EMC Course Notes 2024

Shielding

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Virtually all high-speed electronic devices employ shielding in some form. Computers, cell phones, video games, industrial controls, automotive and avionics systems, etc., all typically come packaged in metal (or metalized) enclosures or have shields located directly over specific components on their printed circuit boards.

Shielded enclosures that are properly designed and installed can be a very effective means of attenuating radiated emissions and protecting products from external sources of interference. In fact, a metallic enclosure with no apertures, seams or cable penetrations can reduce radiated emissions and improve radiated immunity by 40 dB or more.

However, shielded enclosures are a poor substitute for good EMC design at the board level. A single breach of the enclosure (e.g., an unfiltered cable penetration) can eliminate any benefit the enclosure would otherwise provide. In fact, a product in a poorly designed shielded enclosure can radiate more (or be more susceptible) than the same product without the enclosure.

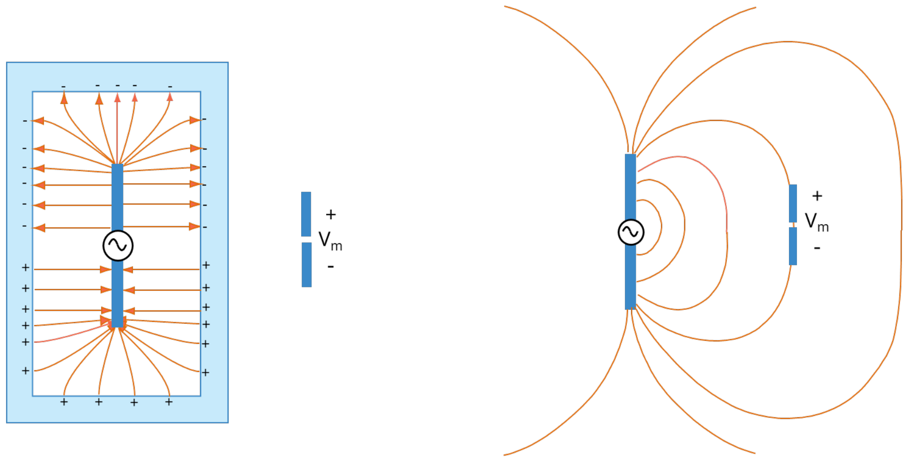
Shielded enclosures generally surround a product's circuitry on all sides. Care is taken to ensure that all apertures are small, and every seam is sealed. On the other hand, it is not uncommon to find shields that do not surround the entire product. Often a shield partially covers only a few circuits. These shields may be penetrated by unfiltered wires and sometimes consist of a single plate of metal that may or may not be connected to ground.

Why are apertures and seams so important in some applications and completely irrelevant in others? The answer relates to the fact that there are various kinds of shielding for different applications. It is convenient to divide enclosure or component shields into three categories: electric-field shields, magnetic-field shields, and shielded enclosures. The best shielding strategy in any given application depends on several factors including the electrical characteristics of the circuit or system being shielded, physical constraints (e.g., size, weight, and accessibility) and cost.

Electric-Field Shields

A perfectly conducting enclosure that completely surrounds a given volume prevents anything within that volume from electrically coupling to anything outside that volume. This type of enclosure is called a *Faraday cage*. Electric fields generated within the volume either terminate on objects within the enclosure or on the inner surface of the enclosure wall, as illustrated in Figure 10.1(a). Free charge on the enclosure relocates itself as needed to exactly cancel the fields within or external to the enclosure.

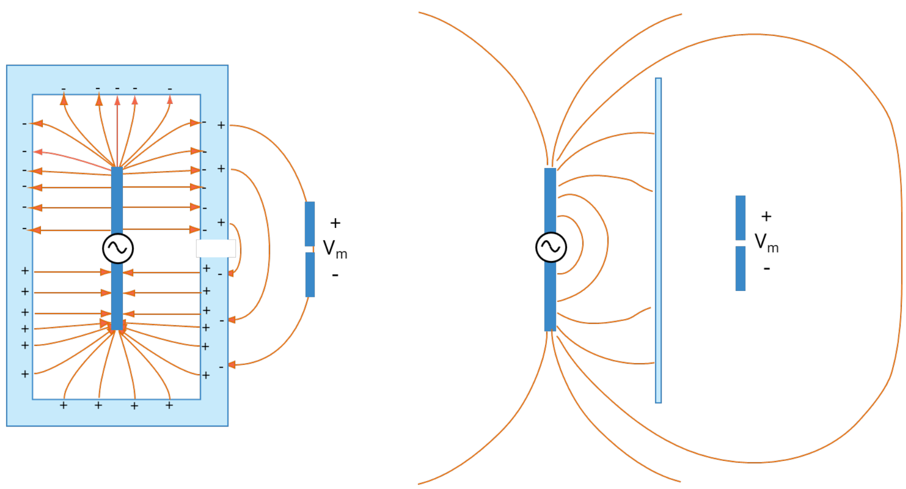
Enclosures that are not perfectly conducting are still good Faraday cages as long as the charges can redistribute themselves fast enough to cancel the internal fields. Most metallic enclosures without significant seams or apertures provide excellent electric-field shielding over a wide range of frequencies.



(a) (b)

Figure 10.1. Electric field coupling/shielding.

Without the shield in Figure 10.1(a) field lines may terminate on other conductors resulting in potential differences between these conductors as indicated in Figure 10.1(b). However, a partial shield [Figure 10.2(a)] or even a simple metal plate [Figure 10.2(b)] can substantially reduce these potentials by altering the path of the electric field lines and preventing the stronger field lines from reaching the victim circuit.



(a) (b)

Figure 10.2. Partial electric field shields.

Key concepts for practical electric field shielding are choosing a location that will intercept the stronger field lines and choosing a suitably conductive shield material. How conductive must the material be? That depends on the frequency or time rate-of-change of the fields. As long as charges can move freely enough to reorient themselves as fast as the field changes, cancellation of the external fields will be achieved.

For static electric fields, almost any material will look like a conductor since the free charge can slowly reposition itself. However, for high-frequency electric fields, the conductivity of the shield material must be high enough to allow the charge to move quickly back and forth.

Intercepting electric field lines with a conductive shield is primarily a matter of visualizing the field lines that are potentially responsible for the unwanted coupling and positioning the shield so that it blocks these fields. Figure 10.3 illustrates three ways that electric-field coupling from an IC to nearby traces could be reduced with shielding. The source IC can be shielded, the victim traces can be shielded, or the shield can intercept the field lines anywhere in between.

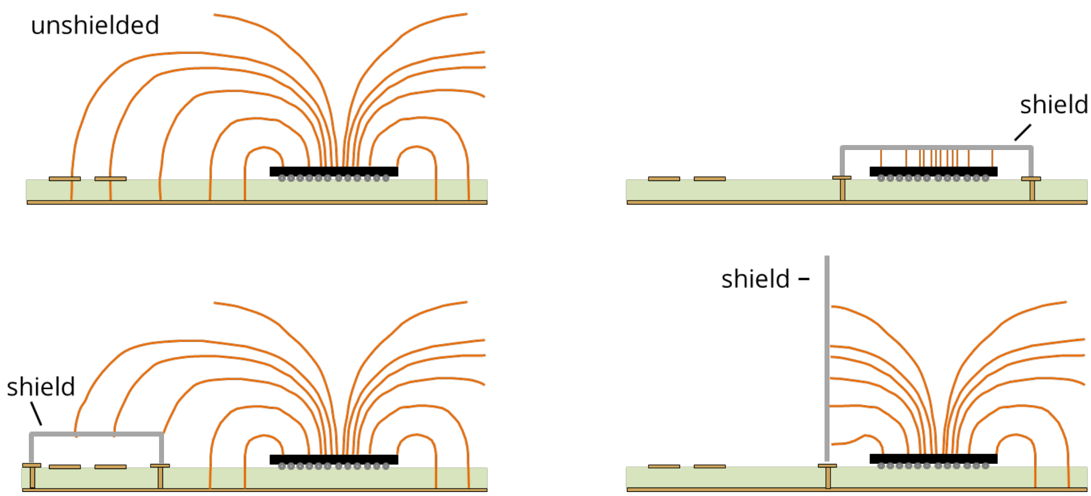


Figure 10.3. Electric field shielding examples.

Note that field lines terminating on a conductor imply there is a negative charge induced at that point. There is a positive charge at locations where field lines emanate from a conductor. If the field is time-varying, there will be a current on the surface of the conductor as these charges move back and forth.

Magnetic-Field Shields

Because there are no free magnetic charges, it is not possible to terminate lines of magnetic flux on a shield. However, it is possible to redirect magnetic flux lines to prevent unwanted coupling. This can be accomplished by electric currents induced in an electrically conductive shield or by altering the path of magnetic flux lines using permeable (μ_r>>1) materials.

Consider the configuration shown in Figure 10.4(a). A vertical magnetic field due to kHz switching currents in an electric motor couples to a small circuit board resulting in interference. Figure 10.4(b) shows the same configuration with a copper plate below the circuit. For any closed-loop path on the surface of the plate that is penetrated by the incident magnetic field, Faraday's law tells us that a tangential electric field must exist on the surface such that,

$$\oint \vec{E} \cdot \vec{d\ell} = \frac{\partial \Psi}{\partial t} \tag{10.1}$$

where the right-hand side of this equation is the time rate of change of the total magnetic flux coupling the loop. However, any tangential electric field on the surface of a good conductor will cause currents to flow in that conductor. These currents will generate their own magnetic flux that opposes the change in the incident flux. In a perfect conductor, the flux generated by these currents would perfectly cancel the incident flux causing both sides of (10.1) to equal zero (i.e., no flux penetrating the conductor and no tangential electric field on the surface).

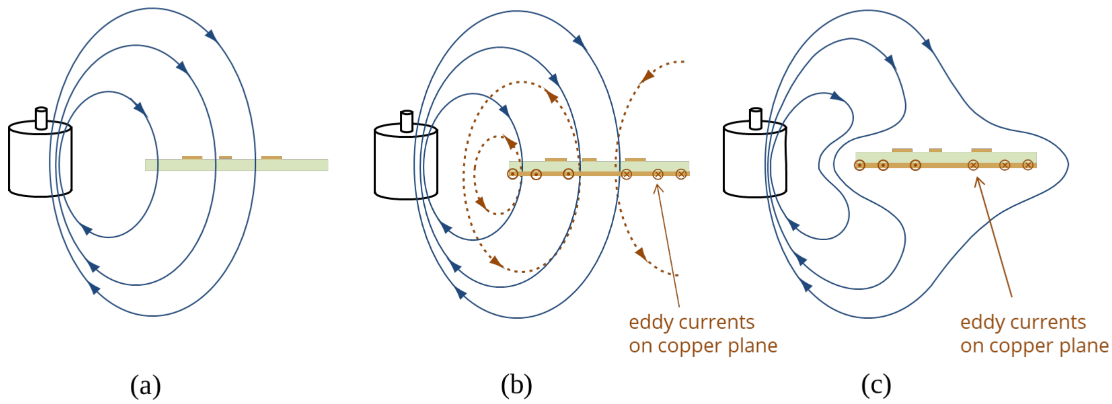


Figure 10.4. Magnetic field shielding with good conductors.

Currents induced in a conducting material by a time-varying magnetic field in this manner are called *eddy currents*. Both the incident field and the magnetic field created by the eddy currents are shown in Figure 10.4(b). The sum of both fields is shown in Figure 10.4(c). Note that the eddy currents cause the magnetic flux to be diverted around the plate and significantly reduce the coupling to the circuit.

To divert a magnetic field with a conductive plate, it is important to be able to develop sustained eddy currents. Since the eddy currents are driven by time-varying fields, a conductive plate cannot divert a static magnetic field. If the field varies too slowly, losses in the conducting plate will cause the eddy currents to dissipate allowing the magnetic flux to penetrate the plate. For this reason, thin conductive materials are generally poor magnetic shields at low frequencies (e.g., below a few hundred kHz). Conductive magnetic shields are also ineffective if they have slots or gaps that interrupt the flow of the eddy currents.

At kHz frequencies or lower, it is necessary to use permeable (magnetic) materials (μ_r>>1) to divert magnetic fields. These materials have a reluctance much less than air, so magnetic field lines can effectively be rerouted by providing an alternative path around the object to be shielded. Figure 10.5 illustrates how a shield made from a permeable material can be used to protect the circuit in the previous example.

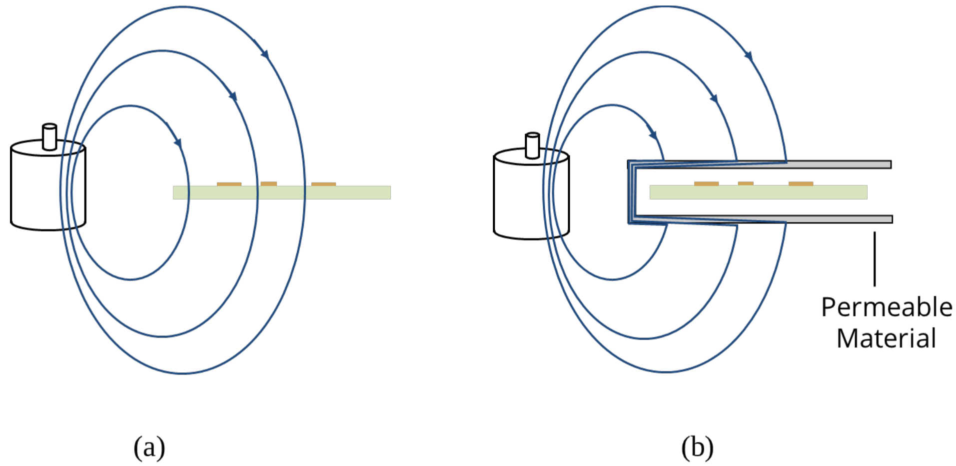


Figure 10.5. Magnetic field shielding with magnetic materials.

Note that it is important for the magnetic material shield to divert the magnetic flux all the way around the object being shielded. A plate of magnetic material above or below the circuit board would provide no shielding at all.

The reluctance of a magnetic flux path is proportional to the length and inversely proportional to the relative permeability times the cross-sectional area. Therefore, materials with higher permeabilities can capture more of the flux and divert it for greater distances. Table 10.1 lists the relative permeabilities of various materials. Materials with high permeabilities generally contain the one of the elements iron, nickel or cobalt.

There is a limit to the amount of magnetic flux that can be diverted by a given material in a particular shielding geometry. When this limit has been exceeded, the material is said to be *saturated*, and the relative permeability drops back to 1. Materials with higher permeabilities saturate more quickly for a given applied field and geometry. Thicker materials can redirect more flux before saturating. When designing a low-frequency magnetic field shield, it's important to consider the maximum strength of the applied magnetic field in order to choose a material and geometry that won't saturate.

Table 10.1. The Relative Permeability of Various Materials

Material	Relative Permeability
Gold, Copper, Aluminum	1
Concrete, Water, Air, Vacuum	1
Austenitic Stainless Steel	1
Ferrite U60 (UHF Chokes)	8
Common Steel	10 - 100
Pure Nickel	600
Ferrite M33 (inductors)	750
Ferritic Stainless Steel (annealed)	1000 - 1800
Pure Iron	5,000
Permalloy (20% iron, 80% nickel)	8,000
Ferrite T38 (RF Transformers)	10,000
Mu-metal	20,000 – 50,000
Supermalloy (recording heads)	100,000

Shielding Enclosures

Many electronic components and systems utilize shielding enclosures. An ideal shielding enclosure with infinite conductivity and no breaks would perfectly isolate (electromagnetically) whatever was inside the enclosure from whatever was outside. Shielding enclosures can help to protect circuits from harsh electromagnetic environments outside the enclosure. They can also contain the fields generated by circuits inside the enclosure, preventing them from interfering with circuits and systems outside the enclosure.

There are various methods for characterizing the effectiveness of a shielding enclosure. Generally, these involve measuring a field quantity at a location outside the enclosure with and without the enclosure in place. The measured shielding effectiveness is then given by,

$$\mathrm{SE} = 20\log\left(\frac{\text{Field measured without the enclosure}}{\text{Field measured with the enclosure}}\right). \tag{10.2}$$

Two well-known standards for measuring the shielding effectiveness of an enclosure are IEEE 299 for large enclosures and IEEE 299.1 for smaller enclosures. These measurements can provide useful information related to the effectiveness of a shielding enclosure, but they do not provide a measure of how well the enclosure will shield any given circuit or system. The geometry, impedance and location of circuits have a significant impact on the magnitude and direction of currents induced on the inside surface of a shielding enclosure.

These currents, as well as the location of the external field measurement, and the position of any external cables or components can have a profound impact on the measured shielding effectiveness.

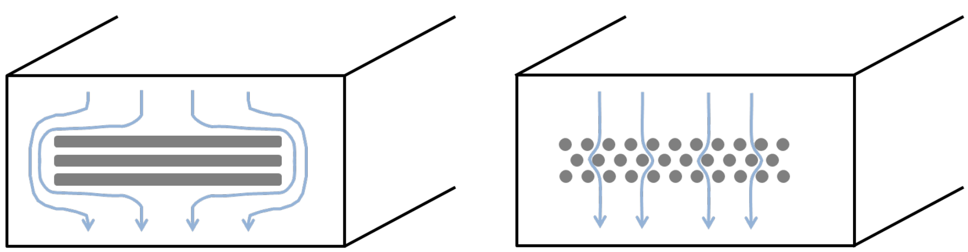
If the material used to construct the shielding enclosure is metal (e.g., copper, aluminum, or steel), virtually no field passes through the wall of the enclosure. Instead, the shielding effectiveness is limited by the power that leaks through the enclosure's seams, apertures, and cable penetrations. When it's necessary for an enclosure to provide a high shielding effectiveness over a wide range of frequencies, it's important to carefully evaluate every seam, every aperture, and every cable penetration to ensure that no significant interfering signals can pass from one side to the other.

Apertures

Apertures are holes in a shielded enclosure such as those required for ventilation, optical displays, plastic components, or mechanical supports. For the enclosure to provide shielding, currents must be able to flow on the surface unimpeded. Fortunately, apertures with maximum dimensions that are much smaller than a wavelength (e.g., < $\lambda/20$) provide very little impedance to the flow of currents on a conducting surface. For this reason, if it is necessary to provide a certain amount of open area (e.g., for air flow), it is much better to accomplish this with many small apertures than with a few large apertures.

Figure 10.6 illustrates the path of currents flowing around two ventilation grids. Note how the grid in Figure 10.6(a) interrupts the flow of current much more significantly than the grid in Figure 10.6(b). In terms of electromagnetic shielding, the grid in Figure 10.6(b) is far superior even though the total open area of both patterns is similar. Note that a shielding enclosure can be very effective even when it has a significant amount of open area as long as each individual aperture is much smaller than a wavelength.

Generally, the amount of energy escaping an enclosure through small apertures is insignificant compared to the energy escaping through seams, larger openings, and wire penetrations. However, if the enclosure is well sealed and it is necessary to further reduce the energy escaping through the apertures, then apertures with sufficient depth can be provided to further attenuate the radiated emissions. Making an aperture extend further into the enclosure creates a small waveguide. For apertures with a small cross-section, the frequencies of the sources within the enclosure are likely to be well below the cut-off frequency of the waveguide. Information on designing apertures to be waveguides below cut-off is provided in a later section.



(a) (b)

Figure 10.6. Two aperture patterns in a shielded enclosure.

Seams

Seams exist wherever two pieces of an enclosure come together. Seams are often a more significant source of leakage than apertures because of their greater length. A seam that is on the order of a half-wavelength long can be a very efficient radiation source, much like a resonant half-wave dipole. It is possible to make inefficient antennas (such as an electrically small wire or loop antenna) radiate much more efficiently by enclosing them in a metal enclosure with a resonant slot or seam.

A seam that optically appears to be well sealed can often disrupt the flow of surface currents significantly, causing a major breach in the shielding enclosure. For example, two metal surfaces simply pressed against each other as illustrated in Figure 10.7(a) or 10.7(b) rarely provide sufficiently reliable contact at high frequencies. Surface oxidation, corrosion and warping on the metal plates reduce the quality of the electrical contact. Screws or rivets [Figure 10.7(c)] can provide a good electrical contact at points, but they do not necessarily improve the connection at locations between fasteners. One technique for reducing the impedance of seams is to overlap both sides of the plates as illustrated in Figure 10.7(d). While this interlocking connection is often better than simple overlap, there will still be areas where the contact is poor. More reliable solutions employ finger stock or gaskets as shown in Figures 10.7(e) and 10.7(f).

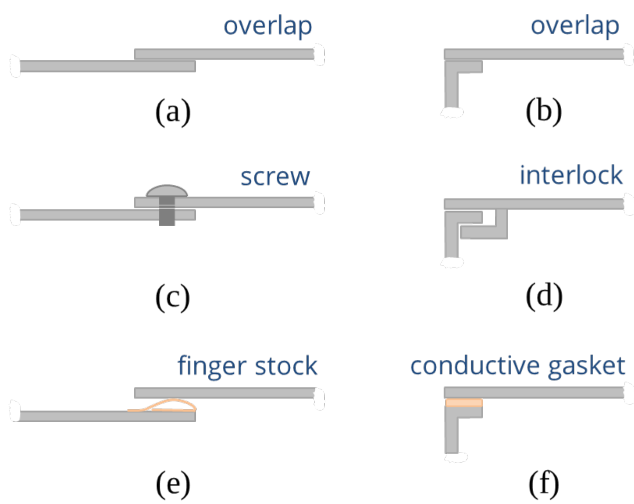


Figure 10.7. Seams in shielded enclosures.

Cable Penetrations

To power and/or communicate with the electronics in a shielded enclosure, it is often necessary to employ wires that pass through the enclosure wall. A single unshielded, unfiltered wire penetrating an enclosure can eliminate any shielding benefit that the enclosure otherwise provided. As illustrated in Figure 10.8, any difference between the voltage on a wire and the voltage on the enclosure drives the wire/enclosure pair like a dipole antenna. Since the wires and the enclosure tend to be among the larger metallic objects in a system, the wire/enclosure pair is often a very efficient antenna at relatively low frequencies. For this reason, it is important to ensure that any wires penetrating the enclosure are either well-shielded or held to the same potential as the enclosure at all frequencies that may be a radiation problem.

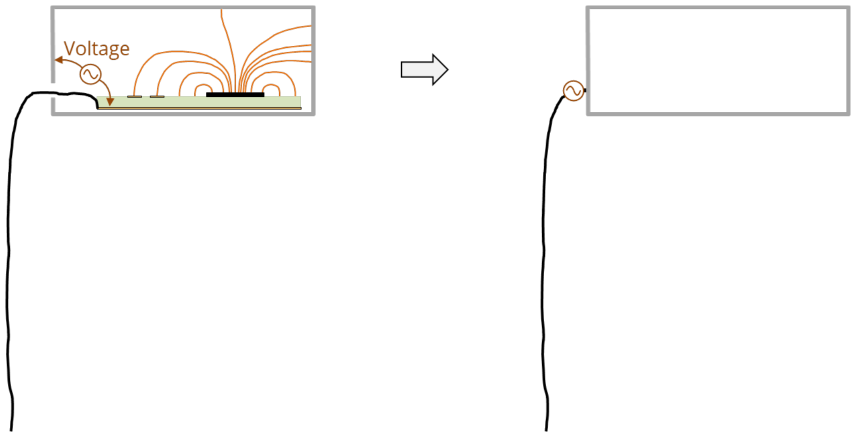


Figure 10.8. Wire driven relative to shielded enclosure.

For the shield on a shielded wire to be effective, it must make a low-inductance connection to the shielded enclosure. This is generally accomplished by using a shielded connector that makes a 360-degree metal-to-metal contact with both the cable shield and the enclosure, as illustrated in Figure 10.9(a).

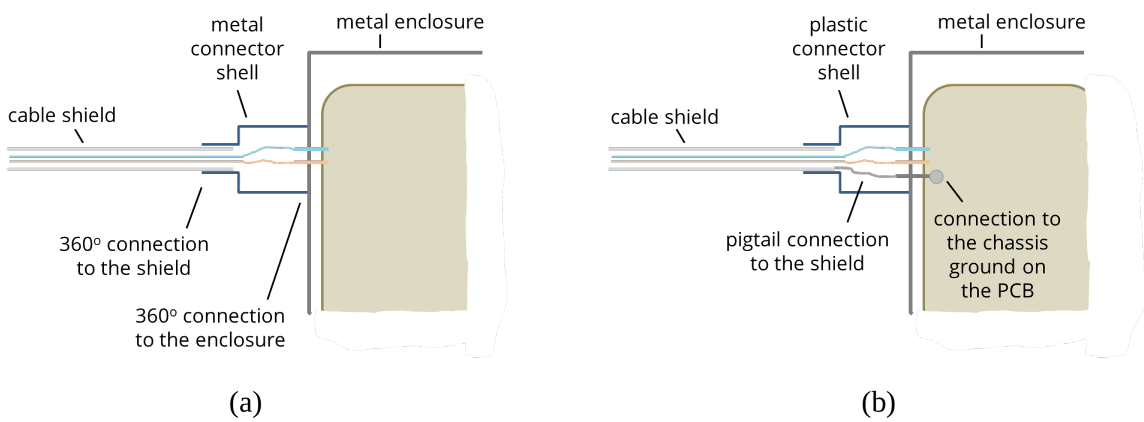


Figure 10.9. Cable shield to enclosure connections.

A *pigtail* connection as shown in Figure 10.9(b) will have significant inductance. It also requires the circuit board to have a chassis ground to facilitate the connection between a connector pin and the enclosure. The inductance of the connection between the cable shield and the enclosure allows a voltage to form between the cable shield and the enclosure. Because of this, pigtail connections are rarely acceptable when the shielding must be effective above a few hundred kHz.

If the wires penetrating a shielded enclosure are not shielded, they must be filtered. Filtering will minimize the voltage between the wire and enclosure at radiation frequencies, while allowing low-frequency signals or power to pass unattenuated. It is usually necessary to locate the filter as close to the connector location as possible to minimize the inductance of the connections and to prevent the possibility of noise coupling to the filtered wire before it exits the enclosure. Examples of filter locations are illustrated in Figure 10.10.

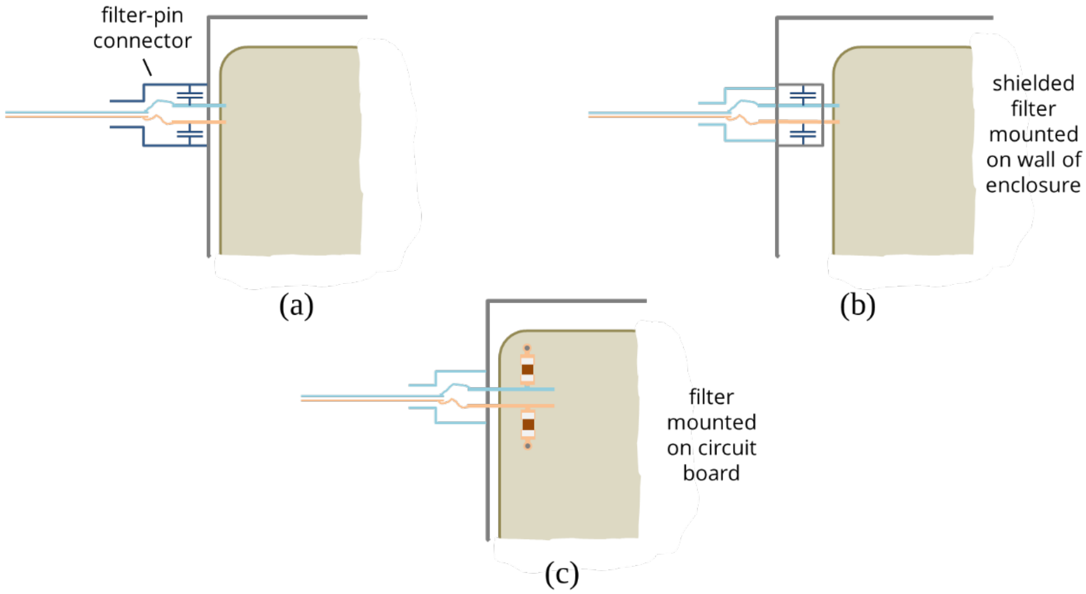


Figure 10.10. Possible cable-filter configurations.

Attenuation Due to Waveguides Below Cut-off

Sometimes it is necessary to have many apertures in a shield for the purposes of ventilation. In large enclosures with very stringent shielding and thermal requirements, it may be necessary to further reduce the amount of energy that can escape through any array of apertures. This can be accomplished by increasing the depth of the apertures so that they resemble small waveguides. At frequencies where the cross-sectional dimensions of the apertures are small relative to a half-wavelength, energy propagating through the apertures will be attenuated in the same manner that energy propagating through a waveguide below the cut-off frequency is attenuated.

Energy will not propagate in a waveguide at frequencies below the cut-off frequency. Instead, the fields decay exponentially. A simple approximate formula for the attenuation provided by an opening with depth, *d*, and maximum height or width, *a*, is:

$$\text{attenuation} = 30\,\frac{d}{a}\sqrt{1-\left(f/f_c\right)^2}\quad \text{dB} \tag{10.3}$$

where f is the frequency of the field and f_c is the cut-off frequency of the opening. The cut-off frequency is approximately the frequency at which the maximum height or width, *a*, is equal to a half-wavelength.

Attenuation in a Rectangular Aperture

For a rectangular aperture with height *a*, width *b*, and length *d*, as illustrated in Figure 10.11, the mode of propagation with the lowest cut-off frequency is the TE_{10} mode.

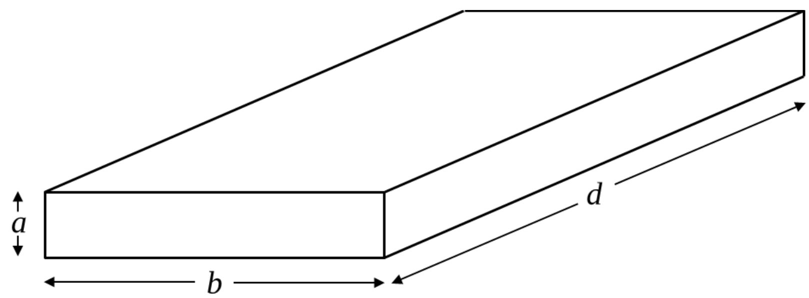


Figure 10.11. Rectangular aperture geometry.

The propagation constant for the TE_{10} mode is given by,

$$\beta = \sqrt{\left(\frac{2\pi}{\lambda}\right)^2 - \left(\frac{\pi}{b}\right)^2} \,. \tag{10.4}$$

At frequencies where the term under the radical is negative, the propagation constant is imaginary, and fields do not propagate. This occurs when $\lambda > 2b$. The cut-off wavelength for the TE_{10} mode is therefore, $\lambda_c = 2b$. The cut-off frequency is,

$$f_c = \frac{v}{\lambda_c} = \frac{v}{2b} \tag{10.5}$$

where v is the velocity of propagation in the aperture dielectric (3×10^8 m/s in air).

Below the cutoff frequency, the magnitude of the field in the aperture decays exponentially,

$$E(z) = E_o e^{-|\beta| z} \,. \tag{10.6}$$

The total attenuation of the field traveling a distance, d, expressed in dB is then,

$$\text{attenuation in dB} = 20 \log_{10} e^{-|\beta| d} = 8.7 |\beta| d \tag{10.7}$$

or, combining Equations (10.4), (10.5) and (10.7),

$$\text{attenuation in dB} \approx 27 \, \frac{d}{b} \, \sqrt{1 - \left(f / f_c \right)^2} \,. \tag{10.8}$$

Attenuation in a Circular Aperture

For a circular aperture with diameter a, and length d, as illustrated in Figure 10.12, the mode of propagation with the lowest cut-off frequency is the TE_{11} mode.

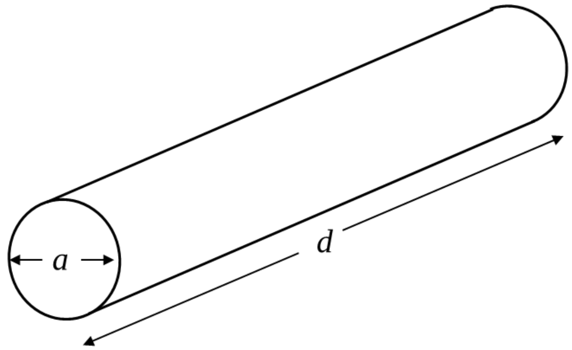


Figure 10.12. Circular aperture geometry.

The propagation constant is given by,

$$\beta = \sqrt{\left(\frac{2\pi}{\lambda}\right)^2 - k_c^{\,2}} = k_c\sqrt{1-\left(k/k_c\right)^2} = k_c\sqrt{1-\left(f/f_c\right)^2} \tag{10.9}$$

where k_c for the TE_{11} mode is,

$$k_c = \frac{3.682}{a} \ . \tag{10.10}$$

Setting the term under the radical in (10.9) to zero, the cut-off frequency is shown to be,

$$f_c = \frac{0.586\, v}{a} \tag{10.11}$$

where v is the velocity of propagation in the aperture dielectric ($3\text{x}10^8$ m/s in air).

Below the cutoff frequency, the magnitude of the field in the waveguide decays exponentially,

$$E(z) = E_o e^{-|\beta| z} \ . \tag{10.12}$$

The total attenuation of the field traveling a distance, d, expressed in dB is then,

$$\text{attentuation in dB} = 20\log_{10} e^{-|\beta| d} = 8.7|\beta| d \tag{10.13}$$

or, combining Equations (10.9), (10.11) and (10.13),

$$\text{attenuation in dB} \approx 32\frac{d}{a}\sqrt{1-\left(f/f_c\right)^2} \ . \tag{10.14}$$

Assumptions and Notes

The derivations based on both rectangular and circular waveguides have a constant in front that is within 3 dB of 30. Other modes of propagation yield different constants, but the lower-order modes dominate so it is reasonable to use a value of 30 for most open aperture cross-sections.

Note that the expression approaches 0 dB as the thickness d of the opening approaches 0. However, even thin apertures will provide some attenuation if their cross-sections are small relative to a wavelength. The approximate expression in (10.3) is not accurate unless $d>>a$.

This model does not account for how the field was set up at one end of the opening or how efficiently it is radiated from the other end. Therefore, by itself, it cannot be used to determine the shielding effectiveness of any particular shield. The attenuation calculated in (10.3) should be added to the shielding effectiveness that would be obtained from the same aperture configuration in a thin shield.

Note that if a wire or second conductor of any kind penetrates the opening, the lowest-order mode of propagation is the TEM mode. Fields at any frequency can penetrate the opening in the TEM mode, so there is no benefit to using a thick aperture if a wire penetrates the aperture.

Some textbooks state that the attenuation in (10.3) should be reduced by a factor of $10 * \log_{10}$(# of apertures) if there are multiple apertures. However, the attenuation in Equation (10.3) is *in addition to* any attenuation provided by a thin aperture or an array of thin apertures. Equation (10.3) cannot be used to calculate the shielding effectiveness directly; and there is no reason to reduce the Equation (10.3) value for multiple thick apertures if you are starting from the attenuation provided by multiple thin apertures.

Cable Shielding

Properly designed and terminated cable shields can significantly reduce field coupling to and from a wire harness. If a harness is being driven by a common-mode voltage, a shield can return common-mode currents to the source without generating radiated emissions. A shield can also block non-radiating fields emanating from a harness or prevent strong external fields from coupling to a harness.

The effectiveness of a cable shield strongly depends on how it is terminated at each end, and the correct shield termination depends on the purpose of the shield. Figure 10.13 illustrates three shield termination options for a cable that is electrically short. A short cable is unlikely to be a significant source of radiated emissions, but shielding may be required to prevent field coupling to or from the cable.

If the cable shield is unconnected at both ends, it provides no attenuation of the field coupling. An incident electric field will terminate on the shield rather than on the wires in the cable, but the wires in the cable are strongly coupled to the shield and will take on the same voltage as the cable shield. A magnetic field that couples to the loop formed by the cable, cable source, cable termination, and ground is unaffected by the presence of an unterminated cable shield. While the presence of an unterminated shield doesn't make the coupling any worse, it doesn't make it any better either.

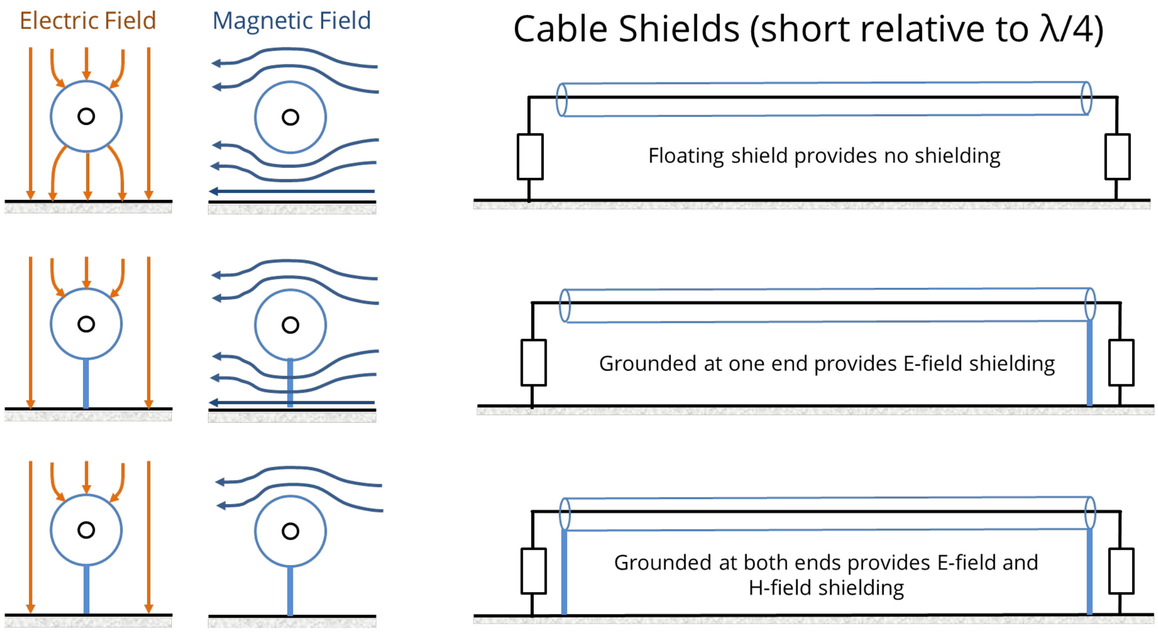


Figure 10.13. Three shield termination options for an electrically short cable.

If the cable shield is connected to the EMC ground structure[1] at either end, it significantly reduces electric-field coupling to or from the cable. Field lines terminate on the shield, but the shield is held to the same potential as the EMC ground. A time-varying field induces currents on the exterior of the shield that flow through the shield's connection to ground. As long as the shield voltage is held constant, no electric-field coupling occurs to the wires inside the shield.

Magnetic field coupling, on the other hand, is not affected by grounding the shield at one end. Magnetic field lines are still able to couple the cable loop, just as they did when the shield was missing or unterminated.

Connecting the cable shield to the EMC ground structure at both ends reduces both electric- and magnetic-field coupling. The additional ground connection allows currents to circulate in the shield-ground loop. This produces a magnetic field that opposes any incident magnetic field. Generally, it is a good idea to bond both ends of a cable shield to the EMC ground structure. However, in situations where only electric-field shielding is required (e.g., a harness passing through a noisy high-voltage environment), grounding at one end is often sufficient.

When a cable is not electrically short, grounding the shield at one end does not prevent electric-field coupling. Holding the potential of the shield to the ground potential in one place doesn't necessarily reduce the electric-field coupling to the shield at other places. The impedance from the shield to ground becomes a function of position and can be very high in places one-quarter wavelength or more away from the grounding point.

On the other hand, grounding an electrically long cable shield at both ends as illustrated in Figure 10.14, provides excellent protection from electric and magnetic field coupling. Currents induced on the shield by external noise sources are confined to the external

[1] The term *EMC ground structure* is defined in Chapter 8.

surface of the shield. There is no longer any inherent mechanism for coupling shield currents to the wires within the shield. Cable shields grounded at both ends tend to be very effective. If any significant coupling occurs, it is generally due to the impedance of an imperfect bond between the cable shield and the ground structure at one or both ends.

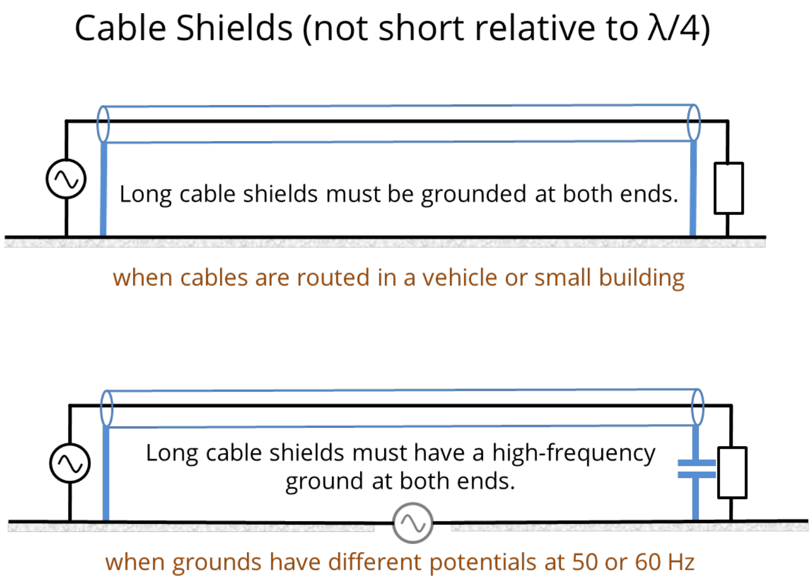


Figure 10.14. Two shield termination options for an electrically long cable.

Occasionally, safety concerns prevent a long cable shield from being grounded directly at both ends. A common example of this would be a cable that extends between buildings with separate electrical services. In this situation, there may be a low-frequency voltage difference between the two ground structures. A few volts of potential difference across a cable shield with 1-Ω of end-to-end resistance would put a few amperes of current on the shield. And while amperes or even tens of amperes of current on a cable shield is not necessarily problematic, it could be dangerous if the cable were suddenly disconnected. This would produce an L dI/dt voltage across the disconnection point. For long cables forming a high-inductance loop, the voltage generated could be high enough to create a spark or to shock the person disconnecting the cable. The energy discharged (proportional to LI^2) could be great enough to cause serious harm.

For this reason, it is common to terminate very long cables with a capacitor to ground at one end, leaving the other end bonded directly to ground. The capacitor blocks the flow of low-frequency power currents, while providing a high-frequency connection.

Where Should a Cable Shield be Connected?

One source of confusion among system designers is how and where to connect the cable shield. A good rule to follow is that a cable shield always connects to the EMC ground structure in a manner that minimizes the connection impedance. Noise currents on the external surface of a cable shield need to be routed to the chassis ground, or to the circuit board's EMC ground.

Cable shields can also carry currents on the internal surface. These could be intentional signal currents (e.g., in the case of coaxial cables), or they could be noise currents generated by common-mode voltages that drive the wires in the shield relative to circuit ground. When signal currents or significant noise currents are flowing on the internal surface of a cable shield, these must be returned to the circuit board's return plane. In this case, it's important to ground the shield to both the chassis and the circuit board. In well laid-out circuit boards, the location of any cable shield connection to the board is always the PCB EMC ground as described in Chapter 8.

Figure 10.15 shows the location of the PCB EMC ground in boards with and without a chassis connection. In each case, the shield is connected to the place where we want to route any noise currents flowing on the external surface of the cable shield. If the board is mounted in a plastic enclosure, the safest place to route noise currents on the shield is to the circuit board's ground (RTN) plane. This is much better than leaving the shield unconnected, which would allow noise currents to come in on the signal wires. If the board is mounted on a metal chassis, the shield currents should be routed to the chassis without coming too far onto the board.

If the connector shell is mounted directly to the metal chassis, as indicated in the far-right illustration in Figure 10.15, there may be no need to connect the cable shield to the board ground. However, if the shield is returning signal currents or common-mode currents induced on the inside of the shield, then both a circuit ground and a chassis ground are required.

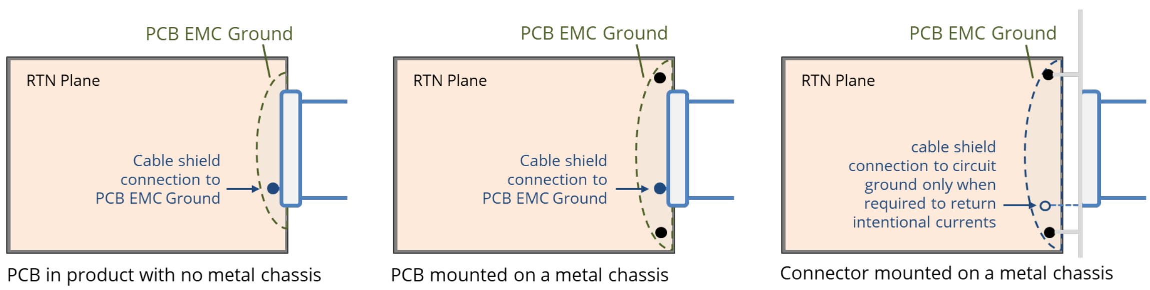


Figure 10.15. Cable shield connections in three different circumstances.

There are many types of products with different grounding strategies, different electromagnetic environments, and different reasons for employing shielded cables. There is no concise rule for cable shield terminations that covers all situations. Cable shields connected to the chassis route incoming noise currents to the chassis. A chassis connection also prevents the cable shield from being driven relative to the chassis. Both are normally highly desirable outcomes, so cable shields are almost always connected to the metal chassis when there is one.

Connections to the circuit board are desirable when currents flowing on the inside surface of a cable shield must be returned to their source. They can also be important when the product has no metal chassis. Since cable shields can make excellent antennas when driven relative to anything else of significant electrical length, any connection between a cable shield and a circuit board return plane should be made in the region identified as the board's EMC ground.

Transfer Impedance

The shielding effectiveness of a cable shield is sometimes quantified using a parameter called the *transfer impedance*. Transfer impedance measurements involve placing a known current on the cable shield and measuring the voltage induced in the signal path. An example of a transfer impedance measurement set-up is illustrated in Figure 10.16.

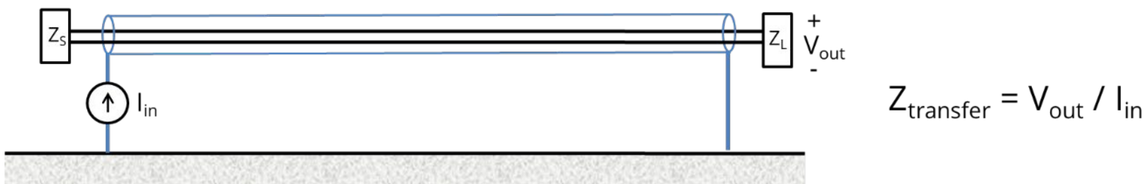


Figure 10.16. Transfer impedance measurement.

For any reasonable cable shield, the measured transfer impedance is very small. It tends to be proportional to the shield and connection resistance at low frequencies. At higher frequencies, connection inductances and imperfections in the shield integrity play a greater role.

In all but the most demanding applications, the published transfer impedance of a cable has little meaning. The actual shielding effectiveness obtained from a cable shield tends to be dominated by the manner and location in which the shield is terminated. The effects of the shield resistance and leakage through a cable shield can only be observed in systems employing shielded enclosures and 360°-shielded connectors. And even in those systems, one can't rely on published transfer impedance numbers, because these numbers often depend strongly on the specific test set-up used to acquire them.

One situation where transfer impedance measurements can be very helpful is when they are made in an independent test lab and used to evaluate shielded cable assemblies from different sources. The quality of shielded cable assemblies from different sources can vary greatly. Some employ full-coverage shields with 360° terminations, while others use pigtails or make no shield connection at all. It can be difficult to tell the difference between a good cable assembly and a poor one by inspection or by making DC measurements. On the other hand, transfer impedance measurements are an excellent way to quickly identify and eliminate assemblies that do not provide adequate shielding.

Plane Wave Shielding

Shields work by reflecting, absorbing, or redirecting electric and/or magnetic fields. They are always located in the near-field of the source or the victim circuit. There is virtually no practical situation where a plane wave is incident on a shield, passes through the shield, and continues to propagate on the other side of the shield. Nevertheless, that is the fundamental assumption of plane-wave shielding theory.

Plane-wave shielding was described mathematically by Schelkunoff[2] in the early 1940s. It cannot be used to evaluate the shielding effectiveness of actual shields or shielding enclosures, but it provides valuable insight related to shielding materials and the way they interact with electromagnetic fields. Plane wave shielding theory allows us to define a

[2] S. A. Schelkunoff, *Electromagnetic Waves*, Van Nordstrum, 1943.

plane wave shielding effectiveness for a slab of shielding material. This shielding effectiveness can be calculated directly from the material properties, and it can be measured by passing plane waves through the material in a lab.

Plane-Wave Shielding Theory

When an electromagnetic wave propagating in one material encounters another material with different electrical properties, some of the energy in the wave is reflected and the rest is transmitted into the new material. For example, consider the electromagnetic plane wave, $\vec{E}_{inc}$, incident upon a semi-infinite slab of material as illustrated in Figure 10.17. The wave propagates in free space in the x direction until it strikes the material, which has an intrinsic impedance, η_s.

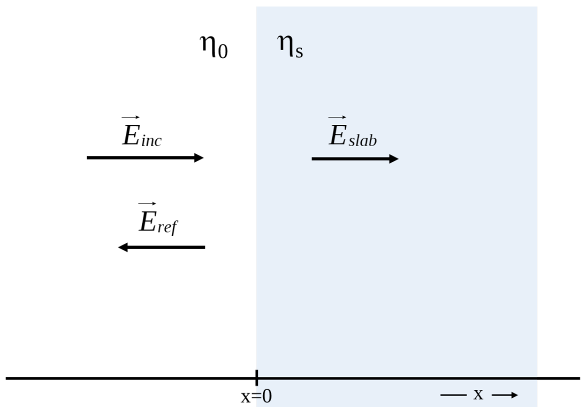


Figure 10.17. Plane wave incident on a shielding material.

The magnetic field in the plane wave is perpendicular to the electric field and has amplitude,

$$\left|\vec{H}_{\text{inc}}\right| = \frac{\left|\vec{E}_{\text{inc}}\right|}{\eta_0} \tag{10.15}$$

where $\eta_0 = \sqrt{\mu_0 / \varepsilon_0}$ is the intrinsic impedance of free space (~377 Ω).

When the plane wave strikes the slab, a reflected wave, $\vec{E}_{ref}$, and a transmitted wave, $\vec{E}_{slab}$, are created. The magnetic field in the shielding material is related to the electric field,

$$\left|\vec{H}_{\text{slab}}\right| = \frac{\left|\vec{E}_{\text{slab}}\right|}{\eta_s}. \tag{10.16}$$

In addition, the boundary conditions on the surface at x=0 require that,

$$\vec{E}_{x=0^-} = \vec{E}_{x=0^+} \tag{10.17}$$

and

$$\vec{H}_{x=0^-} = \vec{H}_{x=0^+} \tag{10.18}$$

where the subscripts x=0$^-$ and x=0$^+$ indicate the fields, just to the left and right of the x=0 surface, respectively. To satisfy Equations (10.15 - 10.18), the amplitude of the reflected field must satisfy the relation,

$$\left|\vec{E}_{\mathrm{ref}}\right| = \left|\vec{E}_{\mathrm{inc}}\right| \Gamma_{\mathrm{E}} \tag{10.19}$$

where Γ_{E} is the electric-field reflection coefficient,

$$\Gamma_{\mathrm{E}} = \frac{\eta_s - \eta_0}{\eta_s + \eta_0}. \tag{10.20}$$

The amplitude of the transmitted field, $\vec{E}_{slab}$, is

$$\left|\vec{E}_{\mathrm{slab}}\right| = \left|\vec{E}_{\mathrm{inc}}\right| T_{\mathrm{E}_1}, \tag{10.21}$$

where

$$T_{\mathrm{E}_1} = \frac{2\eta_s}{\eta_s + \eta_0} \tag{10.22}$$

is the electric-field transmission coefficient at the first interface. Note that as η_s gets closer to η_0, the transmission coefficient increases, and the reflection coefficient decreases. If $\eta_s = \eta_0$, all of the incident field is transmitted.

If the material in Figure 10.17 is lossy, (i.e., $\sigma \neq 0$), the transmitted wave will decrease in amplitude as it propagates,

$$\left|\vec{E}_{\mathrm{slab}}(\mathrm{x})\right| = \left|\vec{E}_{\mathrm{slab}}(\mathrm{x{=}0})\right| e^{-\mathrm{x}/\delta} \tag{10.23}$$

where δ is the *skin depth* of the material. For high-loss materials,

$$\delta \approx 1 \Big/ \sqrt{\pi f \mu \sigma}. \tag{10.24}$$

Now consider the slab of shielding material with finite thickness illustrated in Figure 10.18. An incident field, $\vec{E}_{inc}$, strikes the surface of the shielding material. Some of the power in the field is reflected and some continues into the material. The part that penetrates the material is attenuated before it strikes the second surface at x=t. At that point, once again some of the power is reflected and some of the power is transmitted. If the attenuation is high, the power reflected at the second interface is absorbed and the field transmitted to the region of free space on the right of the slab is given by,

$$\left|\vec{E}_{\mathrm{trans}}\right| = \left|\vec{E}_{\mathrm{slab}}(x = t)\right| T_{\mathrm{E}_2} \tag{10.25}$$

where

$$T_{\mathrm{E}_2} = \frac{2\eta_0}{\eta_0 + \eta_s}. \tag{10.26}$$

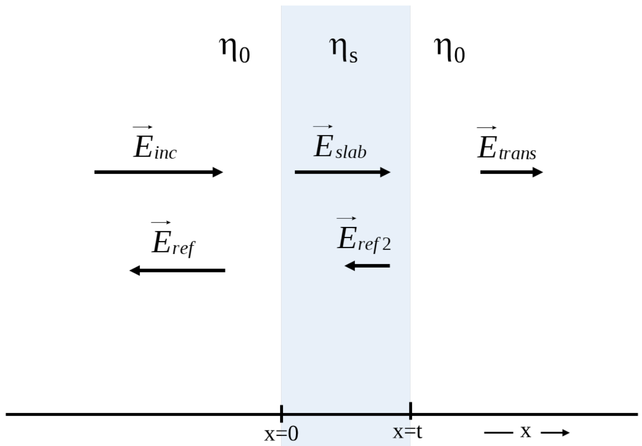


Figure 10.18. Plane wave incident on a finite thickness shielding material.

Combining (10.21), (10.22), (10.23), (10.25) and (10.26); we obtain an expression for the transmitted electric field in terms of the incident field,

$$\left|\vec{E}_{\text{trans}}\right| = \left|\vec{E}_{\text{inc}}\right| \frac{2\eta_s}{\eta_0+\eta_s}\left(\frac{2\eta_0}{\eta_0+\eta_s}\right) e^{-t/\delta}. \tag{10.27}$$

This expression applies to any shielding material that is much thicker than a skin depth. Typically, the best plane-wave shields will be good conductors with a high conductivity, $\sigma >> \omega\varepsilon$. For good conductors,

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \approx \sqrt{\frac{j\omega\mu}{\sigma}} = \sqrt{\frac{\omega\mu}{\sigma}} e^{j\pi/4}. \tag{10.28}$$

For these materials, $\eta_s << \eta_0$ and Equation (10.27) reduces to,

$$\left|\vec{E}_{\text{trans}}\right| = \left|\vec{E}_{\text{inc}}\right| \frac{4\eta_s}{\eta_0} e^{-t/\delta}. \tag{10.29}$$

If we define the *plane-wave shielding effectiveness* of the slab to be,

$$\text{SE} = 20\log\left|\frac{\vec{E}_{\text{inc}}}{\vec{E}_{\text{trans}}}\right|, \tag{10.30}$$

then the plane-wave shielding effectiveness of an infinite sheet of good conductor can be written in the form,

$$\text{SE} \approx 20\log\frac{\eta_0}{4\eta_s} + 20\log e^{t/\delta} = R(\text{dB}) + A(\text{dB}) \tag{10.31}$$

where the total shielding effectiveness is observed to consist of two terms. The reflection loss, R(dB), represents the attenuation due to the reflection of power at the interfaces. The absorption loss, A(dB), represents the attenuation due to power converted to heat as the wave propagates through the material. The reflection loss is independent of the thickness of the shield and depends entirely on the mismatch between the shield's intrinsic impedance and the intrinsic impedance of free space. The absorption loss is directly proportional to the thickness of the shield expressed in skin depths,

$$A(\text{dB}) = 20\log e^{-t/\delta} \approx 8.7\left(\frac{t}{\delta}\right)\text{dB}\,. \tag{10.32}$$

Example 10-1: Calculating Shielding Effectiveness of Copper Foil

Calculate the shielding effectiveness of a sheet of 2-mil copper foil, $\sigma = 5.7x10^7$ S/m, at 100 MHz.

We start by calculating the skin depth in copper at 100 MHz,

$$\delta_{cu} = 1\Big/\sqrt{\pi f\mu\sigma} = 1\Big/\sqrt{\pi\left(10^8\right)\left(4\pi\times10^{-7}\right)\left(5.7\times10^7\right)} = 6.7\ \mu\text{m}.$$

The material thickness (t = 2 mils = 50.8 μm) is clearly much greater than the skin depth, so (10.17) can be used to approximate the shielding effectiveness. The absorption loss is,

$$A(\text{dB}) \approx 8.7\left(\frac{t}{\delta}\right) = 8.7\left(\frac{50.8}{6.7}\right) = 66\ \text{dB}.$$

To calculate the reflection loss, we need to determine the intrinsic impedance of copper at 100 MHz,

$$\left|\eta_{\text{cu@100MHz}}\right| = \sqrt{\frac{2\pi f\mu}{\sigma}} = \sqrt{\frac{2\pi\left(10^8\right)\left(4\pi\times10^{-7}\right)}{5.7\times10^7}} = 3.7\times10^{-3}\ \Omega.$$

Then the reflection loss is,

$$R(\text{dB}) = 20\log\frac{\eta_0}{4\eta_s} = 20\log\frac{377\ \Omega}{4\left(3.7\times10^{-3}\ \Omega\right)} = 88\ \text{dB}.$$

The overall shielding effectiveness is the sum of the reflection loss and the absorption loss,

$$\text{SE} \approx 88\ \text{dB} + 66\ \text{dB} = 154\ \text{dB}.$$

Note that virtually all the incident power is reflected by the shield. 154 dB is a very large ratio, suggesting that the transmitted power is smaller than the incident power by a factor of 10^{15}. In practice, attenuations of this magnitude are neither realizable nor measurable. As a practical matter, most engineering test equipment has a maximum dynamic range of around 80 – 120 dB. Also, the shielding effectiveness of most shielding enclosures is limited by imperfections such as apertures, seams, and wire penetrations.

If the material in Figure 10.18 is not thick relative to a skin depth, some of the energy that reflects off the second interface (at x=t) propagates back into the slab and is reflected off the inside of the first interface (at x=0^+). This energy will then again strike the second interface, and some fraction will be transmitted adding to the total energy transmitted and reducing the shielding effectiveness. The wave may bounce back and forth multiple times before attenuating to the point where it no longer contributes significantly to the transmitted field. If the absorption loss term in (10.31) is less than about 15 dB, the accuracy of the shielding effectiveness estimate is compromised by these multiple reflections.

For conductive materials that are not electrically thick, we can adjust the expression for shielding effectiveness (10.31) by adding a third term to account for multiple reflections resulting in a general expression for plane-wave shielding effectiveness generally known as the Schelkunoff decomposition,

$$\mathrm{SE} = 20\log\frac{\eta_0}{4\eta_s} + 20\log e^{t/\delta} + 20\log\left|1-\left(\frac{\eta_0-\eta_s}{\eta_0+\eta_s}\right)^2 e^{-2\gamma t}\right| = R(\mathrm{dB}) + A(\mathrm{dB}) + B(\mathrm{dB}). \tag{10.33}$$

Note that the multiple reflection loss term has a negative value comparable to the reflection loss for thin materials. When the multiple reflection loss factor is non-negligible, the shielding effectiveness equation in (10.31) is not accurate and shouldn't be used.

Mismatch Decomposition

The Schelkunoff decomposition of plane-wave shielding effectiveness into reflection and absorption components has the advantage of being relatively easy to calculate from the material properties. Nevertheless, this decomposition can be problematic, because the terms do not correlate to quantities obtained from shielding effectiveness measurements in a straightforward or intuitive manner.

A more useful decomposition of plane-wave shielding effectiveness is generally referred to as the *mismatch decomposition*[3]. This decomposition breaks the shielding effectiveness into two components. The first component, representing the reflection loss, is called the *mismatch loss*, and is calculated as,

$$L_M = 10\log_{10}\left(\frac{\left|\eta_0+\eta_{\mathrm{in}}\right|^2}{4\,\mathrm{Re}(\eta_0)\,\mathrm{Re}(\eta_{\mathrm{in}})}\right), \tag{10.34}$$

where η_{in} is the wave impedance looking into the finite-thickness shield material,

$$\eta_{\mathrm{in}} = \eta_0\left(1+\Gamma_{\mathrm{in}}\right)/\left(1-\Gamma_{\mathrm{in}}\right). \tag{10.35}$$

Because η_0 is real, mismatch loss can also be expressed as:

$$L_M = -10\log_{10}\left(1-P_R\right). \tag{10.36}$$

where P_R is the normalized reflected power (i.e., reflected power divided by incident power). Therefore, unlike the reflection loss in the Schelkunoff decomposition, the mismatch loss is directly related to the reflected power.

The second component of the mismatch decomposition, representing the absorption loss, is called the *dissipation loss*, and is calculated as,

[3] A. McDowell and T. Hubing, "Analysis and comparison of plane wave shielding effectiveness decompositions," *IEEE Trans. on Electromagnetic Compatibility*, vol. 56, no. 6, Dec. 2014, pp. 1711-1714.

$$L_D = -10\log_{10}\left(\frac{P_T}{1-P_R}\right). \tag{10.37}$$

where P_T is the normalized transmitted power reaching the load. Note that the mismatch loss (10.36) and the dissipation loss (10.37) add up to the total shielding effectiveness,

$$\text{SE} = L_D + L_M\,. \tag{10.38}$$

Unlike the Schelkunoff decomposition, the mismatch decomposition has only two components that are easily expressed in terms of the amounts of power reflected and absorbed by the shielding material. Additionally, since the power absorbed by a shield will go down if that shield becomes a better reflector, it makes sense to quantify the ability of a shield to absorb power by comparing the absorbed power to the power that is not reflected, as the mismatch decomposition does.

Figure 10.19 shows a plot of the calculated plane wave shielding effectiveness of a very thin (10 µm) copper sheet. The overall shielding effectiveness is approximately 100 dB at 1 MHz and steadily rises from 100 dB to 200 dB from 100 MHz to 10 GHz. The Schelkunoff decomposition terms R, A and B indicate that the reflection loss dominates at all frequencies. The absorption loss is near 0 dB from 1 to 100 MHz and only starts to contribute to the overall shielding effectiveness at frequencies above 100 MHz. On the other hand, the mismatch decomposition tells us that the reflection loss, L_M, and the absorption loss, L_D, contribute equally to the overall shielding effectiveness from 1 MHz to 100 MHz; and that the absorption loss dominates at frequencies above 100 MHz.

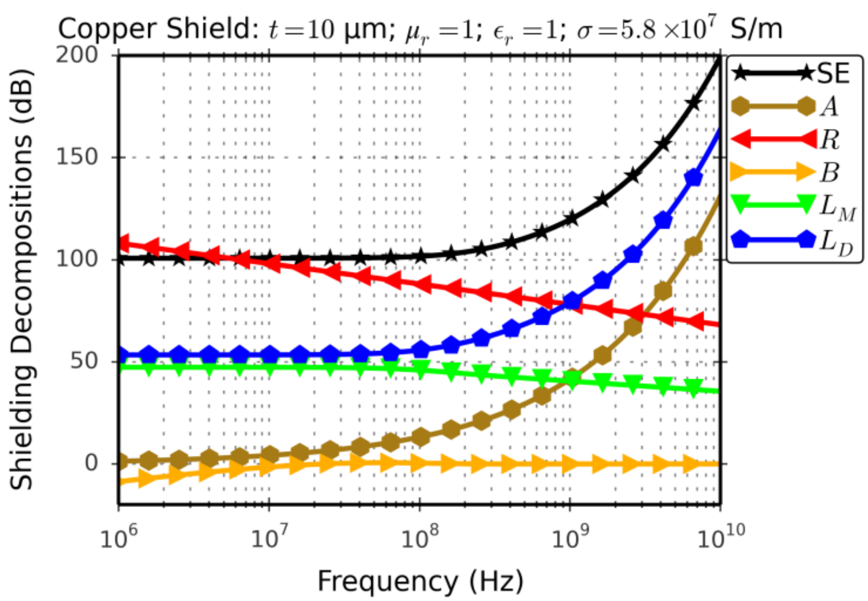


Figure 10.19. Plane wave shielding effectiveness of 10-µm copper.

Figure 10.20 shows a plot of the calculated plane wave shielding effectiveness of a 3-mm thick carbon nanofiber composite material. This material has an overall shielding effectiveness of only about 17 dB. The major advantage of this material is its ability to absorb a significant amount of the incident power so that less is reflected back into the enclosure. Nevertheless, the absorption loss term of the Schelkunoff decomposition is near 0 dB from 1 to 100 MHz. The reflection term minus the multiple reflection term accounts for nearly all the shielding effectiveness. The mismatch decomposition shows that the absorption loss dominates, accounting for 11 dB of the 17 dB total shielding effectiveness.

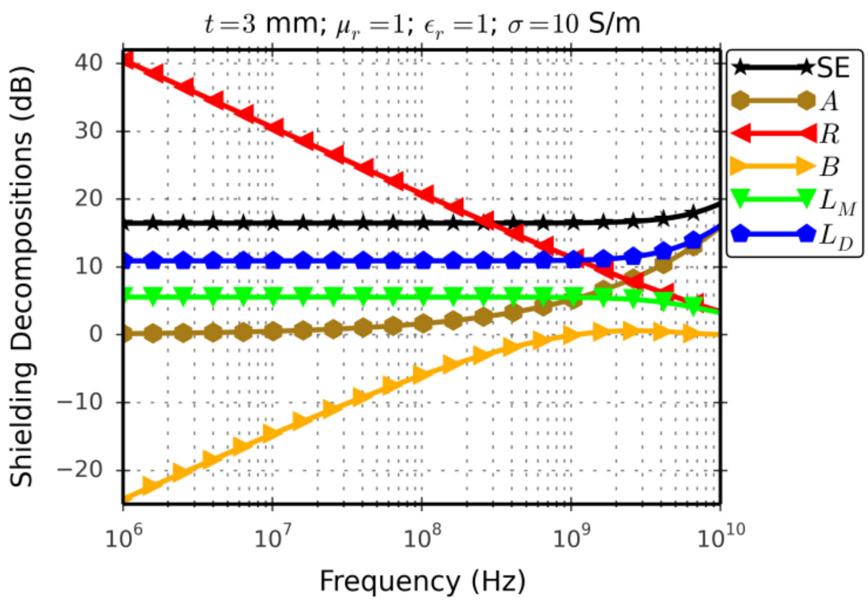


Figure 10.20. Plane wave shielding effectiveness of 3-mm carbon nanofiber composite.

Near-Field Shielding

Plane-wave shielding theory conveniently permits us to calculate a shielding effectiveness value for any shielding material based on its material properties and thickness. Unfortunately, practical shields are never located in the far-field of both the source and victim circuits. Because of this, we are very unlikely to have plane wave propagation on both sides of the material and the calculated shielding effectiveness will not correspond to anything we are likely to measure (except in specially designed test fixtures).

To help understand how near-field shielding differs from plane-wave shielding, consider the configurations shown in Figure 10.21. In Figure 10.21(a), the incident plane wave has been replaced by a small electric dipole source and the shielding material is in the near-field of the source. In Figure 10.21(b), the source is a magnetic dipole, represented by a small loop of electric current.

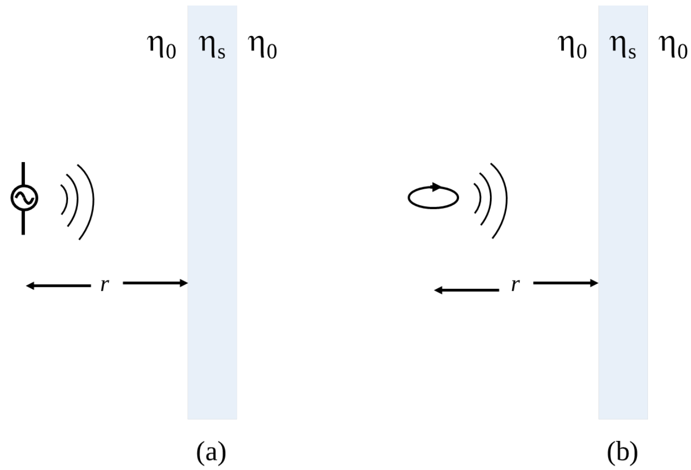


Figure 10.21. Shielding electric and magnetic dipole sources.

Recall that in the near field ($r << \lambda$), an electric dipole source has a strong electric field. The wave impedance in the near-field is approximately,

$$Z_{W_E} = \frac{\left|\vec{E}\right|}{\left|\vec{H}\right|} \approx \frac{1}{2\pi f \varepsilon_0 r} . \tag{10.39}$$

In the near field of a magnetic dipole source, the magnetic field dominates, and the wave impedance is approximately,

$$Z_{W_H} = \frac{\left|\vec{E}\right|}{\left|\vec{H}\right|} \approx 2\pi f \mu_0 r . \tag{10.40}$$

We can estimate the shielding effectiveness of the slab in Figure 10.21, by substituting the wave impedance ($Z_W = Z_{W_E}$ or Z_{W_H}) for the intrinsic impedance of free space, η_0, in (10.33). This yields a new expression for the reflection loss term,

$$R(dB) \approx 20 \log \frac{Z_W}{4\eta_s} . \tag{10.41}$$

The expressions for absorption loss and multiple reflection loss are unchanged. Although this type of shielding effectiveness calculation is a simple approximation that does not correspond to any realizable test structure, it can provide insight relative to the performance of various shielding materials in realistic situations.

Example 10-2: Shielding a Low-Frequency Magnetic Field Source

A transformer generating primarily a magnetic field is located 10 cm from a shielding structure. The shielding structure is made from a 1-cm thick sheet of copper. Estimate the shielding effectiveness of this structure at 1.5 kHz.

If we start by modeling the transformer as a magnetic dipole source, we can estimate the wave impedance at the position of the shield to be,

$$Z_{W_H} \approx 2\pi f \mu_0 r = 2\pi\left(1.5\times10^{3}\right)\left(4\pi\times10^{-7}\right)\left(0.10\right) = 1.2\times10^{-3}\ \Omega.$$

The intrinsic impedance and skin depth of the copper are,

$$\left|\eta_{\mathrm{cu@1.5kHz}}\right| = \sqrt{\frac{2\pi f\mu}{\sigma}} = \sqrt{\frac{2\pi\left(1.5\times10^{3}\right)\left(4\pi\times10^{-7}\right)}{5.7\times10^{7}}} = 14\times10^{-6}\ \Omega$$

$$\delta_{cu} = 1 \Big/ \sqrt{\pi f\mu\sigma} = 1 \Big/ \sqrt{\pi\left(1.5\times10^{3}\right)\left(4\pi\times10^{-7}\right)\left(5.7\times10^{7}\right)} = 1.7\ \mathrm{mm}.$$

The calculated shielding effectiveness is therefore,

$$\begin{aligned}\mathrm{SE} &= 20\log\frac{0.0012}{4\left(14\times10^{-6}\right)} + 20\log e^{10/1.7} + 20\log\left|1+e^{-2\left(10/1.7\right)}\right| \\ &= 26\ \mathrm{dB} + 51\ \mathrm{dB} + \sim 0\ \mathrm{dB} \\ &= 77\ \mathrm{dB}.\end{aligned}$$

Note that in this case the absorption loss plays an important role in the overall shielding effectiveness. Generally, at low frequencies close to a magnetic field source, the wave impedance is low and therefore the reflection loss due to conductive shields is less significant. Absorption loss also decreases with frequency, but not as quickly as reflection loss.

Shielding Effectiveness Measurements

Plane-Wave Shielding Effectiveness

As discussed in the previous section, the concept of plane-wave shielding effectiveness is convenient because it is a function of only the material properties and thickness of a shielding material. Attempts to measure the plane-wave shielding effectiveness generally involve launching a guided TEM wave in a coaxial test fixture containing a sample of the material, as illustrated in Figure 10.22.

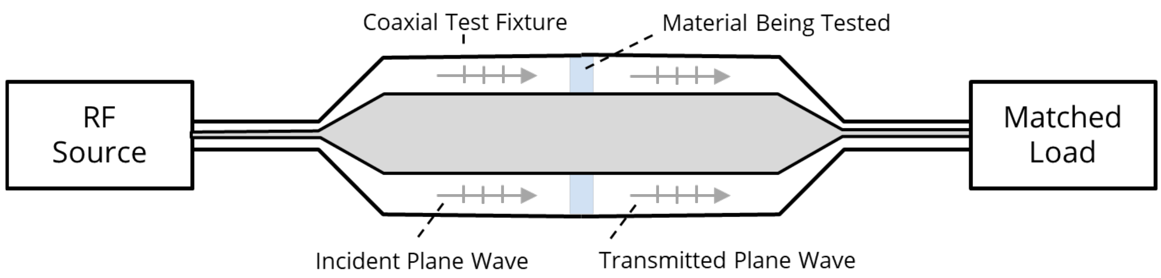


Figure 10.22. Shielding Effectiveness Test Fixture.

The transmission line structure has a specific characteristic impedance (usually 50 Ω). The cross-sectional dimensions are scaled up in the mid-section of the test fixture to accommodate a reasonably sized material sample, which is disk-shaped with a hole in the center. The measured shielding effectiveness is simply calculated as,

$$\text{SE} = 10\log\frac{\text{forward power from the source}}{\text{power received at the termination}} \; . \tag{10.42}$$

When measurements are made with a vector network analyzer, the shielding effectiveness can be conveniently expressed in terms of the s-parameters as,

$$\text{SE} = 20\log\left|S_{12}\right| \; . \tag{10.43}$$

Note that even though the characteristic impedance of the test fixture is 50 Ω, the ratio of |E| to |H| is still determined by the intrinsic impedance of the medium ($\eta_0 \approx 377\,\Omega$ in air).

Other Shielding Effectiveness Measurements

Of course, the shielding effectiveness of an enclosure is likely to be very different from the plane-wave shielding effectiveness of the material from which the enclosure is made. Many factors influence the shielding effectiveness of an enclosure including the size and shape of the enclosure and the type and location of the source. Also, typically power escaping through apertures and seams in a real enclosure is much more significant than any power propagating directly through the enclosure walls.

For this reason, it is usually more practical to define the shielding effectiveness of an enclosure as follows,

$$\text{SE} = 20\log\frac{\text{E-field received from source with no shield}}{\text{E-field received from shielded source}} \; . \tag{10.44}$$

For example, suppose the measured radiated field from an electronic product was measured with no enclosure (or a plastic enclosure) and found to be 52 dB(μV/m). Then suppose that the same product was tested in the same manner with a metallic enclosure and the measured field strength was 38 dB(μV/m). The shielding effectiveness of the enclosure in this particular configuration would then be reported as,

$$\text{SE} = 52\text{ dB}(\mu\text{V/m}) - 38\text{ dB}(\mu\text{V/m}) = 14\text{ dB}. \tag{10.45}$$

This is probably a much lower value than the plane-wave shielding effectiveness of the enclosure walls, but it accounts for the leakage through apertures and seams. It also accounts for the fact that shielded enclosures generally interact with the enclosed sources and the enclosure itself becomes an integral part of the unintentional *antenna*, converting currents into radiated fields.

Note that the shielding effectiveness of an enclosure is a function of the source type, the source location, the receiver location, and other environmental parameters. The measured shielding effectiveness strongly depends on the method used to make the measurement. Therefore, it's rarely helpful to try to compare the measured shielding effectiveness of two enclosures unless both measurements were made in exactly the same way. Also, one should not expect the measured shielding effectiveness of an enclosure to correlate very well with the external field attenuation observed when using that enclosure to shield complex electronic products.

Quiz Question

The shielding effectiveness of an enclosure made of a material that has a plane-wave shielding effectiveness of 60 dB is,

a.) ~60 dB
b.) always less than 60 dB
c.) usually greater than 60 dB
d.) sometimes less than 0 dB

The best answer to this question is (d.) An inefficient radiation source (e.g., an electrically small circuit) can become many orders of magnitude more efficient by coupling to a larger conducting structure. Therefore, it is possible for a shielding enclosure with apertures or seams to increase the radiated emissions from the enclosed sources. In other words, the shielding effectiveness of a metal enclosure can be less than 0 dB at some frequencies. Hopefully, the same enclosure also reduces the efficiency of the strongest sources so that the net effect is a reduction in the maximum radiated emissions. Nevertheless, it is not safe to assume that some shielding is better than no shielding.