3.5.3 STRANDED CABLES AND BRAIDS

A stranded cable consists of a number of wires in close proximity and twisted about each other; it is more flexible than a solid conductor of the same cross-sectional area. Generally for a given cable size both the R_{ac} and the self-inductance of multi-strand conductors are greater than those of solid conductors. This can be ascribed to the significance of the combined skin and proximity effects which increases the resistance as the frequency increases. Furthermore, due to the smaller diameter of the individual wires in a braid, the impedance can be very sensitive to corrosion.

Because of their ineffectiveness at high frequencies, it has on occasion been recommended that stranded cables not be used at frequencies higher than 1.2 kHz. However, in many situations, large cables are required to safely carry currents produced by power faults and lightning discharges; in addition, solid conductors may be unavailable or difficult to install and thus the use of stranded cables may be unavoidable.

3.5.4 STRUCTURAL STEEL MEMBERS

A steel I-beam in the structural framework of a building is another conductor that is frequently used as a ground conductor. Although the resistivity of steel is approximately ten times that of copper and the skin-depth of steel more than three times that of copper, the increased conducting area in steel lowers the resistance at high frequencies to a value comparable to that of copper. This advantage is offset somewhat by the fact that the current tends to flow in the edges of the I-beam and by the surface roughness. The ac resistance will be increased by a factor of four because of this surface roughness and the current redistribution. In addition, the building framework usually offers many paths in parallel, thus lowering both the ac resistance and the inductance between any two points.

3.6 SUMMARY

1. Because of the reactance, the advantages offered by large cross-sectional area conductors are less than they might appear to be from a comparison of the dc resistance values. Hence, the preference for straps.

2. The advantages offered by a large area cable will be somewhat more pronounced for relatively short conductor lengths than for long conductor runs. This is so because inductance increases more rapidly with length than does resistance.

3. Because of lack of dramatic improvement in ac impedance of large cables over smaller cable sizes for long runs consideration of materials (i.e. low resistivity and cost) and labour costs are relatively important and may be the deciding factor.

4. The requirements for low cost, low impedance and current-carrying ability will also be influenced by the amplitude and power spectrum of lighting surge currents, and the radio-frequency noise considerations.

3.7 TRANSMISSION LINE AND FREQUENCY EFFECTS OF GROUND CONDUCTORS

Another factor which influences ground conductor impedance characteristics at high frequencies is the phenomena of transmission line effects. Any ground cable, power or signal, can be viewed as running along in the proximity of a ground plane. Considering the properties of conductors, i.e. R_{skin}, inductance and capacitance, a ground conductor circuit can be represented by these elements distributed as shown in Figure 3.8. At low frequencies the effects of the resistive elements dominate.

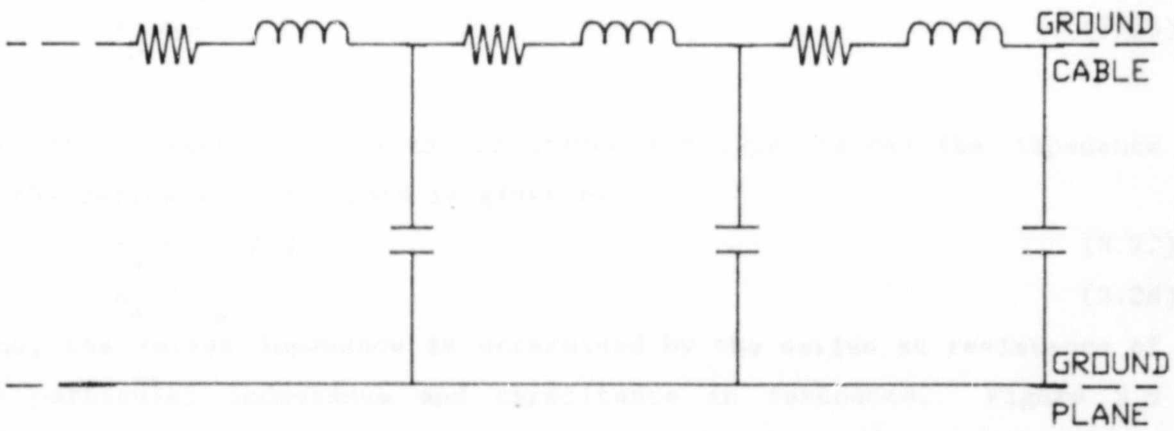

Figure 3.8 Equivalent Circuit of Ground Cable with Distributed Elements

However, as the frequency increases the inductive and capacitive influence begin to dominate until the point of resonance is reached where the inductive and capacitive reactances cancel each other out. The first frequency at which resonance occurs can be determined from:

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{3.23}$$

where L is the total cable inductance and C is the net capacitance between the cable and the ground plane. At resonance, the impedance presented by the grounding path will either be high or low, depending on whether it is parallel or series resonant, respectively. At parallel resonance, the impedance seen looking into one end of the cable will be much higher than expected from $R+j\omega L$.

At parallel resonance:

$$Z_p = Q\omega L \tag{3.24}$$

where Q, the quality factor, is defined as

$$Q = \frac{\omega L}{R_{ac}} \tag{3.25}$$

where R_{ac} is the cable resistance at the frequency of resonance. Combining Equations 3.24 and 3.25

$$Z_p = \frac{\omega^2 L^2}{R_{ac}}. \qquad (3.26)$$

For series resonance (due to inductance and capacitance) the impedance of the series resonant path is given by

$$Z_s = \omega L \,/\, Q \qquad (3.27)$$

$$Z_s = R_{ac} \qquad (3.28)$$

Thus, the series impedance is determined by the series ac resistance of the particular inductance and capacitance in resonance. Figure 3.9 illustrates these resonance effects in ground paths. Thus the impedance behaviour as a function of frequency determines the relative effectiveness of a grounding path.

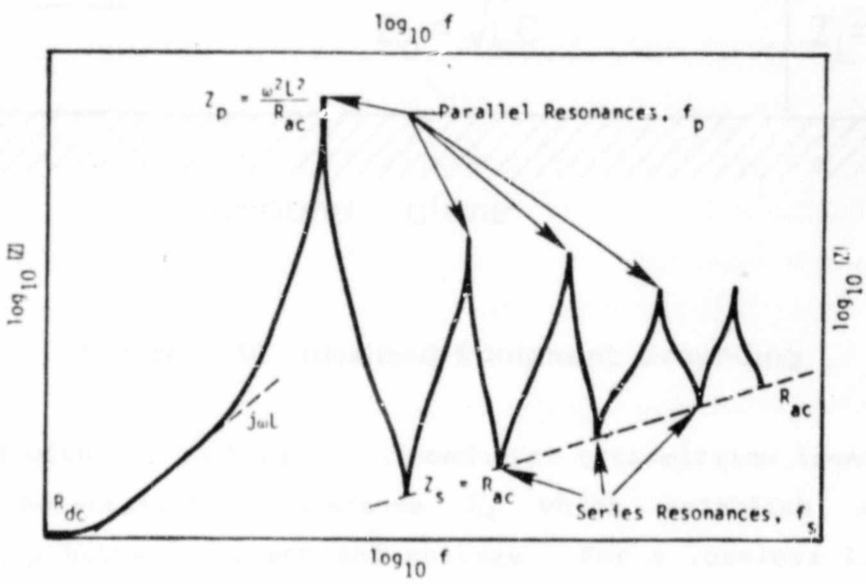

Figure 3.9 Impedance vs Frequency Behaviour of Grounding Conductor (Source: Denny[1])

The high frequency behaviour of a grounding path can be simplified by viewing it as a transmission line as shown in Figure 3.10. At high frequencies every conductor that has appreciable length compared to the wavelength will radiate energy. The energy radiated by the conductor can be much greater than the losses caused by the conductor resistance, if the terminating impedance of the conductor is not equal to the characteristic impedance. In normal transmission, the radiation problem is solved by using another conductor whose electromagnetic field is equal and opposite to that of the first conductor, thus the resultant field is zero everywhere in space. The second conductor is usually parallel or coaxial to the first. However, in the case of a single ground conductor, where only one cable is used for transmitting the current, the current is returned by an apparent conductor formed from the image of the ground conductor, as shown in Figure 3.10.

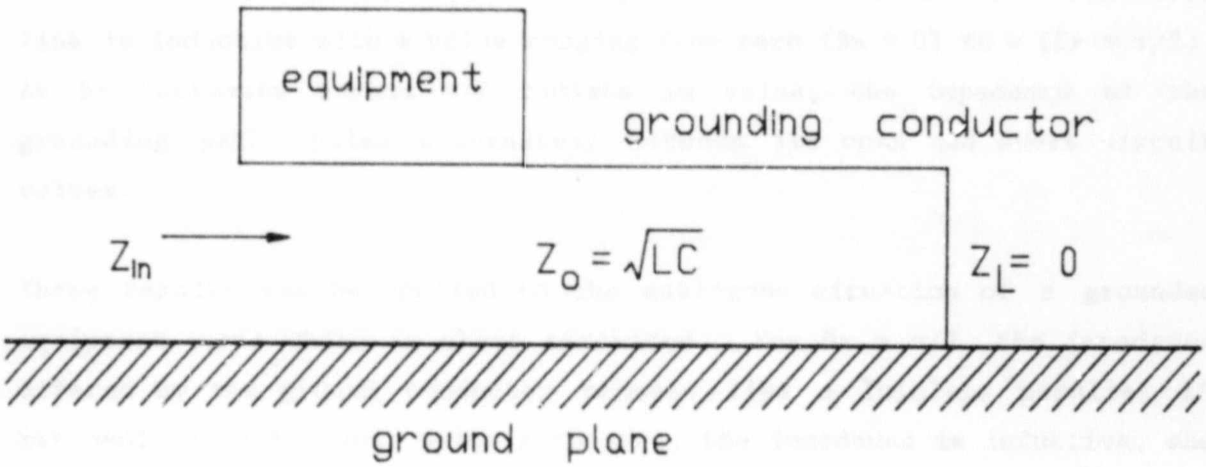

Figure 3.10 Idealised Equipment Grounding

Associated with all such parallel conductor transmission line situations is the characteristic impedance Z_0 which establish a definite relationship between current and voltage. For a lossless line where a resistance is zero or $R/X_L << 1$, the value of the characteristic impedance is equal to $(L/C)^{\frac{1}{2}}$, where L and C are the inductance and capacitance per unit length of the line. However, for the ground conductor the distributive capacitance is not always so well defined. If a ground conductors is routed close to a metallic surface its capacitance can be reasonably well calculated and thus its characteristic impedance can be determined.

For the situation illustrated in Figure 3.10 the input impedance of the grounding path, i.e. the impedance to ground seen by the equipment case, is[55]:

$$Z_{in} = jZ_0 \tan\beta x \tag{3.29}$$

where

β = the phase constant of the transmission line
$= \omega\sqrt{LC}$
$= 2\pi/\lambda$

x = the length of the path from the box to the short

$1/\sqrt{LC}$ = velocity of the travelling wave

Note, for βx between zero and $\pi/2$, i.e. when the electrical path length varies between zero and $\lambda/4$, the input impedance of the short-circuited line is inductive with a value ranging from zero ($\beta x = 0$) to ∞ ($\beta x = \pi/2$). As βx increases beyond $\pi/2$ radians in value, the impedance of the grounding path cycles alternately between its open and short circuit values.

These results can be applied to the analogous situation of a grounded equipment case which is short circuited. For $\beta x = \pi/2$, the impedance offered by the ground conductor behaves like a lossless parallel LC resonant circuit. Just below resonance, the impedance is inductive; and above resonance, it is capacitive; while at resonance the impedance is real and quite high (infinite in the perfectly lossless case). Resonance occurs at values of x equal to integral multiples of quarter wavelength. Thus it is clear that for maximum efficiency ground conductor lengths should be a small portion of the wavelength at the frequency of the signal.

Furthermore, for a maximum energy transfer requirement in transmission lines, the terminating impedance should be equal to the characteristic impedance. When R_t (terminating resistance) is not equal to Z_0, only part of the power is absorbed by R_t and the remaining power is reflected back. In extreme cases, where R_t is zero or infinite, all of the power is reflected back. When R_t is zero, the current is maximum and the voltage is minimum (the voltage being 180° out of phase)

Thus, when ground conductors are used, and the total length of the conductor is not small compared to the wavelength of the frequency of interest, then the characteristic impedance should approximate the terminating or contact impedance between the conductor and the case or ground reference of interest. However, if the source and load impedances do not match the characteristic impedance, the strap will act as a radiator. Since the effort involved in matching impedances could be enormous, it is usual to optimise the conductor length to wavelength ratio to reduce the radiation or pickup capability of the ground conductor.

The circuit resonance behaviour can be related to the antenna effects. Ground conductors at high frequencies can act as antennas to radiate or receive potential interference energy if the conductor length is appreciably close to the wavelength of the current. This fact enables one to derive the desired ground conductor length relative to the wavelength.

From antenna theory, the antenna efficiency is given by[98]

$$k = \frac{R_r}{R_r + R_{loss}} \tag{3.30}$$

where

R_r = radiation resistance

R_{loss} = loss resistance due to any heat loss

$$R_r = 80\pi^2(\ell/\lambda)^2 \tag{3.31}$$

A good measure of performance for a wire is a quarterwave monopole whose radiation resistance $R_r = 36.5\Omega$. The objective in this case is to strive for a very inefficient antenna. A value for k equal to 10 percent or less can logically be defined as inefficient. Thus, for a ground conductor to be an inefficient antenna it should exhibit a radiation resistance of 3.65Ω or less (assuming R_{loss} remains small and constant).
Solving for ℓ/λ in Equation 3.31 gives $\ell/\lambda = 0.068 \approx 1/15$.

The above result simply means that a good criterion for a poor antenna, i.e. a good ground wire, is that the length does not exceed 1/15 of the wavelength of the highest frequency of interest. This result agrees with that given by Kendall[14]. Other researches in the field have quoted different values. The value given by Denny[1] is 1/10, that by White[2] as 1/20 and that by Ott[28] as 1/12. As will be shown in Chapter 5, the transitional region is between 1/10 and 1/20 and thus any value in this region is acceptable. However, the value of 1/15 derived above will be used in this study, although the other values would apply equally well. Hence, a recommended goal in the design of a effective grounding system is to maintain ground wires exposed to potentially interfering signals at lengths less than 1/15 of the wavelength of the interfering signal.

3.8 PRACTICAL CONSIDERATIONS IN THE SELECTION OF GROUND CONDUCTORS

The purpose of this section is to illustrate the application of the equations and results presented in this chapter. In order to begin the design of a ground conductor and/or facilitate selection of correct ground conductors, the designer should consider the following questions:

1. What are the major frequencies for which a ground path must be provided?
2. What is the maximum current that can be expected at the frequencies listed in question 1.
3. What is the maximum voltage drop permissible at the frequencies and currents listed above?

The points listed below guide the choice of a ground conductor and is an application of the equations and information already presented in this chapter. The designer can use the given information to:

1. Select a type of conductor which, based on the information given in the discussion section, would provide a low ac impedance at the relevant frequencies (for example, flat copper strap).

2. Obtain the dc resistance of the ground conductor from Equations 3.1; 3.2; and 3.3.

3. Calculate and obtain the inductive reactance from Equations 3.5; 3.6; 3.7 and from Tables 3.2 and 3.3.

4. Obtain the high frequency ac resistance R_{skin} at the frequencies of interest from Equations 3.13; 3.18; 3.19 and Table 3.4.

5. Calculate the required length of the conductor and compare it to the wavelength of the frequencies of interest, i.e. ensure the conductor length does not exceed 1/15 of the wavelength.

6. Calculate the resonant frequency of the conductor from Equation 3.23

Noting, that ground conductors loose their effectiveness at frequencies above the parallel resonance, the total length of the conductor must be reconsidered in terms of the wavelength of interest and the length to width ratio if the frequency of interest exceed the resonant frequency. Thus, based on the facts, the use of preset values in the design of a ground conductor, such as a 5 to 1 length to width ratio, can cause the strap to act as an open circuit instead of a short circuit. So, in order to insure a low impedance ground path, all of the variables discussed herein would need to be considered.

CHAPTER 4

4.0 INTERFERENCE COUPLING MECHANISMS

4.1 INTRODUCTION

A large number of diverse equipments usually characterise a modern process control environment. To perform all the required tasks and functions and to ensure smooth operation, the control systems, power distribution and signal transmission networks must co-exist and work as an integral unit.

A characteristic feature of a process control installation is that many potentially incompatible signals are present. As noted in Chapter 2, these typically vary from the high power levels to the low signal levels. Furthermore, such signals could range in frequency from a few hertz to several Megahertz. Falling in overlapping frequency ranges, these various signals may interact in an undesirable manner to cause interference which could be damaging. A major objective of interference reduction in modern electronic control systems is to minimise and, if possible, prevent degradation in the performance of the various electronic systems by the interaction of undesired signals, both internal and external. Thus, a clear understanding of the different coupling mechanisms and paths is required which would form the basis for the design of ground systems with the assurance that the interference threat has been reduced to a minimum.

In this chapter, coupling principles are examined which include coupling between conductors, between conductors and equipment and that between equipment. Five different coupling mechanisms are identified. These are common-mode impedance coupling, inductive and capacitive coupling, common-mode radiation coupling and differential-mode radiation coupling. An understanding and appreciation of these various forms of interference is essential when designing the ground system for any facility. Furthermore, the principles developed in this chapter form the basis for ground loop and other interference avoidance techniques presented in later chapters.

4.2 COUPLING MECHANISMS

Coupling is defined as the means by which a magnetic or electric field produced by one circuit induces a voltage or current in another circuit. Interference coupling is the stray or unintentional coupling between circuits which produces an error in the response of one of the circuits[1].

Interference is broadly classified by its coupling means; i.e. as either being conductive or free-space.

- **Conductive coupling** occurs when the interfering and the interfered with circuits are physically connected with a conductor and share a common impedance.

- **Free-space coupling** occurs when a circuit or source generates an electromagnetic field that is either radiated and then received by a susceptible circuit or that is inductively or capacitively coupled to a susceptible circuit.

Two interference modes namely, differential-mode (normal-mode) and common-mode, are defined below and these are important to the understanding of the interference mechanisms described later. Refer to Figure 4.1 and the following extract.

4.2.1 DIFFERENTIAL - MODE INTERFERENCE (NORMAL - MODE)

In a two wire line, the interfering voltage and/or current is of equal amplitude and opposite in phase. This applies to all power mains and signal transmission cables. Differential-mode interference is introduced into the signal channel through the same path as that of the legitimate signal. Closer analysis of the nature of such signals indicates that they often have frequency characteristics which differentiate them from the desired signal.

This type of interference originate primarily from other users on the same power mains. They are usually preceded and/or terminated by a transient. Examples include any source operated for a short period of time such as motor-operated machine-shop equipment, ultrasonic cleaners and electric drills.

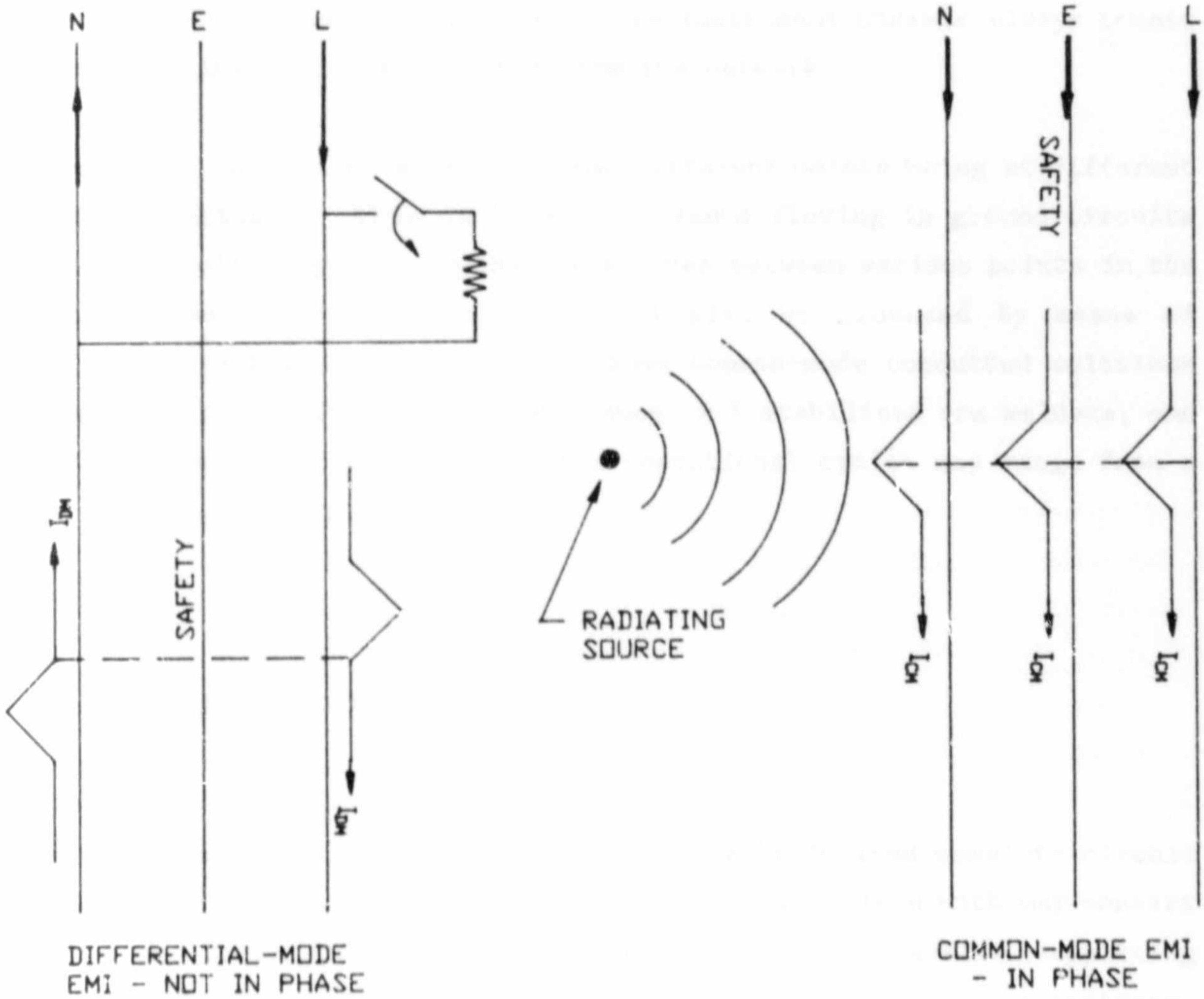

Figure 4.1 Differential and Common-Mode Interference

4.2.2 COMMON - MODE INTERFERENCE

In a two-wire line, the undesirable EMI voltage and/or current in each wire is more or less equal and in phase with respect to each other. The degree of line amplitude balance usually increases with frequency. This interference is normally introduced into the signal channel from a source having at least one terminal which is not part of the legitimate signal channel. In short, the current path for the interference is only partly common with the signal current path. The instrument chassis always counts as one terminal, if not isolated from the network.

Common-mode interference results from different points being at different ground potentials. This is because currents flowing in ground circuits result in volt drops due to the resistances between various points in the plant. Common-mode interference could also be produced by means of electrical pick-up in conductors. Some common-mode conducted emissions usually originate from fluorescent lamps, R-F stabilised arc welders, and diathermy equipment radiations whose operational cycles may range from a few minutes to hours.

4.3 CONDUCTIVE COUPLING

The conducted coupling path is generally a well defined complete circuit transfer path between the source and the susceptor. This path may consist of wiring, power supply, control common, equipment chassis, supporting metallic structure, a ground plane, or mutual inductance or capacitance. Thus any direct connection between two circuits with a return path allows conductive coupling transfer to occur.

Typically, wiring entering a facility provide good conductive coupling paths for interference sources external to the facility. This interference is easily conducted into a particular unit or piece of equipment which, in the case of lightning, could cause extensive damage.

In any process installation the predominant routes for conduction interference currents are power supply lines, control and accessary cables, and ground returns.

Within equipment, interference can also conductively couple between various circuits on the common signal or dc power lines. If one dc power supply is utilised with several circuits operating over various signal voltage and frequency ranges, the operation of one circuit may adversely affect the operation of other circuits. This form of coupling is referred to as common-mode impedance coupling and is examined below.

4.3.1 COMMON - MODE IMPEDANCE COUPLING

The signal reference plane is a common potential coupling path for unwanted signals between equipment and/or circuits. Since practical signal reference planes do not exhibit a zero impedance, any current flowing in such a plane will produce potential differences between various points on the reference plane. Interfacing circuits (equipment) referenced to these various points can experience conductively coupled interference in the manner illustrated in Figure 4.2.

The signal current I_1, flowing in circuit 1 returns to its source through signal reference impedance Z_R, producing a voltage drop V_{n1} in the reference plane. Since the impedance Z_R is common to circuit 2, the extraneous voltage V_{n1} also appears in series with circuit 2 voltage source, V_{s2}. Thus this undesired source produces an interference voltage, V_{n2}, across the load of circuit 2. Similarly, the desired current I_2, may produce interference in circuit 1.

In an industrial facility, where hundreds of control loops exists, the conductive coupling of interference through the signal reference plane of interfaced equipment can occur in a manner similar to that described above.

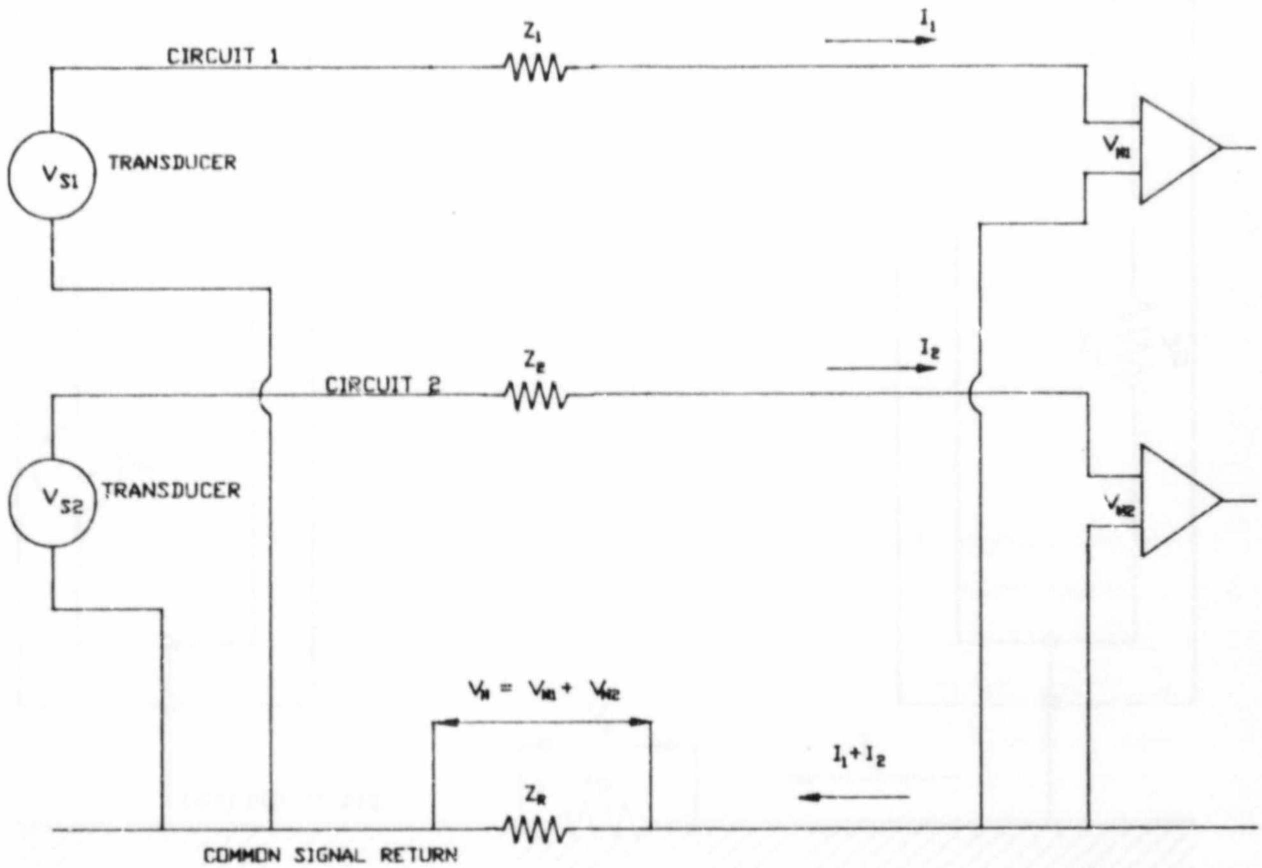

Figure 4.2 Common-Mode Impedance Coupling

Figure 4.3 illustrates the basic circuit concepts of interference between equipment interconnected by cable. In this instance, the existence of the stray current I_R may be the result of the direct coupling of another equipment pair to the signal reference plane, or even external coupling due to an incident field. In either case, the current I_R produces a voltage V_n as a result of the small but finite ground reference impedance, Z_R, which in addition is also frequency dependent. This voltage, V_n appears across both branches of the signal transmission cable and induces a differential-mode signal V_o across the load Z_L. If the signal circuitry sensitivity is below V_o then conductive coupling situation results.

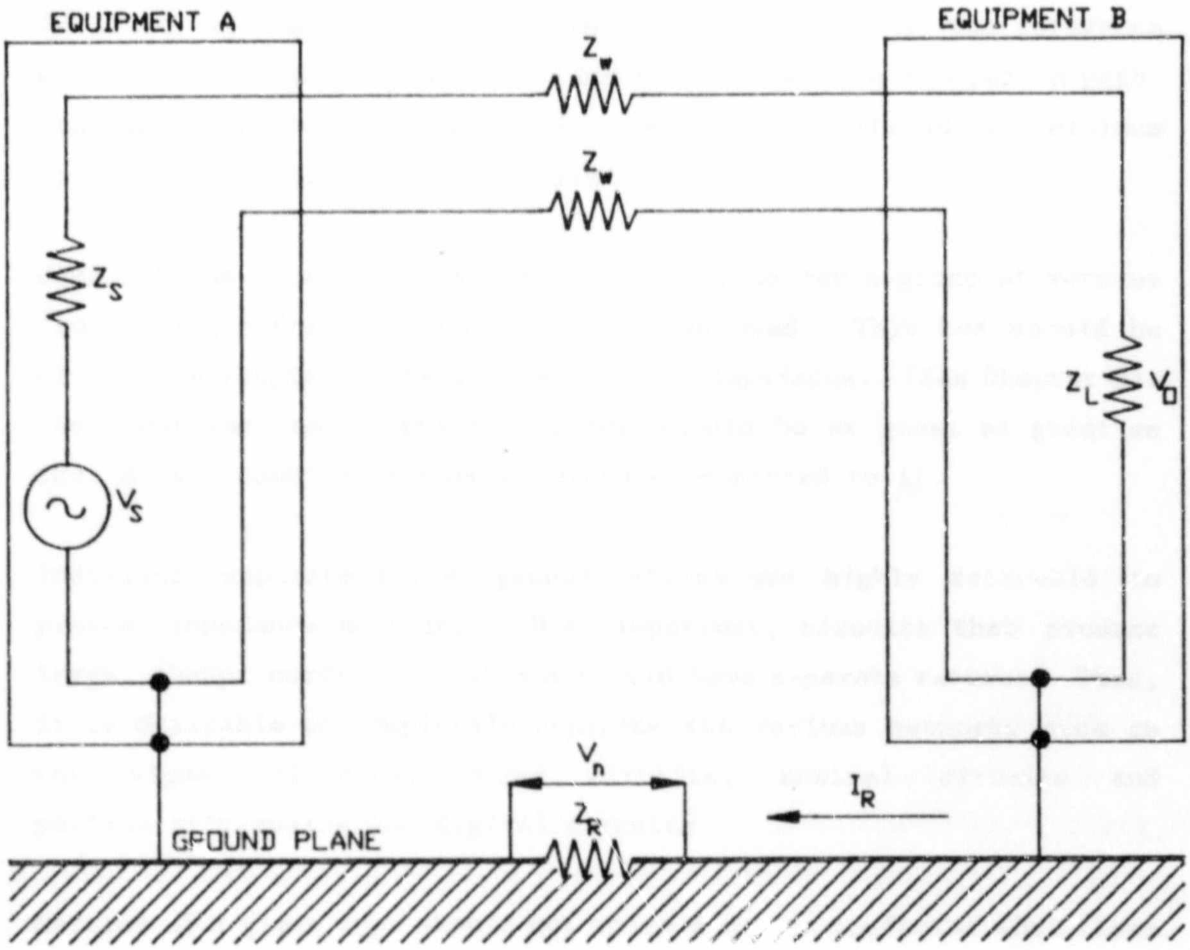

Figure 4.3 Common-Mode Impedance Coupling Between Interconnected Equipment

In an industrial environment, this situation often happens when sensors and data processing equipment are inadvertently grounded to ground references that also carry power currents at 50 Hz. The power current could be several hundred amperes and if the value of Z_R is a few tenths of an ohm, V_n could be tens of volts, often enough to obliterate data signals. Also multiple installed grounds, often at considerable distances from each other, result in inadvertent ground loops. These situations usually occur as a result of inadequate planning for system grounding during the design phase.

A summary of recommended practice for power and signal return circuits is itemised as follows:

1. To avoid impedance coupling minimise the impedance of the reference wire or bus if indeed the wire or bus must be used for the return path. That is, both the resistance and series reactance should be a minimum by using the shortest possible leads.

2. When returns of a given type are combined, as for a group of returns from a single frame, a bus return may be used. This bus should be of minimum length and be flat and of low impedance. (See Chapter 3). The cross section of the return bus should be at least as great as that of the combined return conductors connected to it.

3. Individual separate signal ground returns are highly desirable to prevent impedance coupling. Most important, circuits that produce large, abrupt current variations should have separate returns. Thus, it is desirable to completely separate the various networks such as the signal circuits, power circuits, control circuits and particularly analog and digital circuits.

4. Returns of a given type should not share a common conductor when there is a possibility of circuit coupling.

5. Prevent the source and load ends of the desired signal circuits from being connected to points of different potential, thus preventing the development of ground loops. Conversely, reduce the voltage differential between source and load by lowering the impedance of the path through which interference currents can flow.

6. A chassis or cabinet used as a ground return circuit should be made of metal having high electrical conductivity.

4.4 FREE-SPACE COUPLING

Free-space coupling is the transfer of electromagnetic energy between two or more circuits not directly interconnected with a conductor. Depending on the distance between the circuits, two types of coupling is defined:

- **Far-Field Radiation Coupling** which is caused by radiation of energy by electromagnetic waves; these include free-space separation, signal cabling and grounding systems.

- **Near-Field Coupling** which is due to either inductive (magnetic) and/or capacitive (electrostatic) coupling, according to the nature of the electromagnetic field.

Radiated EMI usually is coupled into network loops bounded by equipment units, interconnecting cables, and the ground plane or safety wires where it can induce both common-mode and differential-mode interference. Some interference transfers can occur over a combination of conducted and radiated paths. Examples are:

a. A source radiates energy into a power cable which conducts it to various circuits that use power.

b. Coupling between two cables involves radiated propagation of energy into and out of the cables.

In this document "radiated coupling" will be used to describe both near field (inductive and capacitive) coupling and far-field coupling. Associated with this coupling is the characteristics of the source. A knowledge of the type of radiator and its characteristics is helpful to the understanding of the radiation coupling mechanism. Two types of sources exists; these are high impedance sources and low impedance sources. Refer to Appendix B for mathematical derivations pertaining to the following extract. Also refer to Figures 4.4 and 4.5.

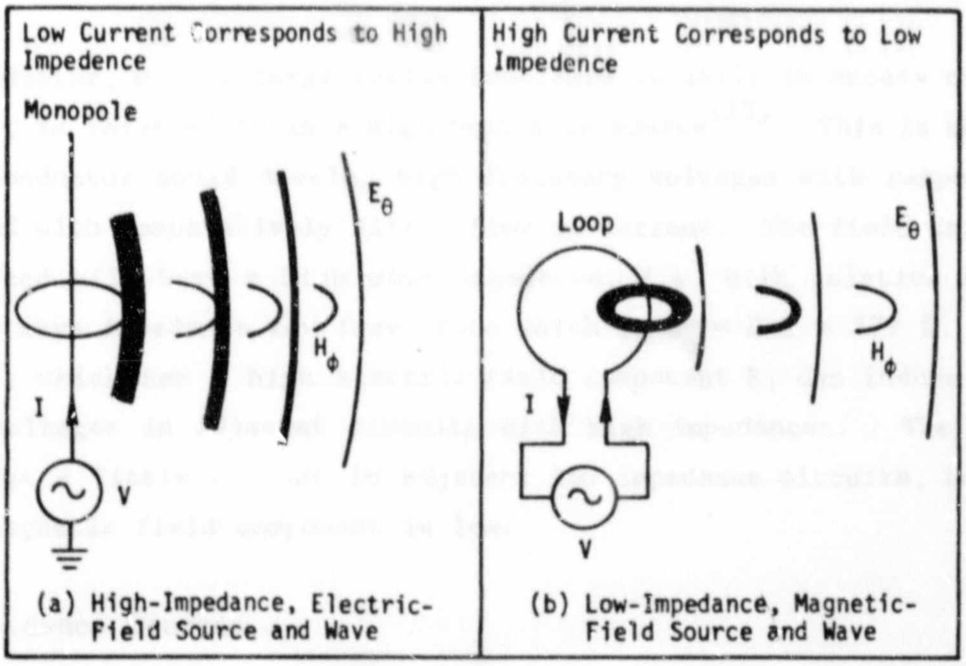

Figure 4.4 Conceptual Illustration of Field Intensities vs Source Type and Distance
(Source: White[(2)])

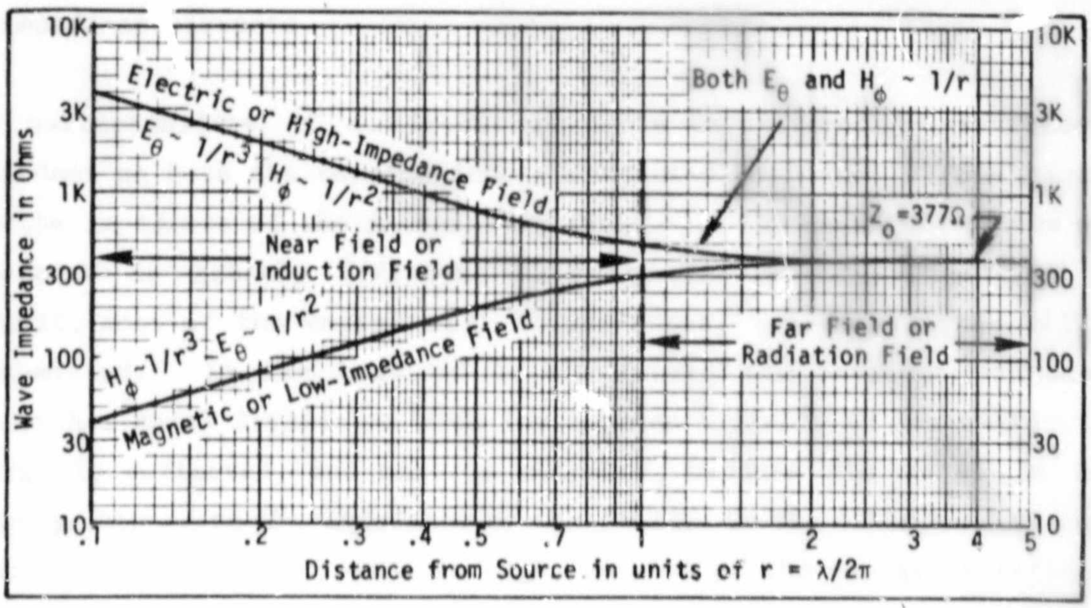

Figure 4.5 Wave Impedance as a Function of Source Distance
(Source: White[(2)])

High Impedance Sources

A conductor, with a large series impedance (usually in excess of 1000 ohms), is referred to as a high impedance source(13). This is because the conductor could develop high frequency voltages with respect to ground with comparatively little flow of current. The field which is produced will have a high wave impedance, i.e. high relative to the plane wave impedance for free space which is $Z_o = E/H = 377\ \Omega$. This field, which has a high electric field component F, can induce large EMI voltages in adjacent circuits with high impedances. The field induces a little current in adjacent low impedance circuits, because the magnetic field component is low.

Low Impedance Sources

If the source contains a large current flow compared to its potential, such as may be generated by a loop, a transformer, or power lines, it is called a current, or magnetic, or low impedance source. Such sources have a large magnetic and a small electric field component. These fields can induce large currents on or in low impedance surfaces or circuits. However, such fields induce little voltage in high impedance circuits.

The susceptibility of a circuit or cable to radiated interference is dependent on both the termination or intrinsic impedance of the circuit and the impedance of the interfering field. If the magnitude of the wave impedance is greatly different from the intrinsic impedance of the circuit, most of the energy will be reflected, and very little will be transmitted. In the case of metals (where the intrinsic impedance approaches zero) irradiated with low impedance fields (H dominant), less energy is reflected, and more is absorbed, because the metal is more closely matched to the impedance of the field. On the other hand the wave impedance of electric fields is high, so most of the energy is reflected for this case.

4.4.1 NEAR-FIELD COUPLING

Two or more wires or other conductors which are close to each other couple energy from one wire to the other wire. This is usually in the form of voltages and currents which are induced by inductive and capacitive coupling mechanisms. The interfering source in this near-field coupling case could be any conductor such as a high level signal line, an ac power line, a control line or even a lightning down-conductor. Such interfering voltages and currents can further be conductively coupled to victim circuits and systems, unless properly grounded.

4.4.1.1 Inductive Coupling (Magnetic Coupling)

Any long, straight current carrying wire has a magnetic field surrounding it and is a means for inductive coupling as shown in Figure 4.6. This situation is essentially a transformer action between the interference source and the sensitive circuit. In addition, dc circuits as well produce changing magnetic fields when current is periodically or intermittently interrupted. Typically, pairs of wires connect transducers in the field with receiving equipment in the control room. If for instance, a high current carrying circuit is placed closed to a low signal level current, such as shown in Figure 4.6, then inductive coupling could exist. Thus, when a current change occurs in one circuit, a changing electromagnetic field through the area of its loop is produced. A voltage will be induced when some of the flux passes through the second circuit. By applying the laws of Faraday and Ampere to this situation, the following expression for the induced voltage in circuit 2 results[3]:

$$V_c = \mu \ell f I \ln (h/s) \qquad (4.1)$$

where

ℓ = length of wire

h = width of victims wire loop

s = distance between circuits

and μ = permeability of the media

f = frequency concerned

I = current in circuit 1.

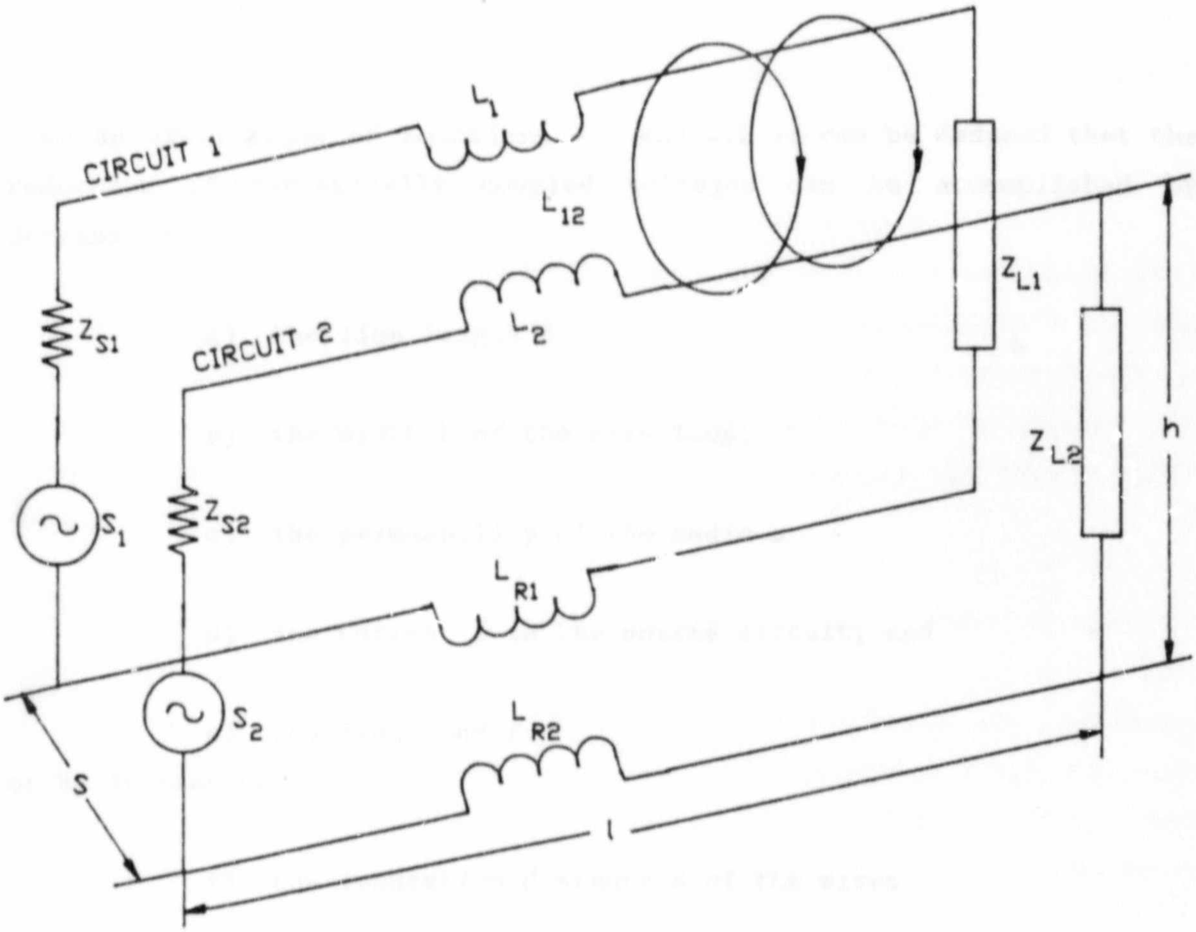

Figure 4.6 Illustration of Inductive Coupling

Incorporating the influence of the self-inductance of the circuits (the mutual inductance of the circuits have been accounted for in Equation 4.1) then the overall induced voltage V_i in circuit 2 is given by

$$V_i = \left(\frac{Z_{L2}}{Z_{S2}+Z_{L2}+j\omega L_2} \right) V_c \qquad (4.2)$$

where

Z_{L2} = load impedance of circuit 2

Z_{S1} = source impedance of circuit 2

L_2 = self-inductance of circuit 2.

The transfer impedance Z_T may be calculated from

$$Z_T = \frac{V_i}{I} \tag{4.3}$$

From an examination of Equations 4.1 and 4.2 it can be deduced that the reduction of magnetically coupled voltages can be accomplished by decreasing

a) the line length ℓ

b) the width h of the wire loop;

c) the permeability of the media μ

d) the current I in the source circuit; and

e) the frequency f.

or by increasing

f) the separation distance s of the wires

g) the source impedance Z_{S2} and

$$\omega L_2 << Z_{S2} + Z_{L2} \text{ ; and}$$

h) the self-inductance of the victim circuit 2 (L_2), where

$$\omega L_2 >> Z_{S2} + Z_{L2}.$$

Methods (a) and (b) can be accomplished by reducing the area enclosed by either the transmitting loop or the receiving loop. This is achieved in principle by using twisted-pair wiring or coaxial cables. In the case of twisted wires, effective inductive decoupling from a practical standpoint may not truly exist because the two wires are not truly spiralled around each other and also because of unequal wire lengths and impedances. However, field tests have shown that cancellation effects of 25 dB (20:1) are possible due to careful uniform twisting; the effectiveness depends on the number of twists per metre and the symmetry of the twisting. Forty to sixty twists per metre result in good noise cancellation[22,13]. However, when manufacturing and economic considerations are taken into account, such figures may be impractical. A minimum value of 14 twists per metre[4] can be adopted if the magnetic ambient level is low.

Reduction of loop area is, however, not always easy to accomplish. In particular, it should be noted that use of a structure (chassis) as a return path may lead to large loop areas. This point is significant when signal grounding is considered in large systems (See Chapters 5 and 7). In general, such loop areas may be minimised by running the positive or live lead directly against the structure and using the shortest possible length of loop.

Method (f) is probably the most obvious and easy means of reducing coupling on a plant. Separating wires that carry interfering currents from those connected to susceptible circuits should be considered standard practice. A helpful factor is that induced voltage decreases in an exponential manner with increasing wire separation. For a more detailed guide to recommended practice and design aids refer to IEEE Guide 518-1982[30].

Figure 4.7 graphically illustrates these techniques and shows the evolution of reducing the coupling as discussed above and also considers the related grounding practice.
Methods (d), (e), (g) and (h) should be considered early in the design stage. Also right angled crossing can help to reduce coupling, although in practice it may not always be possible to achieve.

A common practice to reduce coupling is to shield wires. This is more difficult to implement in the case of magnetic fields than is at first apparent. This is so because shields of 85 percent effectiveness offer adequate electric field shielding but are relatively ineffective against magnetic field interferences.

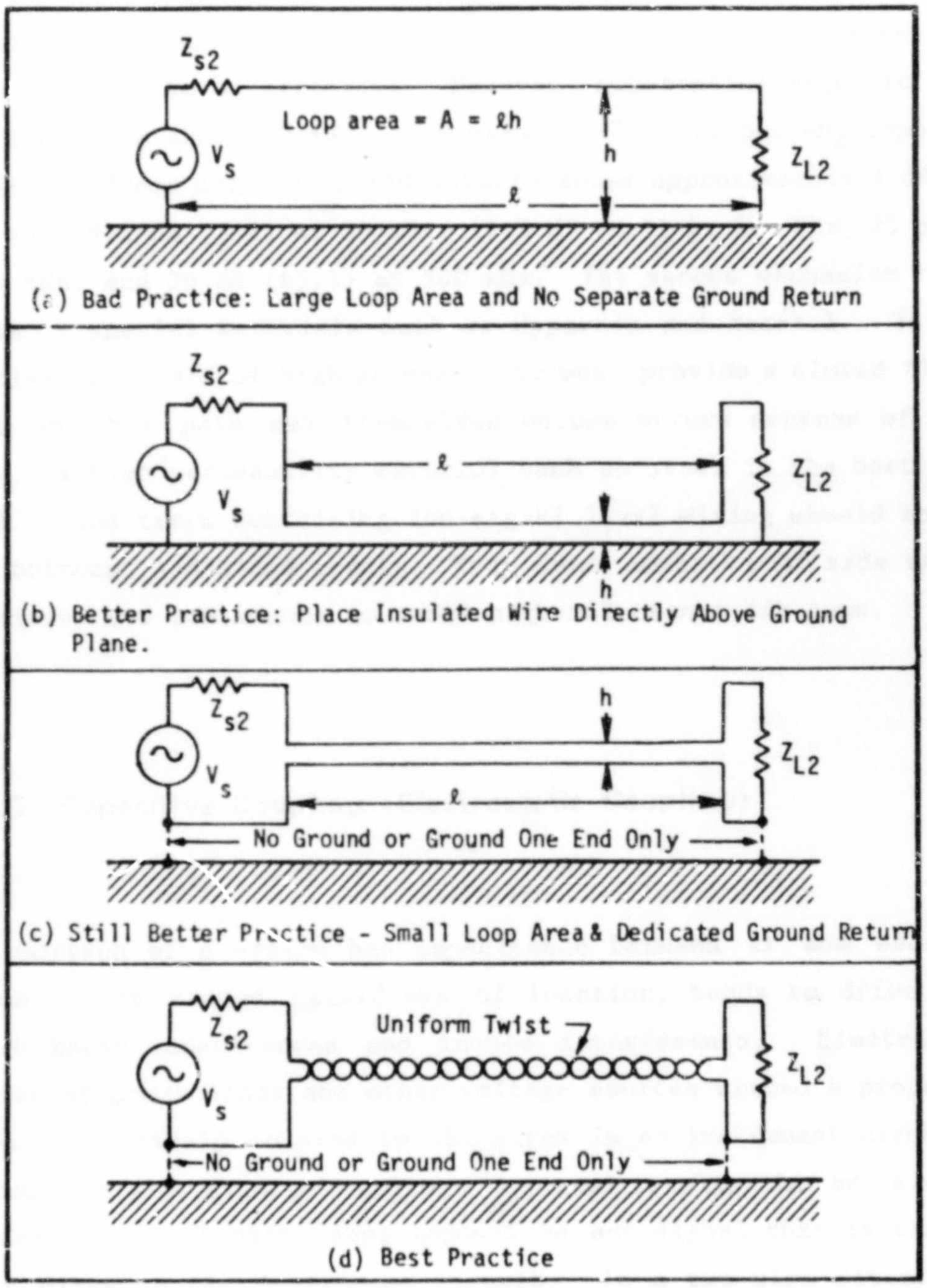

Figure 4.7 Evolution of Reducing Loop Area and Circuit Grounding (Source: White[2])

There are two primary mechanisms by which shielding reduces coupling. In the one case a shield of high conductivity functions by developing a counter flux due to eddy currents exited within the shield. The second mechanism is a shield of high permeability which functions by providing a flux path of low reluctance for the interference field. The first shielding mechanism becomes impracticable against magnetic fields under frequencies of approximately 10 kHz because very thick materials are required for it to be effective. However, substantial magnetic coupling reduction occur above 10 kHz. For example[22], 100 percent copper braid shielding reduces magnetic field interferences approximately 3 dB (1.5:1) at 10 kHz, 9.5 dB (3:1) at 50 kHz, 15.5 dB (6:1) at 100 kHz, 25 dB (18:1) at 200 kHz, and 36 dB (65:1) at 500 kHz. The second mechanism calls for the use of special materials such as Hypernik and Mumetal. To be most effective, a shield of high permeability must provide a closed flux path. Breaks in this path may themselves become severe sources of magnetic fields. A high permeability material such as steel is the best magnetic shield. Thus trays containing low-signal level wiring should have solid metal bottoms, sides and covers. Tray cover contact with side rails must be positive and continuous to avoid high-reluctance air-gaps.

4.4.1.2 Capacitive Coupling (Electrostatic Coupling)

Every portion of a system has capacitance between it and every other portion. Any change regardless of location, tends to drive currents through these capacitances and induce interference. Electric fields radiated by power lines and other voltage sources around a process plant can be capacitively coupled to the wires in an instrument circuit. The coupling to the external voltage sources results in an alternating interfering signal being superimposed on any signal that is transmitted on the wires in the instrument circuit. In a two wire situation, the capacitance between conductors is directly proportional to the distance which the conductors are run adjacent to each other and inversely proportional to the distance between the conductors.

Mathematical relations can be derived for specific situations, but these equations are of academic importance only, for it must be realised that all objects have capacitance to all other objects which result at times in unexpected 'sneak' circuits. However, the effect of capacitive coupling can be considered by examining a typical two wire situation, such as shown in Figure 4.8, where typical capacitive paths are illustrated.

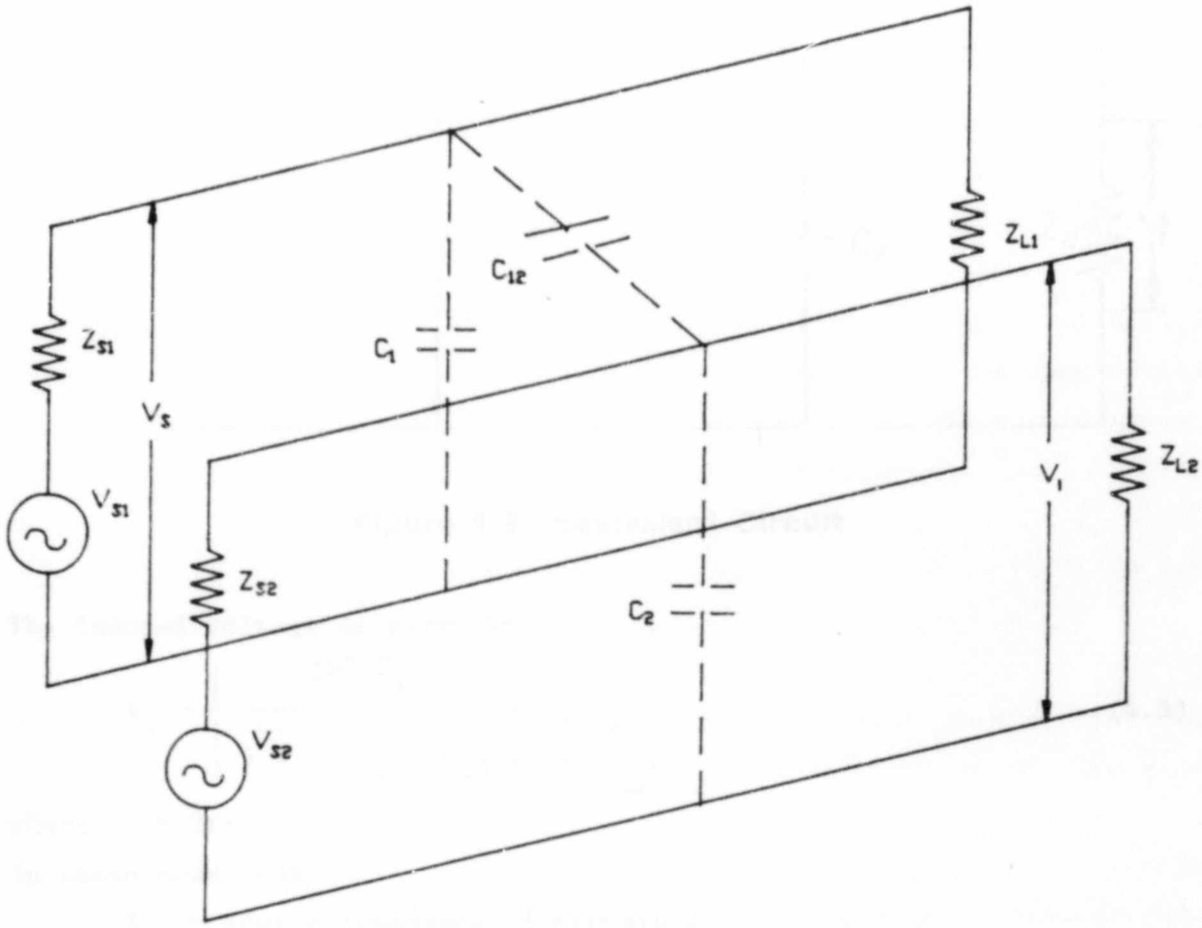

Figure 4.8 Illustration of Capacitive Coupling

The equivalent circuit for Figure 4.8 is given in Figure 4.9 where the parallel combination of Z_{S2} and Z_{L2} has been replaced by the equivalent impedance.

$$Z_2 = \frac{Z_{S2} Z_{L2}}{Z_{S2} + Z_{L2}} \tag{4.4}$$

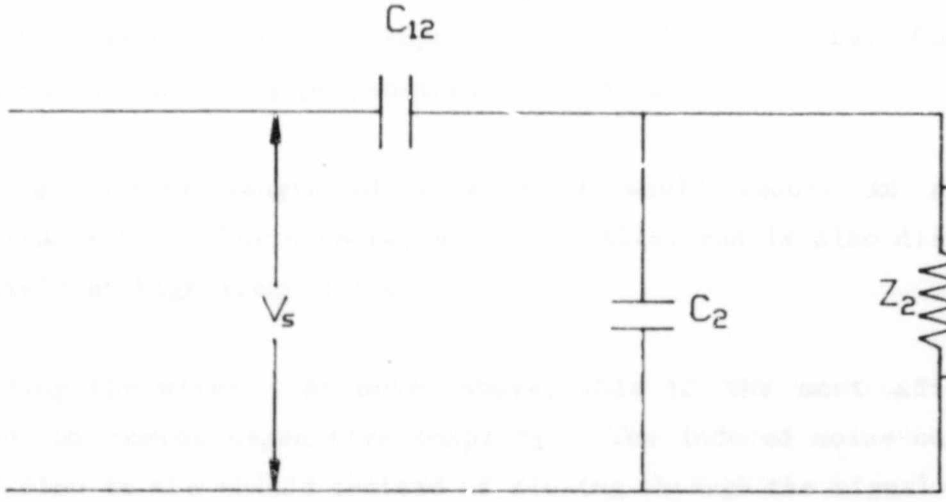

Figure 4.9 Equivalent Circuit

The induced voltage is given by

$$V_i = \left[\frac{j\omega Z_2 C_{12}}{1 + j\omega (C_2 + C_{12}) Z_2} \right] V_S \qquad (4.5)$$

where $\omega = 2\pi f$

In these equations

Z_{S2} = source impedance of circuit 2

Z_{L2} = load impedance of circuit 2

C_{12} = capacitance between the wires

C_2 = capacitance of wire to ground.

Inspection of Equations 4.4 and 4.5 shows that the coupling can be reduced by

a. reducing the source voltage, V_{S1} and/or

b. reducing the inter-wire capacitance, C_{12}.

These can be accomplished by:

1. Use of filters at the source or at the susceptible circuit. It is generally preferable to filter the source side if possible.

2. Increasing the spacing between the wires thus reducing the mutual coupling capacitance C_{12}. Thus, isolate interference source leads from those recognised as pickups for susceptible circuits. Coupling can also be reduced by perpendicular crossing.

3. Using a shorter length of wire which would result in reduced capacitance C_{12}. Furthermore, a long parallel run is also difficult to shield at high frequencies.

4. Shielding the wires. As noted above, this is the most effective method to combat capacitive coupling. The induced noise-currents would flow in the shield instead of flowing through the signal wires. Refer to Chapter 5 for shield grounding considerations.

5. Twisting of the leads which tends to balance capacitive coupling and thus reduce noise levels.

6. Use of balanced lines and balanced circuits. This is effective for sensitive equipment when both wires are maintained at the same impedance (See Chapter 5).

Other factors which influence the interference magnitude voltage are the source and load impedance of the victim circuit and the frequency. The induced voltage will be lower for low source and load impedance. However, the coupling between the two circuits will increase as the frequency increases, since the reactance of the coupling capacity will increase with frequency.

4.4.2 FAR-FIELD RADIATED COUPLING

The production of high frequency, high energy signals is necessary for the transfer of energy between systems in the communications field, where the technology is well developed and understood. However, not so clear is the fact that conductors in an electronic system could radiate energy which escapes and propagates into space. Thus, any current carrying conductor such as a signal line, a power line, or even a ground conductor, does not have to be specifically designed to be able to radiate energy.

These unintentional radiated high-frequency interfering signals may be modulated by the ac power frequency, its harmonics and thus may be troublesome if the radiation is picked-up and demodulated by the sensitive electronic circuits. This pickup and demodulation can produce spurious signals having the ac power frequency as a potential interference to the instrumentation system. Two forms of radiated coupling have been identified and the potential effects of both the coupling paths are examined below.

4.4.2.1 Common-Mode Radiation Coupling

Figure 4.10 is a simplified circuit diagram to illustrate common-mode radiation coupling into circuits. The common signal/load configuration is interconnected by a transmission cable with a cable resistance represented by Z_w. The shaded area defines a close loop formed by the ground plane the signal and load equipment boxes, and the transmission cables. Radiated electromagnetic plane waves incident upon the circuit induces a voltage V_i around the close path on the perimeter of the shaded area. The spatial orientation of the plane of the loop with respect to the electric and magnetic fields of the incident wave determines the amplitude of the voltage induced around the loop.

The induced voltage is indicated by an equivalent source V_i. If it is assumed, as is usually the case, that the transmission cables are closely spaced, two paths for current flow can be identified: ABCDEFA,and ABC'D'EFA. The circuit shows both the signal reference plane (point B) and load reference plane (point E) connected to the ground plane at points A and F respectively.

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