EMC Analysis (NEMCA) Programme is now well established although of fairly recent origin. The Military Committee of NATO agreed the need for a NEMCA Programme in 1971 (ref 1) and the first posts were filled in 1973. The need for an EMC analysis activity arose out of the functions and responsibilities of the Allied Radio Frequency Agency (ARFA). This is one of five Telecommunications agencies located at NATO HQ within the CIS Division of the International Military Staff (figure 1). Among the tasks of ARFA is the coordination of Frequency Supportability Applications for new military systems and equipments to be deployed in NATO Europe and the day-to-day management of the band 225-400 MHz which is a military frequency band in all the European NATO countries. The frequency supportability application process consists of establishing whether the new system or equipment is suitable for operation in the frequency band or bands for which application is made. Agreement by the nations of ARFA implies that, subsequently, efforts will be made to find allotments or assignments of frequencies to allow the newcomer to operate.

1.2 ARFA, which had been established in 1951, found in the late 1960s and early 1970s, that increasing densities of communications equipments, the emergence of new types of coding and modulation schemes and the greater emphasis on system mobility, made it essential to have available EMC analysis in support of its work; in particular on frequency supportability applications and also on frequency assignments.

1.3 National EMC programmes already existed at that time but did not meet NATO's needs. The national programmes were, and still are, principally involved with platform, cosite or intra-system EMC. A new organisation was needed to tackle problems having multinational or NATO-wide aspects. The aims of the NATO EMC programme have hardly changed from its inception but the scope of the programme is much wider. Today it increasingly contributes to the work of other NATO agencies, committees and groups such as the Allied Tactical Communications Agency (ATCA), Sub-Group 5 on the NATO Identification System (NIS), the Allied Naval Communications Agency (ANCA) and the NATO Communications and Information Systems Agency (NACISA).

1.4 To fulfil the original objective of the programme, the Military Committee invested this responsibility with ARFA. In setting up the programme, NATO was faced with the choice of whether to carry out most or all of the work in-house with a substantial team of wide-ranging skills or whether to establish a small team which would contract out the most detailed, calculation-intensive work to existing specialised EMC groups in industry or national defence organisations. The latter course was chosen and the EMC Section of ARFA is a team of only four people with a modest budget which enables them to commission studies requiring larger or different resources.

2. Scope of the NATO EMC Analysis Programme

2.1 In giving terms of reference for the new activity, the Military Committee in MC 177, recognised three different categories of EMC assessment:

a. Equipment EMC aimed at ensuring that all components and modules within an electronic equipment will function properly without causing interference to other elements of the same equipment. This is a national responsibility.

b. Intra-System EMC addresses the electromagnetic compatibility within large installations or platforms (ships, aircraft) with the objective to reduce interference between co-sited systems to an acceptable level so that the performance of these systems will not be degraded. This EMC responsibility rests either with the national authorities or, if the system is procured by NATO, with the NATO agency responsible for procurement. In the latter case some of the work may be performed under the NEMCA Programme. The NATO Airborne Early Warning System is an example.

c. Environmental EMC takes into account the actual deployment of a system under operational conditions including interoperability, Electronic Warfare and flexibility requirements and has to ensure that both the system under consideration and other existing and planned systems sharing the same frequency subband(s) in the same geographical area will be able to perform their functions compatibly. The procuring nation or NATO agency is required to determine the conditions under which a new system can be compatibly deployed. Also included in environmental EMC considerations are aspects such as the equipment mix in a band, frequency assignment or allotment strategies and other questions of this type, which are not associated with procurement of a specific system. In these cases a general policy will be applied, whereby the administration (or agency, or command) with the authority to decide on these questions is inherently responsible for dealing with the EMC analysis implications of such decisions. It is work within this category which forms the bulk of the NATO EMC Analysis Programme.

2.2 Although the NATO EMC Programme is concerned mainly with environmental EMC and communication systems studies, NATO involves itself in certain cases with equipment EMC. For example, the NATO Military Agency for Standardisation (MAS) has published the following series of NATO Standardisation Agreements (STANAGs) on the EMC of aircraft systems and equipments:

STANAG 3456AE	Aircraft Electrical System Characteristics
STANAG 3457AE	Ground Electrical Power Supplies for Aircraft
STANAG 3516AE	EMC for Aircraft Electrical and Electronic Equipment
STANAG 3614	EMC of Aircraft Systems.

Work is in progress in SWG/10 of the CNAD (Conference of National Armaments Directors) to develop a STANAG on inter-ship EMC.

3. Environmental EMC

3.1 Within the NEMCA Programme the environmental EMC assessment sometimes involves establishing by analysis or bench test, the effect of one equipment on another (one-on-one) but this is usually the first step in assessing the system on system (many-on-many) impact. It is rare for a particular system to have exclusive use of a band of frequencies or even of a number of channels (allotment) and given the dynamic character of tactical military systems, absolute compatibility is not possible. Many of the projects within the NEMCA Programme have involved system-to-system analyses aimed at finding whether operating conditions can be established under which a tolerable level of interference is not exceeded and which do not unduly restrict the military mission. With different systems operating in the same frequency band and having to share limited frequency resources, the required compatibility can be achieved by equipment and system design and also by system management. Within the scope of system management, use or access to the radio frequency spectrum contributes in the form of

frequency separation
distance separation
time separation.

In theory at least, one could add polarisation decoupling but in the frequency bands where the need is greatest, below 1 GHz, propagation characteristics and the deployment constraints of the tactical system result in a benefit that is not sufficiently consistent to make this a practical option.

Although spreading codes and frequency hopping sequences are used in order to avoid detection and jamming, their characteristics can also be managed so as to achieve EMC within systems. Indeed the assignment of hop-set sequences (or net numbers) is analogous to the assignment of single frequencies.

3.2 Assessing how well the management of these parameters can achieve the required compatibility of systems in a battlefield environment, involves complicated analyses which necessarily include data on deployment scenarios, frequency assignments, hop-set characteristics, equipment parameters, interference thresholds and propagation statistics or models. Complications arise from establishing the deployment scenarios, which sometimes involves both red and blue forces, and from interpreting the analysis results in operational terms. However, as will be described, many system-to-system EMC analyses performed under the NEMCA Programme have gained acceptance and resulted in, for example, the basis for management of the 225-400 MHz band, the selection of hop-sets for the EW resistant communications systems HAVE QUICK and SATURN, decisions on the use of low power (unlicensed) devices in military frequency bands and have influenced strongly the strategies for frequency management of the VHF (30-88 MHz) band which must accommodate Combat Net Radio and Single Channel Radio Access systems.

3.3 The problems of environmental EMC arise particularly with operations in the military frequency bands below about 5 GHz (figure 2) and most of all the so-called VHF and UHF bands, 30-88 MHz and 225-400 MHz. This is a consequence of these bands being the most suitable for tactical battlefield systems by virtue of propagation and antenna characteristics. Much of the effort of the NEMCA Programme has been and continues to be focussed on these two frequency bands.

3.4 For many years the main military use of the 30-88 MHz band has been for Combat Net Radio (CNR). These are highly mobile systems used in a tactical area in the form of vehicle-mounted and man-pack units. The band has always been over-subscribed in terms of the number of different nets fielded and hence the number of fixed frequency assignments requested. Compatible operation has been achieved to some extent by judicious frequency management at both the allotment and assignment levels. New problems of compatibility have arisen recently by the introduction of digital systems and the fielding of automatic channel selection (ACS) and frequency hopping systems. In addition, the band now has to accommodate Single Channel Radio Access (SCRA) systems from some nations having the same range of parameters. All CNR and SCRA systems are very dynamic in terms of their location and their temporal demand for communications access. Both CNR and SCRA can operate in the same area of the battlefield and may at any time be within co-channel and adjacent channel interference distance of similar (but unfortunately not identical) systems of Allied nations.

3.5 The frequency band 225-400 MHz, with minor sub-band exceptions, has been agreed to be available for exclusive military use in NATO Europe both in peactime and in wartime. This fact alone would be enough to ensure that it is heavily used in terms of systems and assignments. Additionally, of course, this range of frequencies is attractive in terms of both single channel and multi-channel systems and it is the highest available frequency band for tactical systems where links of reasonable length can be established without necessarily having radio line-of-sight. This band is used therefore to support a wide range of narrow and wide band systems which are, for the most part, destined to be operated in the tactical region of the battlefield (including sea and air space) and which must be capable either of rapid deployment or mobile operation (figure 3). All the principal systems operating in this band listed below have been involved in the various EMC analysis projects within the NEMCA Programme.

Air/Ground/Air (A/G/A). Although some military aircraft can use the VHF band almost all communications between aircraft and between ground stations and aircraft must be assigned in the range 225-400 MHz. Some 8500 assignments are presently made. Ground stations may include high-power amplifiers to overcome jamming to the aircraft and may have more than 30 assignments. The single channel A/G/A communications include amplitude modulated clear voice transmissions as well as encrypted voice and data. Fixed frequency operation can take place in the same area and timeframe as slow and fast frequency hopping.

Radio Relay. Trunked radio relay systems provide the backbone communications for the automatic switched tactical area systems. They are digital systems requiring between 0.5 and 1.5 MHz of bandwidth for each point to point connection.

Tactical Satellite Systems. UHF satellite systems operate in sub-bands of the 225-400 MHz band and use fixed and mobile ground stations including aircraft and ships. Channel widths vary from 5 kHz to 500 kHz.

Navy Communications. These are for the most part single channel systems having equipment characteristics similar to A/G/A systems. Like the air force systems, they are now equipped to operate in both fixed frequency and frequency hopping modes.

These systems and others must all share the EM environment compatibly through having to share the same spectrum and sometimes the same channels.

4. Example Projects within the NEMCA Programme

4.1 In this section a few of the projects and activities, both past and present, within the NEMCA Programme will be reviewed. They have been selected to illustrate the range of system EMC problems encountered and how they have been tackled.

4.2 Introduction of frequency hopping systems (NEMCA Project 4).

A project was established in the early 1980s to investigate the effect of introducing frequency hopping systems into the battlefield zone. The early part of the project was aimed at investigating interference mechanisms and establishing the threshold levels with other systems in both the VHF and UHF bands. Much of this work was performed through study contracts let by ARFA. These studies analysed the interference at the equipment-to-equipment level for various types of victim receiver and for a range of hop-rates for the frequency hopping system. In this first part of the study the interference to single channel A/G/A systems, conventional CNR, radio relay systems and domestic television was investigated. These one-on-one results were used to assess the impact of hoppers on the UHF and VHF bands.

The next phase of the VHF band study extended the work to the system level (ref 2). For this it was necessary to describe a range of operational scenarios involving a single Corps area, two and three adjacent Allied Corps and finally that of two Allied Corps and an enemy Corps in which the VHF band was also used for tactical communications. The study produced very valuable though arguably not surprising results. It confirmed the overcrowding of the VHF band that users had long complained about and the likelihood of interference even between conventional fixed frequency systems. The value of the confirmation was that it focussed attention on the problems of frequency management of the VHF band and all subsequent discussions on its use, including the introduction of new equipment designs have been influenced by NEMCA Project 4. It must be borne in mind that by the mid-1980s, large scale exercises had become less and less frequent and in any case it is difficult to obtain field data for several Corps under normal exercise conditions.

The study also showed that for an over-suscribed frequency band, operations in a combat environment could be improved by CNR or SCRA systems using Automatic Channel Selection up to a certain level of channel utilisation. The automatic channel selection exploits any temporal, spatial and spectrum opportunities until prevented by local or widespread saturation. It represents a dynamic refinement of the frequency assignment process.

Concerning the frequency hopping systems, the study showed that, measured in terms of system-wide interference, frequency hopping systems would cause interference to more receivers in an over-subscribed situation where the hoppers could not be given exclusive channels. However the nature of the interference produced by frequency hoppers is different and it can be said that they are more democratic in their manner of interference. In a fixed frequency environment, where co-channel, adjacent channel or intermodulation interference occurs, it will affect a limited number of victims, but it may affect them severely and it will be persistent. Frequency hopping interferers will interfere with more victims but only for the time of a dwell and therefore depending on the hop-rate and dwell time and the number of frequency hopping nets it may be possible to achieve a sort of compatibility where there are more victims detecting interference but fewer suffering to an intolerable extent.

Operational judgements may still have to be made on the protection of long links or of vitally important nets. With conventional fixed frequency systems, this was built into the frequency assignment process thereby avoiding any possibility of co-channel or adjacent channel interference in those particular cases.

They can still be protected from frequency hoppers by excluding certain frequencies from hop-sets or the frequencies interrogated by ACS systems but there is an obvious risk of giving away information to the enemy and there is in any case a limit to the number of candidate frequencies that can be denied to hopping or ACS systems without in turn impairing their operation.

4.3 A very significant outcome of NEMCA Project 4 was the drafting of a Military Committee document by ATCA aimed at improving the situation by management means. MC 297 (ref 3) outlines the need for improved means for assigning VHF radio frequencies in the tactical battlefield area. This need, identified by the Military Committee, was endorsed by the Secretary General of NATO and resulted in the establishment of Project Group 8 under the TSGCEE. PG8 has decided that the need for improved frequency management tools extends beyond CNR and the VHF band and has detailed a programme of work which will lead to the development of a Tactical Spectrum Management System (TSMS).

4.4 Current VHF Band Study (NEMCA Project 7)

Current NEMCA activities in the VHF band are covered under NEMCA Project 7 which is a follow-on to Project 4. It is a study requested by the Allied Tactical Communications Agency and is an attempt to assess the effect of the deployment of Allied Electronic Warfare (EW) assets, e.g. jammers, on VHF band operation. The study takes as its baseline, the scenarios and results of NEMCA Project 4 and adds an EW scenario involving land-based and airborne jammers. The results may well help to indicate in what ways we need to extend and improve frequency management in the battlefield to achieve an acceptable level of compatible operation of Allied communications and EW.

4.5 Introduction of Frequency Hoppers in the UHF Band

Using the results of the first phase of NEMCA Project 4, an evaluation on the impact of introducing the slow frequency hopping system HAVE QUICK into the 225-400 MHz band was performed. The initial operation for HAVE QUICK was for Air/Ground/Air and Air/Air Communications particularly in roles of direct support to air defence. The one-on-one results obtained for potential UHF band victim receivers showed that, given the order of the hop-set size provided by the 225-400 MHz band, and the re-visit and dwell times of the system, HAVE QUICK could share to an acceptable degree of compatibility, the frequencies assigned to other fixed frequency Air/Ground/Air systems. Radio relay and Tactical Satellite systems were judged to be not suitable for sharing. The results of this evaluation were then used to establish the size of the hop-set and select the candidate hop-set frequencies for HAVE QUICK.

4.6 Hop-set Management for Electromagnetic Compatibility

Given that a hop-set must be of finite size, it must be used in such a way that links or nets can be operated independently of one another without interference within the hopping system. The obvious way to do this, in those cases where it is possible, is by establishing an orthogonal mode of operation. This means that all the hop-set sequences switch frequency at the same instant and dwell for the same period in such a manner that at any moment, a particular frequency is only being used by one of the hop-set sequences (or nets as they have become known). If for example, there are 100 frequencies in a hop-set, there will be an upper limit of 100 orthogonal nets that can be obtained whether or not the nets hop through the set in the same sequence. The hop-set sequences, or nets, must be assigned in a calculated manner in much the same way as fixed frequencies to avoid interference. Even if only orthogonal hop-set sequences are allowed, interference can occur on adjacent channels (particularly at co-sites) and it could occur on co-channels if the number of nets to be assigned is greater than the available number of orthogonal sequences.

Work was performed under the NEMCA Programme in 1988-1989 to establish the rules to be applied to the assignment of net numbers (a number which effectively identifies the sequency) for the HAVE QUICK system in NATO Europe. These rules are now incorporated in the SHAPE Standard Operating Procedure (SOP) (ref 4) and the same algorithm will probably be the basis for the net assignment rules which will be established for the SATURN system which will gradually succeed HAVE QUICK over the next few years.

Adjacent channel compatibility within frequency hopping systems can be achieved either by dividing the total number of frequencies available into a number of smaller hop-sets which can then be used independently or by using all frequencies available but arranging hop-set sequences such that, throughout the whole length of the pseudo-random sequence, there will always be a minimum instantaneous frequency separation between certain hop-set sequences. Within a completly orthogonal system there will be no possibility of co-channel interference if no net number is assigned twice and that need not occur until the number of nets required exceed the maximum available for that hop-set size. Even then, compatible operation can be achieved by re-assigning net numbers when the interference distance is exceeded. Assigning net numbers, or hop-set sequences, is therefore analogous to assignment of fixed frequencies.

4.7 NAVY EMC/EMI

The EMC Section of ARFA has, for more than four years now, provided the technical support to Special Working Group 10 which reports to the NATO Naval Armaments Group (NNAG). The work involves the problem of intership EMC/EMI arising out of the formation of multi-nation task forces. The close formation of different ships having different equipments produced a special problem for which national management programmes were not equipped. ARFA assisted in the development of an interference prediction model called NATCAP (NATO Electromagnetic Compatibility Analysis Programme). This was based on an earlier US model called EMCAP. The first version of NATCAP dealt only with radar-radar interference. It uses data on the antenna radiation patters, the radar transmitter spectrum and the victim receiver response to provide a prediction of the probability of interference. Observations were made by several national navies during the exercises "NORTHERN WEDDING 86" and "DISPLAY DETERMINATION 87" to validate predictions made by the model. After NORTHERN WEDDING 86, the model's propagation module was modified under a contract awarded by ARFA. This now allows for the optional use of a near-water path loss calculation which was based on data obtained through ARFA participation in AGARD Working Group 02 dealing with Near Water Propagation (ref 5).

Several NATO nations are now using the NATCAP model for their own purposes. The most recent ARFA contribution to SWG10 has been to obtain national predictions and results of trials, and to make an analysis of the entire data in order to provide validation or to propose further model improvements. Another important development in this work is the NATO Staff Target drafted by SWG10 for an "EMI Prediction Model and Data Base". This would build on the experience of NATCAP to produce a tool to predict interference involving communications, including Tactical Satellite systems, as well as radar. It is intended that the model will be developed in a form which will support detailed frequency assignment planning and be available in a suitable form for use at sea.

4.8 Post-2000 Communications

As well as dealing with EMC problems of current or emerging systems, the NEMCA Programme includes work on future systems that are still very much in the conceptual stage. Staff participate in the meetings of Project Group 6 of the Tri-Service Group on Communications and Electronic Equipment (TSGCEE). PG6 is studying the requirements for tactical communications in the land combat zone, post-2000. ARFA has contributed to their work by preparing a qualitative forecast of the EM environment in which the post-2000 systems will have to operate (ref 6). Factors which ARFA deemed will be important influences on the environment for tactical communications included:

- Spectrum availability. The currently available military spectrum is described in the NATO Joint Frequency Agreement. Tactical bands are available at HF, VHF or UHF in peace and war and can be exclusive or shared with civil users. In the post-2000 period the amount available for military use may be decreased. The range 500-3000 MHz particularly is expected to come under pressure from civil interests at the World Administrative Radio Conference in 1992. The new climate of opinion in Western and Eastern Europe may make future defences of military spectrum more difficult. In summary, for peacetime and wartime operations, the spectrum available will not be greater, it may well be less.
- Many present systems will still be operational beyond 2000. The current lifetime of about 20 years for new systems may be extended due to the good reliability of modern equipments.
- The post-2000 systems will introduce new hop-rates and other ECM resistance features. These may conflict with good EMC, which in principle is most easily achieved with 'like systems', for which better management rules can be developed.
- The new generation of ECM resistant communication systems will generate their own EMC problems for which solutions have

yet to be found. For example as the density of deployment of frequency hopping equipments at co-sites increases, intermodulation will become a problem. The solution is not as straightforward as for fixed frequency systems. Another example is that just as finite propagation time means that a follower jammer cannot jam a system hopping at very fast rates, it also means that the frequency hopping system becomes non-orthogonal and co-channel interference can occur.

The ARFA contribution to PG6 concluded that for the post-2000 systems to achieve a high degree of compatibility they must be designed to:

a. operate in a shared spectrum environment, and

b. avoid co-site interference, including image, harmonic and intermodulation.

Their development must go hand-in-hand with the development of better operational and frequency spectrum management techniques.

4.9 Other current and forthcoming tasks within the NATO EMC analysis programme involve the EMC of Tactical Satellite Systems, the NATO Identification System (NIS) and associated national EMC studies, and the Air Command and Control System (ACCS) which will be the successor to the NATO Air Defence Ground Environment (NADGE).

5. The European Community EMC Directive and its Possible Impact on Military Communications Equipments

5.1 The first part of this lecture dealt with a rather specific area of activity. The previous lectures concerned wide-ranging topics. In this part of the lecture, EMC developments in the European Community (EC) will be reviewed and it will be seen that these developments will have a bearing on almost all the EMC issues dealt with in this lecture series. They even have an impact beyond the European Community area.

5.2 The EC published a Council Directive on EMC in the Official Journal of the European Communities on 23 May 1989 (ref 7). It had been approved by the Council on 3 May 1989. The directive is a document which is binding on governments, and member states of the EC are required to adopt legislation to implement the directive by 1 July 1991. This is done through national legislation. The necessary legislation to apply the provisions of the directive must, according to the present timetable, be applicable by 1 January 1992.

5.3 The basic reason for the EC action is to support the internal market; in other words to permit the free movement of goods. The EC directive on EMC is just one of a large number of measures consequent on the Single European Act, (which is a modification of the Treaty of Rome) taken to ensure that the movement of goods across EC internal boundaries cannot be thwarted by barriers, and this includes technical specifications. It is clear from the EC directive that the legislation and the standards and test methods will apply to a wide range of electrical and electronic appliances. There is the notable exemption given to the radio equipment used by radio amateurs but even then only if the equipment is not available commercially. The direction includes the following list of products as an illustration of what will be covered:

a. domestic radio and television receivers

b. industrial manufacturing equipment

c. mobile radio equipment

d. mobile radio and commercial radio-telephone equipment

e. medical and scientific apparatus

f. information technology equipment

g. domestic appliances and household electronic equipment

h. aeronautical and marine radio apparatus

i. educational electronic equipment

j. telecommunications networks and apparatus

k. radio and television broadcast transmitters

l. lights and fluorescent lamps.

The directive also states that these equipments and appliances, as well as being compatible by virtue of their emissions, must be constructed in such a way as to have an adequate level of electromagnetic immunity.

5.4 As well as the national legislative preparations, work has already started on the development of specifications and test standards. There exists an organisation called CENELEC (European Committee for Electrotechnical Standardisation) which acts on behalf of the European Commission to deal with the technical issues. It is a non-profit making organisation set up under Belgian Law and is composed of a Secretariat and the National Electrotechnical Committees of 18 countries in Western Europe. The 18 countries are the EC countries plus the EFTA (European Free Trade Area) countries. In the context of the composition of CENELEC it is important to remember that not only are not all the EC countries in NATO but not all NATO countries are in either the EC or EFTA (figure 4). CENELEC was given the legal right to set reference European standards on behalf of the EC as long ago as 1983 (EC Directive 83/189). In the matter of EMC, CENELEC tried to do this first by harmonisation of existing national and international standards. This could not be made to work and in order to implement the EC directive, work is now in progress to develop common enforceable standards. As soon as CENELEC started this work, a "standstill" on national EMC standards was imposed. This is standard CENELEC procedure and it means that no new national standards can be developed and no existing ones modified.

The work is being done by two technical committees composed of national representatives:

TC110 on EMC
SC110A on EMC Products.

TC110 is charged to set up comprehensive generic standards covering the aspects of electromagnetic emission and immunity in line with the directive, taking into account existing national and international standards. It will also describe the characteristics of typical EM environment locations for apparatus, which can be correlated with test methods. It will specify recommended emission limits and immunity levels. The prepared standards will contain definitions, requirements test methods, test instrumentation and conformance criteria.

SC110A is charged to consider the harmonisation of product related documents and to participate in the preparation of the dedicated product standards as requested by TC110.

5.5 From the list of apparatus and appliances and the scope of the CENELEC committees it is clear that, even allowing for some lack of definition, most electrical and electronic apparatus will be affected by the directive as well as much of the existing EMC standards and EMC tasking activities. To be covered by the legislation, it is simply necessary that the product is to be made commercially available in the EC or EFTA countries whether made in those countries or outside. CENELEC's work is well underway. The first two generic standards and the first five product oriented standards had, by August 1990, been distributed for comment. The CENELEC Secretariat is confident that most of the generic standards will have been published by the end of 1992.

5.6 The important issue for NATO and also for national defence authorities, is whether the directive and hence the CENELEC standards will appply to military equipment. The issue is complicated and has caused much discussion since the publication of the directive. There is in fact an exemption for military equipment contained within the Treaty of Rome itself in Article 223. This makes reference to munitions of war and thus clearly exempts weapons. It is not clear whether it was intended to exempt military equipment at the time of the original Treaty. A case can be argued for exemption where apparatus is not offered for sale on the open market. Certainly there is still a range of equipment and apparatus, of a non-weapons nature, that is procured by the military and is not commercially available. However, in the category of information technology and to a lesser extent telecommunications, the military use equipments that are the same or similar to those available commercially for civil users. Furthermore, some equipments may be developed originally with the military as the intended user but then became more widely available (e.g. night vision devices).

No exemption can be obtained on the basis of the frequency bands of operation. CENELEC will apply their standards to all frequency bands as they are described in the ITU Regulations, whether they are exclusively civil, shared military and civil, or exclusively military.

For most military electrical and electronic equipment, meeting the CENELEC standards would not be technically onerous. Most military procurement specifications include EMC standards. National ones such as MILSTAM 461 (USA) and DEF-STAN 59-41 (UK) are probably more rigorous in most respects than those that will emerge from CENELEC. In any case the CENELEC standards will be based on national inputs to the CENELEC committees. A problem may arise for organisations such as Defence Ministries, in that, because of their own specialised EMC requirements, it may be necessary for their equipments to be compliant with both their own and the EC standards. These considerations indicate the desirability of having national defence and NATO inputs to the technical committees. Even if military equipments themselves are not subsequently covered by the legislation, operation in radio frequency bands shared with civil users could be affected. Particularly in the area of immunity (susceptibility) from emissions, there may be a need to influence the standards since the design approach for the military and civil equipments is very different because of their intended working environments. Even for operation in exclusive military frequency bands NATO and national defence inputs to the discussions may be needed in order to influence the limits on out-of-band emissions that could be radiated by devices using the adjacent band of frequencies.

5.7 Recent experience in the US has shown the need for vigilance concerning the in-band and out-of-band emissions of so-called "Non-licensed Devices". Many of these correspond to apparatus referred to in Europe as "Low Power Devices". The American and European descriptions cover a wide range of devices such as garage and car door openers, radio controlled toys, baby alarms and cordless telephones. In the US, standards for the in-band characteristics have been established but not for out-of-band emissions. Even the in-band emission levels can be high and do not correspond to everyone's definition of "Low Power". Under the EC directive, these appliances would need to comply with the CENELEC standards but as in the US, individual radio licenses would not be needed.

5.8 Many devices are already available in Europe and it must be expected that their use could become as widespread as in the US, where 13 million garage door openers have been sold. In recent years several of the European national authorities have received requests for approval of Low Power Devices and these included some for operation in the military band 225-400 MHz. This band is not for military use world-wide and for example some garage door openers operate around 310 MHz. In order to assess the risk of interference from LP Devices, a study was performed under the NEMCA Programme at the request to the Joint Civil/Military ARFA meeting. The study was specifically aimed at assessing whether interference to aircraft communications systems could be created by widespread use of LP devices. For the analysis ARFA used the power levels for devices permitted by the Federal Communications Commission in the US. These permitted levels are as follows (and shown in figure 5).

Continuous Emissions

Frequency Range	Maximum Emission Level
216-960 MHz	200 uV/m at 3 metres

Intermittent Emissions

Frequency Range	Maximum Emission Level
174-260 MHz	3750 uV/m at 3 metres
260-470 MHz	3750-12500 uV/m at 3 metres (linearly interpolated)

Relative to a 10 watt aircraft transmitter output, these intermittent emissions are between 63.8 and 55.6 dB down at 225 MHz and 400 MHz respectively. ARFA analysed the situation of low flying aircraft in communication with the ground or other aircraft. Calculations were made to determine the upper limit tolerable for the density of simultaneously active emitters. The analysis assumed an aircraft height of 1000 feet (about 300 metres), wanted signal margins of 0, 10 and 20 dB, and a range of distances between the aircraft receiver and its ground station or other aircraft. The results are shown in figure 5.

Low flying aircraft will tend to be close to their ground station or other aircraft when communicating and of course at the high end of the 225-400 MHz band particularly, the short horizon distance will hinder communication. However the results show that for a within-horizon wanted distance of about 50 km, even if zero dB signal margin is accepted, fewer than 100 emitting devices per sq.km can be tolerated. Judging by usage in the US, this sort of density is quite feasible near large conurbations. Arguably low flying should not occur there but it cannot be ruled out especially in the case of an emergency. In their report to the ARFA Joint Civil/Military meeting, the ARFA EMC Section concluded that low power devices, which would by their nature be used in a non-controlled manner, presented a sufficient hazard to flight safety and recommended that they should not be permitted in the military band, 225-400 MHz. This recommendation was accepted by the Civil/Military meeting.

5.9 The implementation of the EMC directive is not the sole development of Community-wide proportions that will affect NATO and national military planning for equipments and systems using the radio frequency spectrum.

There has been an explosive growth in the demand for mobile communications including car-radios and various paging devices. The growth is expected to continue and the EC has developed a standard for a Pan-European radio mobile cellular system (GSM) which will operate in the 900 MHz band. It is developing a standard for a paging system (HERMES) operating around 170 MHz. Projected numbers are for 12 million GSM subscribers in Europe by the end of the decade and for paging devices to increase from a 1987 total of 1.3 million to 13 million by the end of the 1990s. Based on past experience these projections may well be on the low side.

Several nations have been giving consideration to introducing new ways of managing the radio frequency and there are some suggestions of delegating management of certain bands to commercial organisations ("Deregulation"). At the same time the European Community has published a proposal for the Commission to carry out studies on the use of the spectrum with a view to developing EC views on changes to allocations and spectrum uses (ref 8). It is anticipated that these studies will include EMC analysis.

6.0 Concluding Remarks

The success of the NATO EMC Analysis Programme has been marked by the wide acceptance of its outputs and the direct and necessary support it gives to the frequency management responsibilities of ARFA. It has, as foreseen, filled a gap left by national EMC studies, and thereby contributes also to the work of other NATO Telecommunication Agencies such as ATCA and ANCA, as well as the Tri-Service Group and NATO project offices. There will be a continuing need for the programme as long as there are plans for multi-national operations and multi-national procurements. There will be changes in agreements on the use of the radio frequency spectrum in Europe resulting from equipment and system developments, demands for communications capacity both civil and military, spectrum deregulation, political changes and the increasing influence of the Commission of the European Community in communications and spectrum management. The NATO EMC Analysis Programme under ARFA is well placed to respond to these changes.

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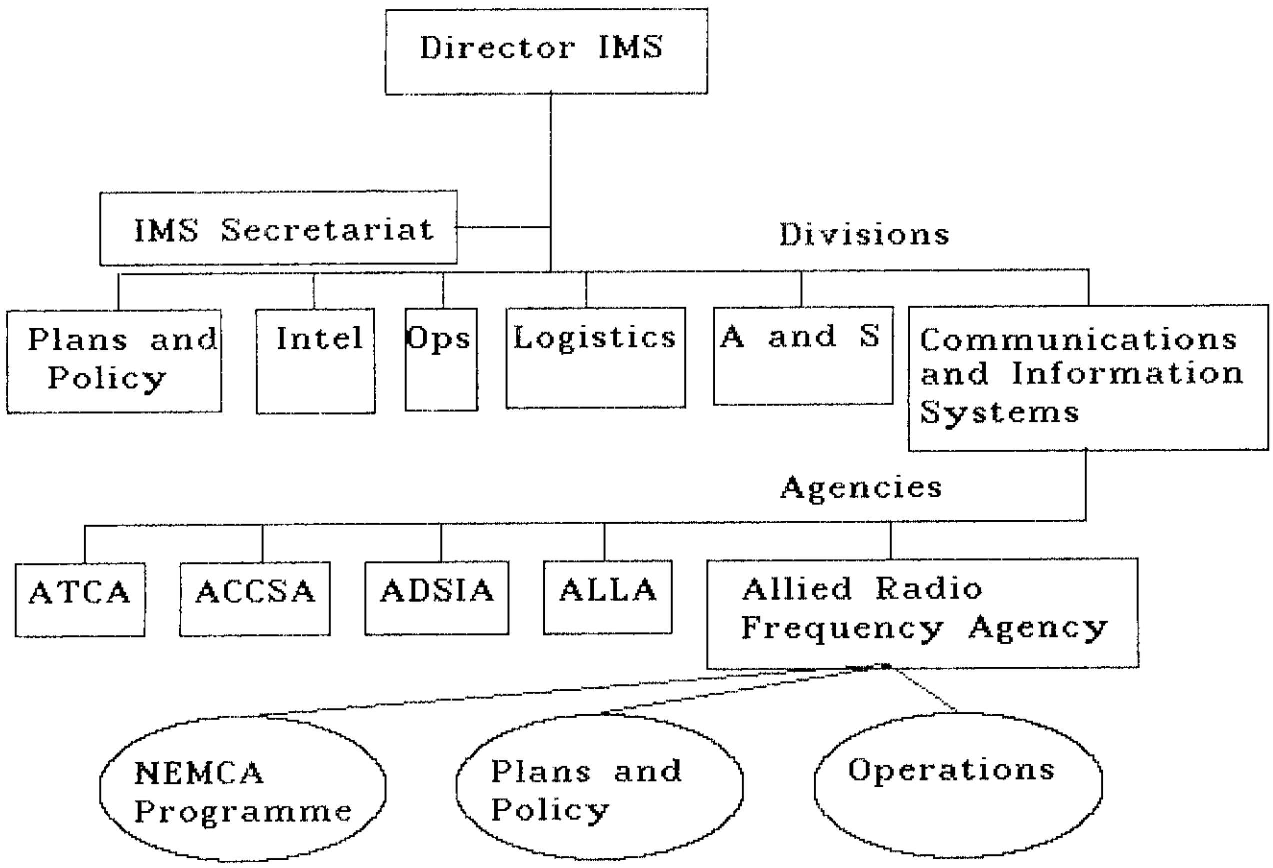
Note:

The NATO references are not contained in open publications. They may be obtained through national representatives to ARFA or ATCA subject to normal security regulations where appropriate.



<u>Figure 1 The International Military Staff</u>

<u>BAND</u>	<u>SYSTEM</u>	<u>DEPLOYMENT</u>
VHF	COMBAT NET RADIO	MOBILE, FIXED
	RADIO RELAY	TRANSPORTABLE
	SINGLE CHANNEL RADIO ACCESS	
UHF	A/G/A SYSTEMS	MOBILE (FAST)
	RADIO RELAY	FIXED
	JTIDS/MIDS	MOBILE (FAST)
SHF	RADIO RELAY	FIXED

<u>Figure 2 Some Military Systems 30MHz-5GHz</u>

SYSTEMS SHARING THE 225-400 MHZ BAND

AIR-GROUND-AIR

AIRCRAFT PWR 20W
GROUND PWR 100W-3kW
BANDWIDTHS 6kHz-50kHz
ANTENNAS: OMNIDIRECTIONAL
LOCATIONS: GROUND- MOSTLY FIXED
AIR - RANGE > 300NM

RADIO RELAY

POWER 20-40W
BANDWIDTHS 500kHz-1.5MHz
ANTENNAS 8-10dB GAIN
LOCATIONS: SOME FIXED
MOSTLY VARIABLE WITHIN CORPS AREA

NAVY

POWER 20-100W
BANDWIDTHS 6kHz-50kHz
ANTENNAS: OMNIDIRECTIONAL
LOCATIONS: SHIPS AT SEA AND IN HARBOUR
AIRCRAFT AT SEA AND OVERLAND

Figure 3

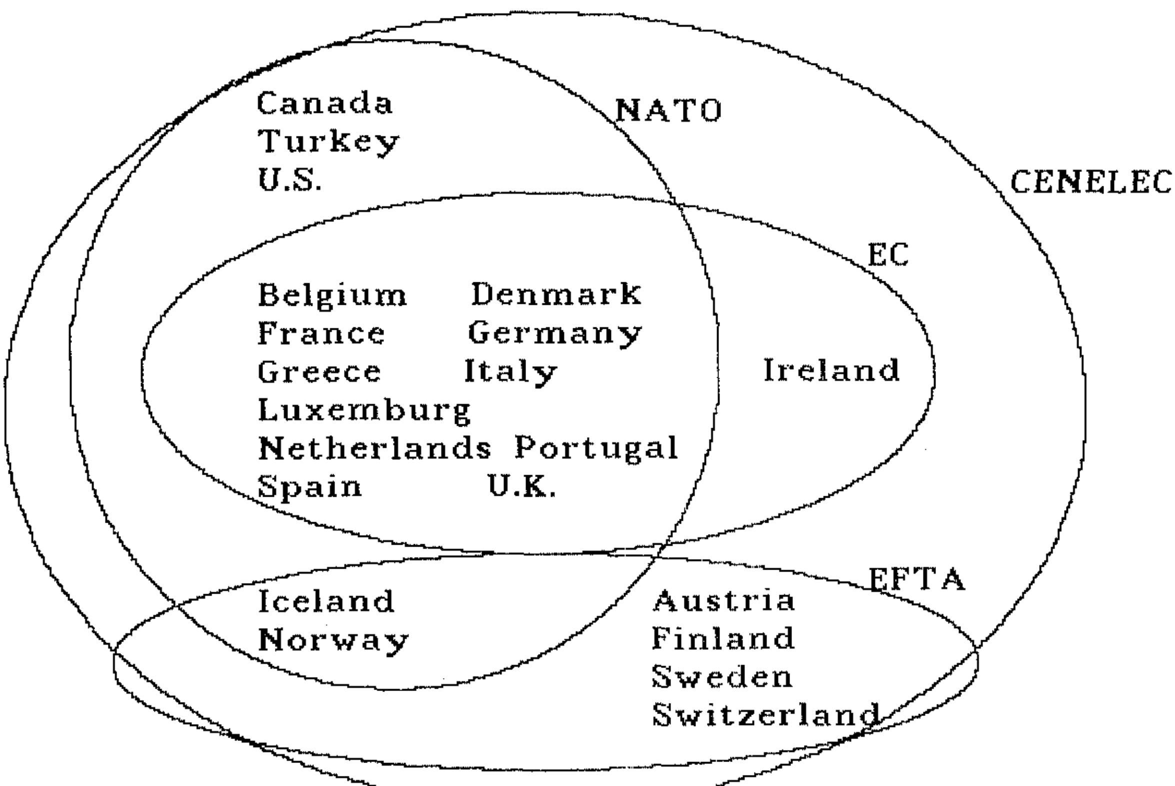


Figure 4 Participation in NATO, EC, EFTA and CENELEC

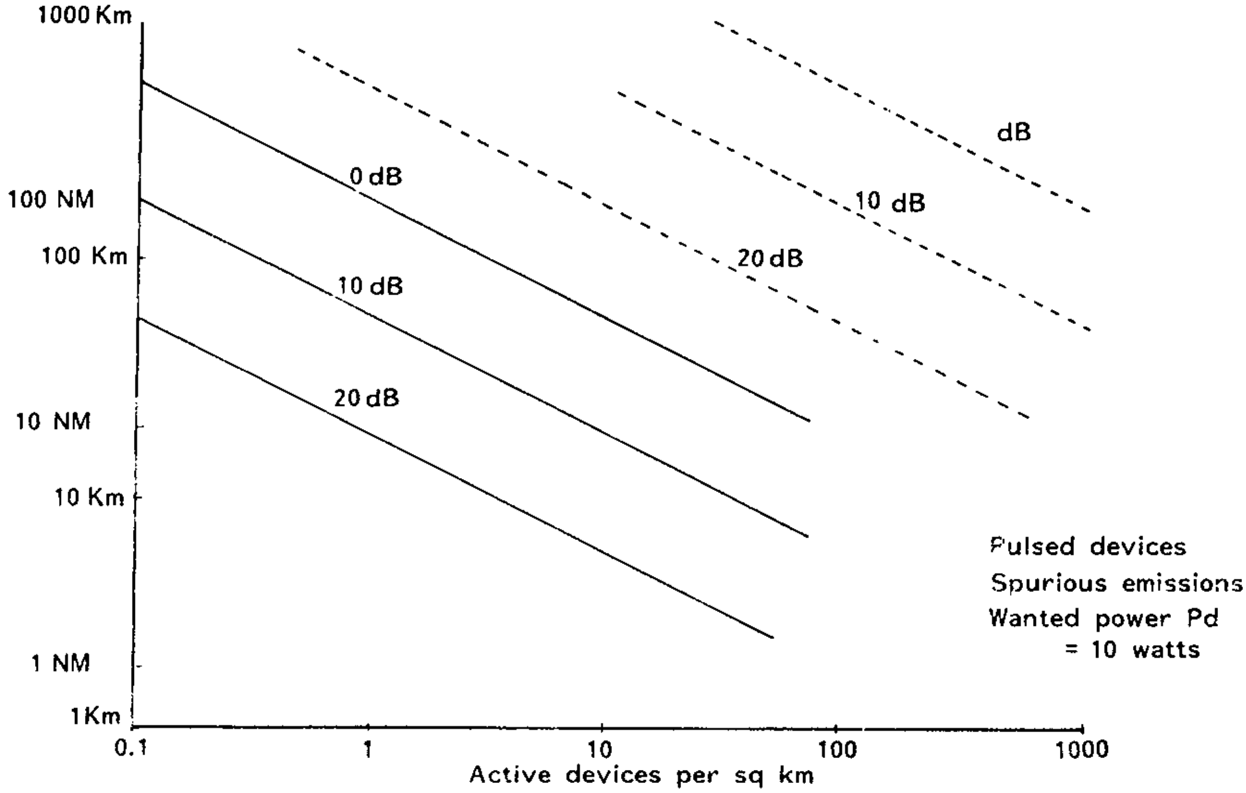


Figure 5 Range vs density of interferers for constant D/U ratio

SPECTRUM MANAGEMENT AND CONSERVATION

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SUMMARY

The lecture first presents a review of the basic concepts of radio spectrum utilisation and the physical mechanisms of electromagnetic (EM) wave propagation of relevance to radio, radar, control and navigation systems. A survey of the uses of the various bands in the radio frequency (RF) spectrum is then presented. Interactions between the propagation mechanisms introduced previously and frequency planning requirements are considered, making reference to the precision of available modelling and prediction procedures; emphasis is given to interference generation potential. Attention is then turned to equipment and system design features which affect their spectral occupancy characteristics, including RF equipment imperfections, signal design, system design and control, and co-siting problems. Finally, the manner in which the bands of the radio spectrum are allocated and controlled is outlined, and areas of possible future concern in both NATO and civil operations identified.

1. INTRODUCTION

1.1 Scope of the Lecture

This lecture is concerned with the use of the radio spectrum by equipments and systems employing "unguided" propagation, ie energy transfer between distinct geographical locations by the physical mechanisms of radiation and near-field (reactive) coupling; it does not consider energy transfer via "guided" propagation, ie over wires, coaxial cables and optical fibres, except where any energy leakage cannot be assumed to be negligible. Some consideration is given to additional spurious radiation from bona fide transmitters, and from electronic systems whose primary function is not to radiate energy, eg computing equipments.

1.2 The Radio Frequency Spectrum

The radio frequency (RF) spectrum is an extremely valuable natural resource and, as with most natural resources, there is the possibility that it may be used inefficiently or in an unauthorised manner (Weisz 1989). Consequently, the disciplines of spectrum management and conservation are intended to ensure that the RF spectrum is used in the most efficient and effective way possible, given the physical characteristics of the available propagation mechanisms, equipment parameters and imperfections, and user requirements. It should be noted that the value of the spectrum as a resource only becomes apparent when it is actually used; then, the users have had to invest in equipment. At that point, they naturally have considerable resistance to any modifications in frequency assignment and control procedures; therefore, any such changes are normally introduced over a considerable period of time.

1.3 NATO Requirements

The NATO operational requirements encompass radio, radar, control and navigation systems of many different types, fixed and mobile, dispersed over a wide geographical area, and radiating in a large proportion of the available frequency bands. Access requirements range from continuous to very infrequent, yet all users require a specified grade of service whenever they choose to access the spectrum. There must also be sufficient flexibility in the spectrum control procedures to allow new systems to be introduced without disrupting existing operations. Similarly, flexibility must also be available to cope with increased levels of activity in times of emergency, tension or war.

Of late, there has been an unprecedented expansion in the number of civil services exploiting the RF spectrum, particularly in the fields of mobile radio and paging; as a result, there is increasing pressure on military users to release some of their exclusive frequency bands in order that these can be made available for civil purposes. This, in turn, creates a situation in which military systems must make more efficient use of reduced spectral allocations.

1.4 The Electromagnetic Compatibility Problem

The main problems in spectrum management and control arise from the need to maintain electromagnetic compatibility (EMC) between a vast number and variety of systems in situations where the degree of natural isolation between those systems is extremely variable. Radio propagation mechanisms can usually only be characterised in a statistical manner; hence, interaction and interference effects between systems are also of a probabilistic nature. Some propagation mechanisms can be effectively confined within national or operational boundaries; others, by their very nature, inevitably cross those boundaries. Thus, spectrum management, control and co-ordination must be applied at both national and international levels.

1.5 Facts of Radio Systems Life

The potential complexity and variety of user requirements for radio services are illustrated by Fig. 1. There are certain general "facts of radio systems life" which impose fundamental limitations upon equipments and systems designed to meet these requirements; these will now be identified.

(i) Efficiency of Radiation

Fig. 2 shows the relationship between wavelength and frequency for EM waves propagating in free space. For most efficient radiation of EM energy, the transmitting antenna should have a physical size which is a significant fraction of the wavelength being used; this allows the antenna to become resonant, or approach resonance, at that frequency. If this condition is not met, radiation efficiency will be low. This has particular implications for systems operating in the lower part of the RF spectrum, as will be discussed later.

(ii) Mobile & Static Terminals

Radiated power levels and antenna efficiencies will tend to be significantly greater for static installations than for mobile terminals. Again, this is especially true of the lower frequency part of the spectrum. The reasons for this are limitations on the physical size of antennas and prime power generation associated with mobiles; there may be additional limitations imposed on mobiles arising from EMC constraints due to the co-location of a number of distinct EM systems operating simultaneously, eg communications transmitters and receivers, navigation and radar, all within a confined physical space.

(iii) Bandwidth Availability

The amount of spectrum available for any given radio system tends to increase as the frequency of operation increases. The fundamental equation governing the amount of information, I, that can be passed in time T over a memoryless communication channel having bandwidth B and received signal-to-noise ratio (SNR) S/N is

$$I = B\,T\,\log_2[1 + S/N] \text{ bits} \qquad (1)$$

where S and N are respectively the total wanted signal and unwanted noise powers within the bandwidth B. From (1), it is clear that, with fixed T and SNR, increasing the bandwidth implies that more information can be passed over the channel. Below are listed typical bandwidths available to a user in different RF ranges:

Frequency Range	Bandwidth
3 - 30 kHz	50 Hz
0.3 - 3 MHz	500 Hz
3 - 30 MHz	3 kHz
0.3 - 3 GHz	25 kHz
3 - 30 GHz	5 MHz
Optical	1 GHz

(iv) Dominant Noise

The dominant type and level of noise experienced by a radio system will depend upon its frequency of operation. In Fig. 3, the relative EM radio noise powers due to natural atmospheric sources, man-made sources and internal sources within a typical receiver are plotted as a function of frequency (CCIR 1978a) (CCIR 1963). It is seen that in the lower part of the spectrum, below a few MHz, atmospheric noise is most significant; in the range up to a few 100s of MHz, man-made noise of various types normally is at the highest level; above this range, internal receiver noise dominates. Thus, depending upon its operating frequency range, a radio system may be "external noise limited" (at the lower frequencies) or "internal noise limited"; in the VHF/UHF range, between say 50 MHz and 1 GHz, internal and external noise levels may well be comparable. Fig. 3 does not attempt to consider co-channel or adjacent channel interference from other radio users of the spectrum. It should also be noted that all noise sources can only be specified statistically and therefore the atmospheric and man-made ranges shown may well vary widely with location, time of day, season, etc.

(v) Range Achievable

The geographical range achievable with a radio system depends upon the frequency of operation in an irregular manner. Fig. 4 illustrates this irregularity in a simplified manner. Although propagation via a satellite transponder is not a natural mechanism, it is included here for completeness. Thus, there is no simple progression in which, say, range increases with frequency of operation. In EMC terms, the implication of this is that the degree of natural isolation between radio services will be highly frequency dependent.

1.6 Format of the Lecture Material

Section 2 of this lecture deals with the propagation mechanisms available for radio services, and with their basic characteristics; the status of analysis and prediction techniques for modelling propagation and interference generation is considered in Section 3. Section 4 examines the various aspects of equipment and system design influencing the EMC of radio services. The means by which the regulation and control of the RF spectrum is effected are described in Section 5, whilst Section 6 discusses areas of possible future concern to NATO and other spectrum users, together with overall conclusions.

2. PROPAGATION AND INTERFERENCE MECHANISMS

To fulfill the user requirements mentioned previously, there are a number of distinct, naturally-occurring, physical mechanisms which allow the propagation of radio waves over a wide range of distances at different frequencies in the RF spectrum. These will now be reviewed briefly (Griffiths 1987) (ITT 1972), and their major properties indicated.

2.1 Mechanism I: Guided Modes

Such modes occur at the lower RF frequencies, typically below a few 10s of kHz. Energy propagates via spherical waveguide modes in the natural cavity formed by the earth's surface on one side, and the lower edge of the ionosphere on the other. The width of this cavity is normally between 70 and 100 km. As with all waveguide modes, propagation losses are low, giving the possibility of RF energy transfer over very long distances; amplitude and phase stability are good - another characteristic of waveguide propagation. A further important property of radio waves in this frequency range is that they have a significant seawater penetration as a result of their long wavelength and correspondingly large skin depth (Watt 1967).

The properties outlined above make this form of propagation useful for terrestrial navigation systems requiring accurate phase measurements, eg OMEGA (Swanson 1983) operating over ranges of several 1000s of miles at frequencies of approximately 10 - 11 kHz. Propagation disruption can occur when the ionosphere becomes disturbed, say due to abnormal solar activity; in this case, phase stability will degrade, as will navigational accuracy. Seawater penetration also makes this band useful for low data rate broadcast communication from surface transmitters to submerged submarines, although communication in the reverse sense presents severe practical problems. The wavelengths in this region of the spectrum can be as much as several 1000s of kilometres; consequently, any practical transmitting antenna can only be a very small fraction of a wavelength in size and therefore highly inefficient. Essentially, such antennas act as probes in the earth-ionosphere cavity.

2.2 Mechanism II: Groundwave

In the frequency range from a few hundreds of kHz to about 25 MHz, beyond line-of-sight (BLOS) ranges can be achieved by groundwave propagation. When a radio wave is launched from an antenna near to the earth's surface, the groundwave propagation mechanism will transfer radiated energy with an efficiency that depends upon the parameters of the ground over which it passes (chiefly conductivity and permittivity), together with the frequency of operation. Maximum ranges of a few 100s of km are achieved over a low-loss, high conductivity, surface at lower frequencies; thus, lower frequency operation over seawater is most effective (Norton 1960).

2.3 Mechanism III: Ionospheric Skywave

The various layers of the ionosphere can act as a distributed refracting mechanism for radio waves in the frequency range from about 2 to 40 MHz, although the available frequency range at any time is dependent upon ionospheric state. Fig. 5 shows the major elements of the ionosphere: it is a stratified medium, with regions of electron density concentration known as "layers". From a propagation viewpoint, the E- and F-layers are most important in providing what is referred to as "high frequency (HF)", or "shortwave", communications; in addition, "sporadic" E-layer modes can extend the available frequency range substantially for limited periods of time over restricted geographical areas. Above about 2 MHz, the D-layer acts primarily as an attenuator of radio waves. Ranges out to world-wide can be obtained via skywave propagation.

As indicated in Fig. 5, the major influence upon ionospheric state is the sun; levels of solar activity directly affect the ionisation of the various layers, and hence the characteristics of radio propagation via those layers. The earth's magnetic field also influences HF propagation. The medium is extremely variable with time, season, solar activity, path orientation, geographical position, etc; typically, radio waves will be simulataneously refracted by more than one layer, giving rise to multipath effects at the receiver (Davies 1966).

2.4 Mechanism IV: Meteor-Burst

Over the approximate frequency range from 25 to 100 MHz, long range radio propagation can occur as a result of reflection from ionised meteor trails. These trails are due to meteorites and micrometeorites entering the earth's upper atmosphere at an altitude of about 100 km and "burning up", thus creating a cone of ionisation which then rapidly disperses. Whilst the trail persists, it is an efficient reflector of radio waves and provides a high quality received signal; as the trail disperses, the received signal deteriorates and may enter a fading regime. Received signal durations are typically 0.5 to 1 second, with inter-trail intervals of a few 10s of seconds. Obviously, data can only be transmitted in relatively short bursts but, during a burst, large bandwidths are available and ranges out to 2000 km can be achieved (Bartholome & Vogt 1968).

2.5 Mechanism V: Ionospheric Scatter

Within the same frequency range of about 25 to 100 MHz quoted above for meteor-burst, ionospheric scatter propagation can also occur. This is due to scattering of radio wave energy from ionospheric irregularities at an altitude of approximately 100 km. Consequently, ranges out to 2000 km can again be achieved. In contrast to meteor-burst, received signals are continuous and relatively weak - a characteristic of many scattering mechanisms (Bartholome & Vogt 1965). Other scatter modes also exist in this frequency range, eg auroral scatter (Thrane 1986).

2.6 Mechanism VI: Tropospheric Scatter

An important propagation mechanism from the frequency management viewpoint, operative over a very wide frequency range from about 30 MHz up to 11 GHz or more, is tropospheric scatter. The troposphere is a turbulent region of the atmosphere a few kilometres above the earth's surface; radio wave scattering takes place as a result of refractive index irregularities in this region. Ranges out to aproximately 400 km are achievable, again with relatively low signal strengths (CCIR 1966)(Ishimaru 1988).

2.7 Mechanism VII: Line-of-Sight

Line-of-sight, sometimes referred to as "space" wave, propagation occurs at virtually all frequencies, although practical line-of-sight (LOS) systems tend to operate in the frequency range from say 30 MHz to a few 10s of GHz. Geographical range is primarily determined by geometrical considerations, with antenna height being critical. Other factors which cause variations in received signal level from that which might be expected from simple geometrical calculations include diffraction, reflection, Fresnel zone effects, refraction and atmospheric absorption. Fig. 6 shows an example of an LOS technique, ie satellite communications, which clearly has the potential for world-wide coverage. Terrestrial systems, on the other hand, have ranges restricted to a few 10s of kilometres (Livingston 1970).

2.8 Mechanism VIII: Ducting

The phenomenon of ducting, in which EM energy is propagated over beyond LOS distances when this would not normally be expected, arises from the presence of naturally occurring waveguides near to the earth's surface. Such waveguides tend to form when weather conditions are very stable, and are experienced particularly over seawater paths. The physical dimensions of the waveguides are relatively small, with the walls corresponding to refractive index discontinuities. Clearly, propagation of radio signals in these waveguides will be most efficient when their wavelengths are comparable with the guide dimensions - typically from a few metres to a few 10s of metres. However, ducting has been noted from frequencies ranging from VHF to 11 GHz or more, over distances of a few hundred kilometres (Rotherham 1984).

Table 1 lists, in broad terms, the internationally designated bands of the RF spectrum and their abbreviated titles, the frequency and wavelength ranges for the bands, the main propagation mechanisms, and typical services operating in each band (ITT 1972).

3. ANALYSIS, MODELLING & PREDICTION TECHNIQUES FOR PROPAGATION & INTERFERENCE

In the previous section, propagation over the complete RF spectrum was described by reference to the various physical mechanisms which can occur. In some cases, these mechanisms are confined within a comparatively limited part of the spectrum; in other cases, the mechanism may straddle several of the bands shown in Table 1. Analysis, modelling and prediction techniques for propagation and interference assessment will now be reviewed, first by reference to those same individual mechanisms, and then by considering situations in which multiple mechanisms may occur simultaneously. A more detailed treatment of some of this material can be found in (AGARD 1979) and (AGARD 1986).

3.1 Guided Modes

A comprehensive discussion of propagation modelling in the ELF/VLF/LF bands is given by (Kelly 1986). As stated previously, propagation in this frequency range is due to spherical waveguide modes in the cavity between the earth's surface and the lower (D-layer) of the ionosphere. Within this cavity, it is possible for a number of different waveguide modes to exist simultaneously, their parameters being dependent upon the physical characteristics of the cavity and its boundaries, together with the manner in which the wave is launched into the cavity. Thus, at a given point on the earth's surface, the total received signal will be due to a combination of waveguide modes. Modelling essentially depends on a calculation of the complex reflection coefficients of the ionosphere, which are themselves functions of ionospheric electron and ion density profiles, and the reflection coefficients of the ground, which depend upon conductivity and dielectric constant; angles of incidence and geomagnetic field conditions must also be specified. Consequently, analysis of system performance must take account of the following parameters:

(a) the eigenangles for each mode, dependent upon the reflection coefficients of ground and ionosphere (note that the ionosphere is anisotropic because of interactions with the earth's magnetic field, causing crosspolar field components on reflection);
(b) transmitter power and antenna type;
(c) receiving antenna type;
(d) the excitation factor for each waveguide mode;
(e) earth radius;
(f) grea t circle distance between transmitter and receiver;
(g) ionospheric height;
(h) frequency of operation;
(i) free-space constants (permeability and dielectric constant);
(j) ground conductivity and dielectric constant.

One of the most sophisticated analysis programs to date is WAVEGUID, developed by the US Naval Electronics Laboratory Center and based on equations formulated in (Pappert 1970). This predicts the vertical electric field received at any point on the earth's surface due to a vertical electric transmitting antenna at another specified location. Fig. 7 is an example of the output of this type of analysis for a seawater path at 24 kHz, showing that the regions of greatest prediction inaccuracy occur around the interference nulls at 2.4 and 3.8 Mm.

Maximum errors in field strength prediction normally do not exceed a few dB, as illustrated by the data of Fig. 7, when the ionosphere and earth's surface can be considered isotropic. When the isotropic assumption cannot be maintained, the analysis procedures required become more sophisticated, with a tendency to larger errors. In these cases, when the path parameters vary spatially, mode-matching techniques coupled with the WKB approximation are used in conjunction with the basic waveguide models, eg (Pappert & Shockey 1974). An alternative to waveguide analysis is termed the "wave hop" technique (Morfitt & Halley 1970); this treats the total received signal as comprising components due to groundwave, 1-hop skywave, 2-hop skywave, etc, and has an accuracy comparable with that of waveguide analysis.

It is also possible to carry out time dispersion and multipath delay calculations using the same propagation models. Multipath is normally the more severe of the two effects and can give rise to maximum time delays of about 0.7 ms.

In any system performance prediction, an acceptable grade of service can be associated with a minimum SNR, which itself will be dependent upon the particular channel encoding scheme employed by the system. Therefore, it is also necessary to model the noise environment which, as indicated in Fig. 3, is dominated by atmospheric sources at the lower end of the RF spectrum. The noise pdf is non-Gaussian and is often taken to be log-normal; consequently, noise clipping can be effective.

As with all natural phenomena, ELF/VLF/LF propagation is the subject of statistical variation. Normal conditions can be predicted with reasonable asccuracy, but abnormal (statistically rarer) effects cannot be modelled with such precision. The physical phenomena which can cause greater divergence between predicted and actual field strengths include:

(a) increased solar activity, such as solar flares, giving rise to sudden ionospheric disturbances (SIDs) and magnetic storms;
(b) polar cap absorption (PCA) events, again due to excess solar particle emissions interacting with the earth's near-vertical magnetic field in polar regions;
(c) transition fading when the day/night terminator crosses the propagation path;
(d) formation of an ionospheric C-layer, below the D-layer, as a result of cosmic ray activity;
(e) ground conductivity and weather effects, which alter the parameters of the lower propagatioboun dary;
(f) propagation over paths exhibiting many and large changes in values of the ground constants.

3.2 Groundwave

The prediction of field strengths due to the groundwave mechanism introduced in Section 2.2 is normally reasonably precise, ie within a few dB, providing that the ground constants can be quantified with sufficient accuracy. The general form of groundwave propagation equation is

$$E = (A\ k\ P^{1/2}) / d \quad mV/m \quad (2)$$

where A is an attenuation factor, k is a further scaling factor which varies with the type of transmitting antenna employed, P is the transmitted power in kW, and d is the path length in km. In one formulation, A itself can be taken to be a function of distance, frequency and conductivity (Norton 1936 / 1937), ie

$$A = (2 + 0.3p) / (2 + p + 0.6p^2) \qquad (3)$$

with p, the "numerical distance", given by

$$p = (0.582\ d\ f^2) / s \qquad (4)$$

where f is the frequency in MHz and s is the ground conductivity in mS/m. Other elaborations to this basic theory have been developed, eg to incorporate the effects of ground permittivity, inhomogeneous paths (Millington 1949), etc.

Fig. 8 is an example of the variation of groundwave field strength with distance and frequency (CCIR 1978b). The groundwave is a comparatively stable propagation mechanism in terms of the amplitude and phase of the received signal. Time dispersion over say a 300 km path can be of the order of 200 ns.

3.3 Ionospheric Skywave

HF, or short wave, ionospheric skywave propagation, which can typically occur anywhere in the frequency range from about 2 to 40 MHz, is one of the more important BLOS mechanisms from a NATO operational viewpoint, particularly for mobile communications. It suffers from the major disadvantage of a highly variable propagation path, whose characteristics are affected by a wide range of physical and system design parameters. Consequently, modelling and prediction precision is somewhat limited. Signal properties depend upon complex interactions, primarily between solar radiation and particle emissions, the ionosphere and the earth's magnetic field.

In general terms, the normal observed properties of an HF skywave path are (Bradley 1979):

(a) at the lower frequencies, waves at all elevation angles (including vertical incidence) are refracted back to earth by the ionosphere;
(b) at the higher frequencies, waves at near vertical incidence tend to propagate through the ionosphere and are not refracted back to earth;
(c) the smaller the launch angle of the waves with respect to the earth's surface, the range achieved;
(d) the higher the frequency of the wave, the greater the height in the ionosphere at which refraction takes place;
(e) the smaller the launch angle of the waves with respect to the earth's surface, the lower the height at which refraction takes place.

From Fig. 3, it is seen that HF radio systems tend to be external noise limited by atmospheric sources. However, a more significant limitation on system performance is co-channel interference due to other bona fide spectrum users; any HF channel is assigned several times over on a global basis, and therefore, because of the world-wide nature of skywave propagation, there is a high probability of different services interacting. Another important feature of HF propagation is the occurrence of long multipath delays of several ms or more. Even at relatively low transmission rates, these multipath effects can cause significant intersymbol interference.

Much of the work on long-term HF modelling has been carried out under the auspices of CCIR, eg (CCIR 1970a) and (CCIR 1978c). Input data required by such programs include:

(a) time of day;
(b) month;
(c) index of solar activity, eg sunspot number;
(d) position of transmitter and receiver terminals;
(e) antenna types used;
(f) transmitter power;
(g) receiver site noise level;
(h) frequency, or frequencies, of operation;
(i) required SNR.

Among output parameter estimates required by a range of users are:

(a) most reliable propagation mode, eg 1-hop F2, 2-hop E, etc;
(b) elevation angle associated with the most reliable mode;
(c) propagation time of the most reliable mode;
(d) how many days per month the most reliable mode can be expected to occur;
(e) transmission loss associated with the most reliable mode;
(f) field strength of the most reliable mode;
(g) the proportion of time for which the received SNR is likely to exceed a specified threshold;
(h) strength of the multipath components;
(i) maximum usable frequency (MUF);
(j) lowest usable frequency (LUF);
(k) optimum working frequency (OWF or FOT).

A typical output format from an HF prediction program is shown as Fig. 9; here, MUF, FOT and LUF are plotted as functions of universal time, with the relevant propagation, noise and system data shown above.

Many different analysis and modelling programs exist, eg CCIR 252-2, HFMUFES4, IONCAP, APPLAB, MINIMUF, CCIR 894, etc. Some are mainframe-based and some PC-based. Predictions are normally made on a "monthly median" basis and, on this basis, they are reasonably precise: MUFs and LUFs can be predicted to better than 1 MHz under normal conditions; signal and noise level estimates to within about 10 dB are also realistic. The major problem with such programs is that on a short-term, or daily, basis, actual conditions may differ significantly from the monthly median values. It is therefore necessary to complement the long-term models with short-term data (Bradley 1979) or some form of real-time channel evaluation (RTCE) (Darnell 1983) if predictions which are reliable on an hour-to-hour basis are to be obtained.

There are still many features of HF systems which can only be modelled approximately, eg auroral effects, sporadic E-layer propagation and co-channel interference (Laycock et al 1988); therefore, on any given day, errors in say SNR estimates of several 10s of dB are not uncommon. Co-channel interference is the most significant modelling deficiency since it represents, in tangible terms, inadequate EMC. Indeed, it is unlikely that there will ever be a precise short-term model for such effects.

3.4 Meteor-Burst

By its very nature, meteor-burst propagation can only be characterised statistically since there is no way of predicting the occurrence of specific meteors. Therefore, parameters such as the distribution of burst types and lengths, the mean time between bursts, and the distribution of received signal and noise levels must be employed. Because of the coherent nature of the reflected energy, particularly in the initial part of the burst before significant diffusion has taken place, transmission losses are about 20 dB lower than for true scattering mechanisms, eg ionospheric scatter (Section 3.5).Fig. 10 shows the amplitude-time profiles for two typical meteor trails.

Because of geometrical considerations, systems with specified terminal locations can only use the small proportion of the total number of meteor trails which have appropriate trajectories. These geometrical restrictions also mean that the area over which energy from a given trail can be received, ie its "footprint", is comparatively small - frequently only a few kilometres across. This implies that the interference generation potential of meteor-burst systems is small, and that frequency sharing between multiple terminals is viable.

A statistically characterised propagation path implies that any modelling will also be of a statistical nature. Inputs required by an analysis program are very similar to those listed for an HF skywave analysis routine in Section 3.3. Noise on meteor-burst links is normally a combination of galactic and man-made; both components can be spatially non-uniform. Models applicable to meteor-burst systems are described by (Brown 1985) and (Ostergaard 1986). Again, there is a problem that the instantaneous conditions experienced by a given user on a specific link may differ substantially from the median conditions predicted by the statistical model.

3.5 Ionospheric Scatter

Ionospheric scatter propagation involves a high transmission loss, coupled with long- and short-term fading mechanisms. As a consequence, high gain antennas and diversity combining are required if transmitted powers are to be kept at reasonable levels. Because the frequency range is approximately the same as for meteor-burst propagation, the noise characteristics are similar to those described in Section 3.4. Modelling of this scattering mechanism can be relatively precise in terms of the received SNRs, since hourly median values of signal strength and fading (assumed Rayleigh) margins can be characterised with reasonable accuracy, ie within a few dB (Griffiths 1987).

In the same frequency range, it is also possible for other scattering mechanisms to occur, eg due to ionospheric irregularities aligned with the earth's magnetic field. These are extremely difficult to predict and model with any precision.

3.6 Tropospheric Scatter

Modelling of tropospheric scatter propagation can, as with ionospheric scatter, be carried out with reasonable precision. Again, median path losses and long- and short-term fading characteristics are required. The wide frequency range over which troposcatter effects can be observed, from a few 10s of MHz to more than 10 GHz, complicates the analysis in terms of noise effects; from Fig. 3, it can be seen that the dominant noise mechanism affecting system performance will vary over this range. An additional complicating factor in the analysis arises from the fact that scattering takes place only a few kilometres above the earth's surface and, therefore, the refractive index irregularities within the "scattering volume" are influenced by local weather systems (Spillard 1990). Many analysis models, eg (Yeh 1960) and (CCIR 1986a), have been employed; they have varying degrees of sophistication, but typically require the following types of input:

(a) transmitter and receiver locations;
(b) frequency of operation;
(c) antenna types;
(d) antenna heights;
(e) transmitter power;
(f) effective earth radius;
(g) surface refractivity;
(h) diversity arrangement;
(i) atmospheric gas absorption;
(j) terrain data;
(k) climatic data.

As stated previously, outputs are typically median signal levels, SNRs, and fading margins. For high rate digital transmission, estimates of time dispersion are also required.

From an EMC viewpoint, troposcatter is a very significant mechanism because of its wide frequency range and BLOS ranges. Although the existence of low-frequency tropscatter has been noted for many years, of late, more attention has been given to the characterisation of troposcatter in the high-HF and low-VHF bands (Darnell et al 1991), where strong signals due to the mechanism have been observed; Fig. 11 shows an example of the diurnal variation of signal level at 47 MHz over a 300 km path. Significant levels of troposcatter have also been measured at 11 GHz (Spillard 1990).

3.7 Line-of-Sight

Of all available propagation mechanisms, line-of-sight (LOS) is the most amenable to accurate characterisation. However, there are a number of additional effects which tend to make the process of path modelling less precise, particularly when mobile terminals are involved, ie (Brodhage and Hormuth 1977):

(a) atmospheric refraction;
(b) terrain features and their seasonal variability;
(c) Fresnel zone obstructions;

(d) reflections;
(e) diffraction;
(f) atmospheric absorption;
(g) rainfall attenuation;
(h) local noise sources and interference;
(i) anomalous propagation modes;
(j) polarisation rotation;
(k) Doppler effects.

Satellite systems are themselves LOS links; in practice, fixed links can be modelled with high precision because many of the factors listed above are not significant, or do not vary. Also, at the higher frequencies above about 2 GHz, system performance is internal noise limited (see Fig. 3); the characteristics of this predominantly thermal noise are stable. Thus, received SNRs in microwave satellite systems can often be predicted to within a dB.

For mobile terrestrial systems operating in the VHF and UHF bands, the situation is very different, and prediction errors of several 10s of dB can be experienced. Hence, the potential for frequency sharing is extremely variable, as are interference effects.

3.8 Ducting

Ducting, as indicated in Section 2.8, can give rise to strong signal propagation over significant BLOS distances. The physical basis of the formation of ducts is extremely complex, imperfectly understood, and thus cannot be modelled accurately. Prediction techniques are at best probabilistic, and indicate the conditions under which ducting is likely to occur, eg with stable meteorological conditions, over seawater paths, etc.

3.9 Multiple Mechanism Paths

As indicated in the preceding sections, analysis techniques having varying degrees of precision are available for modelling individual propagation mechanisms. Operationally, however, frequency bands used by various military and civilian services are such that there is a possibility that more that one mechanism may occur simultaneously. Thus, interactions between radio systems calculated on the basis of a specific propagation mechanism may well prove to be considerably in error if other mechanisms are present.

Taking as an example the low-VHF band, from say 30 to 100 MHz: in this region of the spectrum, there is a possibility that all the propagation mechanisms discussed above, with the exception of spherical waveguide modes, can occur (Darnell 1990). Hence, any EMC modelling, frequency planning and spectrum management procedures should ideally take this into account. Similarly, in the 900 MHz UHF region, LOS propagation, troposcatter and ducting can all be present.

4. RADIO EQUIPMENT & SYSTEM DESIGN & OPERATION FOR ENHANCED EMC

Figure 12 outlines the main elements of radio system EMC which can enhance spectrum utilisation efficiency and ease the spectrum management problem; they are divided into two categories:

(i) equipment / system concept and design;
(ii) in-service measures.

In this section, attention will be concentrated on (i) above.

4.1 Propagation and Noise Modelling

In Section 3 of this paper, the status of modelling techniques in the various frequency bands is outlined. Obviously, there is an on-going requirement to update these analytical models in the light of new experimental data and in response to new system and operational requirements. For example, an increased emphasis on digital transmission schemes brings with it a need to be more precise in the characterisation of time dispersion and multipath effects. Similarly, a situation in which an increased number of mobiles are deployed in a confined geographical area requires greater attention to be given to methods of achieving system isolation, eg via polarisation diversity.

4.2 Signal Generation and Processing

Fig. 13 shows the functional units of what can be termed a "generalised information transfer system". The actual architecture of any practical radio system can be mapped to these units.

In modern radio systems, digital transmission techniques predominate. The advantages of such techniques are:

(a) they allow signals to be regenerated exactly, giving effectively a "distance-independent" SNR, rather than the SNR continuously degrading, as is the case with analogue transmission;
(b) they allow signals to be fully encrypted;
(c) they facilitate channel sharing by multiple simultaneous users;
(d) they enable error control coding to be applied;
(e) they allow exploitation of a wide range of digital signal processing devices and algorithms.

However, digital transmission techniques also have a number of disadvantages, ie:

(f) they tend to be bandwidth-inefficient;
(g) variability of transmission rates as channel conditions change is limited because of terminal equipment characteristics;
(h) some form of synchronisation is normally required.

All the advantages (a) to (e) are also beneficial in EMC terms; similarly, the disadvantages (f) to (h) tend to be detrimental to EMC. Item (f) should be noted especially: a simple calculation serves to illustrate the problem. Consider a toll-quality analogue speech signal with a bandwidth of about 3 kHz; the theoretical (Nyquist) sampling frequency for this signal, f_s, is given by

$$f_s = 2 \times 3000 = 6kHz \qquad (5)$$

Allowing a safety margin for non-ideal sampling and band-limiting, choose

$$f_s = 8\text{ kHz} \qquad (6)$$

It is now required to digitise, or quantise, these samples to about 0.5 to 1% accuracy; therefore, each sample is specified by an 8-bit binary word. The overall data rate for the digitised speech, R, is thus

$$R = 8000 \times 8 = 64\text{ kbits/s} \qquad (7)$$

Fig. 14 shows the general form of the spectrum of this digitised signal; often, the first null is taken as being an indication of the required transmission bandwidth, ie 64 kHz. It is seen, therefore, that there is approximately a x20 expansion in bandwidth requirements in converting from analogue to digital speech; this represents a potentially severe frequency management problem if spectrum is limited - which is normally the case at the lower end of the RF spectrum.

The manner in which each of the functional elements of Fig. 13 influences the EMC characteristics of a radio system will now be discussed.

(i) Data Source

Most data sources have two components, ie "information", which it is essential to preserve during transmission, and "redundancy", which can be discarded. A speech signal is a good example of this where many of its features, such as pitch variation, loudness, high frequency energy, etc, are not essential to its understanding. Also, many digital data sources have states with very different probabilities of occurrence, and hence a predictability which constitutes deterministic redundancy. In principle, providing that the source can be characterised, say in terms of state probabilities, much of this redundancy can be systematically removed.

It should be noted that one of the most effective ways of easing operational demands upon the radio spectrum is simply to minimise the amount of data to be transmitted by pre-planning and eliminating unnecessary messages.

(ii) Source Encoder & Decoder

The functions of the source encoder are to transmit the source signal in form required by the user, eg digitally to facilitate encryption, and to remove as much of its deterministic redundancy as possible. Previously, it was demonstrated that directly digitised speech could typically have a data rate of 64 kbits/s; a source encoder, in the form of a vocoder, can provide intelligible digital speech at rates of between 0.8 - 2.4 kbits/s by making use of an appropriate model of speech production, eg formant, LPC, channel, etc (Darnell 1984). Similarly, if data compression techniques, such as run-length encoding, can be applied to a digital source, the source rate will again be reduced. Encryption can also be considered as a source encoding function.

There are two main benefits to be derived from efficient source encoding:

(i) from Fig. 14, it is evident that the signal bandwidth, and hence spectral occupancy, reduces as the clock rate reduces;
(ii) if the transmitted data rate over the channel is fixed, a reduction in source rate allows more of the channel capacity to be used for say error control - thus enabling higher levels of noise and interference to be tolerated; hence, channel selection criteria become less stringent.

In general, the more that is known of the source characteristics, the more effective can the source encoding procedures be made.

(iii) Channel Encoder & Decoder

The function of the channel encoder is to condition the output of the source encoder in such a way that it can withstand the types of noise, interference and distortion which will be experienced during transmission over the channel. Thus, channel encoding primarily encompasses the functions of modulation, error control coding, synchronisation and multi-user coding.

Of particular importance in the spectrum management context are modulation techniques which optimise energy concentration within a given channel, and minimise any "spill-over" into adjacent channels. The usual measure for digital modulation schemes is termed the "bandwidth efficiency", and is expressed as (bits/s)/Hz. Basic binary modulation methods, eg ASK, FSK and FSK, rate poorly using this criterion; more complex multi-state modulation schemes and partial-response techniques, eg QAM, MSK, GMSK, GTFM, etc, score comparatively highly (Haykin 1988). However, the more complex schemes also tend to be more susceptible to channel perturbations and need to be employed in conjunction with error control coding. On many practical radio channels, therefore, it may be better to use simple and robust modulation methods in conjunction with modest error control coding, rather than complex modulation formats and powerful codes.

The EMC of radio systems can be improved if the number of separate channels required can be minimised. Here, it is important that techniques to allow a given channel to be shared between a number of simultaneous users are exploited to the full; time-division multiple-access (TDMA), frequency-division multiple-access (FDMA) and code-division multiple-access (CDMA) are all valuable here.

As more precise information becomes available on current channel state, so the channel encoding procedures can be made more efficient. With time-varying and dispersive channels, some form of RTCE is required for sensing channel state so that channel encoding algorithms can be adapted responsively (Darnell 1983).

(iv) Diversity Combining

A further aspect of signal design and processing which can potentially enhance EMC is the use of diversity combining This technique requires that two or more versions of the required signal should be available at the receiving site. In a noisy environment, with fading signals, diversity combining is most effective if the noise, interference and fading characteristics of the different versions are independent, or uncorrelated. There are a number of possibilities for obtaining these m (>= 2) versions of the wanted signal, ie

(a) the use of separated receiving antennas (space diversity);
(b) the use of receiving antennas of different basic polarisations (polarisation diversity);
(c) transmission of the signal simultaneously on different radio frequencies (frequency diversity);
(d) transmission of the signal at different times (time diversity);
(e) the use of signals received via different propagation paths (mode diversity).

Diversity combining essentially applies a compensation principle, ie when one version of the received signal is of poor quality, then at least one of the other versions will be of better quality; hence, in comparison with the single received version case, an improved overall signal estimate can be obtained by considering all versions. In practice, this means that a lower mean SNR can be tolerated at the receiver for the same level of performance (Griffiths 1987). Fig. 15 is a plot of the reduction in received SNR, relative to a non-diversity system, which can be achieved with ideal dual (m = 2) and quadruple (m = 4) diversity combining for a range of bit error rates (BERs); in this case, non-coherent binary FSK is assumed. Allowing for imperfect implementation and a degree of correlation between versions, it is seen that at a BER of say 1 in a 1000, the simplest dual-diversity system gains at least 10 dB in SNR relative to a non-diversity system. This means that the transmitter power can be reduced by 10 times, with the obvious spectrum management advantages.

4.3 RF Equipment

No practical radio transmitter is perfect; all will exhibit some degree of non-linearity and also generate noise. Therefore, the following criteria must be reflected in equipment specifications (Schemel 1973):

(a) levels of broadband radiated noise;
(b) direct harmonic levels;
(c) intermodulation product (IP) levels;
(d) frequency / phase stability;
(e) spurious products resulting from frequency synthesis.

Limits for such emissions are specified in (CCIR 1970b). Of particular importance are odd-order IPs with frequencies of

$$f_{ip} = \pm\, p\, f_a \pm q\, f_b \pm r\, f_c \pm \ldots \qquad (8)$$

where p, q, r, etc are integers, and f_a, f_b, f_c, etc are the causal frequencies. Of these odd-order IPs, the 3rd-order products are the most significant since they are at the highest level and tend to fall back into the same frequency range as the causal signals. Although, relative to the intended transmission, (a) - (e) above will normally be at low level, they may nevertheless cause EMC problems for other systems in close proximity.

In the same way, radio receivers also suffer from non-linearity, noise and other imperfections which degrade their performance; chief of these are:

(a) IP generation;
(b) harmonic generation;
(c) synthesiser-generated spurious signals;
(d) reciprocal mixing effects due to local oscillator phase noise;
(e) internal noise;
(f) cross-modulation and blocking.

Particular problems occur with RF equipments when multiple transmitters and receivers (possibly operating in different frequency bands) are co-located, or are in close proximity (Schemel 1973). For example, cross-coupling can occur in the output stages of transmitters via antennas, feeders, common antenna working, etc, causing additional intermodulation.

Antenna directivity can ease coupling problems although, under co-sited near-field conditions, the level of coupling is difficult to analyse. More generally, antenna directivity and polarisation can be used to increase the level of isolation between systems, to allow lower transmitter powers, and to decrease frequency re-use distances - all tending to improve the EMC characteristics of radio systems.

The effect of site/environmental non-linearities must also be considered in a multiple-transmitter/multiple-receiver installation, since these can provide an additional source of harmonic and IP generation. This is particularly severe in a maritime situation where salt water corrosion has for many years been observed to give rise to the "rusty bolt effect", in which metal-to-oxide junctions act as electrical non-linearities when illuminated by EM fields. The spurious components generated by site non-linearities can block what might otherwise be usable channels for nearby receivers. In general, good installation practice and careful maintenance can minimise these effects.

4.4 Operational & Control Procedures

The manner in which a radio system is operated and controlled can have a significant influence upon its EMC characteristics. For example

(a) minimisation of transmitted data reduces spectral occupancy;
(b) patterns of transmission can influence the degree to which a given RF channel can be shared between multiple users;
(c) use of natural terrain features can provide isolation between services, and hence facilitate channel sharing;
(d) radiated power control (RPC), such that system performance is just adequate, eases the problems of spectrum congestion; it should be noted that the implementation of such a procedure requires the availability of RTCE and an adaptive system architecture;
(e) intelligent frequency selection can influence the radio propagation mechanism, and hence the geographical regions over which radio signals can be received.

5. REGULATION & CONTROL OF THE RADIO SPECTRUM

An important source of information concerning the general subject of spectrum regulation and control is (CCIR 1986b); much of the data summarised here is taken from that source.

5.1 Users of the RF Spectrum

The internationally adopted classifications for users of the RF spectrum are shown in Table 2, together with brief explanations of the terminology. Most nations adopt the following order of priority for spectrum allocation (Weisz 1989):

(1) military services;
(2) aeronautical and maritime emergency communications, radio navigation and radio location services;
(3) public safety services, eg police, fire, ambulance;
(4) national telecommunications services;
(5) broadcast radio and TV services;
(6) private user services, eg mobiles;
(7) others.

Any radio emission for any of the services listed in Table 2 must have a type designation; again, these designations are accepted internationally. A 3-symbol code is used,

1st Symbol: type of modulation of the main carrier

2nd Symbol: nature of signals modulating the main carrier

3rd Symbol: type of information to br transmitted.

For example, the designation "A1B" defines amplitude modulated automatic telegraphy by on-off keying, without the use of a modulating audio frequency, whilst "F2A" indicates frequency modulated telegraphy by on-off keying of a frequency modulating audio frequency, or frequencies, or by on-off keying of a modulated emission. Table 3 lists the complete set of designation symbols.

5.2 Spectrum Utilisation and Efficiency

If the RF spectrum is to be managed and conserved effectively, it is necessary to be able to quantify the efficiency of radio system spectrum utilisation. The simple fact that spectral "space" is occupied by a system says nothing about how efficiently the spectrum is being used. In general terms, therefore, "spectrum efficiency (SE)" is represented by

$$SE = \frac{\text{[Information actually transferred by a radio system over a range]}}{\text{[the amount of spectrum space used]}} \qquad (9)$$

System design features which enhance the EMC of radio systems include. antenna directivity, geographical separation, frequency sharing, multiple-access techniques, diversity, RPC, bandwidth-efficient modulation schemes, etc; thus, a measure of spectrum utilisation, or spectral space occupied, should reflect these considerations. Spectrum utilisation (U) is therefore defined as:

$$U = B\,S\,T \qquad (10)$$

where B is the occupied radio bandwidth, S is the geometric space associated with the system and T is the transmission time. S may be a line, an area, a volume or an angular sector, depending upon the nature of the system being considered. Hence, using (9) and (10):

$$SE = \frac{\text{[Information transferred over a range]}}{B\,S\,T} \qquad (11)$$

In this form, the definition of SE is very general; it needs to be refined and applied to each specific system. One way of quantifying the numerator of (11) is to use equation (1) in the form of the channel capacity, C:

$$C = I/T = B \log_2 [1 + SNR] \text{ bits/s} \qquad (12)$$

Then, (11) can be expressed, using (12), as

$$SE = (C\,D) / (B\,S\,T) \qquad (13)$$

where D is a range over which a specified SNR can be maintained.

5.3 Protection Ratios

The performance of any radio system will, to a first approximation, be determined by the SNR achieved at the receiver. In the EMC context, "noise" also has to include interference. The "protection ratio" (P) for a radio transmission is defined as the minimum value of SNR which will provide a defined grade of service, ie

$$P = \frac{\text{[Mean power of wanted signal]}}{\text{[Mean power of noise]}} \qquad (14)$$

As an example of the application of (14), the values of P required for voice communication with different grades of service are listed below.

Grade of Service	P (dB)
Minimum interference threshold	> 32
Good commercial quality	32
Marginal commercial quality	14
Threshold of intelligibility	9
Operator-to-operator (just usable)	5

These ratios assume an interference source which can be reasonably modelled as Gaussian white noise (GWN); if this is not the case, the characteristics of the particular interference source may modify the ratios quoted. Fig. 16 is an LOS radio system example, showing the way in which the ratio P can be used to establish the minimum separation between wanted and interfering signal transmitters; here, d is the maximum range at which the wanted signal can be received with the required grade of service when the interfering transmitter is at the location shown.

The effect of a noise/interfering signal on a wanted signal will depend upon the degree of overlap between their two spectra, ie their frequency offset. This effect is quantified by a "frequency-dependent rejection" (FDR), which is a function of receiver selectivity. Considering the situation shown in Fig. 17: U(f) is the spectrum of the interfering signal, centred on frequency f_u; similarly, W(f) is the passband of the wanted signal receiver, centred on f_w. The FDR, which is a function of the frequency offset, f_o, given by

$$f_o = \mathrm{mod}\,[f_w - f_u] \qquad (15)$$

is itself defined by

$$\mathrm{FDR}(f_o) = 10 \log_{10} \frac{\int_0^\infty W(f)\,df}{\int_0^\infty W(f)\,U(f + f_o)\,df} \qquad (16)$$

The FDR is effectively a measure of the reduction in the interference potential of an interfering transmission as the frequency separation increases. In turn, the FDR can be related to an increase in range over which a desired SNR can be maintained as the frequency separation f_o increases.

Another way of expressing the same concept is by means of a "relative RF protection ratio" (A), where

$$A = P(f_o) - P(0) \qquad (17)$$

Here, $P(f_o)$ is the protection ratio, or effective SNR, when the two signals are separated by f_o, and P(0) the ratio when they both have the same centre frequency.

In order to employ the concepts introduced in this section in a radio system EMC analysis, use must be made of the propagation models described in Section 3, together with the system design parameters outlined in Section 4.

5.4 Regulation & Control of the Radio Spectrum

The International Telecommunications Union (ITU) is an agency of the United Nations with responsibility for overseeing all aspects of international radio spectrum allocation and usage. This is achieved primarily via World Administrative Radio Conferences (WARCs), where member nations of the ITU take part in the decision making. Decisions of the WARC are ratified by member nations, and promulgated by the ITU.

The objectives of the ITU include:

(a) to promote co-operation between member governments;
(b) to allocate the radio frequency spectrum to different user groups and services;
(c) to co-ordinate efforts to eliminate detrimental interference between spectrum users;
(d) to encourage measures for ensuring the safety of life.

Under the administration of the ITU are two major committees, ie the International Telegraph & Telephone Consultative Committee (CCITT) and the International Radio Consultative Committee (CCIR). Both are tasked with studying technical and operational questions, issuing recommendations for equipment, signal and control specifications, and ensuring compatibility between nations; both make extensive use of study groups, working parties, etc staffed by national experts. In general terms, CCITT is concerned with systems and techniques making use of "guided" propagation, whilst CCIR covers systems and techniques employing "unguided" propagation, ie radio.

The CCIR holds a Plenary Assembly every 4 years at which the documents produced by the study groups are ratified as appropriate; these are then published in the form of "green books". Various volumes are published, relating to the areas of interest of the study groups, eg

Volume I:	Spectrum utilization and monitoring;
Volume II:	Space research and radioastronomy;
Volume III:	Fixed service at frequencies below about 30 MHz;
	etc.

Another organisation, operating under the auspices of the ITU, is the International Frequency Registration Board (IFRB), whose main task it is to implement the frequency planning and co-ordination decisions of the WARC.

In addition to spectrum control and regulation at an international level, there needs to be a corresponding degree of spectrum management and enforcement exercised at national level. In the United States, for example, this function is carried out by the by the Federal Communications Commission (FCC) and the National Telecommunications and Information Administration (NTIA); in the United Kingdom by the Department of Trade and Industry (DTI) Radio Regulatory Division and the Ministry of Defence; etc.

NATO itself is not a recognised body at the ITU; therefore NATO's interests must be represented by a co-ordinated position agreed to by the representatives of the NATO nations, with the co-ordinating body being the Allied Radio Frequency Agency (ARFA) (Fitzsimons 1986). There is only one frequency band allocated for exclusive NATO use, ie 225 - 440 MHz.

6. CONCLUDING REMARKS: TRENDS & FUTURE EMC CONCERNS

The topic of "Spectrum management and conservation" is vast; in a relatively brief lecture, it is only possible to outline the essential framework of the discipline. For this reason, a comprehensive list of relevant references has been provided for more detailed reading. In conclusion, it is perhaps worthwhile to indicate a number of specific trends and areas which may well give rise to radio system EMC concerns in the future.

6.1 Pressure on Spectrum Allocation

As indicated in (Darnell 1991) and (Fitzsimons 1991), there will be increasing pressure placed upon the spectrum allocation process by the unprecedented increase in civil radio systems, particularly for portable and mobile purposes. Civil demands will create a situation in which bands previously employed primarily for military purposes may have to be released.

To some extent, this situation will be eased by the progressive exploitation of higher frequency bands, eg 30 - 70 GHz; this is made possible by the ability to fabricate devices capable of significant power generation at these higher frequencies. However, there will undoubtedly be a still greater impetus in the direction of radio system designs providing improved spectral efficiency (see Section 4) and an increased use of civil communications facilities, such as PTT networks, by the military. The requirements for security and encryption, which previously gave rise to dedicated miltary digital systems, will be met in part by future civil digital systems. Therefore, mobile terminals compatible with say the Integrated Services Digital Network (ISDN) may well fulfill important operational roles.

6.2 Channel Sharing Techniques

With limited spectrum space available to satisfy an increasing number of more sophisticated requirements, techniques for channel sharing are assuming greater significance. In the military UHF band, single channel bandwidths have reduced from 100 kHz to 25 kHz and less over the past few decades as a result of improved system design - the main elements of which are better selectivity, lower speech digitisation rates and greater frequency stability; this is a trend which can be expected to continue.

In the satellite communication context, effective frequency sharing within the transponder bandwidth, via various multiple-access techniques, has been practised by NATO and other operators for many years (Campanella 1989). The main techniques employed are variants on basic TDMA to take account of different demand levels and contention problems, FDMA and CDMA. In FDMA, the chief EMC disadvantage is the generation of IPs as a result of satellite non-linearity. Any future proliferation of Very Small Aperture Terminals (VSATs) will require more sophisticated channel sharing techniques and control protocols (Naderi and Wu 1988), including multi-beam antennas and composite multiple-access schemes, eg FDM/TDMA.

In the field of terrestrial mobile radio, much activity is directed at civil cellular networks, such as the pan-European system; here, channel re-use is vitally important since it has a direct effect on the viability of the system (British Telecom Research Laboratories 1990). Multiple-frequency trunking techniques, in which mobile users, under the control of a base station, are shifted dynamically between individual channels in an allocated set are employed in land mobile services. One factor which is likely to be more widely exploited for improved frequency re-use, especially for military purposes, is the detailed nature of the terrain over which the system is operating. The more widespread availability of digital terrain maps will allow predicted propagation and interference levels for an interconnected LOS network of mobile terminals and base stations to be computed in real-time. This, in turn, will allow adaptive selection of antenna polarisations, operating frequencies, transmission rates, etc. For BLOS systems, the application of integrated RTCE techniques will enable adaptive frequency selection, variation in transmission rate, RPC, etc to be carried out.

6.3 Introduction of Wideband Systems

Increasing numbers of wideband (frequency-hopping or direct-sequence spread-spectrum) systems are being introduced into frequency bands previously only occupied by more conventional narrowband systems. The levels of mutual interference between these two classes of systems have been a matter of study by NATO for a number of years (Fitzsimons 1991). Further work is required to quantify this situation more precisely in terms of co-channel and adjacent-channel interference over the complete range of propagation and operational scenarios.

In the HF band, a similar problem will exist as more adaptive BLOS systems, operating in response to RTCE data, are introduced into an environment comprising largely non-adaptive systems. The frequency management implications of this evolution need to be evaluated.

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Band No.	Frequency Range	Wavelength Range	Band Name	Range & Services
2 3	30 - 300 Hz 0.3 - 3 kHz	10^4 - 10^3 km 10^3 - 10^2 km	ELF VF	Several Mm: specialised low data rate communications, eg for submarines; sea-water penetration significant
4	3 - 30 kHz	10^2 - 10 km	VLF	Several Mm: radio navigation services; low data rate telegraphy for submarines
5	30 - 300 kHz	10 - 1 km	LF	A few Mm: radio navigation and broad-casting
6	0.3 - 3 MHz	10^3 - 10^2 m	MF	Several 100s of km: broadcast, coastal radio and mobile communications
7	3 - 30 MHz	10^2 - 10 m	HF	World-wide: via ionospheric skywave; shorter ranges via groundwave; data and voice communications; broadcast,mobile and amateur services
8	30 - 300MHz	10 - 1 m	VHF	Line-of-sight primarily: also out to 2000km via scatter mechanisms; data and voice communications; broadcast, mobile and amateur services
9	0.3 - 3 GHz	1 - 0.1 m	UHF	Line-of-sight primarily: also out to 400km via troposcatter; data and voice communications; broadcast, mobile,amateur, satellite and long-range radar services
10	3 - 30 GHz	10 - 1 cm	SHF	Line-of-sight primarily: also out to 400km via troposcatter; satellite services, radar
11	30 - 300GHz	10 - 1 mm	EHF	Line-of-sight: high atmospheric attenuation; research and experimental mainly, eg for mobile communications
12	0.3 - 3 THz	1 - 0.1 mm		Research

NOTE: The band number "N" covers the frequency range 0.3. 10^N to 3. 10^N Hz; the upper limit is included in each band, whilst the lower limit is excluded.

Table 1

List Of Services

Broadcast :	Radio propagation intended for direct reception by the public, eg AM and FM radio, VHF and UHF TV
Fixed:	Point-to-point service, eg HF , satellite, microwave radio relay
Mobile:	Service between mobiles, or between fixed stations and mobiles
Aeronautical Mobile:	Fixed / aircraft, or aircraft / aircraft
Land Mobile:	Base / land mobile, or land mobile / land mobile; including portable units
Maritime Mobile:	Ship / shore, or ship / ship
Satellite:	Normally earth-satellite-earth, for a variety of applications
Standard Frequency & Time Signals:	Radio transmission of stated high-precision frequency and time standards, intended for general reception, eg MSF and WWV
Radio Navigation:	Radio transmission for position determination for navigation and obstruction warning
Radio Location:	Radio transmission to determine position, velocity, etc, eg radar and radio altimeter
Radio Astronomy:	Astronomy based upon the reception of radio waves of cosmic origin
Amateur:	Radio systems for technical / non-commercial interest
Industrial, Scientific &Medical	Significant radio emissions from ISM systems must lie within specified frequency bands

Table 2:

RADIO EMISSION DESIGNATIONS

1st SYMBOL

N Unmodulated carrier

Main Carrier AM

A DSB
H SSB, full carrier
R SSB, reduced carrier
J SSB - SC
B ISB
C Vestigial SB

Main Carrier FM

F FM
G PM
D Simultaneous AM & angle modulation

Pulse Emission

P Unmodulated pulse sequence
K PAM
L Pulse width/duration modulation
M Pulse position/phase modulation
Q Angle modulated carrier during pulse period
V Combination of above pulse modulation types, or other

Other

W Carrier modulated by combination of 2 or more of amplitude, angle or pulse
X Cases not otherwise covered

2nd SYMBOL

0 No modulating signal
1 Single channel containing quantised or digital information without use of modulating sub-carrier
2 Single channel containing quantised or digital information with use of modulating sub-carrier
3 Single channel containing analogue information
7 2 or more channels containing quantised or digital information
8 2 or more channels containing analogue information
9 Composite system with 1 or more channels containing quantised or digital information, together with 1 or more channels containing analogue information
X Cases not otherwise covered

3rd SYMBOL

N No information transmitted
A Telegraphy - for aural reception
B Telegraphy - for automatic reception
C Facsimile
D Data transmission, telemetry, telecommand
E Telephony (including sound broadcasting)
F Television (video)
W Combination of any of the above
X Cases not otherwise covered

Table 3

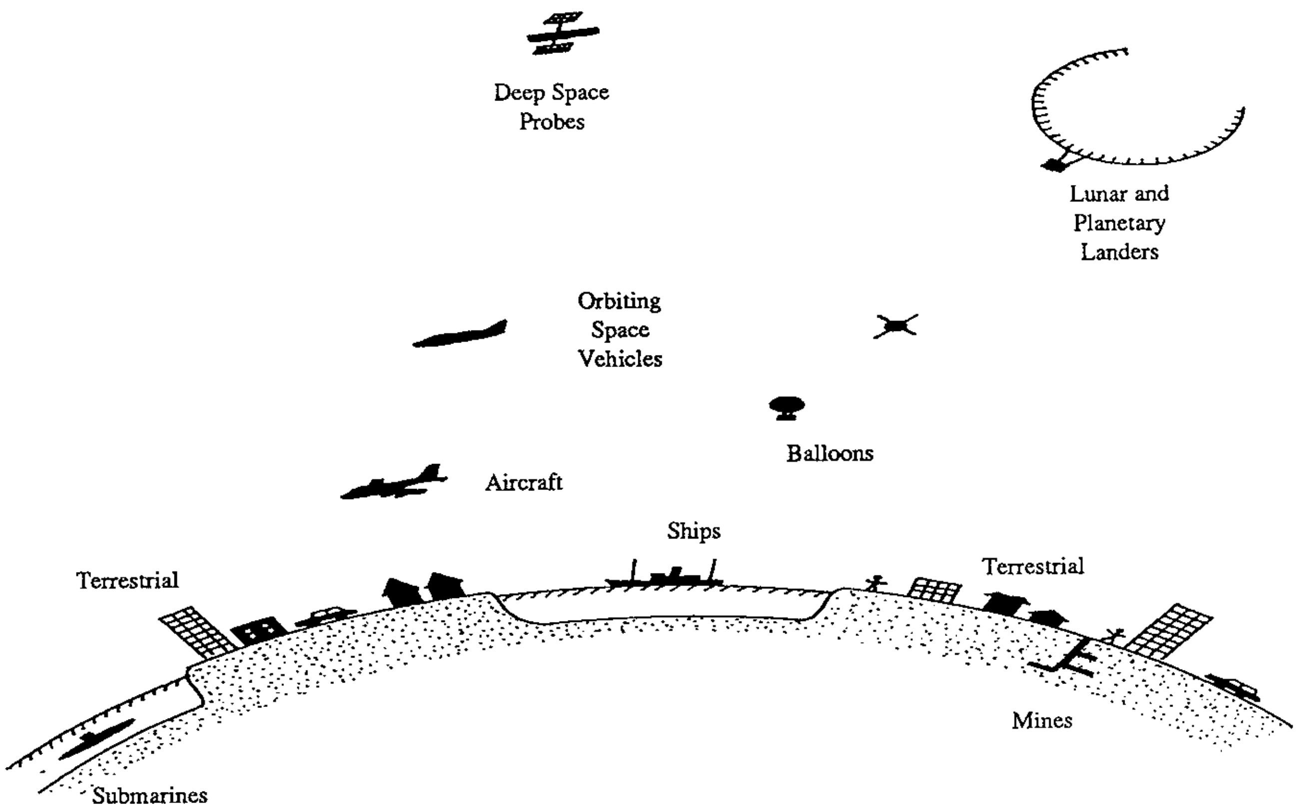

Figure 1: Variety of Radio Services

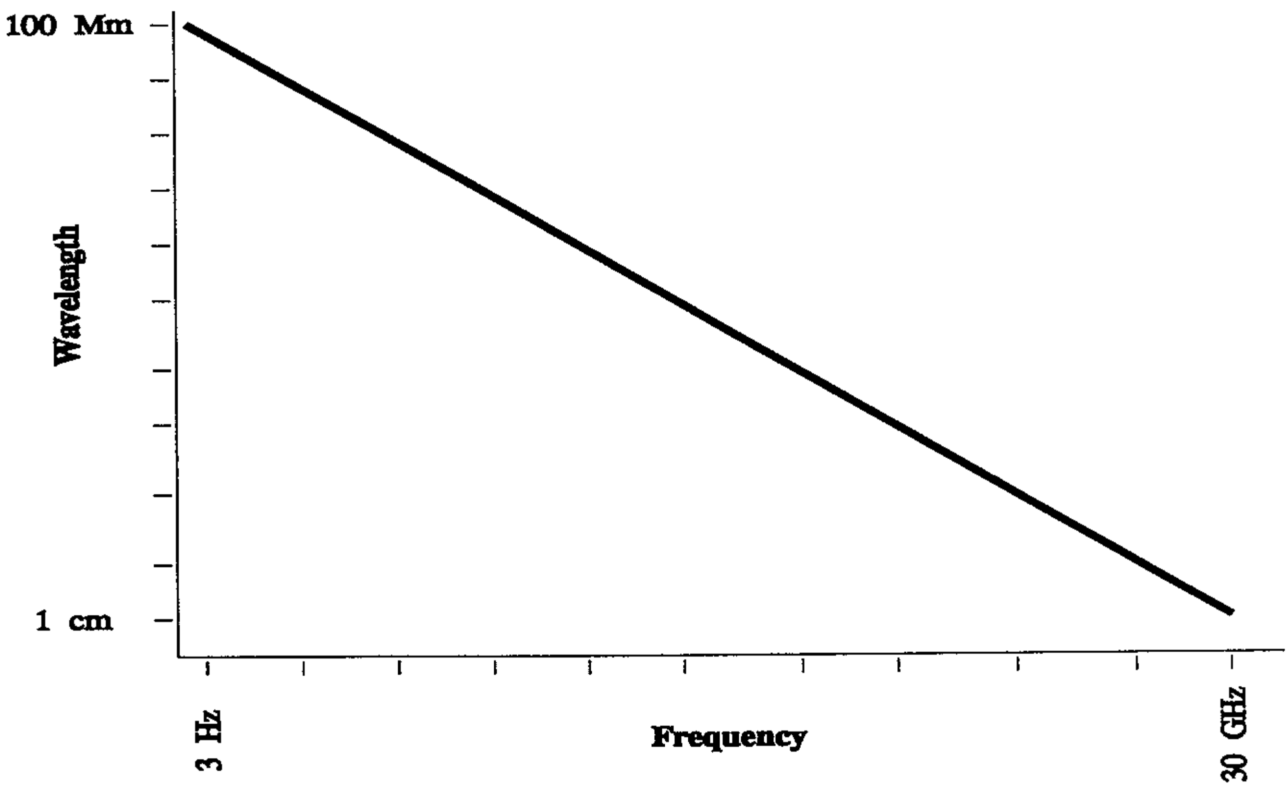

Figure 2: Relationship between wavelength and frequency for EM waves

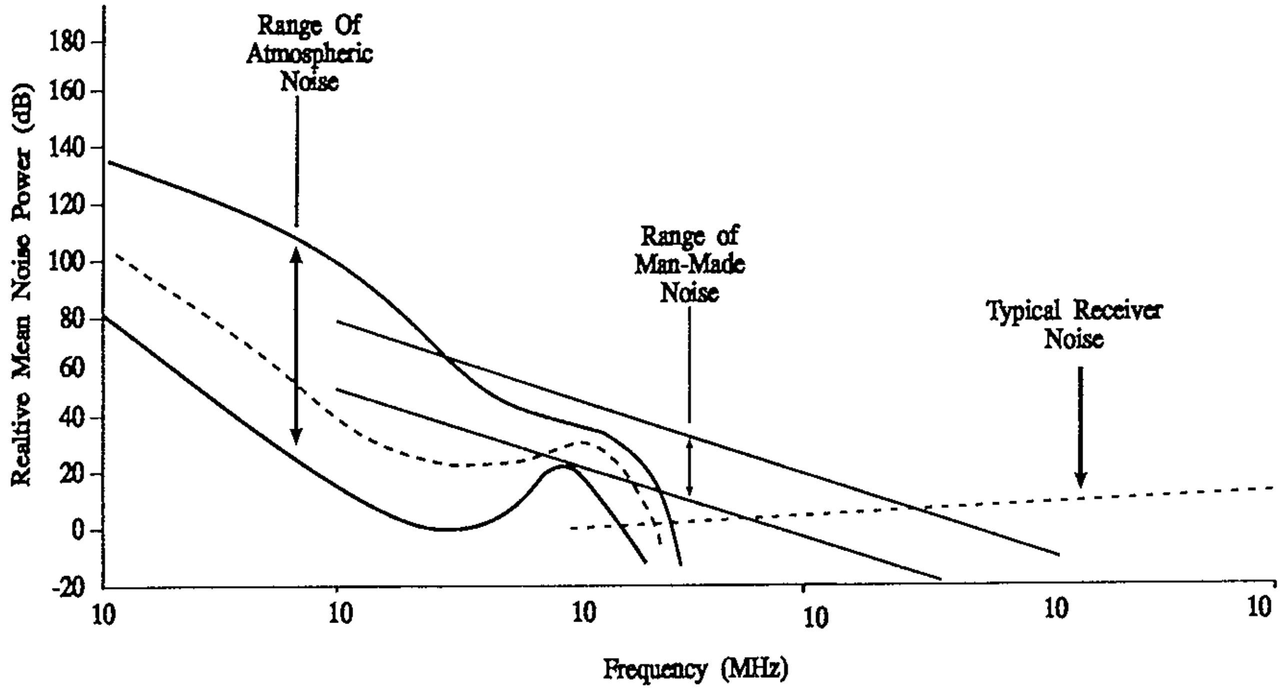

Figure 3: Variation Of Noise Power Level With Frequency for various sources

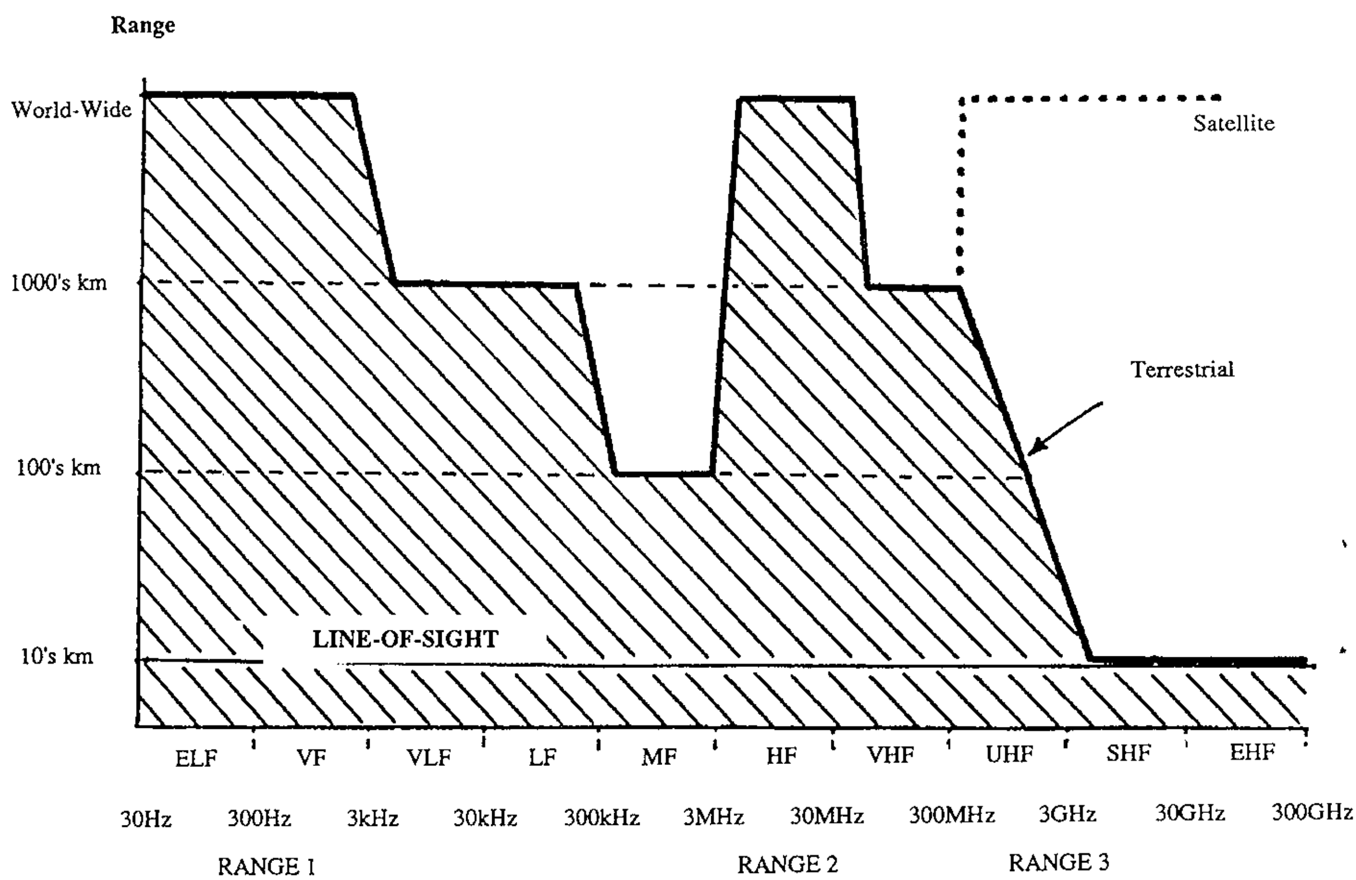

Figure 4: Variation of Radio System Range With Frequency

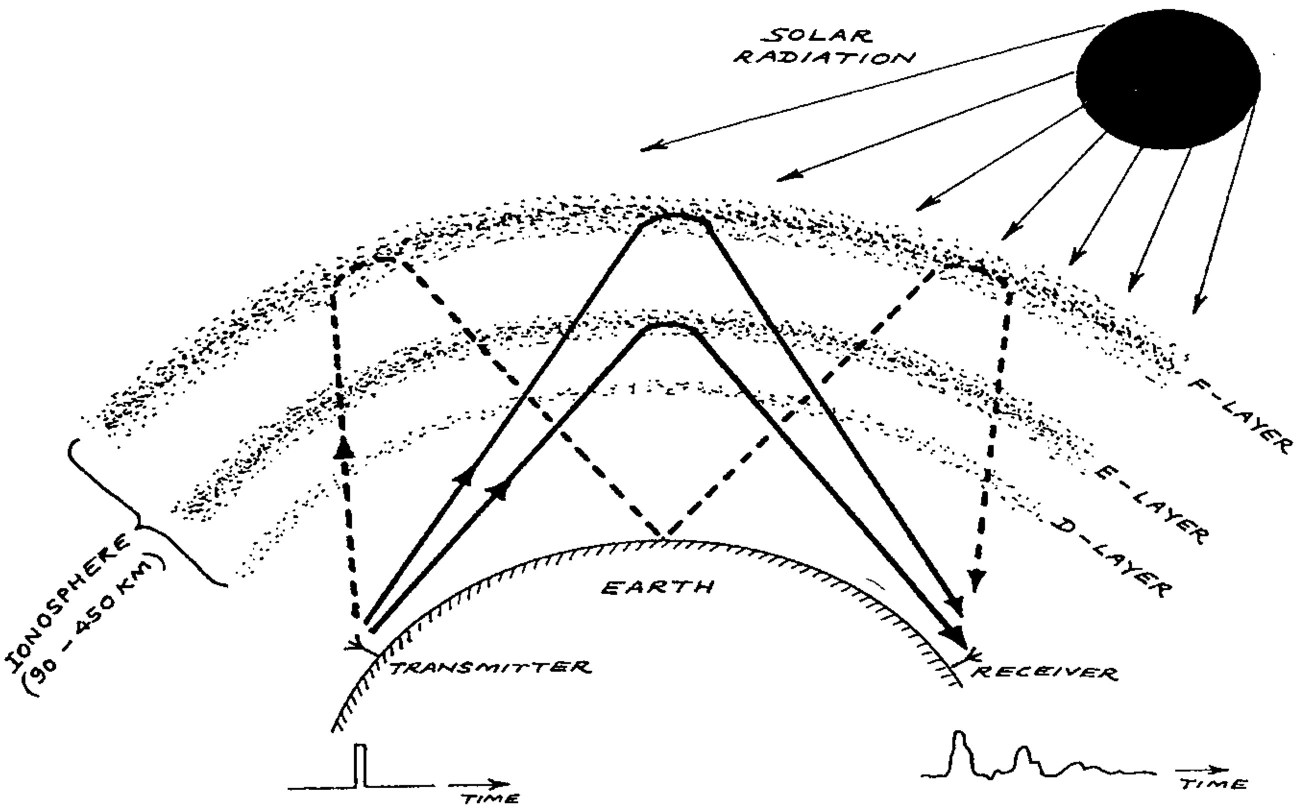

Figure 5: Ionospheric Skywave Propagation

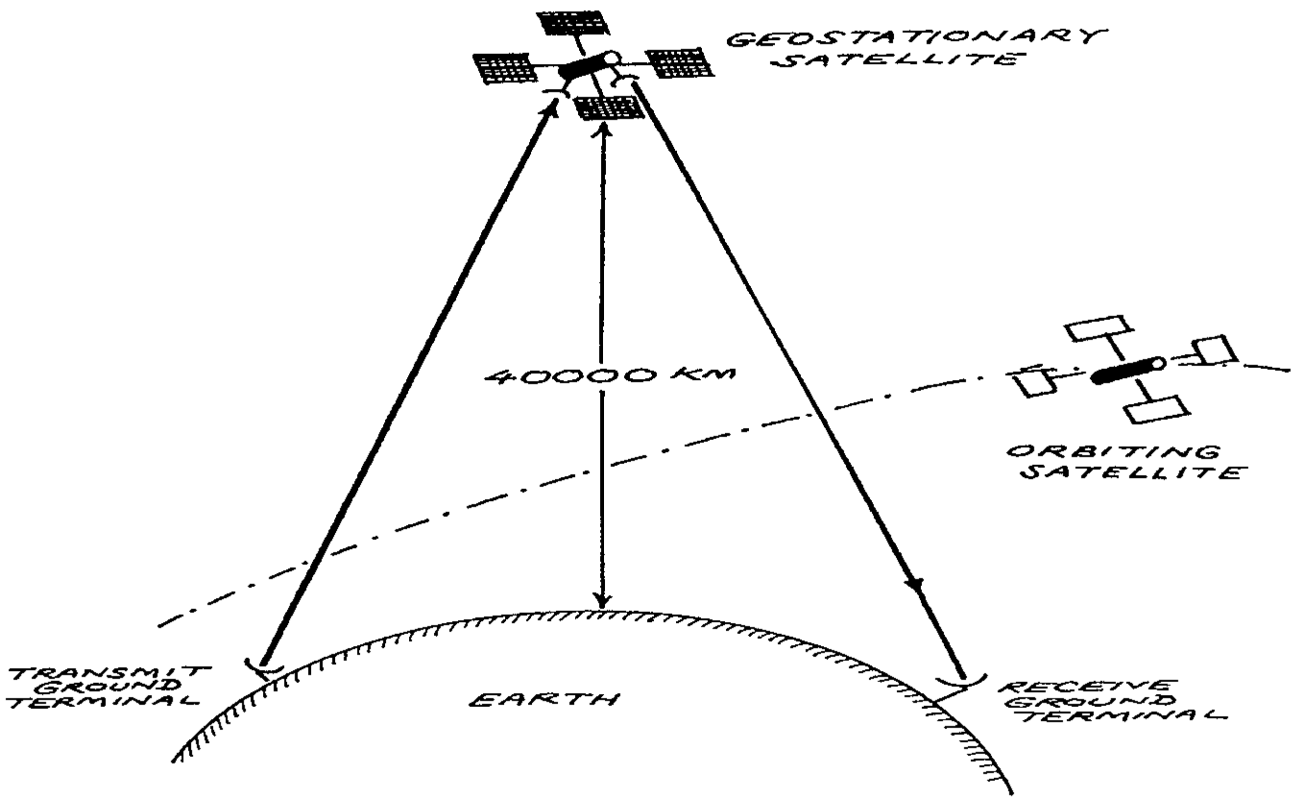

Figure 6: Satellite Communications LOS System

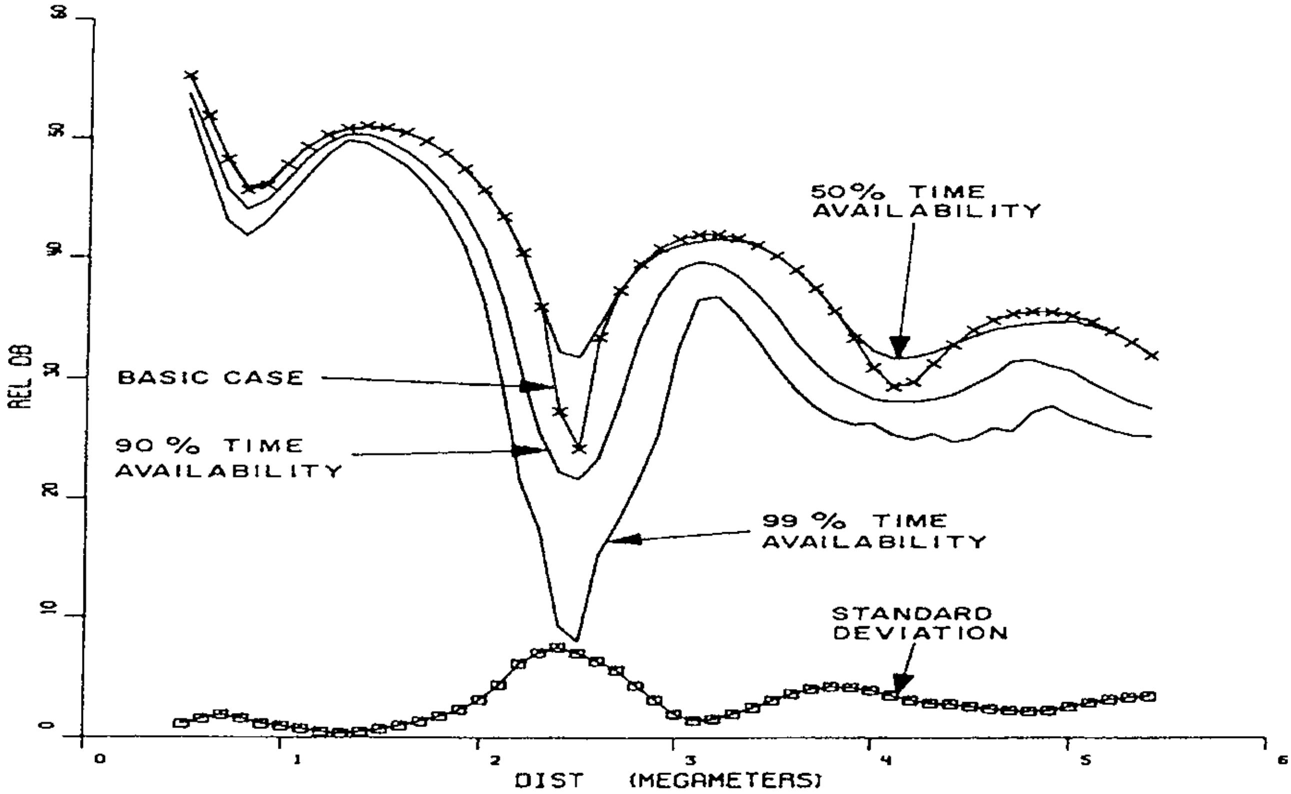

Figure 7: VLF Field Stength variations with Distance (Kelly 1986)

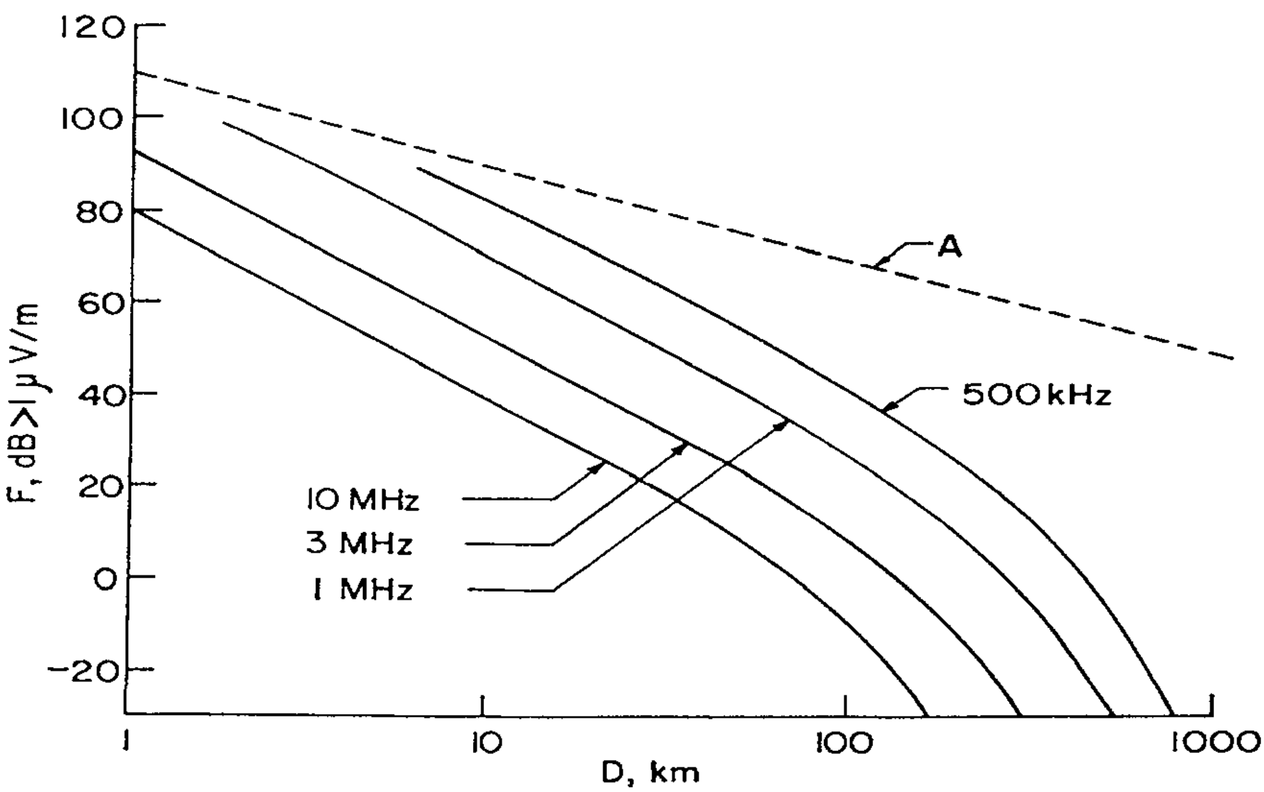

Figure 8: Groundwave Field-Stength variation with Distance (CCIR 368-2)

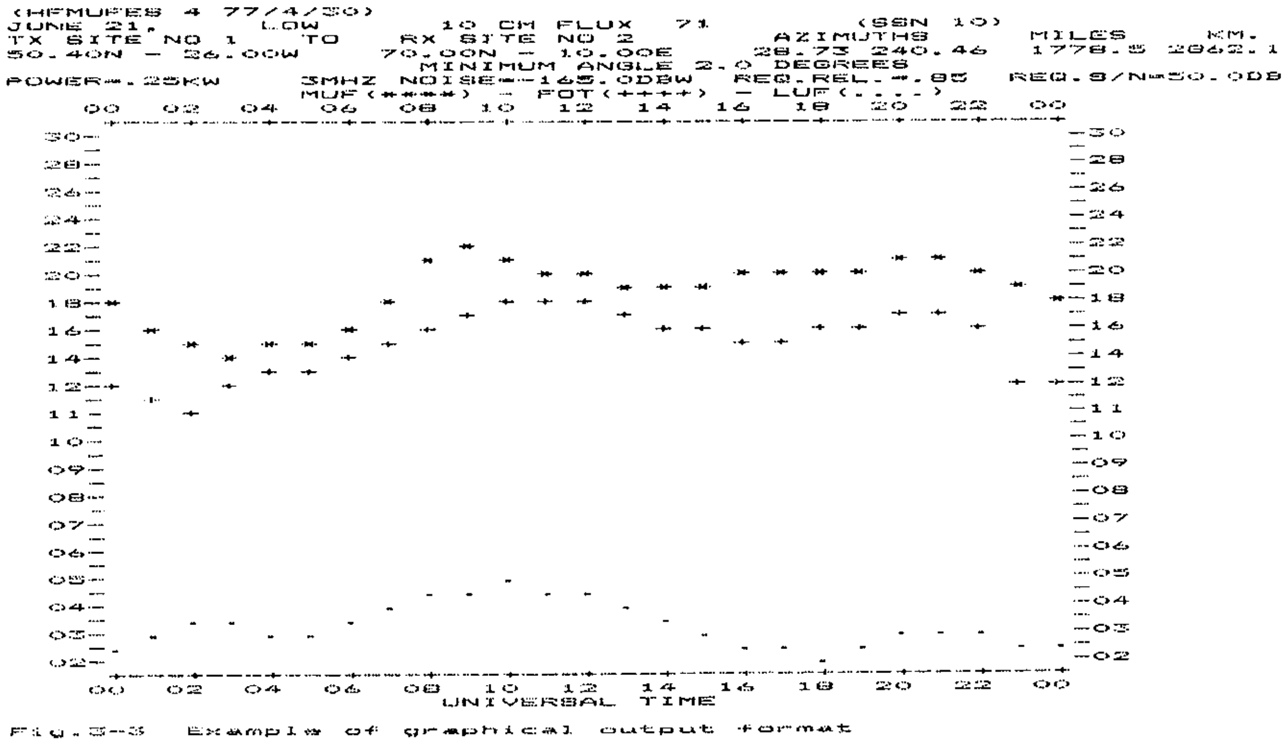

Figure 9: Example of HF Prediction Program Output

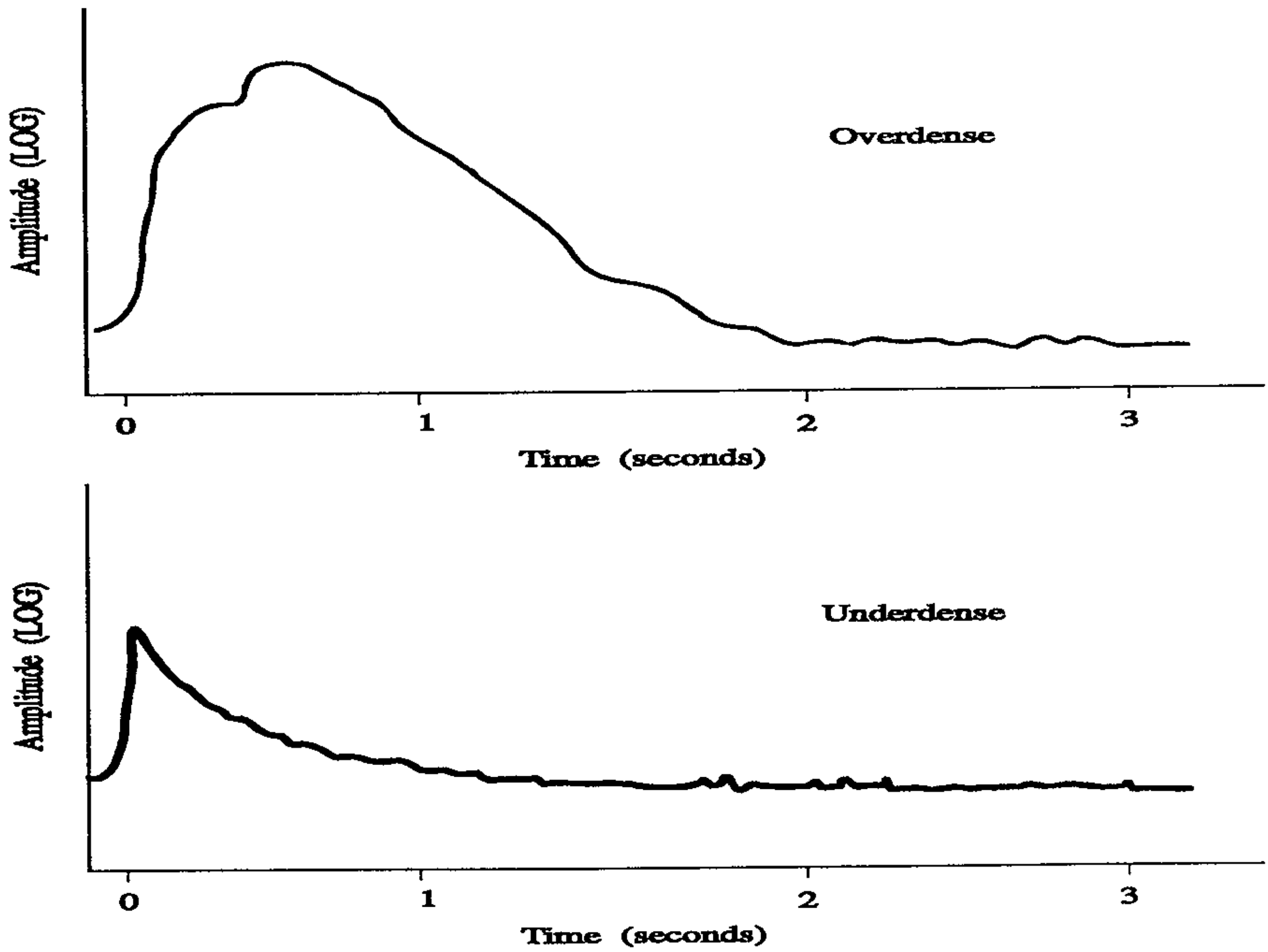

Figure 10: Amplitude-Time Profiles of signals via meteor trails

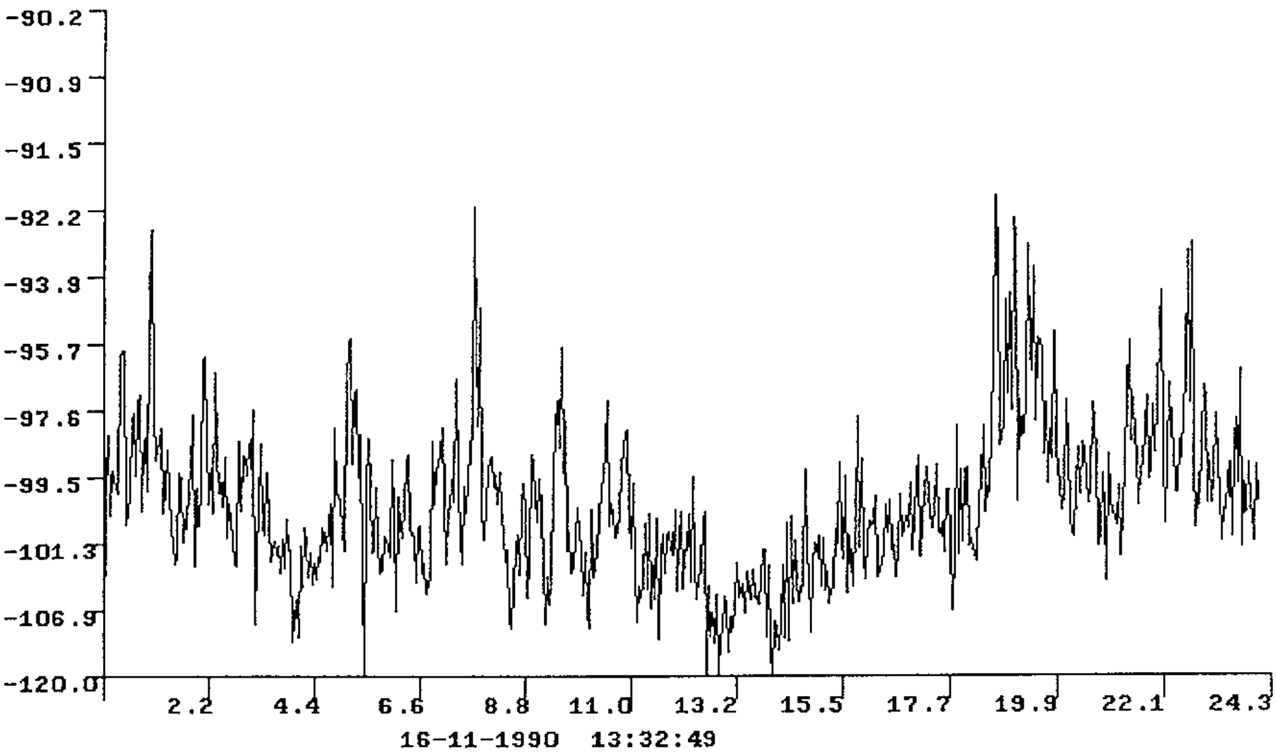

Figure 11: Diurnal variation of 47MHz Tropscatter received signal level

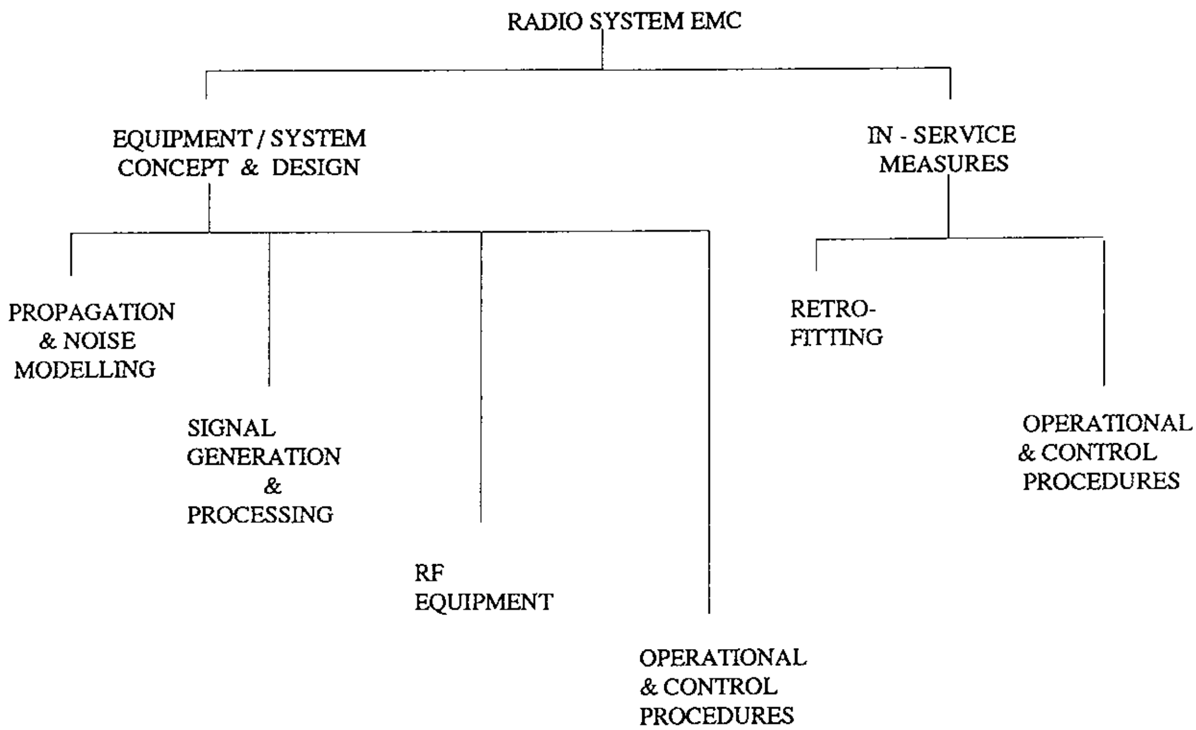

Fig. 12 Elements of radio system EMC

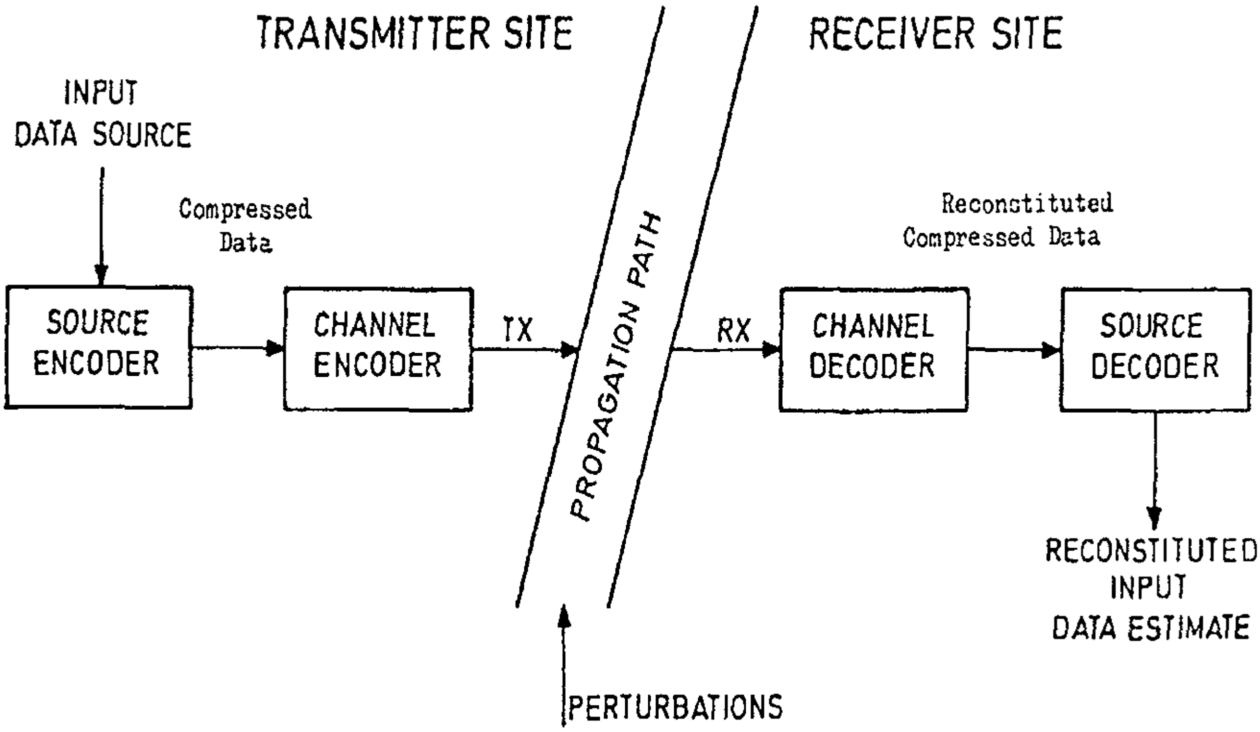

Figure 13: A Generalised Information Transfer System

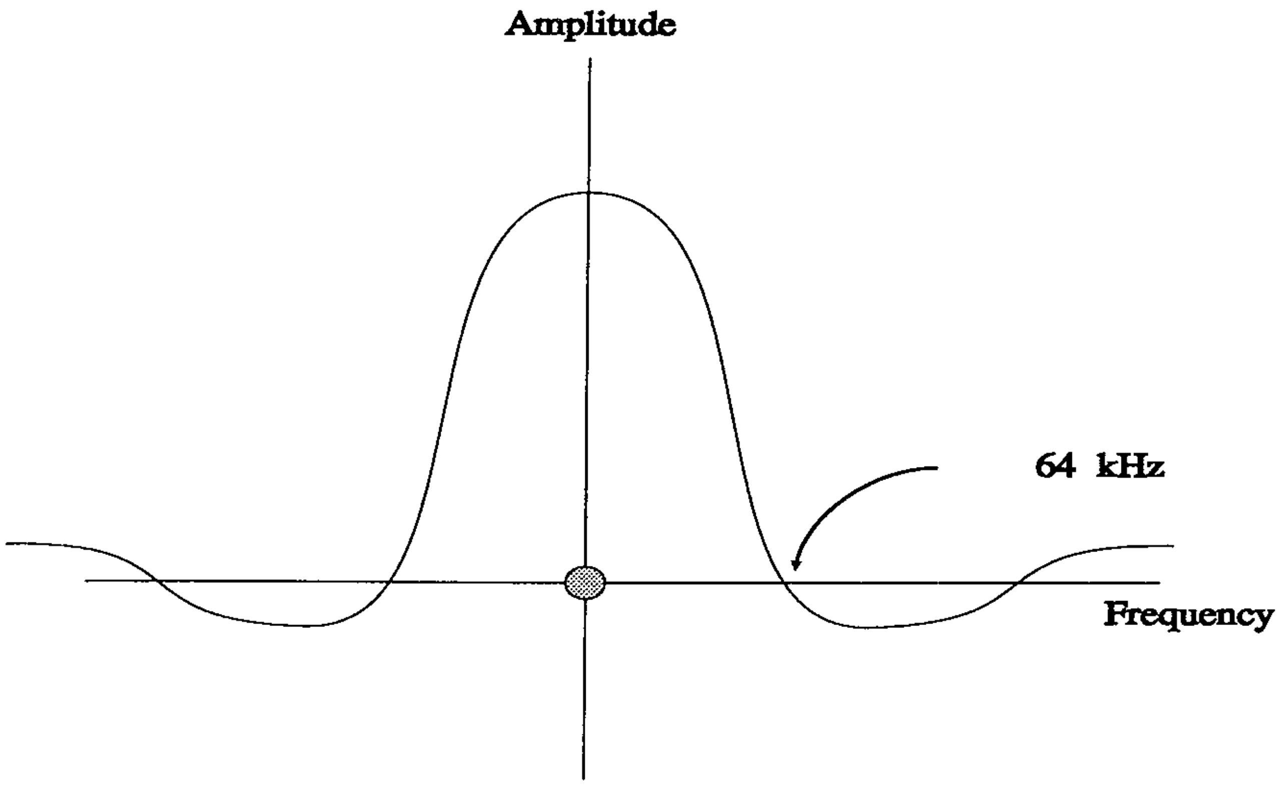

Figure 14: Spectrum of 64 kbits/sec digital signal

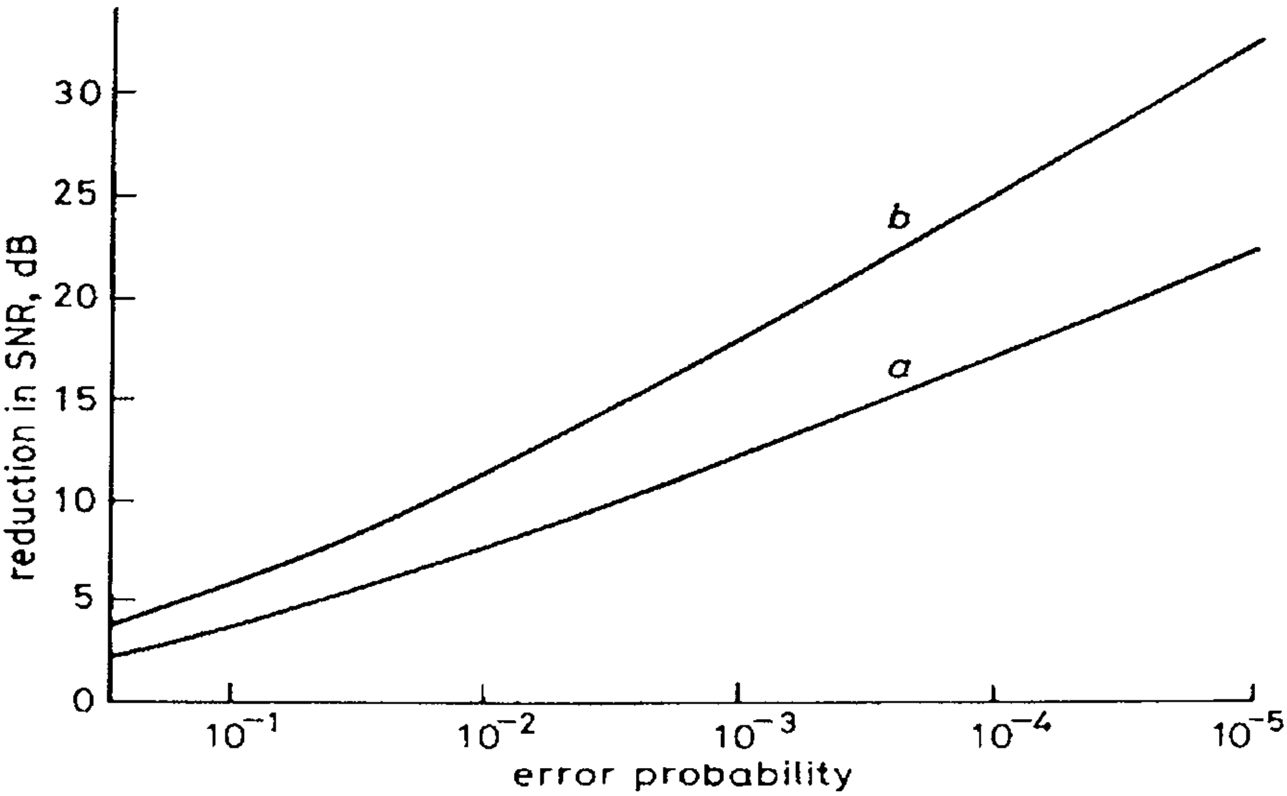

Figure 15: Reduction in SNR required at receiver using 2nd and 4th order diversity

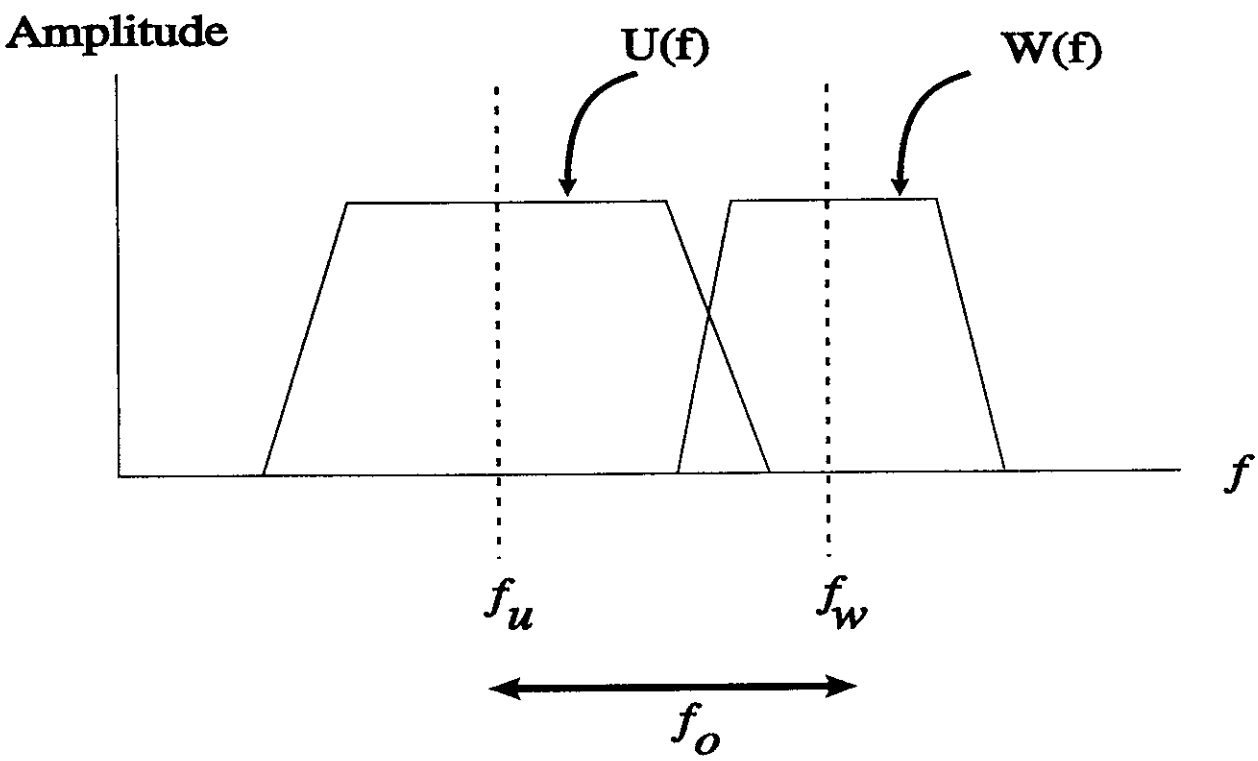

Figure 16: Illustration of the case of the protection ratioFigure 17: Illustration of frequency dependent rejectionFigure 17:

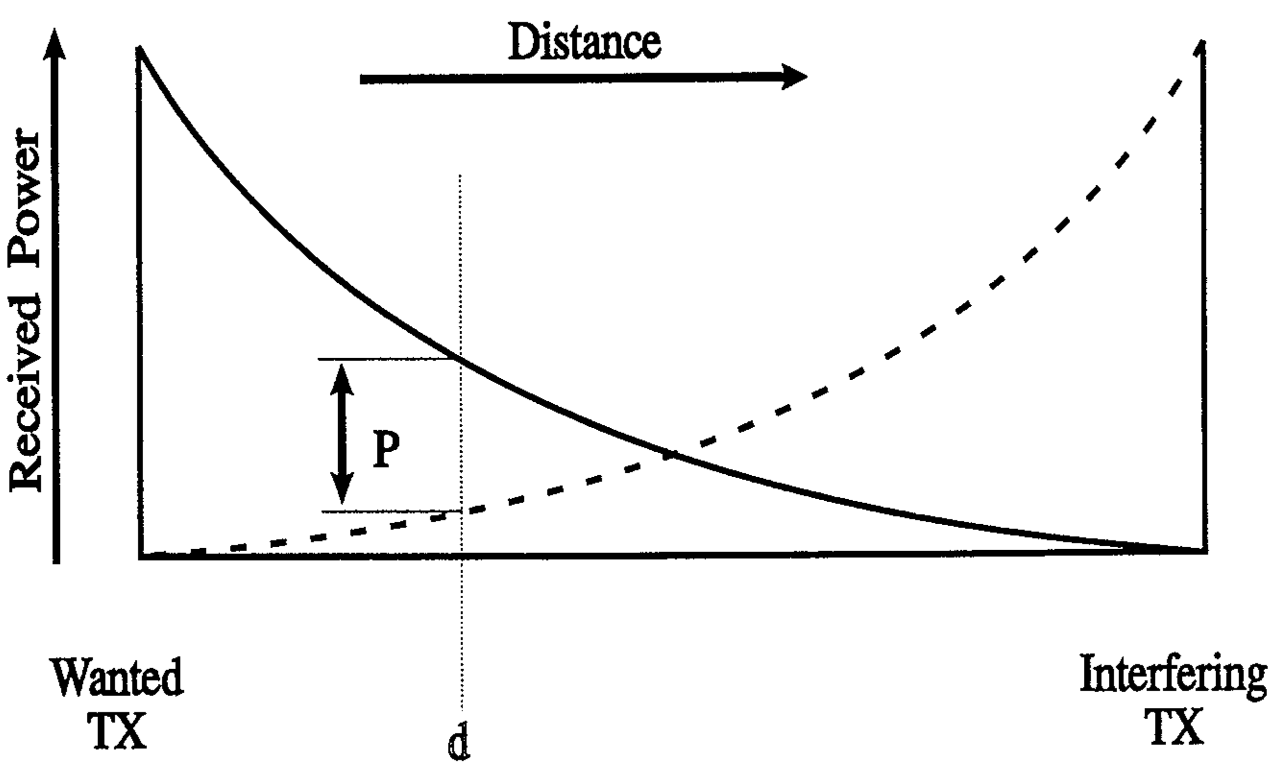

Figure 17: Illustration of frequency dependent rejection

MEASUREMENT ENVIRONMENTS AND TESTING.

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Summary.

This lecture will describe the various methods used to assess both the emission (interference generation) performance of electronic equipment, and the immunity of electronic equipment to external electromagnetic interference.

The measurement methods attempt to simulate realistic operating conditions for the equipment under test, yet at the same time they must be repeatable, and practical to operate. This has led to a variety of test methods being developed, each of which has its limitations. The lecture will concentrate on the most common measurement methods such as open-field test sites, screened enclosures and TEM cells, and describe the physical justification for the methods, their limitations and measurement precision.

Ways of relating similar measurements made by different methods will be discussed, and some thoughts on future measurement improvements will be presented.

1. Introduction.

1.1 Historical Background.

The requirement for EMC testing arises from the need to control the electromagnetic environment. The control is necessary in order to ensure that systems operating in that environment can do so without causing or suffering from electromagnetic interference. The history of relatively large scale electrical and electronic technology, with the large scale use of electrical power and communications technology, is confined to the twentieth century. The timescale has also coincided with other mass technological developments such as land and air transport based on various forms of internal combustion engine. Whenever the use of technology expands in this way, the users of the technology become less concerned in the technology itself and more concerned with their own use of the technology as laymen. Regulation of the use and performance of the technology then becomes necessary in order to ensure a safe and equitable service. The regulation is in the hands of national and international bodies which develop and promote the use of standards for equipment design and use. Part of the regulation process is the development of performance standards and the test methods required to ensure compliance with the standards.

The development of EMC standards began in earnest with the setting up of CISPR, the International Special Committee on Radio Interference in 1934. Various manifestations of electromagnetic interference had been apparent before that date. Interference between early radio systems was one impetus towards the development of narrow bandwidth tuned cw systems as opposed to inherently broadband spark systems which took place during and immediately after the First World War. The parallel large scale developments of broadcast radio and motor transport, which took place during the nineteen twenties, resulted in the requirement for suppression of impulsive noise from electrical ignition systems on vehicles and, as such, represents the first contact by the public at large with an EMC problem. This is also an early example of an interference problem between two engineering systems, one of which is neither an overtly electrical or tele-communications system. The quasi-peak detector, developed by CISPR to quantify the audio annoyance factor of this impulsive interference problem, and described below, lingers on today in EMC measurement specifications. The early EMC specifications did not need to address disturbance to other than radio and analogue equipment as the thermionic valves and electro-mechanical switching used for the processing of signals required substantial energy inputs before malfunction occurred.

The situation, as it exists today, has arisen from this historical background. In effect, two almost completely separate EMC communities exist. The commercial and domestic electronics industry has slowly evolved an EMC community as indicated, primarily in the United States through the Federal Communications Commission (FCC), and Europe with (Western) Germany taking a leading European role through the VDE. This community has received a substantial boost recently on both sides of the Atlantic with the development of the European Community Directive on EMC, now to be fully implemented in 1995.

In contrast, the rapid developments of military electronics that occurred from 1934 onwards with the parallel development of radar in several countries led to substantial military EMC problems. As a result a military EMC community developed which achieved maturity much earlier than the civilian sector. The US Military Standards for EMC are widely accepted throughout NATO and form the basis of other national standards, for example the British Defence Standards on EMC. While trying to achieve the same aims, the test methods and specified electromagnetic field levels are very different for the two communities, the military standards in general being rather more stringent.

1.2. Considerations Leading to Test Methods.

Let us now consider the overall philosophy underlying EMC testing. The problems arise because unwanted energy couples from one system to another causing various levels of malfunction from slight performance degradation through to catastrophic failure. Each problem has an interference source and a victim. A simplistic approach is to state that no interference sources should exist, that is, systems should be so designed that interference is not generated. Bearing in mind that electromagnetic waves, the medium by which interference is propagated, are generated by any time varying electrical signal, such a requirement would place impossible design requirements on the screening and filtering of any system. Practical equipment will generate some interference. Potential victim equipment also has to be capable of operation in the presence of naturally occurring interference, and so must be designed with some level of immunity to interference. Thus, in order to achieve electromagnetic compatibility, two aspects can be identified, interference emission and immunity to interference. Each equipment must be assessed for each aspect.

The levels of interference generation allowable and the and the immunity required of equipment are set by the regulating authorities through specialist committees. These consider the likely operation of the equipment in terms of its surroundings, known external threats and known potential interference victims. Such factors as the distance to the nearest other equipment and the other types of energy propagation paths available (power and signal distribution networks) are considered. The interference levels are based on reasonable operation of the equipment. For example, it must be possible to operate a broadcast radio receiver in the same premises as a personal computer, and so the interference generated by the computer must allow an adequate signal to (noise + interference) ratio for the radio. However, it would be unreasonable to expect to operate a portable radio placed on top of the computer VDU.

The complexity and uncertainty of the environments in which electronic equipment is operated necessitates some simplification and partitioning for test purposes. The testing is first split into interference emission assessment and assessment of immunity to interference. (The term immunity is used interchangeably with the term susceptibility, with the military EMC community generally using the latter term.) The next split considers the energy propagation process, and this is simplified into radiated and conducted interference assessments. The case of capacitive/inductive coupling which forms an overlapping area between radiated and conducted interference is not considered separately. Four types of test are thus identified as shown in Table1.1 below.

Table 1.1. Types of EMC Test.

Conducted Emission	*	*Conducted Immunity*
*		*
Radiated Emission	*	*Radiated Immunity*

The types of signals are also considered. For measurements of emissions, the frequency domain analysis is always undertaken using special receivers, irrespective of the nature of the waveforms causing the emissions. These could be a set of cw oscillators in a communication system, or a set of harmonically related signals from a digital clock as illustrated in Fig 1.1. Both of these have line spectra. The emissions could also have random or noise like properties as generated by analogue or digital data signals. These have continuous spectra as illustrated in Fig 1.2. The problems associated with the wide range of signal types are addressed in the section on receivers.

For immunity measurements both time and frequency domain signal specifications are used, the time domain specified signals being restricted to conducted immunity tests. For example, a specification may require an equipment to be subjected to a modulated cw carrier swept or stepped across a given frequency range, or the waveform and repetition rate of an impulsive signal may be specified in the time domain. The two approaches are possible in the immunity case as the signals are directly under the testers control.

1.3. The Test Method Design Philosophy.

The design philosophy of EMC test methods merits some consideration before the various test methods are described. Most tests are of the "go, no-go" type where an indication greater than the specified level of interference is considered a failure and one below a pass, although in reality, some judgement may be used for results close to the specification level. In performing an EMC test we are assessing the probability of an interference problem existing. It is therefore more accurate to say that test procedures are performed rather than measurements, and that the outcome of the procedures indicates that either an EMC problem may occur, or that one is not likely to occur. The level of confidence in these assertions depends on the difference between the pass-fail level and the level of interference actually observed. This is not generally quantified.

In designing a test method we are attempting to mimic a realistic operational environment for the Equipment-under-Test (EUT). In general, the working environment of an EUT can only be defined in the broadest sense and so the test environment is, at best, a simplification of the real environment. It is only necessary to mimic the electromagnetic environment of the EUT so other factors such as temperature or humidity need not be included. If the ambient electromagnetic environment is present in the test, the assessment of interference may be subject to interference. In conducted interference measurements this is relatively straight forward, as both power and signal cables can be filtered. In radiated interference measurements exclusion of the ambient requires screened enclosures. If these are anechoic then they present no problems. At low frequencies however, anechoic performance cannot be obtained, and the test environment is degraded. Screened and filtered measurements are also required if signal security is an issue.

A further issue is that of measurement repeatability and precision. No two test environments are identical. There is, for instance, a multiplicity of screened enclosure sizes on the market, and test houses tend to buy the enclosure that fits their premises and budget. While the specifications may be quite detailed in its description of the layout and dimensions of the antennas etc used inside the enclosure, they reflect the multiplicity of sizes usually by only inferring a minimum size through minimum antenna to wall spacings. Open-field test sites used for radiated emission measurements are also of variable construction because of minimum rather than actual conducting groundplane size specifications. These constructional variabilities lead to poor repeatability of measurements between facilities, particularly in screened enclosures where enclosure resonances cause considerable variability. These effects are discussed in more detail in the sections describing the different measurement environments.

For the measurement of emissions, the test environment need only accommodate the emissions from the EUT in as realistic a way as possible. The emissions are then detected by an appropriate transducer and passed on to a receiver for subsequent analysis. The EUT is exercised throughout its range of operation. The frequency range of observation is determined by the assessment of the threat that the EUT poses. For example, a commercial/domestic radiated emission measurement is normally performed in the 30MHz to 1GHz range. The upper limit is set by the use of the radio spectrum in the likely environment of the EUT, as most terrestrial radio services operated by unskilled users are in this range. A plot of the ambient spectrum in a typical urban location at York, England is shown in Fig 1.3. The spectral energy present in most digital electronic equipment is also becoming insignificant above this frequency, although new systems with clock frequencies in excess of 100MHz are required to be tested to higher frequencies in the USA. Below 30MHz it is observed that most interference energy propagates in the form of conducted interference as the efficiency of the radiation process is reduced as the frequency reduces. In this frequency range conducted interference assessments are made. The military EMC community has more stringent requirements, and requires both radiated and conducted emission measurements to be made over a wider frequency range, although the conducted tests still occupy a lower frequency range than the radiated ones.

In the case of immunity assessments, the requirement to mimic the operational environment poses a further problem. The immediate electromagnetic environment of the EUT can be constructed in the same way as that for emissions assessment. For immunity assessments, the threat fields must also be present. In operation, the EUT will be illuminated by a variety of threat fields simultaneously. Each will have its own direction of arrival, polarisation, modulation and intensity. Except for specialist military threat simulators, the multiplicity of threat fields is simplified to either a single pulse train or modulated carrier swept over the predetermined frequency range. Clearly, this is a considerable simplification of the actual threat, but the economics of the test process generally precludes any more complex effort.

As with the emission assessment, the EUT must be exercised over all its functions. This is much more time consuming for immunity. For emission assessment, a frequency scan can be taken relatively quickly for each EUT state, as for each state, the emission spectrum is usually stable, and the scan rate is limited by receiver considerations. The receiver output data is presented in graph form, and a pass-fail judgement can be made quickly by comparison with the specification. For immunity assessment, the EUT must be assessed at all frequencies for each EUT state with sufficient dwell time for the tester or a data logger to observe any effect. Often the observation may be subjective. For example, the interference effect may be a degradation of a video display or audio signal, or it may be an instrument reading error comparable with the inherent accuracy of the instrument. Such assessments take time and are operator dependent.

2.0 Survey of Test Methods.

In this section the various types of test methods are reviewed. The section is divided according to the test types identified in section 1.2 with the addition of a section on receivers.

2.1. Radiated Emission Measurements.

Two techniques will be described for performing radiated emission measurements, the Open-Field Test Site (OFTS) and Screened Enclosures. The former is used primarily by the commercial/domestic EMC community, while the latter is used mainly by the military EMC community.

Specifications for radiated emissions are defined in terms of the radiated electric field incident upon a receiving antenna. The signal received by the antenna is measured as a voltage in a receiver operated with a specified bandwidth and detector type, and is related to the incident field by the antenna factor (A.F.) of the antenna defined as;

$$\text{A.F.} = E/V \quad m^{-1}$$

where E is the magnitude of the incident field and V is the receiver input voltage. The antennas are calibrated along with their interconnecting cable, and many receivers give their output as incident electric field rather than input voltage, the antenna factor being stored in a look-up table in the receiver.

A plan view of an OFTS as defined in CISPR Publication 22 [1] is shown in Fig 2.1. The site should be on level ground as far away from buildings and other potential reflecting objects as possible. The EUT to antenna spacing is fixed at 3m, 10m, 30m, 100m, with the first two being the most common. These distances are chosen to be representative of typical distances over which interference may occur. At 30MHz, the antenna to EUT spacing is $\lambda/3$ for a 3m site and λ for a 10m site where λ is the wavelength. The 3m site, in particular, is marginal for a 377Ω wave impedance assumption. The maximum EUT size is limited by the Rayleigh Range criterion of a minimum EUT to antenna spacing of $2D^2/\lambda$ where D is the maximum dimension of the EUT. This criterion is determined by the need to make measurements in the Fraunhoffer diffraction defined far-field of a radiating system. For measurements at 1GHz (λ = 0.3m) on a 3m site, the maximum EUT dimension set by the Rayleigh Range criterion is 0.7m.

The minimum clear area is defined by the EUT to antenna spacing by an ellipse with the antenna and the EUT at the foci, and the major and minor axes being 2.0 and $\surd 3$ times the focal distance of the ellipse. This geometry gives an EUT to ellipse perimeter to antenna distance of twice the focal distance and thus defines the maximum spurious reflection from objects outside the ellipse. In the worst case of a large specular reflector on the perimeter, tangential to the ellipse, the spurious reflected signal would be no more than half the direct signal. Within the ellipse the ground is required to provide a specular reflection and thus must be flat to within $\lambda/16$ at the shortest wavelength as a rule of thumb. As the highest frequency of operation is 1GHz, this requires a surface level to within about 20mm. The conducting groundplane is present to give a repeatable ground reflection, independent of the ground moisture content, over as much of the site as possible. Its minimum size, as specified in BS6527 [2], is shown in Fig 2.2. If possible the conducting groundplane should cover the whole site. It must be present over the area shown in order to provide;

i) a defined specular reflection at the mid- point between the EUT and the antenna.

ii) a defined reflection coefficient beneath the antenna to stabilise the antenna- ground interaction.

iii) a defined reflection coefficient beneath the EUT to stabilise the EUT-ground interaction.

On outdoor sites the conducting groundplane is made from a mesh material in order to aid drainage. A mesh size also less than $\lambda/16$ at the shortest wavelength (20mm) is required so that the mesh behaves as a continuous reflector.

A side view of the site is shown in Fig 2.3 which illustrates the various interactions between the antenna, the EUT and their images. From this it can be seen that the total field present at the antenna is the phasor sum of the direct wave and the ground reflected wave. The value of this sum depends on the path length difference between the direct and reflected paths in wavelengths. For a given antenna height this is dependent on frequency. If the value of the sum is plotted for a set of frequencies against antenna height, a result similar to that of Fig 2.4, recorded by FitzGerrell [3] at NIST Boulder, is obtained. The measurement specification calls for the maximum value of the resultant interference pattern to be recorded for each frequency. In the frequency range from 30MHz to 300MHz a broadband Biconical Dipole is often used as the antenna. From 300MHZ to 1GHz a log-periodic array is used. In some cases more precise measurements of the most significant emission frequencies are made with resonant $\lambda/2$ dipoles.

The measurements are repeated for both horizontal and vertical polarisation of the antenna, and the EUT is rotated in the horizontal plane to ensure that the worst case radiation is received. In many cases, the major repeatability problem is caused by the layout of power or signal cables effecting the radiated fields. For a typical 19" (0.5m) dimension equipment with a 2m long mains cable, the EUT is between $\lambda/4$ and 8λ in length. Individual circuit cards and sub-assemblies do not radiate in their own right to any great extent, however they do excite the whole structure including the cables which then radiates. The radiated fields are then highly dependent on the cable layout. Some specifications call for the cable to be placed to ensure maximum radiation. As the radiation efficiency of any particular layout is highly frequency dependent, this procedure needs to be repeated for each frequency examined.

The quality of an OFTS is assessed by measuring its site attenuation. This is a measure of the signal transmission between two resonant $\lambda/2$ antennas placed on the site at the antenna and EUT positions as shown in Fig2.5. The maximum value of the measured signal as one of the antennas is scanned vertically from 1m to 4m is compared with that received when the two connecting cables are joined together. This is the attenuation introduced by transmission over the site. The site attenuation can be computed for balanced dipole antennas over an infinite perfectly conducting groundplane by defining the self and mutual impedances of the two antennas and their images. The resultant 4x4 impedance matrix is inverted to compute the response. For a site to be acceptable, the measured site attenuation must lie within a $\pm$3dB envelope of the computed figure. The theoretical site attenuation is shown in Fig 2.6. The points near the curves are measured values of the NIST site at Boulder, Colarado [3]. This site has a 60m square groundplane, and thus represents the best likely attainable performance. The $\pm$3dB figure is a recognition that most sites are constructed in unfavourable locations with nearby buildings and other reflecting objects. It defines the precision of the measurements made on the site and the repeatability of measurements made on different sites. Because of this factor many test houses only issue unrestricted clearance to EUTs with emissions more than 6dB below the specification limit.

Screened enclosures are used for radiated emission measurements in circumstances when the presence of the ambient radiation is undesirable. Screened rooms are used by the military EMC community over a frequency range from below 10kHz to above 18GHz [4,5] for all radiated emission measurements. The commercial EMC community uses them for diagnostic measurements and in some cases screened enclosures can be made to simulate an OFTS.

Enclosure sizes vary in the main between 3m x 3m x 2m and 20m x 20m x 20m with some exceptions on either side of these limits. The larger enclosures, capable of housing full size vehicles or aircraft, are found in relatively few locations usually associated with national facilities or large manufacturers. The majority of enclosures have floor areas less than 10m x 10m.

The operation of the enclosure as a measurement environment is determined by the size of the enclosure relative to the wavelength of the radiated emission to be measured. At microwave and UHF frequencies, the enclosures can be made anechoic by covering the walls, ceiling and possibly the floor with radio absorbent material (RAM). The use of RAM in anechoic chambers used for antenna measurements is well known. The typical reflectivity level of -40dB associated with RAM in this application is not required for EMC measurements where the acceptable precision is lower. Good performance can be obtained with a RAM reflectivity level no better than -20dB. As a result the depth of RAM required is about $\lambda/4$ and this sets the lowest attainable frequency of anechoic operation for a given enclosure size. For most small chambers the RAM depth is limited to about 0.5m giving a lowest anechoic frequency of around 150MHz. The larger chambers have RAM depths up to 2m giving claimed anechoic performance down to 30MHz. The eventual limitation on RAM depth is mechanical support and the economics of enclosure design. In order to use the low frequency RAM a large enclosure is required with consequent increased wall area. Medium size chambers have been built with RAM on the walls and ceiling leaving a conducting floor, and these are able to be used as OFTSs above about 100MHz [6].

The frequency range below that at which an enclosure can be made anechoic can be divided into three regions dependent on the principal means of energy propagation within the room [7]. The lowest frequency range, up to a frequency at which the maximum linear dimension of the enclosure is approximately $\lambda/10$, can be treated as a quasi-static problem. No appreciable phase shift occurs in wave propagation across the enclosure, and the coupling of equipment to an antenna can be analysed in terms of mutual capacitive and inductive elements in an equivalent circuit. Such a circuit is shown in Fig 2.8 and is equivalent to the EUT and antenna layout inside a screened enclosure shown in Fig 2.7. In the capacitive coupling case the antenna is a 41" active rod antenna as specified in Def Stan 59-41 and Mil Std 462 for the measurement of radiated electric fields. The surface of the enclosure and the conducting bench that the EUT is placed on is an equi-potential surface.

The next frequency range is that up to 30MHz, the upper frequency of use of the 41" rod antenna. The conducting groundplane and its extension can be regarded as the inner conductor of a coaxial transmission line that supports a Transverse Electro-Magnetic (TEM) wave propagating from the EUT to the antenna. If the EUT is regarded as a set of electric and magnetic dipoles, then these couple into the TEM mode through their mutual capacitance and inductance with the transmission line. The bench bonded to the screened room wall is a short circuit, and the active rod antenna is a capacitive load at the end of the line. The change in width between the bench and the groundplane extension is a change in line impedance with its associated discontinuity capacitance representing the stored energy associated with the field perturbation at the discontinuity. Equivalent circuits for the two coupling modes are shown in Fig 2.9 and the associated frequency responses for a typical screened enclosure are shown in Fig 2.10. The resonance at 35MHz is associated with $\lambda/4$ resonance of the transmission line system, and the coupling null at 17MHz is due to mode competition between the direct capacitive coupling from the EUT to the antenna dominant at low frequencies and the TEM coupling dominant at higher frequencies.

In the frequency range from 30MHz to 200MHz a different enclosure set-up is required as shown in Fig 2.11. A biconical dipole antenna is used to measure the fields. If the enclosure is not anechoic in this frequency range, cavity resonances will be present. The frequencies of these resonances for an empty rectangular enclosure of dimensions **a** x **b** x **c** is given by;

$$f_{pqr} = 150((p/a)^2 + (q/b)^2 + (r/c)^2)^{1/2} \text{ MHz}$$

where p,q,r are integers, one of which may be zero. In addition, TEM resonances of the conducting bench are present. The bench perturbs some of the cavity resonant frequencies, and the EUT and antenna also modify the resonant frequencies. The frequency response of a small enclosure of dimensions 2.25m x 2.25m x 4.5m with a conducting bench and biconical antenna is shown in Fig 2.12. The excitation source is a short dipole placed at the EUT position on the bench.

Consideration of the low frequency operation of screened enclosures for radiated emission measurements shows that the measurements are not directly equivalent to measurements performed on an OFTS. The energy propagation mechanisms present in the enclosure are different to electromagnetic radiation. The various resonance phenomena can impose measurement uncertainties of up to $\pm$40dB unless steps are taken to remedy their effects. A number of studies are underway in the UK and elsewhere aimed at evaluating ways of reducing the resonance effects using various forms of dissipative dielectric and magnetic materials positioned at field maxima inside the enclosures [8]. The eventual aim is to reduce the resonance induced coupling variations to $\pm$5dB. While it may seem that the use of an enclosure for radiated emission measurements is prone to considerable error, it should be remembered that the oscillating charges and currents that cause radiation in free space are those that excite the various coupling mechanisms in a screened enclosure. The enclosure measurement, in principle could be used to evaluate free space radiation. The assumption here is that the presence of the enclosure does not interfere with the oscillating sources. Recent experimental work has shown this to be the case for magnetic dipole sources but not for electric dipole sources [9]

2.2 Conducted Emission Measurements.

The measurement technique used for making conducted emission measurements is determined by the type of current flow on the conductors. Both common mode and differential mode emissions require assessment. Common mode currents flowing on a conductor set such as a power cable are an important source of radiated emissions, and in some circumstances, measurements of common mode currents can replace radiated emission measurements.

Common mode currents are measured using some form of current transformer. In its simplest form it comprises a toroidal core of high permeability material, through which the cable carrying the common mode current is passed. The time varying common mode current induces a time varying magnetic flux in the toroid. A sense coil is wound around the toroid, and the induced voltage, a measure of the changing flux, is measured by the receiver. The system is shown in Fig 2.13. The frequency response of the system is shown in Fig 2.14. Above the frequency where the reactance of the sense coil is greater than the input resistance of the receiver, the system is frequency independent. It is characterised by a transfer impedance, measured in dBΩ, which is the relationship between the received voltage and the source common mode current.

Current probes measure the total common mode current on a cable. They cannot distinguish between currents on an individual conductor within a cable, unless that conductor is passed through the probe in isolation. Care must be taken to ensure that the current probe toroid is not saturated by large direct or power frequency currents as the change in differential relative permeability reduces the transfer impedance. Current probes can be obtained for frequency ranges up to 1GHz. The normal design value of transfer impedance is 1Ω (0dBΩ) thus eliminating the need for numerical conversion between measured voltage and common mode current. As a current probe is essentially a transformer in which the cable to be sensed forms a single turn primary winding, the effect of the current probe is to insert a small series impedance into the cable. The value of this impedance depends on the receiver input impedance, the turns ratio and the other parasitic effects associated with a transformer. In general the inserted impedance is less than 1Ω. In a multi-conductor cable, the distribution of the series impedance is not easy to determine as it depends on the termination impedances of each conductor and the cross-talk between them.

As, at higher frequencies, the common mode current on cables is a primary source of radiated emissions, a common mode current measurement can replace a radiated emission measurement. This is an attractive option in view of the problems associated with making reliable and repeatable radiated emission measurements. The main problem with a current probe measurement at higher frequencies arises from standing waves set up on the cable. These can be largely eliminated if the common mode current on the cable is terminated in an absorbtive load. This can be accomplished using an absorbing clamp as shown in Fig 2.15. This comprises a conventional current probe immediately followed by a series of ferrite rings made from a ferrite material which exhibits loss at radio frequencies. The cable to be assessed is placed on a bench and the clamp is moved along the cable. The maximum value of the residual current standing wave is measured.

Differential mode currents are measured using a network inserted into the cable known as a Line Impedance Stabilisation Network (LISN). This is a simple circuit which manages to fulfil three requirements.

i) It defines the impedance seen by the differential mode currents on the cable. This is necessary in order to perform a repeatable measurement. The r.f. impedance of a domestic 240V supply varies between 0Ω and 200Ω resistive and ±200Ω reactive. A suitable standard impedance in this range is 50Ω, the input impedance of most measurement receivers.

ii) It reduces differential mode interference propagating away from the EUT thus isolating the EUT.

iii) It reduces other r.f. currents on the cable reaching the EUT thus reducing interference with the measurement.

A circuit diagram of a LISN is shown in Fig 2.16.
Its performance in terms of the impedance seen by the EUT is shown in Fig 2.17. As the differential mode currents are terminated by a standard impedance they are assessed in terms of r.f. voltage at the output of the LISN. The switched resistors ensure that each line is terminated in the same impedance irrespective of which line the receiver is switched to. The layout of a conducted interference measurement according to Def Stan 59-41 is shown in Fig 2.18.

2.3. Measurement Receivers.

In this section the instrumentation used for emission measurement is considered. As the interference waveforms are indeterminate and, for the most part, we are seeking to protect radio services, emission levels are defined in the frequency domain. The interference is then measured on a receiver of defined bandwidth and detector function. The receivers used all use the superheterodyne architecture. They are classified into two types;

i) Spectrum analysers. These are relatively cheap and provide a quick visual display of their input spectrum. Most spectrum analysers operating in the frequency range below their first IF frequency (about 2GHz) have only an input low-pass filter, with no tuned pre-selection before the first mixer. This means that all the incoming energy is presented to the mixer. If the interference is impulsive, then, while the mean energy may be low, the instantaneous peak energy can be sufficient to overload and possibly damage the mixer device. Spurious signals can thus be generated within the analyser, and unless steps are taken to identify them the results of emission measurements from spectrum analysers are suspect.

ii) Measurement Receivers. Measurement receivers have tuned pre-selector filters before the first active stage and thus limit the input energy to that stage. This may take the form of a tuned filter, a switched filter bank or a combination of both. They do not suffer to the same extent as spectrum analysers from spurious responses and they have a wider dynamic range. The requirement to tune or switch the pre-selector filter means that the tuning rate of a receiver is less than that of a spectrum analyser, and traditionally receivers have been manually tuned with a meter indication of the received signal strength. Most receivers also have audio de-modulation facilities and a speaker which are not generally found on a spectrum analyser.

Modern design has resulted in a synthesis of the best of both receiver types by use of computer control, data storage and display. Receivers now have VDU display of the spectrum and spectrum analysers have pre-selector options giving them receiver-like performance. The add-on pre-selectors cost more than the basic spectrum analyser!

The main conceptual difficulties with receivers lie in the selection of measurement bandwidth and detector type. Receiver bandwidths are selectable with values in the range 100Hz to 1MHz. The signals to be measured can be classified into two types, broadband and narrow band. A broadband signal is one with a spectrum extending beyond any filter bandwidth available on the receiver. The amount of energy incident on the receiver detector varies with the measurement bandwidth and thus any measurement must have a defined bandwidth. Random noise and impulsive signals are generally broadband.

A narrowband signal is one with a spectrum narrower than the receiver bandwidth. Thus most analogue radio signals and individual clock harmonics are narrowband signals. The energy incident on the detector is independent of the receiver bandwidth.

Clearly the definitions lead to classification difficulties. Strictly speaking, the broadband definition should be sub-divided into continuous spectrum noise like signals and line spectrum repetitive or impulsive signals. This is important when considering the detector input energy to bandwidth relationship. For a continuous spectrum noise like signal, the detector input energy is proportional to the bandwidth and hence a bandwidth increase of a factor of ten results in a detector input power increase of 10dB. In the case of an impulsive signal with a series of phase related spectral lines, the bandwidth increase of ten results in an increase of ten in the number of spectral lines incident on the detector. As the spectral lines are phase related, the detector input voltage is increased by a factor of ten (20dB) and hence the detector input energy increases by 20dB.

Two detector types are in common use for EMC measurements, the peak detector and the quasi-peak detector. The peak detector is more common in military usage whereas the quasi-peak detector is used by the commercial EMC community. Both detector types comprise a conventional diode detector followed by a charge -discharge circuit with defined charge and discharge times. The peak detector has a very short charge time and a relatively long discharge time. It thus measures the peak voltage of any input signal. For sinusoidal and other similar narrow band signals it can be scaled to read the r.m.s signal level.

The quasi-peak detector is designed to have an output which responds both to the amplitude of the input signal and to the repetition rate of impulsive signals. Thus an impulsive signal with a higher repetition rate but the same amplitude than another signal will give a greater output. The design of the quasi-peak detector is configured in this way to indicate the subjective annoyance value of repetitive signals on audio and video signals. It is not adequate however to indicate bit error rates in digital systems. As the output of the quasi-peak detector is much less than the peak input signal, the associated r.f. mixer and IF stages must be capable of reproducing the full amplitude of the impulsive signal. Thus a considerable signal overload factor relative to the indicated output level must be built into these stages. The quasi-peak detector functions by having detector charge and discharge times related to the IF bandwidth. The IF bandwidth is related to the receiver input frequency as shown in the table below.

Table 2.1. Quasi-Peak Detector Characteristics.

Input Frequency Range.

	10-150kHz	0.15-30MHz
peak bandwidth	100-500Hz	7-10kHz
quasi-peak bandwidth	200Hz	9kHz
charge time	45ms	1ms
discharge time	500ms	160ms
IF overload factor	24dB	30dB

Input Frequency Range.

	30-300MHz	0.3-1GHz
peak bandwidth	100-300kHz	100-500kHz
quasi-peak bandwidth	120kHz	120kHz
charge time	1ms	1ms
discharge time	550ms	550ms
IF overload factor	43.5dB	43.5dB

The relative responses of peak and quasi-peak detectors are shown as a function of pulse repetition rate in Fig 2.19. Calibration of a quasi-peak detector is made using a standard r.f signal generator generating a sinusoidal signal. Under these circumstances the two detector types give the same indication.

2.4 Radiated Immunity Measurements.

In this section the basic techniques of performing radiated immunity measurements are discused.

Radiated immunity assessments require *that the EUT be illuminated by a propagating* electromagnetic wave of the required frequency and field strength. As the frequency is stepped or swept over a substantial frequency range and field strengths of up to 200V/m are required it is normal to perform these assessments in shielded enclosures. Once again, the mode of operation of the enclosure is determined by the frequency range of the assessment. At frequencies where the enclosure can be made anechoic, the EUT can be illuminated by a wave generated by a conventional antenna as in free space. The r.f. power required P_t can be evaluated by considering the required threat field E and the gain G available from the transmitting antenna. If the intrinsic impedance of free space is Z, and the EUT to antenna distance is R then,

$$E^2/Z = Pt \cdot G / 4\pi \cdot R^2$$

Examination of this equation indicates a possible advantage in using high gain antennas. This is not the case as two further factors need consideration. The first is that the antenna only exhibits its full gain at distances beyond the Rayleigh range, a diffraction limited effect. If the antenna maximum dimension is D and the wavelength is λ then

$$R > 2D^2/\lambda$$

is required to realise G. As the gain of an antenna of radiating aperture say D^2 is

$$G = 4\pi \cdot D^2/\lambda^2$$

it can be seen that the greater the gain G the greater the distance R needs to be. The second effect is the relationship between the 3dB beamwidth of the antenna θ and hence the area of the EUT illuminated by the antenna and the antenna gain. The approximate relationship is,

$$G = 4\pi/\theta^2$$

The higher the gain of the antenna, the narrower the beamwidth, and hence for a given size of EUT, the further away the antenna needs to be to ensure that the whole EUT is illuminated. It is left as an exercise for the reader to show that the transmitter power required to illuminate a given size of EUT is independent of the antenna gain if the above illumination criterion is followed.

There is considerable advantage therefore in minimising the antenna to EUT distance, and hence screened enclosure size, by using a low gain antenna, provided that the Rayliegh range criterion and the minimum distance requirement for a 377Ω wave impedance ($R > \lambda/2\pi$) are satisfied.

At lower frequencies, where the enclosures cannot be made anechoic, cavity resonances can cause considerable measurement uncertainty. Problems also arise at even lower frequencies because the 377Ω wave impedance criterion results in an unacceptable EUT to antenna distance. Various techniques are used to overcome the problems including at the lower frequencies the use of capacitor and inductor like field generators that produce localised high impedance (electric field) or low impedance (magnetic field) waves.

A better technique for producing plane electromagnetic waves at low frequencies is to use a Transverse ElectroMagnetic (TEM) wave generator such as the Crawford Cell. Such a device is shown in cross-section in Fig 2.20. The cell comprises a tapered transmission line of square cross-section with the inner conductor in the form of a flat septum. The tapered cross-section maintains a constant characteristic impedance on the transmission line from the coaxial input to the working volume and on to the matched load. The cell geometry is chosen to give a 50Ω characteristic impedance for the transmission line (voltage to current ratio for a wave propagating on the line) while the air dielectric ensures an electric to magnetic field ratio of 377Ω. Within the working volume, the fields mimic those of a plane wave. The Crawford cell operates from d.c. to an upper frequency set by the onset of higher order modes than the wanted TEM mode. These occur at a frequency when the cross-section of the cell becomes $\lambda/2$, but their onset can be delayed by careful use of absorbing material in the cell. The maximum size of EUT is limited to one occupying one third of the septum to wall dimension if acceptable field perturbation is to be maintained. Thus the maximum frequency of operation of a Crawford cell is set by the size of the EUT, the larger the EUT the lower the maximum frequency.

The electric field E inside the cell can be calculated from the voltage V applied to the cell and the septum to side spacing d.

$$E = V/d \text{ Volts/m}$$

and the magnetic field H is

$$H = E/377 = V/377.d \text{ A/m}$$

For a 50Ω characteristic impedance cell, the power P_c required for an electric field E is

$$P_c = (E.d)^2/50 \text{ Watts}$$

Other types of TEM cell are used including parallel plate transmission lines. The principles of operation are identical. In the case of the parallel plate lines, the fields are not confined as in a Crawford cell and the surroundings can cause considerable field perturbation.

2.5 Conducted Immunity Measurements.

The techniques used for conducted immunity assessments are based on the same current and voltage transducers used in conducted emission assessments. Detail changes in design to accommodate the greater power requirements of immunity assessment may be made. Again, two operation modes can be identified, common mode and differential mode. In the case of immunity measurements we are not restricted only to the frequency domain, and threat currents or voltages in the form of impulses can be used.

One aspect of conducted immunity assessment is now starting to replace radiated immunity assessment on large systems such as aircraft. The power and space requirements of radiated immunity assessment of a large system can be prohibitive. As a rule it is recognised that the energy propagation path into susceptible systems is via currents induced onto cable bundles by the incident wave. The first part of this energy transfer process can be avoided if interfering currents are directly injected onto the appropriate bundles. This bulk current injection technique is useful if the relationship between the incident threat field and the induced current is known. The relationship can sometimes be deduced by numerical modelling or by illuminating the large system with a low incident field and measuring the induced currents. An assumption of linearity can then be made to deduce the induced currents from a full threat field. Neither of these techniques is completely satisfactory, as incorporation of the non-linear behaviour of the susceptible systems is difficult in both cases. Fig 2.21 shows a test system for bulk current injection.

3.0 Conclusions and Future Directions.

In this paper I have attempted to give an overview of some of the current measurement and test techniques used for EMC assessment. The treatment is necessarily incomplete, and no mention has been made of ElectroStatic Discharge measurements. New measurement techniques are under development in several areas. Much work is underway on screened enclosure techniques at low frequencies below the anechoic limit frequency. The large uncertainties caused to both immunity and emission measurements are not acceptable and resonance damping techniques and enclosure/antenna system calibration techniques are under development. These are at relatively early stages but may be incorporated into revised standard in the future.

The problems of EMC assessment of large systems are also attracting attention, particularly in the commercial EMC community where large distributed computer systems within buildings and electronic telephone exchanges represent current problems. The move to low energy low voltage logic families and poorly screened plastic equipment enclosures enhances problems, and the screening performance of whole buildings is an important consideration when considering these distributed systems. The advent of optical data transmission has not proven to be the solution of EMC problems as some originally thought. While considerable amounts of data can be transmitted this way, power still goes along copper, as does interference. The systems at the end of the fibres are still electronic.

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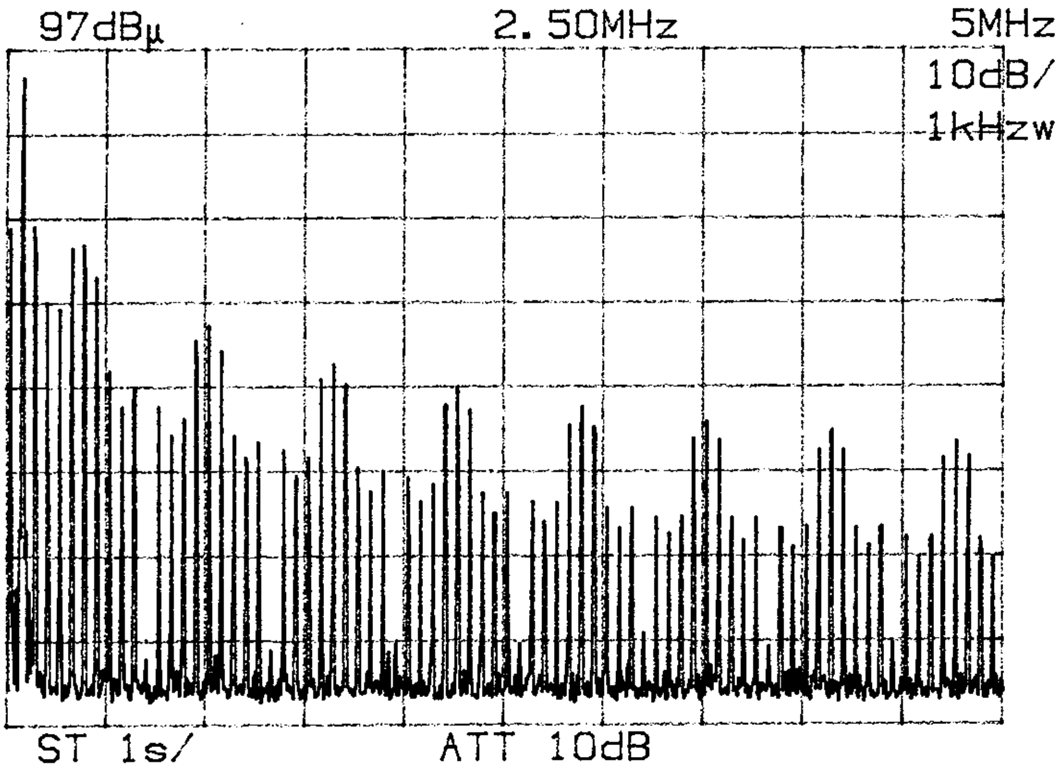


Fig 1.1. Spectrum of Harmonics from a Digital Circuit.

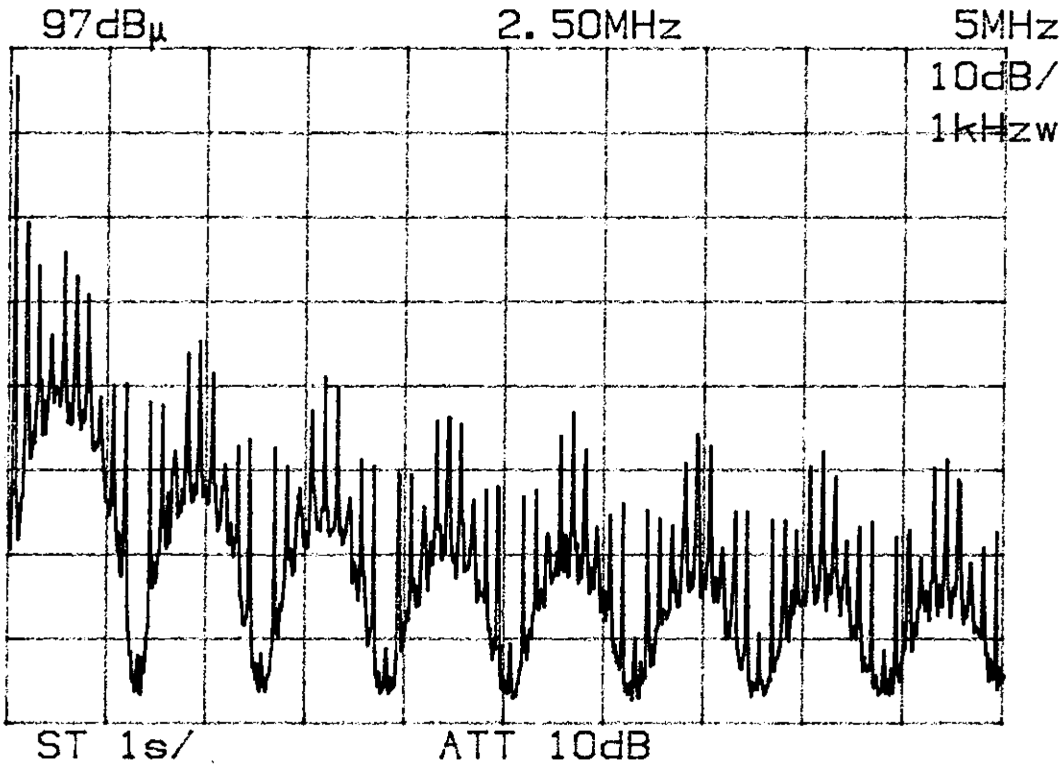


Fig 1.2. Continuous Spectrum of Digital Data.

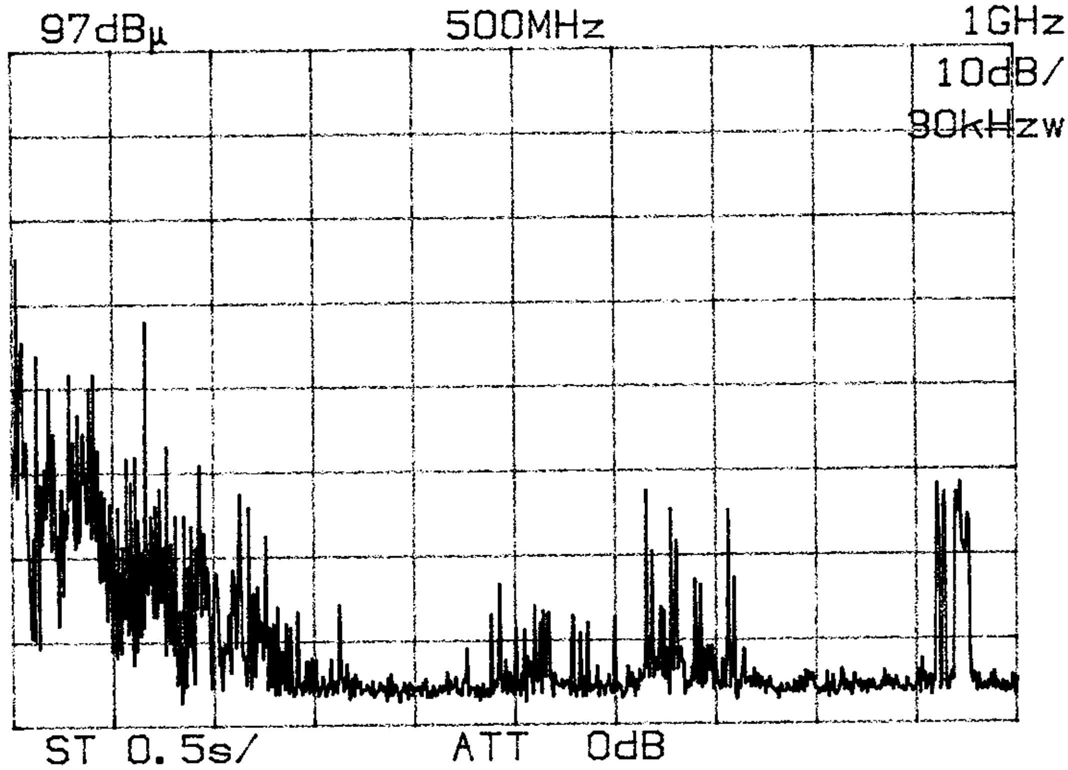


Fig 1.3. Ambient Radio Spectrum to 1GHz at York, U.K.

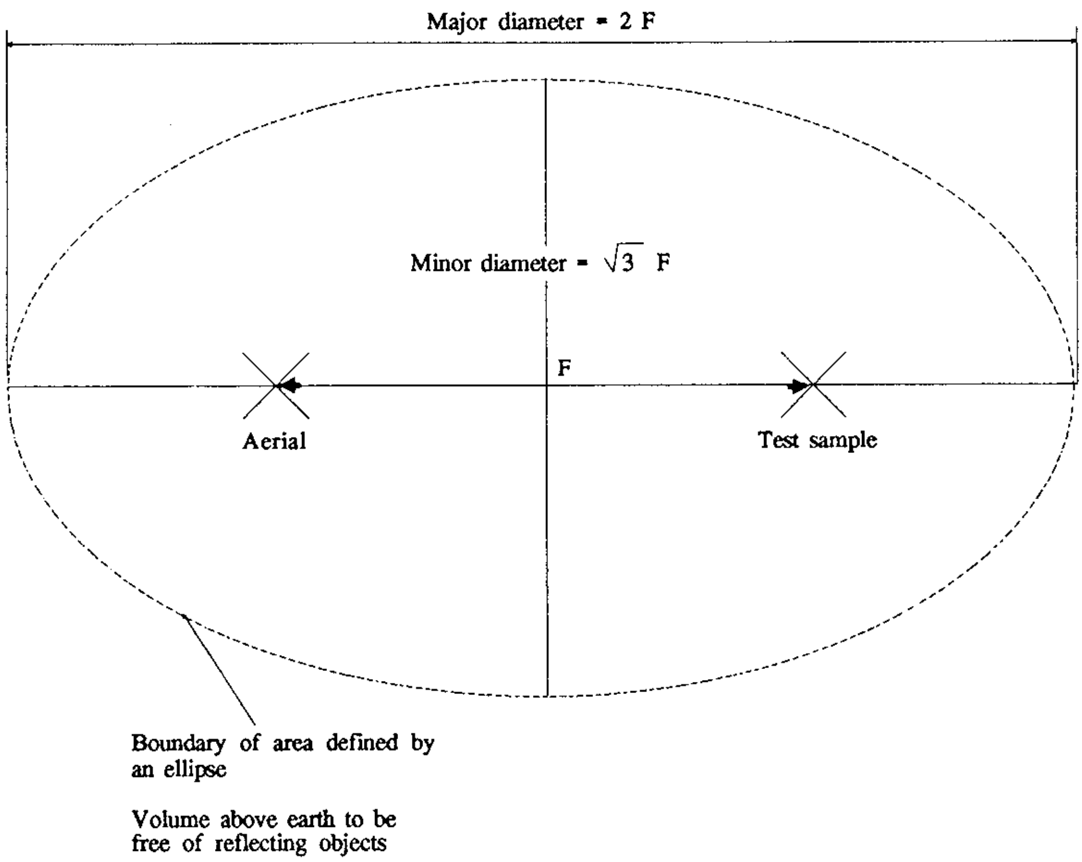
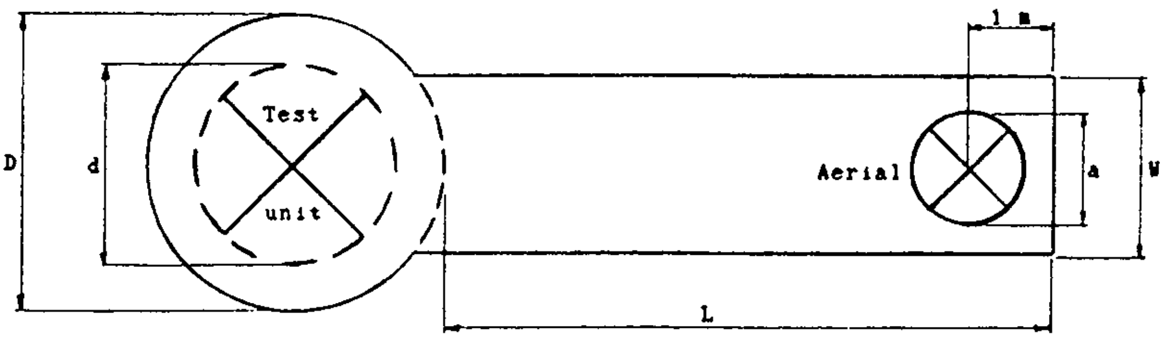


Fig 2.1. Plan View of an OFTS.



D = d + 2 m, where d is the maximum test unit dimension;
W = a + 1 m, where a is the maximum aerial dimension;
L = 3, 10, or 30 m.

Fig 2.2. OFTS Groundplane Minimum Area.

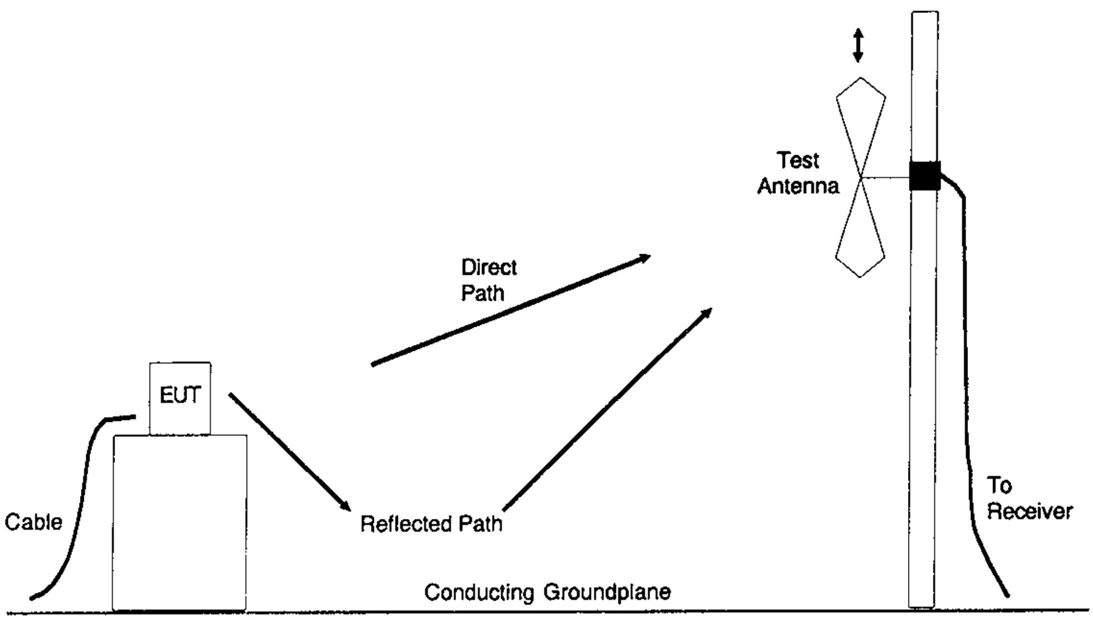
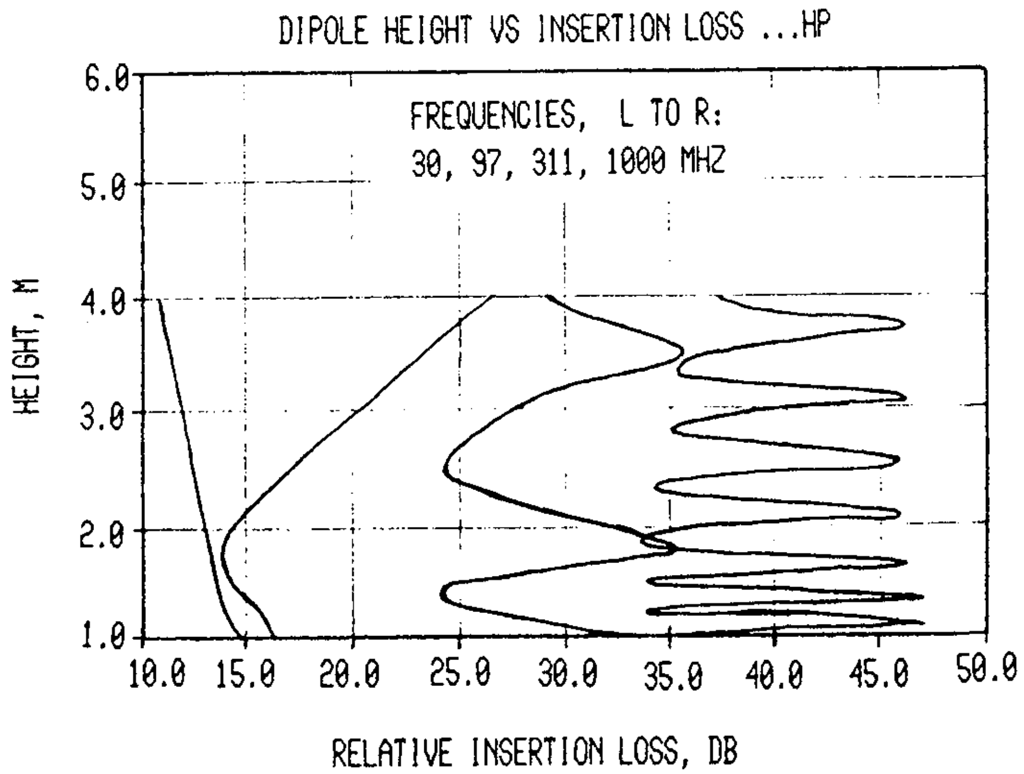


Fig 2.3. Side View of an OFTS.



Selected relative insertion loss data calculated for horizontally polarized, dipoles over perfect ground. Site attenuation is the single minimum insertion loss value of each curve. Separation distance is 3 m; transmitting dipole height is 2 m.

Fig 2.4. Vertical Interference Patterns on an OFTS.

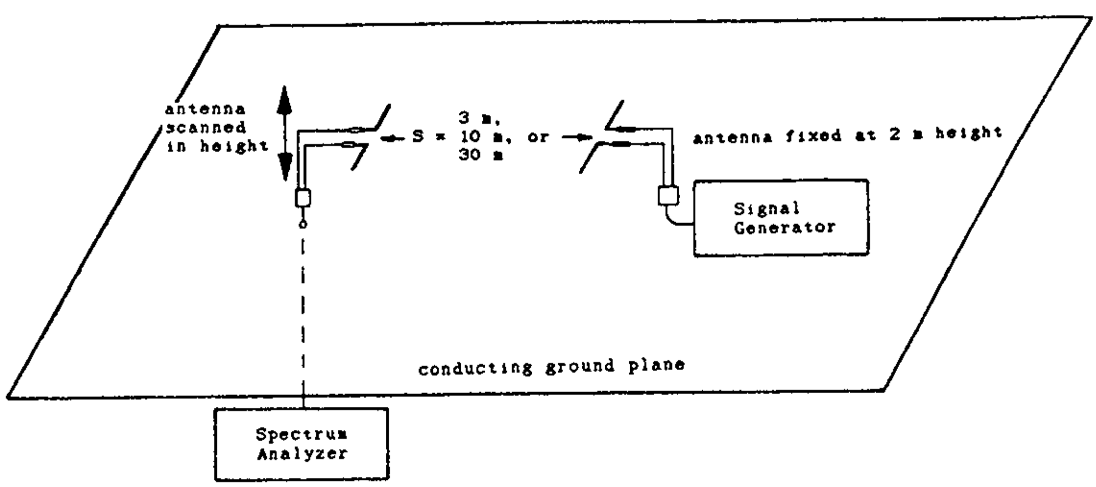
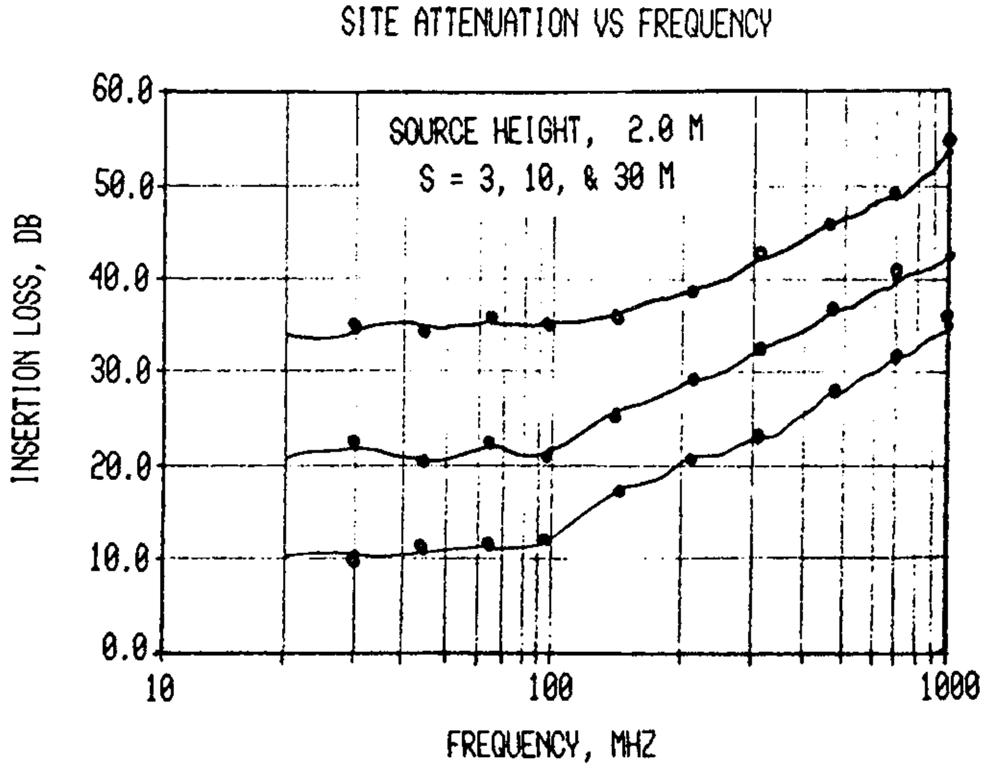


Fig 2.5. Antenna Positions for OFTS Site Attenuation.



attenuation for horizontally polarised dipoles over perfect ground. Solid curves are calculated, points are measured data

Fig 2.6. Theoretical and Measured Site Attenuation.

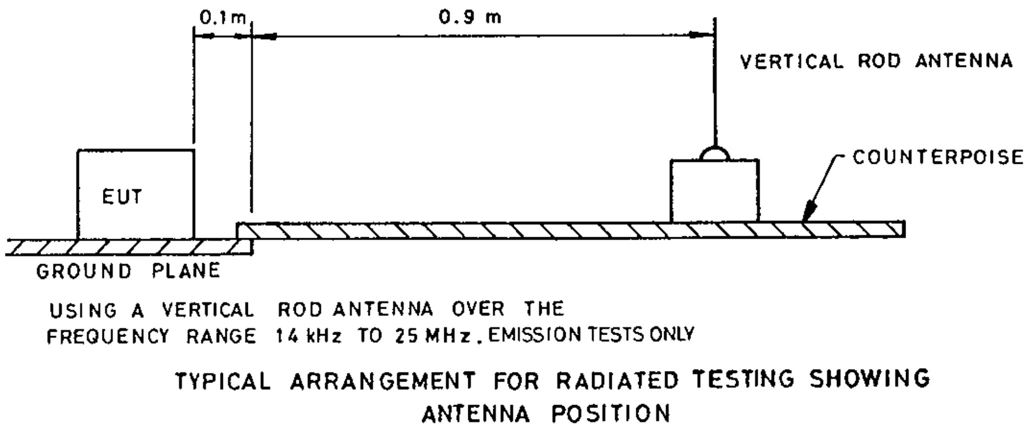


Fig 2.7. EUT Position in a Screened Enclosure.

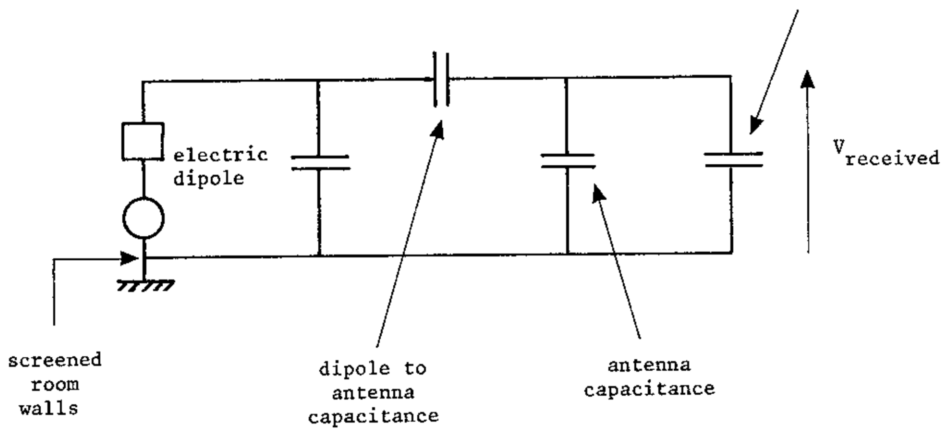
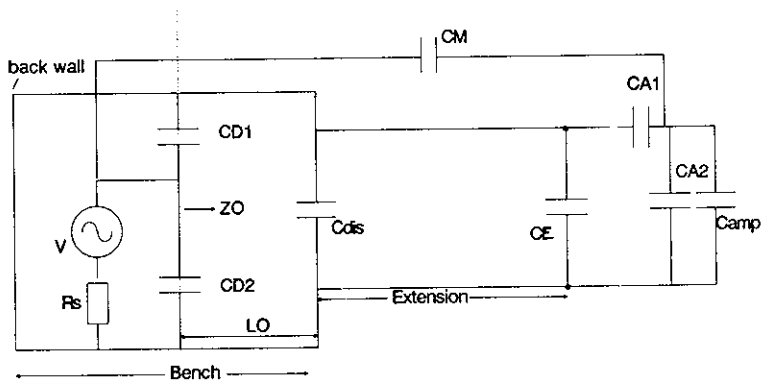
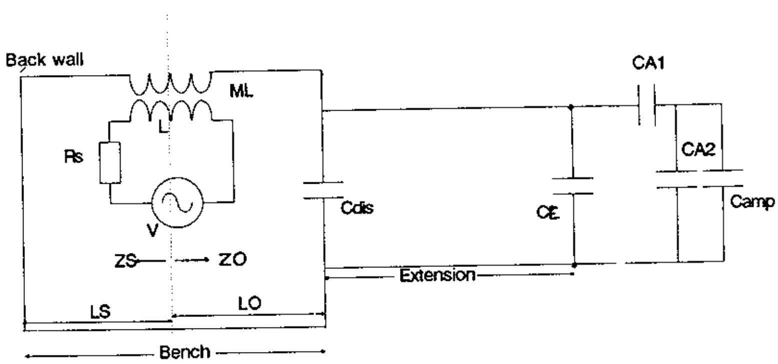


Fig 2.8. Equivalent Circuit for EUT to Antenna Capacitive Coupling.

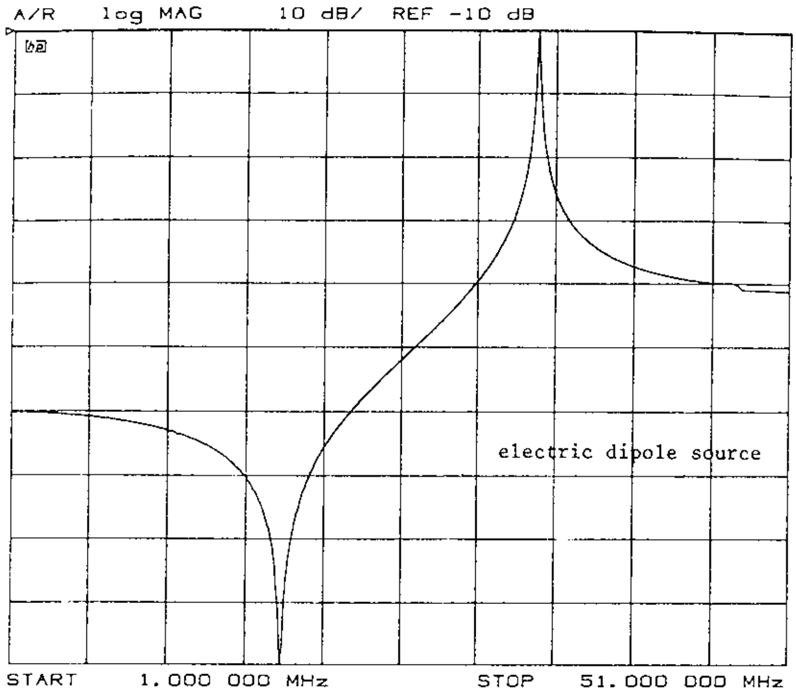


a)electric dipole source



b)magnetic dipole source

Fig 2.9. Equivalent Circuits for TEM Mode EUT to Antenna Coupling.



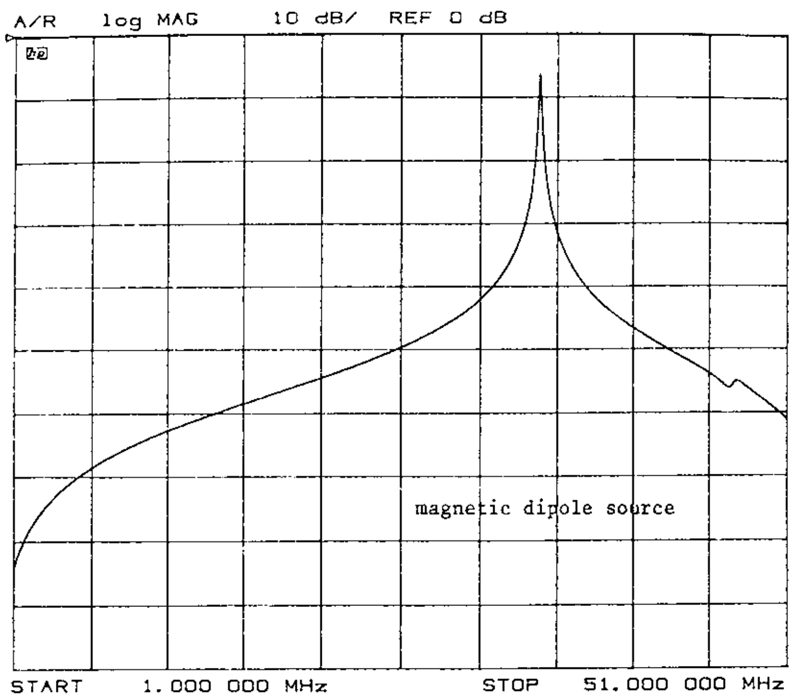


Fig 2.10. Frequency Responses for Circuits of Fig 2.9.

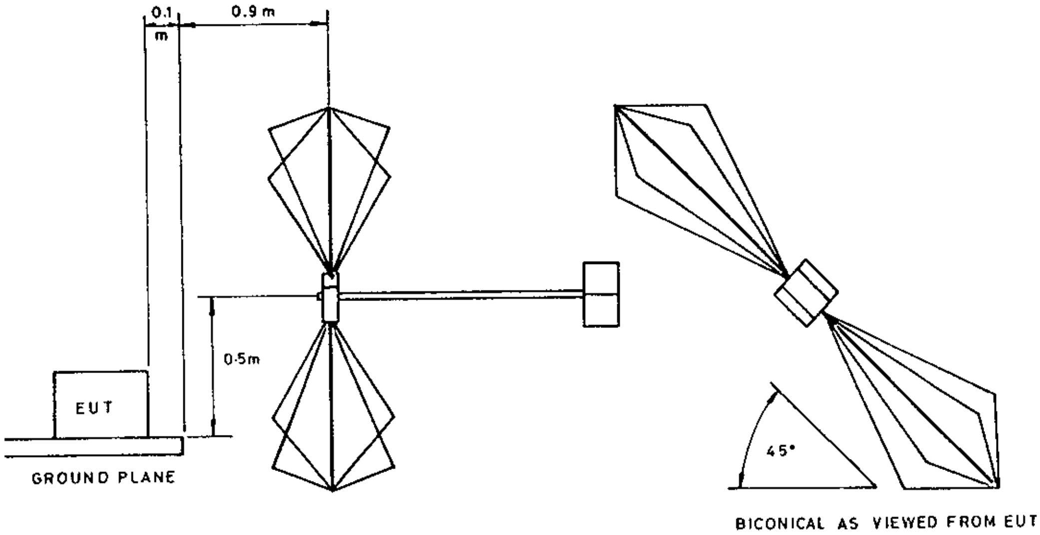


Fig 2.11. EUT and Antenna for 30 - 200MHz.

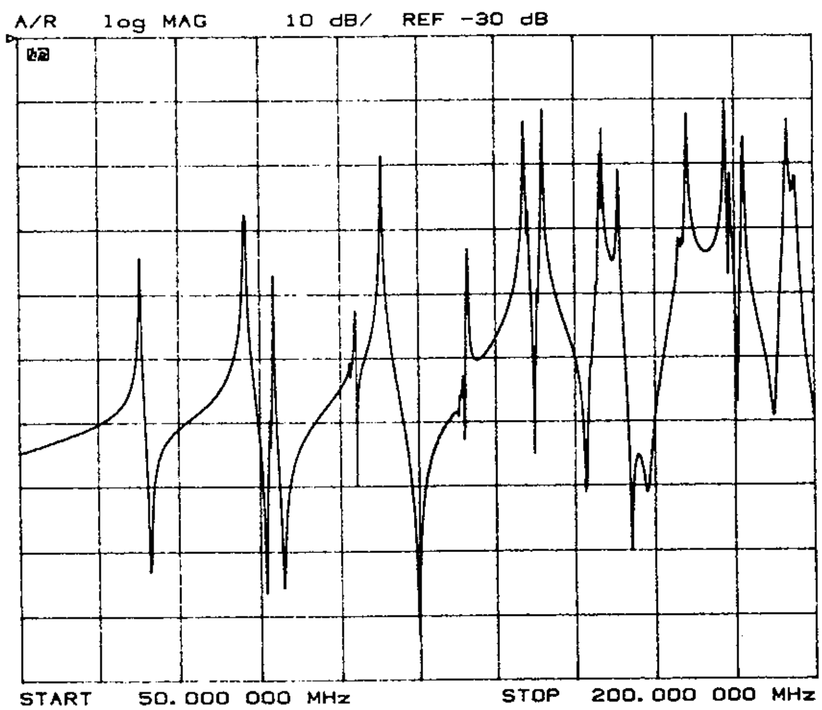


Fig 2.12. Frequency Response of Screened Enclosure 30 - 200MHz.

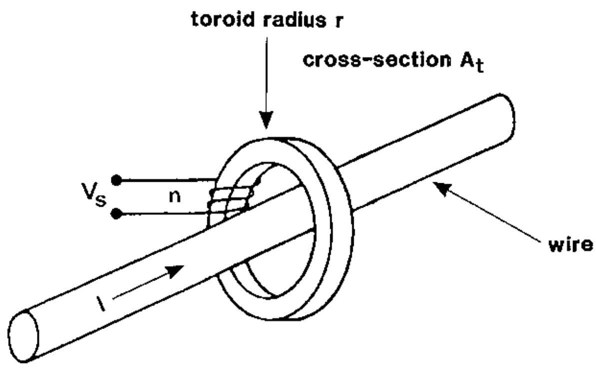


Fig 2.13. Current Transformer.

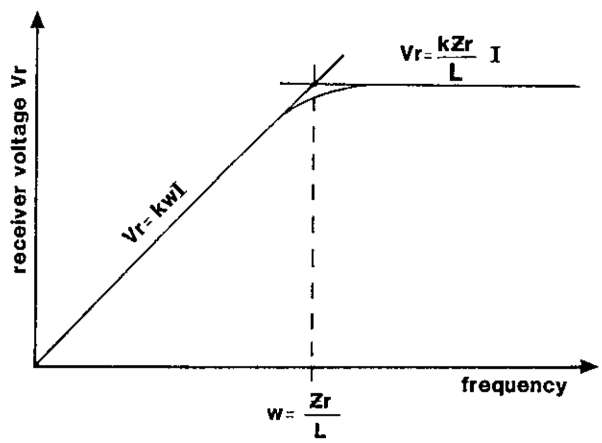


Fig 2.14. Frequency Response of Current Transformer.

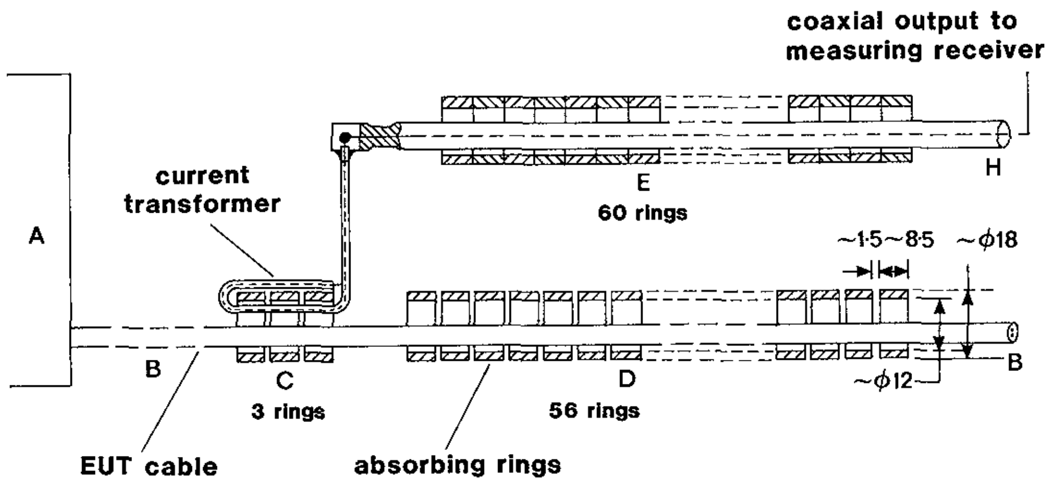


Fig 2.15. Absorbing Clamp Construction.

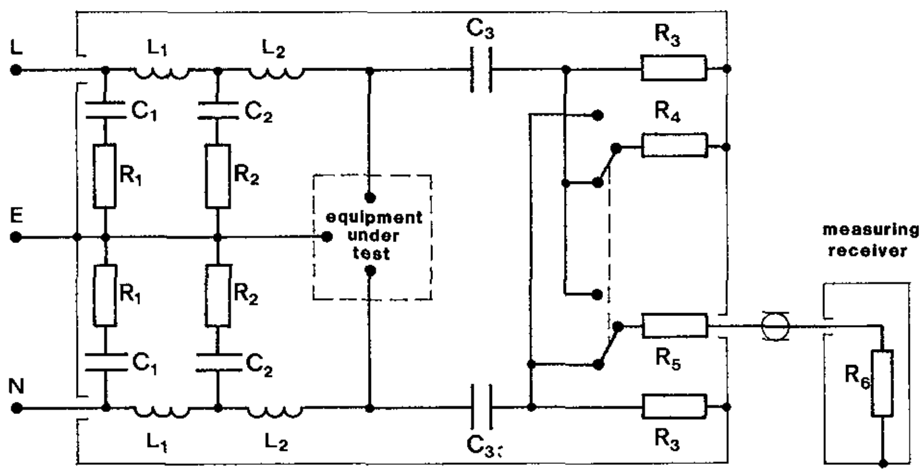


Fig 2.16. LISN Circuit Diagram.

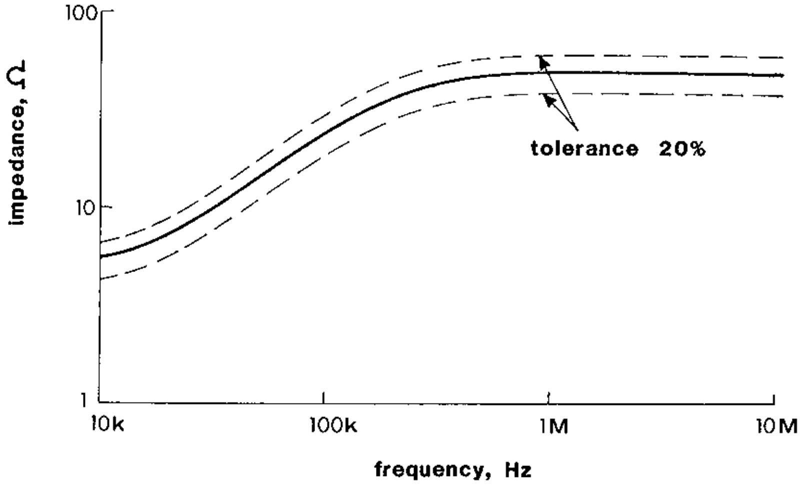


Fig 2.17. LISN Impedance Performance.

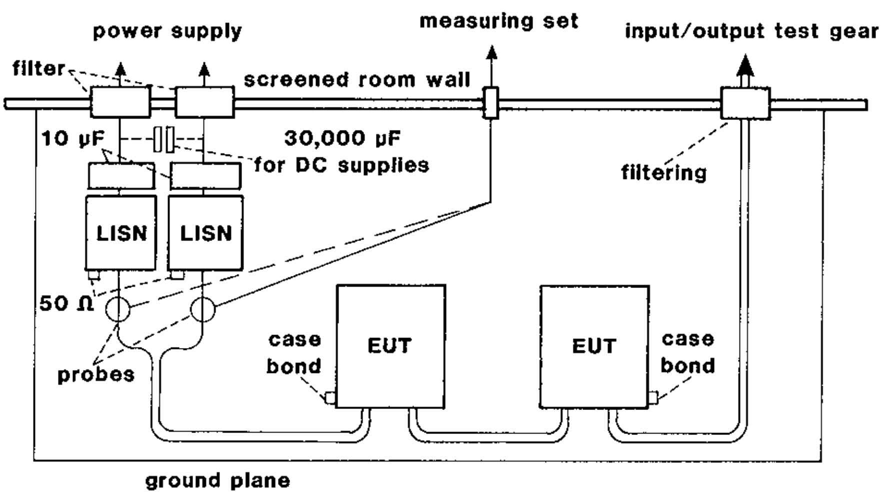


Fig 2.18. Conducted Interference Measurement Layout.

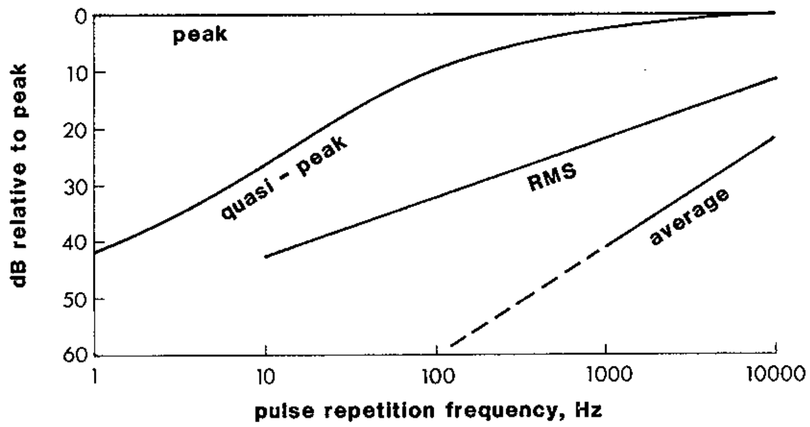


Fig 2.19. Detector Responses.

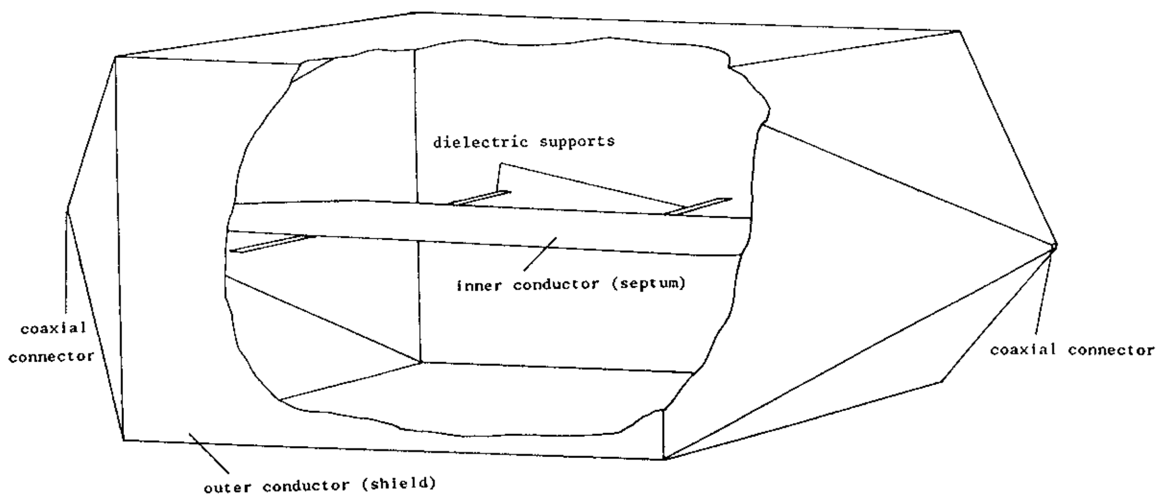


Fig 2.20. Cross-Section of a Crawford Cell.

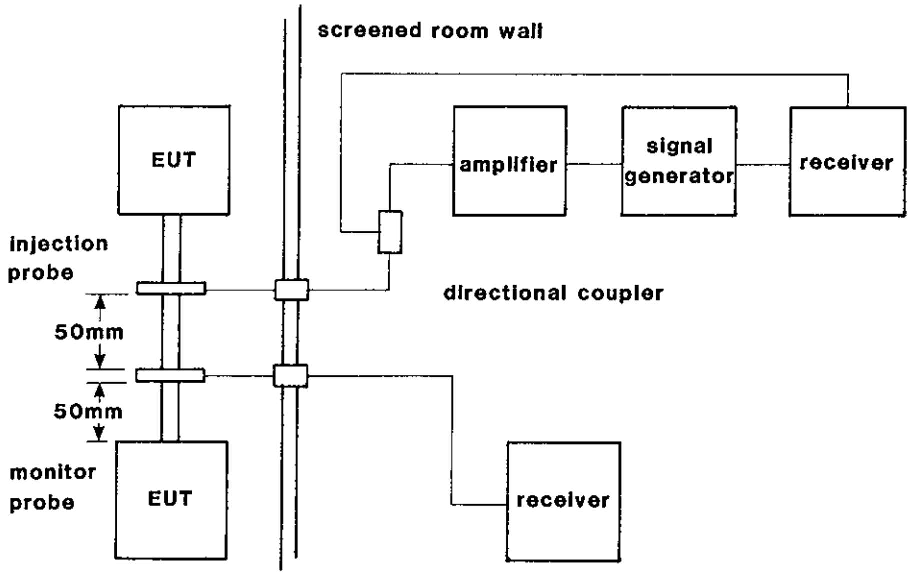


Fig 2.21 Bulk Current Injection Test System.

Design Principles for effective EMC.

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Effective EMC design is directly related on minimising the different effects in EMC : EME, EMS and the coupling/transmission paths.
A short overview will be given of the most important factors concerned with EMC design. It follows design parameters and design rules for each item.

0. Introduction

Effective EMC design is directly related on to minimising the effects of some important factors concerned with EMC : *emission*, *susceptibility*, coupling paths, propagation paths... . At this stage, two remarks must be made :

- a degradation in the good working of a system or circuit is always a cumulative effect of noise and interfering signals (unwanted signals). Without any interference, a normal design is made for an acceptable Signal/Noise value (S/N), with N = the inherent system noise (bandwidth-limited white noise) for analogue systems/circuits and with N = the noise immunity level or the noise margin for logic families.
  With an interfering ambient, the interfering signals I are added to the (also) unwanted noise N.
  So the real signal/noise ratio becomes :

  $$S/N \rightarrow S/N{+}I = \frac{S/N}{1 + I/N} \cdot$$

  It is necessary that $I/N < 1$ into obtain a non-disturbed system by interference, but achieving $I/N \lll 1$ gives a costly overdesign.

- As a system, it should be understood as any combination of two or more systems, subassemblies, racks, PCB's or even chips on a PCB for high speed/high frequency circuits.

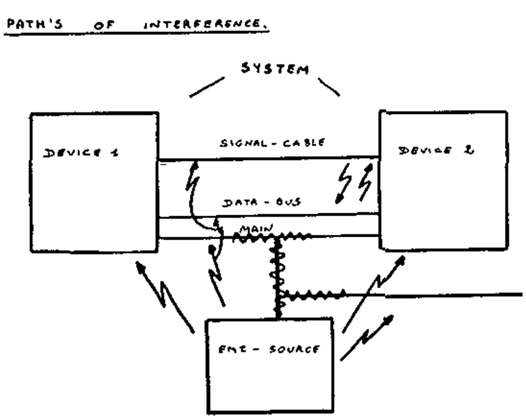


Fig. 0.1 Schematic of a system.

This means that for some topics, very general design rules may be given. Examples are about grounding, crosstalk, field-to-loop/loop-to-field coupling, ... for large systems or PCB-design. The only difference is that the larger the system is, the lower the frequencies where some effects occur. This means in practice that for a large computer system, the same rules and effects may be handled as for a high speed PCB-design.
As typical parameters, the wavelegth/Dimensions or the pulse duration/wire length ratios are the constants in the general design rules.

In this paper, some general design rules will be discussed in this manner, so that they may be applied for each appropiated design in real life.

1. General overview

It has been mentionned that the overall EMC quality of a system is a cumulative effect of a lot of items, also internally of a system or a PCB, because f.ex. at PCB-level or back-phase level, small circuits or components may also be regarded as "a system or subsystem".
An overview is summarized in the next table, and each point will be discussed in more detail:

- common impedance coupling between systems
- common mode coupling to EM-fields
- coupling of ground noise into the systems/circuits
- differential mode coupling to EM-fields
- balanced/unbalanced systems and twisted pairs
- filtering of power lines
- filtering of data/signal cables
- crosstalk on wires
- reflections on wires
- shielding
- decoupling capacitors
- surge protection
- non-linear effects and HF behaviour of components
- spectrum management
- mechanical design and lay-out

Two points should be mentionned :

1.1. Reciprocity of emission/susceptibility

The reciprocity theorem is a very general theorem in circuit theory. For EMC, it means that a loop will act with exactly in the same manner for emission and susceptibility. The same may be said for the bandwidth of a system (even a logic one). This means that a good design for susceptibility will also have a good behaviour with a low emission level. EMI-sources are always possible EMS-receivers, and vice versa.
In practice however, it is more difficult to have a high EMS-value, due to the fact that for emission control, the choice of components, circuits, PCB-lay-out, ... is fully under control of the designer, but that for immunity control, the noisy ambient is NEVER exactly known in detail. So, normally in practice, the real problem is a design for a good <u>SUSCEPTIBILITY</u> of the system. But, measurements and validation of a system are easier for EMISSION.

1.2. Source/victim/coupling path control

Control of EMC may be done at different levels of the EMC transmission path at the source level, victim level or at the level of the coupling path.

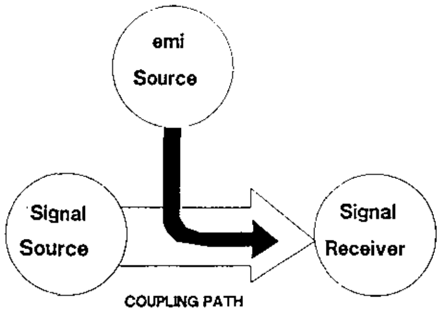


Fig. 1.1. EMC transmission path

At the source and victim level, two worlds of design are concerned with :

- the electrical/electronical one, concerning the choice of components, circuit design, technology, subassemblies, frequency, range, ...
- the mechanical one, concerning the choice of PCB-system (ex. double side or multi-layer), lay-out on PCB, mechanical construction, mechanical attachment of cables, ...

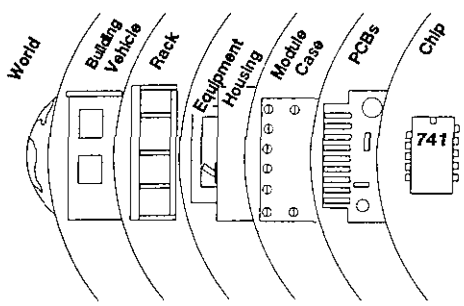


Fig. 1.2. EMC-levels of a system.

At the coupling and transmission path itself, it is more concerned with mechanical properties and choices, such as boxes and housing, connectors, filter mountings, cable mounting, rack mounting of PCB's, ...

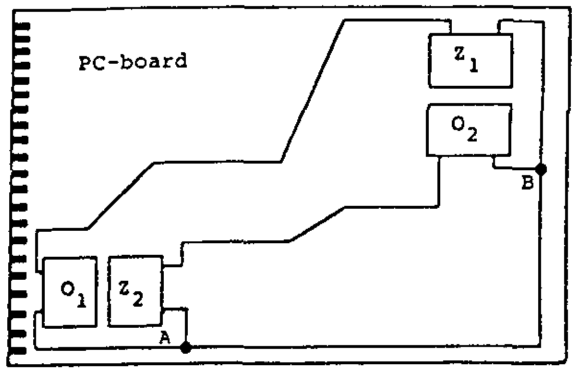
A good EMC designed system has been carefully designed at all three points (source, receiver, path) of a "system".

2. 15 points to remember about EMC design.

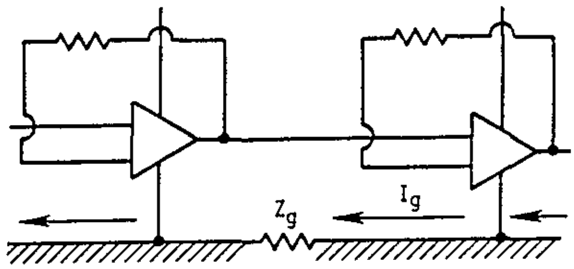
In this section, the previous mentionned 15 items will be discussed in some more detail, with the links to the EMC-noise balance in the system and the typical parameters concerned with to minimise EMC problems.
As stated in the introduction, it is important to minimise the interfering signal level I compared with the own noise level N of the system considered. In this way of thinking, signals I stands for all unwanted signals in a system : this may be the ambient noise coupling into part of a circuit, but also reflections on a transmission line due to mismatch of a load.

2.1. Common impedance coupling.

Common impedance coupling is the effect that currents from different circuits or components are flowing through the same conductor, creating unwanted voltage draps over this common wire. Examples are sketched in fig. 2.1.1.



PCB-example (lay-out)



Circuit-example (op-amp design)

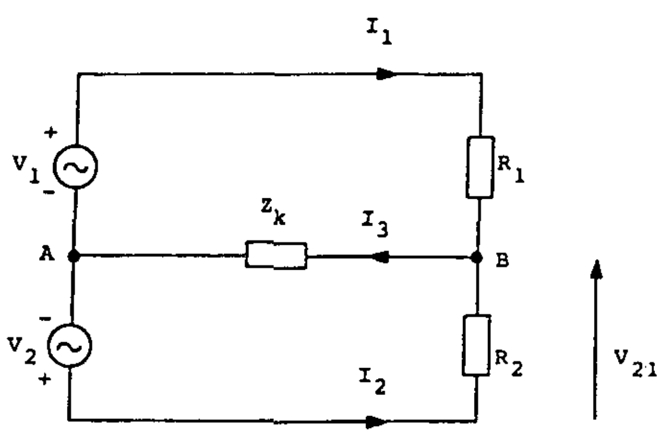


Fig. 2.1. Examples of common impedance.

As mentionned in another paper of this lecture series, a wire or track on a PCB has not only a resistance, but also an inductance. The latter one may be estimated as 1 nH/mm.
Here is an induction L_{AB} = 250 nH.
In the case of TTL logic, a positive going edge on Z_1/O_1 can provide upto 10 nA in 10 nsec.
So, a voltage V_{AB} is induced over L_{AB}.

$$V_{AB} = \mathcal{L}_{AB}\frac{\Delta I}{\Delta t} = 250.10^{-9}\,\frac{10.10^{-3}}{10.10^{-9}} = 250\,m\,Volt.$$

Added to some other minor effects, this voltage can cause a false triggening of the circuit 02. When ESD effects are concerned with, even without direct effect on the components, this voltage drop due to the discharge current flowing through some common impedance of the system will cause a bad working of the system. (Remember for ESD : $\Delta I/\Delta t$ = 1A/10 nsec).

An equivalent circuit diagram for a common impedance coupling was given in the figure 2.1.1.
It is easily found that :

$$V_{21}\,(V_2 = s) = \frac{R_2\ Z_k}{R_1 R_2 + (R_1 + R_2) Z_k} \cdot V_1$$

The voltage V_{21} is reduced by reducing the common impedance Z_k. Normally, circuit-designers and layout-designers (PCB/mechanical) are working in different departments, without any direct supervision officer. For the case of common impedance, a very important problem is the ground-reference, nearly never drown on logic schematics. Some examples of good and bad design are shown in fig. 2.1.2.

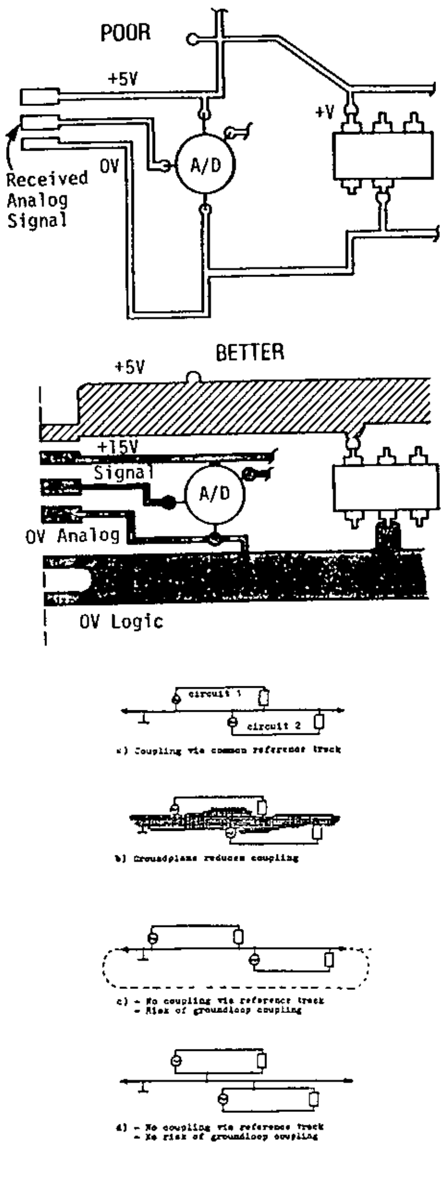


Fig. 2.1.2. Examples of common impedance.

Rules : 1. Don't use high speed (large $\Delta I/\Delta t$ values) or large bandwidth (high frequency swings) for the "sources" if there is no real need to do it.

2. Don't mix ground and power supply lines for different circuits or critical parts of it.
Don't mix (sensitive) analog and (high speed) logic circuiting.

3. Don't clean-up circuit lay-out by combining the latest residues of gates, flipflops,... of different circuit parts together in one package.

BE AWARE OF AUTOMATIC CAD LAY-OUT !

2.2. Common mode coupling

As referenced in fig. 2.2.1., common mode and differential mode working of a system is directly related to the paths currents are flowing through.

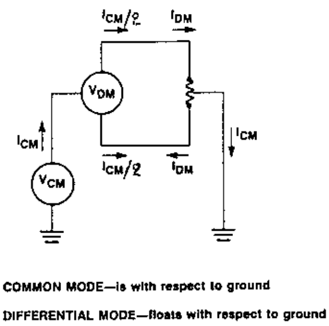


Fig. 2.2.1. Common/Differential mode.

It should be stated very clearly that the difference between both modes has to do with the currents mentionned. This means that for one current path, the system may act as a DM system, and for an induced current path, the system is acting as a CM system. Most common mode effects have to do with "a" ground or earth reference, somewhere in the system. Examples are given in fig. 2.2.2.

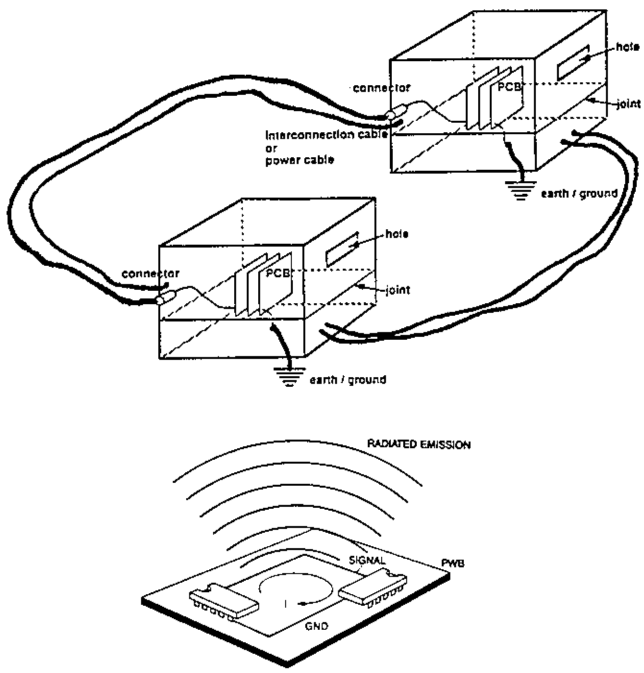


Fig. 2.2.2. Examples of CM in practice.

The common modes occuring may be sketched principally as in fig. 2.2.3.
It is seen that CM coupling is a coupling into or from the ambient electromagnetic fields. The CM effect is created by the large surface area contained within the signal wires and the ground reference. A ground look is created, and a voltage is induced due to the incoming electromagnetic wave, or a radiated field is generated due to CM currents (sec PCB).

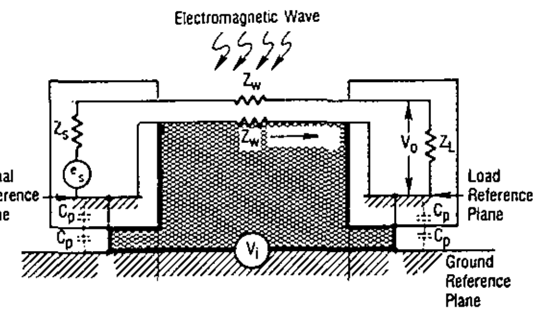


Fig. 2.2.3.

For emission : near field : $H \doteq \frac{1}{d^3} \cdot A.I$

far field : $H \doteq f^2 \frac{1}{d} A.I$

$\doteq f \frac{1}{d} A.V$

For reception : $V_i \doteq \mu_o A \frac{dH}{dt} \doteq \mu_o A f E/120\Pi$
(far field)
with : A = surface or loop area
f = frequency
d = distance to the loop
I = current in the loop
V = voltage over the loop
E/H = electric/magnetic field

Both for emission and reception, it follows that minimising loop area, voltage level (current level) or field level and frequency will minimise the CM coupling effect.

Referrint to far field conditions, there is a 20 dB/decode slope for the relationship between E-field and voltages conerned. However, it should be mentionned that this curve will act with a flat envelope (and sharp dips) when the dimmensions of the loop and the wavelength are of the same order of greatness. It is referred to section 2.2.4. for an example of this relationship.

Rules : 1. Have a good control about (unknown) ground loops. Minimise the loop areas, f.ex. by using multilayer PCB's instead of double sided ones, or good controlled earth wires. If appropiate, use 3-wire systems.

2. Minimise frequency range (or rise/fall-times of logic circuiting). For I/O cables, filtered connectors may be used to achieve this.

3. Minimise signal levels if possible.

As an example, high speed logic should be placed as close as possible to the connector.

2.3.4. Coupling of ground noise.

In both sections, we discussed effects where in the ground a noise voltage was induced. However it is not shure that this voltage is also really and fully coupled into the electronic system itself. Two cases must be considered, an unbalanced and a balanced one.

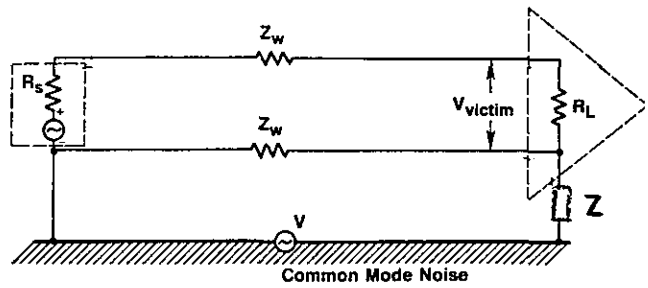


Fig. 2.3.1. Unbalanced system.
(Ref. D. White/ICT)

Analysing this system, the coupled victim voltage due to the induced voltage is given by :

$$V_{victim} \cong \frac{Z_w}{Z} \times \frac{R_L}{R_L + R_S + Z_w}$$

Two parameters are important :

° V_{victim} is directly proportional with R_L, so a low load impedance is interesting.

° V_{victim} is inversely proportional with Z (Z should be high).
Remark however that in most cases, a *parasitic capacitance* is bridging this impedance Z, so that at higher frequencies, this coupling impedance Z becomes small, closing the loop, instead of opening it.

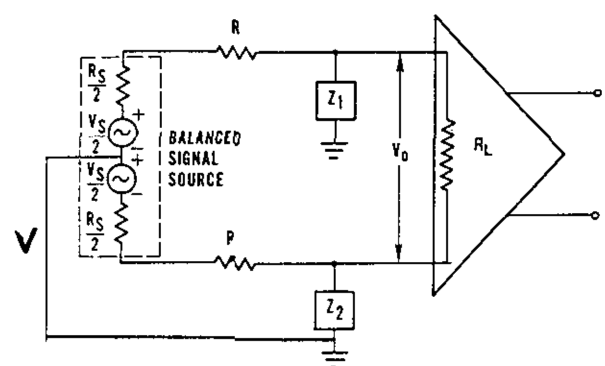


Fig. 2.3.2. Balanced system.

In fig. 2.3.2., a balanced system is schematically given. Z_1 and Z_2 are simulating some unbalances. Analysing this system, the coupled victim voltage is given by :

$$V_{victim} = V\left(\frac{Z_1}{Z_1 + R + R_{s/2}} - \frac{Z_2}{Z_2 + R + R_{s/2}}\right)$$

It follows that for a full balance ($Z_1 = Z_2$), no voltage is induced in the system.

Rules : 1. Use a balanced (or symmetrical) system (and even 3-wire cables)

2. Otherwise, open the ground loop by:

° floating systems, boxes, PCB's
° isolating systems (transformers, opto-isolators,...)
° feed-thru capacitors to pass wires into a box

Be aware for safety earth connections!

3. Low impedance design at load level, and high impedance design at source level (≡ current loops!)

2.4. Differential mode coupling.

Referring to section 2.2., there is no physical restriction why for the differential mode (DM) there should be no radiated field or no susceptibility from incoming fields. However, the loop *areas concerned with are much smaller. This is* schematically shown in fig. 2.4.1.

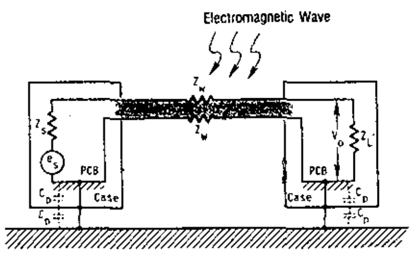


Fig. 2.4.1. Differential mode coupling.

The coupling from an incoming plane wave electrical field into an induced voltage in a loop with dimensions lxs is given in fig. 2.4.2.

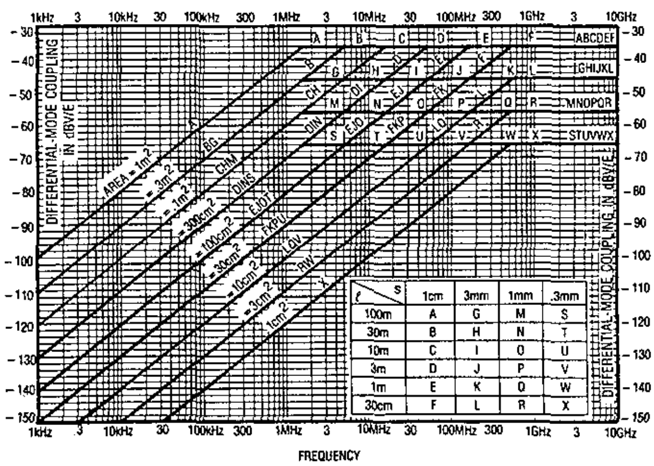


Fig. 2.4.2. DM coupling E-field/loop voltage (ref. D. White/ICT)

The only difference with the coupling of fields in a CM-systems is the loop area, so, in practice CM-coupling is normally the dominant one.

Rules : 1. Minimising the loop area by controlling the separation distance between wires and using twisted pairs (sec. 2.5)

2. Minimising frequency range or rise/fall times

3. Shielding of the DM-wires.

2.5. Coupling of DM signals.

Induced DM signals are directly coupled into a system, in the same way as the signals or data. This means that the only way of acting is to minimise the DM coupling itself. The best action to take is to use twisted pair cables, so that even induced voltages are cancelled, due to the opposite loop-directions. A twisted pair is sketched in fig.2.5.1

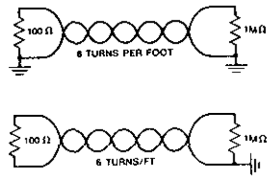


Fig. 2.5.1 Twisted pair

An apropiated use of twisted pair cables, with respect to the same configuration with normal parallel wires gives easily a gain of - 40 upto - 60 dB of less coupling for DM problems.
Better is to use shielded twisted pair cables.
It should also be mentionned that a lot of loops are introduced in a design by currents flowing through a set of conductors. Examples are given in fig. 2.5.2., with loops created by a (bad) power supply/ground reference design, and by connecting *larger systems together.*
For the later case, couples of wires are forming a kind of folded dipole. When the length becomes comparable with the half-wavelength of the frequency components of a complex signal, very efficient E-field antenna's are created !

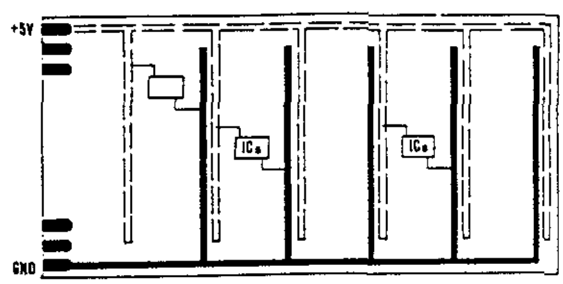
Rules : 1. Minimise the coupling itself by minimising loop area (twisted pair) and shielding.

2. Use feed-thru capacitors to pass wires into a box.

3. Be aware of the existance of wire-like, dipole and folded-dipole antenna-models.

2.6. Filtering of power lines.

The difference between power line filtering and data-line filtering is the separation of the frequency ranges in the spectrum.
For normal situations, filtering power lines is much easier. Typical line-filters are shown in fig. 2.6.



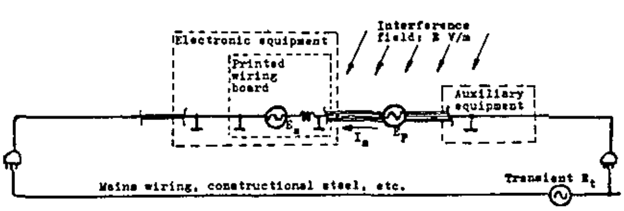


Fig. 2.5.2.

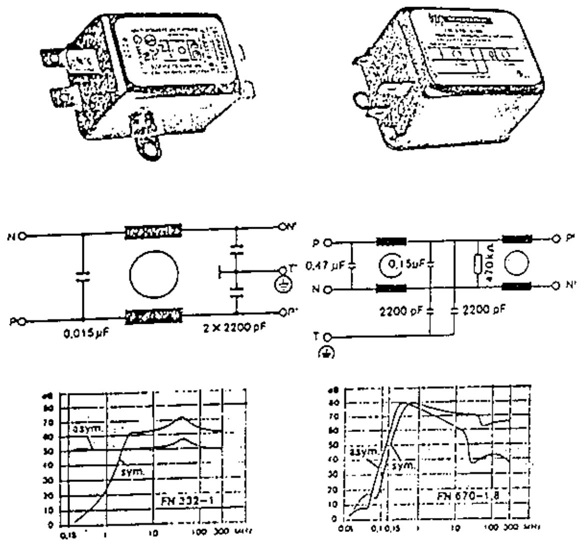


Fig. 2.6.1. Power line filters.

There are three points to mention in order to make a good choice :

° All normal rules concerning source and load impedances from circuit theory must be used, choosing filters with series or shunt branches at their input/output sections. It should be noted that the input section for emission is the output section for susceptibility ! Characteristics and specifications from data sheets are taken for a 50 Ω source/ 50 Ω load impedance. So, no one of the published characteristics may be directly used for normal power line applications!

° Look out for the saturation of the cores of the inductances by the current. Look out for the leakage currents from the power wires to the neutral/earth wire.

° Make a carefull mechanical mounting of the filter, avoiding unfiltered wires passing over filtered ones, or creating radiating loops inside of a system. This is illustrated in fig. 2.6.2.

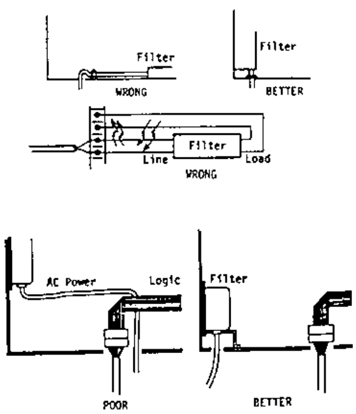


Fig. 2.6.2. Mounting of power line filter.

Rules : 1. Make a choice of power line filters, taking into account impedance levels, currents and leakage currents.

2. Proper mounting of the filter in the system.

2.7. Filtering of data/signal lines.

For the filtering of data/signal lines, in a lot of cases, interfering and useful signals are in the same frequency-range. It follows that filtering will be a hard job, invoking filter theory. Examples of the principal filter circuits are given in fig. 2.7.1.

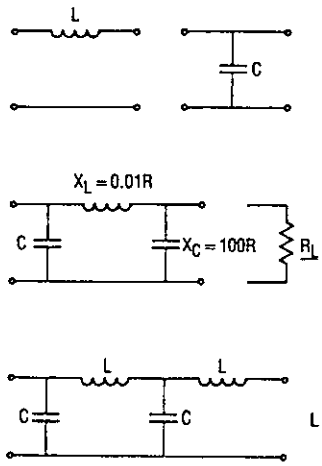


Fig. 2.7.1. Basic filter components.

In fact, all kind of filter types may be used (ex. Butterworth, Elliptic,...) in order to obtain good filtering characteristics.
For more common cases, some simple tricks may be used, such as the use of ferrite beads. These are highly inductive rings, which may be placed over wires, PCB-tracks, included in connectors, ...
Fig. 2.7.2. gives some examples of the application of these ferrites, for CM and DM coupling of signals. For separating DM signals from interfering CM signals, a ferrite ring may be placed over both wires.

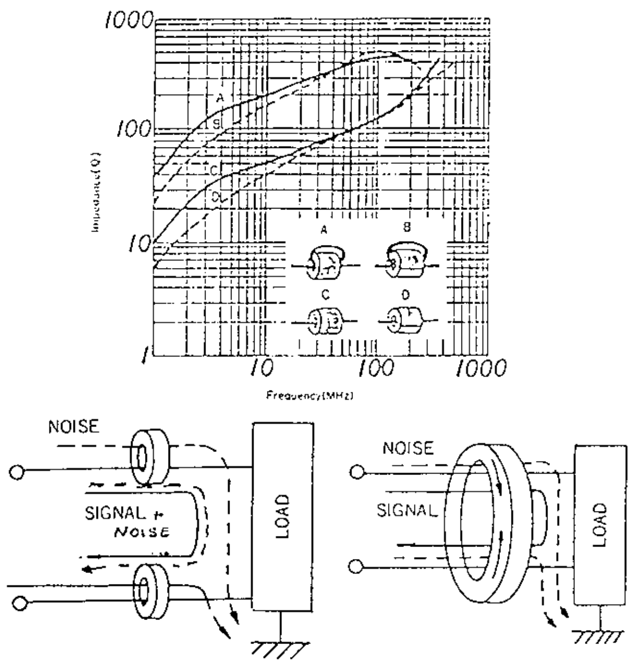


Fig. 2.7.2. Ferrite beads.

For DM-signals, of which both wires are passing through one ring, no flux is generated and the signals are not affected.
So, in this way it is possible to separate DM from CM signals.
In fig. 2.7.3., the mounting of (filtered) connectors is shown, creating a ground loop inside of the system (bad) or giving a good shielding effect (good).

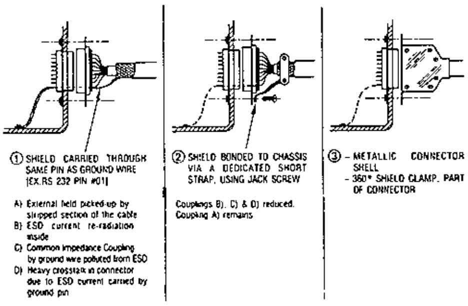


Fig. 2.7.3. Mounting of connectors.

In figure 2.7.4., the construction of filtered pins in connectors is shown.

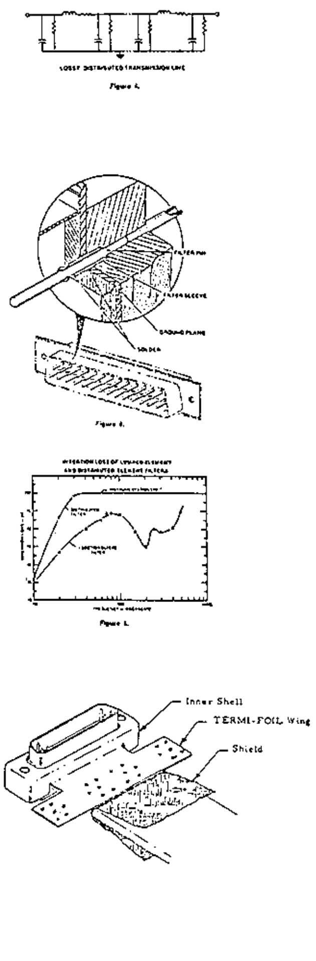


Fig. 2.7.4. Filtered connectors

As allready mentionned earlier, filtering of cables may be used to avoid interfering signals coming into a system (susceptibility), but also to reduce radiation from loops (emission).

Rules : 1. Use specially designed filters for special applications.
2. Use standard ferrites for simple and small dimension applications.
3. Filters, filtered connectors,... should be mounted properly.

2.8. Crosstalk.

For crosstalk discussion, it is assumed here that a low frequency model will be used. However, basic characteristics of crosstalk are obtained, and some general conclusions are obtained.

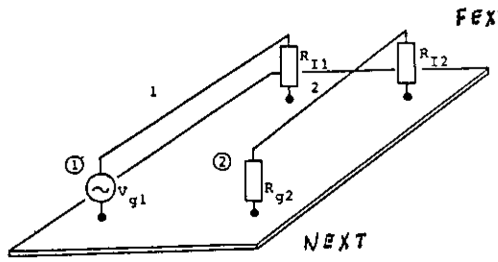


Fig. 2.8.1. Crosstalk of two wires.

The coupling of two adjacent wires or PCB-tracks can be modelled by the following modes: a capacitive and an inductive coupling. They are analysed as follows :

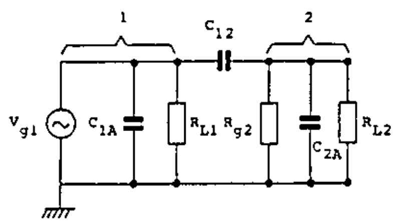


Fig. 2.8.2. Capacitive crosstalk.

It is easily seen that the coupling is given by

$$\frac{V_o}{V_{g_1}} = \frac{sC_{12}.R/2}{1 + s\,\frac{R}{2}\,(C_{12} + C_{2A})} .$$

This acts as a high-pass filter. Three important parameters are concerned with :

° frequency (or the derivative of the voltage waveform)

° coupling capacitor C_{12}

° large value of the capacitor C_{2A}, invoking the influence of a closely coupled ground-plane.

It should also be noted that the capacitive coupling gives the same coupling voltage at both ends of the victim wire.

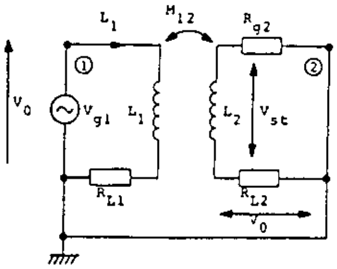


Fig. 2.8.3. Inductive coupling.

For the inductive coupling part of the cross-talk, the circuit of fig. 2.8.3. may be used for analysing it. It follows

$$\frac{V_o}{V_{g1}} = \begin{cases} \dfrac{s\,M_{12}\,I_1}{2} & \omega L_2 \ll R \\[2ex] \dfrac{M_{12}\,R\,I_1}{L_2} & \omega L_2 \gg R \end{cases}$$

This acts as a high pass filter. Three important parameters are concerned with :

- ° frequency (or the derivative of the current waveform)
- ° coupling mutual inductance M_{12}, influenced by a closely coupled ground plane
- ° level of current, invoking the impedance levels at source and load sides.

It should be noted that the inductive coupling gives an opposite phase coupled voltage at both ends of the victim wire.
Combining both effects, it follows that the near end crosstalk (NEXT) and the far end crosstalk (FEXT) are not identical. An example is shown in fig. 2.8.4.

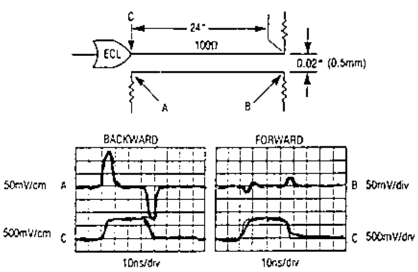


Fig. 2.8.4. Example of crosstalk.

Rules : 1. Use a closely coupled ground reference (plane).
Multilayer PCB's are giving a big advantage for minimising crosstalk.

2. Use a good separation between critical signal wires.
A good example is the isolation given by altering signals and ground in flat cables and connectors.

3. Don't use high frequency speed technology if it is not really required by the system.

2.9. Reflections.

The interconnection of digital circuits and databus systems takes an important place in system-design of modern electronic equipment. Two typical examples are shown in the next figure 2.9.1.

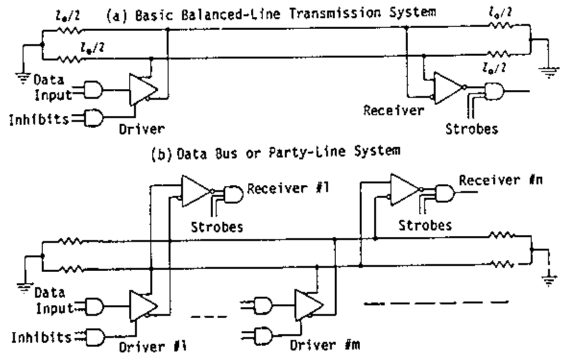


Fig. 2.9.1. Examples of "long-lines".

It follows that some transmission line effects must be considered. This means that $\lambda/l < 20$ or $t_{rise}/\tau < 10$ will give raise to standing wave patterns, reflected power and impedance mismatches. In the time domain, reflection may introduce "ghost" bits, generating false triggering of a circuit.
Graphical methods are used, such as Smith chart, time/distance chart and Bergeron charts (allowing non-linear loads).
Depending on the ratio of load and characteristic impedances, an extra propagation delay or ringing on a line is generated. An example is shown in the picture of fig. 2.9.2.

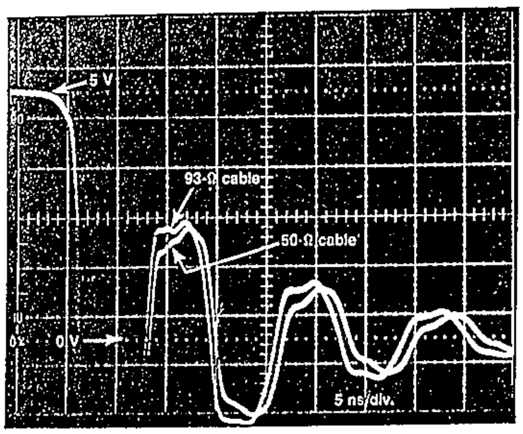


Fig. 2.9.2. Example of ringing due to line-reflections.

Three parameters are involved with this effect:

- ° Characteristic impedance of the line.
- ° Source output impedance.
- ° Load input impedance.

Altering these impedances may change the observed effects. Matching the source impedance is normally done by putting a resistor in serial with the source output. Matching the load is obtained by putting shunt resistors between the signal line and the ground/power supply wires. Also clamping diodes are used, avoiding the generation of reflections. It follows also that any changement of the characteristic impedance will introduce some "reflection-like" noise.
Examples are corners in PCB tracks, transitions in connectors, cable/connector/PCB transitions, pins of IC packages, via's in multilayer PCB's, ...

Rules : 1. Buslike structures must be matched at their ends.

2. Be aware of any sharp transition in a transmission line-like signal path.

3. Make a carefull choice of connectors for high speed/high frequency circuiting.

4. Use cables, back-planes,... with a known characteristic impedance. For PCB-design, this means that microstrip design techniques must be used.

5. When matching loads/sources ..., be aware of the EXTRA power consumption caused by these extra components.

2.10. Shielding.

Shielding of a system is putting a barrier on the radiating path of a system : it may be done to avoid the emission of radiated electromagnetic waves, or to avoid the penetration of incoming waves.
A typical box or housing is sketched in fig. 2.10.1.

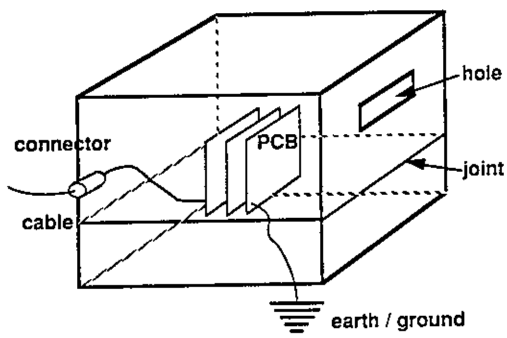


Fig. 2.10.1. Example of housing.

Referring to shielding theory (see another paper of this lecture series), it follows that the shielding effectiveness value of flat materials is depending on :

° material constants (σ, μ)
° frequency
° distance of the source to the shielding wall
° kind of the source wave (E or H).

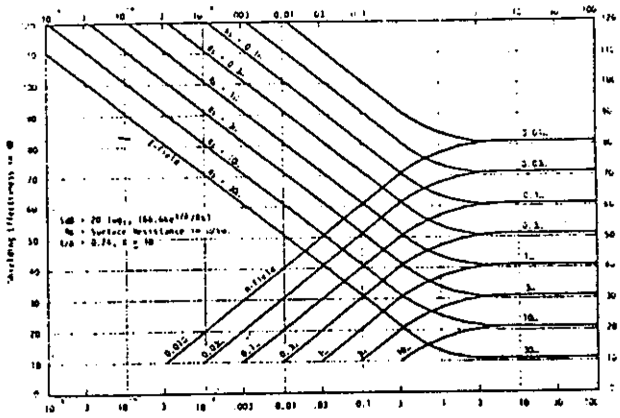


Fig. 2.10.2. SE for flat materials.
(Ref. D. White/ICT)

However, a lot of other (degrading) effects are coming in with the application of real boxes :

° joining structures between two parts of a box.

° joining structures between box and other parts (connectors)

° holes

° displays, switches, ...

As a basic point should be stated that good shielding is only maintained when a good conductive contact is realised througout the whole system. For holes and slots, grids, meshes, honeycomb conductive structures are used, to create "shielded holes".
Fig. 2.10.3. gives a typical SE characteristic for a grid-structure.

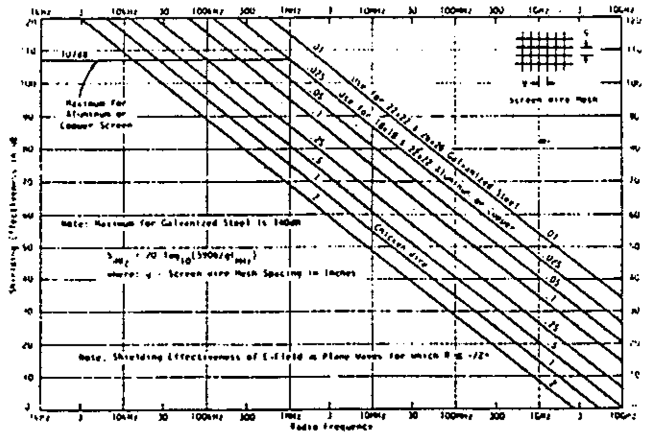


Fig. 2.10.3. SE of a grid.
(Ref. D. White/ICT)

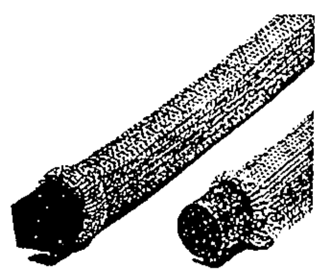
For joining parts of a system, or clean metal contacts are used, or gaskets giving a continuous contact over all the parts. Important on the long run is the effect of galvanic corrosion, degrading the contact impedance over the system, and as a consequence, a degradation of the SE of the whole housing or box .

Global SE is determined by the weakest shielding point.

Rules : 1. Make the exact choice of the shielding box (and material) referring to the basic characteristics (ex. kind of field, distance to source, ...).

2. Take care of the joining structures, also on the long run for galvanic corrosion.

3. Take care of holes, openings,...



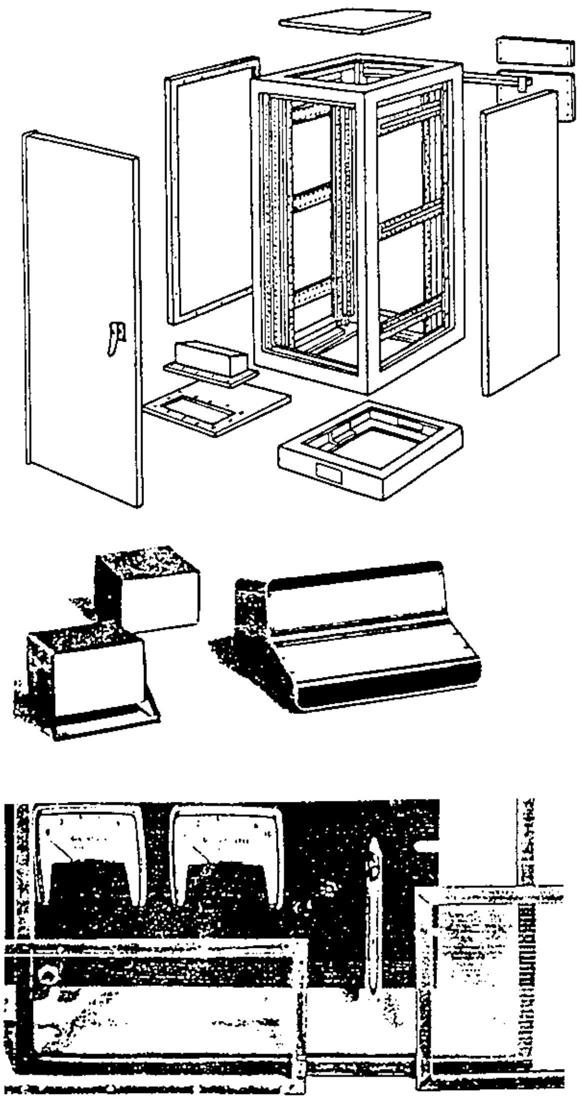


Fig. 2.10.4. Examples of shielding boxes and gaskets, ...

2.11. Decoupling Capacitors.

When logic circuitry is triggered, changement of the state of the transistors is going on inside a circuit. It means that a lot of transient currents are created, and these transient currents are flowing through the power supply lines. Two effects are important :

- ° the voltage drop over these lines, creating a common impedance effect for other circuits, and a drop of supply voltage for the own circuit.
- ° a high frequency radiation because of the current spikes flowing in the loop created by the power line and the ground.

An example of current during state transitions is given in fig. 2.11.1.

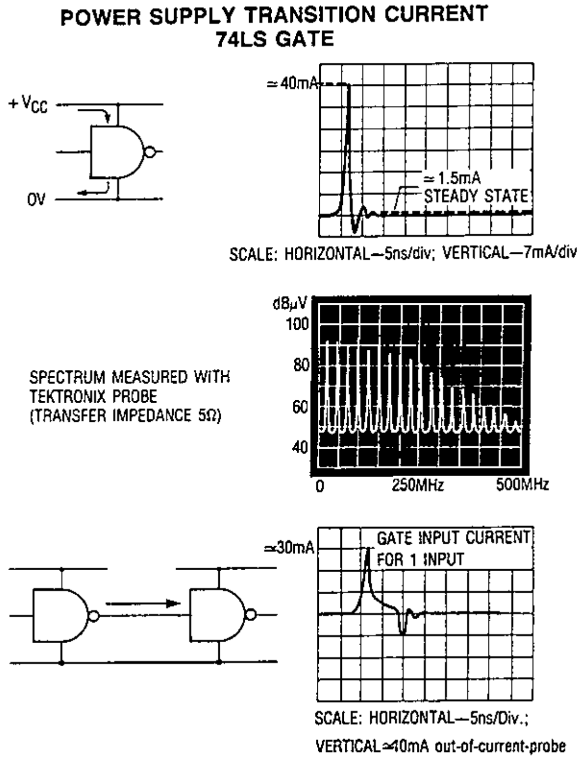
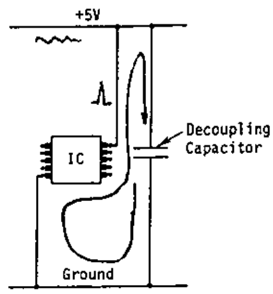
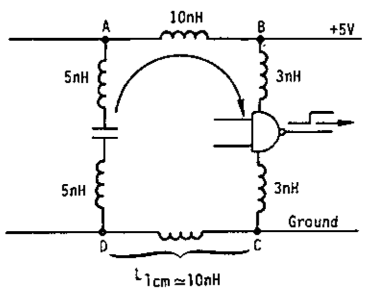
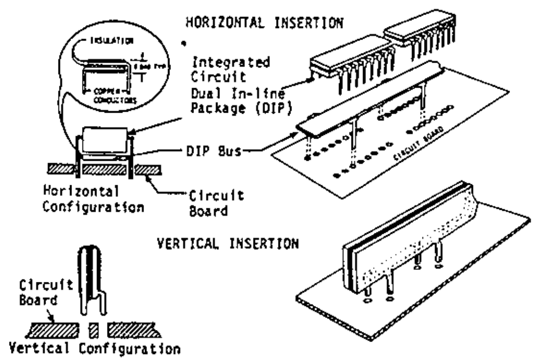


Fig. 2.11.1. Current during state transitions.

By putting a decoupling capacitor nearly a circuit, the high current spikes are retained to the loop formed by this capacitor and the circuit itself. It follows that the longer wires coming from the power supply are carrying only a nearly DC current (with much less radiation).







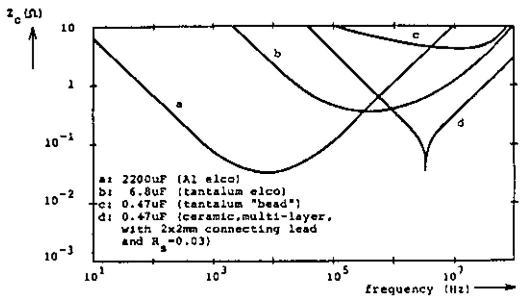


Fig. 2.11.2. Decoupling capacitors.

Three points must be mentionned :

° Regarding radiation problems, the emplacement of the decoupling capacitors must be very close to the circuits itself (and one for each ic).

° Remember that each strip or lead of 1 mm equals an inductance of 1 nH. This means that, in combination with the capacitor, a serial LC-oscillating circuit is obtained. Over the resonance frequency, the circuit behaves inductivally, so no decoupling effect is obtained.

° The same acts for the decoupling capacitor itself.

In the next table, some typical capacitor values are given.

Logic Family	Current Requirements Gate Switch	Gate Drive*	dV = 20% of NIL	dt = Rise Time	Decoupling C = ΔI/(dV/dt)
CMOS	1 mA	1 mA	200 mV	50 ns	500 pF
TTL	16 mA	8 mA	80 mV	10 ns	3000 pF
STTL	30 mA	20 mA	60 mV	3 ns	2500 pF
LSTTL	8 mA	11 mA	60 mV	8 ns	2500 pF
ECL-10K	1 mA	6 mA	20 mV	2 ns	700 pF

* Based on fanout of five gates.

Fig. 2.11.4. Table of decoupling capacitors.

Rules : 1. Choose the right value and type of capacitor.

2. Capacitors must be placed as close as possible to the circuits.

3. General card decoupling capacitors must be doubled by small capacitors at the circuit level.

4. Use distributed types if possible (busbars).

5. Take care during lay-out phase of putting power supply lines and ground closely together, and to connect the capacitors to the right lines/wires.

2.12. Surge protection.

Surges are heavy overvoltages on power lines, data cables, connectors, ...
They are directly related to conductive problems. They are oviginated from the switching of heavy loads, coupling into lines and cables from EMP, lightning, ESD, ... Depending on their origin, these overvoltages (over electrical overstress) have an amplitude of 100's upto 1000's of volts.
Surge protection is done as close as possible to their source-incoupling path, or as close as possible to the incoming path at a system. Protection is done by short-circuiting the overvoltage waveforms on itself (between both wires) or between wires and ground.

An example of installing such devices is given in fig. 2.12.1.

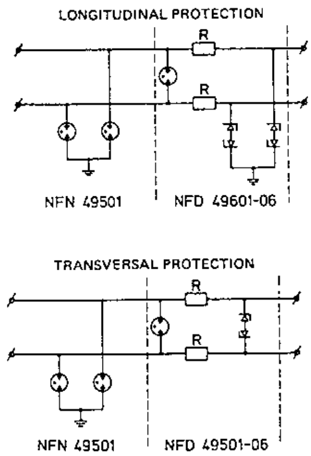


Fig. 2.12.1. Principle protection for electrical overstress.

A wide variety of protecting devices are available, depending on the range of overvoltage (and energy) concerned with :

° gas tubes and neon ampules
° gas diodes
° transzorb
° non-linear resistors (Varistors)
° Zener-diodes
° metal oxide varistors (MOV's)
° silicon avalanche diodes
° capacitors and RC-snubbers.

Furtheron, after removing the heavy overvoltage with one or more of these devices, filtering, deglitching, ... may be done in the same way as allready described for normal filtering techniques.

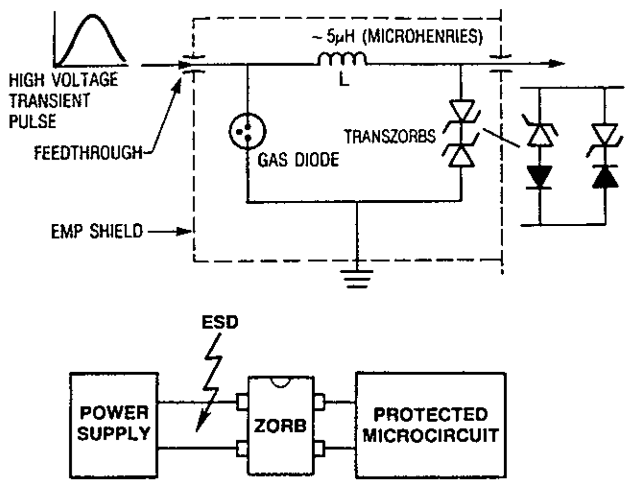


Fig. 2.12.2. Typical protection devices.

In the next figure, an example is given of a good installation of the protection devices, on a good grounded plate and as close as possible to the incoming wires, avoiding that the overstress spikes may enter the shielded box.

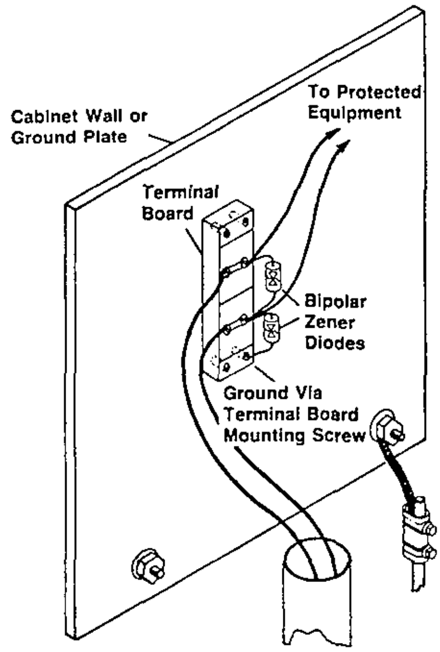


Fig. 2.12.3. Installation of protection device (Ref. D. White/ICT).

Rules : 1. Choose one/more appropiated devices, depending on the overvoltage level and spike duration.

2. Mounting of the devices as close as possible to the incoming wires.

3. Avoid that disturbed wires are really entering the system.

2.13. Non-linear effects and HF-behaviour.

The first effect to be mentionned is concerning the HF-behaviour of components.
It should be clearly understood that :

° *each lead has an inductance of 1 n H/mm*

° everywhere there are two conductors, parasitic capacitances exist.

So, all components have a rather complex behaviour at higher frequencies.

Non-linear effects are the effects of demodulation of LF signals modulated on HF-carriers. This occurs because of the non-linear behaviour of pn-junctions in diodes, transistors, ...
There is only one method to avoid this effect: short-incuiting for the unwanted HF signals the incoming path to the components.
Examples of these techniques are shown in fig. 2.13.1.

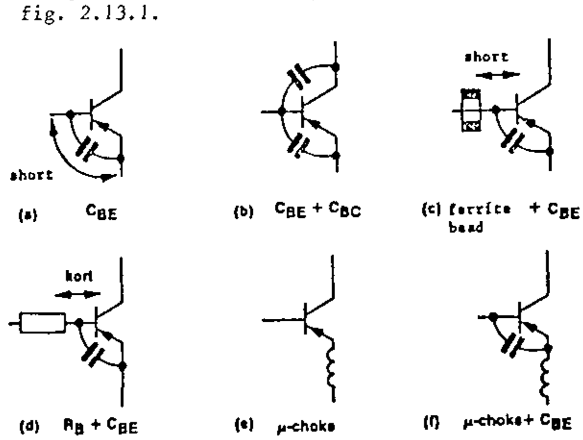


Fig.2.13.1 Avoiding HF signals to enter components.

The effect may be described by :

$$V_{out} = A \cdot V_{in} + B \cdot V_{in}^{2} + \dots ;$$

$$V_{in} = V \cos\omega t$$

*Even without the demodulation effect of broad*cast in signals (or pulse modulated carrier of a radarsystem), another effect occurs :

$$V_{out} = A \cdot V \cos\omega t + B \cdot \frac{V^2}{2} (1 + \cos 2\omega t) + \dots$$

This shows that harmonic frequencies are generated, and also that a DC-shift (of the bias-current) is generated.
The latter influences internally the good working of components and integrated circuits.
This is shown in fig. 2.13.2. for a diode-circuit.

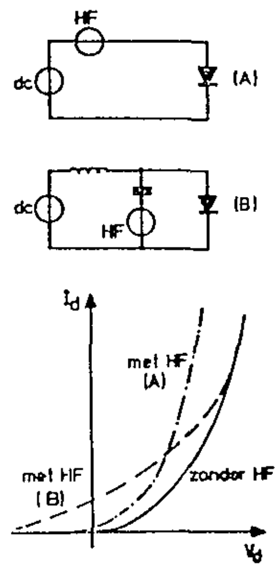


Fig. 2.13.2. DC-shift of diode-circuit due to HF demodulation.

Rule : Avoid that HF-signals may enter the components by short-circuiting the incoming paths, as close as possible to the component-leads.

2.14. Spectrum management.

Spectrum management is the technique to avoid- by a good choice of carrier frequencies for telecom systems, local oscillators, ... that interferences are created at the beginning of the concept of electronic systems.
It is referred to another paper of this lecture series for more details concerning techniques.

2.15. Mechanical design and lay-out.

Good EMC design of a global system is directly depending on both good electronic circuit design, and on the final mechanical design, construction and system/PCB-layout. Examples were allready mentionned in the previous sections handling about CM/DM coupling with electromagnetic fields, shielding, crosstalk ... . Layout and design rules were given concerning separation of wires, ground plane references, ...
In this section, some other examples will be shown about mechanical design and system layout.

But first of all, it should be stated very clearly that even the best electronic circuit design and system concept will act only on its optimal level if a good mechanical design and/or layout has been done.
Two important points must be mentionned :

- It is a basic requirement that the system design engineer is supervising both electronic and mechanical design.
- Concerning PCB-layout, use automatic CAD-routers only for simple and/or non-critical parts of the circuits, but start with the critical paths by a computer-assisted, but manual routing.
  The same rule acts if automatic CAD-routers are putting long connection tracks for rather simple connection paths.
  The reason is that automatic CAD-routers are working very carefully, consequently following the rules of the algorithm, but they are NOT really thinking !

As a first example, multilayer PCB concepts are shown. Typical is the close coupling of power supply/ground plane reference.

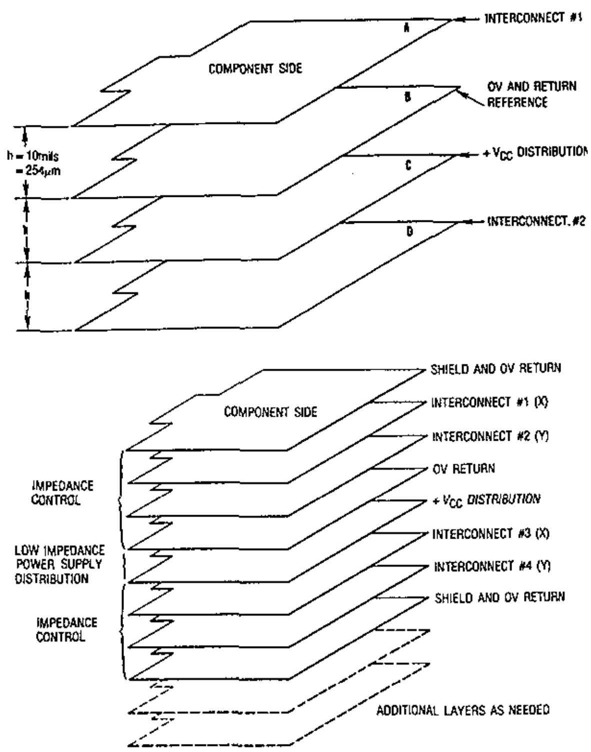


Fig. 2.15.1. Some multilayer PCB's.

A second example concerns the placement of components on a PCB. High frequency and high speed logic circuitry should be placed as close as possible to the connector edge. This arrangement minimises the radiation of the PCB, because the loops concerned with these high frequencies are as small as possible.

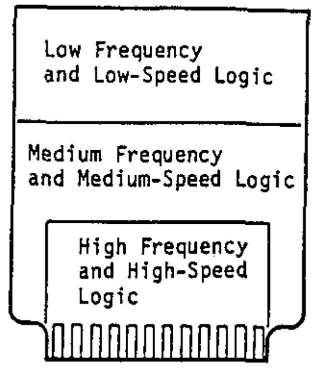
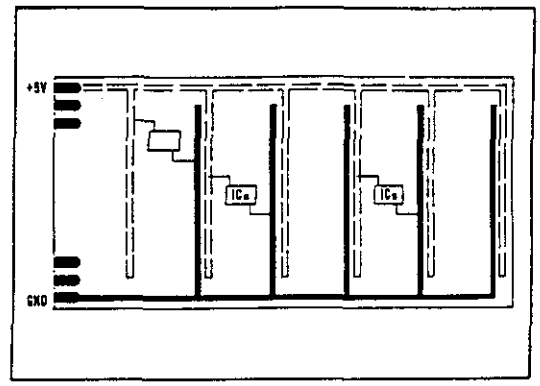
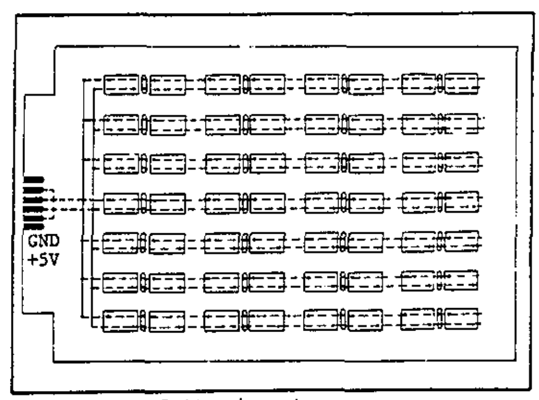


Fig. 2.15.2. Component placement on PCB.

A third example shows a bad and a better layout of the power-supply/ground return distribution on a PCB.



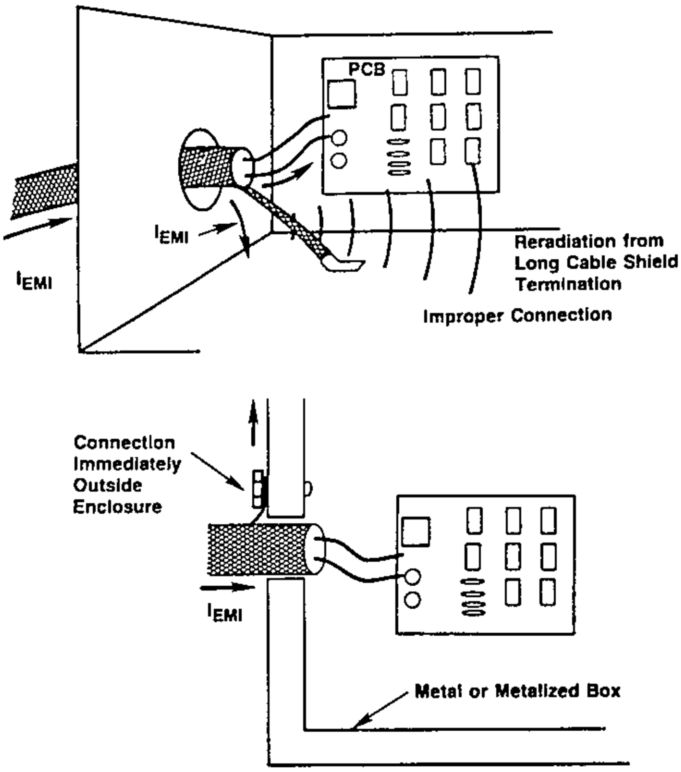
A Bad Layout Giving High Inductance and Few Adjacent Signal Return Paths, Which Leads To Crosstalk.

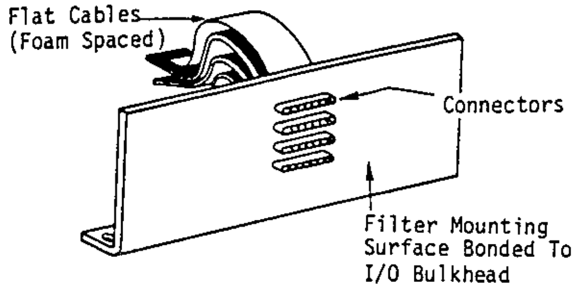


- Better Layout
Reduce Power Distribution and Logic-Return Impedances, Trace Crosstalk and Board Radiation.

Fig. 2.15.3. Power distribution on PCB (ref. D. White/ICT).

The fourth examples shows a bad and a better mounting and coupling of interconnection cables, showing also the famous "pig tail", creating a loop inside of shielded systems.





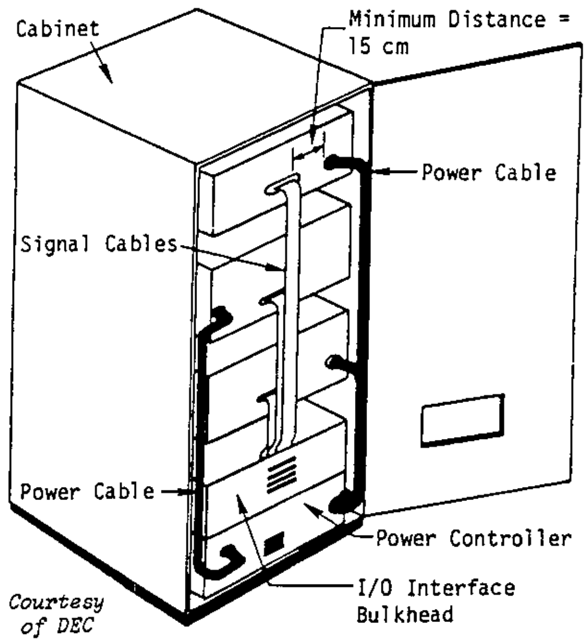
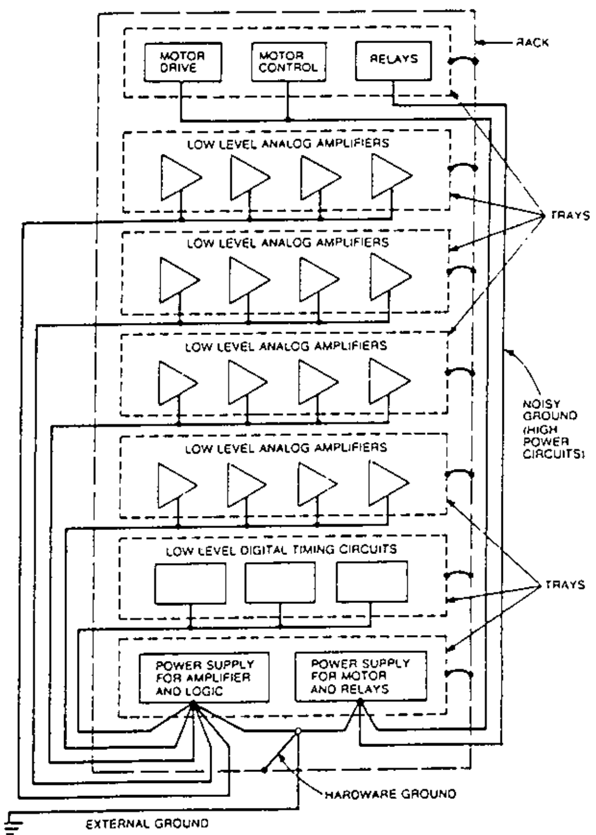


Fig. 2.15.4. Mounting of cables.
(ref. D. White/ICT).

And as the fifth series, some examples are given of general system design.



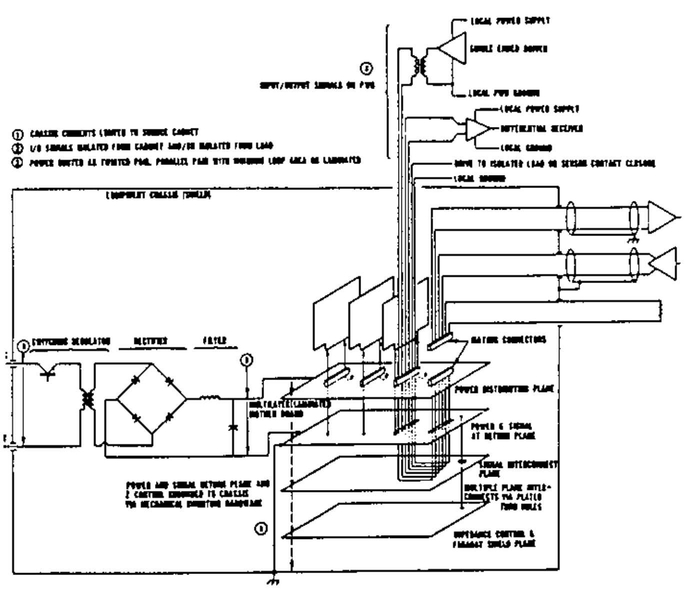


Fig. 2.15.5. General system design examples.
(ref. D. White/ICT).

Rules : 1. Make a very carefull mechanical design, taking into account wire separation, ground reference, pig-tails, ...

2. A global supervision is needed for both electronical and mechanical design by the system design engineer.

3. Be aware of AUTOMATIC CAD-routers for PCB-layout.

3. How To Tackle EMI-problems ?

In this section, some rules for good shielding practice will be discussed.

3.1. How to tackle the problem ?

1. Define the type of EMI-signal
2. Define the type of the problem (R or C)
3. Choose the solution : EMI, EMS or shielding/ filtering
4. Have an idea of the signal levels
5. Localise the real problem
6. Is there also an ESD problem ?
7. Other features needed ?
8. Choice of solutions and materials
9. (Re)design
10. TEST AGAIN !

We will discuss these 10 points now more in detail.

3.1.1. Define the type of EMI-signal :

- what is the frequency-range/frequency-band
- is it a small-band/broadband signal
- what is disturbed : a system, a circuit,...
- in what application field (industrial, consumer,...)

3.1.2. Define the type of the problem :

- conduction or radiation
- intersystem or intrasystem
- is the source natural or man-made
- is it in the near-field/far field region of the source
- E-field or H-field problem

3.1.3. Solution ?

- Redesign : in many cases, a good redesign (e.g. PCB's) might resolve the problem, in order to minimise the emission level itself, or the susceptibility level.
- EMI : shielding to avoid that an EMI-signal may leave a system. (see EMI-regulations). This solution is not suitable for natural and intentionally man-made EMI-sources.
- EMS : shielding the "receiver" against disturbing signals.

3.1.4. Signal levels :

- what is the signal-level (or field-strenght) of the EMI-signals
- what is the sensitivity-level (susceptibility) of the receiving system or circuit
- set requirements to achieve.

3.1.5. Localise the real problem :

- component in a circuit. circuit in a system: design of the system.
- shielding problem of a whole enclosure
- problem about holes, apertures for displays, connectors, buttons, doors, cooling-in lets, ...)
- power line/data cablefiltering.

3.1.6. ESD ? (Electro-static discharge) :

Try to find-out if there should be any ESD-problems. ESD-sparks may cause an EMI-problem. And, EMI/EMS solutions use normally the same kind of materials.

3.1.7. Other features ?

- air-cooling filters
- optically transparent
- anti-dust sealing
- air-pressure sealing
- oil-leakage
- moisture exclusion
- shock-proof
- temperature range
- fixation (see also galvanic corrosion)

3.1.8. Choice of sòlution :

- redesign of the electronic system
- choice of components
- power line filters
- shielded/filtered connectors
- box/enclosure for a whole system
- shielding the source-circuit or the sensitive circuit only
- hole slot shielding by : cut-off wave guide
  gaskets
  seals
- interconnections of systems by coding of signals (double) shielded cables twisted/ balanced wires.

3.1.9. Redesign of the system

Is an EMI-redesign possible, without redesigning the whole system (mechanically compatible) ?

3.1.10. Test again, and ???

THINK ALWAYS

E M C

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14. Abstract

Two aspects of the current electromagnetic (EM) environment having great significance for NATO systems are:

(a) electromagnetic interference (EMI) arising from both natural and man-made sources;

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The Lecture Series ~~will first set~~ EMC in a NATO operational context. Major EMI generation mechanisms ~~will then be~~ reviewed and their characteristics outlined. The manner in which EMI energy couples into EM systems ~~will be~~ is discussed, together with the analysis and modelling tools available to assist in computing such interactions.

Modern EMC testing methods and environments ~~will then be~~ examined and their limitations indicated. The relationship between NATO and civilian EMC requirements ~~will be~~ is examined in the light of the European Community EMC Directive. Consideration ~~will also~~ be given to the problems of spectrum management and conservation as they affect systems intentionally radiating EM energy, e.g. radio, radar, etc.

Finally, design principles and techniques for EM systems with effective EMC characteristics ~~will be~~ presented. A Round Table Discussion will enable attendees to interact in some detail with the lecturing team.

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