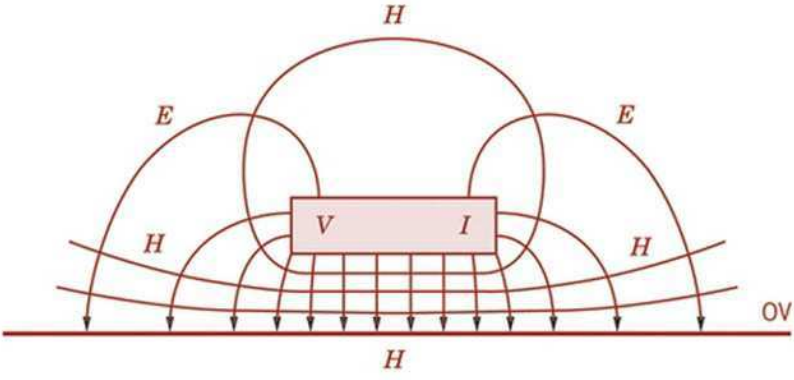
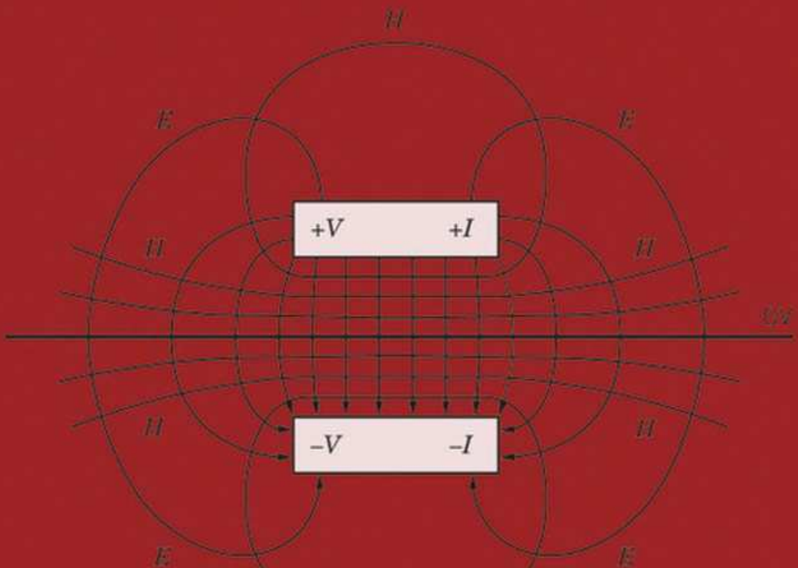
FIFTH EDITION



GROUNDING AND SHIELDING

CIRCUITS AND INTERFERENCE



RALPH MORRISON

GROUNDING AND SHIELDING

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GROUNDING AND SHIELDING

CIRCUITS AND INTERFERENCE

FIFTH EDITION

Ralph Morrison

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PREFACE TO THE FIFTH EDITION

The first edition of *Grounding and Shielding* was published in 1967. The fourth edition was published in 1997. It is hard to imagine that another decade has passed and that I would be writing on the same subject for a fifth time. This fifth edition is a total rewrite because I wanted to present ideas in a different order and also to present new insight that was gained through teaching seminars and consulting. For example, I wanted to add material on printed circuit board design, a subject that was only touched upon in the fourth edition. I also left out material that was no longer relevant.

This book is intended for those interested in the real world of electronics. The new engineer can easily read this material although he or she may not recognize the importance of the approach or why some material is stressed. Engineers and circuit designers in the analog, digital, or power world who have seen a few problems will appreciate the need to understand interference and all it represents. There is some mathematics involved but it is not necessary that every equation be understood to get a great deal from the content of the book.

I have spent a career in electronics that started with vacuum tubes. I remember the first silicon diodes and the first transistors. I also remember thinking that 10 MHz would be impossible using solid sate devices. So much for impossibilities. Today, logic designers are considering clock speeds of 24 GHz. This time I believe it is possible.

In my early years I worked on handling low-level analog signals and I found out the importance of my physics background. That is what led me to write the first edition. Today, when I see the problems of digital design or facility layout I realize again how important physics is in understanding all electrical design. This is the reason I am writing a fifth edition. There are many new problem areas that need to be discussed. The book starts out with some very elementary physics. I hope the reader does not skip this material because it is at the heart of what is to follow.

There are several ideas that I want to emphasize in this book. The first idea is that a schematic or circuit diagram is only a rough plan or outline. It provides a basis for circuit analysis but it cannot present any geometric information. For example, component size and orientation are missing as are lead lengths and lead dress. As circuits get more complex this geometry is critical. The second idea I want to present is that all components are field operated. A capacitor stores electric field energy and an inductor stores magnetic field energy. A transistor requires a field in the semiconductor material. A transformer works by using both electric and magnetic fields. I want to show that the conductors in our circuits are there to place these fields into the various components. Third, I want to teach that a field in a component means the presence of field energy. Moving this energy around is the purpose of the conductors we use in our circuits. Trying to do this at a GHz is not simple. Understanding transmission lines is key to understanding how energy is moved in all circuits from dc to 100 GHz. Fourth, I want to teach that good engineering is a compromise. There are no perfect solutions. The issue is to solve the problem with available material in an economical manner. The problem is often that we expect too much from circuit symbols. These symbols are really just a starting point in design.

I want to thank the people that were helpful in making this book possible. My wife Elizabeth is a writer in her own right and I have gained from her many comments. I thank my editor George Telecki, who recognized the need for this book and offered me encouragement at every step along the way. The good words put in by Henry Ott and Rick Hartley gave important support. I would like to thank Dan Beeker of Freescale for his help and assistance. He recognizes the need for engineers to continue their education so that they can best use what technology has to offer.

My hope is that this book will provide the reader with a useful viewpoint, one that will solve problems. I hope that the approach taken in this book can find its way into the educational system where I feel it is strongly needed. I further wish that readers who agree with me will work with academia to make the needed changes.

My many thanks to all of you who have bought the early editions of this book. I trust you can gain by reading this fifth edition.

Pacifica, CA RALPH MORRISON

CHAPTER 1

Voltage and Capacitance

1.1. INTRODUCTION

How does a circuit work? One answer is to do a sinusoidal analysis using Kirchhoff's laws. Another answer is to write a set of logic statements. These responses provide a small part of the answer. The full answer is buried in a mountain of details. In this book we are going to look at some of this detail but in a non-circuit way. We will take this new approach because circuit diagrams and circuit theory by their very nature must leave out a lot of pertinent detail. This detail is important for performance at high frequencies or for performance involving very small signals. It is also important when radiation, interference, or susceptibility are involved. Wire size, connection sequences, component orientation, and lead dress are often critical details. I like to call these details "circuit geometry." These details in geometry are closely related to how well a circuit works. Geometry is an issue in analog circuits, power circuits, and especially in digital circuits where clock rates are always rising.

When a circuit is put to practice there are many details that we take for granted. The components will most likely be connected together by strips or cylinders of copper. They will be soldered into eyelets or onto copper pads. Traces will go between layers on a printed circuit board using *vias*. These are the details in a design that are not questioned. There are details of a more subtle nature such as the thickness of a trace or ground plane or the dielectric constant of an epoxy board. In most cases we do not question how things are done because we tend to rely on accepted practice. Circuits built this way in the past have worked, so why make changes?

Taking things for granted is not always good engineering. Note that digital clock rates have changed from 1 MHz to 1 GHz in 20 years. That is three orders of magnitude! Imagine what would happen if automobile speeds went up one order of magnitude. That is 600 mph or jet aircraft speed. Even a modest increase in automobile speed would require extensive changes to the design of our roads and cities, not to mention extensive driver training.

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

In electronics, an increase in speed does not pose a safety hazard. There are, however, differences in performance that should be understood. Often an effect is not sensed until the next generation in design is introduced. Understanding these effects requires an understanding of basic principles. The details we will look into do not appear on a circuit or schematic diagram. We will address many of these details that affect performance.

Electronics often makes use of power from the local utility. For reasons of safety the utility must earth one of the power conductors. Electronic hardware must interface with this power and also share this same earth connection. The result is often interference. I will discuss the relation between power distribution and circuit performance in later chapters.

A circuit diagram is only a plan or an organization of ideas. Circuit theory provides a basic overview of circuit performance. Circuit symbols are a part of the problem. They are necessarily very simple representations of complex objects. Every capacitor has a series resistance and an inductance and every inductor has series resistance and shunt capacitance. These considerations only begin to tell the entire story. For example, at high frequencies, dielectrics are nonlinear. For magnetic materials, permeability falls off with frequency. Thus circuit symbols can only convey limited information. Further, we do not have symbols for skin effect, transit time, radiation, or current flow patterns. A straight line on a diagram may actually be a very complex path in the actual circuit. In short, a schematic diagram provides little information on physical structure and this can limit our appreciation of what is actually happening.

The performance of many circuits or systems is closely related to how they are built. This is true for analog circuits as well as computers. I have often commented that it is not a question of whether there is an electrical connection but where the connection is to be made. In an analog circuit it is often important to know which end of a shield is grounded, not whether it is grounded. Here is good question! How should a circuit board ground plane be connected to the surrounding chassis? The answer to this question is not available from a schematic diagram. We will discuss this problem in the digital circuits chapter (Chapter 7).

A repeated theme discussed in this book relates to how signals and power are transported in circuits. This approach will lead to an understanding of many issues that are often poorly understood. In order to discuss the transport of electrical power and signals, the electric and magnetic fields must be discussed. To begin this discussion we introduce the electron. Don't despair. The time spent reviewing this area of physics will make it much easier to understand the ideas presented later in this book.

1.2. CHARGES AND ELECTRONS

Circuit theory allows us to relate circuit voltages and the flow of current in a group of interconnected components. For RLC networks (Resistor-

Inductor-Capacitor) this analysis is straightforward using Kirkkhoff's laws. The processes I want to discuss do not involve this approach. To understand the more subtle aspects of circuit performance we will use basic physics to explain many details that are often left to chance. Our starting point may seem a bit remote, but please read on. We first discuss the atom.

Atoms are composed of a nucleus of protons and neutrons surrounded by shells of electrons. The electrons have a negative charge and the matching protons in the nucleus have a positive charge. In a neutral atom the positive and negative charges are exactly equal. Each electronic shell is limited to a fixed number of electrons. The number of electrons in the outer shell says a lot about the character of the atom. As an example, copper has just one electron in its outer shell. This outer electron has a great deal of mobility and is involved in electrical activity. Because protons are comparatively heavy and because the shells of electrons shield them, they are not directly involved in the electronics we are going to consider.

Molecules are formed from atoms that bond together and share outer shell electrons. For an insulator, this bonding limits the outer electron mobility. Typical insulators might be nylon, air, epoxy, or glass. If two insulators are rubbed together, such as a silk cloth and a rubber wand, the difference in the mobility of the outer electrons will allow the transfer of a few electrons from the rod to the cloth. In this case, the silk cloth with extra electrons is called a negatively charged body. The rod is said to be positively charged. We will call the absence of negative charge a positive charge. In reality the positive charge stems from the immobile protons in the nucleus of atoms that do not have matching outer shell electrons. The absence of negative charges is the same as if there were fictitious positive charges on the surface of the insulator.

Experiments with charged bodies can demonstrate the nature of the forces that exist between charges (electrons). If one charged body repels another it is actually the fields of electrons that are involved. If you remember your physics class, these forces can be demonstrated using pith balls that hang by a string. Here the charges are attached to small masses and we can see the pith balls attract or repel each other.

The number of electrons involved in any of these experiments is extremely small. To illustrate this point I want to paraphrase the writing of Dr. Richard Feynman.[1] If two people are standing a few feet apart, what would be the force of repulsion if 1% of the electrons in each body were to repel each other? Would it be a few pounds? More! Would it be greater than their weight? More! Would it lift a building? More! Would it lift a mountain? The answer is astounding. The force would be great enough to lift the earth out of orbit. This is why gravity is called a weak force and the force between electrons is called a strong force. This also tells us something about nature. The percentage

[1] *The Feynman Lectures on Physics*, Volume 2, p. 1-1. Addison-Wesley Publishing Company, Inc. Copyright 1964, California Institute of Technology.

of electrons involved in electrical activity is extremely small. We know that the forces in a circuit do not move the components or the traces. Obviously, since electrical forces are so large, electrical activity involves a very small percentage of the available electrons.

1.3. THE ELECTRIC FORCE FIELD

When we encounter forces at a distance we us the expression *force field*. We experience a force field at all times as we live in the gravitational force field of the earth. Every mass has a force field, including the earth. The earth has the dominant field because the earth is so massive. The result is that each mass on earth is attracted toward the center of the earth. The attraction forces between individual objects on the earth are so small that they are very difficult to measure. On the earth's surface the force field is nearly constant. We would have to go out into space to see a reduction in the force of gravity.

The electrical force field is similar in many ways. Every electron carries with it an associated force field. This force field repels every other electron in the area. If a group of extra electrons are located on an isolated mass we call this mass a *charged body*. We refer to the extra electrons as a *charge*. If this mass is a conductor, the extra electrons will move apart until there is a balance of forces. On a conducting isolated sphere the electrons will move until they are evenly spaced over the entire outer surface. None of these excess electrons will remain on the inside the conductor. For a perfect insulator any extra electrons are not free to move about. Extra electrons on the inside of this material are called *trapped electrons*. It is also possible to have trapped absences of electrons. We can call these *holes*.

1.4. FIELD REPRESENTATIONS

The electric force field in a volume of space can be measured by placing a small test charge in that space. A test charge can be formed using a small mass with a small excess of electrons on its surface. The force on this test charge has a magnitude and direction at each point in space. Having direction the force field is called a vector field. To be effective this test charge must be small enough so that it does not influence the charge distribution on the objects being measured. Performing this experiment is difficult, but fortunately we can deduce the field pattern without performing an actual test.

It is convenient to represent a force field by lines that follow the direction of the force. For an isolated conducting charged sphere, the lines of force are shown in Figure 1.1.

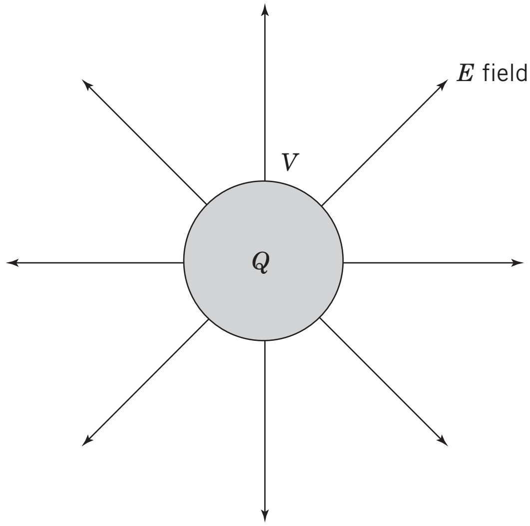


Figure 1.1. The force field lines around a positively charged conducting sphere

Note that the field exists everywhere between the lines. The lines are simply a way of showing the flow or shape of the field. In any practical example, the number of extra electrons that form the charge Q on the surface of a conductor is small compared to the number of electrons in the conductor. For a practical surface charge, the number of electrons is still so large that we can consider the charge as being continuously distributed over the surface of interest. This is the reason we will not consider the force field as resulting from individual electrons. From here on we will consider all charge distributions as being continuous. The total charge on the surface of the sphere in Figure 1.1 is Q. The charge density on the surface of the sphere is

$$\frac{Q}{\mathrm{A}} = \frac{Q}{4\pi r^2}. \tag{1.1}$$

We will use the convention that a line starts on a unit of positive charge and terminates on a unit of negative charge. This unit can be selected so that the graphical representation of the field is useful. If the total charge is doubled, then the number of lines is doubled. For representations in this book, no attempt will be made to relate the number of lines to any specific amount of charge. In general we are interested in the shape of the field, areas of field concentration, and where the field lines terminate.

In Figure 1.1 the force on a small test charge q in the field of Q is proportional to the product of the two charges and inverse to the square of radius r or

$$f = \frac{qQ}{4\pi\varepsilon_o r^2} = \frac{qQ}{\varepsilon_o \mathrm{A}} \tag{1.2}$$

where A is the area of the sphere. The constant ε_o is the permittivity of free space. Equation (1.2) is known as Coulomb's law. The force per unit charge or f/q is a measure of the electric field intensity. The letter E is used for this measure. The force field around a group of charges is referred to as an E field. Mathematically the E field around a charge is

$$E = \frac{Q}{4\pi\varepsilon_o r^2}. \tag{1.3}$$

The E field falls off as the square of the distance r. Equation (1.3) could also be written as

$$E = \frac{Q}{\varepsilon_o \mathrm{A}} \tag{1.4}$$

where A is the surface area of the sphere at the distance r. In Figure 1.1, the force field intensity E decreases as the field lines diverge. The forces are greatest at the surface of the sphere. Note that the field lines do not enter the sphere. This is because there are no excess charges inside the conductor. The field lines must terminate on the sphere perpendicular to its surface. If there were a tangential component of force on the surface the charges on the surface would be accelerated. If there were an absence of electrons on the surface, this absence of charge would also be accelerated. Remember the absence of negative charge can be considered the presence of a positive charge. For conductors, the mobility of a group of electrons is no different from the mobility of an absence of electrons. Except for the direction assigned to the force field, we will assume that positive and negative charges behave the same way. Figure 1.1 shows a sphere with a positive charge Q. If the charge were negative (the presence of electrons), the field lines would be shown with the arrows pointing inward.

The field lines in Figure 1.1 start at the surface of the sphere. If the charge Q were located at the center of the sphere and the sphere were removed, the field pattern at every value of r would be unchanged. A point charge Q implies an infinite charge density, which is impossible. Often it is mathematically convenient to consider the fields from point charges even though this can't exist.

1.5. THE DEFINITION OF VOLTAGE

A test charge q in the field of a charge Q experiences a force given by Eq. (1.2). The work required to move the test charge a small distance Δd is $f(\Delta d)$.

The work to move it from infinity to a point r_1 is the integral of force times distance from infinity to r_1. If we follow one of the field lines, the force is always tangent to the field lines. The work is

$$W = \int_{\infty}^{r_1} f \cdot dr = -\frac{qQ}{4\pi\varepsilon_o r_1}. \tag{1.5}$$

If we divide both sides of this equation by q we obtain the work per unit charge. This term has the familiar name *volts*. In equation form the voltage is given by

$$V = -\frac{Q}{4\pi\varepsilon_o r}. \tag{1.6}$$

DEFINITION

Voltage difference. The work required to move a unit charge between two points in space in an electric field.

In Eq. (1.5) we can make the assumption that the voltage at infinity is zero. This allows us to assign a voltage to points in space. In a circuit, the work required to move a unit charge between two conducting surfaces is called a potential difference or a voltage difference. It is important to realize that potential differences do exist between points in space. Of course it is difficult to place a voltmeter in space to get a measure of this voltage.

The voltage difference between two points in space is

$$V_2 - V_1 = \frac{1}{4\pi\varepsilon_o}\left(\frac{1}{r_2} - \frac{1}{r_1}\right). \tag{1.7}$$

N.B.

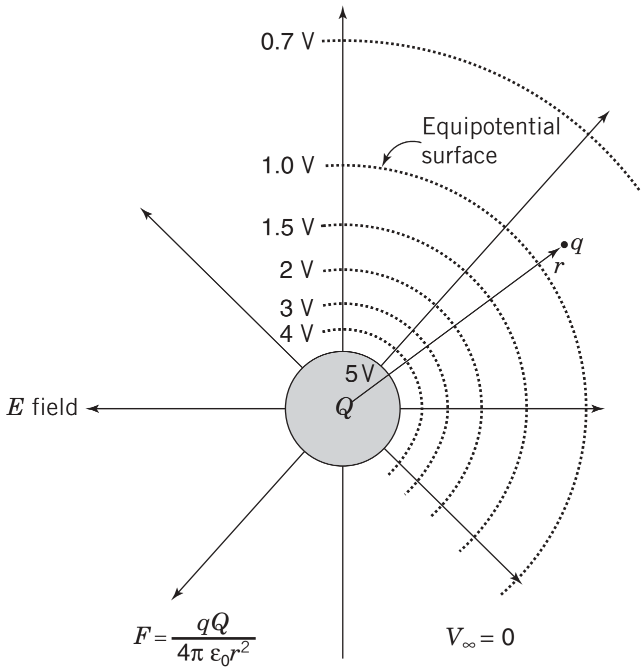
A voltage difference cannot exist without the presence of an electric field.

In the presence of conductors, an electric field cannot exist without charges on the surface of these conductors. These charges are not apparent from a schematic diagram. When a circuit is in operation, there are surface charges everywhere there are voltage differences. These charges are independent of the moving charges in the circuit that we call *current*.

1.6. EQUIPOTENTIAL SURFACES

As the word implies, an equipotential surface is a surface of equal voltage. No work is required to move a test charge on this surface. Figure 1.2 shows equipotential surfaces around the charged sphere in Figure 1.1. Note that these surfaces are also spheres and the surfaces are always perpendicular to the field lines.

> **N.B.**
>
> Conducting surfaces are equipotential surfaces regardless of their shape. This assumes that the surface charges are not in motion.



Q is the charge on the sphere
q is small unit test charge
ε_0 is the dielectric constant of free space (permittivity)

Figure 1.2. Equipotential surfaces around a charged sphere

In practice, a conducting surface is an equipotential surface even when the charge distribution is not uniform. This is true even when there are areas of positive and negative charge on the same conductor. It takes no work to move charges on the surface of a conductor. If work were required there would be a tangential electric field and this means that free charges would have to be in motion.

1.7. THE ELECTRIC FORCE FIELD BETWEEN TWO CONDUCTING PLATES

Consider two conducting plates separated by a distance h. On the top plate there is a charge Q and on the bottom plate there is a charge $-Q$. This configuration is shown in Figure 1.3.

If we ignore edge effects, the force field can be represented by equally spaced straight lines that run from the top plate to the bottom plate. In this configuration the net charge in the system is zero. There is no loss in generality if we assume that all of the field lines stay in the volume between the two plates. Since the lines do not diverge, the force on a test charge q is constant everywhere between the plates. In other words, the electric field intensity E is a constant between the plates. If the charge density on the plates is made equal to the charge density of the sphere in Figure 1.1, the force field between the plates will have the same intensity as the force field at the surface of the sphere. The work required to move a unit charge between the plates will then be force times distance or

$$W = \frac{Q}{4\pi\varepsilon_o r^2} \cdot h = Eh. \tag{1.8}$$

This work W is the potential difference between the plates. If the bottom plate is assumed to be at 0 volts, the upper plate will be at a voltage $V = Eh$. Note

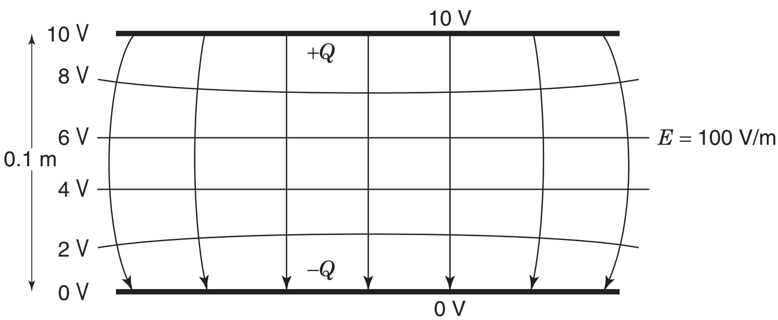


Figure 1.3. The force field between two conducting plates with equal and opposite charges. Some edge effects are shown.

the E field has units of volts per meter. If the voltage between the two planes of Figure 1.2 is 5 volts and the spacing in meters is 1 cm, the E field intensity between the plates is 500 volts per meter.

1.8. ELECTRIC FIELD PATTERNS

Figure 1.4 shows a printed circuit trace over a conducting surface. Such a surface is often called a ground plane. There are many types of ground planes and we will talk about this later. The spacing between these two conductors might be as small as 0.005 inches or 1.3×10^{-4} meters. A typical logic voltage might be 5 volts. The E field intensity under the trace would be 38,000 volts per meter, a very surprising figure.

The E field lines terminate and concentrate on the surfaces between the trace and the ground plane. Since field lines terminate on charges it is obvious where the surface charges are located. Remember this is a static situation. Note this surface charge distribution is not considered in circuit theory. The path taken by the charges to achieve this distribution is also not considered.

FACTS

- The charge distribution on the surface of the conductors is not uniform.
- There is no potential gradient along the ground plane as the charges are not moving.
- The charges concentrate at the interface between the circuit trace and the ground plane.
- There is some E field above the circuit trace.
- Surface charges concentrate on the sharp edges of the circuit trace.
- If the voltage were to change slowly, new charges must arrive by moving on the ground plane.

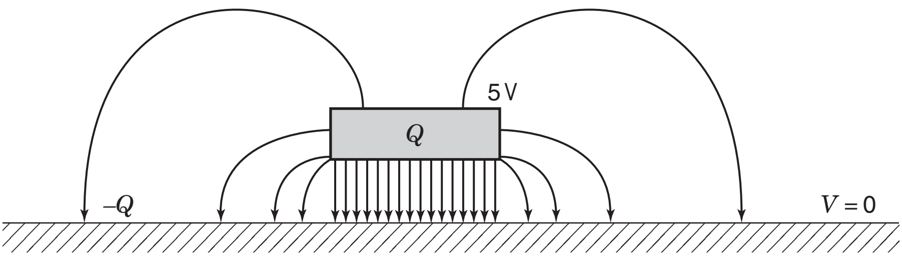


Figure 1.4. The electric field pattern of a circuit trace over a ground plane

Consider the field pattern when there are two traces over a ground plane. This pattern is shown in Figure 1.5. If the voltages are of opposite polarity, the charge distribution on the ground plane will reverse polarity under the traces. Again, the ground plane is an equipotential surface.

Consider the field pattern around a section of shielded cable as in Figure 1.6. In Figure 1.6a the shield S fully encloses the center conductor A and no electric field escapes. In Figures 1.6b and c there is a hole in the shield that allows some of the electric field to terminate on conductor B. The field lines that terminate on conductor B imply a charge distribution on conductor B. In Figure 1.6c conductor B floats in space. Note there is still a charge distribution on this conductor but the net charge on the conductor is zero. This floating conductor would not be at zero potential but it would still be an equipotential surface. The grounded conductor by definition would be at zero potential over its entire surface even though there is a net charge on its surface.

If the voltage on the shielded conductor in Figure 1.6 is slowly changed, the field intensity changes everywhere. The amount of charge on conductor B in Figure 1.6b must now change. This change in charge must flow in the connection to the ground plane. The charge on conductor B is called an *induced charge*. Any current that flows to conductor B is called an *induced current*. In the case of Figure 1.6c, the charge density on the surface of the floating conductor must change. This means that induced currents must flow locally on this conductor. Note there is no path for new charges to reach the isolated conductor.

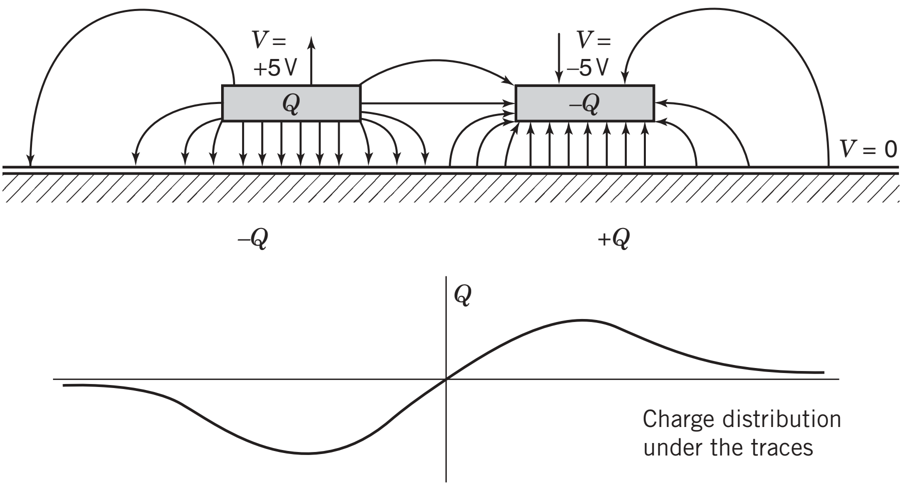


Figure 1.5. The electric field pattern around two traces over a ground plane

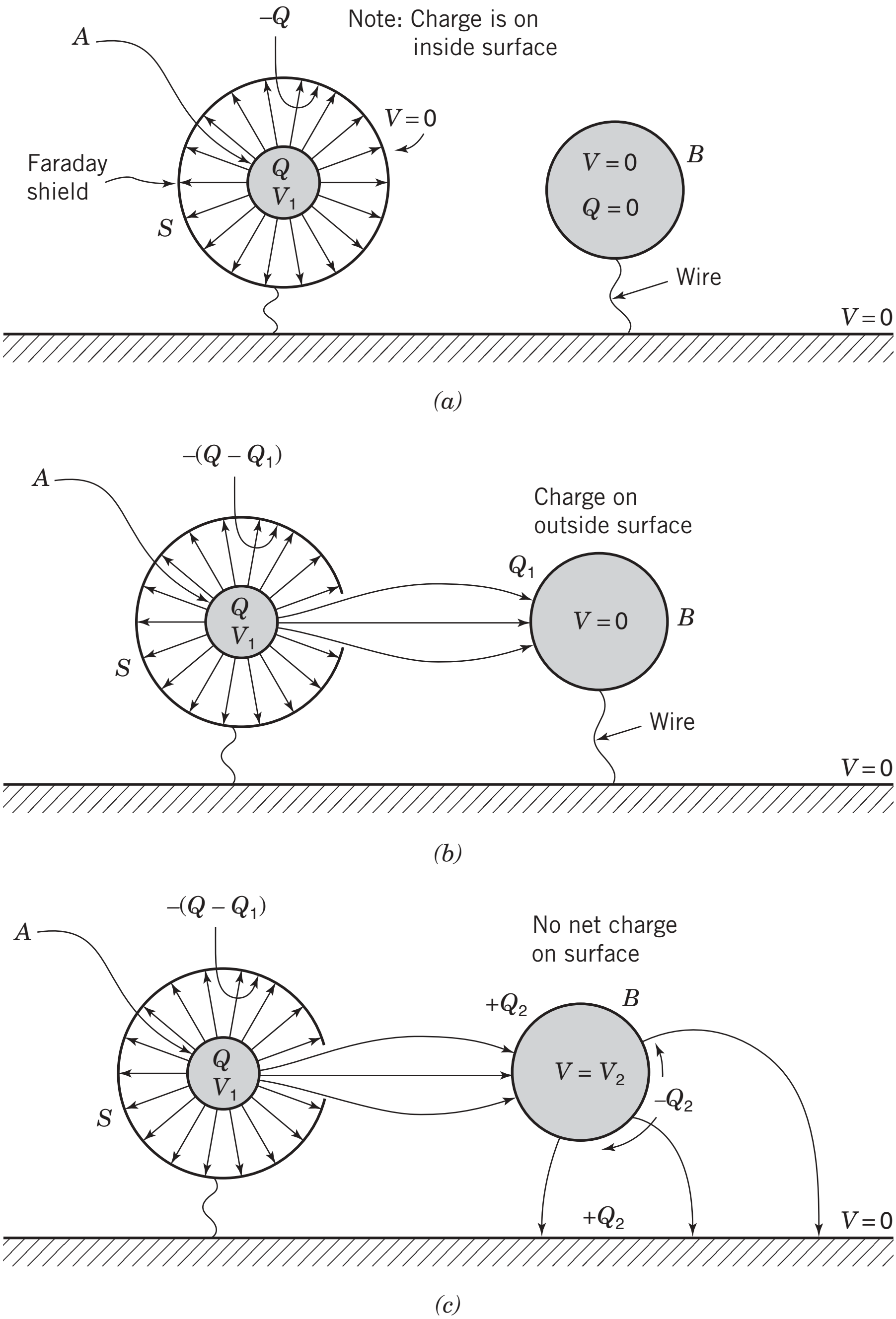


Figure 1.6. Field configurations around a shielded conductor

N.B.

Surface currents can flow on a conductor that is not connected to a circuit. There simply needs to be a changing electric field in the space around the conductor.

The electric field in Figure 1.6a is totally contained. The internal field can change and there are no induced currents on nearby conductors. This containment of the electric field is called *electric shielding*. The outer conductor A is often called a *Faraday shield*. A conductor with an outer conducting sheath is called a *shielded cable*. Later we will discuss a shielded conductor called *coax*.

N.B.

The electric field lines in Figure 1.6 terminate on the inside surface of the cable. If the voltage changes slowly, the resulting change in field causes current flow on the inside surface of the shield. Ideally this field does not penetrate into this shield and get to the outside surface.

N.B.

Shielding has nothing to do with external connections to the shield conductor. If the electric field is contained the shield is effective. The shield need not be at "ground potential" to be effective.

1.9. THE ENERGY STORED IN AN ELECTRIC FIELD

It takes work to move a charge in an electric field. In Figure 1.3 the work required to move a unit charge between the plates is the voltage difference between the plates. As more charge is moved across the space, the voltage between the plates increases. The work done on the charges is stored as potential energy. Where is this energy stored? Since it is not stored inside the conductors or on their surfaces the only place that is left is in the space between the plates. The same problem exists with gravity. When a weight is lifted in a gravitational field the added energy is stored in this field, not in the mass.

The force on a charge dq is $E\,dq$ where E is the force field. Assume the top plate has a charge q and the bottom plate has a charge $-q$. The E field from Eq. (1.4) is $q/\varepsilon_o A$. The work dW required to move an increment of charge dq across the distance h is

$$dW = fhdq = \frac{q}{\varepsilon_o A} hdq. \tag{1.9}$$

By integrating Eq. (1.9) over charge from 0 to Q, the total work W required to move a charge Q is

$$W = \int_o^Q \frac{qdq}{\varepsilon_o A} = \frac{Q^2 h}{2\varepsilon_o A}. \tag{1.10}$$

Since $E = Q/(\varepsilon_o A)$, the work W can be written as

$$W = \frac{1}{2} E^2 \varepsilon_o A h = \frac{1}{2} E^2 V \varepsilon_o \tag{1.11}$$

where V is the volume of the space between the two conductors.

N.B.

Every volume of space that contains an E field stores field energy.

N.B.

Every circuit with voltages stores electric field energy.

Somehow space has a quality like a spring that can be used to store potential energy. The conductors seem to provide a handle for holding onto this spring. We can't see it or feel it, yet we can do work with the energy that is stored in space. Without the restraint of these conductors the field energy would have to leave the area at the speed of light.

In practical circuits the electric field patterns are complex and the intensity of the field varies over space. To calculate the total stored energy, space can be divided into small volumes of near-constant field intensity. The important fact to remember is that the energy stored per unit volume is proportional to the square of the field intensity. In many practical problems the region of high field intensity is all that is important as this is where most of the energy is stored.

The potential at a point in space is a number, not a vector. The potential at every point is the work required to move a test charge against the forces created by every element of charge in the system. This voltage value is a summation based on Eq. (1.6). The electric field vector can be determined by locating the direction in which the voltage change is maximum. Stated mathematically, the gradient of the voltage is the electric field.

1.10. DIELECTRICS

We next consider the effect dielectric materials have on electric fields. Typical dielectrics are rubber, silk, Mylar®, polycarbonate, epoxy, air, and nylon. Up until now we have considered the electric field in a vacuum. Consider the two plates in Figure 1.3. If the space between the plates is filled by an insulating dielectric, it takes less work to move a charge Q from one plate to the other. This means the force field inside the insulator is reduced. This reduction factor ε_R is known as the relative dielectric constant. The force field between the planes in the dielectric medium is given by

$$E = \frac{Q}{\mathrm{A}\varepsilon_o \varepsilon_R}. \tag{1.12}$$

If the space is first filled with air, then a voltage V results in a charge Q. When a dielectric material is inserted between the plates, the voltage must drop to V/ε_R. If the voltage is again increased to V, the amount of charge on the surface would increase by the factor ε_R.

N.B.

The relative dielectric constant of air is 1.0006.

N.B.

Dielectric materials are used in capacitors to increase the charge stored per unit voltage.

1.11. THE D FIELD

It is convenient to discuss two measures of the electric field. The voltage between two points defines the E field intensity. A second field measure, called the D field, relates directly to charges. In a vacuum the E field and D field patterns are exactly the same. In a region where there are dielectrics, the E field intensity changes at every dielectric interface. The D field starts and stops on charges but does not change intensity at a charge-free boundary. In Figure 1.1, if the charge Q were located in a dielectric medium the E field would be reduced by a factor equal to the relative dielectric constant. The new E field is then given by Eq. (1.13):

$$E = \frac{Q}{4\pi\varepsilon_o\varepsilon_R r^2}. \tag{1.13}$$

The energy stored in the field of this charge is reduced by the square of the relative dielectric constant.

Figure 1.7 shows the field pattern between two planes where half of the space has a dielectric constant of 8. If the total spacing is 10 cm and the dielectric material has a dielectric constant of 8, the E field pattern must adjust so the total voltage difference is 10 V. The voltages in terms of the E field are $E/8 \times 5$ cm and $E/1 \times 5$ cm. E in the open space is obviously 8.9 V/cm and inside the dielectric it is 1.1 V/cm. The voltages are 8.9 V across the air space and 1.1 V across the dielectric. The electric field intensity in air is now 8.9 V/5 cm = 1.78 V/cm. Before the dielectric was inserted the field intensity in air was 1 V/cm. This means that the added dielectric has increased the charge Q on the plates by 78%. Note that the majority of energy is stored in the air space, not in the dielectric.

In Figure 1.7, the D field is continuous from the top to bottom plates. If E equals D/ε_o in the air space, then in the dielectric

$$E = \frac{D}{\varepsilon_R\varepsilon_o}. \tag{1.14}$$

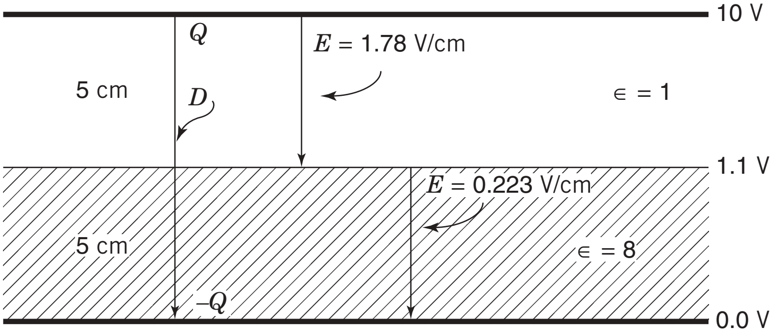


Figure 1.7. The electric field pattern in the presence of a dielectric

N.B.

The D field originates on charges and is not affected by the presence of a dielectric.

N.B.

Field energy is stored in the E field.

In high-voltage transformers, an oil dielectric is often used to reduce the E field around conductors. This reduction in the E field reduces the chance of arcing. The oil also helps to conduct heat away from the windings.

1.12. CAPACITANCE

The ratio of charge to voltage is capacitance C. The unit of capacitance is the farad. A capacitor of one farad stores one coulomb of charge for a voltage of one volt. In Figure 1.1 the voltage on the surface of the sphere is associated with a stored charge Q. The voltage V is $Q/4\pi r\varepsilon_o$. The ratio Q/V equals $4\pi r\varepsilon_o$. If the sphere is located in a dielectric medium, the voltage V is reduced by ε_R and the ratio Q/V is $4\pi r\varepsilon_o\varepsilon_R$. The capacitance of a sphere in a dielectric medium is

$$C = 4\pi r\varepsilon_R\varepsilon_o. \tag{1.15}$$

For the parallel planes in Figure 1.3, the voltage between the conducting plants is the E field times the spacing h. The voltage from Eq. (1.12) is $V = Qh/(\varepsilon_o\mathrm{A})$. If there is a dielectric present, the ratio Q/V is

$$C = \frac{\varepsilon_o\varepsilon_R\mathrm{A}}{h}. \tag{1.16}$$

Capacitance is a function of geometry. So far we have discussed two simple geometries, the sphere and parallel conducting planes. In most practical circuits the geometries are complex and the capacitances are not simply calculated. It is important to recognize that capacitance is controlled by three factors. It is proportional to surface area, inversely proportional to the spacing between surfaces, and proportional to the dielectric constant.

N.B.

Capacitance is a geometric concept. All conductor geometries can store some electric field energy, therefore they all have capacitance.

The idea of capacitance can be extended into free space. Consider a cube in space oriented so that electric field lines are perpendicular to two of the cube's faces. If there were equal and opposite charge distributions on the two opposite faces of the cube, the field lines would be no different. The voltage between the faces is the E field times the distance across the cube. Since we have an equivalent surface charge and a potential difference the ratio is capacitance.

N.B.

In the absence of conductors, free space has the ability to store electric field energy. A volume of space has a capacitance.

The factor ε_o is called the permitivity of free space and it is equal to 8.85×10^{-12} farads per meter. Consider the capacitance of a printed circuit trace over a ground plane. If the spacing *h* is 5 mm and the trace is 10 mm wide by 10 cm long, the trace area is 100 mm^2. The value of A/*h* is 20 mm or 20×10^{-3} meters. If the relative dielectric constant is 10, then the capacitance between the trace and the ground plane is equal to $(A/h)\varepsilon_R\varepsilon_o = 177 \times 10^{-12}$ farads or 177 picofarads (pF).

It is interesting to calculate the capacitance of the earth as a conductor. The radius of the earth is 6.6×10^6 meters. Using Eq. (1.15) the capacitance is 711 μF.

1.13. MUTUAL CAPACITANCE

A mutual capacitance is often referred to as a leakage capacitance or parasitic capacitance. The electric field pattern in most practical circuits is complex. The voltage on any one conductor implies a self-charge and induced charges on all the other conductors. For small component geometries a large percentage of the field energy may be parasitic in nature.

The ratio of charge to voltage on any one conductor is called a *self-capacitance*. Examples of self-capacitance are shown in Figures 1.1 and 1.4. An example of a mutual capacitance is shown in Figure 1.6b. The ratio of the charge induced on a second conductor to voltage on a first conductor is called a *mutual capacitance*. This measure of capacitance requires a test voltage be placed on one conductor and all other conductors must be at zero potential.

The mutual capacitance C_{12} is the ratio of charge induced on conductor 2 for a voltage on conductor 1. It turns out that $C_{12} = C_{21}$.[2] All mutual capacitance values are negative as the induced charge for a positive voltage is always negative. A simple geometry showing a few mutual capacitances is shown in Figure 1.8.

A voltage V_1 is placed on trace 1 and traces 2, 3, and 4 are at zero volts. The ground plane is also at zero volts. The capacitances C_{11}, C_{12}, C_{13}, and C_{14} are the ratios V_1/Q_1, V_1/Q_2, V_1/Q_3, and V_1/Q_4. Mutual capacitance C_{32} would be the ratio V_3/Q_2.

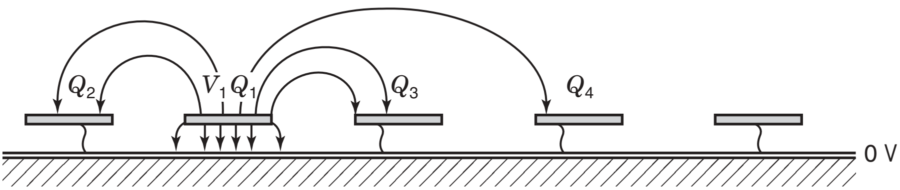
Mutual capacitances are a function of circuit geometry. These capacitances often limit or determine circuit performance. In an integrated circuit amplifier, mutual capacitances are an integral part of the design. They may define circuit bandwidth and circuit stability.

1.14. DISPLACEMENT CURRENT

Figure 1.3 shows two conducting plates. If a charge Q is placed on the top plate, a charge $-Q$ must exist on the lower plate. The ratio of charge Q to the voltage on the top plate is the capacitance of this geometry. This geometry is typical of many small commercial capacitors.

If the charge stored on the capacitor plates increases linearly with time the voltage difference V will also increase in a linear manner. A fixed constant current source can serve to provide this increasing charge. The equivalent circuit is shown in Figure 1.9.

For convenience we will use a standard circuit symbol for this capacitance and label it with the letter C. The current flow in this circuit can be looked at in two ways. First viewpoint: The electrons flow onto the plates of the capacitor but they do not flow through the dielectric. Second viewpoint: In a loop



$$C_{11} = \frac{V_1}{Q_1} \qquad C_{12} = \frac{V_1}{Q_2} \qquad C_{13} = \frac{V_1}{Q_3}$$

Figure 1.8. The mutual capacitances between several traces over a ground plane

[2] Measuring a small mutual capacitance can be difficult. One method of making a measure is to use a changing voltage (a sinusoidal voltage) and observe the changing charge as current flow. The current can be measured as a voltage across a series resistor. Leakage capacitances as low as 0.1 pF can be measured this way. This capacitance measurement requires very careful shielding.

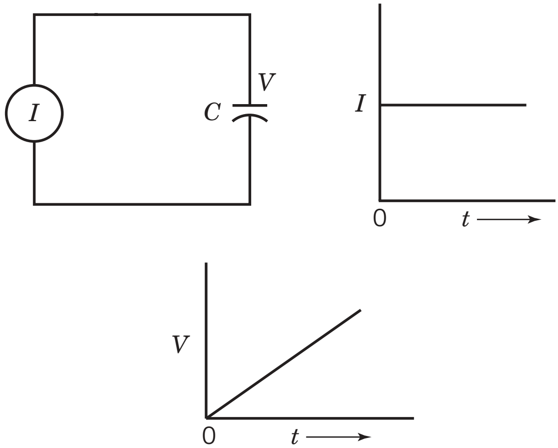


Figure 1.9. A capacitor driven from a constant current source

analysis the current flows through the capacitor. Which viewpoint is correct? They both are correct providing we interpret the changing field in the dielectric correctly. As the charge Q accumulates on the plates, the D field in the dielectric also increases. This changing D field is equivalent to a current flow. We can view a changing D field as a displacement current. This statement is one of Maxwell's wave equations.

N.B.

A changing D field in space is equivalent to a displacement current flowing in space.

It is important to realize that we cannot have physical laws that work some of the time. Physical laws must work the same way all of the time at all frequencies. Circuit theory does not deal with the electric fields but they are present if there are voltage differences. Later we will discuss electromagnetic radiation. Field energy or radiation that leaves a circuit involves both an electric and a magnetic field. In a capacitor the changing electric field is a current that has an associated magnetic field. As you will see in the next chapter a changing magnetic field requires an electric field. In other words, these changing fields go hand in hand. In effect, energy can be carried across the capacitor plates as radiation and this requires the presence of both an electric and a magnetic field. Obviously it is very cumbersome to view the operation of a capacitor in terms of this radiation.

1.15. ENERGY STORED IN A CAPACITOR

The energy stored in the field of a capacitor from Eq. (1.9) is $\frac{1}{2}E^2\ \varepsilon Ah$. If we substitute $E = V/h$ and remember from Eq. (1.4) that $E/\varepsilon A = Q$, then we can write the energy E as a function of charge and voltage:

$$E = \frac{1}{2}QV. \tag{1.17}$$

We can use the ratio $C = Q/V$ in Eq. (1.14) to obtain two equivalent equations for energy storage:

$$E = \frac{1}{2}CV^2 \tag{1.18}$$

$$E = \frac{1}{2}Q^2/C. \tag{1.19}$$

1.16. FORCES IN THE ELECTRIC FIELD

Field energy exists in the space between individual electrons. In Figure 1.1 the electrons moved apart on the surface of the sphere until they were uniformly spaced. In effect, they arranged themselves to store the least amount of field energy. This is characteristic of nature in so many ways. In fact, all static field configurations represent a minimum of field energy storage within the geometric constraints provided. For a set of charges on conductors there is only one field configuration possible and that field configuration stores a minimum of field energy.

N.B.

If there is a way, nature will follow a path that will reduce the amount of potential energy stored in a system.

If the plates of a capacitor are moved closer together, the capacitance increases. Equation (1.19) shows that if the charge Q is fixed, then a larger capacitance stores less energy. This means there is a force acting on the plates trying to reduce the spacing as this reduces the potential energy stored in the system. It helps to recognize that work equals force times distance. The derivative of work with respect to distance is simply force.

Acoustic tweeters work on this principle. Parallel metal plates can be made to move air by placing an audio voltage between the plates. Potential differences in the order of several hundred volts are required. The E field must be biased at dc so that there is no doubling of frequency.

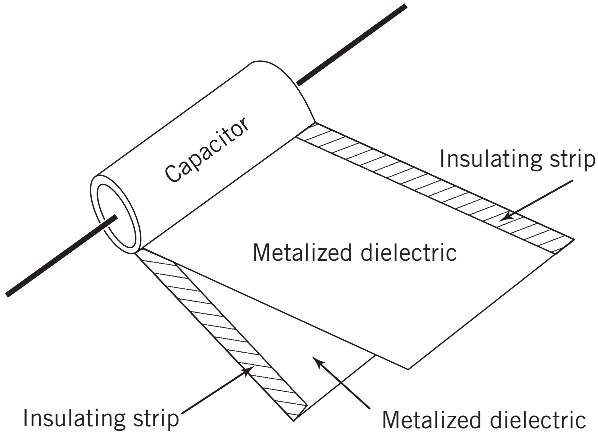


Figure 1.10. A typical wrap and foil capacitor

1.17. CAPACITORS

Capacitors are components that store electric field energy. A circuit engineer has a wide choice of component types and values to choose from. There are many construction styles and types of dielectric and voltage ratings. Capacitance values can cover a significant range, covering 12 orders of magnitude from a few picofarads to 1 or 2 farads. Typically electrolytic capacitors in the range 100 μF are used to store field energy for general circuit use. Capacitors in the range 0.01 μF can be used on circuit boards for supplying local energy to operate digital circuits. Many capacitors are of the wrap and foil type as shown in Figure 1.10. Connections are made to the foil on the opposite ends of the cylinder. In many capacitors the conductors are vacuum deposited on the dielectric. An example of this technique is metalized Mylar®. Conductor surfaces can be made irregular to increase the effective surface area, which increases the capacitance. Electrolytic capacitors use an electrolyte as a dielectric. This type of capacitor must be used where there is a polarizing voltage or the capacitor will conduct like a poor resistor.

In Chapter 7 more will be said about the role capacitors play in making it possible to operate digital circuits at high clock rates.

CHAPTER 2

Magnetics

2.1. MAGNETIC FIELDS

We are all familiar with the magnetic field of the earth. A compass needle responds to this field to provide us with a navigational aid. We have all experimented with magnets and noted the forces that exist in the space between poles. If it were not for magnetic effects we would not have motors, generators, or transformers. These devices are basic to our entire civilization. It is easy to forget that this same magnetic phenomenon is at the heart of operating all of our electronic circuitry. The story starts again with the atom.

The electrons in atoms have spins that create local magnetic fields. In a few elements the atomic structure is such that atoms can be aligned to allow a net magnetic field to leave the atoms. The iron in a mineral called magnetite can produce a magnetic field that can also deflect a compass needle.

The flow of electrons is another way to generate a magnetic field. These electrons can be flowing on a conducting surface, along a circuit trace, or in free space. The simplest geometry to consider is a long circular conductor. To demonstrate that a current generates a magnetic field, thread a conductor through a piece of paper. Iron filings placed on the paper will line up in concentric circles around the conductor when it carries a direct current. A small compass needle in the vicinity of the conductor will also align itself with the magnetic field pattern.

The letter H is reserved for the magnetic field generated by a current. Figure 2.1 shows the shape of the H field around a long straight conductor carrying a current I. In the figure the field lines form circles around the conductor. In this representation the closer the field lines, the more intense the field. These field lines are also called lines of magnetic flux. Note the field intensity for this geometry is constant along a path that is concentric with the conductor.

The field intensity (flux density) is equal to the number of lines that cross a unit area perpendicular to the lines. Note the magnetic field is a vector field. At every point in space the field has a magnitude and a direction.

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

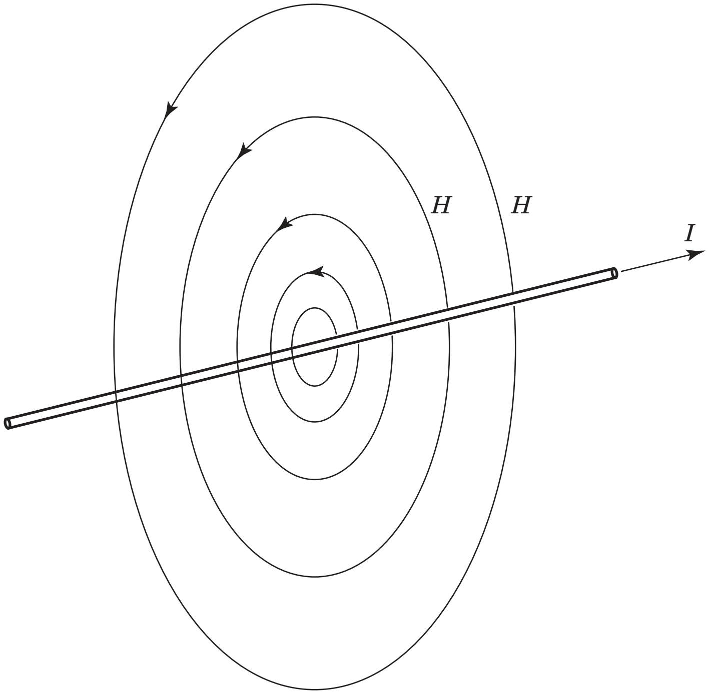


Figure 2.1. The H field around a current-carrying conductor

The magnetic field is a force field. This force can be exerted only on another magnetic field. If two parallel conductors both carry current, the resulting force will try to move the conductors together.[1] The direction of the force, the direction of the current flow, and the direction of the field lines are all perpendicular to each other.

2.2. BIOT AND SAVART'S LAW

Biot and Savart's law states that the integral of the H field intensity in a closed loop path around a conductor is equal to the current threading that loop:

$$\oint \mathrm{H}\, \mathrm{d}l = \mathrm{I}. \tag{2.1}$$

The simplest path to use for this integration is one of the concentric circles in Figure 2.1 where H is constant and r is the distance from the conductor. Solving for H we obtain

[1] The force is the direction to reduce the energy stored in the circuit inductance. This is discussed later in this chapter.

$$H = \frac{I}{2\pi r}. \tag{2.2}$$

From this equation we see that H has units of amperes per meter. In this geometry the H field falls off linearly with distance. The value of H is constant at a distance r from the conductor. For a long conductor the H field falls off linearly with distance. Eq. (2.1) is often called Ampere's law.

2.3. THE SOLENOID

The magnetic field of a solenoid is shown in Figure 2.2. Note the field intensity inside the solenoid is nearly constant while outside the solenoid it is very low. Using Ampere's law the integral of H over a closed line of flux is

$$\oint H\, dl = nI\, l. \tag{2.3}$$

This integral is not easily performed as H is not constant outside of the solenoid. The important thing to note is that the H field is proportional to the current and to the number of turns in the solenoid. If the only H field intensity that contributes to the integral is inside the solenoid, then $Hl = nI$.

2.4. FARADAY'S LAW AND THE INDUCTION FIELD

When a conducting coil is moved through a magnetic field a voltage appears at the open ends of the coil. This is illustrated in Figure 2.3. The voltage depends

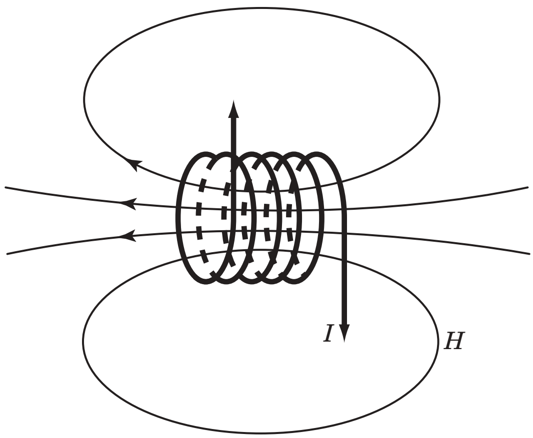


Figure 2.2. The H field around a solenoid

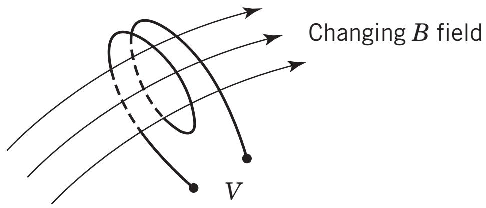


Figure 2.3. A voltage induced into a moving coil

on the intensity of the magnetic field, the number of turns in the coil, and the rate at which the flux that threads the coil is changing. No voltage results if the coil is oriented so that the total flux through the coil is constant.

Remember there are two measures of the electric field. The D field is related to charge and the E field is related to force on another electric field. The magnetic field also has two measures. As we have seen, the H field is proportional to current flow. The field representation that induces voltage is called the B or induction field. The relationship between B and H fields is given by

$$B = \mu_R \mu_0 H \tag{2.4}$$

where the factor μ_0 is the permeability of free space and μ_R is the relative permeability of the medium. Air has a relative permeability of unity. The factor μ_0 is equal to $4\pi\ 10^{-7}$ teslas per ampere per meter. The relative permeability of iron can vary from 500 to 100,000. For an area of constant field intensity the magnetic flux ϕ is simply the product BA, where B is in teslas, A is the area in m^2, and ϕ is the flux in webers. We shall refer to this flux as "lines." The voltage induced in a conducting coil is

$$V = n\frac{d\phi}{dt} \tag{2.5}$$

where n is the number of turns in the coil. This equation can be written in terms of the B field as

$$V = nA\frac{dB}{dt}. \tag{2.6}$$

N.B.

The relationship between voltage and the B field does not depend on the magnetic material involved. The coil can be in air or wrapped around a magnetic material.

Equation (2.5) is known as *Faraday's law*. If the induction flux B increases linearly, a steady voltage V must exist at the coil ends. The inverse is also true. If there is a fixed voltage on this coil, then the B field intensity in the coil must increase linearly. This is known as *Lenz's law*. These laws are very difficult to observe for simple coils in air at frequencies below a few megahertz.

In electrostatics the E field was the force field. In magnetics the B field is the force field. The force on a current loop in a magnetic field is proportional to the current and the B measure of the field produced by the current.

2.5. INDUCTANCE

DEFINITON

Inductance. The ratio of magnetic flux generated per unit current.

It is a difficult problem to calculate the total magnetic flux in a typical geometry. A practical way to measure inductance is to use Faraday's law. We begin by looking at the coil in Figure 2.2. Equation (2.6) states that the magnetic field B will increase at a constant rate if a steady voltage is applied to the coil. Both the H field and the induced voltage are proportional to the number of turns. Therefore the voltage V is proportional to n^2. For a coil in air $B = \mu_0 H$ and Eq. (2.6) can be rewritten in terms of a changing current as

$$V = n^2 A k \mu_0 \frac{dI}{dt} = L \frac{dI}{dt} \tag{2.7}$$

where k relates to the geometry of the coil. The factor $n^2 A k \mu_0$ is the inductance L of the coil. The unit of inductance is the henry. Equation (2.7) states that if V = 1 volt, then for an inductance of one henry the current will rise at the rate of one ampere per second. A henry is a large unit of inductance. Typical circuit inductors range from a few microhenries to a few millihenries. The unit abbreviations are mH and μH. In Chapter 7 when we discuss decoupling capacitors, inductance in the picohenry range will be important.

As mentioned before, the inductance in Eq. (2.7) is for a coil in air. Most commercial inductors are constructed by associating the coil with a magnetic material. We will discuss this construction in the next sections.

2.6. THE ENERGY STORED IN AN INDUCTANCE

In the electric field case, work was related to moving a small test charge in the electric field. In the magnetic field case, a test magnetic field (unipole) does not exist. Pushing on a charge in a magnetic field is complicated because the force results in a circular motion. One way to calculate the work stored in a magnetic field is to use Eq. (2.7). The voltage V applied to a coil results in a linearly increasing current. At any time t the power P supplied is equal to VI. Power is the rate of change of energy or $P = d\boldsymbol{E}/dt = VI$ where $\boldsymbol{E}$ is the stored energy in the inductance. Since the voltage $V = LdI/dt$ the stored energy in an inductance L is

$$\boldsymbol{E} = L\int_o^I IdI = \frac{1}{2}LI^2. \tag{2.8}$$

N.B.

An inductor stores field energy. It does not dissipate energy.

The presence of a voltage V on the terminals of an inductor implies an electric field. The movement of energy into the inductor thus requires both an electric and a magnetic field. This is very similar to what we found when we placed field energy into a capacitor. By moving charge into the capacitor we created a magnetic field. We have just shown that both the E and B fields must be present to move energy into an inductor. To remove energy from an inductor or a capacitor both fields must again be present.

Faraday's law requires a voltage when a changing magnetic flux couples to a coil. This voltage means electric field energy must be present. This energy is stored in a distributed manner between every conducting element pair on the coil. When a steady current flows in the inductor the magnetic flux is constant. This means that the voltage is zero and there is no electric field energy stored. When the circuit is opened the current starts to diminish. The result is a changing flux that creates a voltage and this voltage begins to place energy into the interwinding capacitance of the coil. In effect the current continues to flow but now it flows in the distributed capacitance of the coil. The magnetic field energy begins its conversion to electric field energy.

N.B.

The energy stored in an inductor cannot simply vanish. It must go somewhere.

The field energy in the inductor is $\frac{1}{2}LI^2$. The field energy in a capacitance is $\frac{1}{2}CV^2$. Consider a 1-mH inductor carrying a current of 0.1 A. Assume the shunt capacitance equals 100 pF. The stored energy is 5×10^{-4} joules. When this energy is fully transferred to the capacitance the voltage must be 3,116 volts. The natural frequency of the inductance and its own parasitic capacitance is about 500 kHz. The energy transfers from the inductor to the capacitor in $\frac{1}{4}$ cycle or in 0.5 μs. Mechanical contacts cannot open very far in 0.5 μs so the result is the voltage breakdown of air. For a relay contact this sudden rise in voltage results in an arc. The energy that was stored in the inductor now goes into light and heat. If the switch is a semiconductor, the resulting voltage would probably destroy the component. There are several ways to absorb the stored magnetic field energy and avoid a high voltage. A reverse diode across the coil can provide a path for interrupted current flow. Another technique is to place a capacitor across the inductance. This will lower the natural frequency and at the same time reduce the voltage.

N.B.

The field energy in an inductance cannot be dissipated in zero time.

2.7. MAGNETIC FIELD ENERGY IN SPACE

To solve for the magnetic field energy stored in a region of space we can invent an increment of magnetic field from an increment of current flowing in a closed superconducting loop. We can use this increment of magnetic field to add field energy into a main superconducting loop. When the incremental loop of current is moved a distance x in the B flux of the main loop, the work done on the main loop is BHAx, where HA is the flux from the incremental current loop. This work increments the current in the main loop, which increases the field intensity. To build up the field energy in the main loop we must bring the energy across in small increments. At the beginning no work is required as the initial B field is zero. When the B field is maximum the work per unit current is $\boldsymbol{W} = \mathrm{BHA}x$. The average work done to store energy in the B field is one-half this value. Since $\mathrm{B} = \mu_0\mathrm{H}$ in a vacuum, we can write the energy stored as

$$\boldsymbol{E} = \frac{1}{2}\mathrm{B}^2\boldsymbol{V}/\mu_0 \tag{2.9}$$

where the volume $\boldsymbol{V} = \mathrm{A}x$ and μ_0 is the permeability of free space.

Space can store magnetic field energy. In this sense every volume of space has an inductance. This is the counterpart to the electrostatic case, where every volume in space has a capacitance.

There is magnetic flux around an isolated conductor carrying current. The flux per unit length defines an inductance per unit length. Figure 2.4 shows the inductance of round copper wires as a function of length. The important thing to appreciate is that this inductance is essentially independent of conductor diameter. At a frequency of 10 MHz a 20-inch-long #19 conductor looks like 120 ohms. A 20-inch-long #0000 conductor looks like 48 ohms. Three parallel #19 wires spaced a few inches apart would look like 30 ohms. It is obvious that the amount of copper is not the issue. The geometry of the copper makes

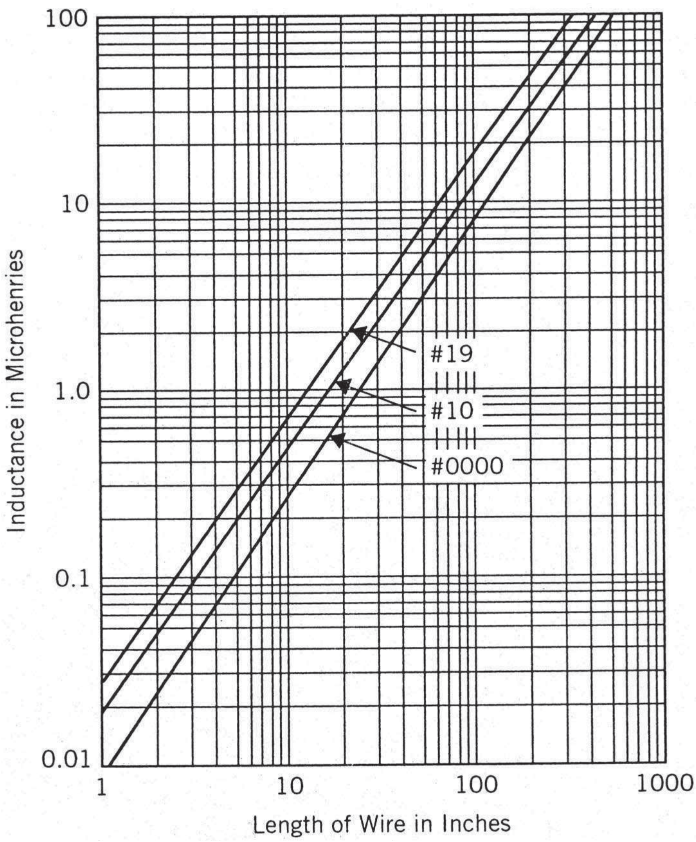


Figure 2.4. The inductance of round copper conductors

all the difference. At high frequencies it is essentially impossible to short two points together. This important point will be discussed later in the book.

> **N.B.**
>
> Heavy conductors are not the solution to limiting potential differences. The reason is simple. They do not get rid of fields.

2.8. THE MAGNETIC CIRCUIT

Magnetic materials are needed to design motors, generators, and transformers. These materials are also used to make practical inductors. To understand the role of magnetic materials we start with Figure 2.5 where a conducting

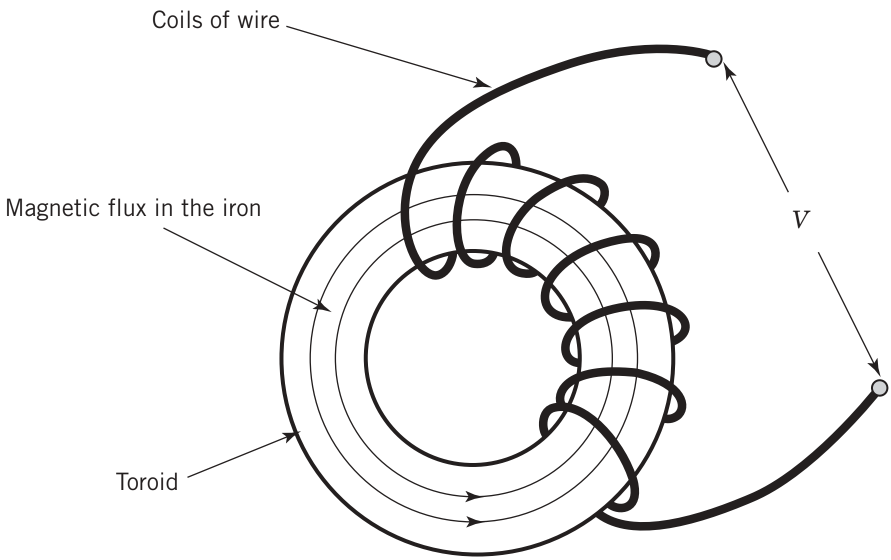


Figure 2.5. A coil wound on a toroidal core of magnetic material

coil is wound on a simple toroid made of magnetic material. Assume a steady voltage V is applied to the coil starting at time $t = 0$.

From Eq. (2.5) the voltage depends on the rate of change of the B field or $dB/dt = V/nA$. The maximum B field intensity in a typical power transformer core is about 1.5 teslas (15,000 gauss).[2] If V is a constant, we can calculate how long it will take the field intensity to increase to this level. Since $B = tV/nA$, the time t for B to reach 1.5 teslas is

$$t = \frac{nAB}{V} = \frac{1.5nA}{V}. \tag{2.10}$$

If the area A is $1\,cm^2$ (10^{-4} square meters), $V = 1$, and $n = 100$ turns, the time t is 0.015 seconds. This time is independent of the type of magnetic material in the core.

The H field associated with this B field can be determined by using Ampere's law. If we assume the H field is constant in the toroid and if we choose a closed path that threads all of the turns of the coil, then

$$\oint H\,dl = 2\pi rH = nI, \tag{2.11}$$

$$I = \frac{2\pi rH}{n}. \tag{2.12}$$

[2] The unit of magnetic induction used in transformer design is the gauss. One gauss is 10^{-4} teslas.

To calculate I we need to relate B to H in the magnetic material. Using Eq. (2.4),

$$I = 2\pi r B / n\mu_R\mu_0. \tag{2.13}$$

If we assume $r = 0.1\,\text{m}$, $\mu_R = 1$, and setting $\mu_0 = 4\pi 10^{-7}$, the value of I is 7,500 amperes. Obviously this level of current is not acceptable. If the relative permeability of the core material is 50,000, the current reduces to 0.15 amperes, a very practical number.

The current required to supply the B field is called a magnetizing current. Without the presence of a magnetic material, this example shows that the circuit is of no practical use. If the permeability of the material were infinite, the magnetizing current would be zero. An ideal transformer requires no magnetizing current.

It is interesting to calculate the energy stored in the magnetic material and to determine the inductance of this geometry. The volume V of the toroid is approximately $2\pi \cdot 10^{-5}\,\text{m}^3$. For an example, set B maximum equal to 1.5 teslas. Using Eq. (2.9), setting $\mu_0 = 4\pi \cdot 10^{-7}$, if $\mu_R = 50{,}000$, the energy $\boldsymbol{E}$ is $4.5 \cdot 10^{-3}$ joules. The inductance using the relationship $\boldsymbol{E} = \frac{1}{2}LI^2$ is 0.4 henries. This is not a very practical inductor because it can store only $4.5 \cdot 10^{-3}$ joules and the core is nearly saturated at a current level of only 150 mA.

2.9. A MAGNETIC CIRCUIT WITH A GAP

We next consider the effect of placing an air gap in the magnetic path in Figure 2.5. This is shown in Figure 2.6. When a voltage V is applied to the coil, the B field will increase by Faraday's law. The gap has no influence on the buildup of the B field. Remember the B field is continuous around the magnetic path. The H field in the magnetic material is $B/\mu_0\mu_R$. In the gap the H field is B/μ_0. Note that the H field is much larger in the gap.

The H field in the gap must be B/μ_0 or 1.5 teslas divided by $4\pi \times 10^{-7}$ or 1.19×10^6 amperes per meter. For 100 turns the current requirement is reduced by a factor of 100. The current required to establish this H field in a gap of 10^{-3} meters is 11.9 amperes. The current required to establish the H field in the magnetic material is essentially the same as before, or 150 mA. The total current is therefore 12.05 amperes.

The energy stored in the gap is $\frac{1}{2}B^2\boldsymbol{V}/\mu_0$ where the gap volume is $1\,\text{cm}^2 \cdot 0.1\,\text{cm} = 10^{-5}\,\text{m}^3$. This energy is $1.12\ 10^{-5}/4\pi \cdot 10^{-7} = 8.75$ joules. Without a gap the energy stored was $4.5 \cdot 10^{-3}$ joules, a small number. The inductance using Eq. (2.8) is 0.241 henries, a significant number.

N.B.

Magnetic field energy is largely stored in air, not in a magnetic material.

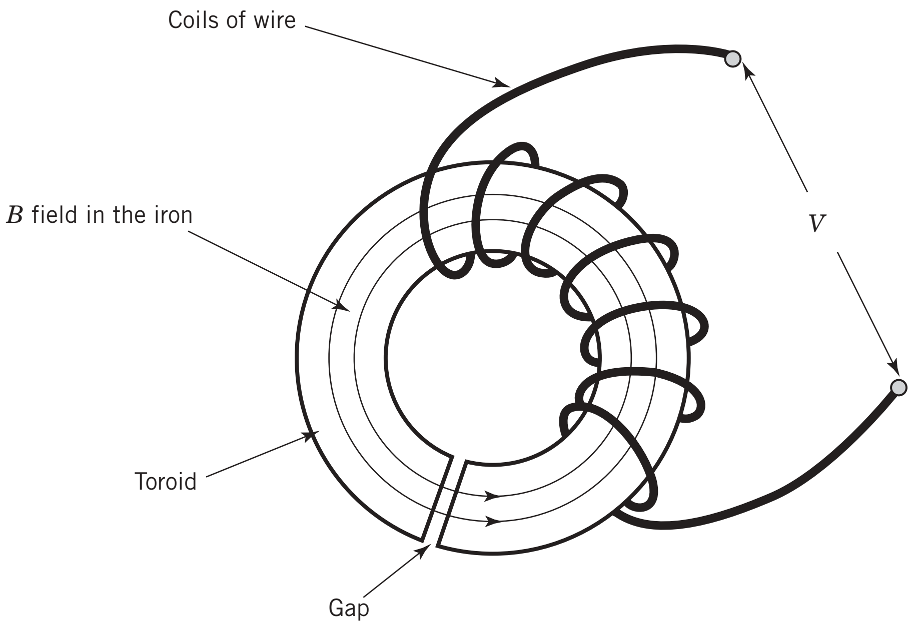


Figure 2.6. A magnetic circuit with an air gap

N.B.

For a geometry involving air and dielectrics the majority of electric field energy is stored in the air. See Figure 1.7.

The toroidal shape of the magnetic material serves to focus the B field into the gap. The magnetic field follows the permeable material as this path takes so much less energy than using the nearby air space. This is another example of nature configuring a field to store the least amount of energy. In this example, the B field is near maximum. If the magnetic material becomes saturated, the ability to store energy is lost. When saturated the magnetic material has a relative permeability of one.

The toroid with a gap in Figure 2.6 is called an *inductor*. The quality of the inductor at different frequencies depends on the type of magnetic material and how the coil is wound.

2.10. SMALL INDUCTORS

Inductors in the microhenry range are often wound on a small cylindrical core. The flux path is in the space around the inductor. Since very little energy is stored in a magnetic material the energy storage for these inductors is in the surrounding space.

A conductor that is threaded one or more times through a ferrite bead forms an inductor. The inductance is in the order of nanohenries. One nanohenry at one gigahertz is only 6.28 ohms. This level of inductance is effective only in very-low-impedance circuits. If there is any low-frequency current in the conductor, the core may be saturated. The presence of the core may space the conductors, thus reducing some of the coupling. In this case, the magnetic material is not needed. It is important to note that the permeability of most magnetic materials falls off rapidly above a few megahertz.

2.11. SELF- AND MUTUAL INDUCTANCE

The magnetic flux generated by a current can couple into nearby circuits. The flux associated with this coupling is called *leakage flux*. The ratio of the leakage flux that couples into a second circuit to the generating current is called a *leakage* or *mutual inductance*. The symbol L_{12} represents the flux coupled to circuit 2 from a current flowing in circuit 1. The symbol L_{11} represents a self-inductance. The inductor in Figure 2.6 has a self-inductance of 0.241 henries. In this case, all the flux of the inductor couples to its own coil.

In Figure 2.6 there is an H field close to the current-carrying conductor. This is H field leakage flux that does not use the core as the magnetic path. This means that circuits that are very close to the toroid can couple to this field. Around the gap the leakage flux can be more substantial. Many inductors are wound using "cup cores." In this geometry the gap is in the very center of the core and the leakage flux is well contained.

The circuit symbol for inductance implies that the field energy is stored inside the inductor. Small-valued inductors (microhenries) often store a fraction of their field energy in the space around the component. Some of this field radiates and is not returned to the circuit. This is a situation not covered by circuit theory. We will discuss this radiation in a later chapter.

2.12. TRANSFORMER ACTION

When a steady voltage is connected to the coil in Figure 2.5 the B flux increases linearly with time. From Eq. (2.10) if $n = 100$ turns and the voltage is 1 volt, the B field will reach an intensity of 1.5 teslas in 15 ms. If the voltage is 10 volts, the time would be 1.5 ms. When the B field reaches 1.5 teslas we can reverse the voltage polarity. After this reversal the flux intensity starts decreasing. In another 1.5 ms the field intensity is again zero. After another 1.5 ms the flux intensity is –1.5 teslas. At this point, if there is a second voltage polarity reversal, the flux would again return to zero in an additional 1.5 ms. This full cycle would take 6.0 ms. If this cycle is repeated over and over, the resulting waveform is called a *square wave* at a frequency of 166.6 Hz. The B field we have discussed is independent of the permeability of the magnetic material in the core. The flux pattern and the coil voltage are shown in Figure 2.7.

The square-wave voltage in Figure 2.7 has no dc component. In theory, if a dc component were applied, the core would saturate in a few cycles. In a practical situation a slight dc offset can be accommodated. The penalty is an asymmetrical flow of magnetizing current and some distortion in the voltage waveform.

N.B.

A magnetic material is needed to limit magnetizing current.

We now consider a core with two coils as shown in Figure 2.8. The two coils are called the primary and the secondary windings, respectively. If a voltage is applied to the primary coil, the B field we have just described couples to both coils. An oscilloscope placed on the second coil would show a square wave at 166.6 Hz. Since the original square wave was 20 V peak-to-peak, the secondary voltage would be the same. If the number of turns on the secondary were increased to 200, the voltage would be 40 V peak-to-peak. If the number of turns on the primary and secondary were both doubled, the voltages would not change. The differences would be that the maximum B field would be 0.75 teslas. Since the magnetizing current is proportional to the H field and inversely proportional to the number of turns the magnetizing current would be reduced by a factor of four.

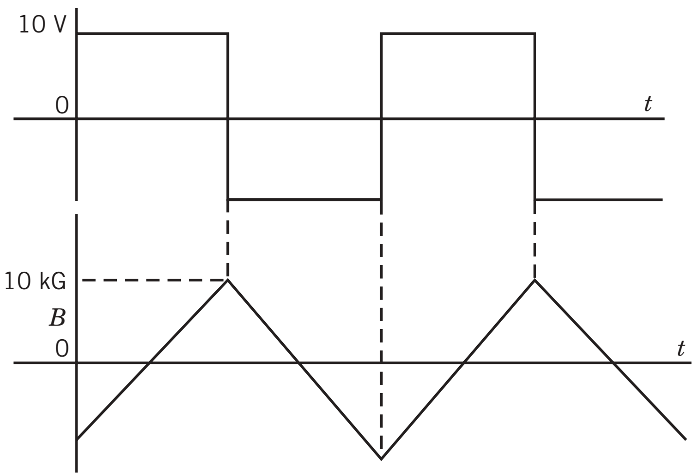


Figure 2.7. The flux pattern for a square-wave voltage applied to a transformer coil

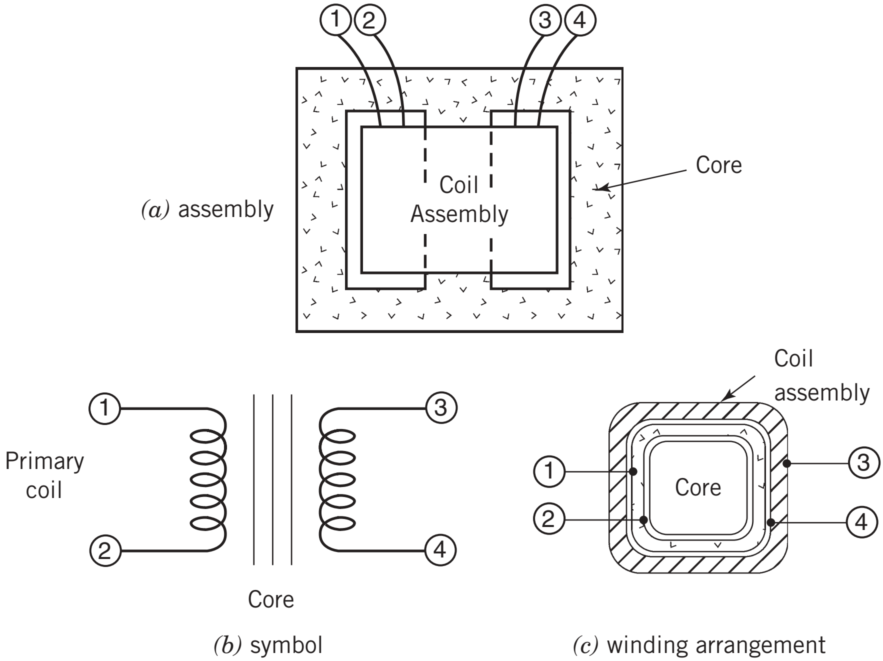


Figure 2.8. A core with two coils that form a simple transformer

If a load resistor is placed on the secondary coil, the current that flows by Ohm's law is V/R. To keep the net H field constant this same number of ampere-turns must flow in the primary coil. If there are 100 primary turns and 200 secondary turns, a secondary current of 0.1 A requires a primary current of 0.2 A. The current in the primary coil must now equal the magnetizing current plus the load current. In a well-designed transformer the magnetizing current is held to a few percent of the maximum load current. It is convenient to accept the idea that the magnetizing current creates the B field in the magnetic material.

The energy stored in the magnetic material is stored in what is called the magnetizing inductance. Ideally current flowing in an inductance produces no heat. There is some heat loss because the magnetizing current flows in the primary coil resistance. There are eddy current losses in the magnetic material as there are currents associated with a changing magnetic field. See Section 2.14. Consider the heat loss in a 10-kW transformer. A 1% loss (100 watts) would be a good design. This amount of heat can cause a significant rise in temperature in the center of the transformer. Usually some form of forced ventilation is required for this size of transformer.

In a transformer, secondary load impedances are reflected to the primary side multiplied by the turns ratio squared. In the example above, the turns ratio is 1:2. A 100-ohm load on the secondary appears as a 25-ohm load to

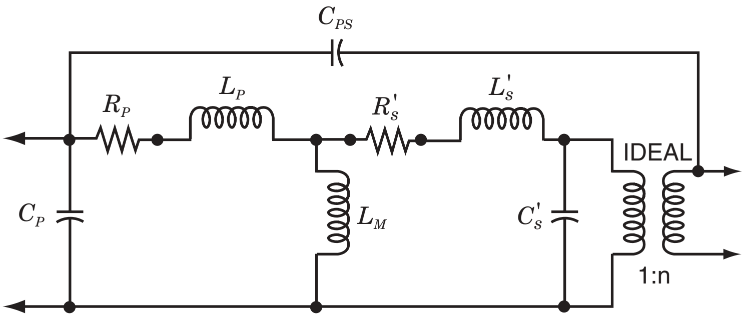


Figure 2.9. The equivalent circuit of a transformer

the primary. To show this is true consider a 10-volt secondary voltage and a current level of 0.1 amperes or a power level of 1 watt. On the primary, the voltage is 5 volts and the current is 0.2 amperes. The primary voltage source sees a resistance of 5 V/0.2 A = 25 ohms.

A practical transformer has series leakage inductance and shunt capacitance associated with coil construction. These reactances appear as loads to any primary voltage. Reactive loads are reflected across the transformer by the turns ratio squared. This means that the secondary leakage inductance is multiplied by the turns ratio squared and the secondary capacitance is divided by this turns ratio squared. The equivalent circuit of a transformer is shown in Figure 2.9. The transformer symbol in the figure represents an ideal transformer with a step-up turns ratio of 1 : *n*. The magnetizing inductance is labeled L_m. The resistances of the coils are labeled R_P and R_S respectively. There are capacitances associated with the primary and secondary coils labeled C_P and C_S. The inductances associated with the leakage flux are labeled L_P and L_S. The primes indicate that the values have been corrected for the turns ratio squared and referenced to the primary side of the transformer. The secondary leakage inductance and resistance are divided by n^2. The secondary shunt capacitance is multiplied by n^2.

The load current and the magnetizing current both flow in the primary leakage inductance. Only the load current flows in the secondary leakage inductance. This means that the leakage flux is load dependent. In facilities with large distribution transformers, load currents can be hundreds of amperes. The leakage flux from these transformers can interfere with some types of computer monitors or magnetic memories. It is possible to limit the leakage inductance in the design of a transformer by interleaving the primary and secondary coils and by using more core material that allows fewer turns per volt.

The shunt capacitances associated with transformer coils are related to the storage of electric field energy. If there are conductors and a voltage difference, there must be an E field. Every conducting element along the conductor stores field energy with every other conducting element depending on separation and voltage difference. Winding a coil so that the starting turns are near the ending turns will greatly increase the stored electric field energy. In some power transformers the start and end points are separated to avoid a possible voltage breakdown. Most of the field energy is stored in air so the parasitic capacitances are not greatly influenced by a dielectric. The capacitances between the primary and secondary coils pose a complex problem that will be discussed later. In Figure 2.7 this mutual capacitance is represented by C_{12}.

The magnetizing inductance of a transformer can be measured at the primary connections if the secondary coils are open circuited. The result will vary depending on voltage level and the frequency selected. The leakage inductance can be measured from the primary winding if the secondary coils are all short circuited. The test voltage must be a small fraction of the normal operating voltage.

When a voltage is impressed on the primary coil of a transformer there is an electric field. As the current builds up there is a magnetic field. The transfer of energy across the transformer requires the presence of both fields. As we have seen, placing energy into a capacitor or into an inductor requires both fields. A transformer requires both an electric and magnetic field to transfer energy from the primary side to the secondary side.

2.13. HYSTERESIS AND PERMEABILITY

The relationship between the B and H fields in a magnetic material is not linear. The B/H curves for a typical material are shown in Figure 2.10. The ratio of maximum B to maximum H is one measure of permeability. Typically B/H curves are shown for a sinusoidally varying B field. Note that this measure of permeability varies with B maximum. B/H characteristics will also vary with frequency and with primary voltage waveform. This means that a permeability figure is an approximate measure for a complex relationship.

There are a large number of magnetic materials for use in magnetics. On one hand, high-silicon transformer steel does not have a high permeability for small values of B. Mumetal®, on the other hand, has a very high permeability at low flux densities. Mumetal® is difficult to use and it is expensive. Ferrite materials have excellent permeability at high frequencies. Manufacturers of magnetic materials provide the designer with hysteresis curves that are related to the expected uses of that material.

The nonlinear relation between B and H means that if the B field is sinusoidal, the magnetizing current is not sinusoidal. As a magnetic material approaches saturation the demand for magnetizing current increases significantly. For a small power transformer the IR drop from magnetizing current can distort the voltage waveform at the secondary. In this case neither the B nor the H fields are sinusoidal.

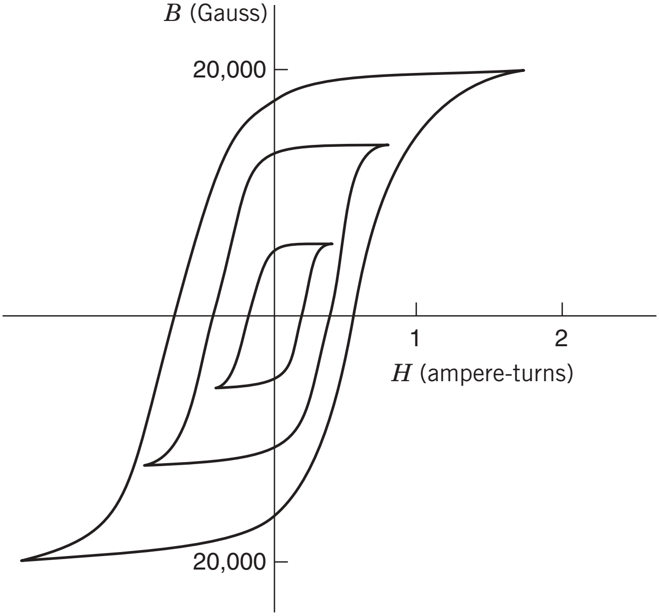


Figure 2.10. Typical hysteresis curves

In a transformer the point of maximum flux density occurs at zero voltage crossings. If there is core saturation, there will be excess magnetizing current flowing just before these voltage crossings. This can be used as a signature for this class of interference If there is voltage distortion at the peak of voltage, the problem is usually related to the peak current demanded by rectifier capacitors.

2.14. EDDY CURRENTS

Consider a closed circular path inside a conducting magnetic material. If the field crossing this loop changes, a voltage must result. This voltage in the material results in eddy current flow that dissipates heat. This induced current flows in circles around the changing lines of flux. One technique to limit this current flow is to construct the magnetic core from thin insulated laminations. Typically, at 60 Hz, 15-mil laminations are used, and at 400 Hz the laminations are 6 mils thick. These thin laminations break up these current paths without significantly reducing the effective core area.

For transformer applications above 400 Hz the preferred core material is ferrite. A ferrite is a made from finely powdered magnetic material suspended in a filler. When the mixture is fired the resulting material is similar to a ceramic insulator. The eddy current losses in this type of magnetic material are

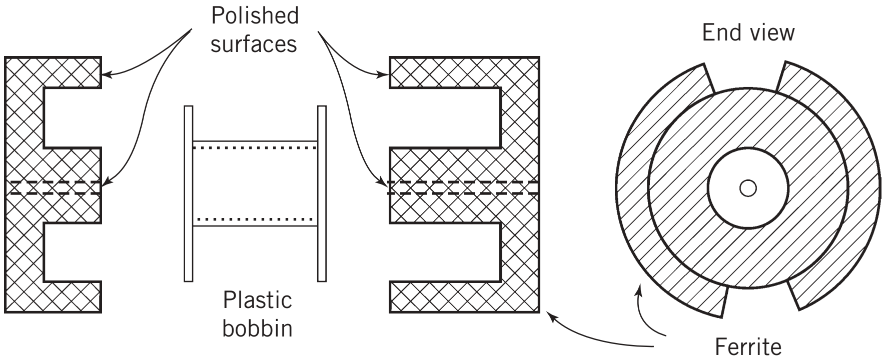


Figure 2.11. Ferrite cup-core construction

quite low. Dc-to-dc converters use ferrite core transformers because of their excellent high-frequency characteristics. Figure 2.11 shows a typical ferrite cup-core arrangement.

Transformer coils are wound on a bobbin that is mounted on the center leg of the core. Cup cores can also be supplied with a built-in center gap. This configuration is ideal for building inductors. As we have seen, a gap reduces the effective permeability of the core but provides for energy storage. In transformer applications a gap is not desired. To limit the gap the mating surfaces are carefully machined and polished. These cups are always supplied as mating pairs.

2.15. THE TRANSPORT OF ELECTRICAL ENERGY

It is usual to refer to conductors as carrying energy. This idea is reinforced in circuit theory by considering voltage and current and pairs of conductors. Conductors in themselves cannot carry or store energy. We have already discussed voltages and currents and their associated fields. We have seen that these fields can store energy. These same fields can also transport energy. Conductors can serve only to direct the path of energy flow. We have seen that both fields are necessary to place energy into a capacitor or inductor or to transfer power across a transformer. It also takes both fields to transport energy between two conductors. The two fields are present for a flashlight at dc or for a power distribution system at 200 kW. The fields associated with a flashlight are shown in Figure 2.12.

> **N.B.**
>
> Power is not carried in conductors. Power is carried in the space between and around conductors in the electric and magnetic fields.

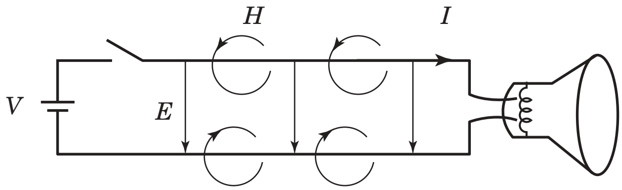


Figure 2.12. The electric and magnetic fields associated with a flashlight

> **FACT**
>
> The purpose of conductors is to direct the path taken by fields.

Information can take the form of a changing voltage. This change implies a changing electric field. A changing electric field implies a displacement current and an associated magnetic field. The magnetic field can be very small but it must be present.

> **N.B.**
>
> The movement of electrical information requires the presence of both an electric and magnetic field.

2.16. POYNTING'S VECTOR

The flow of electrical energy requires the presence of both electric and magnetic fields. Both fields are present if the energy is carried in free space or between two conductors. At any point in space the power crossing an increment of area is equal to the vector cross product of the E and H fields. This product is called *Poynting's vector*. It is a vector as it has intensity and direction at all points in space. The total power crossing a surface is the integral of Poynting's vector over that surface. Figure 2.13 shows Poynting's vector P for two conductors carrying power. The vectors E, H, and P are always at right angles to each other. The E field has units of volts per meter, the H field has units of amperes per meter, and the product has units of volt-amperes per meter squared.

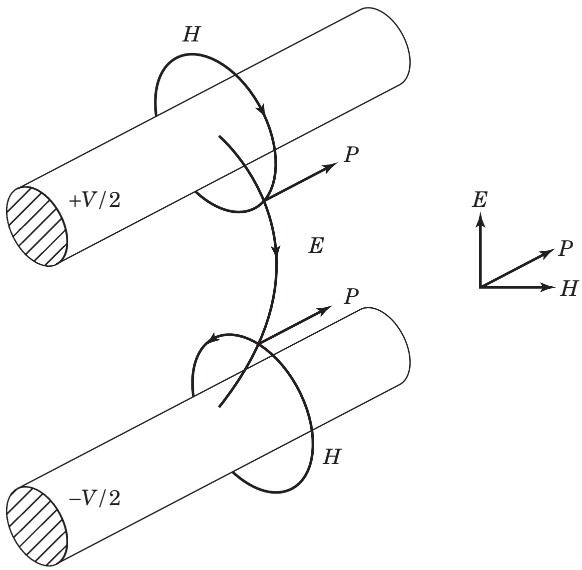


Figure 2.13. Poynting's vector for two parallel conductors carrying power

2.17. TRANSMISSION LINES INTRODUCTION

Transmission line theory was first developed for radio engineers to explain how to transport energy from a transmitting tube to a transmitting antenna. These transmission lines were often just pairs of open wires. The goal in design was to avoid reflections and to transport the maximum energy to the antenna for radiation. The theory is often presented using lumped parameter symbols that represent a distributed inductance and capacitance for the conductors. These symbols tend to reinforce the idea that energy is stored in the components, but this is not the case. The fields associated with open transmission lines concentrate between the conductors but some field energy does go out into space. As we will discuss later, some of this energy can escape as radiation.

The traces on a printed circuit board are small transmission lines. These lines radiate a small amount of energy. Because there are thousands of these traces on a single board this radiation must be considered. This subject will be covered in Chapter 7. The velocity of the wave is $(LC)^{-1/2}$, where the inductance and capacitance values are measured per unit length for a pair of traces.

Why does field energy follow pairs of conductors? The answer is very simple: It is easier to follow this path than to jump into space. For power transmission at 60 Hz very little energy leaves the space between the conductors and radiates. The fields follow the conductors wherever they go. At 400 Hz the inductance per foot limits the distribution of power to a few hundred feet. Above this frequency power must be transported inside a conducting conduit. This conducting conduit is called *coax*.

The energy flowing from a battery to a load flows in fields. When a new load is added to the circuit, a change in the field must propagate back to the battery. The battery then corrects the field and more energy is supplied. This means that there is a changing field in the space between the battery and the load. By Faraday's law, circuits that share the same space as the battery circuit will couple to this changing field. This coupling is called *interference*. This interference can be avoided if the field from the battery is confined to a known small volume. Without a field view of energy transport this coupling process would not be apparent. This is exactly the situation in an automobile. Demands for power are transported in fields. If these requests are step functions, the changing field can introduce interference into nearby cables and thus into hardware.

N.B.

Fields transport electric energy at all frequencies including dc. Changes in load can cause interference that can couple into nearby circuits.

2.18. TRANSMISSION LINE OPERATIONS

Consider the battery, switch, and transmission line in Figure 2.14. At the moment the switch closes, an E field appears across the line and charge begins to flow in the first increment of line capacitance. This flow of charge is a current that creates a unit of magnetic field associated with the first increment of inductance. This increment of inductance involves both the outgoing and return conductors. In the second increment of time the second increment of capacitance receives charge and the next increment of magnetic field is generated. There is a steady current flow as the same amount of charge is supplied for each increment of time. The net effect is a wave of field energy traveling down the transmission line filling the space with an E and H field. In a typical transmission line the wave velocity is about one-half the speed of light.

A fixed voltage source and a fixed current flow implies that a transmission line looks like a resistance. This resistance is called the *characteristic impedance* of the line. The term *impedance* is usually reserved for sine wave analysis but it is generally used by engineers for transmission lines without regard to waveform.

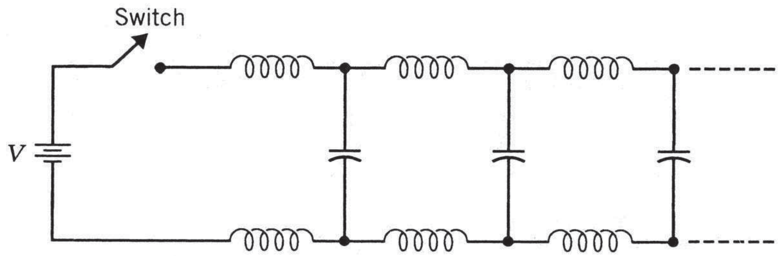


Figure 2.14. A battery, a switch, and a transmission line

2.19. TRANSMISSION LINE FIELD PATTERNS

The B and H field patterns around two parallel conductors carrying power are shown in Figure 2.15a. These conductors might be traces on a printed circuit board. A single conductor over a conducting plane is shown in Figure 2.15b. The field pattern above the conducting plane is the same as the pattern in Figure 2.15a. Notice how the E field terminates on the conducting plane beneath the conductor. This field termination indicates the pattern of current flow along this surface. This pattern is the same whether the voltage is for a logic signal or for power distribution. The characteristic impedance of the transmission line in Figure 2.15b is one-half that of Figure 2.15a.

Only one wave is shown in Figure 2.15. Any number of waves can use a transmission at the same time. Note that energy can be flowing in both directions at the same time.

N.B.

Energy cannot be canceled out. Energy must be dissipated in a resistance, radiated, or stored in another component. It must go somewhere.

The energy supplied to a transmission line continues to flow until it reaches the end of the line. This energy cannot be lost or destroyed. If the end of the transmission line is open, this energy cannot spill out into space. This forces a reflection, which means the energy continues to flow by returning back along the same transmission path. The reflected wave must cancel the current at the end of the line. Poynting's vector for the return wave has the same polarity of E field but the H vector must be reversed. This means the voltage at the end of the line must double. Before and after the reflection, the battery continues to supply energy to the line. When the reflected wave reaches the battery, the total voltage is incorrect. At this point the battery current goes to zero and a second reflection folds the wave energy back onto the line. If there were no

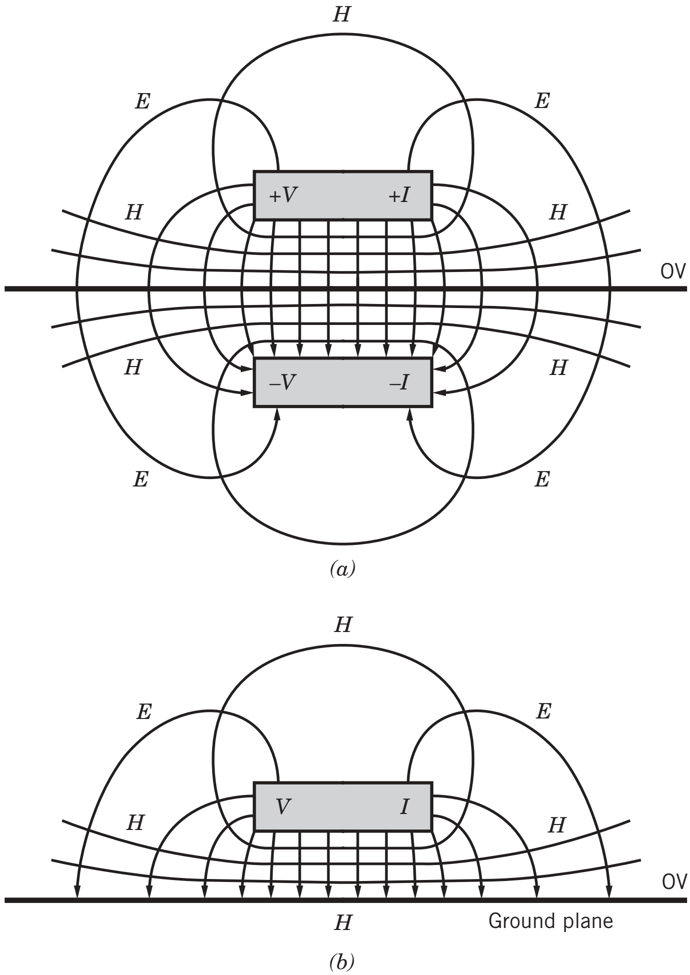


Figure 2.15. The B and H field patterns around a transmission line

losses, the voltage at the end of the line would appear as a square wave of double voltage. In the center of the transmission line a staircase voltage would be apparent. The voltage at the battery terminals would stay constant. In effect the field energy supplied to the line during the round trip simply sloshes back and forth. In practice the wave rapidly loses its character and in a few cycles the voltage settles to a steady value. The waveform for reflections assuming no losses or smearing is shown in Figure 2.16.

If the transmission line is terminated in a short circuit, the first reflection must cancel the voltage. The current behind the reflected wave is doubled. The second reflection from the battery adds new energy to the line. The current supplied from the battery is now triple what it was originally. After a second round trip the current is five times the initial value. This staircase of current

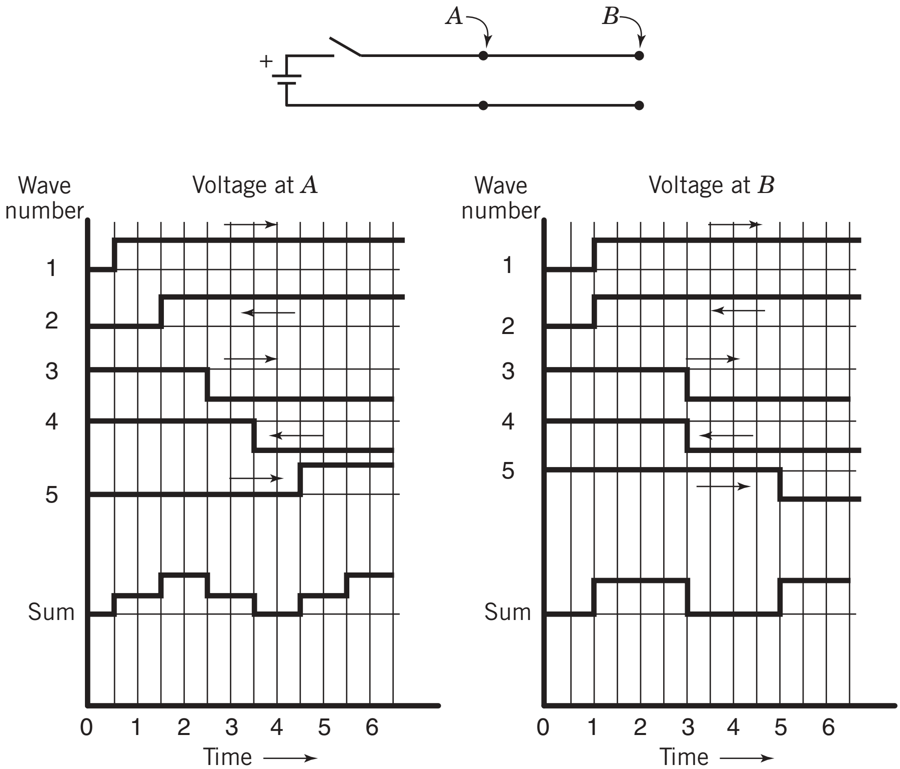


Figure 2.16. The voltage waveforms on an ideal open circuit transmission line

flow continues until a fuse blows or the wires melt. This staircase of current is how a short circuit develops.

> **N.B.**
>
> A voltage source cannot sense an open circuit or a short circuit until a reflected wave returns to the source.

When a transmission line is terminated in its characteristic impedance the first wave reaches the termination and there is no reflection. In typical digital circuits the terminations are nonlinear and the reflections are complex. The logic voltage must settle to an acceptable level value in one clock time so that the next logical operation can proceed. This settling process limits clock rates. More will be said about digital signals in Chapter 7.

2.20. INTERFERING FIELDS

In the next chapter we will discuss utility power and its effects on electronics. The radiation of power is discussed in Chapter 5. In this section we will discuss how fields couple to circuits. These fields result from both power distribution and various transmitters. We will discuss details of how these fields enter electronic hardware in Chapter 6.

All conductor pairs are transmission lines. Conductors that form pairs can include the earth, sheets of metal, conduits, shielded cables, open wires, building steel, telephone lines, and the power lines. All of these conductor pairs are irregular transmission lines with changing characteristic impedances. Many of their terminations are short circuits.

Field energy from remote sources will couple to these odd transmission lines because less energy is required to use a conductor geometry than to travel in free space. This energy literally bounces around reflecting from all the irregularities. Light entering a room does the same thing. It reflects and gets absorbed at all the many surfaces. The light intensity at any one point in a room would be very difficult to calculate. Similarly the field intensity in an arrangement of conductors is also very difficult to calculate.

Field coupling from local sources depends on the whether the source is electric or magnetic in character. Close to changing high voltages the coupling is usually capacitive in character. Near changing high currents the coupling is usually magnetic in character. At frequencies above 1 MHz the coupling from remote radiators can be approximated using either field intensity. At this point in our discussion it is sufficient to say that almost all coupling is proportional to loop area. Field coupling is discussed in more detail in Sections 5.8 and 5.9.

CHAPTER 3

Utility Power and Facility Grounding

3.1. INTRODUCTION

A large part of any engineering effort involves the use of utility power. Utility power can affect the performance of analog and digital systems. The environment we work in is full of power-related electromagnetic fields. This same environment contains the radiation from many distant transmitters that are also powered by a utility. Because power fields share the same physical space as analog and digital circuits or systems this energy can enter electronic hardware. For this reason design engineers need to understand the role utility power plays in electrical interference.

Power presents different problems on a spacecraft, in an automobile, in an aircraft, or in a laboratory. In this chapter we will look at electrical power for facilities. Many of the interference processes that are discussed in this chapter can be applied to every power-operated system.

3.2. HISTORY

When electrical power was first introduced there were few rules and regulations. There were frequent electrical fires and people received electric shocks from poor wiring. During lightning storms, power wiring provided a path for lightning to enter facilities. Under pressure from insurance companies and the National Fire Protection Association the National Electrical Code (NEC) was initiated. This code provided standards that brought needed control to a growing power industry.

Electrical power is central to our modern society. When the power fails our need is immediately apparent. Maintaining quality power is important. Local governments have accepted the NEC and now enforce its rules as the law. The code is modified regularly to accommodate new materials, methods, and procedures. The code represents good practice that has stood the test

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

of time. The code is not a scholarly work with a rationale behind each rule. Often a practice is accepted because its rejection would represent a significant hardship for the power industry.

Broadly the NEC provides rules that keep electrical installations safe from electric shocks and electrically caused fires. If a fault condition occurs, no one will receive a shock and a breaker will open the power circuit. The *listed* materials that are accepted for installation must provide service in all sorts of weather for an extended time period. Power wiring must be earthed so that lightning that couples to the power grid can find an earth path before it enters a residence or facility. The utilities ground their transmission lines to limit lightning damage to distribution hardware. Their interest is to provide uninterrupted power to their customers.

There are many ways to use the rules outlined by the NEC. Some approved practices are better at limiting electrical interference than others. The code makes no recommendations as to how the rules are to be followed. It is for this reason that engineers must understand the mechanisms of interference. With this knowledge decisions can be made about how hardware and facilities are to be designed.

3.3. SEMANTICS

Key words used by a power engineer are clearly defined by the NEC. These words must be clearly defined or the rules that follow can be misinterpreted. Circuit engineers do not have a controlling organization behind them and their language keeps evolving. The word *ground* to a power engineer means a connection to earth or its equivalent. The word "ground" to a circuit engineer may mean a power common on the secondary of a transformer or a reference conductor in a floating circuit. In order to discuss the role of power wiring in electronics, we will use the power industry's definitions given in this chapter. Later when we use these words to describe circuits we will be less restrictive. When we use a word in the power sense it will be in Italics. Here is a list of 12 key power-related words:

1. *Ground*. A connection to earth or its equivalent.
2. *Equipment ground*. All protective conductors that might contact power wiring. This includes conduit, equipment housings, racks, receptacle housings, trays, bare wires, and green wires. A green wire in power wiring is *equipment ground*. It does not carry power current. No other power conductor can be colored green.
3. *Grounded conductor*. A power conductor that carries current that is nominally at zero volts. It is usually colored white.
4. *Ungrounded conductor*. The power or "hot" conductor carrying voltage. The color code can be black or blue but never green or white.

5. *Neutral*. In three-phase power, the current-carrying conductor at zero volts. The *grounded conductor* is often called a *neutral* if it was derived from one leg of a source of three-phase power.
6. *Isolated ground*. An *equipment ground* conductor that returns separately from a receptacle to a *service panel* or a *service entrance*. It is always earthed.
7. *Service entrance*. The power entry point into a facility.
8. *Grounding electrode system*. All of the interconnected non-power conductors in a facility including building steel, computer floors, equipment grounds, rebars, gas lines, guy wires, etc.
9. *Feeder circuit*. A group of conductors carrying power to *branch circuits* with a protecting breaker.
10. *Branch circuit*. A power circuit providing power to various loads with a protecting breaker.
11. *Separately derived power*. Power taken from an auxiliary power source or a distribution transformer where a new *neutral* is supplied. This *neutral* is connected to the nearest point on the *grounding electrode system*.
12. *Listed* equipment. Hardware that has been tested and approved for electrical installation.

3.4. THE EARTH AS A CONDUCTOR

The earth is a complex conducting object. Resistances between two points on the earth can vary from 10 ohms to megohms. The highest resistance areas might be on a block of granite, in the dry desert, or on a lava bed. The lowest resistance areas might be in damp soil or at the seashore. By burying copper conductors and treating the surrounding soil chemically the resistance of an earth connection can be as low as one ohm at a few hundred cycles. Typical wet earth contact resistances are in the order of 10 ohms. The contact resistance can be measured by applying a voltage between two earth connections. One-half of the resistance observed can be attributed to each connection.

Lightning is an earth-seeking phenomenon. To provide lightning protection the NEC requires that for power entering a facility the *grounded conductor* (*neutral* conductor) must be earthed at the *service entrance*. The resistance to earth should not exceed 25 ohms. In poor soil situations this resistance may not be attainable. In this case two connections are sufficient to satisfy the code. When lightning strikes the power conductors an earth connection provides a path for lightning outside of the facility. If the utility supplies power underground, the earth connection is still required. This practice forces all facilities to be alike with respect to earth grounding.

The power definition of an earth connection assumes that the measure is made at 60 Hz. The discussion in Chapter 2 makes it apparent that a low-impedance earth connection using long round conductors is not generally practical above a few kHz.

Currents flowing in the earth and in the *grounding electrode system* of a facility are not easily controlled. Electrical interference is often blamed on this current flow. Any attempt to restrict this current flow by isolating a *grounding* area in a facility is illegal. The code prohibits having two *grounding electrode systems* in one facility. The reason is very simple: The resistance between two 10-ohm earth connections could be 20 ohms. If a fault condition connects an *ungrounded* or "hot" conductor to the second *grounding electrode system*, a breaker may not trip. A 20-ohm load at 120 V is a current of only 6 amperes. In this situation there could be power voltages between conductors that would normally be at *ground* potential. A fault condition could cause a significant shock hazard that might go unnoticed.

3.5. THE *NEUTRAL* CONNECTION TO EARTH

The NEC requires that the entry *neutral* or *grounded conductor* be earthed at the *service entrance*. This *neutral* may not be *regrounded* inside a facility. This insures that inside a facility, load currents do not flow in the *grounding electrode system*. The fault protection system in a facility relies on this restriction.

The NEC requires that an *equipment grounding conductor* follow each power conductor and be connected at each power receptacle. This grid of conductors is a part of the *grounding electrode system*. *Equipment grounding conductors* connect all receptacles, housings, conduit, and equipment racks together. An *equipment grounding conductor* must follow every power conductor where it is a part of the permanent facility installation. *Equipment ground* must surround all open power wiring. If a "hot" power conductor should fault to the *equipment ground*, the fault path must be low impedance to guarantee that a breaker will trip within a few power cycles. This is the reason why *equipment grounds* and power conductors must be run in parallel and close together. Any large loop in the fault path has inductance and this can limit the fault current.

N.B.

The code allows conduit to be used as an *equipment ground* provided the conduit and hardware are *listed* and approved. In industrial facilities it is preferred practice to run a green *equipment grounding conductor* and still use the conduit as *equipment ground*.

Equipment grounding conductors are often connected to earth at many points. Examples might be water pipes, boilers, motor housings, building steel, and metal siding.

N.B.

There can be only one earth connection for a *neutral* power conductor in a facility.

3.6. GROUND POTENTIAL DIFFERENCES

A signal cable that is connected to a signal source at a remote point is associated with the ground at that remote point. When this unterminated but *grounded* cable is brought to a local circuit, the potential difference between ground points can be observed on an oscilloscope. A potential difference means that there are electromagnetic fields crossing this area between the cable and the earth. The voltage is simply the field intensity times the loop area formed by the cable, the earth, and the oscilloscope. If somehow the area could be made zero, there would be no voltage to measure. In the case of the earth, the currents associated with fields are distributed beneath the surface of the earth and there is no way to reduce this area to zero. If there is a continuous conductive plane involved, the loop areas and thus the voltage coupling can be controlled. The subject of ground planes is covered in Chapter 8.

In a facility there are potential differences between racks and between pieces of hardware. These potential differences are again the result of electromagnetic fields in the area. There will be surface currents on all open conductors as a result of these fields. It is usually incorrect to use Ohm's law and assume a current flow based on a voltage and resistance measure. As an example, the noise voltage between two racks could be 1.0 V. The dc resistance between bonded racks could be one milliohm. By Ohm's law, current flow would be 1,000 amperes. This is obviously not the case. The same thing happens on the surface of a printed circuit board. A voltage between ground points represents mostly field coupling, not current flow. More will be said about this in Chapter 7.

Measuring resistances between ground points with an ohmmeter has its complications. Digital or active voltmeters may malfunction if there is power current flowing in the conductors. Often the contact resistance of the metering leads is greater than the resistance to be measured. One method of measurement is to use a four-terminal approach. A known current is applied to two outer contacts and the voltage difference between two inner points is observed. This method avoids the problem of contact resistance. The resistance is the inner voltage divided by the current.

The resistance between two points depends on how the current distributes itself on the conductor. As an example the resistance between two points on a square conducting plate will depend on the location and area of the contact points. Since the current does not flow in the corners the shape or thickness of the corners cannot affect the resistance.

N.B.

The resistance between two points depends on how and where the current is inserted.

3.7. FIELD COUPLING TO POWER CONDUCTORS

Consider power transmission lines carried on power poles. A field moving in the direction of the transmission line can couple voltages to this line. There are two modes of coupling. The field that crosses the loop formed by the two conductors generates a normal mode voltage. The field that crosses between the pair of conductors and the ground plane causes a common-mode voltage. This coupling is shown in Figure 3.1.

Normal-mode coupling adds to the transmitted signal. Common-mode coupling affects a group of conductors and appears as a ground potential difference. Ground potential differences can be attributed to current flow alone but usually the voltages are the result of field coupling to the meter leads. Often there is current flow but the potential difference cannot be used to calculate this current.

A common-mode signal is the average signal coupled to a group of conductors. Normal-mode coupling affects each conductor pair separately.

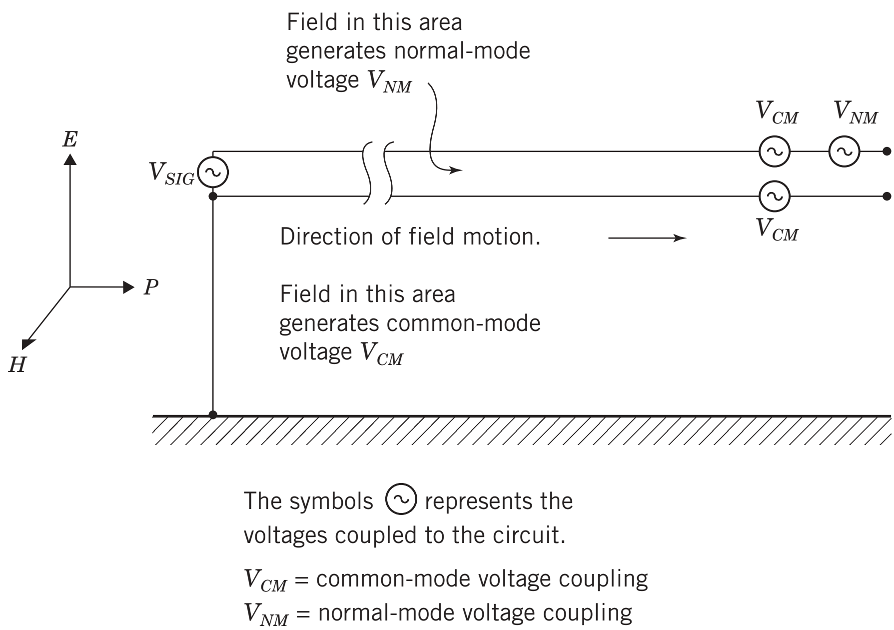


Figure 3.1. Normal-mode and common-mode field coupling

Normal-mode is also called difference mode, differential mode, or transverse mode. Common-mode is also called the longitudinal mode. More will be said about common-mode voltages in Chapter 4.

Common-mode fields couple to power conductors as well as signal conductors. For power this coupling includes the *neutral* and the *equipment grounding conductor*. Coupled fields move energy in both directions. There are reflections that occur along the line as well as at the ends. These reflections do not affect the power transmission but they do add to the field energy that is brought to the hardware via the power connection. Power line filters can reflect common-mode and normal-mode interference at frequencies above 100 kHz. The filtering of power line harmonics requires large and expensive filters and is generally not practical.

There are many ways for field energy to couple into hardware. Power line filters can limit this coupling but they are only a part of the story. In Chapter 6 the coupling mechanisms are discussed in detail.

3.8. *NEUTRAL* CONDUCTORS

Utility power is generated three-phase because the power generated for a balanced load is constant over the entire cycle. Constant power generation means the torque on the rotor of a generator is a constant over each rotation. Heavy equipment such as large fans and industrial motors are connected directly to the three phases. For most low power applications power is taken from phase to *neutral*. The loads are arranged so that each phase supplies the same amount of power. If the loads were linear and balanced the *neutral* currents would average zero through out each cycle.

Most electronic devices use a rectifier system that stores energy in filter capacitors. The capacitors demand current near the peak of voltage. These peaks of current occur at different times for each phase with the result the *neutral* currents cannot balance to zero. This *neutral* current is rich in harmonics and it flows in the reactance (series inductance plus the series resistance) of the *neutral* conductor. The resulting voltage drop in the *neutral* conductor can be seen as a potential difference between *the equipment ground* and the *grounded conductor*. This voltage is in series with leakage capacitance in the power transformer and causes current to flow in the input common conductors. See Figure 3.2.

If a distribution transformer handles loads with high harmonic content, the magnetic leakage field around that transformer must carry this same harmonic content. This means the nearby B field is rich in harmonics. Any high-frequency content in the field can more easily couple to nearby conducting loops. If the building steel forms a nearby loop, these induced currents can circulate in an entire facility. It is good practice to break up any nearby conducting loops of steel by using insulators.

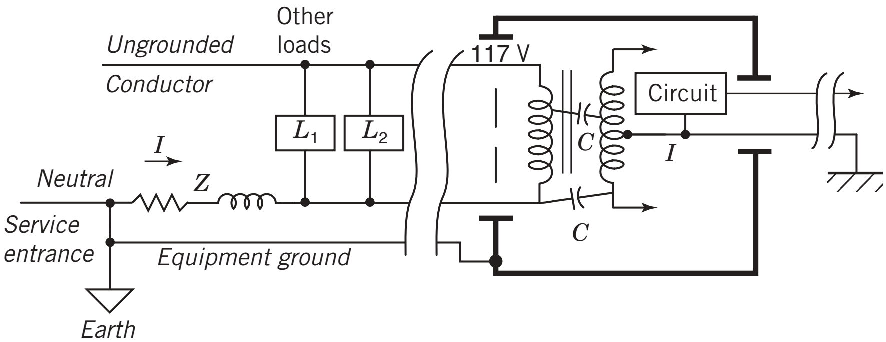


Figure 3.2. The coupling in a transformer resulting from neutral voltage drop

In three-phase power, if the harmonic content is high, the *neutral* current can be greater than the line current. *Neutral* conductors that carry high harmonic current can overheat. If the power is carried in a delta connection, then there is no *neutral* conductor to consider. This approach is more expensive as it requires an added *separately derived* source of power. This added transformer must have a *grounded neutral.* In effect a *separately derived* source of power is like a new *service entrance*. Of course, the loads on this secondary can again generate a *neutral* voltage drop.

3.9. *K*-FACTOR IN TRANSFORMERS

Electronic loads are typically very nonlinear. Rectifier systems require large amounts of current at the peak of voltage. This nonlinear current can be described in terms of harmonics of the fundamental. The effect on transformers is to increase eddy current losses in the transformer iron and increase heating losses in the wiring. Transformers used to supply electronic loads must be rated to handle the harmonic content or they will overheat. This problem occurs often enough that a special factor is used in specifying these transformers. This factor is known as the *k* rating. A high *k* rating requires transformers be built with a higher grade of transformer steel and with larger wire sizes. To limit skin effect the wire may be laminated.

There are several ways to calculate the *k* rating. The simplest method is to use Eq. (3.1). In this equation n is the harmonic number, I_n is the current at this harmonic, and I_T is the total load current:

$$k = \sum_{n=1}^{n} n^2 \left(\frac{I_n}{I_T} \right)^2 . \tag{3.1}$$

If there is no harmonic content, the *k* factor is unity. Harmonics out to the 25th should be considered in calculating *k*. Typical *k* factors for electronic loads are around 8 while factors above 30 are rare.

3.10. *UNGROUNDED* POWER

There are a few installations where *ungrounded* power is required. An example might be power used to heat a large crucible. If there is a fault and power is interrupted, the crucible might be lost. The procedure is to detect the fault and turn the power off when the crucible is empty. Facilities that use *ungrounded* power must have personnel on duty that are properly trained to handle any fault condition that might occur.

The power supplied on shipboard in the navy is floating. Floating power is electrically noisy. When power is switched on or off to a load the distributed capacitances in the entire system must store different amounts of energy. The transients that result enter every piece of equipment through the power transformers. The approach used to avoid this interference is to add a distribution transformer that supplies *grounded* power to sensitive electronic hardware.

3.11. A REQUEST FOR POWER

What happens when a power switch is closed? The answer may surprise you. Before the moment of closure the electric field between the *grounded* and *ungrounded conductors* stores electrostatic field energy. At the moment of closure a wave at half voltage starts moving at half the speed of light in both directions. The current in the wave depends on the characteristic impedance of the line. When the wave reaches the load a reflection takes place based on the impedance of the load. Meanwhile a wave of half voltage propagates back toward the *service entrance*. The wave progresses until it reaches a branch connection. At this point there is another reflection and the wave propagates along two paths. The process of branching and reflecting soon fills the entire wiring grid in the facility with electrical activity. The electrostatic energy stored in the wiring begins to move toward the load to supply it with power.

The electrical activity immediately after the switch closure involves numerous reflections that combine to propagate a wave that eventually reaches the power generator. The transient wave at the *service entrance* no longer has a steep wave front. Because of these numerous reflections the transient wave loses its amplitude. It is interesting to measure the transient voltages on a power line at different distances from a switch closure. Near the switch the voltage drops to half value for a few nanoseconds. At a nearby panel the voltage transient might be a volt lasting a microsecond. At the *service entrance* the transient is hardly visible. A wave must eventually reach the power generator so that it can adjust its output voltage to accommodate the new load.

Consider the electrostatic energy that is available in a facility. Assume a capacitance of 50 pF per foot of wiring. In 1 microsecond a wave will propagate out a distance of 500 feet. The network of wiring within this radius might consist of 3,000 feet of wiring. The wiring capacitance might be 0.15 μF. The energy stored in this capacitance when the voltage reaches 120 V is about 10^{-3} joules. If this energy were dissipated in 1 microsecond, the power level would be 1,000 watts. If the load request is 100 watts, the voltage at a distance of 500 feet would sag by about 5%.

The energy that bounces around the power wiring circuitry moves in the space between power conductors. Some of this energy is confined to the space inside metal conduit. Some of this energy adds to the ambient field in a facility.

Very-high-speed voltage changes on the power line have no effect on the rectified and filtered supplies inside the hardware. There are still several coupling mechanisms that allow fields to enter the hardware that involve the power transformers. If the power filters are not installed correctly then this is an entry point for interference. This entry mechanism is discussed in Chapter 6.

3.12. EARTH POWER CURRENTS

Power wiring that distributes power between facilities is often 4-wire. This means the *neutral* conductor is carried along with the three phases. The *neutral* is earthed along the distribution path as a protection against lightning. The amount of *neutral* current that flows depends on load balance and on harmonic content. The *neutral* current that flows divides between the earth and the *neutral conductor*. The current density between earth points is low where the current spreads out in the earth. A wide current path means a low impedance path. A typical earth resistivity might be 1,000 ohm-cm. Using this figure the resistance across a cube of soil 10 meters on a side is only 1.0 ohm.

In areas using electrical power there are metal objects buried in the earth that reduce the impedance between points in the earth. Gas lines, metal fences, and building steel are good examples of buried objects. Building steel is earthed at many points for lightning protection. *Neutral* current flow will follow these conductors as they provide a lower impedance than earth. Note that the *neutral* current flowing in building steel could be associated with power flowing to a nearby facility.

3.13. LINE FILTERS

Electronic hardware is often supplied with line filters. These filters place capacitors between the power leads and *equipment ground*. *Equipment grounds* form a grid that can be earthed at many points. This means that reactive filter currents can use parallel paths to return to the *neutral* at the *service entrance*.

One function of a line filter is to keep interfering fields from entering the hardware. The filter works by providing a high-impedance path into the hardware and a low impedance path around the hardware for high-frequency energy. Filter capacitors circulate interference currents as well as 60-Hz currents in the *equipment ground* grid in a facility. All the return paths lead to the *service entrance* where *equipment grounds* connect to the *neutral* conductor and earth. If there are many parallel paths for current flow, the field intensity will be low. The number of filters is proportional to the number of hardware items involved. In large systems the reactive currents returning to the *neutral* tie at the *service entrance* can be amperes.

Line filters can provide a local source of immediate energy for hardware. If the hardware requires energy in a step manner, the capacitors in the local power filter can supply this leading-edge energy. This stops a high-frequency request from propagating a steep wave out into the facility. This energy is not available if there is a series inductor on the load side of the filter.

N.B.

The 60-Hz current in a filter capacitor is reactive, which means it is not in phase with the voltage. This fact can be used in troubleshooting to identify the source of current flow.

Power line filters are discussed again in Section 6.7.

3.14. *ISOLATED GROUNDS*

An *equipment grounding* conductor is provided in parallel with power conductors to all hardware using power. The line filter in each piece of hardware connects capacitors from both sides of the line to this *equipment ground*. In racks of hardware the *equipment grounds* also connect to the rack. In standalone hardware there is no rack to interconnect the *equipment grounds*. In this case the *equipment grounds* are interconnected at the power receptacles.

In the *isolated ground* case the *equipment grounds* are not tied together even at the receptacles. Instead, each *equipment ground conductor* goes back separately to a panel or to the *service entrance*. These parallel paths are long and therefore inductive. If there are signal paths that interconnect the hardware, power filter currents will flow in this cross connection. This can couple interference into a signal process depending on the nature of the interconnecting cable. This problem is shown in Figure 3.3.

The noise generated in one piece of hardware can contaminate an entire facility. The rationale for allowing *isolated grounds* is to keep the contaminating current from one piece of hardware from flowing in the *equipment ground conductors* of other hardware. Unfortunately the current flowing in an *isolated conductor* can create fields that fill an entire facility. The problem is not the current flow but the fields that are created by the current flow. In nearby conductors the fields imply surface currents. If every *equipment ground* has a long return path, the fields in the facility are much greater. The fields associated

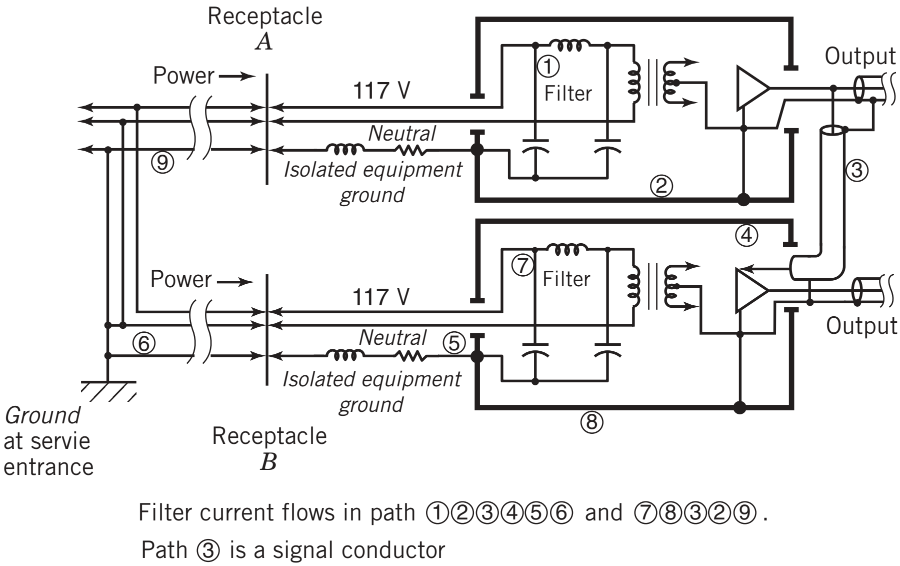


Figure 3.3. Filter current flowing in signal interconnections

with a grid of *equipment grounding conductors* are less intense. If the current spreads out on a ground plane the field intensity can be further reduced.

> **N.B.**
>
> A grid of *equipment ground conductors* behaves like a pseudo-ground plane.

A single *equipment ground conductor* running a long distance is a high impedance. If this impedance is in series with a line filter, it limits filter performance at high frequencies. This in turn allows line voltage interference to enter the hardware and this defeats the purpose of the line filter. This is another reason to use the standard grid of *equipment ground* connections.

In standalone hardware, power connections are often made to separate power receptacles. In the *isolated grounding* arrangement the loop formed by the *equipment grounds* can be large. Assume a signal path between the two pieces of hardware closes this conducting loop. Transient fields from nearby current interruptions in motors can couple to this loop and induce common-mode voltages. At a signal interface these voltages can be great enough to damage hardware. The standard *equipment ground* configuration limits this loop area. The result is a more reliable installation.

3.15. FACILITY *GROUNDS*—SOME MORE HISTORY

The ideas behind a central facility *ground* probably started in the early days of electronics. In designing hardware it was found that using a water-pipe ground reduced hum coupling in analog amplifiers. There was evidence to suggest that a better earth ground reduced the interference further. The exact reasons were not debated because the solution seemed obvious.

It was common practice in circuit design to use a single-point grounding scheme. The idea was to control cross coupling due to the flow of current in common conductors. The common conductors included power-supply conductors, input signal leads, output common leads, shield connections, and the *equipment ground*. These conductors were stacked up on a common grounding stud mounted to the chassis. The order of connection was carefully considered. This arrangement of grounds is called a "star connection." As an example, a star connection insures that output current does not flow in an input common conductor. Also current flowing in an *equipment ground* would flow directly to the hardware chassis without flowing in a signal conductor. A star configuration works fine if the dimensions are a few hundred centimeters and the frequencies of interest are under 20 kHz. In today's digital world this practice cannot function. It cannot function at the circuit card level or at the facility level.

N.B.

The use of a star connection has its place. The connection of the *neutral*, the *equipment grounds*, and the earth at the *service entrance* is a star connection.

Circuits designers like to think of a ground as being able to "absorb" noise. After all, capacitors connecting circuit points to this ground seem to "drain" the noise away. It seems obvious that the earth is the ultimate current sink. Capacitors have this ability to bypass noise.

In the 1950s, the electronics industry received a big boost from aerospace and military activity. Many large electronic facilities were built in this period. The idea of an earth ground was extended from hardware design to an entire facility. A typical facility might contain hundreds of electronic devices. If a good ground made an instrument quiet, then a "really good earth ground" would make a facility really quiet. As a result many facilities were built with a very expensive and extensive grounding scheme. The star connection idea was extended to a grounding well where an earth connection was made using large amounts of copper and chemically treated soil. All signal shields were assembled at several nodes and carried as one large conductor to the earth ground. A similar treatment was made for all *equipment grounds* and signal commons. In these schemes the power was still supplied to a *grounded service entrance*. This earth connection was separate from the ground well.

Bringing all the *equipment grounds* via a collecting node to this one star *ground* point added a significant loop area to the fault protection path. It was argued that this departure from the code was needed to provide for a quiet facility. This practice added a loop area to every return path for filter current. This adds to the interference fields in a facility rather than making it quieter.

The premise that noise currents flow into earth and never return is of course not good engineering. I call this the "sump theory of electronics." Circuit theory requires a return path for currents that go into the earth. Where the return paths are, nobody knows. Engineers seem to brush this issue aside, assuming "Somebody knows more than I do." Once a complex grounding scheme is installed it is difficult to disable it. It is also very difficult to experiment with an entire facility to prove or disprove its quality.

N.B.

A facility is not a circuit. It is a complex of conductors where fields associated with various voltages and currents follow the paths that store the least amount of field energy.

A central grounding conductor that "carries all interference current to earth" is a giant upside-down antenna. The interference fields around this ground are propagated through an entire facility. To stop this radiation the facility must be built as a solid sealed metal box. This is a difficult building construction problem.

3.16. LIGHTNING

When it rains, water droplets strip electrons from air molecules and carry this charge to earth. The charge in the atmosphere results in an electric field that surrounds the earth. The field strength at the earth's surface is on average about 100 V/m. We generally pay little attention to this field because wherever we walk on the earth the potential goes to zero. Our bodies are good conductors in constant touch with the earth. For this reason we do not sense this voltage gradient. With sensitive instrumentation this voltage gradient can be measured.

In areas of weather the electric field intensity increases and the field strength around sharp objects can be quite high. If the air ionizes around this object, the ionization path extends the sharp point upward. The voltage gradient between the clouds and the tip of the ionization path increases and this accelerates the ionization process. Eventually a narrow ionized path is established from earth to a cloud of ions. At this point there is an avalanche of charge that flows down the prepared ionized path. The electric field in the ion cloud stores the electric field energy that uses the ionized path as a conductor. It is as if a long conductor were discharging an immense capacitor storing charge.

The volume in space that is discharged extends about 150 meters around the ionized path. At the speed of light it takes about half a microsecond for this field to collapse. The pulse of current that flows can vary from a few thousand amperes to a maximum of 100,000 amperes. The first pulse of current widens the ionized path. After this current pulse stops, the field pattern readjusts in the cloud and a second and third pulse of current can follow using this same path. When the path breaks up or the voltage gradient is too low the lightning activity stops.

N.B.

Lightning is the mechanism that keeps the charge in balance in the atmosphere all over the earth.

3.17. LIGHTNING AND FACILITIES

When lightning strikes the earth the current spreads out on the surface of the earth. Because of skin effect the electromagnetic field cannot penetrate deeply into the earth. The voltage gradient near the strike can be great enough to electrocute cattle standing in a wet field.

Utilities should bring services into a structure at the same point. Examples might be power, cable, and telephone lines. If separate grounding points are used, a nearby lightning strike can cause large ground potential differences between these utilities. If these circuits are brought together into the same hardware, there can be a severe shock hazard or the chance of damage to the hardware.

Lightning and circuit theory have little in common. The electromagnetic field around the pulse is complex. It is this rapidly changing field that controls the path the lightning will take. Our best tool is to describe this current path in terms of inductance. At the moment the main pulse reaches the building the voltage to earth is near zero. The energy that flows through the building can be thought of as a wave in a coaxial transmission line. The lightning path is the center conductor. The outer conductor (return path) is the displacement current flowing in the surrounding space. A displacement current results from a rapidly changing E field.

The voltages in various conductor geometries can be calculated assuming a rate of change of current in an inductance. The pulse of current rises in about 0.5 μs. A 10-inch-long conductor has an inductance of 1 μH. A 50,000 ampere pulse develops a voltage of 100,000 volts in this inductance. In a building the voltage could easily reach a million volts per floor.

When voltages of this magnitude are involved, air can break down and conduct. When lightning hits a steel structure such as a building under construction, one can see lightning jump between girders rather than follow a horizontal beam out to a vertical beam. The reason is simple. A right-angle path is inductive and the resulting voltage will ionize the air. The air breakdown starts at the beam intersection and moves out into the surrounding space. When lightning hits the roof of a building it should be directed to the walls of the facility and to earth on parallel "down" conductors. The added lightning paths should be lower impedance than air ducts, plumbing vents, or antennas.

Where possible, down conductors should provide straight paths without bends or turns. If there must be turns, they should have a reasonable radius. A wide strip of metal is a far better down conductor than a large round conductor. In buildings with sheet metal sidings the sidings provide a far lower impedance path than a group of down conductors. In this case the down conductors may actually be redundant. If sheet metal is used, it should be earthed at multiple points along the bottom edge to limit arcing to earth near the floor of the facility.

Lightning can develop a high induction field near a down conductor. In a building with a steel framework, electronics should be positioned away from this steel to avoid possible damage. To illustrate this problem the H field at a distance of 1 meter from a 50,000 ampere pulse is almost 8,000 amperes per meter. The B field is $\mu_0 H$, which equals 0.01 teslas. The flux in a conducting loop of 0.01 m^2 is 10^{-4} lines. If the flux rises in 0.5 μs, the induced voltage is 200 volts. The induced voltage near sheet metal siding carrying this same current is probably reduced by a factor of 100.

N.B.

Lightning current does not need to flow in a circuit to do damage. A nearby pulse can induce sufficient voltage into a conductive loop to damage circuitry.

CHAPTER 4

Analog Circuits

4.1. INTRODUCTION

The expression *analog circuit* can mean any electronics that is nondigital. An rf transmitting circuit can be considered an analog circuit. This chapter is devoted to analog circuits that operate below 100 kHz. This includes instrumentation amplifiers, signal conditioning, audio circuits, medical amplifiers, and power supplies.

The availability of integrated circuits has simplified many aspects of analog design. The basic problems of measurement and instrumentation are not addressed by this new technology. Problem areas include system interconnections, interference field coupling on long signal lines, rejecting ground potential differences, and circuit stability. The general problem of handling analog signals is called *signal conditioning*, which includes filtering, providing offsets, bridge balancing, controlling gain, common-mode rejection, transducer excitation, and calibration. The circuitry for this conditioning is often integrated into any required amplification. If a signal has sufficient resolution, much of the conditioning can be handled in software, which eliminates the need for circuitry. This chapter treats the problems of handling signals so that interference does not introduce errors.

4.2. INSTRUMENTATION

Measurements of temperature, strain, stress, position, and vibration are an important part of many development efforts. Aircraft landing gears, missile housings, helicopter rotors, and turbine engines are examples of structures that undergo extensive testing. Transducers associated with these measurements are mounted on a test structure or vehicle. Some of these transducers require external excitation while others are self-generating. Some transducer types are electrically bonded to the structure under test and others are

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

electrically floating. The signals generated by these transducers must be conditioned and then recorded for later analysis. The amplification can be on the structure under test or it can be located at a separate location.

There are many ways for an analog signal to be compromised. As an example, it makes no sense to use digital signal filters if the signal has already overloaded the circuitry. If shields are not terminated correctly, software cannot remove the resulting interference. If common-mode signals are not rejected properly, significant signal errors are the result.

N.B.

If the answer is unknown, it can be very difficult to verify that the measurement is valid. For example, signals that overload an input stage and are then filtered are just noise.

There are semantics problems in discussing analog electronics. Here is a glossary of key words that will help in understanding the discussions that follow.

DEFINITIONS

Balanced signal(s). Two signals measured with respect to a reference conductor whose sum is always zero. For example, a center-tapped transformer produces a balanced signal. The centertap can be called the reference conductor and the voltages on the transformer terminals $+V_{sig}$ and $-V_{sig}$ are a balanced signal. The signal generated by active arms in a Wheatstone bridge can be a balanced signal.

Common-mode voltage. The average interfering voltage on a group of signal conductors measured with respect to a reference conductor. Usually a ground potential difference. In telephony a common mode signal is called a *transverse-mode signal*.

Difference signal. The voltage difference of interest.

Differential signal. The voltage difference of interest.

Instrumentation amplifier. A general-purpose differential amplifiers with bandwidth from dc to perhaps 100 kHz and variable gains from 1 to 5,000. This instrument might provide transducer excitation and signal conditioning.

Normal-mode signal. The signal of interest. In telephony the normal-mode signal is called the *transverse-mode signal*.

Reference conductor. Any conductor used as the zero of voltage. In a power supply with voltages of 0 V, +15 V, –15 V, and –5 V, the conductor labeled

0V is the reference conductor. If a signal is measured with respect to a ground, then the ground becomes the reference signal conductor. In an analog circuit there may be several reference conductors. Because there are electromagnetic fields in the area, these reference conductors may differ in potential.

Signal common. A signal reference conductor.

Signal ground. A signal reference conductor.

Unbalanced signal. A single signal voltage measured with respect to a reference conductor. A single-ended signal.

The material in this chapter stresses handling analog signals. The interface between analog and digital signals is discussed in Chapter 7.

4.3. HISTORY

The first signal amplifiers were designed using vacuum tubes. The dc operating voltages often exceeded 200 volts and filament current had to be associated with each tube. Since vacuum tubes are not available as complimentary devices (e.g., NPN and PNP transistors) early amplifiers were ac amplifiers. Low-frequency signals were often mechanically or electrically modulated and amplified as a carrier. Strain gages were excited directly with a carrier signal. The amplification of carrier signals allowed the use of transformers that provided isolation between input and output circuits. After amplification the signals were demodulated and filtered. The end result was amplification with a limited bandwidth. Bandwidth using carrier techniques rarely exceeded 1 kHz. In this approach crosstalk was very difficult to control.

In telephony, dc amplification is not needed and transformers were used to provide isolation and to convert single-ended audio signals to balanced signals and vice versa. Balanced signals can be transported over long distances and the effects of common-mode interference are canceled. In audio work, transformers provided input-to-output isolation, which eliminated circuit common connections (ground loops). In instrumentation where dc gain was required, carrier techniques were used where transformers again provided isolation. Feedback around these transformers was usually not provided. Obviously circuits that relied on vacuum tubes and transformers left a lot to be desired.

When semiconductors were introduced it became possible to amplify signals with a dc component using well-balanced matched pairs of transistors. Techniques evolved that made signal isolation possible without the use of transformers. Today it is possible to handle microvolt signals or 100 V signals with bandwidths from dc to many megahertz. Millivolt signals can be amplified in the presence of 300-V common-mode voltages. Providing signal integrity without the use of signal transformers in the presence of interference is the subject of this chapter.

4.4. THE BASIC SHIELD ENCLOSURE

The circuit in Figure 4.1 has potentials labeled as follows: The input lead is V_1, the output lead is V_2 and a signal common or reference conductor *equipment ground* is labeled V_4. The conducting enclosure labeled V_3 is floating. It is convenient to label this as the reference conductor at zero volts. Every conductor pair has a mutual capacitance that we label C_{12}, C_{13}, C_{24}, and so on. When these capacitances are drawn out as a circuit in Figure 4.1b it is immediately apparent that if the circuit is an amplifier there is feedback from the output to the input. The circuit is shown in Figure 4.1c.

It is common practice in analog design to connect the enclosure to circuit common. This enclosure connection is shown in Figure 4.2. When this connection is made, the feedback is removed and the enclosure no longer couples signals into the feedback structure. The conductive enclosure is called a *shield*. Connecting the signal common to the conductive enclosure is called "grounding the shield."

Most practical circuits must provide connections to external points. To see the effect of making a single external connection, open the conductive enclosure and connect the input circuit common to an external ground. This ground can be any structure, earth or hardware common. Figure 4.3a shows this grounded connection surrounded by an extension of the enclosure. This

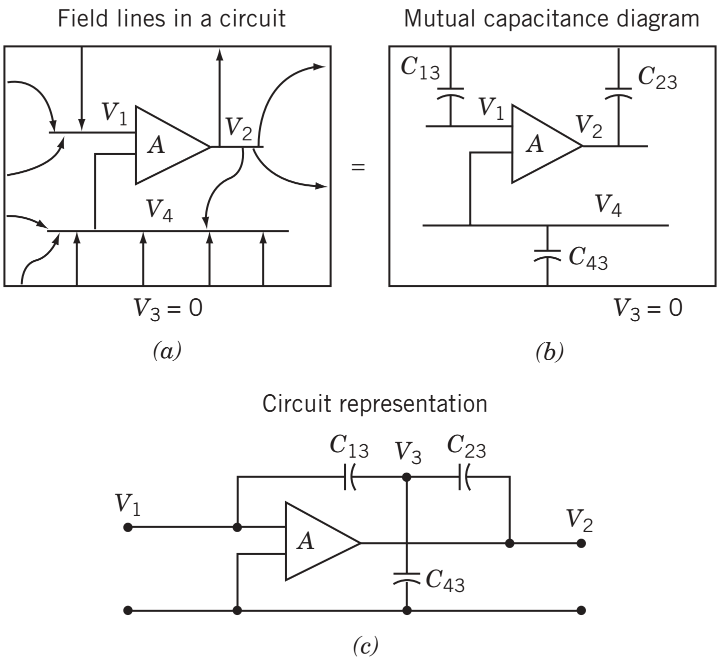
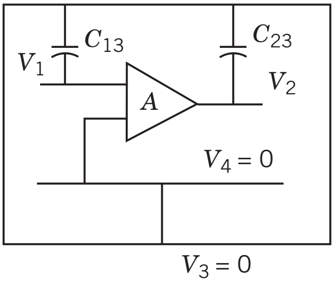


Figure 4.1. Parasitic capacitances in a simple circuit



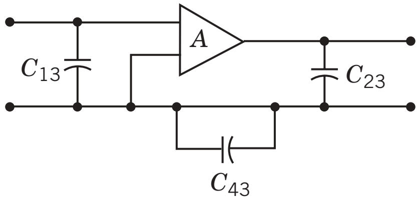


Figure 4.2. Grounding the shield to limit feedback

extension over the input conductor pair is called a *cable shield*. A problem is caused by the location of the connection between the cable shield and the enclosure in Figure 4.3a. The electromagnetic field in the area induces a voltage in the loop ① ② ③ ④ ① and the resulting current flows in conductor ① ②. If this conductor is the signal common, this lead might have a resistance R of 1 ohm. In this case every milliampere of coupled current would develop a millivolt of interference signal that will add to any desired signal. Our goal in this chapter is to find ways of keeping interference currents from flowing in any input signal conductor. To remove this coupling, the shield connection to circuit common must be made at the point where the circuit common connects to the external ground. This connection is shown in Figure 4.3b. This connection keeps the circulation of interference currents on the outside of the shield.

There is only one point of zero signal potential external to the enclosure, and that is where the signal common connects to an external hardware ground. No part of the input shield should be connected to any other ground point. The reason is simple. If there is an external electromagnetic field, there will be current flow in all exposed conductors. An incorrect shield connection will allow this current to flow in conductors inside the enclosure.

N.B.

An input circuit shield should connect to the circuit common where the signal common makes its connection to the source of signal. Any other shield connection will introduce interference.

4.5. THE ENCLOSURE AND UTILITY POWER

When utility power is introduced into an enclosure, a new set of problems results. The power transformer couples power as well as fields from the

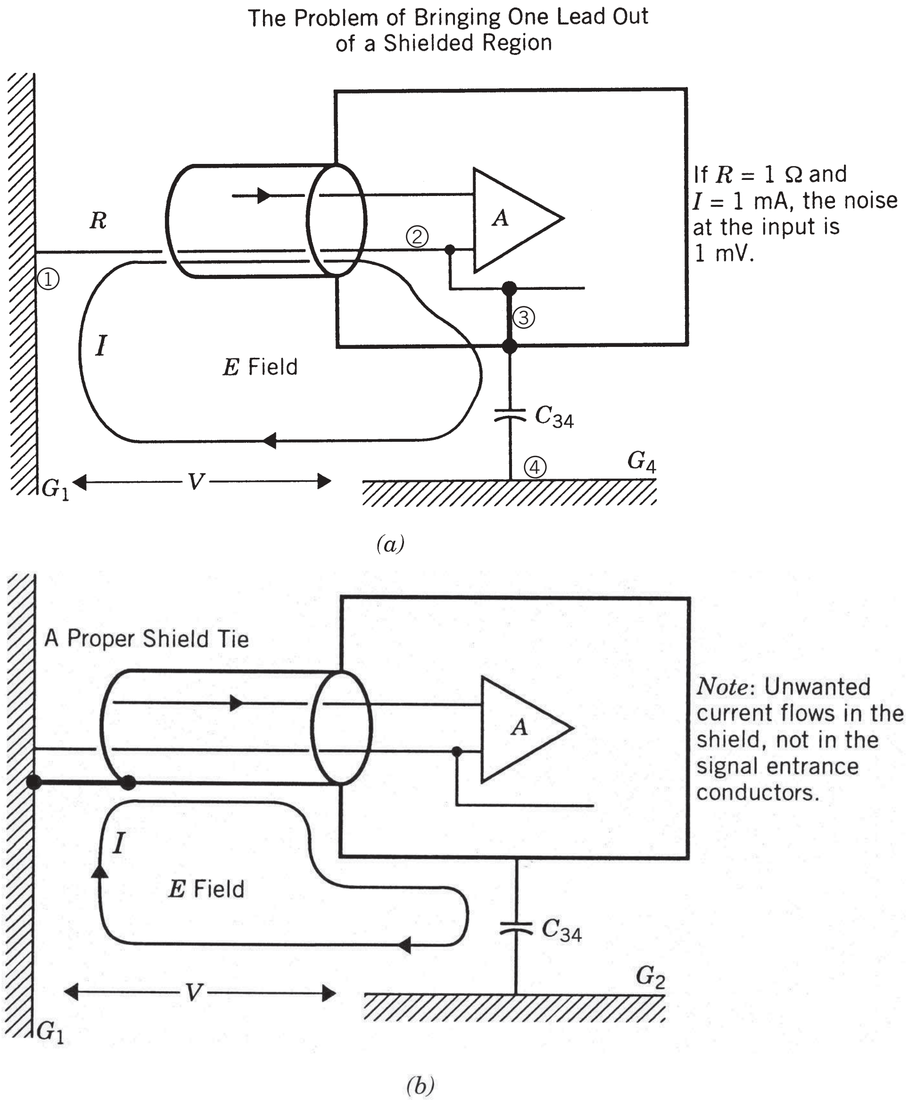


Figure 4.3. A single input connection from the circuit common to an external ground

external environment into the enclosure. The obvious coupling results from capacitance between the primary coil and the secondary coil. Note that the secondary coil is connected to the circuit common conductor. The basic offender is the 120 V on the *ungrounded* conductor. The reactive coupling is shown in Figure 4.4. Unwanted current now flows in the loop involving the utility ground, the primary voltage, and the input signal common.

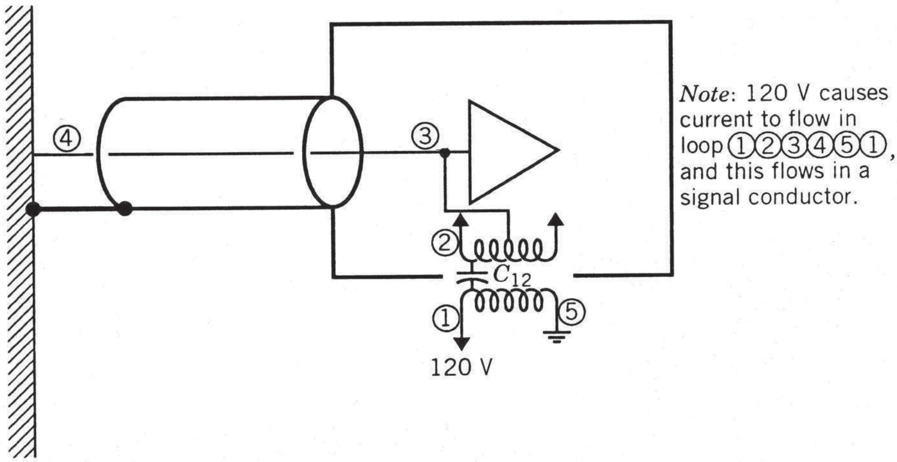


Figure 4.4. A transformer added to the circuit enclosure

The construction of power transformers usually begins by winding a primary coil on a bobbin. After a layer of insulation. the secondary coil is wound over the primary coil. This practice places one end of the primary coil next to one end of the secondary coil. Typically the interwinding capacitance is a few hundred picofarads. At 60 Hz this is a reactance of about 10 MΩ. If the "hot" lead is next to the secondary coil the resulting current at 60 Hz is 12 μA. For many applications, this level of current flow is not a problem. Of course if the *grounded* conductor is wound next to the secondary coil, the current flow at 60 Hz would be smaller.

The issue is often not the current flow at 60 Hz but the noise current flow at higher frequencies. This noise can result from three basic sources: the electromagnetic fields in the area, pulses or signals on the power line generated by nearby hardware, and neutral voltage drops. If the input common conductor is long, current flow in the lead impedance can cause a significant voltage drop. This voltage drop is a normal-mode signal that adds to the desired signal. In some cases, power line voltage spikes on signal conductors can be great enough to damage hardware. This type of interference often can be limited by power line filters or by diode clamps placed across the input terminals. Output circuits can be damaged as well as input circuits.

Out-of-band currents that flow in the input common conductor can generate signals that can interfere with the performance of both analog or digital hardware. Transient bursts can overload analog electronics resulting in offsets or "pops." The pop is often the recovery transient after an overload has occurred. If the out-of-band signal is a steady-state carrier, then there may be signal rectification resulting in dc offsets. To limit this coupling, small rf filters are placed at the input to analog amplifiers. A typical low-pass filter might be a series 100-ohm resistor and a shunt 500-pF capacitor located at each input base or gate. In digital circuits a voltage pulse can cause a logic error or in some cases damage the circuitry.

At this point in our development there are two circuits entering the circuit enclosure. The input common conductor and the power transformer are both associated with external grounds. Consider the case where only the output common and the power transformer enter the enclosure. The power transformer now circulates current to ground in the output common conductor. This current rarely causes problems because:

1. Output signal levels are usually greater than a few volts.
2. The voltage drop in the output common is not amplified.
3. The signal output impedance is low.
4. Output cable runs are usually relatively short.

The output signal common can be inside a shielded cable or it can be the shield itself. In this latter case the shield should be treated like a signal conductor, not an extension of the enclosure. If a two-conductor shielded cable is used, the shield can be terminated (grounded) at one or both ends. The preferred connection is where the signal common terminates. It is often practical to transport the output signal on open wires (ribbon cable) as long as there are no sensitive circuits in the vicinity.

4.6. THE TWO-GROUND PROBLEM

The circuit in Figure 4.4 has one grounded input conductor plus a power transformer connection. If the input and output circuit common are both brought out of the enclosure and grounded, the result is the familiar "ground loop." The fields in the external environment will couple to the loop formed by the common conductor connecting between two ground points. Currents flowing in this loop can flow in the transducer source impedance. For many applications the interference created by this ground loop can be larger than the signal of interest.

Ground loops can result when signal cables go between pieces of hardware that are connected to *equipment ground*. For example, in bench testing, the item under test and an oscilloscope form a ground loop. It is good testing procedure to isolate the oscilloscope with a "cheater plug" to avoid this loop. This plug simply breaks the *equipment ground* connection provided by the manufacturer. This third-wire safety connection is required by the NEC. If an isolated oscilloscope is connected to the "hot" or *ungrounded* power line, the shock hazard is obvious. It is good practice to place a warning label on equipment that is used *ungrounded*.

There are many systems made up of interconnected pieces of hardware. These systems may be analog or digital in nature. If the hardware is rack mounted, then the *equipment grounds* connect the hardware together. For digital hardware, most of the problems discussed in this chapter will not be an issue because the signal levels are several volts.

4.7. INSTRUMENTATION AND THE TWO-GROUND PROBLEM

The basic analog problem is to condition a signal associated with one ground reference potential and transport this signal without adding interference to a second ground reference potential. Consider the arrangement in Figure 4.5. The input and output circuits have been separated so that the input common grounds at the source of signal and the output common grounds at the termination of the output signal. This unbalanced signal source represents the most difficult problem in instrumentation. The ground potential difference between the two enclosures causes current to flow in the unbalanced source resistor R_1 and the impedance Z_1 in enclosure 2. The current is limited by the value of Z_1.

The balanced signal source is usually a strain gauge. This type of signal is discussed in Section 4.8. Low-impedance sources such as thermocouples are discussed in Section 4.10.

Input signal conductors should be guarded right up to the input base or gate of the input amplifier. The guard shield in this arrangement is not a circuit conductor. It is an electrostatic shield. On a circuit board, added traces can be used to guard the signal instead of a shielded cable.

In this approach the second enclosure has a high input impedance differential amplifier that provides all the needed gain. There is circuitry between the input leads and the amplifier input. This circuitry is really a part of the input enclosure so it must be well guarded. It includes diode clamps, high-impedance conductive paths for gate or base currents, and any input filters. The detailed treatment of the input circuit is not shown.

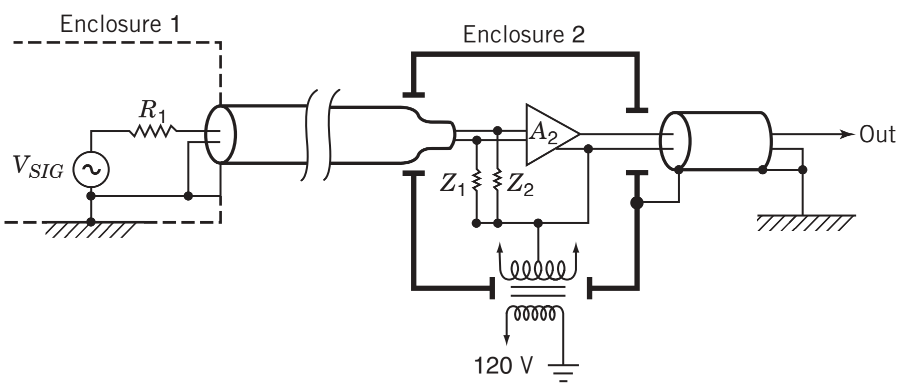


Figure 4.5. The two-circuit enclosures used to transport signals between grounds

Consider instrumentation in Figure 4.5, where the gain is 1,000 and the input unbalanced resistance R_1 is 1,000 ohms. If the output error limit is 10 mV and the input error is 10 μV, the current in the unbalanced resistance is limited to 10 nA. If the common-mode voltage is 10 V, the impedance Z limiting common-mode current flow must be 1,000 megohms. General-purpose instrumentation is designed to have an input impedance of 1,000 megohms on both inputs. This way the unbalanced resistance can be on either input connection.

N.B.

One thousand megohms at 60 Hz is the reactance of 2 pF of capacitance.

The ability to reject a common-mode signal is called the *common-mode rejection ratio* (*CMRR*). If a common-mode signal is 10 V and the resulting output signal is 10 mV, the rejection ratio is 1,000 to one or 60 dB. If the amplifier gain is 100, the error signal at the input is 0.1 mV. The ratio of 0.1 mV to 10 V output is 100,000 to one or 100 dB. This figure is the CMRR *r*eferred *t*o the *i*nput (rti). In the previous example, the CMRR referred to the input is a million to one or 120 dB measured at 60 Hz with a 1,000-ohm input line unbalance.

At 60 Hz a capacitance of 2 pF has a reactance of about 1,000 megohms. To maintain this same reactance at 600 Hz the leakage capacitance would have to be held to 0.2 pF. These numbers show how difficult it is to reject higher frequency common-mode signals using this type of circuitry.

N.B.

A guard shield should be connected to the input signal conductor at the point where the signal connects to the external reference point. In a multichannel system the best practice is to provide each signal with its own guard shield.

The input guard shield is necessary in areas where there are nearby conductors that are not at the input ground potential. In Figure 4.5 the guarding near R_1 may not be needed as all nearby conductors are at the input reference potential. Inside enclosure 2, most of the conductors are referenced to the output common. In this area the input guard shield must surround the input signal line. Any leakage capacitance is a part of the impedance Z_1 or Z_2.

The technique we have just discussed uses a differential amplifier that is located a distance from the transducer. When amplification is provided at the transducer then several different approaches are possible:

1. Provide an rf link using a modulator and demodulator (analog or digital).
2. Provide optical coupling using a digital data link.
3. Use a current loop to couple to a remote differential amplifier.

The first two methods raise the isolating impedance Z to infinity. These methods can be used to transport signals from an aircraft to earth or between two buildings or between computers. Current-transmitting loops are used in many industrial applications. The differential method shown in Figure 4.5 is often used in general-purpose test bays. When all of the electronics is located in the test bay it is convenient to reconfigure the testing by moving transducers. Providing electronics at each transducer or at groups of transducers is more complicated.

Amplifying the signal near each transducer can solve the common-mode interference problem. If the full-scale signal level is increased to 100 mV or greater and the new signal source impedance is below a few ohms, the new common-mode interference problem is not difficult to solve.

If an input signal is preamplified inside the guard shield enclosure and carried by cable to a second amplifier in this enclosure, the CMRR requirement in the second amplifier is reduced by the preamplifier gain. In this case the impedance at the interface between the two amplifiers can be quite low. If the pre-gain is 100, then the CMRR in the second amplifier may only have to be 10^3. This approach is practical when there is no need for signal conditioning at the transducer. Any reduction in the required CMRR does not change the requirement for a high Z in the common-mode path. This impedance is needed to limit the common-mode current flow just as before.

4.8. STRAIN-GAUGE INSTRUMENTATION

Strain-gauge instrumentation is a two-enclosure problem as shown in Figure 4.6. The signal source is a symmetric Wheatstone bridge. The input enclosure contains the gauge resistors as well as the source of transducer excitation. In this diagram all four bridge arms are active, which means they are mounted on the structure being tested. In many applications only one arm of the bridge is active and the remaining arms are located near the source of excitation. All bridge arms must be inside the input guard enclosure. If the excitation source is centertapped, then only two bridge resistors are required.

The capacitance from a conducting surface to a gauge element can be several hundred picofarads. It is safe to assume that there will be a potential difference between the surface under test and the instrumentation output ground. If the gauge is not grounded to the surface under test, this potential difference will add signal to the input circuit. If there is only one active arm, the interference will be the greatest. The best protection against this type of

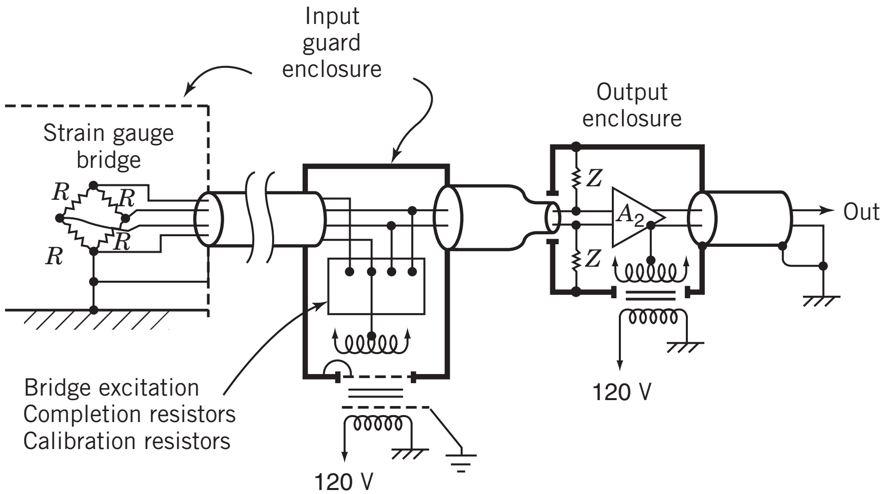


Figure 4.6. The basic strain-gauge circuit

interference is to ground the bridge at the structure and connect the guard shield to this ground.

A strain-gauge bridge can require up to 10 signal conductors. These conductors provide for excitation, remote excitation sensing, calibration, signal, and the guard shield. One conductor group is required for each signal. The guard shield should be carried through any interface without a break. Connecting all the guard shields together at an intermediate point can defeat the purpose of the guard shield.

When the gauge elements of the bridge undergo stress or strain the unbalanced resistance can be as high as 50 ohms. If the interference error limit is $10\,\mu V$, the interference current flow is limited to $0.2\,\mu A$. If the common-mode voltage is 10 V, the impedance Z allowing common-mode current flow must be $5\,M\Omega$. This is the reactance of 3 pF at 10 kHz. This mode of interference coupling is maximum for a full-scale signal. In effect the stress or strain modulates the interference. This problem can be resolved only by limiting common-mode current flow in the gauge arms.

N.B.

A resistance strain gauge generates millivolts of signal.

4.9. THE FLOATING STRAIN GAUGE

An electrical connection from a strain gauge to the structure is not always convenient. If the structure is not used as the reference conductor, then there can be current flow from the structure through leakage capacitance through the gauge elements to the input cable shield. If the coupling is symmetrical, then this current flow creates a common-mode signal that is rejected by the amplifier. If one arm of the gauge is active, then the reactive coupling is definitely unbalanced and any coupled interference will be normal mode. This signal is amplified, not rejected.

The input guard shield and one side of the signal are shown connected to the structure, in Figure 4.6. If there is no connection to the structure, the guard shield should still be connected to the low side of the excitation. In a relatively quiet environment and because the strain gauge is basically a balanced system, the structural connection of the shield and the signal as shown in Figure 4.6 may not be required.

Most instrumentation provides an internal path for input base or gates currents. This allows the inputs to be left open without causing the instrument to overload. These paths are often in the 100-megohm range. Even if these paths are provided it is still best to avoid making measurements where the input leads have a high-impedance to ground. It is possible for the inputs to saturate through a high impedance leakage connection. When this happens the inputs are in overload, but because of feedback the problem may go unnoticed.

Making measurements on a floating structure is also not recommended. If possible, a grounding strap should be used so that the potential difference is controlled.

It is very difficult to mount a gauge and control the coupling capacitance to the structure. If one gauge element is mounted closer to the structure than another, the reactive current flowing in the gauge resistances will not be symmetrical. In a noisy environment this lack of symmetry will introduce a normal-mode signal. This interference can be rejected only by limiting the bandwidth of the system.

In applications where it is difficult to ground the gauge and guard shield to the structure under test, one central point can be used as a ground for a cluster of gauges. This works when the interference field in the space near the transducers is not intense.

N.B.

A high impedance is needed to reject common-mode signals. This impedance may not allow operating from high-impedance signal sources.

There are two common-mode signals that must be rejected by the input amplifier. We have just discussed the ground potential difference mode. The second common-mode signal is one-half of the excitation supply. If the excitation level is 10 V, then this common-mode signal is 5.0 V. The CMRR for this signal should also be 120 dB. As an example, if the excitation level is 10 V, the error signal or offset that is introduced should be less than 5 μV. This requirement is usually not mentioned in a list of specifications. Since the excitation level is constant this is a static specification.

4.10. THE THERMOCOUPLE

A thermocouple is formed by joining two dissimilar metals and bringing them back as conductors to a reference temperature. The voltage between the two conductors at the reference temperature is a measure of the temperature at the junction. This reference temperature can be an ice bath or a temperature-controlled surface. If a reference temperature is not used, a temperature compensation circuit can be used in the instrumentation. This compensation circuit corrects the voltage based on the ambient temperature. In either case the resulting voltage is a measure of the junction temperature. At the reference point the dissimilar metals are connected to copper wires that carry the measured voltage to the instrumentation.

The thermocouple junction is often bonded to a conducting surface to obtain a good measure of temperature. Theoretically this bonding point is where the input guard shield should tie. In practice the guard shield is usually connected where the thermocouple conductors transition to copper. This is satisfactory because the input unbalance resistance is low and bandwidth is not of concern.

If the thermocouple is used to measure the temperature of a fluid, then the junction does not contact a conducting surface. The input signal guard shield should still connect to one side of the signal. One solution is to connect the shield to one of the thermocouple leads at the transition point to copper. The input leads should not be left floating because there is a chance of overload if a leakage path occurs.

It is good practice to filter every thermocouple signal. If necessary a resistor-capacitor filter (RC) can be added at the input to the instrumentation. Because of possible aliasing errors it is necessary to filter the signal before any digital sampling. For a floating thermocouple the mid-point of a balanced RC input filter can be used as a grounding point. This type of filter attenuates both normal and differential mode noise. The impedance of the RC filtering circuit can be 1.0 kΩ. If the capacitors are 0.1 μF, the cutoff frequency will be 1.59 kHz. A higher resistance can be effective but the dc drift of the instrumentation might limit accuracy.

4.11. THE BASIC LOW-GAIN DIFFERENTIAL AMPLIFIER

A simple low-gain differential amplifier is shown in Figure 4.7. This type of amplifier can be applied when the input signal has a low source impedance and signal levels are above 0.1 V full scale. The circuit has inputs labeled V_{IN1}

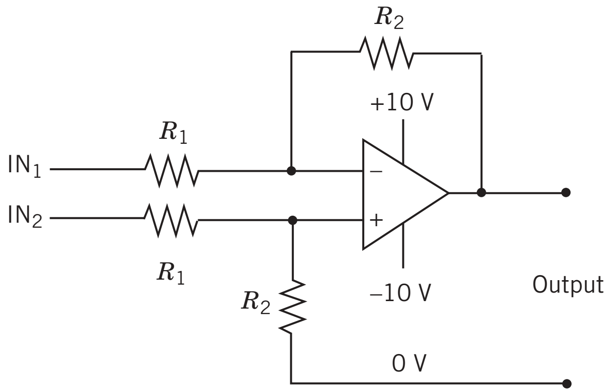


Figure 4.7. The basic low gain differential amplifier

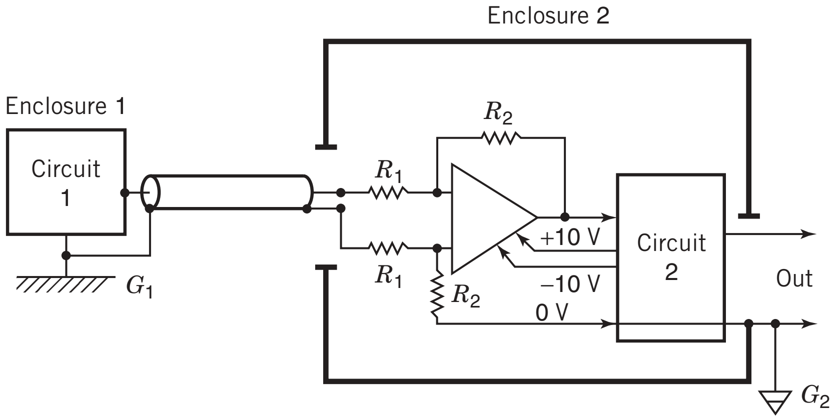


Figure 4.8. The low gain differential amplifier applied to the two-ground problem

and V_{IN2}. The gain from V_{IN1} to the output is $+R_2/R_1$. The gain from V_{IN2} is $-R_2/R_1$. If the same signal is applied to the two inputs, the gain is near zero. If a difference signal V_{DIFF} is applied between the two inputs, the output signal is V_{DIFF} R_2/R_1. This circuit provides gain to a difference signal and rejects the average or common-mode signal. If either input is at zero volts (grounded), the gain to the other input is the ratio of resistors R_2/R_1. The sign of the gain is plus or minus depending on which input is used.

The differential amplifier can be used as the input circuit for the second enclosure as shown in Figure 4.8. The added differential amplifier uses the power supplies available in the second enclosure. If the output common of the second enclosure is taken as the reference conductor, then the input ground potential of the first enclosure becomes the common-mode signal. Ideally the gain for the common-mode component of the signal should be zero. In this application the differential amplifier is called a forward-referencing amplifier. This amplifier re-references the signal found in the first enclosure to the signal common found in the second enclosure.

The common-mode rejection of the circuit in Figure 4.7 depends mainly on the ratio of feedback resistors. If the resistors are equal and their ratios are matched to 1%, the CMRR will be about 100:1. This means a 1-volt common-mode signal at 60 Hz will generate a 10 mV error. This is a 0.1% error compared to a 10 V output. Typical resistor values might be 10 kΩ. This same circuit can be applied to isolating video signals if the integrated circuit amplifier has adequate bandwidth. In this application the feedback resistors should be about 1 kΩ. In video applications the signal is usually limited to 2 volts peak-to-peak.

N.B.

Source impedances must be considered when common-mode rejection depends on having an accurate ratio of feedback resistors.

The gain of the differential amplifier in the above example is unity. If gain is provided by the differential amplifier or by circuits that follow, a higher CMRR should be provided. For example, if the gain in the amplifier is 10, the CMRR should be 1,000:1 to limit the signal error to 0.1% of full scale.

A CMRR depends on the feedback factor (loop gain) of the integrated circuit. See Section 4.18. At frequencies above a few kHz the CMRR will generally fall off proportional to gain. It is good design practice to verify the CMRR at high frequencies if this performance is needed.

The resistances R_1 and R_2 in Figure 4.7 limit the amount of common-mode current that flows in the input common conductor. If this value is 10 kΩ and the common-mode voltage is 1 volt, the current in R_1 is 0.2 mA. If the input common lead has a resistance of 1 ohm, the interference coupling is 0.2 mV. This coupling has no relationship to the CMRR of the forward referencing amplifier. If the feedback resistors are 100 kΩ, the current flow would be reduced by another factor of 10.

Common-mode rejection does not reduce coupling from the power transformer. To limit this coupling, shielding can be added to the power transformer. This shielding is discussed in the next section.

4.12. SHIELDING IN POWER TRANSFORMERS

A basic transformer shield consists of a single wrap of foil between coils. This foil must be insulated at the overlap to avoid a shorted turn. An effective shield can be made from a thin layer of either copper or aluminum. A connection to the shield can be made by taping a bare wire to the shield or by soldering to an installed copper eyelet. The shield lead is usually brought out for an external connection. A shield in a 10-W transformer might limit the mutual capacitance from the primary coil to the secondary coil to about 5 pF. This shield is shown in Figure 4.9.

A single shield between the primary and secondary coils can help to limit reactive current flow in the input signal common. The proper connection for this shield is to *equipment ground*. If the shield were connected to the enclosure or to the signal common, the power fault path would be through the input cable and this is not acceptable. A single shield can not limit the current flowing in the loop ① ② ③ ④ ⑤ ⑥ ① or in the loop ⑦ ③ ④ ⑤ ⑥ ⑦. It would take two additional shields to control this current (see below). The single shield helps when there is a potential difference between the *equipment ground* and the *grounded* or *neutral* power conductor. This potential difference is often the result of harmonic current flow in the facility neutral conductor.

A power transformer used inside of hardware requires three shields to limit most of the unwanted power current flow. The shield next to the primary coil would connect to *equipment ground*. The center shield would connect to the enclosure and the shield around the secondary coil would connect to circuit common. An example of two shields is shown in enclosure 1 of Figure 4.6. To limit the mutual capacitances to around 0.2 pF the shields would have to "box" the coils. (A "box" shield fully encloses each coil.) These shielded transformers must be hand crafted and as a result they are expensive. Boxed shielded transformers are rarely used in today's electronics.

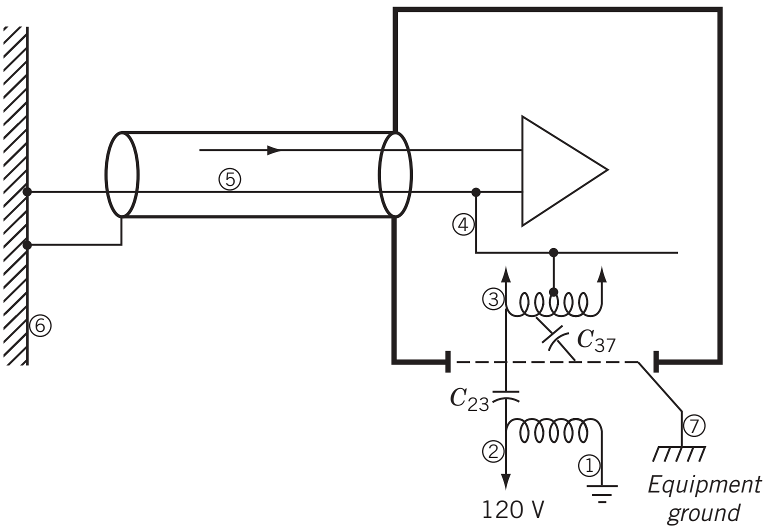


Figure 4.9. The single shield applied to a power transformer

There are applications where the circuit common grounds through a series resistance. An application might be a floating power supply. Three shields as mentioned above can be used to control the flow of reactive currents in the transformer mutual capacitances. In a transformer with three boxed shields the leakage capacitances can be held to approximately 0.2 pF. At 60 Hz this is a reactance of $13 \cdot 10^9\,\Omega$. The circulating current at 120 V 60 Hz is about $9.2 \cdot 10^{-9}$ A. This is a voltage drop of about 9.2 µV in a resistance of 1,000 ohms. Fortunately, there are circuit techniques that can float a power supply without the use of these expensive multishielded transformers.

4.13. CALIBRATION AND INTERFERENCE

There are many approaches to calibrating instrumentation. Bench calibration can test for parameters that on-line testing cannot measure. On-line calibration just prior to a test can verify basic operation and provide data needed to correct for some of the errors. Measurement errors include factors such as linearity, gain, offset, amplifier input noise, common-mode rejection, temperature coefficients, excitation accuracy, signal losses in a cable, rise time, and settling time. Calibration can correct for errors in gain and offset but not for errors caused by common-mode signals. Because of the large number of possible errors, each error contributor should be small compared to a full-scale signal. Specifications often call for error levels of 0.1% of full scale.

N.B.

No amount of calibration can eliminate errors caused by interference.

N.B.

A reminder: A microampere of unwanted current flowing in a one-ohm resistance generates one microvolt of error.

4.14. THE GUARD SHIELD ABOVE 100 kHz

The guard shield should protect the input signal up to the input bases or gates. The presence of the guard shield in the instrumentation can couple high-frequency fields into the enclosure. Even if these signals are out of band they can cause errors that result from overload and signal rectification. It is good practice to connect the shield to the enclosure at frequencies above 100 kHz through a series RC circuit. The circuit ideally should be located outside of the instrument but it is usually placed at or near the connector. Typical values are R = 100 Ω and C = 0.01 μF. This circuit limits the intensity of the field that enters the enclosures. This RC circuit is shown in Figure 4.10.

4.15. SIGNAL FLOW PATHS IN ANALOG CIRCUITS

The basic components that make up an analog circuit are integrated circuits, feedback resistors, resistive voltage dividers, diode clamps, zener diodes for voltage control. RC filter components and power supply components. Typically these components are interconnected on a two-sided printed circuit board. The dc power can be supplied from a nearby rectifier system and/or voltage regulators. On some circuit boards the power is supplied locally from a power transformer. Note there is usually no ground plane available when using a two-layer board. If care is taken in layout, a ground plane is not needed. External signal and power connections can be made using connectors or solder pins. When digital circuits are involved a ground plane may be a necessity. See Section 7.13.

Here are a few rules that will help in analog board layout:

- Maintain a flow of signal and signal common from input to output. The area between the signal path and the signal reference conductor should be kept small.
- Components associated with the input should not be near output circuit components.
- Power supply connections (dc voltages) should enter at the output and thread back toward the input. This avoids common-impedance coupling (parasitic feedback).
- The greatest attention should be paid to the input circuit geometry. Lead length for components connecting to the input path should be kept short. Another way of describing this requirement is to interconnect the components to minimize the amount of bare copper connected to the input signal path.

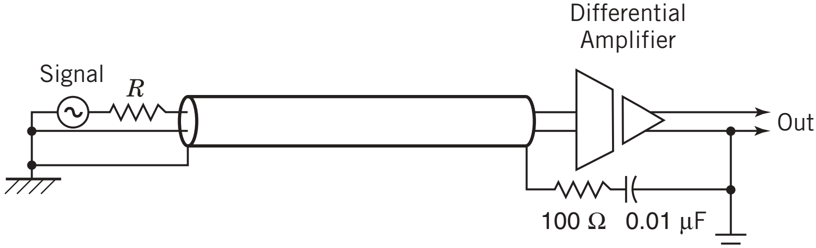


Figure 4.10. The RC bypass on the input guard shield

- Feedback summing points are critical. Keep lead lengths short at these nodes.

4.16. PARALLEL ACTIVE COMPONENTS

When transistors or FETs are paralleled for added performance, there is a good chance for an instability. If an oscillation should take place, it can be high enough in frequency to go undetected. An oscillator of this type can overheat components and/or limit their effective gain. Radiation from this type of oscillator can interfere with nearby circuits. If the circuit is marginally stable, the oscillations might occur after a long warmup or for some values of load.

A problem often occurs when two or more elements of an active component type are connected together. An example might be paralleling power transistors by tying bases, emitters, and collectors together. It is good practice to place a resistor in series with each emitter and add a series resistor to each base. Typical emitter resistors might be 10 ohms and typical base resistors might be 1,000 ohms. These resistors are often called *suppression resistors*. A typical circuit is shown in Figure 4.11.

> **N.B.**
>
> If more than one element of a component is paralleled to a second component, series resistors should be used.

4.17. FEEDBACK STABILITY—INTRODUCTION

Integrated circuit amplifiers are usually supplied with a very high forward gain at dc. Negative feedback is usually required to make these devices practical.

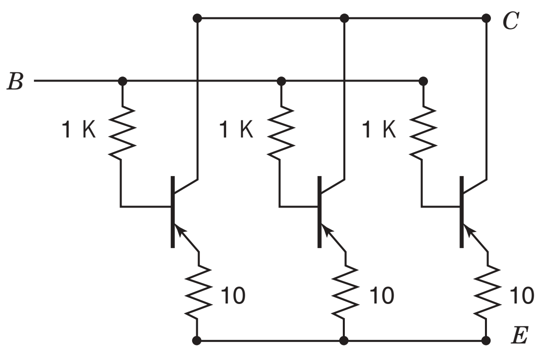


Figure 4.11. Adding suppression resistors to parallel circuit elements

This type of amplifier is supplied with internal compensation, which means there is internal shaping of the open-loop frequency response. Without this compensation, feedback will usually result in oscillation. Even with compensation there are applications where circuit stability can be marginal. To understand the problem a brief review of feedback theory is needed. Because stability is not guaranteed, it is a good idea to test every feedback circuit to make sure it is unconditionally stable.

Here is a glossary of terms that will help in understanding this section.

DEFINITIONS

Closed-loop gain. The gain after feedback is applied.

Feedback factor. The forward gain in excess of the closed-loop gain.

Negative feedback. Subtracting a fraction of an output signal from an input signal and amplifying the difference.

Open-loop gain. The gain before feedback.

Positive feedback. Adding a fraction of the output signal to reinforce the input signal. This usually results in an oscillator.

4.18. FEEDBACK THEORY

A basic feedback circuit is shown in Figure 4.12. The input signal is the sum of a fraction of the output signal βE_{OUT} and the input signal E_{IN}. This summation usually takes place in a differential input stage or by a resistive divider. The gain of this feedback circuit is

$$E_{OUT}/E_{IN} = -A/(1 + A\beta). \tag{4.1}$$

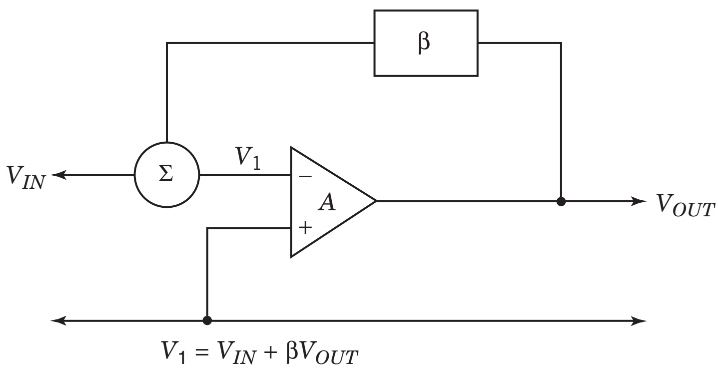


Figure 4.12. The basic feedback circuit

If A is negative and large, then the gain is very close to $-1/\beta$.

In an internally compensated amplifier the forward gain falls off proportional to frequency. The bandwidth of the closed-loop amplifier is approximately the frequency where the open-loop gain is equal to $1/\beta$. As an example, consider an amplifier with an open-loop dc gain of -10^6. If the closed-loop gain is -100 and the bandwidth is 100 kHz, the open-loop gain of the amplifier had to start down around 10 Hz.

The phase shift associated with any circuit is closely associated to the attenuation slope. For a slope that is proportional to frequency the phase shift is 90 degrees For a slope that is proportional to the square of frequency the phase shift is 180 degrees. In the compensated feedback amplifier of Figure 4.11 the phase shift of the open-loop amplifier is approximately 90 degrees over the range 10 Hz to 100 kHz. The phase shift of the closed-loop amplifier is 90 degrees divided by the feedback factor. For example, if the closed loop gain is 100, the feedback factor at 1 kHz is 10^2. This means the phase shift at 1 kHz is approximately 0.9 degrees. At 100 kHz the feedback factor is unity and the phase shift has increased to about 45 degrees. At frequencies above 100 kHz where the attenuation slope is proportional to frequency the phase shift will be 90 degrees.

In Eq. (4.1) the gain A must not have a phase shift greater than 180 degrees before $A\beta$ reaches unity. If this condition is not met, the circuit will oscillate. This condition is known as the Nyquist criterion. If the phase shift approaches 180 degrees as $A\beta$ approaches unity, the result is an amplifier with a very large peak in its amplitude/frequency response. The transient response of this amplifier will have a large overshoot with many cycles of ring-down. This is an indication that the circuit is marginally stable. This ringing condition indicates there is a problem that needs correction.

Feedback systems have limited gain to any internal interference. If a 60-Hz signal is injected at an internal point, the gain to this signal is the closed-loop gain reduced by the gain ahead of the injection point. If the closed-loop gain is 100 and the gain preceding the injection point is 1,000, a 0.1-V interfering signal would by multiplied by 100 and divided by 1,000. The result would be an output signal of 0.01 V. If the output stage has a linearity error of 1 percent this error is reduced by the feedback factor. If the feedback factor is 100 the resulting linearity error would only be 0.01%.

4.19. OUTPUT LOADS AND CIRCUIT STABILITY

If the feedback amplifier in Figure 4.12 is connected to a capacitive load, the result will be additional phase shift in the forward open-loop gain. A capacitive load might be a signal cable having a capacitance of several hundred picofarads. When feedback is applied around this block of gain an instability can result. The problem is most severe if the closed-loop gain is unity. It is good practice to place a parallel LC circuit in series with the output if this output is intended for general use. The L can be 10 μH and the resistor 10 or 20 ohms.

This circuit is shown in Figure 4.13. At frequencies above 100 kHz the output impedance is the resistor in series with the output impedance.[1] This resistance is usually sufficient to avoid any instability. If a 10- or 20-ohm resistance in series with the output poses no problem, then the inductor is not necessary.

It is good practice to test every output circuit for stability. A square-wave signal should be used to drive the amplifier. The preferred test is as follows: A square-wave should be used to drive the output through a 100-ohm series resistor. If there is ringing at the output terminals the circuit should be revised. Note the drive signal must not overload the output circuit. For this reason it is good practice to observe stability using small square-wave signals. Large signals can sometime reduce the loop gain and the ringing (instability) may not be apparent.

N.B.

A resistive load applied to the output of a feedback amplifier can reduce the open loop gain. This can often provide a margin of stability. For this reason stability tests should be made with the output unloaded.

4.20. FEEDBACK AROUND A POWER STAGE

If an integrated circuit amplifier drives a power output stage it is desirable to include the power stage in the feedback loop. The added phase shift (delay) in this output stage can cause instability. A simple way to avoid this problem is to provide a feedback path at high frequencies directly to the output of the integrated circuit. This arrangement is shown in Figure 4.14. At high frequencies the feedback path is through C_1 and R_1.

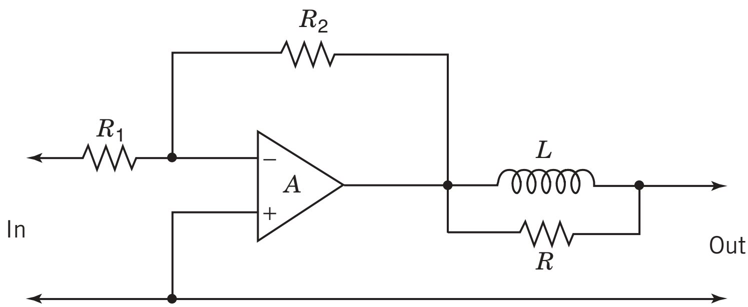


Figure 4.13. An LR stabilizing network

[1] The output impedance of an amplifier with negative feedback is the impedance of the output stage reduced by the feedback factor. Because the feedback factor falls off linearly with frequency the output impedance rises with frequency. A rising output impedance means the output impedance looks inductive. Any capacitive loading is in series with this inductance represents a series resonant circuit. If the phase shift at the feedback sensing point ever reaches 180 degrees and the gain is greater than unity, the circuit will oscillate.

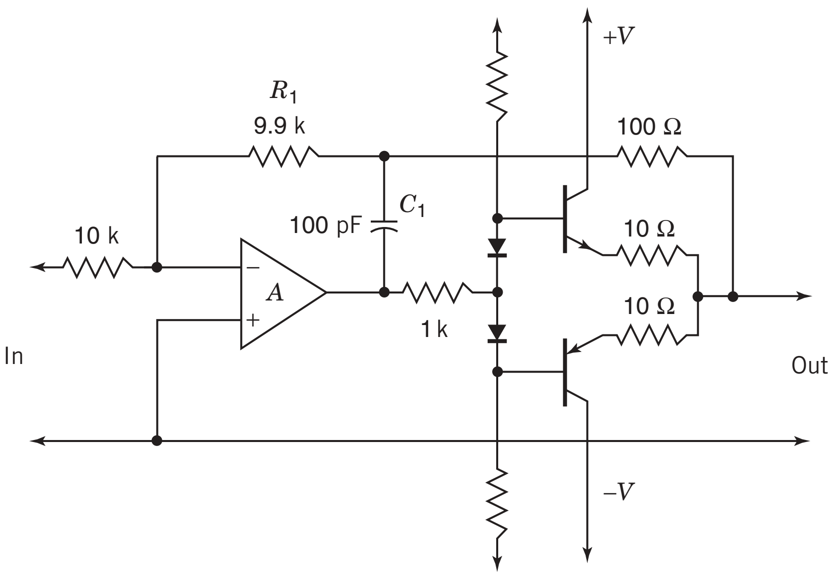


Figure 4.14. Feedback around a power stage

4.21. CONSTANT-CURRENT LOOPS

Instrumentation is available that transmits signal data over long lines using a constant current source. The standard range of current is 4 to 20 mA. The 4-mA current value can represent the zero of signal. A constant current source implies that the current flow is independent of the load and loop resistance or any external voltage induced in the current loop. A precision series resistor can be used to convert the current to a voltage at the receiver. This voltage is then amplified by a differential amplifier located in the second enclosure. If the loop is opened for any reason there is no feedback. Under this condition the circuit will overload.

N.B.

A current loop provides a high source impedance.

4.22. FILTERS AND ALIASING ERRORS

Signals that are sampled must be filtered to avoid aliasing errors. If the sample rate is 10 kHz, there can be no signal content above 5 kHz. If there is signal content, there will be foldover. As an example, a modulated 8-kHz signal will result in a 2-kHz output. Note that 2 kHz is the difference between 8 kHz and the sample rate of 10 kHz.

N.B.

In movie Westerns, stagecoach wheels may appear to rotate backward. This is a form of aliasing error. To eliminate this effect the number of frames per second would have to increase.

An anti-aliasing filter that has a 60-dB-per-decade slope must start attenuating the signal at 500 Hz to provide 60 dB of attenuation at 5 kHz. This analog filtering can take on many forms. Elliptical filters are often used to provide a flat frequency response before there is attenuation. This type of filter has a significant transient overshoot. Bessel-type filters have a very soft knee with no transient overshoot. Because the filter character is known, the resulting data can be digitally corrected for any in-band attenuation. The type of filter selected should match the type of data that is expected. Noisy data can excite transient overshoot in a steep filter and this can result in signal overload.

N.B.

If the data sample rate is reduced, the analog aliasing filter cutoff frequency must be also reduced by the same factor.

4.23. ISOLATION AND DC-TO-DC CONVERTERS

The transformers used in dc-to-dc converters provide a degree of isolation not provided by 60-Hz transformers. These converters are not free from problems, however. Converters are smaller and less expensive than 60-Hz transformers and for this reason they have found wide acceptance in electronics hardware design.

In a typical dc-to-dc converter power supply, utility power is rectified and energy is stored in an electrolytic capacitor. This stored energy is modulated at a frequency above 50 kHz and a transformer couples stored energy to the circuits of interest. The modulator produces a square wave voltage that is rectified in the secondary environment. Energy is supplied to the transformer once per modulation cycle and this means that large energy storage capacitors are not needed on the secondary. The problem with these converters has to do with coupling across the transformer at the fundamental and harmonics of the modulation frequency. Specifically, the leading edges couple voltage spikes across the transformer.

Shielding in a modulation transformer is not practical. The capacitances that would be introduced would make the circuit unusable. If the primary peak voltage is 170 V, the square waves on the primary coil can be 340 V peak-to-peak. If the rise time is 1 μs, the current flowing between the primary and secondary coils depends on the mutual capacitance between the coils. If this capacitance is 10 pF, the current spike has a peak amplitude of 3.4 mA. This magnitude of current flowing in an analog circuit common is generally not acceptable.

There are several ways that this current spike can be reduced. If the transformer is built with a centertapped primary (so that there are positive and negative going voltages), then equal and opposite voltages are coupled through equal capacitances to the secondary coil. Often bifilar wound coils can provide this balance. Under these conditions the currents can cancel. This cancellation is not perfect but the currents can be reduced by a factor of 10 to 100. Sometimes a trimming capacitor might be necessary. If the square wave of voltage is generated with respect to *equipment ground*, any filtering of the current pulse must be with respect to this *ground*.

A way to reduce high-frequency coupling to a secondary circuit is to use two converter transformers in cascade. The second transformer is connected to a coil on the first transformer. This added transformer can be balanced (centertapped) and can operate at a low voltage. The coupling between the two transformers can be referenced to an output circuit common. The low voltage limits the high-frequency currents that might flow in an input common. This arrangement is shown in Figure 4.15.

The power line voltage can be rectified and energy stored in a capacitor. This energy can then be modulated and used for a power supply. During part of the cycle, energy for the modulator can come from the power line as well as from the storage capacitor. To limit this demand, a line filter is needed. This filter places capacitors between the power line and equipment ground. This means that some of the modulation current can enter the *equipment grounding* grid. The loop areas involved in this current flow are not well defined. The result is there can be fields generated by the modulator that create interference for an entire facility. For this reason a dc-to-dc converter should be designed so that this current is limited. In facilities with many pieces of electronic hardware this noise current level can cause troublesome interference fields.

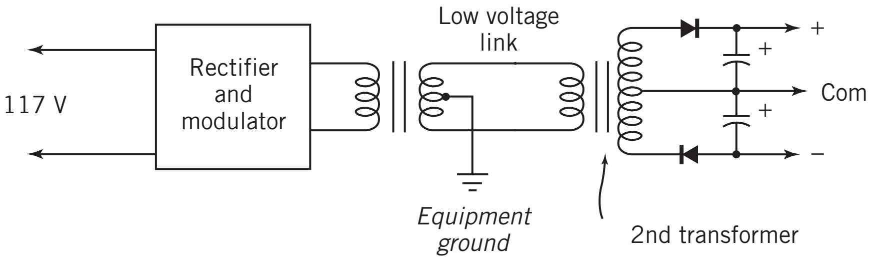


Figure 4.15. Using a second transformer to isolate switching noise

A step-down transformer is often used to reduce the ac voltage before rectification. In this situation the dc-to-dc converter can operate at a lower voltage and the spike levels are reduced. The penalty is the cost and size of the step-down power transformer.

If care is taken, a dc-to-dc converter is a good way to generate isolated power supplies for transducer excitation. Multiple secondaries on a common transformer can be used to generate a group of isolated excitation supplies.

4.24. CHARGE CONVERTER BASICS

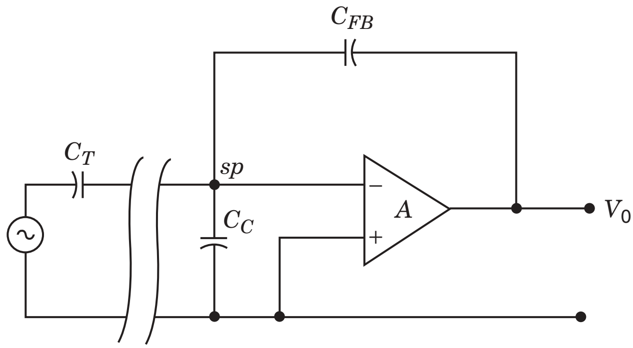
In vibration analysis the sensing transducer is often a quartz crystal. This crystal is electrically equivalent to a capacitor. When the transducer is accelerated, a force is exerted across opposite faces of the crystal. This force generates a voltage on the faces of the crystal. This is called a piezoelectric effect. The acceleration can be measured by sensing the voltage or measuring the charge generated in the capacitance. The relationship between charge and voltage is $V = Q/C$, where C is the transducer capacitance. The relationship between charge and acceleration is a specification provided by the manufacturer.

The voltage on the transducer can be amplified by a high-impedance amplifier. The input cable capacitance attenuates the input signal and this makes calibration a function of cable length. The preferred method of amplifying signals from piezoelectric transducers is to measure charge generation, not the voltage generation. The input cable capacitance does not attenuate the charge, making calibration much simpler. The charge is first converted to a voltage and the voltage is then amplified. This type of instrument is called a *charge amplifier*.

The basic feedback around an operational amplifier usually involves two resistors. The voltage gain is simply the ratio of the two resistors. If the resistors are replaced by capacitors, the gain is the ratio of reactances. This feedback circuit is called a *charge converter*. The charge on the input capacitor is transferred to the feedback capacitor. If the feedback capacitor is smaller than the transducer capacitance by a factor of 100, then the voltage across the feedback capacitor will be 100 times greater than the open circuit transducer voltage. This feedback arrangement is shown in Figure 4.16. The open circuit input signal voltage is Q/C_T. The output voltage is Q/C_{FB}. The voltage gain is C_T/C_{FB}. Note that there is essentially no voltage at the summing node.

N.B.

A charge converter does not amplify charge. It converts a charge signal to a voltage.



C_C = Cable capacitance C_T = Transducer
C_{FB} = Feedback capacitor
The summing point *sp* is a virtual ground.
The circuit transfers the charge generated on C_T and places it on C_{FB}.

Figure 4.16. A basic charge amplifier

The input cable capacitance is connected between the summing point and signal common. Its effect is to increase the noise generated by the input stage of the charge converter. The cable capacitance does not change the conversion from input charge to output voltage.

If the transducer capacitance is 0.01 μF and the feedback capacitor is 100 pF, the gain to an input voltage is 100. To provide a low-frequency response to 1 Hz the resistance across the feedback capacitor would have to be $10^{10}\,\Omega$. This can be provided by a 100-MΩ resistor and a feedback voltage divider of 100:1. This divider arrangement is shown in Figure 4.17. The path for input bias current is 100 MΩ. These high impedances require that the input stage to the amplifier must be an FET.

The noise level referred to the input for a charge converter is usually below 10^{-13} coulombs rms in 100 kHz bandwidth. Consider the coupling between the summing point and a nearby power supply voltage. If the coupling capacitance is 3 pF and the power supply has 30 mV of ripple, the noise coupled to the input is equal to 10^{-13} coulombs. To limit this coupling, the power supplied to a charge converter must be well regulated.

The smallest practical feedback capacitor in the charge converter circuit in Figure 4.15 is about 100 pF. One picofarad of capacitance represents a 1% gain error. Consider a gain of 100 following the converter. Capacitance from the summing point to this amplified signal can cause the circuit to malfunction. If the gain is 100, the mutual capacitance must be held below 0.01 pF. This small value requires that the charge converter be carefully shielded electrostatically. The circuit is usually protected by two small metal boxes that cover the components and circuit traces on both sides of a circuit board. The connection from the transducer to the charge converter must use special

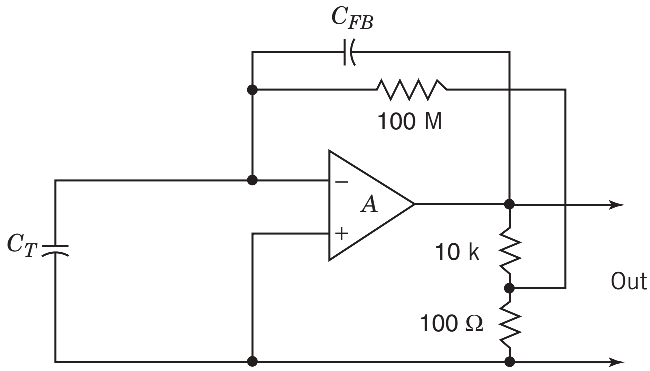


Figure 4.17. The resistor feedback arrangement to control the low-frequency response

low-noise cable. This includes the signal connection inside any instrumentation. See Section 6.3.

> **N.B.**
>
> If the charge amplifier is a plug-in module, a dedicated coaxial input connector must be used. The input signal must be carefully shielded along its entire path.

> **N.B.**
>
> Transducers add weight to the item they are testing. A smaller transducer generates less charge.

The resonant properties of the transducer must be higher than any test frequency. If the transducer requires an insulating mount, this adds mass and another degree of freedom to the measurement. For high-frequency vibration testing it is desirable to avoid this insulation. This means the transducer common is grounded to the structure. This grounding requires that a differential amplifier stage be placed somewhere in the signal path. The stage can be a forward referencing amplifier in the terminating electronics or it can be provided in the measuring instrument. In this case the instrument must have two power supplies, one for the input charge converter and the second supply for the differential stage and any additional circuitry. The instrument would then be called a *differential charge amplifier*. Note the input signal is not balanced or differential in character.

4.25. GUARD RINGS

There are applications where the resistivity of a circuit board can be a problem. A grounded ring can be placed around a key circuit point to collect unwanted board current. A second solution involves a series of grounded vias located in a circle. If this protection is not adequate, the circuit may have to be built on sapphire insulators.

4.26. THERMOCOUPLE EFFECTS

In dc amplifiers the input conductors require special attention. Every joint or connection is a possible thermocouple. When every microvolt counts the following rules apply:

- Maintain circuit symmetry for both inputs.
- Keep the input leads very close together.
- Do not place any heat near the input leads.
- Do not blow ventilation air across these leads.

4.27. GUARD SWITCHING

Switching that involves relay contacts or semiconductor elements is not perfect. The leakage capacitance or resistance between open contacts is often the problem. In some applications, one pF across an open switch is large enough to couple unwanted signals. To limit this coupling a center set of contacts is often provided. These center contacts are connected to points in the circuit that prevent cross coupling by guarding the signal. In complex switching schemes guard contacts may have to be switched to limit unwanted coupling.

N.B.

Floating switch contacts are usually not effective in providing isolation.

N.B.

When signals originate from different ground reference potentials, then all input signal leads should be switched (multiplexed) and this includes the guard shield. The guard shield for each signal must continue to protect the signal through any switching matrix.

4.28. DIGITAL CONTROL

There are many parameters in an instrument amplifier that can be controlled digitally. The digital commands that control functions in the input guard enclosure must cross that boundary without compromising the common-mode rejection specification or adding interference. Control signals can enter optically, differentially, or by use of mechanical relays. In most applications the control link is not in operation during the operation of the instrumentation. This simplifies many aspects of the interference problem.

Transducer excitation functions that are controlled in the guard environment can include voltage or current levels, excitation modes, and current limiting. For strain gauges there are many possible modes of shunt and series calibration. The user has many choices to make in selecting hardware that will meet needs in this area.

Calibration requirements can take on many different forms. It is often desirable to inject a known signal into all inputs just prior to a test run. Signal substitution can introduce problems of crosstalk as this signal must enter every data channel. Switching this signal on and off can be difficult.

There are many ways to handle drift errors. The preferred method is to let a computer measure the offset at the output and correct the data at a later time rather than make electrical changes to the input. Sometimes the output circuitry can be adjusted to accommodate an input offset if the range is not too great.

Functions that relate to the output enclosure do not impact the common-mode rejection process. These functions include some gain ranges, some types of offset, and filter settings.

CHAPTER 5

Radiation

5.1. HANDLING RADIATION AND SUSCEPTIBILITY

Engineering is the study of applying scientific principles to the real world. When electromagnetic radiation and susceptibility are involved, many assumptions are needed in order to obtain useful results. It is possible to apply theory to an antenna design but it is very difficult to apply antenna design to the radiation from a printed circuit board. It is difficult to calculate the structure of a wave that penetrates a round hole in an infinite conducting plane. If the hole is in a metal box, then a solution might be approximated. If the metal box has internal electronics, then a picture of the internal field becomes impossible. It might be suggested that the field in the box be measured using a sensor of some sort. The difficulty here is that the sensor becomes a part of the problem.

> **N.B.**
>
> Almost every practical problem involving radiation or susceptibility requires an approximate solution.

In the real world, radiation and susceptibility problems are complex. Holes are not round and centered, radiation does not arrive perpendicular to surfaces, and the fields are not plane waves. Computers can be used to work on these problems but it is nearly impossible to provide an accurate statement of the conductor geometry so the programs can function. In fact, engineering economics usually requires that this approach not be taken. To understand what happens in practical problems we can apply only basic principles and a worst-case analysis. With this view in mind, we can examine performance based on round holes, infinite planes, and perfect plane waves.

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

N.B.

Even a worst-case analysis should provide for a margin of error.

5.2. RADIATION: WHAT IS IT?

In Chapter 1, the electric field and the magnetic field were discussed. We saw that both fields were present as energy was transferred into a capacitor or inductor, or through a transformer. If one field changed, the other was automatically present. A changing electric field implies a displacement current and a magnetic field. A changing magnetic field implies an electric field. An electric field implies that there are voltages.

When a pair of open wires is placed across a voltage source there are usually no matching source and terminating impedances. As a result there will be many wave reflections that make the round trip over the length of wire. In a practical situation there are heating losses and some radiation. Depending on the length of the line, the reflections that occur will attenuate within a very short time. If an oscilloscope probe is connected to this line, it becomes a part of a line termination. The processes we are describing are real but they occur in a very-high-speed world. In many applications the wave action that takes place in the first few nanoseconds is of little interest. In digital circuitry these first nanoseconds are important. To understand how energy is moved around in a circuit and into space we begin by seeing how energy is coupled to a single pair of conductors.

A simple way to place energy into an open transmission line is to use a voltage having a source impedance equal to the characteristic impedance of the line. When the voltage is connected to the line, a wave of half voltage will propagate down the line. At the open end of the line a reflected wave will result that cancels the current. The forward and return wave continue to add energy to the line. When this reflected wave reaches the source, the voltage across the matching impedance at the source is zero and all wave action stops. The line is now charged to the full voltage. In effect we have radiated energy into the transmission line. The elapsed time was one round trip.

In a transmission line, the flow of energy requires the presence of both fields. At an open circuit or short circuit the energy in motion is not lost. It is simply reflected and returned down the same transmission line. As stated before, any number of waves can use a transmission line at the same time. If the transmission line is terminated in its characteristic impedance, the energy is dissipated in the terminating resistor and there is no reflection. In the example above when the wave reached the termination there is no reflection and the line is at its static condition.

If there is a fork in the transmission line, the energy that reaches the branch point splits and takes two paths and some of the energy is reflected.

If there are multiple forks it is easy to see that there will be a complex of waves traveling in both directions. The reflected waves that thread backward will again reflect at each discontinuity. In theory the number of reflections becomes infinite.

The propagation of electromagnetic energy can be compared to the wave action in a pond. When a stone hits the water a complex wave propagates outward. At the point of entry the wave action eventually attenuates but the wave continues to propagate outward. A stone entering a pond must increase the height of the water in the entire pond. The only way that potential energy can distribute itself over the entire pond is by wave action that reaches the entire surface.

Consider a tapered transmission line where the spacing between the two conductors is increasing. The characteristic impedance increases as the conductors diverge. What happens when energy is sent down this type of line? Wave energy will reflect at each incremental change in the characteristic impedance. This reflected energy is re-reflected at each incremental change in characteristic impedance. The energy that flows outward is a combination of the initial wave and all of this re-reflected energy. The energy that takes the longest path has the greatest delay. It is not difficult to see that the forward wave has a leading edge that broadens with time and distance.

Consider a circuit or antenna that develops an electromagnetic field. The field around this conductor geometry leaves the area as if there were a large number of diverging transmission lines. The characteristic impedance of the diverging lines is continuously increasing. It is not too great a step for us to consider the wave action without the presence of these diverging conductors. The wave that travels outward is continuously reflected. The reflected energy is continuously re-reflected forward. The result again is an outgoing wave of energy that has a broadening wave front.

At a distance from the source the reflecting processes begin to attenuate. This is because the characteristic impedance is nearly constant. The E and H fields working together leave the initial geometry at the speed of light. Within a microsecond the field has progressed about 1,000 feet. If we assume a step voltage source, after a microsecond the electric field near the source has stabilized and the magnetic field is essentially zero. To an observer at 1,000 feet both fields are present as the wave goes by.

The process of returning field energy to a circuit is a little like removing the stone from the pond. The energy that is removed must eventually come from the entire pond. The energy stored at a remote point cannot be redistributed until a wave propagates to that remote point. Since energy must move in both directions from every point of storage, there must be a reflected component to the wave at every point in the pond.

The analogy with a transmission line can help in understanding how waves return stored energy to a circuit. Consider a section of open transmission line that stores an electric field. These two conductors form a capacitor storing a charge Q at a potential difference V. We can remove energy from this line by applying a terminating resistor to the line. If this resistor matches the characteristic impedance of the line, the voltage at the moment of connection drops to half value. A wave propagates forward that is at this half voltage.

The current in this wave is equal to $V/2Z$. A wave flows outward but the energy flows back into the terminating resistor. When the forward wave reaches the open end of the transmission line a reflection must cancel the current. To maintain the flow of energy in the line the return wave must also cancel the remaining voltage. All during the time the wave is traveling out and back, energy is being supplied to the terminating resistor. The energy dissipated in the resistor is $1/2I \cdot 1/2V \cdot 2t = 1/2VIt$ where t is the transit time in one direction. Since $Q = It$ and $C = Q/V$, the dissipated energy equates to $1/2C\ V^2$, the initial energy stored in the transmission line. When the return wave reaches the terminating resistor all wave action stops. If the velocity of the wave is $c/2$, the time it takes to dissipate the energy is $4 \cdot l/c$ where l is the length of the transmission line.

Consider an open transmission line that is tapered. If this line is charged to a voltage V, the line acts like a capacitor. When the line is terminated the result is an outgoing wave with energy returning to the termination. As the wave progresses outward there is a continuous reflection as the characteristic impedance changes. The reflected wave is continuously re-reflected for the same reason. The net result is that most of the stored energy is returned to the termination and some of the energy continues outward.

If we take the next step and remove the conductors we have a static electric field where the field intensity falls off with distance. When the circuit generating the field is terminated, energy begins to flow toward the termination. The continuous nature of the double reflection process sends some of the energy away from the source. This energy will leave the circuit and not return. This continuous process of reflection and re-reflection appears as a delay. The leading edge of the returned energy is spread out over time.

If the pond in the analogy is infinite in scope, then the first wave created by the stone never stops moving. When the stone is removed from the pond the second wave also never stops moving. The two waves never catch up with each other, with the result that some of the energy leaves and never returns. This energy is obviously radiated.

Propagation and continuous reflection occurs whether the wave action is a step function or a sinusoid. One way to view radiation from a sinusoid voltage is to consider the waveform as being the sum of a series of voltage steps. Each step generates a pattern of reflection and re-reflection that has just been described. The beautiful thing about an analysis using sine waves is that the waveforms at all points in the geometry are also sine waves. This is certainly not true when step functions are used. It is important to note that sinusoids are used in most communications systems and that Maxwell's equations are best handled using these same sinusoids.

N.B.

Every transition in voltage or current radiates some energy.

Consider a low-frequency sinusoidal electromagnetic field created by a circuit. The time it takes to return most of the energy to the circuit is short compared to the time of one cycle. In effect the returned energy is in phase with the generating source. As the frequency increases, the time it takes for energy to make a round trip becomes a significant fraction of the sine wave period. The returned energy now has a component that is delayed by 90 electrical degrees. This is the radiated component of the field energy.

5.3. THE DIPOLE ANTENNA

The field pattern around a length of conductor (antenna) can be approximated by summing the field pattern from each segment of the conductor. Assume a sinusoidal current is introduced at the base of the antenna. The magnitude of the current in each segment depends on its position along the antenna. At the tip of the antenna the current is zero and at the base the current is maximum. The distribution of current along the length of the antenna is considered a portion of a sine wave. An antenna driven at its base with respect to a ground is called a *half dipole*. This configuration is shown in Figure 5.1.

The electric and magnetic fields around a dipole are a function of the driving sinusoidal voltage, the directional angle, the distance r from the antenna, and the wavelength. The wavelength λ is the distance the wave travels in one cycle. For example, at 1 MHz a wavelength is 300 meters. The electric field intensity perpendicular to the antenna can be approximated by Eq. (5.1):

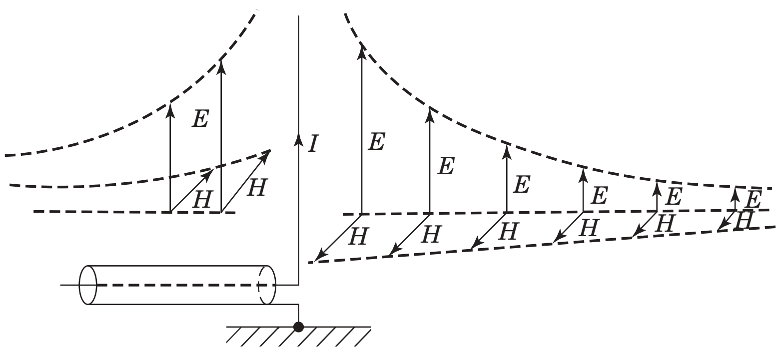


Figure 5.1. A half-dipole antenna

$$E = k_1(\lambda/2\pi r)^3 + k_2(\lambda/2\pi r)^2 + k_3(\lambda/2\pi r). \tag{5.1}$$

The magnetic field can be approximated by

$$H = k_4(\lambda/2\pi r)^2 + k_5(\lambda/2\pi r). \tag{5.2}$$

If the antenna is one-quarter wavelength long, the peak E field near the antenna is $4V/\lambda$ volts per meter where λ is the wavelength.

5.4. WAVE IMPEDANCE

Using Eqs. (5.1) and (5.2), the ratio of E/H at a remote point is simply k_3/k_5. Since the units of the E field are volts-per-meter and the units of H field are amperes-per-meter, the ratio has the units of ohms. The ratio k_3/k_5 is 377 ohms. This does not imply that an ohmmeter can be used to measure the character of space. Since both E and H can be measured over any distance d and the two fields are perpendicular to each other, it is correct to say that E/H equal 277 ohms per square. This is sometimes written as $377\Omega/\square$.

At a distance $r > \lambda/2\pi$, the ratio E/H is 377 ohms. At a shorter distance the ratio E/H will rise proportional to $1/r$. For example, if the frequency is 1 MHz, then $\lambda/2\pi$ is 47.7 meters. At half this distance or 23.8 meters the wave impedance is 754 ohms. The distance $\lambda/2\pi$ is called the *near-field/far-field* interface distance. This impedance value simply means that the E field dominates near this radiating source. This fact will be important when we consider shielding against the penetration of electromagnetic energy.

The value of Poynting's vector integrated over the surface of a sphere at any distance r from the radiating source must yield the same total power. At half the near-field/far-field interface distance the E field increases by a factor of $2\sqrt{2}$ and H increases by a factor of $2/\sqrt{2}$. The ratio of E/H is doubled and the power crossing the surrounding sphere is constant.

The radiation from a current loop takes on the same form as Eqs. (5.1) and (5.2) but with the roles of E and H interchanged. These equations are:

$$H = g_1(\lambda/2\pi r)^3 + g_2(\lambda/2\pi r)^2 + g_3(\lambda/2\pi r) \tag{5.3}$$

and

$$E = g_4(\lambda/2\pi r)^2 + g_5(\lambda/2\pi r). \tag{5.4}$$

For large values of r the ratio of E/H is again a constant equal to 377 ohms. The near-field/far-field interface distance occurs where $r = \lambda/2\pi$. For values of r less than this value the wave impedance gets smaller. At one-half the distance the wave impedance is 188 ohms. At this half distance the value of E increases by $2/\sqrt{2}$ and the value of H increases by $2\sqrt{2}$. The reduction in wave impedance near the radiating source is the reason it is difficult to shield against the penetration of this field energy. The field near a loop of current where the H field dominates is called an *induction field*. At power frequencies the field dominated by current flow is always an induction field. The interface distance at 60 Hz is about 500 miles. The wave impedance at a distance of a few inches would calculate to be a few microhms. This number may not be very meaningful but it does clearly show why shielding this type of field is very difficult.

5.5. FIELD STRENGTH AND ANTENNA GAIN

If a transmitter were to transmit power equally in all directions, the power crossing a spherical surface at a distance r from the radiating source would be

$$P = E \times H\ A = 4\pi r^2 EH. \tag{5.5}$$

Since the ratio of E to H is 377 ohms, E can be written as

$$E = (30P)^{1/2}/r \tag{5.6}$$

where P is in watts, E is in volts per meter, and r is in meters. For example, the E field 1 km from a 1-MW transmitter is 5.47 volts per meter.

In most applications, field energy is directed at some target. In radar, the beam is directed outward by a parabolic surface and the beam angle can be a few degrees. In television broadcasting the energy is directed at the population and not upward to the sky. Obviously this directivity greatly reduces the amount of power required to provide field strength at the target. In the case of radar, if the beam has a solid angle of 1 degree, the power required to produce an E field at a target is reduced by a factor of 360. In the example above, the power requirement would be 2.7 kW. The ratio of 360 degrees to the radiated solid angle is called the antenna gain. The field strength at the target is the same as if a 1-MW transmitter generated a field that propagated uniformly in all directions.

DEFINITIONS

Antenna gain. The ratio of perceived radiated power to actual radiated power. In the example above, the antenna gain is 360.

Effective radiated power. That power level that would provide a required field strength through a solid angle of 360 degrees.

In problems involving susceptibility the peak field strength at the target is all that matters. In a radar signal the pulse duty cycle may be 1%. This makes it practical to use kilowatts of average power to produce field strengths equivalent to a gigawatt transmitter. The susceptibility problem relates to the field strength as if the transmitter used a gigawatt.

The effective radiated power from a list of transmitters is given in Figure 5.2.

5.6. RADIATION FROM LOOPS

In this book we are often interested in unintentional radiators. These radiators are the conducting loops that carry signals and power in circuits. For example, when a logic transition sends a signal to a nearby gate, current flows in the loop formed by the voltage source, the logic trace, and the ground plane. If the energy comes from a local decoupling capacitor, there is a current loop formed by the decoupling capacitor, the ground plane, and a circuit trace. Another circuit loop might involve the drive transistors and an associated dc-to-dc converter transformer. Another loop might involve the connection between a triac and a motor. We are interested in the combined field from all of these radiators.

The electric field pattern from a loop carrying a sinusoidal current depends on many factors, including reflections from nearby conductors and the angles from the center of the loop. In the spirit of worst-case analysis, we consider the

Application	Frequency Range	Effective Radiated Power
VLF navigation	10–300 kHz	300 kW
AM radio	0.5–1.5 MHz	50 kW
Fixed HF	3–30 MHz	10 kW
Hams	3–30 MHz	750 W
Land mobile	3–30 MHz	100 W
VHF TV (low)	50–80 MHz	200 kW
FM radio	80–120 MHz	100 kW
VHF TV (high)	150–250 MHz	250 kW
UHF TV	400–900 MHz	5 MW
Radar—military	0.2–100 GHz	10 GW
Radar—ATC		1 GW
Radar—harbor		100 MW

Figure 5.2. A table of common radiators

maximum field strength E at a distance r from a radiating loop. See Figure 5.3. Field strength is proportional to loop area, to current level, and to the square of frequency. The field strength falls off linearly with distance when the field is measured beyond the near-field/far-field interface distance. At 100 MHz the interface distance is 0.47 meters. In equation form the field strength E beyond the interface distance is

$$E = 6 \times 10^{-3} IA\, f^2/r \tag{5.7}$$

where E is in dB μV/meter, I is the current in mA, A is the loop area in cm^2, f is the frequency in MHz, and r is the distance in meters from the loop. If the distance is 1 meter, the loop area is 1 cm^2, the sinusoidal current is 100 mA, and the frequency is 100 MHz, the radiation is approximately 60 dB μV/meter. The loop we are describing may be rectangular, square, or round.

N.B.

Knowing the wave impedance, the magnetic field strength can be calculated from the electric field strength.

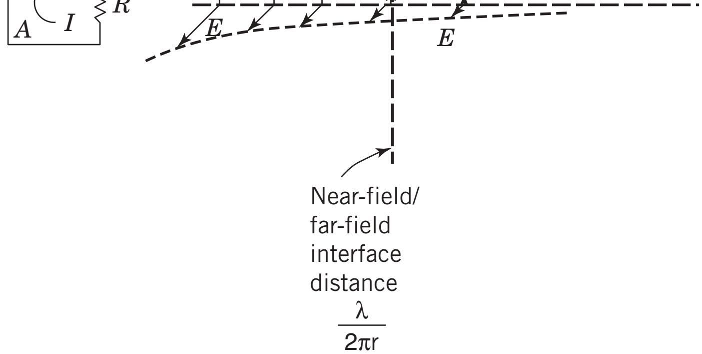


Figure 5.3. The radiated field from a conducting loop

If the dimensions of the loop exceed one-half wavelength, then there may be field cancellation. In the spirit of worst-case analysis, the maximum dimension allowed for any one loop is one-half wavelength. This limitation disallows any field cancellation.

5.7. E-FIELD COUPLING TO A LOOP

When a radiated field is associated with any circuit loop there is a voltage induced in that loop. For analysis purposes, this voltage can appear anywhere in series with one of the loop conductors. If the voltage is in series with a signal loop, then the interference is normal mode. If the interference involves a return conductor sharing several signal conductors, the coupling may be common-mode.

The maximum coupling occurs when the direction of field propagation is parallel to the cable direction. In this arrangement the H-field flux of the interfering signal crosses the loop at right angles. The H field converted to a B field can be used to calculate the induced voltage. The rate of change of the magnetic flux yields the voltage from Eq. (2.5).

The E field can also be used to calculate the induced voltage in a loop. The E-field calculation is simpler because the conversion of units is not involved. At an instant in time the E field has a different intensity at the ends of the cable. This difference is maximum when the cable length is one-half wavelength. The voltages at the cable ends are the E field times the cable spacing d. The maximum induced voltage coupling is twice the peak value of E times d. If the half wavelength is greater than the cable length, the coupling is proportional to the fraction of wavelength. This coupling is shown in Figure 5.4.

Since the field coupling is proportional to cable length and cable spacing it is proportional to loop area. In the spirit of worst-case analysis, if the cable length is greater than one-half wavelength, the half-wavelength figure is used to calculate the voltage.

5.8. A NOTE ON SINE WAVE ANALYSIS

Circuit theory and radiation theory both rely heavily on sinusoidal analysis. For communication transmitters and receivers this analysis fits the application. When nonsinusoidal signals are involved the analysis requires that the spectrum of the signal be considered. Digital logic and transient effects are in this category. For repetitive signals the harmonics can be determined. An analysis then can be made using each harmonic signal. When the responses to the harmonics are added together the result is the response to the original waveform. This assumes a linear system. For impulses or single events the analysis gets more difficult. Often it is sufficient to be able to predict the magnitude of a result at a key frequency rather than actually calculate the spectrum of the

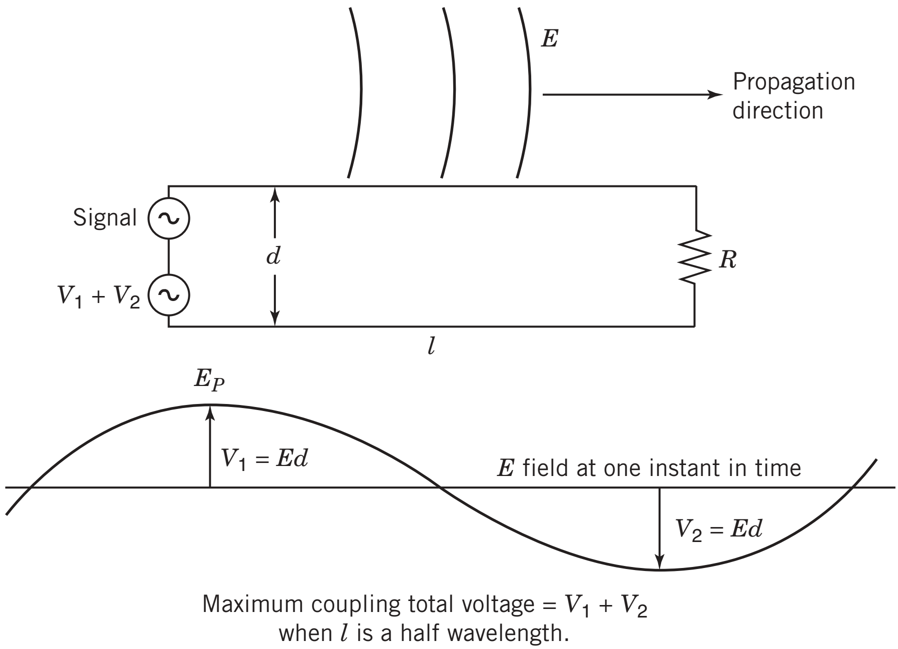


Figure 5.4. Electric field coupling to a pair of conductors

result. The following section outlines this technique whether the signal is a single pulse or has a repetitive waveform.

5.9. APPROXIMATIONS FOR PULSES AND SQUARE WAVES

The radiation generated by a circuit may be the result of a square-wave clock or digital logic. Square-wave voltages are also found in power supplies that use dc/dc converters. The effect these signals have in coupling to a circuit or in radiation from a circuit depends largely on rise time and amplitude as well as loop area. To show this is true we start by examining square waves.

A square wave of voltage or current generates a square wave of E or H field. The sine wave frequencies and their amplitudes that make up a square wave can be derived from a Fourier analysis. These amplitudes are shown in Figure 5.5. The fundamental frequency has an rms amplitude of $2A/\pi$. The sine waves that make up the square wave include all the odd harmonics of the fundamental. The amplitude of the third harmonic is 1/3 the fundamental and the amplitude of the fifth harmonic is 1/5 the fundamental, and so forth. These amplitudes and frequencies are plotted on logarithmic scales in Figure 5.6. Note that the harmonic amplitudes lie along a straight line. This line has a slope of 20 dB per decade. We shall refer to this line as an envelope of peak amplitudes.

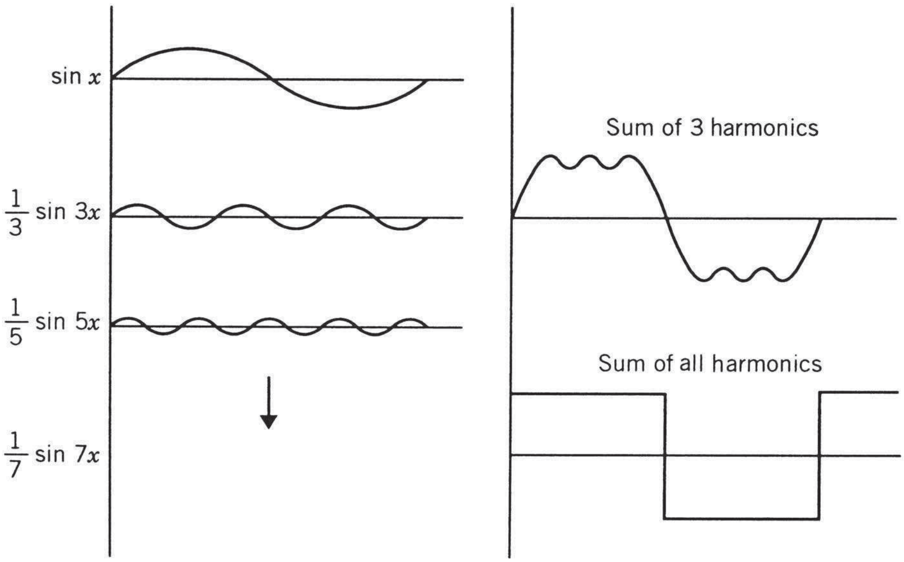


Figure 5.5. The harmonics that make up a square wave

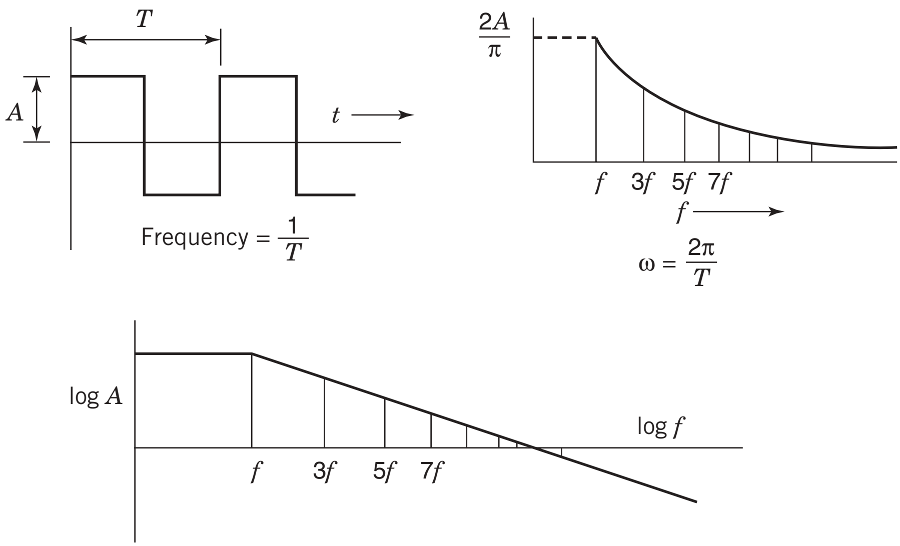


Figure 5.6. The harmonics of a square wave plotted on logarithmic scales

When the square wave has a finite rise and fall time the Fourier analysis is a bit more complex. The harmonic content is shown in Figure 5.7. In this case the harmonic amplitudes vary. When the harmonics are plotted on logarithmic scales a general form emerges. The harmonic amplitudes are less than a linear

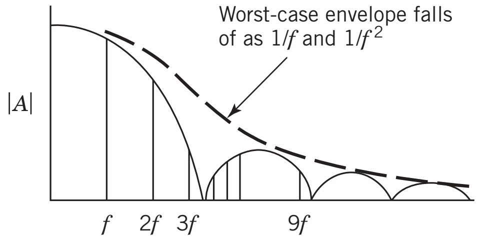


Figure 5.7. The harmonics that make up a square wave with a finite rise time

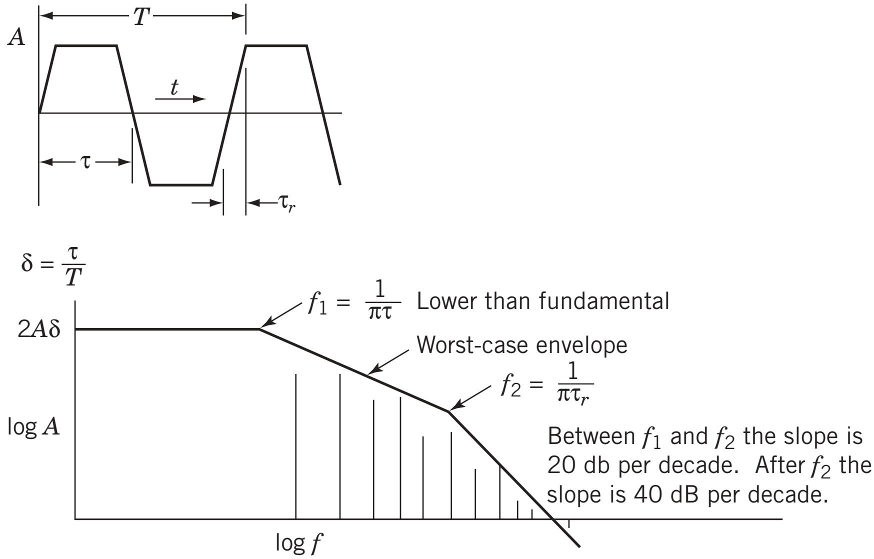


Figure 5.8. The harmonics of a square wave with finite rise time plotted on logarithmic scales

envelope out to a frequency $1/\tau_r$ where τ_r is the rise time of the square wave. Above this frequency the amplitudes are contained by an envelope that falls off as the square of frequency. On a logarithmic plot this second envelope has a slope of 40 dB per decade. This plot is shown in Figure 5.8.

For repetitive pulses with a short duty cycle and a finite rise time, a Fourier analysis shows that the harmonics less than the frequency $1/\pi\tau_r$ have an amplitude less than $2A\delta$ where δ is the ratio of pulse time to duty cycle time or $\delta = \tau/T$. A logarithmic plot of the harmonics with an enclosing envelope is shown in Figure 5.9. A worst-case envelope for harmonic amplitudes has a slope of 20 dB/decade from a frequency $1/\tau$ to $1/\tau_r$. Above this frequency the harmonic amplitudes fall off at 40 dB per decade.

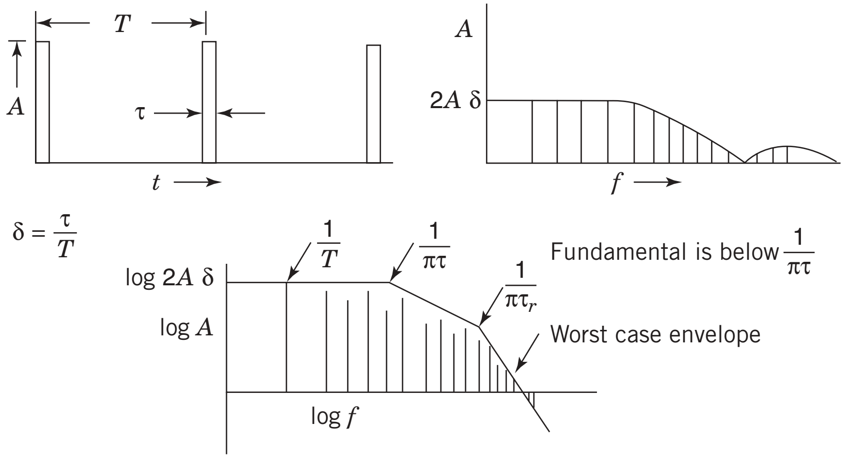
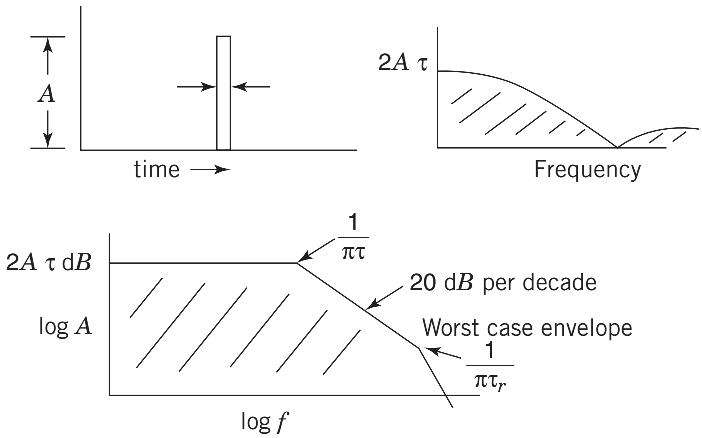


Figure 5.9. The frequency spectrum for repetitive short pulses



If A is in volts and τ is in μs then $2A\,\tau$ is in volts per megahertz

Figure 5.10. The frequency spectrum of a single pulse with a finite rise time

For a single pulse there is frequency content at all frequencies. The amplitudes are a constant out to a frequency of $1/\pi\tau_r$. This is shown in Figure 5.10.

When a square wave of voltage, current, or field couples to a circuit the time response can be calculated by summing the responses to each harmonic. In most circuits the coupling process is proportional to frequency. Because the harmonic amplitudes are falling with frequency and the coupling is increasing proportional to frequency the two effects cancel. The result is a reconstructed square wave using harmonic content out to a frequency of $1/\pi\tau_r$. When there is a finite rise time the harmonics above the frequency $1/\pi\tau_r$ are attenuated in the coupling process.

To analyze how a circuit responds to a complex waveform, a sinusoidal voltage can be selected at a frequency $1/\pi\tau_r$ where τ_r is the rise time. The rms voltage amplitude of the sine wave should be set equal to $2A/\pi$ where A is the peak voltage amplitude of the wave form. This frequency and amplitude can be used in Eq. (5.7) to determine whether there will be circuit damage or a radiation level out of specification. This all assumes the exact waveform of the result is not needed. This approach works for single events or repetitive waveforms.

The two most damaging pulses are lightning and ESD (electrostatic discharge). A switch closure or the arcing at a contact opening also can be considered a pulse. The frequency that characterizes a pulse-like event is again $1/\pi\tau_r$ where τ_r is the rise time.

In the next chapter we will discuss how electromagnetic energy enters an enclosure and how it couples to circuitry. Since the substitute interference is a sine wave and only an approximation, it is wise in design to provide for an additional safety factor of 3 or 10 depending on criticality.

N.B.

Lightning pulses have a rise time of approximately 0.5 μs. A peak current level of 100,000 amperes is worst case. The sine wave frequency to use is 640 kHz.

N.B.

An ESD pulse has a rise time of about 1 ns. A peak current of 5 amperes is typical. The sine wave frequency to use is 300 MHz.

N.B.

Pulse-like events often have different rise times and fall times. The shortest time should be used in an analysis.

5.10. RADIATION FROM A PRINTED CIRCUIT BOARD

The radiating field from a printed circuit board might come from hundreds of sources. Each radiating source has a different orientation and the field measurement might be made at any point around the board. In the spirit of worst-case analysis, the fields from each source can be added together to produce a maximum field strength. The frequency to consider is $1/\pi\tau_r$ where τ_r is the minimum rise time. The radiating loop areas to consider are logic loops, clock loops, and every decoupling capacitor loop. If the number of gates changing state per clock time in an integrated circuit is 25, then the current demand from the decoupling capacitor is multiplied by this number. This information can be used in Eq. (5.7) to obtain a figure for the external E field for each loop. In the spirit of worst-case analysis, the sum of all the E fields provides an indication of the expected radiated field near a printed circuit board.

5.11. THE SNIFFER AND THE ANTENNA

A very practical tool to measure the magnetic fields near apertures or conductors can be built using a section of coaxial cable. The tool, called a *sniffer*, is shown in Figure 5.11. The center conductor is connected to the outer conductor to form a shielded loop. The induced voltage is proportional to the loop area A and the rate of change of magnetic flux.

If the frequency is known, the H field can be calculated from the induced voltage. This is not intended as a calibrated tool but as a simple way to locate sources of radiation.

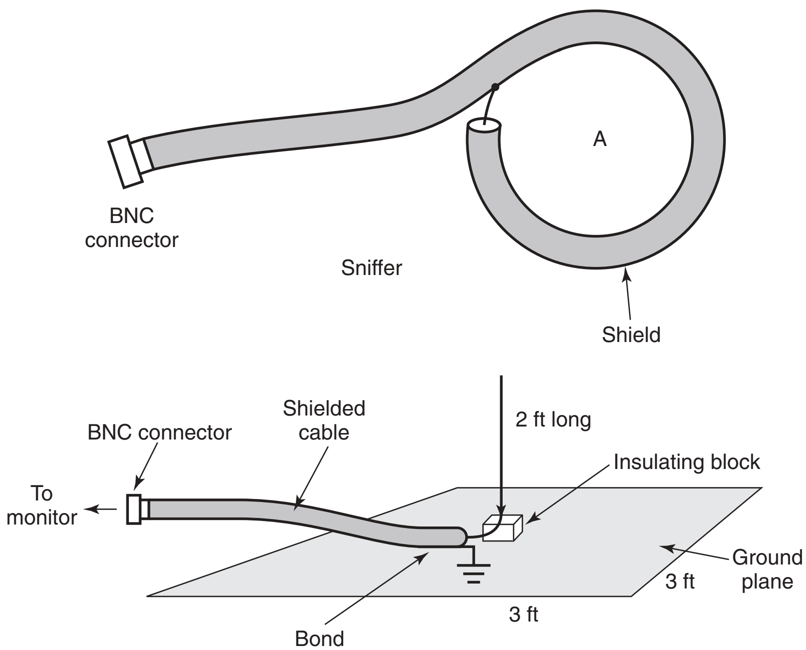


Figure 5.11. A sniffer and a test antenna

Another simple tool that can be built is a small antenna. It too can be used to locate sources of radiation. This structure is also shown in Figure 5.11. The E field can be calculated by noting that the voltage sensed is proportional to the ratio of half wavelength to antenna half wavelength. The maximum voltage that can be sensed is the E field at a quarter wavelength times the length of the antenna. A smaller version can be built for higher frequencies.

N.B.

Before measurements are made the ambient field in the area must be considered.

5.12. SOLAR MAGNETIC STORMS

During periods of solar magnetic activity there can be significant amounts of earth current. These earth currents can be large enough to upset the performance of a power grid. A surge of earth current from a geomagnetic storm apparently blew out a key transformer in Quebec on March 13, 1989, causing six million people to lose power for nine hours.

For a power grid, earth currents can flow under certain fault conditions. If a detector senses an excess of ground current, there may be no way to tell whether this is from a fault or a solar flare. The proper procedure is to protect the hardware and disconnect the load.

Oil pipelines can form large conducting loops. In Alaska some of these loops involve stantions that connect to the earth below the frost line. During periods of solar activity currents as high as 200 amperes have been reported. Smaller current levels are observed on a continuous basis. Since these currents can create corrosion problems insulators are used to break up long sections of pipe to limit this current.

5.13. RADIATION FROM THE EARTH

The magnetic field associated with earth is perhaps the best-known earth radiator. Recently there has been evidence to suggest that very-low-frequency changes in the earth's magnetic field might be associated with pre-earthquake activity.[1] If this is the case, satellites may provide data that can be used to predict where on earth an earthquake might occur. The mechanisms that could generate these low-frequency fields are as yet unknown.

[1] *IEEE Spectrum*, December 2005. Web sites: http://www.quakefinder.com; http://www.science.nasa.gov/headlines/y2003/liaug-earthquakes.htm.

CHAPTER 6

Hardware

6.1. CABLES WITH FOIL SHIELDS

The conductors that interconnect electronic devices are often grouped together into what we call a cable. In analog work aluminum foil is often used as a shield around a cable that goes between a transducer and signal conditioning hardware. The foil has a folded seam that runs the length of the cable. The outside of the aluminum foil is anodized to provide protection against corrosion. Because it is difficult to terminate the foil at the cable ends, a drain wire is provided inside the cable. This drain wire is multistranded and bare so that it can make contact with the foil along the cable length. If the foil should break, the drain wire connects the segments. The drain wire is used as the shield connection.

An aluminum foil over a group of conductors provides an excellent electrostatic shield at low frequencies. In analog work a shield should be connected at one end to the reference conductor preferably where it connects to a ground. See Section 4.4. If the drain wire is connected at one end, this condition is met. If the drain wire is connected to hardware at both ends, then significant interference can result. Electromagnetic fields in the area will induce current flow in the resulting loop. Since the drain conductor is next to the signal conductor bundle, this current will couple interference to any signal being carried by the cable. In a noise-free environment or if the cable run is short, this may not be a problem.

A foil seam does not allow current to flow freely around the cable. Also the foil does not form a very stable geometry. For these reasons foil shields should not be used where the characteristic impedance of the cable needs to be controlled. As we shall see later the termination of shields at a hardware interface can be critical. A cable terminated by a drain wire allows field energy to penetrate the hardware at the connector. More will be said about this topic in Section 6.18.

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

6.2. COAXIAL CABLES

The term *coax* is generally applied to cable where the characteristic impedance is controlled. A center conductor surrounded by a shield with a controlled geometry is called coax. For applications from dc to about 1 MHz the characteristic impedance of a cable may not be important. Above this frequency coaxial cable has wide application. Specifications relating to signal loss at high frequencies are supplied by the manufacturer.

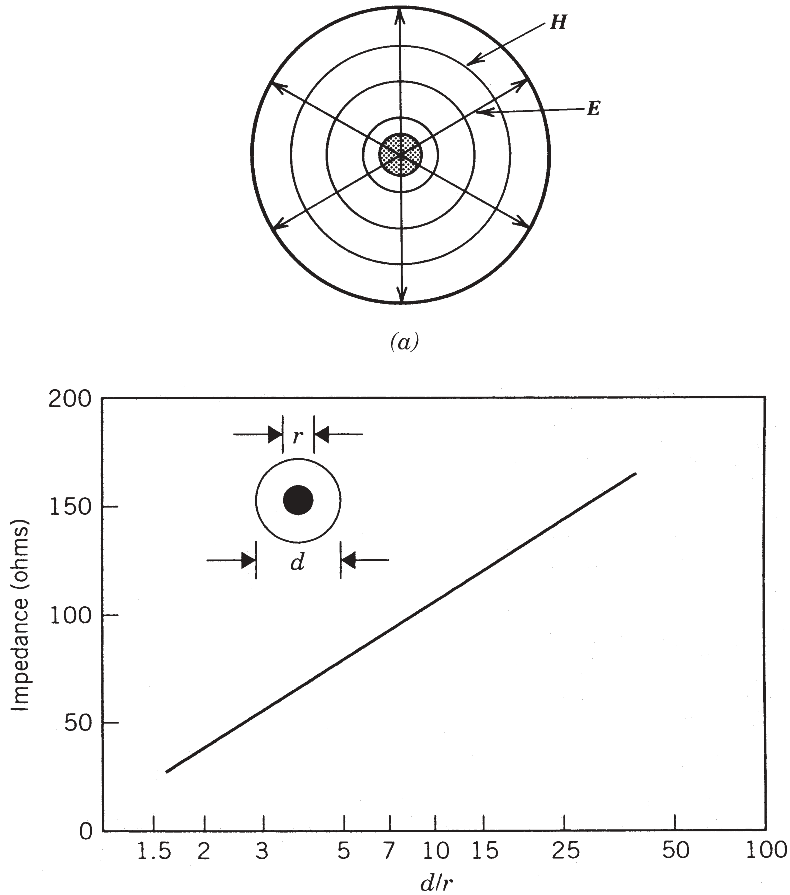
The characteristic impedance of a transmission line is a function of the conductor geometry and of the dielectric constant. To transport power without reflections the source impedance and the terminating impedance must match the line impedance. To transmit high power the terminating impedance should be low. In general, high power requires raising the voltage level. Unfortunately, increasing conductor spacing to accommodate a higher voltage is in the direction to raise the characteristic impedance. In many applications the power level is incidental as the intent is to transfer information at a reasonable voltage level. Obviously the cable from a transmitter to an antenna should be selected to carry power.

Coaxial cable that is used to carry video signals must operate over long distances. Because of distributed reflections it is undesirable to use a dielectric material in the cable. The coaxial cable used by cable companies has a very smooth inner surface that avoids surface reflections. The center conductor is spaced by a nylon cord that is coiled around the center conductor. The center copper conductor is rigid so that it will not easily kink or bend.

The characteristic impedance of a coaxial cable depends on the ratio of conductor diameters. This is shown in Figure 6.1. This table assumes an air dielectric. If a dielectric is present, the capacitance per unit length is proportional to the relative dielectric constant. Since the characteristic impedance equals $(L/C)^{1/2}$ the impedance depends on the inverse square root of the relative dielectric constant.

The characteristic impedance of open parallel conductors is given in Figure 6.2. As the spacing increases the characteristic impedance increases. Assume the characteristic impedance of two parallel conductors spaced by 60 mm is 50 ohms. The characteristic impedance of one conductor spaced 30 mm from a ground plane is simply one-half of this value or 25 ohms. Note the field pattern above the ground plane is the same as one-half the field pattern between two conductors.

The characteristic impedance of circuit traces over a ground plane can be calculated based on the equation $Z = \sqrt{L/C}$. In logic structures the source and terminating impedances are nonlinear, spacings and trace thicknesses can vary as much as 15%, and path lengths are small compared to a wavelength. Typically, traces over a ground plane have a characteristic impedance of about 50 ohms. If the logic levels are 3 volts, the initial current before the first reflection can be assumed to be about 30 mA.



(b)

Figure 6.1. Characteristic impedance of a coaxial geometry

In analog work it is common practice to leave a signal cable unterminated. If the expected bandwidth is greater than a few kHz, it is wise to check the frequency response of the cable. Without a termination there can be peaking to the amplitude response. A series RC termination can shape this frequency-amplitude response. The easiest way to determine the values of R and C is to test the line with a square-wave signal at about 3 kHz. An overshoot of about 6% is acceptable. The capacitor should be set to the smallest value that will allow the resistor to control the overshoot.

L/I	*Z* (Ω)	*H/h*	*Z* (Ω)
1.1	53	0.6	37
1.5	115	1.0	79
2.0	158	2.0	124
2.5	188	2.5	138
3.0	212	3.0	149
4.0	248	4.0	166
5.0	275	5.0	180
10.0	359	10.0	221
30.0	491	30.0	287
100.0	636	100.0	359

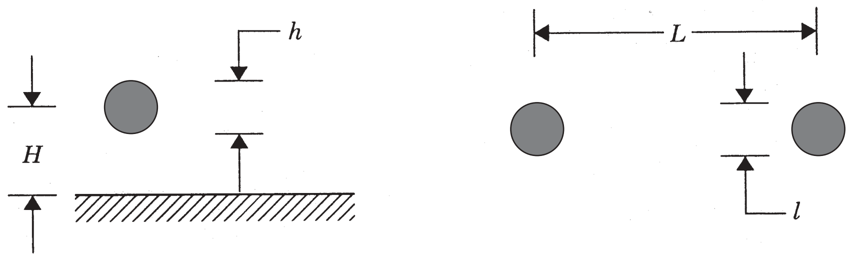


Figure 6.2. The characteristic impedance of parallel conductors

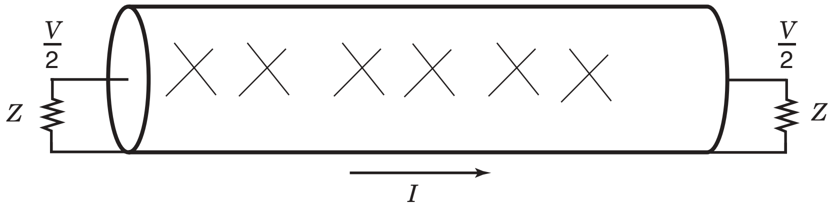
6.3. LOW-NOISE CABLES

Applications that require low-noise cables include piezoelectric transducers. When signal cables are flexed, charges are generated on the surfaces of the dielectrics. This is known as a *triboelectric effect.* To a charge amplifier these charges are noise. To limit this effect, special conductive materials are added around the dielectric. These low-noise cables are available from the manufacturers of piezoelectric transducers.

6.4. TRANSFER IMPEDANCE

Cables often couple to external electromagnetic field energy. As we have said before, current will take this route because it stores less field energy. The coupled energy flows between the cable shield and other nearby parallel conductors. If some of the surface current flow can find a way to the inside surface of the shield, there will be field on the inside of the cable. This mechanism is called *transfer impedance.*

The field that couples into a cable through the shield carries energy that moves in both directions inside the cable. If the cable is terminated at both



$$Z_t = \frac{V}{I}\ \Omega \text{ per meter}$$

Figure 6.3. Transfer impedance test for a coaxial cable

> **N.B.**
>
> If an interfering current flows on the inside surface of a coaxial shield, there is an unwanted field inside the coaxial cable.

ends, then one-half of the coupled energy is dissipated in each termination. If a current I on the shield causes an interfering voltage V at each termination, the ratio of $2V/I$ is said to be the transfer impedance of the cable. This value is normalized for one meter of cable. The test for transfer impedance is shown in Figure 6.3.

At low frequencies the current in a cable shield uses the entire cross-sectional area. The IR voltage drop in the shield produces an internal field. For solid conductors at frequencies above 10kHz the currents tend to stay on the outside surface so there is little internal field. For braided cables for frequencies above a few megahertz the current that gets to the inside surface increases with frequency. The transfer impedance is a function of both the tightness of the weave and the fineness of the braid.

The undulation of the braid adds inductance to the forward path. The current flow in each conductor strand makes many contacts. On the average, conductors heading to the outside surface make contact with conductors that are directed toward the inside. As a result, some of the current transfers to the inner surface. If there are two insulated braids, then the coupled energy is reduced. There is a transfer impedance into the space between the braids. This field is then transferred to the center of the cable by a second transfer process. The lowest transfer impedance is achieved by a using a solid conductor for a shield. Mechanical flexibility can be provided if the shield is corrugated.

Transfer impedance has units of dB ohms per meter. A 0dB figure means 1 ohm per meter while a 20dB figure implies 10 ohms per meter. The transfer impedances for a few standard cables are shown in Figure 6.4. It is interesting to note that the transfer impedance for some braided cable is high enough to make the shield ineffective above a few 100MHz. Obviously the amount of

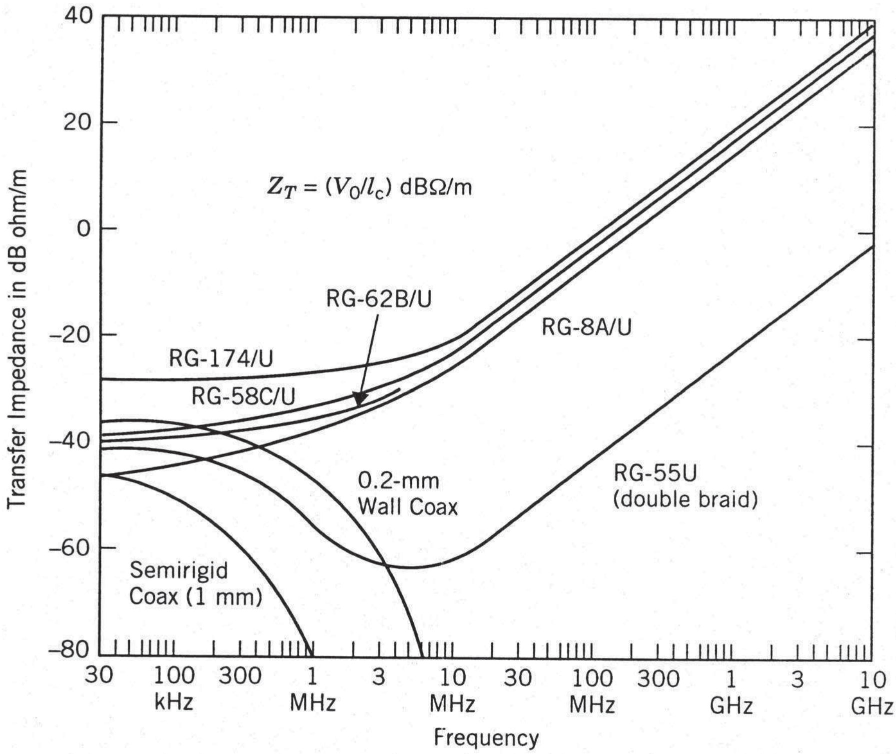


Figure 6.4. The transfer impedance for a few standard cables

coupling increases with cable length. For cables longer than one-half wavelength the coupling tends to cancel. In a worst-case analysis this cancellation cannot be relied on. The maximum cable length used in any calculation should be one-half wavelength.

In applications where the cable runs are short, the cable type may not be important. As we shall see later, the treatment of the cable at the connectors is often far more critical than the type of cable that is selected. See Section 6.18.

6.5. WAVE GUIDES

Electromagnetic fields can propagate in free space at all frequencies. For frequencies below 1 MHz the antennas that are used to propagate energy into space get very large. When two conductors are present it is easy to move field energy from point to point in the frequency range dc to perhaps 100 MHz. Waveguides are the preferred method of transport above a few hundred megahertz.

A waveguide is a hollow cylindrical conductor. Consider field energy where the half wavelength is the size of the waveguide opening. This wave can establish a pattern of reflections on the inside surface of the guide that will allow it to propagate down the guide. At specific higher frequencies the waveguide can support the flow of field energy in various field patterns or modes. As the frequency increases, the number of permitted modes increases until there is no longer a modal restriction. Waveguide propagation requires no center conductor. If a conductor is added the wave guide becomes a poor piece of coax.

An electromagnetic wave that has a half wavelength greater than the waveguide opening is attenuated by the guide. Under this condition the waveguide is said to be operated beyond cutoff. The attenuation of a wave that appears at the entrance of a waveguide is given by

$$A_{WG} = 30\,h/d \tag{6.1}$$

where the attenuation factor A_{WG} is in dB.

If this ratio h/d is 3, the attenuation factor is 90 dB. A waveguide beyond cutoff provides a significant attenuation factor. We will use waveguide attenuation when we discuss ways to shield an enclosure.

An FM radio operates around 100 MHz. The half wavelength is about 1.5 meters. Wave energy from an FM station can easily propagate into tunnels and into underground parking structures. An AM radio station might broadcast at 1 MHz. At this frequency the half wavelength is 150 meters and this energy will not propagate into these structures. If an insulated conductor is added at the roof of a tunnel, it becomes a section of coaxial cable. This one conductor will allow the fields from AM stations to enter the tunnel. This one added conductor will also allow radiated power to exit an enclosed structure.

6.6. ELECTROMAGNETIC FIELDS OVER A GROUND PLANE

A simple picture of field energy propagating along an infinite conducting surface involves a plane wave. Ideally if the conductor has zero resistivity the surface current that flows does not dissipate energy. The only restriction is that the E field must be perpendicular to the surface. The H field must be directed along the conducting surface, which requires a surface current. Any horizontal E-field component would require infinite surface current.

If an electromagnetic wave arrives perpendicular to a perfect conducting surface, the wave simply reflects. The reversal of the E field at the conducting surface is similar to the reflection at the end of a shorted transmission line. The reflected wave cancels the voltage at the surface but leaves the current unchanged. For a plane wave reflecting off of a perfect conducting surface there must be current flow to support the H field at the surface.

For conductive metallic surfaces more than a few millimeters thick, plane waves are reflected and essentially do not penetrate the surface. Plane wave energy will usually enter an enclosure through an aperture involving a hole or a seam, not through the enclosure walls.

> **N.B.**
>
> For conducting surfaces more that a few millimeters thick very little plane wave energy can propagate through the conductor.

6.7. SKIN EFFECT

Electromagnetic field energy that penetrates a plane conductor is attenuated exponentially with depth. This attenuation factor A is

$$\mathrm{A} = e^{-\alpha h} \tag{6.2}$$

where h is the depth of penetration. The term α is equal to

$$(\pi\mu\sigma f)^{1/2} \tag{6.3}$$

where μ is the permeability, σ is the conductivity, and f is the frequency in Hz.

For copper the relative permeability is 1. In Eq. (6.3) μ is the permeability of free space, which is $4\pi \times 10^{-7}$ henries per meter. The conductivity of copper is 0.580×10^{8} amperes per volt-meter. Note that a henry equals a volt-second/ampere. At $f = 1\,\text{MHz}$ the attenuation factor for copper is

$$\alpha = 15.13/\text{mm}. \tag{6.4}$$

When h is equal to $1/\alpha$ the attenuation factor is $1/e$. The attenuation factor A at this depth expressed in decibels is $20 \log 1/e = -8.68\,\text{dB}$. The depth where the field intensity is reduced by the factor $1/e$ is defined as *one skin depth*. For copper at 1 MHz one skin depth is 0.066 mm. In two skin depths the attenuation factor is 17.3 dB.

> **N.B.**
>
> In digital circuits, logic and clock currents flow on the surface of the traces.

The skin depth of copper can be calculated at other frequencies by noting it is inversely proportional to the square-root of frequency. For example, the skin depth at 60 Hz is 0.066 mm increased by the factor $(10^6/60)^{1/2}$ or 0.855 cm. The skin depth of iron can be calculated in a similar fashion by correcting for conductivity and permeability. Note that the permeability of iron is very nonlinear and falls off with frequency. Permeability depends on both alloy type and field intensity.

> **N.B.**
>
> The conductors used in high-voltage power transmission have a steel core center. This type of cable has added tensile strength that allows the supporting towers to be spaced further apart. This increased spacing reduces the cost of transporting the power.

The skin depth analysis we have discussed assumes an infinite conducting plane and a plane wave. The equation for skin depth in a cylindrical geometry (round copper wire) is more complicated. The skin depth $1/\alpha$ is used as a good approximation for all geometries.

6.8. OHMS PER SQUARE

When current flows uniformly in a conductor the voltage drop is simply IR. Consider the conducting square in Figure 6.5. The resistance of a square of material of thickness h is

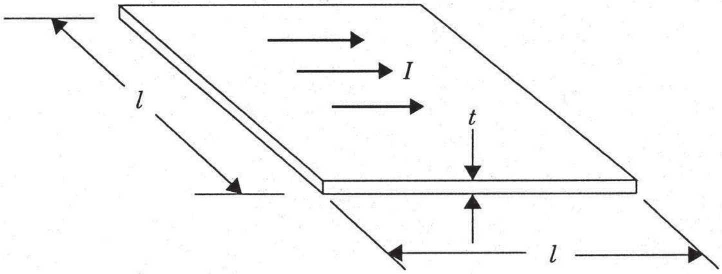
$$\mathrm{R} = \rho l/A \tag{6.5}$$

where ρ is the resistivity, A is the cross-sectional area, and l is the length of a side. Since A equals lh, the resistance equals

$$\mathrm{R} = \rho/h. \tag{6.6}$$

The resistance R is independent of the length or width l.

> **N.B.**
>
> The resistance of a square of conducting material is independent of the size of the square.



$$R = \frac{\rho l}{A} = \frac{\rho l}{lt} = \frac{\rho}{t}$$

R does not depend on dimension *l*.

Figure 6.5. Current flowing across a square

	Copper			Steel		
Frequency	$t =$ 0.1 mm	$t =$ 1 mm	$t =$ 10 mm	$t =$ 0.1 mm	$t =$ 1 mm	$t =$ 10 mm
10 Hz	172 μΩ	17.2 μΩ	17.2 μΩ	1.01 mΩ	101 μΩ	40.1 μΩ
100 Hz	172 μΩ	17.2 μΩ	3.35 μΩ	1.01 mΩ	128 μΩ	126 μΩ
1 kHz	172 μΩ	17.5 μΩ	11.6 μΩ	1.01 mΩ	403 μΩ	400 μΩ
10 kHz	172 μΩ	33.5 μΩ	36.9 μΩ	1.28 mΩ	1.26 mΩ	1.26 mΩ
100 kHz	175 μΩ	116 μΩ	116 μΩ	4.03 mΩ	4.00 mΩ	4.00 mΩ
1 MHz	335 μΩ	369 μΩ	369 μΩ	12.6 mΩ	12.6 mΩ	12.6 mΩ
10 MHz	1.16 mΩ	1.16 mΩ	1.16 mΩ	40.0 mΩ	40.0 mΩ	40.0 mΩ

Figure 6.6. Ohms per square for copper and steel

The square of material can be the size of a room or the size of a postage stamp and the resistance is the same. Ohms per square is often abbreviated as Ω/□.

To use the idea of ohms per square, the current must flow uniformly across the surface. If the surface is a rectangle, the resistance in the direction of current flow is equal to the sum of the resistances of each square. If the flow is across the shorter dimension of the rectangle, the resistance is that of parallel squares.

The resistance of a square of conducting material depends on skin depth. At 1 MHz the skin depth of copper is 0.066 mm or 2.6 mils. This means that in many applications most of the copper is not used for conduction. The ohms per square for copper and steel as a function of thickness and frequency are shown in Figure 6.6.

N.B.

Skin depth limits current penetration for high-frequency current flow. Thickness is provided so that the surface is robust enough to be practical.

Figure 2.4 shows a graph of inductance for long isolated copper conductors. The reactance at 1 MHz of 100 inches of #00 wire would be about 60 ohms. A square of copper 100 inches on a side only 1 mm thick using the data in Figure 6.6 would have a resistance of 360 microhms.

N.B.

The only way to consider low impedances for large distances is to spread the current flow out over a large conducting surface area.

N.B.

A lightning pulse of 100,000 amperes flowing evenly over a sheet of copper 1 mm thick covering an area of 1,000 square feet would cause a voltage drop of about 1.5 volts. This same pulse flowing in 100 inches of #00 wire would result in a voltage drop of over 5,000,000 volts.

6.9. FIELDS AND CONDUCTORS

The current that flows in a conductor flows because there is an E field in the conductor. The electrons are accelerated by the E field but they give up their energy when they collide with the atoms in the conductor. As a result, the electrons reach an average velocity that we interpret as current. Consider a #19 copper conductor that has a resistance of 8 ohms per 1,000 feet. In one meter the resistance is 26.2 milliohms. To support a current flow of 1 ampere the voltage drop is 26.2 millivolts. This is an E field of 26.2 mV/m. In a 10-volt circuit where the spacing is 1 cm the E field between the conductors is 1,000 V/m. The ratio of tangential E field to the perpendicular E field is approximately 38,000 to 1. This supports the idea that there is very little tangential E field in most of the circuits we consider.

Consider a large conducting plane where current enters and leaves at two points. The concentration of current at the points of contact depends on the area of the connection. Because the current concentrates at the points of contact the ohms-per-square concept no longer applies.

At high frequencies a significant electromagnetic field can exist near the points of contact. This field will be greater if the contacts are perpendicular to the plane. The field at the contact point is dominated by the H field and therefore it appears as a series inductance.

6.10. CONDUCTIVE ENCLOSURES—INTRODUCTION

In the next sections we will discuss how electromagnetic energy can enter or leave an enclosure. The energy can couple through and around cable connections, through apertures, and through the conducting enclosure itself. For a near induction field a conducting enclosure may be an ineffective shield allowing direct field penetration.

In Chapter 2 we discussed filtering an input circuit to limit interference coupling. In some analog circuits the effects of radiation are subtle. Out-of-band fields can cause loss of loop gain and add offsets. If local filtering is effective, then enclosure shielding may be unnecessary. This chapter discusses field coupling mechanisms without regard to application.

The radiation that enters an enclosure will couple to internal electronics. Before we discuss enclosure shielding, a story might help to set the stage.

STORY

Consider a metal tub with a lid. Place a small battery-operated FM radio in the tub and note that it receives signal. Now place the lid on the tub. Place your ear near the lid and listen for signal. The radio should stop playing. Now place a section of insulated wire into the tub with a few feet of wire dangling on the outside. The radio should again receive signal.

N.B.

It takes only one unfiltered conductor to violate a shield enclosure.

The rule is simple. Do not attempt to shield an enclosure unless you intend to block all sources of field penetration.

N.B.

A boat with many holes will sink if there is only one *unplugged* hole.

6.11. COUPLING THROUGH ENCLOSURE WALLS BY AN INDUCTION FIELD

Consider a copper or aluminum enclosure. For power-related induction fields (60 Hz and harmonics) there is little reflection loss and the field enters unattenuated. There can be attenuation when the enclosure is made from a material having permeability. The best material is mu-metal. The high permeability is obtained by annealing a magnetic alloy in a magnetic field in an inert atmosphere. The physical size of a mu-metal part is limited by the magnets that generate the field in the oven. After a part is annealed it cannot be punched, drilled, or bent or it loses its permeability. Mu-metal enclosures have been used to enclose CRT monitors to attenuate power-related induction fields. When these fields couple to the electron beam they produce interference patterns on the screen. This shielding is not required for plasma display monitors.

Shielding against induction fields can be accomplished by using layers of steel and copper. For example, signal transformers can be housed in nested cans of iron and copper.

The permeability of some magnetic materials at can be as high as 100,000. This permeability is measured at a maximum flux density for the material. For this same material the permeability may drop to 1,000 at 100 gauss, and at the milligauss levels the permeability may be as low as 2 or 3. This low permeability shows how difficult it is to shield against low-intensity power frequency induction fields.

Magnetic fields can be reshaped by the presence of magnetic materials. In some situations this reshaping can reduce magnetic coupling in a critical region. An example is shown in Figure 6.7. Do not expect a significant attenuation factor with this approach.

If a cable is coated with a layer of permeable material, an external field can be diverted so that the magnetic flux does not cross the area between two signal conductors. The attenuation factor is usually less than 20 dB depending on the field strength and the permeability of the magnetic material. Twisting the signal pair also can be used to reduce this type of normal-mode coupling. The coupling for a half twist cancels the field coupled in the next half twist. Obviously twisting is not available if the cable is coax.

6.12. REFLECTION AND ABSORPTION OF FIELD ENERGY AT A CONDUCTING SURFACE

When an electromagnetic field impinges on a conducting surface, two mechanisms occur. Part of the arriving energy is reflected. The fraction that is not reflected enters the conductor and is attenuated by skin effect. For conductors that are more than a few millimeters thick this field energy is simply converted to heat. For thin conducting layers some of this energy may actually penetrate the conductive barrier.

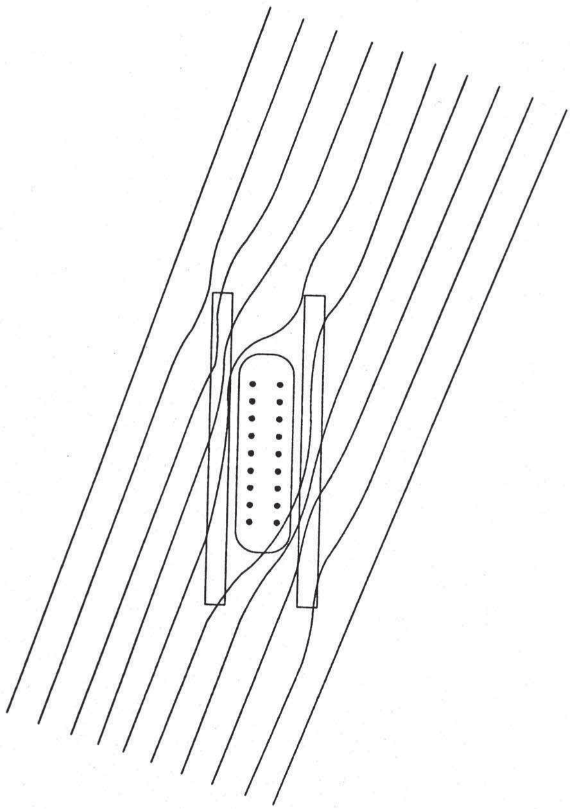


Figure 6.7. Diverting a magnetic field

The reflection loss can be approximated by Eq. (6.7) where Z_W is the wave impedance

$$R_{dB} = 20\log Z_W/4Z_B \tag{6.7}$$

and Z_B is the barrier impedance. Note that for near induction fields, Z_W is very low and there is essentially no reflection. The reflection loss can never become negative as this would imply a reflection field greater than the arriving field.

The wave energy that enters an enclosure through the walls can be characterized by either the arriving E or H field. If the reflection loss is 40 dB and the thickness is 2 skin depths, the arriving field is attenuated by a total of 57.4 dB. If the arriving E field is 40 dB volts per meter, the E field inside the enclosure is 40 dB − 57.4 dB or −17.4 dB volts per meter. In the spirit of a worst-case analysis this field strength is not a function of enclosure volume, field direction, or polarization. In equation form

$$F_{INT}(dB) = F_{EXT}(dB) - R(dB) - 8.68n(dB) \tag{6.8}$$

where F_{INT} is the internal field, F_{EXT} is the external field, R is the reflection loss, and *n* is the number of skin depths of the material. For pulses or square waves the skin depth is calculated based on the frequency $1/\pi\tau_r$ where τ_r is the rise time.

Fields generated inside of an enclosure can radiate out through the enclosure walls. If the radiator has a low wave impedance there may be little reflection loss. In this situation the enclosure may not be effective in attenuating field energy. Equation (6.7) applies for both directions of wave transport.

6.13. INDEPENDENT APERTURES

Field energy can enter an enclosure through an aperture in any wall. The wave pattern on the inside of the enclosure depends on many factors including the presence of internal hardware, the shape of the aperture, the polarization and direction of the field, and the size of the enclosure. A worst-case analysis assumes that the field intensity is not attenuated if the aperture is greater than a half wavelength in dimension. It is common practice to assume that the wave impedance inside the enclosure is the same as the arriving field.

For apertures greater than a half wavelength in dimension it is best to assume no attenuation. For dimensions smaller than a half wavelength the field attenuation is assumed to be the ratio of half wavelength to aperture opening. As an example, if the half wavelength is 20 cm and the aperture opening is 2 cm, the attenuation factor is 20 dB or a factor of 10.

A single aperture might be a seam. The dimension to consider is the seam length. Even if the seam appears to be optically tight, it is still an aperture. The reason for this assumption relates to surface current flow. If the seam interrupts the flow of surface current, it acts as an aperture. Closing a seam requires the use of a conducting gasket. See Section 6.20.

Multiple apertures that allow the free circulation of surface current around each aperture will allow multiple points of field entry. The field strength inside the enclosure is assumed to be the sum of individual penetrations. If an external field is 10 V/m and one aperture attenuates this field by 40 dB and another aperture attenuates the field by 46 dB, the two field strengths are 0.1 V/m and 0.05 V/m. The sum of these fields is 0.15 V/m. Note that the dB measure of field strength is not additive. A field strength of 0.15 V/m is −16.5 dB V/m. The initial field is 20 dB V/m. The attenuation factor expressed in dB is 36.5 dB.

DEFINITION

Independent aperture. An opening in an enclosure that allows the free flow of surface current around the enclosure.

N.B.

When multiple independent apertures allow field entry, the internal field can never be greater than the external field.

6.14. DEPENDENT APERTURES

Arrangements of apertures that do not allow the free flow of surface current are called *dependent apertures*. An example might be a group of ventilation holes. The radiation that penetrates a group of dependent apertures is the same as if there were one aperture.

A wire mesh or screen is considered a set of dependent apertures as current cannot flow freely around each opening. The aperture size for a screen is the dimension of one opening. There are two restrictions: (1) The conductors that make up the grid must be bonded at each crossing. For example, screening made from aluminum conductors can oxidize. When this happens the aperture openings are not controlled. (2) The screen or mesh must be bonded to a conducting surface along its entire perimeter. If there is no bond, the perimeter becomes the aperture.

Consider an enclosure made from conducting panels that are held together by screws. The seam between two screws is considered an aperture. Because surface currents cannot flow freely around each screw, the apertures are considered dependent. For this reason the seams appear as one aperture where the maximum spacing between screws is the dimension of the aperture.

If a 20-dB improvement in field penetration is required, the number of screws would have to be increased by a factor of 10. The practical solution to this problem is to close the apertures using a conducting gasket. To be effective a gasket material should make a continuous connection along the seam. Gasket contact areas should be plated so that there is no chance of oxidation.

If the enclosure is made of sheet metal panels, the edges can be bent to form a flange. This flange can be used as a waveguide opening. If the flange makes many wide surface contacts, the aperture openings are small but they are deep. This type of aperture is considered a waveguide beyond cutoff. Even if the openings are independent the attenuation can be significant. See Eq. (6.1).

If panels are made of molded plastic with conductive surfaces, then flanges can be a part of the design. These flanges can be used with screws or gaskets to form waveguides beyond cutoff. If ventilation holes are plated and extended in depth, then waveguide attenuation is available.

6.15. HONEYCOMBS

Honeycomb structures are often used to ventilate an area and provide for field attenuation. The honeycomb is formed from conducting hexagonal cylinders that are bonded (flow soldered) together. Consider an impinging field at 100 MHz with an E-field intensity of 20 V/m. A half wavelength at this frequency is 1.5 meters. The field at each opening is attenuated by the ratio of the opening to the half wavelength. If the openings are 1.5 cm, the field strength at each opening is 0.2 V/m or −14 dB V/m.

A wave propagates down each honeycomb cell. If the cells are 4.5 cm long, the attenuation in each cell using Eq. (6.8) is 90 dB. The field entering the enclosure from each cell is thus −104 dB V/m. The cells in a honeycomb are independent apertures as current can circulate freely inside each cell. The field that penetrates the enclosure is the sum of the fields from each cell. If there are 20 cells, the internal field strength is increased by 26 dB to −74 dB V/m. Converting from decibels, the internal field is 0.0002 V/m.

N.B.

If the perimeter of the honeycomb is not bonded correctly to the mounting surface, then wave energy can enter through any resulting aperture. In the above example, a 1.5-cm opening would allow the internal field strength to be 0.2 V/m. The honeycomb would be ineffective.

Most honeycomb filters are supplied with mounting hardware and a gasket. The mounting surface should be plated to avoid oxidation. The surface must not be painted or anodized. If the honeycomb is removed for cleaning, the gasket may need replacement if there is any question about a good contact over the entire surface.

6.16. SUMMING FIELD PENETRATIONS

The fields that enter an enclosure can enter through apertures, directly through the skin, or directly on conductors. These conductors might be used for input, output, shielding, control, or power. These same conductors can carry fields out of an enclosure.

N.B.

Nature does not read labels or color codes. Energy can flow in both directions on every conductor entering an enclosure.

A conductor that enters an enclosure with a voltage V creates an E field that is determined by conductor spacing. Conductors that carry power can direct field energy originating from other hardware or the environment into the hardware. These same conductors can carry field energy to other pieces of hardware. For this reason it is important to consider every conductor that enters or leaves an enclosure. If the conductor is carrying current, the H-field intensity and Faraday's law can be used to determine coupling to nearby loops. If the E field is known, then capacitive coupling can be used to calculate coupling.

Up to this point we have considered fields that penetrate through the skin or through apertures from one interfering source. In a worst-case analysis, these fields are simply added together. If fields enter on conductors, they too must be included. If fields have different spectrums, they must be treated as completely separate problems. The fields from different sources can be added together using their rms measure. The total field intensity is the square-root of the sum of the field intensities squared.

6.17. POWER LINE FILTERS

Power line filters are used to restrict the flow of interference into and between pieces of hardware. Filters consist of series inductors in the *grounded* and *ungrounded* power conductors and shunting capacitors. Capacitors can go between power conductors or from power conductors to *equipment ground*. The NEC prohibits filter components from being placed in series with the *equipment ground* as this may limit the flow of fault current. Filters are often purchased as ready-made components that are mounted on the hardware by the user. It is common practice to manufacture a filter assembly that includes an "on" switch, a power breaker, a power cord connector, and even an "on" light.

Power filters have published specifications that show the attenuation of voltages on the power line. This attenuation data should include both common-mode and normal-mode interference. Normal-mode filtering attenuates voltages between the power conductors and common-mode filtering attenuates voltages between power conductors and *equipment ground*.

If the unfiltered power conductors are brought into the enclosure to attach to the filter, these conductors can radiate directly into the hardware and the filter is bypassed. The correct geometry for mounting a filter is shown in Figure 6.8.

Electrical interference can be carried by any conductor pair and one of the conductors can be *equipment ground.* This means that the power conductors and the *equipment ground* conductor must be kept out of the enclosure. This safety conductor should terminate on the inside of the filter enclosure without electrically entering the hardware enclosure. If filter currents must

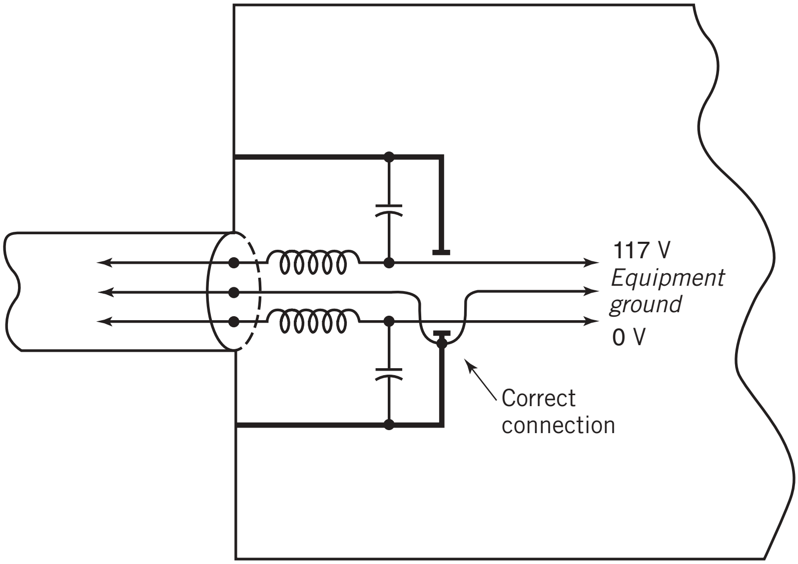


Figure 6.8. The location of power conductors and a line filter

flow from inside the filter enclosure to the outside of the filter enclosure to get to *equipment ground*, there must be fields outside of the filter enclosure and this violates the intent of the filter.

As a side note, it is important to realize that at high frequencies current cannot flow through a piece of metal. Because of skin effect, currents must find a path on conducting surfaces. Consider a filter capacitor that terminates on the inside of a metal can. If the current must flow to an external connection, it must flow through a hole in the can to get to the outside surface. If this current flows on the outside surface of the can, there is an external field. This is what the filter is supposed to eliminate. The presence of this field implies an impedance has been placed in series with the capacitor. In other words, the filter is compromised.

There are many factors to consider in filter design. The field pattern generated inside a filter enclosure must not couple across the filter circuit. This may require some internal partitioning. All inductors have a resonant frequency. Above this frequency the inductor looks like a capacitor. In this frequency range the filter looks like a capacitive divider. All shunt capacitors have a series inductance that is also in the direction to limit filter performance.

There are many filter types. The filter can be L shaped, pi shaped, T shaped, or a combination of the three. Filters can be applied to one or both power conductors. Simple line filters are L shaped with the inductors facing the source of power. The line-to-line capacitor on the load side provides a low-impedance source for step demands in energy. The inductors on the power side of the filter limit current flow for pulses on the line. The L sections should be reversed if the interference comes from the load side of the filter.

Power line filtering starts at frequencies generally above 100 kHz. This limits the size of the filter elements and provides filtering at frequencies where there might be radiation. Line filters are not intended to limit harmonic distortion. Filters that perform this function are very large and expensive.

N.B.

Filters that attenutate signals over a wide frequency range are usually built in sections. This obviously adds to the cost.

If the filter uses a plastic housing or if the filter snaps into an opening using spring clips, the *equipment ground* connection to the hardware must be made separately. If the *equipment grounding* jumper is brought inside the hardware enclosure, it should be kept short to limit loop area and radiation. This method of mounting a filter is not recommended.

For best performance, filter housings should bond to the enclosure. Painted or anodized surfaces should not be used. To avoid oxidation these surfaces should be plated. In some applications it may be necessary to use a conducting gasket between the filter housing and the enclosure.

6.18. BACK SHELL CONNECTORS

Conductors that enter an enclosure through a connector can carry interference. This is no different than the problem of interference entering from the power line. Connectors are available with internal filters. The performance of filters in a connector is limited because of the available space.

If an arriving cable is shielded, most of the interference current is carried on the outside surface of this shield. Field energy will enter the hardware through the connector if the shield has any openings at the connector. If the braid is bunched to form individual conductors and then terminated on mounting screws, interference current will flow on the surfaces of these formed conductors. This allows field to couple to the cable conductors and enter into the enclosure. Ideally shield currents should stay on the outside surface of the shield and flow in a smooth manner to the outside surface of the enclosure. This is the purpose of a back shell connector. It terminates the shield through 360 degrees at the connector. The connector is then mounted using a gasket to close any possible aperture. The termination of a shielded cable using a back shell is shown in Figure 6.9. In a noisy environment this is the only way to keep field energy from entering an enclosure at the connector.

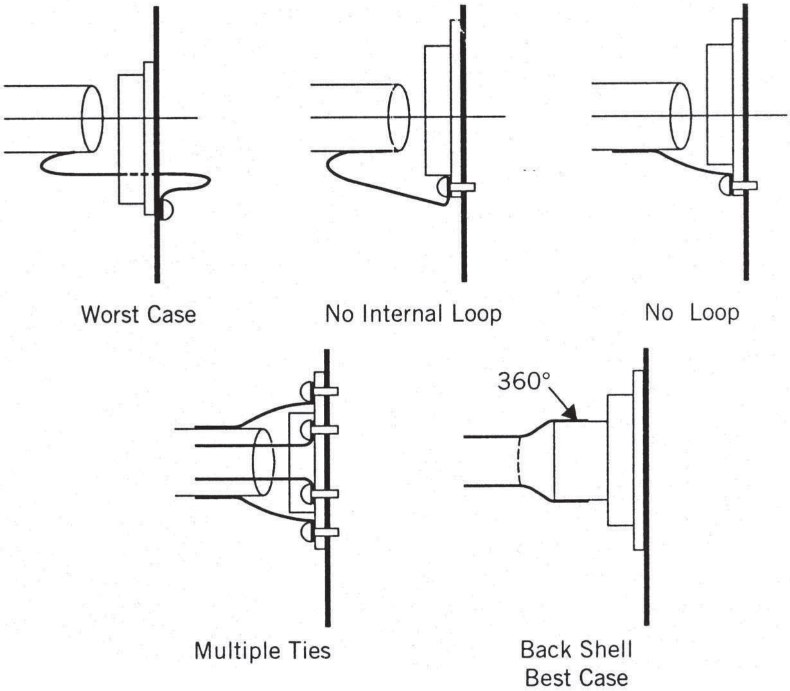


Figure 6.9. Methods of terminating the braided shield on a cable

N.B.

BNC and TNC connectors terminate the coax shield so that current flows uniformly to the terminating enclosure surface.

6.19. H-FIELD COUPLING

The fields that enters an enclosure can couple into a circuit through conductive loops. The largest loops often involve cables between pieces of hardware. If the E field is known, the H field can be determined from the wave impedance. Once the H field is known, the B field can be calculated using the relationship

$$B = \mu H \tag{6.9}$$

where μ is the permeability of free space or $4\pi\ 10^{-7}$. The B field flux ϕ is the B field intensity times the loop area in meters squared. The voltage is given by

$$V = d\phi/dt. \tag{6.10}$$

If the original E field is given as an rms value, the voltage V will also be an rms value.

Example 1. Consider an E field of 10 V/m at 100 MHz inside an enclosure. The H field assuming a plane wave is 10/377 = 0.027 A/m. The B field intensity is found from Eq. (6.8) or B = 3.33×10^{-8} teslas. If the coupling loop is 0.01 m^2, then the flux is 3.33×10^{-10} webers. The induced voltage is $2\pi f$ times this flux or 0.2 V.

Example 2. An ESD pulse strikes an enclosure 10 cm from an aperture 10 cm long. What is the voltage induced in a loop 10 cm^2 near this aperture? Assume a 5-A pulse where the rise time implies a frequency of 300 MHz. Since $2\pi r$H = 5 A, H is equal to 7.96 A/m. B is equal to 10^{-5} teslas. At 300 MHz the half wavelength is 0.5 meter. The wave is attenuated at the aperture by a factor of 5. The field intensity B inside the enclosure is 0.2×10^{-5} teslas. The flux in webers is BA or 0.2×10^{-8} webers. The voltage induced at 300 MHz is 3.7 V. If this voltage adds to a logic level it could damage an integrated circuit.

6.20. GASKETS

Conductive gaskets are used to close apertures. They are often made by embedding filamentary lengths of stainless steel in a carrier. These pieces of metal overlap so that the material is a conductor. When the gasket is under pressure it makes many connections to the conductors involved to close the aperture. Gaskets are available as patterned parts or as strips of various widths. Some gaskets are in the form of a metal braided rope with sharp edges. This type of gasket can be placed in a preformed groove to connect two pieces of metal. When the metal parts are assembled the gasket closes the aperture.

N.B.

Surface preparation is key to the proper installation of a gasket.

N.B.

Gaskets often deform when installed. Gaskets that are removed when equipment is serviced may have to be replaced.

> **N.B.**
>
> The shielding effectivity of gasket material is supplied by the manufacturer. Shielding effectivity is the ratio of field penetration before and after the gasket material is in place. The manufacturer should supply information on how the test was performed.

Metal cloth is available as a shielding material. To close an aperture the cloth must be bonded around its perimeter. Since cloth is somewhat fragile it should not be used where it might be damaged.

6.21. FINGER STOCK

A form of gasket involves finger stock. It is often necessary to shield the aperture around a door. The seam length on a door is long so it is necessary to make many contacts between the door and the door jamb. Finger stock as the name suggests provides many contacts in parallel. If the individual connections are deep, the openings take on the character of waveguides. Typically beryllium copper fingers make contacts that are about a quarter of an inch wide with eighth-inch separation. The stock should be mounted inside a folded cover so that the fingers do not catch on clothing.

6.22. GLASS APERTURES

A thin conductive layer can be plated onto glass to attenuate electromagnetic fields. Unfortunately, these same conductive materials also attenuate light. For this reason any solution is usually a compromise. Contact must be made to the conductor around the perimeter of the glass so that this glass aperture can be closed.

A fine wire mesh embedded in glass can block radiation. Again contact must be made to the mesh around the perimeter to close this aperture. If this mesh is used to cover a computer monitor, Moiré patterns may make the solution ineffective.

> **N.B.**
>
> If the optical path includes a waveguide beyond cutoff, the radiation can be effectively controlled in either direction.

If the opening is 10 inches wide, a 10-inch-long hood bonded to the opening can provide 30 dB of field attenuation out to 1 GHz.

6.23. GUARDING LARGE TRANSISTORS

The collectors or drains for power devices are often connected to the transistor housing. When these devices are used to handle large amounts of power it is necessary to provide for the dissipation of heat. One method is to mount these transistors on a large conducting surface. To avoid an electrical connection a thin insulating gasket is added between the transistor and this surface. To increase the heat conducting area, a thin layer of thermal conducting paste is applied to the gasket.

The capacitance from the transistor housing to the conducting surface can allow current to circulate in the *equipment ground.* To limit this current, the mounting gasket can be metal with insulation on both sides. This metal forms a guard shield that can be connected to circuit common. This shunts most of the parasitic current back to the circuit rather than out into the facility. A typical circuit is shown in Figure 6.10. The leakage capacitance around the gasket

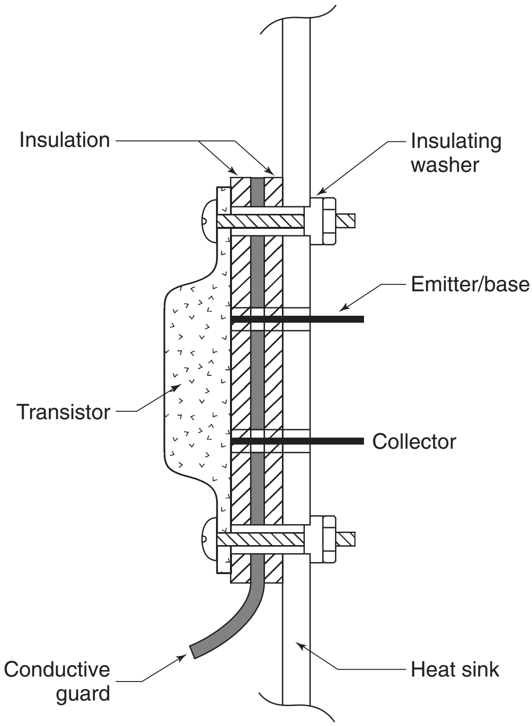


Figure 6.10. A guard gasket applied to a transistor

might be 5 pF while the capacitance without the guard might be 50 pF. If the circuit is arranged so that the drains or collectors are at ground potential, this guarding is not required.

6.24. MOUNTING COMPONENTS ON SURFACES

At high frequencies skin effect keeps currents on the surface of conductors. When components are mounted on conducting surfaces, the way the surfaces mate is important. If the current must concentrate at several points, the result is an inductive connection. At low frequencies this inductance is of no concern. Above a few MHz this inductance can influence the performance of the component by introducing feedback or cross coupling. The important point to consider is that there should not be a concentration of current flow if there is a choice. As an example a component that is riveted in place functions differently than a component that is soldered into position.

The current that flows on the outside of a component housing can only enter the component proper through a hole in the enclosure. This is because high-frequency currents cannot get inside the housing any other way.

Circuit paths that require sharp changes in direction are inductive. A smooth transition is less inductive than a sharp bend. In some cases the roughness of a surface can change the performance of a circuit. Manufacturers will often provide information on acceptable geometries for their components. If this information includes surface and contact quality, these recommendations should be followed.

6.25. ZAPPERS

A zapper is a testing device that generates high-voltage pulses. These pulses can be used to test a piece of hardware to see if it is susceptible to radiation. There are two modes of operation. In the first mode of operation the zapper probe is in contact with the hardware and a pulse of current flows from the probe tip to the hardware. In the second mode the probe tip is placed near the hardware and the zapper generates an arc to a conducting surface. The pulse rate and the intensity of the pulse are both variables. In the contact mode the H field dominates near the point of injection. In the arcing mode the E field is dominant.

There is a close relationship between radiation and susceptibility as both are related to loop areas in the circuitry. In general, if a device is not susceptible to radiation, it probably will not radiate. The test proceeds by using the probe on points such as cables, connectors, seams, displays, and controls. The test should proceed in 1,000-volt steps from 1,000 to 15,000 volts. The midrange around 7,000 volts is important as this is where the intensity of the field energy is apt to peak. If at any time there is a malfunction in the unit under test, it is wise to stop the test and consider some sort of repair.

If the device under test is insulated or isolated in any way, a discharge path should be provided. This should be done so that there is no accumulation of charge from repeated pulses. A resistor of 100 megohms can provide an adequate discharge path. Without this path there can be damage to the circuit that is not related to a susceptibility test. As an example, hardware that is powered from a transformer has no discharge path unless one is provided. Battery-operated hardware also needs a discharge path or the hardware will assume the potential of the zapper.

CHAPTER 7

Digital Electronics

7.1. INTRODUCTION

This chapter covers material that might be considered the analog aspects of digital design. Topics include rise and fall times of clock signals, transmission lines, delay times, radiation, and ground planes. More specifically, this chapter covers circuit boards, cable layout, and ways to optimize digital performance. Circuit boards are often called PCBs (printed circuit boards), PWBs (printed wiring boards), or simply PC boards.

The art of circuit board manufacture has changed significantly in the last decades. There was a time when traces were on tenth-inch centers on a two-sided board. For the bandwidths involved, circuit boards were an economical way to interconnect components and avoid the expense of hand soldering. As the digital clock rates increased, it became necessary to incorporate ground planes to make the circuits function. To add these planes and to accommodate high trace counts, manufacturers of circuit boards have had to develop an amazing art. The very computers that use these multilayer boards make it possible to lay out and manufacture these complex designs. Circuit boards can be built with as many as 70 layers. The manufacturer of a PC board can do magic in interconnecting components, but the design and layout is still the responsibility of the board designer. There is a lot of detail involved in these circuit board designs. There are many ways to create problems in layout especially at high clock rates.

To reduce cost, provide a viable design, and make a board manufacturable, it is wise to work closely with a manufacturer. It is important to find out what he can readily handle and where he has limitations. It is important to find out what makes a board reliable and what features are expensive. Here is a partial list of mechanical parameters that a board manufacturer needs to consider:

Grounding and Shielding: Circuits and Interference, Fifth Edition, by Ralph Morrison

- The grade and type of board material. FR4 is standard glass epoxy.
- How many board layers are needed, including the number of power and ground planes.
- The board size and its relationship to panel size.
- The use of solder masks, conformal coatings, legends, and impedance control coupons.
- Is lead-free manufacturing a requirement?
- Base copper thickness.
- Plating requirements.
- Drilling diameters and pad sizes.
- Trace widths and spacings, layer thickness, overall board thickness.

A multilayer board with symmetry about the center plane reduces cost and is less apt to warp with temperature change. Leaving excess copper on the various layers is preferred as it adds to mechanical stability. Excess copper can be used to improve electrical performance if the designer is aware of the technique.

It takes experience and knowledge to generate a good electrical design that can be easily manufactured. It is much more than a list of components and their interconnection. Here is a partial list of things a designer must consider:

- Component orientation and location
- Number, type, and location of decoupling capacitors
- Connector pin assignments, including ground and power pins
- Connector locations
- Assigning traces to the various layers
- Trace width and spacing; pad size
- Clock signal generation and routing
- Control of characteristic impedance
- Trace interconnection between layers (vias)
- Trace routing, the treatment of stubs, and transmission line terminations
- Board material based on clock rate and impedance control

Digital designs include clocked and unclocked logic. In this chapter the emphasis will be on clocked logic. Logic states change at clock transitions and must settle to a new set of states in one clock cycle. Clock rates have been rising steadily, and at the time of this writing, clock rates in excess of 24 gigahertz are being considered. This exceptional clock rate requires an approach to design that is considerably different from that required by a 10-MHz clock rate. Success at high clock rates rests on an understanding of how signals and energy propagate and reflect on transmission lines.

7.2. CIRCUIT BOARD MATERIAL

Circuits are assembled on an insulated board that allows the interconnection of components by strips of copper, which are called *traces*. The majority of circuit boards built today are made from one or more layers of FR4, an epoxy resin bonded glass fabric. This material has low moisture absorption, high surface and volume resistivity, and low dielectric losses. The maximum clock rate for this material is about 4 GHz. (FR stands for *fire retardant*.)

Dielectric losses are a function of frequency. Because digital signals are rich in harmonics the effect of dielectric loss is to limit rise time. For high-frequency sinusoidal circuits the effect is to attenuate the signal. The attenuation factor for the dielectric is known as a loss tangent or tan(δ). At a frequency f in gigahertz the attenuation in dB per inch is given by

$$\alpha = 2.3\, f\, \tan(\delta)\sqrt{\varepsilon_{eff}} \tag{7.1}$$

where ε_{eff} is the effective dielectric constant at that frequency. The loss tangent for FR4 is between 0.02 and 0.03. The loss tangent for FR408 is between 0.01 and 0.013. This material can be used in digital circuits up to 8 GHz. Nelco 4000-12 can be used up to 12 GHz. Isola 640-300 through 345 can be used up to 40 GHz. Other manufacturers include Matsushita and Rogers.

7.3. THE TWO-SIDED CIRCUIT BOARD

A two-sided board is made by etching copper plate from a clad glass epoxy board to form traces and pads for component mounting. A hole pattern is then drilled that allows components to be mounted to the board. On a quality board the holes are copper plated and solder coated as required.

> **N.B.**
>
> A good design should specify parameters that are easily handled by the manufacturer.

A two-sided circuit board provides for the interconnection of components including signals, ground, and power traces. In this construction the loop areas formed by circuit traces and the common returns can be large. Often the number of traces requires the components to be widely spaced. Boards built with these large circuit loop areas will radiate and be susceptible to interference from external fields. Circuits with large loop areas should be avoided. There are techniques that allow two-sided boards to operate over ground-plane islands. In this case the radiation problem can be controlled. See Section 7.5.

One design approach on two-sided boards is to place the common or ground conductor on the perimeter of the board. The power conductor may be placed on the perimeter of the board on the reverse side. This approach is used to reduce the number of traces but fails to limit signal loop areas. Another approach involves a ground grid constructed from buss wire on the component side of the board. This helps a little, but this technique is usually unsatisfactory.

It may be possible to associate a ground return trace or power trace with every logic line. This solves the problem of providing a transmission line (path) for every signal. This is usually impractical as it doubles the number of traces. In most complex circuits, a ground plane is the best way to provide controlled transmission lines for every signal. This ground plane can be added by laminating two epoxy boards together. This construction allows for two new circuit layers that can be used for a ground plane, a third trace layer, or a power plane layer. With four layers the new problem is the interconnection of components and traces between layers.

7.4. MULTILAYER CIRCUIT BOARDS

A multilayer board allows the use of one or more ground planes, possible power planes, and separate layers for the connecting traces. To increase the component density, trace widths can be made as small as 5 mils on outer layers and 4 mils on inner layers. Smaller trace widths are possible, but they can add to the board cost.

The decision to use multilayers must be carefully considered as it increases the cost of manufacturing and in particular the cost of prototyping. In most cases a prototype cannot be built without going through a full manufacturing cycle. Computer simulation can check out logic function and limit design errors, but it may not be able to predict problems relating to radiation, interference, and logic settling time. The material in this chapter is intended to show the designer how to avoid many of these performance problems and produce a good prototype.

DEFINITION

Via. A conducting hole in a printed circuit board that allows a trace to cross from one layer to another. A "blind via" is visible from an outside layer of the board. A "buried via" is located between inner layers and is not visible.

Multilevel boards can be built with separating layers of insulation called *prepreg*. Consider an eight-layer board with prepreg between layers 2 and 3 and between layers 6 and 7. The vias formed on layers 1,2 or 6,7 are blind. The vias located in the center four layers are called buried vias. The structure of vias on a mutilayer board is shown in Figure 7.1.

> **N.B.**
>
> A through-hole is used to mount components. It can also serve as a via.

There are several ways to lay out a four-layer board. The first method buries the ground and power plane. The layer assignments are:

A. Layer 1: traces
B. Layer 2: ground plane
C. Layer 3: power plane
D. Layer 4: traces

The second method allows extra copper to be used in the trace layers:

A. Ground plane
B. Trace and islands of power
C. Trace and islands of power
D. Power plane

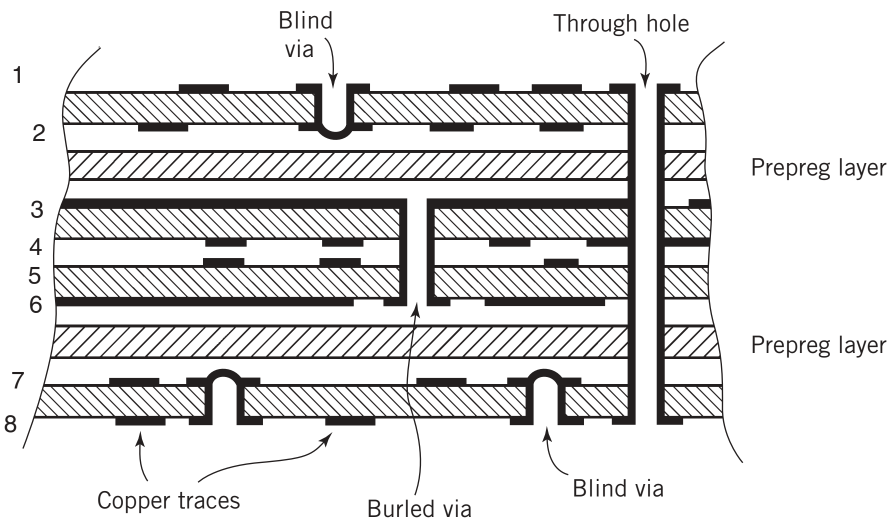


Figure 7.1. The structure of hidden and buried vias

The third method allows extra copper on the outer trace layers:

A. Trace and islands of power
B. Ground plane
C. Power plane
D. Trace and islands of ground

There are many choices to make when the number of layers is six or greater. In general, it is preferable to place ground or power planes between every trace layer. This allows tighter control of characteristic impedances and also limits crosstalk and radiation.

7.5. GROUND PLANES AND DIGITAL CIRCUIT BOARDS

A ground plane is not a shield. Basically a ground plane confines the fields of logic transmission. It is the return path for transmission line current. Without a nearby return path for each logic trace the signal loop areas become large and the fields are not confined. These loops are inductive and this limits clock rate, adds delay to the logic signal, and allows energy to cross couple to other circuits. Every current loop is a source of radiation that adds to the susceptibility problem.

N.B.

A ground plane is a construction technique that allows circuits to operate at high clock rates.

Each logic trace uses the space between the trace and the ground plane to transport field energy. This field energy carries the signal information. If the trace is very close to the ground plane, the electromagnetic field is well confined. Figure 2.15 illustrates this field pattern of a trace over a ground plane. The closer the trace is to the ground plane, the less crosstalk there will be. The field would also be confined if adjacent traces were ground or power traces. The ground plane method automatically provides a near return path no matter where the trace is routed. A ground trace or a power trace both provide an adequate return path but the power trace must be properly decoupled to be effective. A nearby decoupling capacitor between the power trace and the ground plane satisfies this condition. Decoupling is discussed in Section 7.8.

N.B.

On a practical basis the field pattern and the resulting current pattern on the ground plane are not frequency dependent. Most of the current flows under the trace at dc or at 100 MHz. The difference at high frequency is simply skin effect.

N.B.

There is no benefit in etching away excess copper from a circuit or ground layer. Excess copper should be grounded or decoupled, not left floating.

The coupling between traces is proportional to the percentage of flux a trace generates in the space under a second trace. If these fields were confined by coax, there would be no cross coupling. Another way to approximate cross coupling is to consider the frequency associated with rise time, the logic voltage, the capacitance between two traces, and an impedance level for the receiving logic. The cross-coupling signal level for short trace runs is small in most practical circuit board designs.

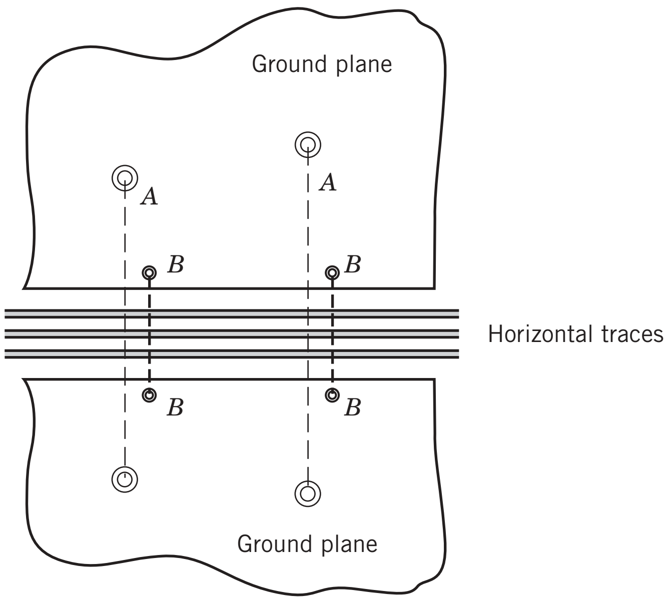
N.B.

As a rule of thumb, on an average board the crosstalk between outer traces is less than 5% if the trace spacing is twice the trace width. On traces between conducting planes the spacing can be one trace width.

N.B.

Cross coupling depends on signal direction. If the culprit signal is first coupled at the termination end of a trace, the coupling level is usually greater.

A ground plane can be broken up into islands as long as all logic traces in the area are associated with a nearby return path. If a logic trace crosses over a split in a ground plane and there is a close parallel ground or power trace, the logic loop area is controlled. An adjacent power trace must be properly decoupled to be effective. This practice is shown in Figure 7.2. If a trace



Vertical traces "*A*" on reverse side.
Vias "*B*" connect ground plane sections
Near "*A*" traces.

Figure 7.2. Crossing a gap in a ground plane

crosses over a split ground plane and there is no controlled return path, the path will be inductive and there will be radiation.

Excess copper that is not associated with nearby logic traces still should be connected to digital common. When traces cross over a conducting area that is not grounded the fields are still confined. The problem is that there are impedance discontinuities (wave reflections) where the trace enters and leaves the control of the floating conducting surface.

When high-speed logic is confined to the inner workings of circuit components, it may be possible to operate the remaining board on logic with longer rise times and at lower clock rates. This approach reduces power consumption and it may be practical to use a two-layer board where a ground plane is not required.

Skin effect limits the penetration of logic current in a ground plane. One-ounce copper plating means one ounce of copper per square foot of surface area. The thickness of 2-oz copper is about 0.3 mm. The ohms per square for this thickness of copper at 10 MHz is 390 μΩ. The skin depth at 10 MHz is only 0.02 mm. This means that very thin layers of copper are effective in carrying logic signals. The thickness that is provided has to do with the practical aspects of manufacture, not with providing conductivity. The characteristic impedance of a trace over a ground plane is a function of the trace surface geometry. The trace thickness plays a minor role in characteristic impedance. The inner copper of a trace does not contribute much to the conductivity.

N.B.

A line of closely spaced vias can limit the effectivity of a ground plane. Current should be able to flow freely around each via.

7.6. CLOCKED LOGIC

In logic systems a transition to a new logic state is related to a clock signal. This transition can start at the clock's leading or falling edge. It takes time for a new set of logic states to settle. This settling time relates to the following factors:

- The rise and fall time of the clock
- The time it takes to pull energy out of a decoupling capacitor
- The switch impedance transition time
- The time it takes to propagate energy to the next logic element
- The time it takes for energy stored in a transmission line to dissipate
- The time it takes for reflections to attenuate

N.B.

All transient effects must settle to an acceptable level before the next clock transition occurs. The worst-case settling time determines the maximum clock rate.

Integrated circuits (ICs) have specifications that relate to external clock transition time (clock leading or falling edges). If the leading edge is too slow, then internal logic may not function in the right sequence, leading to errors. Some integrated circuits regenerate the clock or use an internal clock to avoid this type of error. In some devices the internal clock can be locked to a lower-frequency external oscillator (phase lock loop). If a crystal oscillator is used, the internal clock rate can be made quite accurate.

N.B.

Clipped sine waves do not usually produce a useable clock as the leading and falling edges are too slow. The preferred method is to buffer the signal with a Schmidt trigger to provide a fast leading edge.

Clock signals must connect to many points in the circuit. In some systems it may be preferable to have parallel clock drivers rather than drive all circuits from one source. This approach limits cross coupling and avoids voltage drops in the clock lines. It takes time for clock and logic signals to propagate. These delays become critical as clock rates rise.

A clock line is a transmission line. The initial current that flows depends on the voltage and the characteristic impedance of the transmission line. Adding loads at the end of this line cannot change this initial current value. If two lines are connected to the source, the initial current is double. If there are branches in one transmission path, there will be reflections and the voltage that goes forward will be reduced. If loads are distributed along a single line, then the voltage reaching the last load will be attenuated by numerous reflections. Reflections are not a problem if the voltage can settle along the clock's path before the next clock transition. Stated another way, reflections are not an issue if the lines are short and the rise time is long. Make sure that transmission line stubs are short.

N.B.

The clock must not transition until the logic signals have stabilized.

Consider what happens for a clock signal with a fast rise time. A simple branch point has one-half the characteristic impedance of the line. The reflected wave energy plus the forward transmitted energy must equal the arriving energy. The result is that the forward wave voltage is reduced by $\sqrt{3}$ or 57%. This attenuated clock level is usually too low to operate logic. For lower clock rates the required clock voltage will appear after a series of reflections.

To limit radiation from a transmission line, four conductors located on the corners of a square can be used. If the diagonal conductors are paralleled, the result is a low radiation line. This geometry does not require a ground plane to control radiation. The lines can be traces on two sides of a board. One of the pairs can be at ground potential.

One clock line may supply many ICs if the components are "daisy chained" and the individual ICs do not terminate the transmission line. If the clock line is long, a termination at the end of the chain might be effective. In general, stub lengths should be kept below one-eighth the distance a wave travels in one rise time.

7.7. THE TRANSMISSION OF A SINGLE LOGIC SIGNAL

A trace over a ground plane is a simple transmission line. The transmission line we considered in Section 2.17 consisted of a switch, a battery, and a pair of conductors. In the practical world, the switch is replaced by a transistor and the source of energy is usually a decoupling capacitor. The theoretical switch closed in zero time, but the practical semiconductor switch takes time to transition between states. A logic switch can be technically considered an impedance that changes as a function of time.

The decoupling capacitors that are used are placed across the power supply at or near the switch. Their role is to provide local energy that can be sent down a logic transmission line when a switch is closed. Without a local capacitor the energy request would have to travel back to the power supply. When many circuits make demands on a power supply line and it is not adequately decoupled the voltage can sag to near zero. This sag may last a microsecond, but in today's systems many logical decisions need to occur in this time interval.

After the switch closes, the local drain of energy creates a slight sag in voltage. This slight sag is a request to the power supply to resupply charge to the capacitor. This must take place over the next clock cycle.

The logic levels from a device can be no higher than the power supply voltage at the device. If other integrated circuits share this power supply voltage, a drop in voltage is propagated on these other logic lines. The result is that the logic signals are cross coupled. This cross coupling and the time it takes for the power supply voltage to recover are the reasons it is necessary to provide sufficient local energy storage (decoupling).

N.B.

A source of field energy is stored in the capacitance between the ground and power plane. This is discussed in the next section.

The velocity r of a wave on a transmission line is less than the speed of light. The equation for velocity in meters per second is given by

$$v = 1/\sqrt{LC} \tag{7.1}$$

where L is the distributed inductance in henries per meter and C is the distributed capacitance in farads per meter. Since capacitance is inversely proportional to the dielectric constant, the velocity of a wave is inversely proportional to the square root of the dielectric constant. The velocity along a circuit trace is typically about 6 inches per nanosecond. The distance a signal travels during a logic rise time provides a good reference point. If a transmission line is shorter than one-fourth of this distance, then transmission line terminations are usually not required. For example, consider a 100-MHz clock with a rise time of two nanoseconds. If the traces are less than 3 inches long, there is little need to terminate the lines. A clock bus that is longer than a wave travels in a quarter rise time should be terminated.

The outer surface traces on a PC board are separated from the ground or power plane by a layer of glass epoxy. For these outer traces most of the electric field associated with a signal is located in this dielectric. From Eq. (7.1) the velocity of the wave depends on the distributed capacitance, which is directly related to the dielectric constant. For glass epoxy the relative dielectric constant is about 4.4 at 1 MHz and about 3.6 at 2 GHz. This means that the higher harmonics of a square wave will travel faster than the lower harmonics. For traces on a PC board these timing differences are very small. For longer lines this difference in velocity adds to the distortion of the waveform (adds risetime).

There are many additional factors involved in logic propagation:

- A logic switch is a nonlinear source of current with a finite rise time.
- The voltage supplied to a logic element also connects to other logic elements. This means that some of the initial energy really comes from these transmission lines and logic lines in parallel.
- The gate or base receiving the logic is also nonlinear and usually does not terminate the line in its characteristic impedance. In some logic families the termination might look like a capacitor.
- Decoupling capacitors take time to release their energy. Energy in zero time is infinite power. See Section 7.12.
- The board material is a nonlinear dielectric that affects the transmission time and contributes to losses.
- Transmission line energy must reflect until the energy is lost in heat or is radiated. If the line is terminated, there is no reflection and the wave heats the terminating resistance.
- The characteristic impedance of a transmission line is affected by the presence of all nearby conductors whether they are grounded, connected to logic, or floating.
- When a logic state returns to zero, the energy stored in the line must dissipate or radiate.

As a result of these factors it is not easy to predict with accuracy the resulting waveforms and the time it takes for the logic level to settle. It will vary for each trace and for each component. Fortunately, in digital logic the waveforms are not critical. The only thing that matters is that the voltage be within accepted limits at the time the clock signal changes state. It is good practice in digital design to establish an error budget that includes variations in power supply voltage, common-mode levels, noise, reflection levels, temperature, humidity, line losses, and termination effects. This error budget must be considered in addition to the maximum and minimum logic threshold levels.

In practice an integrated circuit may have many internal logic transitions at each clock edge. The logic levels that are generated may go to many circuits (fan out). Performance at this level of complexity is best handled by building circuits and running tests. The manufacturer will usually recommend the highest clock speeds and indicate the size and location of required decoupling capacitors. We will discuss this aspect of design later.

N.B.

When a logic state returns to zero, the energy stored on the line must be dissipated. The energy can be lost as heat in the copper and in the dielectric or it can radiate. The decoupling capacitor cannot be recharged using this energy.

Consider a board with 1,000 traces with an average capacitance of 2 pF. If 50% of the traces carry a 5-V logic level, the energy stored is $\frac{1}{2}CV^2$ or 25×10^{-9} joules. If the clock rate is 100 MHz, then the energy must be stored and dissipated every clock period. This means there is 2.5 watts of heat and radiated power. If 10% of the traces are active at any one time, this is still a significant amount of energy to dissipate. If 10% of this energy is radiated, then the board may not meet a radiation regulation.

Consider a wave with a very short rise time. When this wave reaches unterminated logic, the wave reflection tries to double the voltage at the end of the line. This excess voltage might cause damage to circuitry or perhaps cause logic to malfunction. If this occurs, some sort of resistive termination may be required. This resistor can be in series with a small capacitor so that the termination is not held over the entire clock period. A single RC termination works for the rising and falling edge of the logic signal. Another solution to this leading edge reflection problem is to terminate the line in a zener diode.

When the logic level returns to zero the reflection at the receive end of the line would then force the line to be negative with respect to common. This voltage can be clipped by a forward diode connected between the logic line and common. These clamping elements must be fast enough to be effective. If diodes are not fast enough, then a resistive termination may be required.

Note that diodes have capacitances that must be considered a part of the termination.

It is interesting to see what actually happens when a steep wave front reaches the end of a line terminated in a high impedance. At the end of the line the voltage rises until some input device breaks down and conducts and provides clamping action. The voltage rises until the current in the clamp equals the transmission line current. This requires a small reflected wave and a forward current defined by the characteristic impedance of the line. As an example, if the reflected voltage is +0.6 volts and the characteristic impedance is 60 ohms, the current in the reflected wave is –0.01 amperes. When this small wave reaches the source it is re-reflected. The re-reflected wave is –0.6 V at a current of –0.01 amperes. When this re-reflection reaches the termination end of the line, the voltage is below the firing threshold and the current flow is zero. At this point the reflections stop. For short transmission lines where the rise times are long enough this reflection process will not take place.

7.8. DECOUPLING CAPACITORS

The capacitance selected for a decoupling capacitor depends on the number of logic transitions that require energy per clock period. If the lines are short and the rise time is a few nanoseconds, the energy requirements relate to the transmission line capacitance rather than the characteristic impedance of the line. If energy must be supplied to 20 gates whose capacitance is 2 pF the decoupling capacitor can be as small as 400 pF. The voltage sag in this case is only 5%. A larger value is used for reasons that are discussed later.

It is apparent that if energy must come from a decoupling capacitor in less than a nanosecond, the connecting traces must be very short. A low series inductance requires surface-mounted components. Wide parallel paths without sharp turns are preferred. The number and size of the decoupling capacitors required on a board vary with the logic family and the clock speed. At high clock rates it may be necessary to decouple every integrated circuit. If this is not required, it is acceptable to space capacitors uniformly around the board.

Decoupling capacitors should be placed at the power and common pins of a component. If the power and common pins are on opposite sides of a socket, a long circuit trace is needed to connect to a decoupling capacitor. This length adds inductance to the decoupling circuit. If the power and common pins are connected to the power and ground planes, then the initial energy can come from these planes. (See Section 7.9.) This helps to overcome the delay caused by the lead inductance to the decoupling capacitor.

Decoupling capacitors always should be connected so that current that is supplied to the capacitor does not flow in traces that take energy from the capacitor. The right and wrong practices are shown in Figure 7.3.

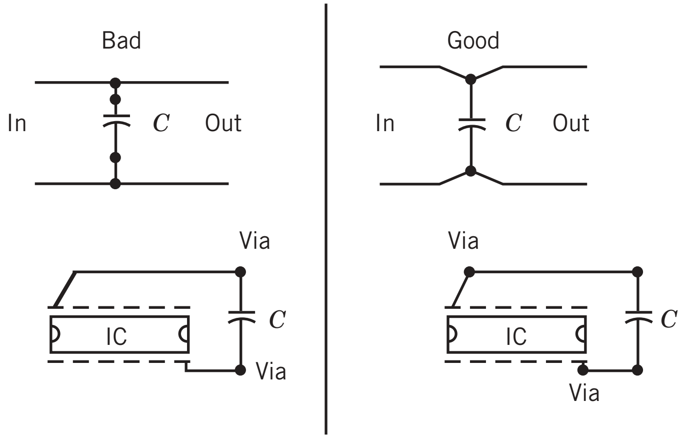


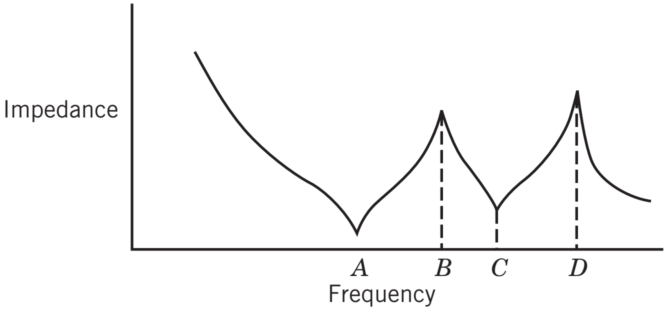
Figure 7.3. Traces connecting to decoupling capacitors

In some integrated circuits the ground and power pins are on opposite sides of the component. This geometry makes it difficult to connect to a decoupling capacitor without a long trace lead. In this situation it is good practice to connect the power and ground pins directly to the ground and power planes. In this geometry decoupling energy is available from the capacitance between the ground and power planes.

A wide range of decoupling capacitors are available on the market. Ball-grid array capacitors have parasitic series inductance as low as 60 pH, while a single component will have an inductance greater than 600 nH. This inductance is measured at the terminals and does not allow for any connection length. The natural frequency of a capacitor of 0.01 μF in series with 600 pH is 65 MHz. At 0.001 μF the natural frequency increases to 112 MHz. The natural frequency is $1/2\pi\sqrt{LC}$ where L is in henries and C is in farads.

For a decoupling capacitor at its series resonant frequency the impedance is equal to the skin effect resistance of the circuit. Above this frequency the impedance looks like an inductance. If a second capacitor value is used in parallel with the first capacitor there will be both a series and parallel resonant circuit to consider. Consider adding a large parallel capacitor. At some frequency it has a series resonance with its self-series inductance. Above this frequency this capacitor looks like an inductance. A second smaller parallel capacitor placed across the first capacitor can resonate with this inductance to form a high-impedance parallel resonant circuit at a higher frequency. A high-impedance power supply at some upper frequency is not desirable. To avoid this resonance in the decoupling structure it is advisable to use decoupling capacitors of the same value. The resonance properties of parallel capacitors are shown in Figure 7.4.

If an integrated circuit must supply logic to a group of logic lines where the characteristic impedance is a consideration, the current demand can be



A. Resonance of largest capalitor C_1 with series L_1

B. Parallel resonance of L_1 with second largest capacitor C_2

C. Series resonance of C_2 and L_2

D. Parallel resonance of C_3 and L_2

Figure 7.4. The resonant properties of three different parallel capacitors

high. Larger decoupling capacitors will be required in this application. If the logic is fast enough and the characteristic impedance of the traces is 50 ohms and if on average 10 lines are active per clock cycle, 3-V logic must supply 10 watts to the board.

N.B.

Manufacturers of integrated circuits will often specify the size, type, and number of decoupling capacitors that must be placed on the perimeter of a component. This one size should be used on the entire board.

N.B.

The "on" resistance of the switches in an integrated circuit dissipates power in the integrated circuit. This should be small compared with the power delivered to the transmission lines.

7.9. THE POWER PLANE

A power plane is a surface of copper that supplies an operating voltage to the components on a circuit board. Decoupling capacitors are still required between the ground and power plane to supply energy for logic transmission. When several power or ground planes are used they should be connected together where the decoupling capacitors mount and where the components mount. The capacitance between the ground and power plane adds to the decoupling process.

N.B.

Circuit traces can be placed on layers between the power and ground planes. The fields associated with these traces are tightly confined so there is little chance of radiation. These traces must be connected to the circuit using vias and or through-holes.

N.B.

A power plane functions just like a ground plane except that it is associated with a dc voltage.

The distributed capacitance between the power and ground plane stores energy. Several benefits result from using a power plane:

- There is no need for power traces.
- The capacitance between the power plane and the ground plane can supply some of the energy to support the transmission of logic signals.
- The fields associated with traces between the ground and power plane are tightly confined. These traces will contribute very little to any board radiation.
- Crosstalk between traces is reduced. Trace spacing need only be equal to trace width.

7.10. THE GROUND AND POWER PLANE CAPACITANCE

When an integrated circuit is mounted to the ground and power planes it is mounted directly to the interplane capacitance. This capacitance is in parallel with nearby circuit decoupling capacitors but it has a very different geometry. Some of the energy stored between the planes is available before a nearby mounted capacitor can be accessed. The majority of the energy stored between the planes is located at a distance from the logic that is being served and is not immediately available. It takes time for a request to propagate to the energy and for the energy to return to the integrated circuit and then travel down the logic trace. At very high clock rates this delay must be considered.

The capacitance between the ground and power plane can be increased if the ground and power planes are closely spaced. This assumes there are no traces between the planes. In most applications this is still not a large enough capacitance to supply the needed energy and decoupling capacitors are still needed. If the dielectric constant is increased in this space, then more energy can be stored. If traces use this region, the spacing must be increased and this reduces the capacitance. With a higher dielectric constant, line losses on longer trace runs become an issue. Note that a higher dielectric constant reduces the velocity of propagation. There is more energy available but it takes longer to bring it to the transmission lines needing the energy. If the clock rate is low enough, this delay may not be an issue.

7.11. USING VIAS

Traces can carry a signal between layers by using a via (Figure 7.5). It is important to consider the path the electromagnetic energy must take when a via is used. If a via carries a trace to the opposite side of a ground or power plane, then the field must "flip" to the other side of the plane. This transition creates a minor impedance discontinuity. If the via carries a trace to a layer that uses a different ground or power plane for its return path, then the impedance discontinuity can be a problem. The easiest way to see the difficulty is to consider the return path for current flow. If there is a nearby via connecting the ground planes, the signal loop area is controlled. If there is a nearby decoupling capacitor associated with the ground or power planes, then the signal loop area is also controlled. If the return path is a distance away, the field associated with the signal transmission must fan out in the space between

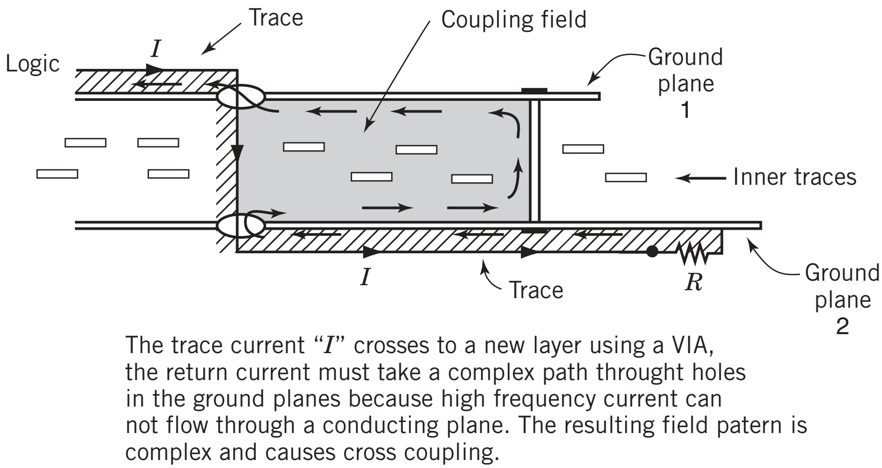


Figure 7.5. A via carrying a trace between layers on a multilayer board

planes, causing a significant discontinuity and adding to crosstalk. Remember that fields cannot penetrate a conductive plane. The only way field energy can get from one layer to another is through some sort of opening. Fields can use a ground or power planes as a reference conductor. A connection between planes can be a decoupling capacitor or a simple via.

> **N.B.**
>
> For clock rates above 1 GHz the reflections caused by vias can create problems.

7.12. DECOUPLING CAPACITORS AS TRANSMISSION LINES

The delay in getting energy from a capacitor to a load is usually attributed to a series inductance. This inductance involves lead length and circuit loop area. It is interesting to consider this delay in terms of transmission line phenomena rather than apply linear circuit theory.

A capacitor can be viewed as a short segment of transmission line. It has two parallel conductors separated by a dielectric. The characteristic impedance of this geometry is low as the intent is to have a high capacitance per unit length. The traces that connect to the capacitor are transmission lines with a higher characteristic impedance. This geometry is shown in Figure 7.6. We start the process by assuming there is a voltage V on the capacitor. After a switch connects a load to this conductor geometry, a series of reflections connects the capacitor to the load resistor. In this configuration we can assume that the transmission line representing the capacitor is open circuited on one end.

When the switch closes, a wave is sent to the load Z_3 and a second negative wave is sent toward the capacitor along Z_2. These waves have an amplitude equal to $VZ_2/(Z_2 + Z_3)$ and $-VZ_3/(Z_2 + Z_3)$. When the wave that propagates along Z_2 reaches the junction between Z_1 and Z_2, there is another reflection. The fraction of the wave that reaches the open end of Z_1 reflects again and continues forward. The reflection processes occur at every impedance mismatch in both directions. In Figure 7.6, time is represented vertically. The notation τ represents a transmission fraction and the notation ρ represents a reflection fraction. For example, when a wave V_1 traveling along Z_1 reaches Z_2, the reflection term is $V_1(Z_2 - Z_1)/(Z_2 + Z_1)$ and the transmission term is $2V_1Z_2/(Z_1 + Z_2)$. At the open end of the transmission line the impedance is considered infinite. In this case there is only a reflection. To simplify the process, the value of Z_3 has been set equal to Z_2.

There are reflections at points numbered 0, 1-2, 2-3, and 3-Z_3. It is obvious from Figure 7.6 that there will be an infinite number of reflections. The second wave to reach the load Z_3 is delayed by the time it takes for a wave to make

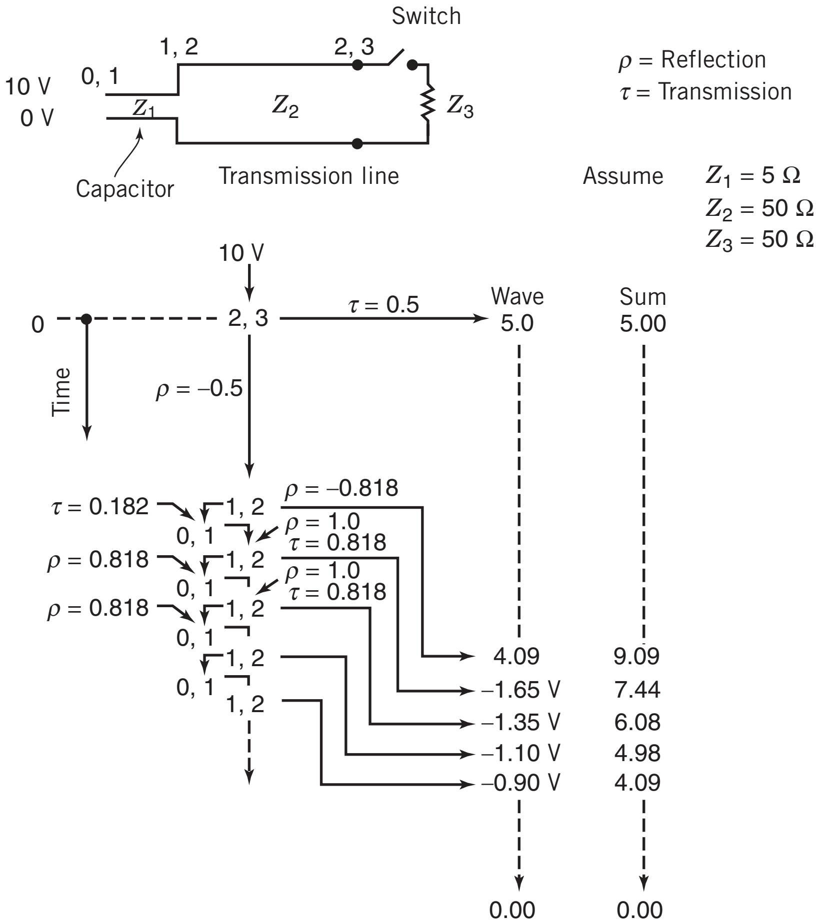
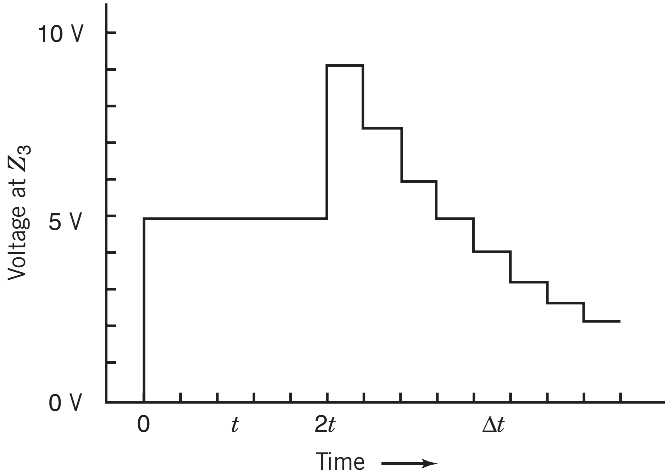


Figure 7.6. A capacitor as a transmission line

the round trip from point 3 to 2 and back to 3. After the first reflection the voltage at point 3 will rise to a voltage nearly equal to the initial voltage V. Eventually the capacitor will be fully discharged into the impedance Z_3. The discharge pattern is shown in Figure 7.7.

If Z_3 does not equal Z_2, then some fraction of every wave that returns to Z_3 reflects back to the capacitor. Every reflection starts an infinite series of reflections in the capacitor. The result is a very complicated sequence of events that is best handled by a computer.

The voltage at Z_3 is a set of step functions. For most practical geometries this process occurs so fast that these details are not visible. Is it correct to say that this delay is the result of inductance, or is it the result of transit time? After all, inductance per unit length was used in calculating the characteristic impedances of the two transmission lines. There is no simple answer.



Note: t is the time it takes for a wave to propagate from the capacitor to the load Z_3.
Δt is the time it takes for a wave to propagate across the capacitor and return.

Figure 7.7. The voltage at the impedance Z_3 as a function of time

The energy stored in the capacitor is eventually dissipated in resistor Z_3. This energy is transported by reflections on transmission lines. This is a viewpoint that is not considered in circuit theory. Remember the analogy with the pond when energy was added by dropping a stone in the water. Wave action had to carry the energy to all points in the pond. This transport cannot be easily explained using sine waves.

7.13. CHARACTERISTIC IMPEDANCE CONTROL

Impedance is a sinusoidal concept involving the phase angle between current and voltage. If the current and voltage are in phase, the term *resistance* is used. If the current lags by 90 electrical degrees, the term *inductance* is used. If the current leads by 90 electrical degrees, the term *capacitance* is used. In handling communication transmitters on a PC board the control and matching of transmission line impedances plays an important role. A mismatch in impedance can directly relate to transmission efficiency.

In logic the intent is not to radiate. The step functions that are used contain a wide energy spectrum that extends well below the clock frequency to 10 times the clock frequency. Any energy that is put out on a transmission line must dissipate. The process involves reflections, resistive heating, dielectric losses, and some radiation. Termination by its very nature requires dissipation over the entire clock cycle in the termination resistance. Reflections add to the radiation but these signals die down exponentially.

> **N.B.**
>
> The characteristic impedance of a transmission line is a function of the dielectric constant and this varies about 20% from 1 MHz to 1 GHz.

For a given transmission line it is possible to relate geometric anomalies to impedances at a particular frequency. Since there are so many variables there is no simple way to use this information to relate to the impact a specific anomaly will have on a given logic transmission line. For example, a measurement might indicate that a via may look like 25 ohms at 2 GHz. At 100 MHz the effect of the via on logic may not be sensed. The effect on a step function may be a slight change in waveform.

The following list illustrates the problems of controlling characteristic impedance:

- Glass and epoxy have different dielectric constants. The distribution of materials on a board may not be uniform so that the characteristic impedance of a transmission line will vary along its length and in different parts of the board.
- The characteristic impedance of a transmission line is a function of nearby traces and components. Equations and tables used to obtain these values do not address this fact.
- Water has a dielectric constant of 60 and any moisture absorption on the surface of a board will affect the transmission line impedance.
- The presence of coatings or legends will affect the dielectric constant and thus the characteristic impedances of traces. If the traces are embedded in a dielectric, this will not happen.

> **N.B.**
>
> All of the above-listed factors affect the velocity of propagation.

A basic question in digital design involves the decision to control the characteristic impedance by testing every board. If logic lines are unterminated, it would seem that impedance control is unnecessary. Since characteristic impedance does control reflections and board losses it is a good idea to design the board with the same character over the entire board. This means that inner traces will have different spacings than outer traces. For logic boards it may be unnecessary to ask the manufacturer to control the impedance of a test trace as circuit variability is far greater than board variability.

There are many ways to obtain the characteristic impedance for various conductor geometries. There are free calculators on the Internet and there are services that will provide information for a fee. Tables and equations are available for every possible arrangement including embedded traces, surface traces, inner traces, offset traces, stacked traces, parallel traces, unsymmetric traces, and so forth. It is wise to keep the characteristic impedances in the range 40 to 70 ohms. For lower impedances the radiation problem is increased and some ICs may have trouble supplying the current. For higher impedances there are problems with crosstalk and radiation. In cases where the logic must drive a controlled bus that has a specified impedance, greater care should be taken in testing the boards. If tight impedance control is necessary, a correction should be made for nearby traces. In general, nearby traces will increase the characteristic impedance by about 5%.

7.14. RADIATION FROM DIGITAL BOARDS

The traces and components on the outer surface of a board are most apt to contribute to radiation. The fields associated with traces over a ground plane are confined mainly to the volume under the trace. Only a small fraction of the field is above each trace. As an example, if there are a thousand traces, this field should be considered. The simplest approximation is to consider the loop areas of all the traces that are active at any one time. The loop area is the space between the trace and the ground plane times the trace length. The frequency to consider is $1/\pi\tau_r$ and the amplitude to consider is the power supply voltage. The field from all component loops should be added to this field. This includes decoupling capacitors and component connections. Equation (5.7) should then be used to determine the worst-case radiation in the far field. The H field can be calculated using the wave impedance of free space.

The fields associated with edge traces are tightly contained if they are between conducting planes. To further limit radiation, edge trace runs should be kept short and if necessary a guard ground trace can be added to the perimeter of that layer. If there is room on the board, then traces should be kept two widths from the edge of the ground or power plane.

7.15. MEASUREMENT PROBLEMS—GROUND BOUNCE

The loop areas formed by an oscilloscope probe and the probe shield are large compared with the loop areas on the circuit board. This means that logic levels are observed together with any logic and transient phenomena. The loop area at the probe tip can be reduced by moving the reference connection to a point near the signal point. A loop area of one cm^2 is often equal to the loop area of 1,000 traces.

The radiated field near the board will use the probe braid as a conductor and leave the circuit on this conductor. The transfer impedance of the probe will allow this radiation to add to the observed signal. To test for this coupling the probe tip can be connected to the probe common on the board. If there is no signal, then there is no transfer impedance coupling.

A test that is often performed is to connect the oscilloscope between two ground points on a circuit board. If a voltage is detected, it is often blamed on an inadequate "ground." The ohms per square for a copper ground plane even at 100 MHz is only a few milliohms. One volt associated with the ground plane would imply 1,000 amperes of current flow. The only way to explain this voltage is to recognize that there must be a changing electromagnetic field in the area near the board. This can be verified by creating a loop with the same area and using it to sense and measure the field. The loop should be oriented the same way as the oscilloscope probe and it should be left floating. A sniffer probe using a shielded cable is shown in Figure 5.11.

When ground potential differences are observed the effect is called ground bounce. The term is unfortunate because it suggests ground current. To reduce this effect the loops that are creating the field must be made smaller. Another problem may be attributed to decoupling capacitors, which may have to be less inductive or relocated. If the signal that is sensed is the result of probe transfer impedance, then again the radiated field from the board is creating the signal. A better shield on the probe may be necessary to limit this coupling.

N.B.

Fields near a board (ground bounce) cannot be reduced by using thicker copper ground planes.

7.16. HIGH CLOCK RATES

The demand for fast logic has prompted manufacturers to introduce new logic families. These logic families run at clock rates where there is little time for trace reflections to attenuate. One approach to limit reflections is to match impedances at the receive end of the logic transmission (parallel termination). As an example, at 24 GHz the clock period is only 42 ps. In this time period the wave would travel about 0.5 inches. Assume a rise time of 4 ps. In one-quarter of this time the wave would travel a little over a tenth of an inch. To avoid reflections in this situation the receiving logic must terminate the line in its characteristic impedance.

Matching impedances implies dissipating transmitted energy in resistors. Assume there are 100 logic transitions per clock cycle and the voltage level is 3 V. If the terminating resistors are 50 ohms, the dissipation is 18 W. If the logic is active 50% of the time, the board must dissipate 9 watts. This can pose a difficult heat sinking problem.

N.B.

Some circuit board manufacturers can embed terminating resistors in vias or through-holes.

Assume the source impedance matches the line and the transmitted voltage is half the power supply voltage or 1.5 V. If the receive end is an open circuit, the reflected wave doubles. When this reflected wave reaches the source it is absorbed and the reflections are over. There are two problems with this approach. Each transmission line requires a matching resistance and the settling time is the time it takes for one round trip. Now assume that only the receiving end of the line is terminated. The full voltage reaches the termination and there is no reflection. The only disadvantage is that any interference coupled to the line is reflected at the source because there is no matching impedance. In this approach the transmit end can fan out on more than one line.

To limit dissipation, the transmitted logic could be a pulse. This pulse could trigger a setting in the receive logic that would then be detected at the clock transition. If the pulse were one-eighth of a clock period, the dissipation would be reduced by a factor of eight.

When the logic family provides internal terminating or source impedances, the traces carrying logic must match this characteristic impedance. This means that the designer should specify all of the parameters that control the characteristic impedance of the trace. These parameters include trace width, thickness, spacing, and the relative dielectric constant. The characteristic impedance also will be a function of whether there is a ground plane, whether the traces are between two conducting planes, and whether there are nearby traces.

N.B.

Nearby traces change the characteristic impedance of a transmission line.

Where possible, logic lines should be straight lines without branches or vias. If there are turns, they should have a radius. Transmission line anomalies introduce reflections that affect waveforms. Any reflected energy is directed back to the driving source. If the driver does not match the line impedance, this reflection will be re-reflected and it will add to the logic waveform. A small amount of re-reflected energy is acceptable depending on the logic family and the trace length.

N.B.

Vias that pass a trace through one ground plane are less of a problem than vias that cross several planes.

N.B.

At clock rates where line terminations are required, care needs to be taken in controlling characteristic impedance and limiting discontinuities along the transmission path.

A technique that is used in testing logic transmission is called an *eye pattern*. A logic stream consisting of a pattern of bits is monitored at various circuit points. The pattern that is viewed is an envelope of signals covering three logic transitions. The oscilloscope is synched with the clock signal. The envelope that is displayed includes delays, overshoots, undershoots, and reflections. In a good response there is a pattern representing ones and zeros with a clear region between transitions. These clear regions are the "eyes." If the response is poor, the eye will fill in with transitions, which means that the logic is suspect. In critical applications eye patterns can be improved by using better board material.

7.17. BALANCED TRANSMISSION

Balanced logic transmission is used when common-mode coupling is expected. Logic is transmitted on a pair of conductors so that line A is the inverse of line B. The receive logic will respond only if A ≠ B. For these balanced signals a ground plane is not required. This is because the field is confined to the space between the balanced lines. The trace lengths for the balanced logic pair should be equal. Clock and logic signals may have to be routed to accommodate transit time differences.

The difference signal is called the *even mode* and the common-mode is called the *odd mode*. Since these processes are logical there is no meaning to the term *rejection ratio* that is used in linear systems. The odd mode either is rejected or it is not. A simple balanced logic line is shown in Figure 7.8.

The characteristic impedance of traces used for balanced transmission depends on the trace orientation and on whether there are conducting planes associated with the traces. If the traces are not associated with conducting planes, the characteristic impedance is undefined for unbalanced signals. If there are conducting planes, the characteristic impedance will be different for normal-mode and common-mode signals. The impedances will also depend on

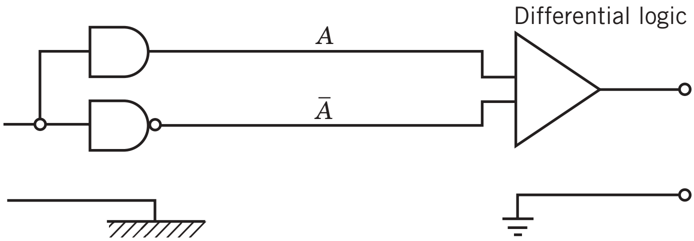


Figure 7.8. A balanced logic system

whether the traces are side by side or one above the other. If the logic does not terminate the transmission line, then the characteristic impedance of these lines does not need to be controlled.

7.18. RIBBON CABLE AND CONNECTORS

Ribbon cables are insulated, parallel, evenly spaced conductors. The cable provides a very orderly way to carry a group of conductors between two points. Ribbon cables can be soldered into place or terminated in connectors. Once conductors leave a circuit board they are no longer closely spaced or close to a ground plane. The conductor length and increased spacing adds to the loop areas and this increases the chances of radiation and susceptibility.

It is good practice to provide more than one ground conductor in a ribbon cable carrying logic. If every third conductor is a ground, then every logic conductor has a nearby return path. This scheme will not be effective unless the ground conductors are individually carried through the mating connectors and are terminated on ground planes at both ends of the cable.

Ribbon cables are available with a copper backing on one or both sides of the cable (microstrip or stripline). This backing can serve as a ground plane provided the plane is properly terminated at the ends of the cable. This copper should not be considered a shield but an extension of the ground plane. A single connection to this sheet of copper at either end of the ribbon run will negate the value of the plane.

Ribbon cables frequently cross open areas. This increases the chances of common-mode coupling. To limit this coupling the cables should be routed along conducting surfaces. Excess cable should not be coiled as this also increases the opportunity for common-mode coupling.

A connection between a chassis and a circuit board ground will do little to reduce the field in the area. To limit these fields, apertures may require closure or the circuits should be designed with more care. When the cable is routed on the conducting surface the coupling is reduced independent of where a ground connection is made. For short cable runs the cable routing may not need to be controlled.

The preferred way to locate multiple connectors on a circuit board is on one edge. If the connectors are located on opposite edges, the entire board together with cables becomes a path for common-mode coupling. Grounding the board to a chassis will do little to control this flow.

7.19. DAUGHTER BOARDS

It is important to control the logic transmission fields that cross between a main board and a plug-in board or *Daughter Board*. The problems discussed for ribbon cables apply. Every logic line should have a ground or decoupled power pin as a neighbor. If this is not done, there is the chance of radiation or of susceptibility.

The logic pin and its associated ground pin should terminate on the same side of the board. If this is not done, the return current path is long and there can be problems. There is no way for current to flow through a ground plane. The current must travel to the edge to get around to the other side. This means that the transmission line will have a significant discontinuity. This discontinuity is inductive and it can radiate and be susceptible to crosstalk or interference.

7.20. MIXING ANALOG AND DIGITAL CIRCUITS

It is often necessary to interface analog and digital circuitry. At digital speeds the differential approaches suggested for analog coupling are hardly appropriate. There are several schools of thought, which include separate grounding schemes and even separate boards. Somewhere the analog ground must come in contact with the digital ground. The problem is usually related to an analog-to-digital (A/D) converter. If 14-bit accuracy is required, the error for a 10-volt full-scale signal is only 0.5 mV. If two grounds are involved, the noise voltage that is sensed can easily exceed 10 times this figure. If the two grounds are connected together by a single conductor, the loop area that is formed is great enough that the ground potential difference will still be a problem.

The best solution for interfacing analog and digital functions is to use a common unbroken ground plane. It is important to make sure that the fields associated with analog and digital functions do not share the same physical space. The following six rules apply:

1. Analog circuit components must not be intermixed with digital circuit components.
2. Make sure that the analog fields do not share the same physical space as the digital fields.
3. The pin assignments in connectors should separate function so that the analog fields and digital fields are separated.

4. The A/D converter should have an internal forward referencing amplifier.
5. Orient the A/D converter to limit field coupling.
6. The circuits should pull energy from different decoupling capacitors.

7.21. OPTICAL ISOLATION

Optical isolation can be provided by a light-emitting diode (LED)/transistor pair. The transistor acts as a switch when the LED emits light. This arrangement allows external logic to be ohmically isolated from the circuit receiving the logic state. Components are available with a number of these isolating pairs.

The leads carrying the logic to the optical isolator can carry field energy from the external electrical environment. If these leads are brought into an enclosure, they can radiate around the optical isolators and couple to the internal circuitry. This means that optical isolators should be treated just like a power line filter. The isolators must straddle the interface between the outside environment and the inside of the circuit enclosure. The conductors carrying the logic must be kept outside of the enclosure. This is difficult to implement if the optical isolators are mounted on the very circuit board that is affected.

7.22. GOLD PLATING

There are many techniques for making connections to printed circuit boards. The reliability of these connections is an important consideration. In one approach the connections are formed as fingers on the edge of the board. The board is then plugged into a fixed connector. In other approaches connectors are soldered to an array of pads.

Gold plating is usually applied to the conductors that make contact through a connector. This is done to avoid the oxidation that can build up on plated surfaces. This often requires gold deposition on selected parts of the board. In many applications where the voltages are greater than 12 V the gold may not be required. For analog signals (low frequency) or for low logic levels, gold is the only way to guarantee a good connection over time. There are cases where equipment stays idle for long periods of time or where a logic line rarely changes state. It should not be necessary to reinsert the card or connector to establish a good connection. The only way to avoid oxidation is to use gold plating.

The thickness of the plating depends on how often the connection is to be broken. For frequent disconnects the gold plating should be thick enough to tolerate the expected wear.

7.23. GHZ NOTES

Special attention must be given to circuit designs that operate at GHz clock rates.

1. Decoupling capacitors must be resupplied energy as rapidly as this energy is depleted.
2. Vias do not have low characteristic impedances. If energy must transfer between several ground/power planes then vias may not be adequate. Care must be taken so that field energy is not forced to transfer between planes at the board edges.
3. The time it takes to obtain energy from a ground/power plane geometry depends on planar spacing not on the dielectric constant.
4. The self inductance of a capacitor does not adequately measure the time it takes to move energy in and out of a capacitor.
5. The transfer of energy over short sections of typical circuit trace may take nanoseconds. This is the result of impedance mismatches. This time can be greater than one clock period.

CHAPTER 8

Facility Hardware

8.1. GROUND PLANES

A ground plane is a conducting surface. The size of a ground plane depends on application. For a printed circuit board the size of the plane might be as small as a square inch. A large printed circuit board ground plane might be 12 by 18 inches. In an electronics facility a constructed ground plane might be the size of a room. Ground planes include the side panels in racks, a strip of aluminum foil, or for that matter, the earth under a building. A metal box can be considered a ground plane although the conducting surface is not flat. If the conducting surface extends from a floor onto a wall, it is still a ground plane.

In Chapter 7, the printed circuit board ground plane was used as a return path for circuit currents. Stated another way, the fields that carry logic or signal information are carried between traces and the ground plane. If a conducting surface is flexible and it turns a right angle, it still functions as a ground plane.

> **N.B.**
>
> It is not practical to use the ground plane in a facility to provide a return path for logic or signal information. The ground plane in a facility can be involved only in limiting interference coupling.

A facility ground plane does not attenuate or eliminate electromagnetic fields. The only role a conductive plane can have is to reflect an arriving wave that has an E field that has a component that is tangent to the plane. Waves with a vertical E field can propagate along the surface of the plane without attenuation. The associated H field that is parallel to the conducting surface simply causes a surface current to flow.

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When a cable is routed parallel to a ground plane the coupling to interference fields will be common-mode in nature. The coupling will be maximum when the wave and the cable run in the same direction. The coupling is proportional to the loop area formed by the cable and the ground plane. For sinusoids the maximum coupling occurs when the cable length is one-half wavelength. This assumes the ground plane is continuous along the path of the cable. If the cable run is greater than one-half wavelength, then the coupling for a half wavelength is used for a worst-case analysis. If there were no loop area, there would be no coupling.

8.2. A FACILITY GROUND PLANE USING STRINGERS

One form of constructed ground plane in an electronic facility consists of conducting stringers that form an open grid. Typically this grid is supported on stantions that raise the floor about three feet above a concrete surface. A raised-floor ground plane is shown in Figure 8.1. The stringers are usually on 30-inch centers.

There are several benefits to having a raised floor. Cables can be routed between racks under the floor and this allows for a clean-looking installation. The space under the floor can form a plenum chamber that can be used to supply cooling air for the electronics. The floor tiles that are supported by the stringers are usually made slightly conductive to dissipate any possible

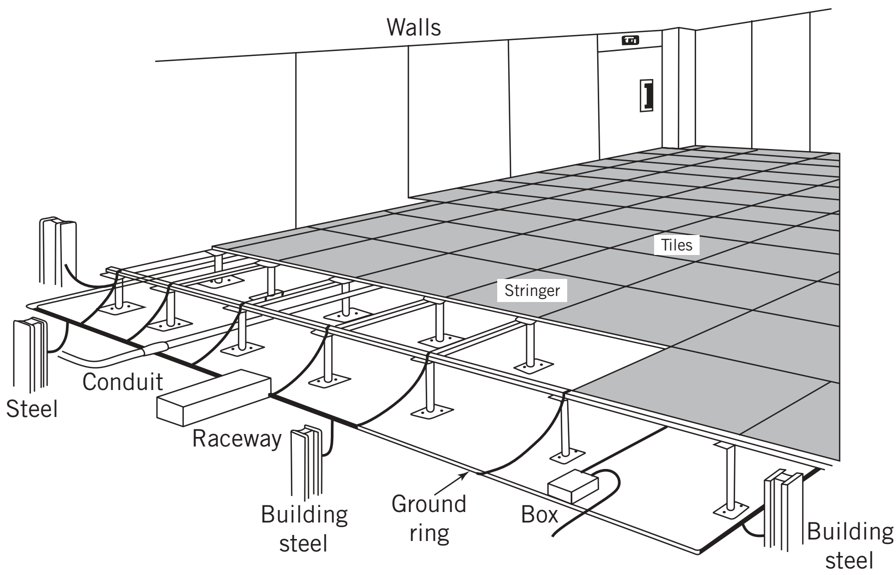


Figure 8.1. A typical ground plane in an electronic installation

electrostatic charge buildup on personnel. A typical human body has a capacitance to the floor of about 300 pF. A time constant of one-tenth second means the floor tiles should provide a resistive path of about 300 megohms. These tiles usually have a resistivity of about 10^7 ohm-centimeters.

In Section 6.8, we discussed the ohms per square of a sheet of copper. At a frequency of 1 MHz the ohms per square for 1 mil copper is 369 μohms. To make the stringer ground plane look as good as a sheet of copper the resistive connections between the stringers should also be microhms. This is accomplished by manufacturing the stringers with plated surfaces and using spring-type washers to bolt the stringers together. This way the contact area is under constant pressure. The intent is to maintain a good connection over time for a range of floor loads and temperature changes.

A ground plane constructed of well-bonded stringers provides excellent protection against lightning. If the current pulse is distributed evenly across the ground plane, the potential difference across the plane can be held to a few volts. This means that the ground plane must be multiply connected to building steel and to the grounding electrode system of the facility. These connections are shown in Figure 8.1.

The stringer ground plane is connected to each rack of electronics, which in turn is connected to *equipment ground* for every piece of equipment. The line filters in equipment have shunt capacitors that connect to equipment ground. The stringer system connects to this equipment ground and provides multiple paths for these power filter currents. For this reason the stringer system is effective in limiting the intensity of interference fields associated with this current flow.

Each tile that fits between the stringers must make contact with the stringers around the perimeter of the tile. When tiles are removed and then replaced the hardware that provides this connection must be reinstalled or the protection against charge buildup will not be present.

N.B.

The best way to control ESD (Electro Static Discharge) is to control humidity. If the humidity is above 30%, there is little chance of charge buildup. The best way to control humidity is in a central air conditioning unit. Added humidifiers often leave areas without adequate protection.

N.B.

Rotating floor polishers should not be used on floors near electronic hardware. There is a risk of generating ESD.

In Section 6.13 we discussed dependent apertures. This type of aperture does not allow surface current to flow freely around an opening. This is the case with a stringer system. The stringers act like a wire mesh except that the conductors and openings are much larger. For a wire mesh to be effective it has to be bonded to an enclosure around its perimeter. In the case of a ground plane made from stringers there are no controlled conductors around the perimeter to use for aperture closure. For this reason external fields can be present on both sides of the ground plane.

For a ground plane to be effective in a facility, all equipment racks should be bonded to the ground plane at their base. In effect the racks should be extensions of the ground plane. Of even more significance is the routing of connecting cables. The usual procedure is to drop the cables to the concrete floor under the racks. The loop areas between the cables and the stringer ground plane can be large. These loop areas allow common-mode coupling for fields that exist under the stringer floor. The ground plane is ineffective in controlling this coupling. The loop area between the stringers and any interconnecting cables has led engineers to add a ground plane on the concrete surface under the installed ground plane. This ground plane is often made of wire mesh or even of metal foil. To be effective this added ground plane should follow the cables into the racks. Because this is rarely done and there are few problems it seems logical to conclude that common-mode coupling to cables in a typical installation is not a problem.

N.B.

A ground plane is a conductive geometry that can limit common-mode field coupling to interconnecting cables.

8.3. OTHER GROUND PLANES

The rebars in a concrete slab might be considered a ground plane. Rebars are buried in the concrete and multiple surface connections are not provided. Because the bars are not welded at every crossing they do not constitute a proper grid structure. The presence of this steel should have little impact on the electrical performance of a facility. Rebars are simply more conductive material like water pipes, gas lines, steel beams, or metal trays. These conductors modify electromagnetic fields but they are not surfaces that should be expected to limit common-mode coupling.

A tray used for transporting cabling is made of sections that are bolted or riveted together. This type of construction makes the tray ineffective as a ground plane for controlling radiated fields. A continuous conductive liner can be added to the tray to form a ground plane. At the ends of the tray, the liner must be bonded across its entire width to a continuation of the ground plane.

This liner is not a shield. It must provide a continuous conducting surface for current flow between hardware items located at the ends of the tray. This is the only way a tray can be an effective extension of the ground plane.

Racks of hardware are often bolted together. This bond can be unreliable unless the connecting surfaces are prepared and the bolts are under spring pressure. A conducting surface can be added to the floor of connected racks to form a ground plane. This added surface must be bonded across its width at the ends to the rack framework to be effective. This added conductor should not concentrate the current flow at points of connection.

8.4. GROUND PLANES AND REMOTE SITES

Remote sites often house electronics in a trailer. If utility power is provided to the site, there can be a safety issue. If the site is on the desert floor, then an adequate local earth connection may not be available. The vehicle frame becomes the *grounding electrode system*. All metal objects external to the vehicle should be connected to the vehicle. If this is not done, there is a possible shock hazard. If there is a power fault or if there is a lightning strike nearby, there can be a potential difference between the vehicle and these conductive external objects. For example, a metal floor mat and a nearby metal fence should be bonded to the vehicle chassis. If this bond is not present, then any potential difference can be dangerous.

8.5. EXTENDING GROUND PLANES

To extend a ground plane from one room to another it is sufficient to make a series of spaced connections through the separating wall. A typical solution would be to connect the stringers together using #10 wire on 6-inch centers across the width of the room. The conductors might be welded to the stringer or bolted onto a prepared surface.

To extend a ground plane to a second floor the planes must connect by an extension up one wall. In this case cables routed between floors must cross to the other side of the plane. This crossing should pose no problem. The plane on the wall can be made of a wire mesh as long as the conductors are bonded at every intersection. The mesh should be welded or bolted to the stringer system along the width of the room.

It is not practical to attempt to use one ground plane that extends between buildings. An accepted compromise is to place cables inside a large conducting conduit and bond the conduit to the ground planes on each end. The fields associated with the lightning will probably stay on the outside of the conduit. This reduces the risk of lightning following one of the cables into the hardware. Optical fibers are an effective way to transport signals between buildings without having to consider the potential difference between facilities.

8.6. SEPARATELY DERIVED POWER

The power in a facility enters all electronic hardware. Even with hardware filters the power leads still carry interference fields. In particular the potential difference between neutral and equipment ground can be troublesome. These voltages are often harmonics of the power frequency and are not easily filtered. To avoid this type of problem a separately derived source of power can be supplied to a block load in a facility. This allows a new neutral to be created that is *grounded* to the *grounding electrode system* of the facility. The potential difference between the new *neutral* and *equipment ground* for any connected load is again zero. This separately derived source of power is often called "clean" power.

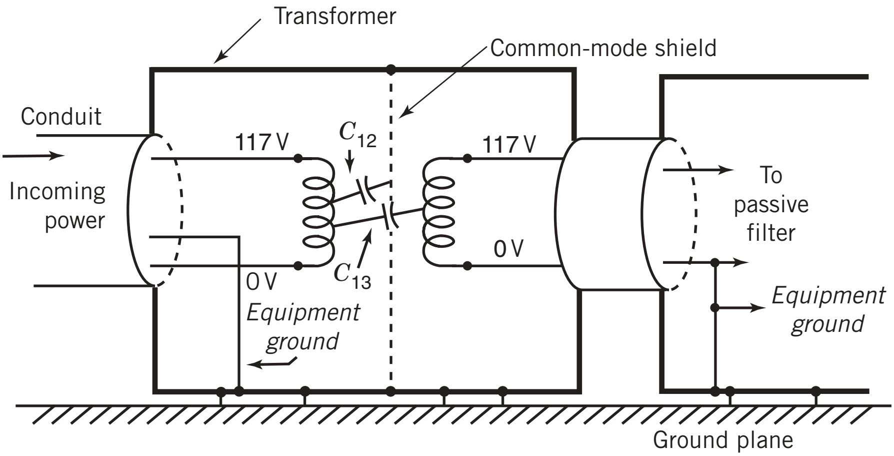
The transformer used for this arrangement is often a part of a computer power center (CPC). The NEC allows this type of power center to be mounted on the ground plane along with the hardware that is powered from this source. The center also provides breakers, line filters, and surge protection. A single-phase version of this transformer and its shields are shown in Figure 8.2c. In a three-phase transformer there would be nine shields. In this figure we show only three shields in a single-phase version. It is worth discussing how the multiple shields in this transformer function.

In Figure 8.2a the shield is called a common-mode shield. The shield stops common-mode current from flowing in the parasitic capacitance between the primary and secondary coils. With the shield connected to the nearby *equipment ground*, the reactive common-mode current flows from the primary coil to the input conduit. There is still some leakage capacitance around the shield indicated by C_{13}.

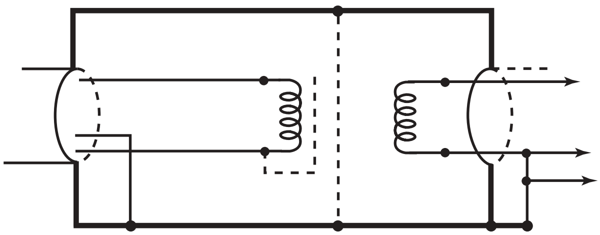
There is a problem with this shielding approach. Notice that some of the common-mode current flows in some of the turns of the primary through C_{12} and C_{13}. This current causes current flow in the secondary coil by transformer action. To stop this coupling mechanism a second electrostatic shield is added. The shield is shown in Figure 8.2b. The shield is connected to one side of the primary circuit. Now the common-mode current does not flow in turns of the primary coil.

Common-mode signals can be generated on the load side of the transformer. To stop these signals from propagating by transformer action toward the source of power, a third shield is added that is connected to one side of the secondary. This shield is shown in Figure 8.2c.

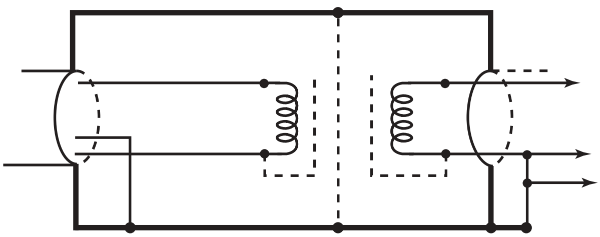
In a CPC transformer, all of the shields are connected internally. The user has no choice in how the connections are made. The shields are effective to about 100 kHz. Above this frequency, mutual capacitances allow leakage current to bypass the shields. To limit this coupling, a passive line filter is placed after the transformer secondary. The leakage current that crosses the transformer flows on the inside of the connecting conduit and is attenuated by the passive filter.



(a) One shield



(b) Two shields



(c) Three shields

Figure 8.2. The transformer used in a computer power center

N.B.

The *k* factor for a CPC transfomer is discussed in Section 2.9.

A separately derived source of power using a multishielded transformer need not be a part of a ground plane or a CPC. To be effective it should be installed close to the loads it is to service.

The leakage fields from facility transformers can couple to nearby structural steel and circulate current throughout an entire facility. If necessary, nearby loops of steel can be opened by using an insulating plate. The preferred method is to mount the transformer on a wooden structure.

Transformers should not be mounted to a conductive surface by placing screws through the core material. Any conductive path that includes the magnetic flux of the transformer is a shorted turn, causing the transformer to overheat. If screws are used they should not form a conducting loop through the core. This rule applies to all transformers from a few watts to kilowatts.

8.7. SURGE PROTECTION

Surge protection is needed to limit transient overvoltages on power supplied to critical systems. These transients can be the result of a lightning strike or from an interruption in current in an inductive load. The best practice is to provide protection at points along the power distribution path. The first line of defense can be at the *service entrance*. Surge protection is usually provided as a part of a CPC.

The least expensive surge protection components are metal-oxide-varistors (MOVs), which are usually placed directly across the power line. These components are available in a variety of power ratings and voltages. The varistor starts to conduct in nanoseconds but it has two disadvantages: Repeated surges tend to limit its effectivity and the peak voltage across the device can be two or three times its firing voltage. In critical applications MOVs are used in parallel with silicon clamping devices (zener diodes). These devices have a sharper knee but they are not as robust as the MOVs. A commercial surge suppressor is preferred over installing individual components. They should be a part of a facility design. If MOVs are used, they should be replaced on a routine basis.

8.8. THE ISOLATION TRANSFORMER

When there is interference, circuit designers often turn to adding a transformer in the power path to somehow solve the problem. These added transformers are called *isolation transformers*. All transformers provide a form of isolation because primary coils are not ohmically connected to secondary coils.

Commercially available isolation transformers provide ohmic isolation but with added electrostatic shields. Shield leads are brought out of the transformer so the user can decide on the connections. In some isolation transformers the primary and secondary coils are brought out inside of shielded braid. As an example, the primary coil is brought out inside of the primary shield. As we shall see, connecting these shields in a useful manner poses several problems.

An isolation transformer is often added to a rack full of hardware. The transformer is usually 120 V in and 120 V out. In powering a rack, the shield next to the secondary coil should be connected to one side of the secondary. The question is often raised, Should one side of the secondary coil be grounded to the rack? The answer is complicated because the *grounding* of the rack is usually made by the *equipment ground* connection at the normal power connection. Most isolation transformers are not approved for use as a source of *separately derived* power. If the secondary of the isolation transformer is left floating, there is a risk of shock. If the secondary hot lead should fault to the rack, there is no mechanism to detect the short. The NEC requires that the rack be *grounded* because it is considered electrical *equipment*. This *grounding* occurs where the power plug-strip is mounted. If the isolation transformer supplies the power plug-strip, how does the fault protection system function in conjunction with a floating power source? Before power is isolated with an added transformer these aspects of safety should be resolved.

Transformers that are mounted on the inside of hardware are not covered by the NEC. The codes that govern this aspect of hardware design are covered by Underwriters Laboratory (UL) or its equivalent. The manufacturers of consumer goods must follow these rules and get acceptance before placing a product on the market. Many special hardware items are designed for rack mounting and are not UL approved. An isolation transformer used to power a rack of hardware may be in this class.

N.B.

Long shield connections to a transformer make the shields ineffective.

8.9. SCREEN ROOMS

A screen room is a large metal enclosure. It is designed to restrict the flow of electromagnetic energy either into the enclosure or out of the enclosure. In one application a screen room can provide a field-free environment so that hardware can be tested for radiation. A screen room also can be used to process secret data so that all electrical activity is confined to that room.

A single uncontrolled conductor that enters the enclosure can destroy its electrical integrity. This means that power conductors, telephone lines, and any testing cables must be correctly shielded and filtered at the wall of the screen room. It is good practice to have a single controlled area for power entry and cable entry. The room must have a door and it must be ventilated. These potential apertures must be closed electrically or the room will not be electrically tight.

A large part of the cost of building a screen room is the sheer mass of the steel used to form the room. This steel is needed to attenuate near induction fields. Typical screen rooms are built of steel panels that are welded together. This type of construction provides a path for surface currents on the inside and outside surfaces of the room. All inside room edges use welded fillets so that surface currents do not have to make sharp turns to transition between walls. Without these fillets, surface current would penetrate the walls. The steel that is used should have permeability at low flux levels or these fields will not be adequately attenuated in this part of the spectrum.

It is good practice to limit surface currents on the outside walls of a screen room. To achieve 80 dB of attenuation the walls must have a thickness of at least 10 skin depths. It is easier to limit surface current than to add thickness to the walls. One way to limit outside surface currents is to concentrate all electrical connections in one area.

The equipment ground connection to the screen room provides adequate electrical safety. A second ground only invites the flow of additional surface current on the outside wall. If lightning were to hit the room, the field that would enter the room would hardly be noticed. There is no reason for a second grounding conductor to earth the screen room.

If fiber optics are used for communication through a wall, the steel support for the fiber should be removed for the last 6 feet. This one conductor can violate the room. The entry point for the fiber should be through a long tube that provides waveguide attenuation. To provide a waveguide attenuation the tube can run parallel to the wall for 5 or 6 inches.

The entry point for air should be a honeycomb structure. The air vent should be an insulator for the last 6 feet. This eliminates another ground connection and it removes the chance that lightning will use this path to get to the screen room.

N.B.

The lighting in a screen room must be from filament-type lamps. Fluorescent lamps are electrically too noisy. Cell phones should be turned off.

Conduit carrying power should not be routed along the outside wall of the screen room. This invites near-field penetration into the screen room. This rule includes the primary power entry. The power should enter the screen room perpendicular to the wall.

The power filters used in a screen room can be quite large and expensive. They can be specified to filter energy in one or both directions. Because they must provide filtering over a wide spectrum, they are usually made using several sections that are electrically isolated from each other.

The *equipment grounding conductor* cannot be filtered. It should be terminated inside the filter so that interference currents carried on the *equipment ground* stay on the right side of the filter boundary.

Equipment should not be operated near the walls of the room. Distance raises the wave impedance and allows the screen room to provide protection against field penetration.

N.B.

If power-related fields are not of concern, then the screen room need not have thick walls of steel.

8.10. MOTOR CONTROLLERS

Motor controllers are circuits that vary the power supplied to a motor to control its speed or torque. These controllers work by connecting the motor to the power line for a fraction of each power cycle. The basic control element is the *triac*. A triac is made from parallel oppositely directed signal-controlled rectifiers (SCRs). Each SCR is made from four layers of doped silicon (PNPN). If the inner P layer or gate is connected to the outer N layer, the device functions like two back-to-back diodes. In other words, there is no conduction in either current direction. If the inner P layer or gate is more positive than the outer N layer, the device functions as a normal diode. For the triac, one gate can control the forward conduction of both diodes in midcycle. When the diodes are turned on for the entire cycle the load is fully connected to the power line.

When the triac is placed in series with the power line it can be used to connect the power to a load for a fraction of the power cycle. Once the triac is turned on, the gate loses control of the triac. The gate cannot regain control until the current in the triac returns to zero. After turn-on, the triac current must reach a minimum hold value or the triac will not stay turned on. If the current in the triac is zero, the gate can then regain control after a specified time. The gate is usually controlled by an optical transistor to isolate the control circuit from the utility line.

Controllers are units that include triacs, gate drivers, and line filters. Units often contain circuitry that limits line transients and provides filtering for the gate signal. When the triac is turned on in midcycle a step voltage is supplied to the load. This steep leading edge demands energy from the distribution system and transmits a step voltage to the load. If these high-speed processes are not properly controlled, the result can be interference that invades an entire facility. This is the problem addressed in this section.

When a triac is turned on, the rise time can be as short as a microsecond. At the peak voltage of a 120-V line this is 170 volts per microsecond. This rise time characterizes the frequency of interest as 318 kHz. From an interference viewpoint there is harmonic content well beyond 10 MHz and this high-frequency content must be considered.

The interference process can be simplified by viewing the movement of field energy the moment after the triac switches on. The interference takes place in the next few microseconds and during this short time period the line voltage is essentially constant. The power connections are transmission lines and the events are similar to what occurs on a digital logic board. The energy that is sent forward must come from the electric field energy that is stored between the power conductors in the facility. To keep this request from propagating back into the facility it is a good idea to supply a local decoupling capacitor. This capacitor must be placed between the *ungrounded* and *grounded* power conductors near the triac. Small inductors placed in the power leads ahead of this decoupling capacitor can limit interference field energy from propagating back into the line. These inductors should not saturate at full line current and their natural frequency must be well above 10 MHz. It is also good practice to decouple the *grounded* conductor at this same point. This means placing a capacitor between the *grounded conductor* and the green wire and conduit.

The controller problem is similar to the issues discussed for a power line filters. If the filter geometry is not correct, fields will couple around the filter components and enter the hardware, thus violating the filter. If the filter enclosure is not bonded to an enclosure, fields will cross the boundary through any unclosed apertures. The same problem exists with a controller. The changing field in the controller can be carried out of the controller on any lead and this includes *equipment ground*. Fields can couple around the controller filter components if the input and output leads share the same conduit or terminating box. Fields can radiate directly out of the controller if they are not confined to an enclosure.

The decoupling capacitors discussed earlier do not control the steep voltage wave that is transmitted toward the motor. When this steep wave reaches the motor cage, energy will radiate through the cage openings into the facility and cause current flow on all conductors, grounded or not. To limit this radiation the wave sent to the motor should have a controlled rise time of about 100 microseconds. This again requires an LC filter on the power output leads of the controller. The inductors must not saturate at maximum line current and their natural frequency should be above 10 MHz.

The initial current through the triac must be high enough to guarantee that the triac fires and stays on at all voltage levels. The load filter just discussed may limit this initializing current. To guarantee a sufficient leading edge current a shunt input resistor or capacitor may be required ahead of the load side filter.

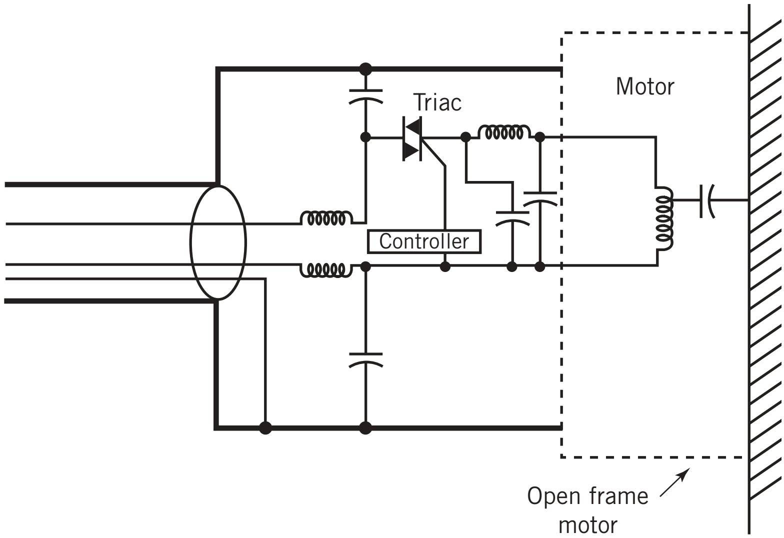


Figure 8.3. A simplified controller diagram

> **N.B.**
>
> Signal or control leads that enter the controller can carry field energy in or out of the controller. These leads are best filtered at the enclosure boundary to limit field crossing this interface.

The current supplied to a motor is not in phase with the voltage. When the current reaches zero the triac disconnects the power line from the motor. The voltage across the motor is not zero at this moment and this means that there is electric field energy stored in the parasitic capacitance of the motor. This energy will ring down in the parallel inductance of the motor. Normally this transient will be characterized at a frequency where radiation is not a problem.

A simplified diagram of a controller with the basic filtering elements is shown in Figure 8.3.

8.11. A TRUE STORY

A facility was designed to test motors in a low-temperature liquid-fuel environment. On a regular basis the fuel and the motor under test were removed from the test tank. Wiring that normally powered the motors was left coiled up outside of the tank with the load end disconnected inside the tank. A lifting crane was being used to move equipment near the tank. The crane was not grounded and it developed a static charge from a dry wind. As the crane neared the tank it discharged into the coiled wiring. The discharge path went from the wiring into the tank, exploding the vapors in the empty tank. The resulting explosion did a lot of damage.

RULE

Near any fueling point all conducting materials should be connected together before any vapor is allowed to escape. If possible, these conductors also should be earthed.

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