Cables, Transmission Lines, and Shielding for Audio and Video Systems

by Jim Brown
Audio Systems Group, Inc.
http://audiosystemsgroup.com

Every cable (that is, a pair of wires, or a wire surrounded by a shield) has series inductance and resistance, parallel capacitance, and dielectric loss in the insulation between conductors.

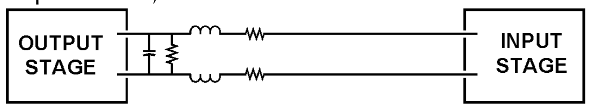

Fig 1a – Two devices connected by a balanced cable

For high frequency signals, the cable will behave as an infinite number of tiny inductors, capacitors, and resistances that comprise a ***transmission line*.** Fig 1b and 1c show a circuit model that can help us understand how the signal travels along the line. In the model, G is a conductance, having the units of mhos (1/ohms) When an impulsive signal is impressed upon the line (for example, part of a digital or video signal), it charges the tiny capacitances through the tiny inductances, with loss contributed by the tiny resistances, and each of the tiny capacitances are charged and discharged in sequence as the signal moves along the line.

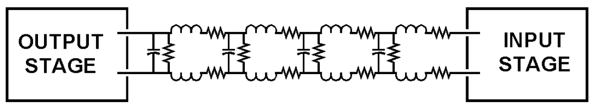

Fig 1b – Fundamental transmission line model

With each charging and discharging of the capacitance, energy will be traded back and forth between the magnetic field established by the cable's inductance, and the electric field established by the capacitance between the two conductors. In addition, a small amount of the energy will be converted to heat within the two resistive components -- that is, the wire resistance and the dielectric losses (conductance) in the insulation between the conductors.

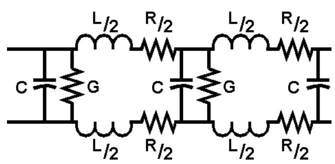

Fig 1c – Fundamental transmission line model with circuit parameters

When transmission line effects are minutely small, as they are in the vast majority of cables carrying audio signals, the pulses travel at the speed of light. A relatively short length of cable shows no measurable transmission line effects for low frequency signals. But if the cable is long enough (or the frequency components of the signal are high enough), transmission line effects will begin to appear with increasing frequency or length. When that happens, the time constant associated with the R, L, C, and G will result in a somewhat lower velocity of transmission along the line, while the resistance and dielectric loss cause some attenuation of the signal that increases with increasing frequency. The ratio between these basic parameters, established by the cable's physical construction, establish its ***characteristic impedance***, (Z_O) and in combination with the loss parameters, the velocity at which the signal will move along the line (called its ***velocity of propagation***) (V_P). Mathematically,

$$Z_O = \sqrt{\frac{R + j2\pi fL}{G + j2\pi fC}}$$

At high frequencies, the equation simplifies to $Z_O = \sqrt{\frac{L}{C}}$

For any given cable type, Z_O and V_P are essentially constant above about 50 kHz, but at low frequencies, vary considerably with frequency before settling to their high frequency values! The more complex form of the equation must be used with very low frequency transmission lines – for example, audio frequency telephone lines that are tens of miles long. For all other uses of transmission lines, the simplified equation is entirely adequate.

The ***velocity of propagation*** (V_P) (the value at high frequencies) in practical cables ranges between about 65% and 85% of the speed of light, depending on their construction, and the ratio between the cable velocity and the speed of light is called the ***velocity factor** (VF)* for the cable. Mathematically,

… $VF = 1/\sqrt{\varepsilon}$ where ε is the dielectric constant relative to free space.

If the receiving end of a transmission line is terminated in a resistance equal to its characteristic impedance, the signal will simply travel to the end of the line and be dissipated in that resistance, and the signal will be cleanly received. Such a line is said to be ***matched***. If the termination has a different value, or if it has capacitive or inductive components, some of the signal will be reflected from the end of the line and travel back toward the receiver. Such a line would be described as ***mismatched***. When this happens, the pulses or video waveform that make up the signal will be distorted by the (algebraic) addition of the reflection). A reflection will also occur if two transmission lines are connected in parallel (that is, if there were a Tee connection to split one line into two lines). [Note: there are some very special techniques that can be used to connect two lines in parallel, using matching stubs or baluns. Such techniques are outside the scope of this tutorial.]

The characteristic impedance of a transmission line is symbolized by $\mathbf{Z_O}$. A matched transmission line has the convenient property of disappearing in the circuit model – the output stage sees a pure resistance of $\mathbf{Z_O}$, so exactly half of the output voltage appears at the input to the line. This voltage will travel to the other end of the line, and will be attenuated by the line. Whatever is left will be dissipated in the $\mathbf{R_{IN}}$. In other words, the effect of the line is to simply attenuate the signal. Another way of looking at it is to say that a proper termination makes the line act as if it were infinitely long.

Fig 1d – A matched transmission line "disappears" – the output stage simply sees $\mathbf{Z_O}$ and the load sees the output voltage attenuated by the loss in the line

But, as usual, it isn't quite that simple. Because the loss in the cable varies with frequency, the cable will modify the frequency response of the system. The attenuation of virtually all real cables increases with frequency, so they act as low pass filters, with the steepness of the rolloff being proportional to the length of the cable. (See Fig 2a) Some video distribution amplifiers and line drivers include variable pre-emphasizing equalization that can be adjusted to compensate for the rolloff in cables of various lengths and types. The phase shift produced by these equalizers can distort the signal, and must be taken into account in their design and adjustment.

In a transmission line, the energy moves through the dielectric – that is, the insulating material between the conductors – as an ***electromagnetic field***. In any electromagnetic field, the energy is continuously being traded (on each quarter cycle of the waveform) between the electric field (that is, charging the capacitance) and the magnetic field (established by the inductance). In general, the larger the dielectric, the more efficiently it can transport the field, and the lower will be the loss in the cable. Obviously, the dielectric <u>material</u> matters too – some materials offer much lower loss than others. Fig 2b shows the loss in 100 ft of low loss coaxial cables of different sizes. The cable with the greatest loss is the "mini-coax," which is about 0.1" o.d., and has a $\mathbf{Z_O}$ of 75Ω. Next is RG-58, 0.195" o.d., also 50 Ω. Next are an RG-59 and RG-8X, at 0.22" and 0.24" o.d. respectively. Next are three 0.4" o.d. cables – 9212 is RG-11 optimized for high power at HF, 9913F is a flexible RG-8 with a foam dielectric for transmission and reception, 3227 is a very stiff cable optimized for use at cell phone sites (800 – 1,900 MHz). The RG-59 and RG-11 cables are 75 Ω; the others are 50 Ω.

Fig 2b shows two other effects relative to cable construction. The primary difference between 9212 and 9913F is their dielectric material – 9913F is designed for constant flexing and has a special foam dielectric that provides superior performance at VHF and UHF, while 9212 uses a conventional dielectric optimized for use with high power transmitters below 30 MHz. The increased loss below 3 MHz for the RG59 occurs because its center conductor is copper-coated steel (and thus a higher DC resistance). Other RG59's have significantly less loss.

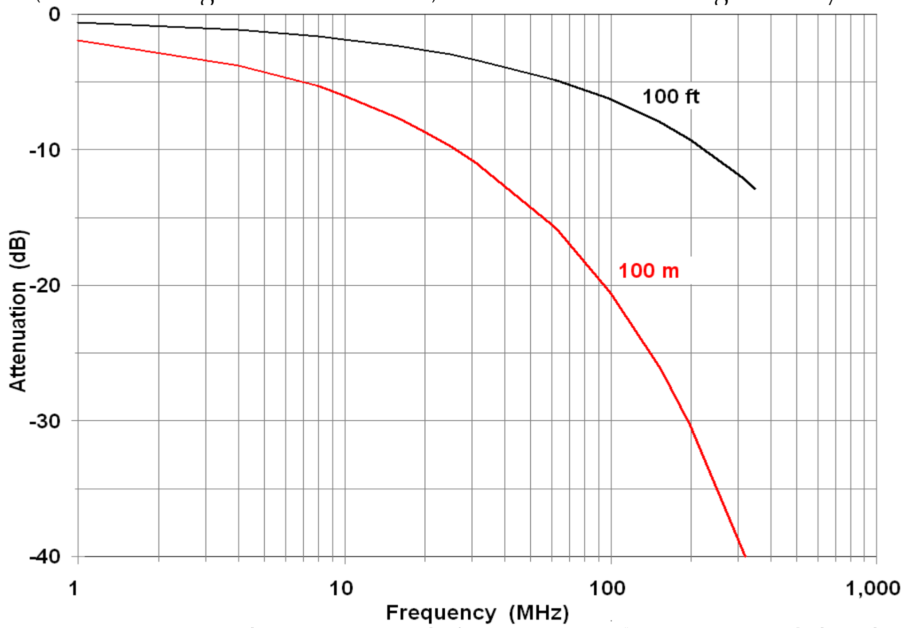

Fig 2a – Loss in transmission lines varies with frequency and increases with line length. This data is for a premium CATx data cable (Belden Mediatwist)

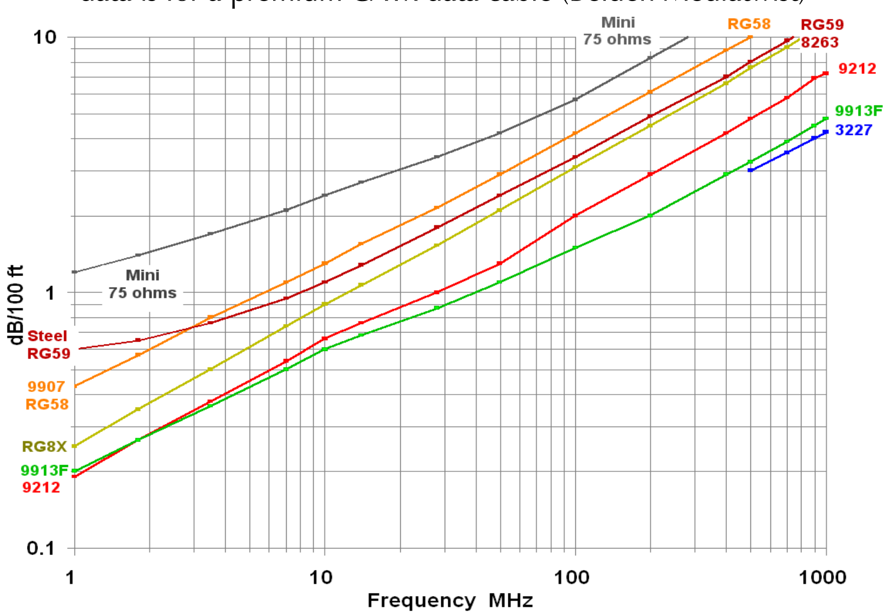

Fig 2b – Size matters – Larger cables have less loss

Coaxial Cable Types – What's All this RG-stuff? The RG-numbers for coaxial cables date from at least the days before WWII. When they were developed, coax was used for radio, and radio more or less stopped at 50 MHz. In those days, coax was coax – if you wanted to handle more power or wanted lower loss, you used bigger cable! We learned that RG-58 and RG-59 were the "regular" 50 ohm and 75 ohm cables respectively, RG-8 and RG-11 were the big stuff. The 50 ohm impedance was a good match to a "ground plane" antenna, and became the standard for two-way radio (that's why it's used for wireless mics – the engineers come out of that world, and so does the test equipment!). The 75 ohm impedance was a good match for dipole antennas, and later became the standard for video and for TV antenna systems.

An important digression: We need to define the term "***baseband***" as a descriptor for audio and video that is not modulating (being carried on) a radio frequency carrier. "***Baseband***" audio refers to audio as it comes out of a microphone or mixer – it's plain old audio as we know it. Likewise, "***baseband***" video is what comes out of a camera or other signal source, or as it would be fed to a video display. The spectrum of "***baseband***" audio is 20 Hz to 20 (or 100) kHz, and the spectrum of ***baseband*** video is from near dc to tens of MHz. We must analyze the circuits (amplifiers and cables) that carry these signals in terms of how they behave at these "***baseband***" frequencies. On the other hand, when these signals are carried on a radio carrier, we must analyze the cables and electronics in terms of their performance at the frequencies of the radio carriers.

Today, there are nearly 50 different RG-59 cables in the Belden catalog, a dozen or so each of RG-58's, RG-8's, and RG-11's. RG-58 is typically 0.19" o.d., RG-59 about 0.24" o.d., RG-8 about 0.4" and RG-11 about 0.45" o.d. RG-6 (.275" o.d.) and RG-8X (.242" o.d.) are a sort of "super RG-59 and RG-58 respectively, with roughly 20% less loss than their smaller cousins. And all of these cables are roughly 20% smaller in their plenum-rated versions.

But that's only the beginning of the story about coaxial cables. The more important differences between them fall into several basic categories:

1. The composition of the cable's dielectric. Cables that must have low loss at VHF and UHF typically have some form of foam dielectric. Foam is not the right choice for all cables – it deforms easily, so is not ideal for cables that must withstand lots of handling. Foam also can absorb moisture, and when it does, the loss increases drastically. This can make its use problematic when it is exposed to the elements.

2. The composition of their center conductor. Cables that will be used only at high radio frequencies often have a center conductor of copper-coated steel. Because skin effect causes the current to be concentrated on the outer skin of the center conductor at radio frequencies, these cables can have fairly low losses at HF. At low frequencies where skin effect is much less pronounced, the high resistance of their center conductor makes them perform poorly with ***baseband*** video (see Fig 2b). On the other hand, video requires a cable with a "beefy" copper center conductor to carry the 60 Hz components. For audio, the resistance of the center conductor doesn't matter, because audio inputs have fairly high impedances.

3. The composition of their shields. Because both video and (sometimes) audio are transmitted unbalanced, the shield often carries noise current. Both video and unbalanced audio circuits benefit greatly from having the lowest possible resistance in the shield to minimize the IR drop from the hum component of that current. Cables like Belden 8241F (single copper braid shield) and Belden 8281 (double copper braid shield) offer the best performance.

   On the other hand, cables that will be used in RF signal distribution for receiving systems (wireless mics, MATV and CATV systems) don't need good low frequency performance, but they often do need very good shielding. Double and quad-shielded cables with foil and braid shields work fine. They are less costly and are lighter weight than the higher density copper braid shields that don't provide as much shielding at higher frequencies.

4. Their physical handling properties. In some applications, flexibility is important, in others it is not. Some cables that will be suspended over a long span have a steel messenger cable bonded to their jacket. Foil shields don't generally hold up to handling very well.

5. Their resistance to UV radiation. The outer jacket of most cables can decompose when exposed to UV radiation and migrate through the shield into the dielectric. That can seriously degrade the cable's loss performance. Cables with a "non-contaminating" jacket have a chemical composition to resist that decomposition. The letters "NC" in a catalog indicate this type of cable.

6. Their power-handling capabilities. This is of interest only when the cable will be used with high power radio transmitters; all but mini-coax will handle hundreds of watts.

Table 1 – Coaxial Cable Applications Matrix

Application	Z_O	Shield	Center conductor	LF Loss	VHF Loss
Baseband video – NTSC	75Ω	Beefy copper	Beefy copper	critical	not critical
Baseband video – high res	75Ω	Beefy copper	Beefy copper	critical	critical
Baseband audio	any	Beefy copper	doesn't matter	critical	doesn't matter
Wireless mics, MATV, CATV	75Ω	Double/quad foil/braid	copper, copper coated steel	doesn't matter	very important
AES-3id	75Ω	copper	copper	doesn't matter	important
Infrared radiators	50-75Ω	copper	copper	somewhat important	doesn't matter

Note: The listing of 75Ω for wireless mics, MATV, and CATV is not a mistake. That's what should be used here, because 75 ohm cables have far less loss than 50 ohm cables and are ¼ the cost of 50 ohm cables.

What about standing waves? Anyone working with sound systems knows about acoustic standing waves in rooms. A sound source excites a room, and depending on the room geometry and the wavelengths of the sound, peaks and valleys of energy will be established at various points within the room as the sound bounces around it. In simple terms, the most pronounced standing wave patterns occur when the wavelength of the sound corresponds to some multiple of quarter-wavelengths in the acoustic path, and when the surfaces involved are most highly reflective.

A transmission line is simply a two-dimensional electrical equivalent of standing waves in a room. When the transmission line is terminated in $\mathbf{Z_O}$, there is no reflection, so there are no standing waves. The more the termination differs from $\mathbf{Z_O}$, the more energy will be reflected. At some frequencies the line will be resonant, and pronounced maxima and minima will occur.

Are the standing waves in transmission lines important to audio and video engineers? The answer is generally no – except when they distort the waveform of digital data or video. We'll address that issue later. Standing waves are mostly important to engineers designing radio transmitting systems, and to some extent, very high performance receiving systems. Standing waves can be **used** to create tuned traps – resonant lengths of transmission line (called stubs) that can be used to select or reject a fairly narrow range of frequencies.

To understand how a stub works, consider a short piece of transmission line that is precisely one-quarter wavelength at some frequency, and that it is an open circuit at the far end. If a sine wave at that frequency is generated at the sending end, it will travel to the far end and be perfectly reflected back to the sending end, where it will be 180 degrees out of phase with the generator. If there is NO loss in the line, these two signals will cancel, and the total voltage will be zero. The line looks like a short circuit, but only at that frequency – or, more correctly, it looks like a series resonant circuit tuned to its resonant frequency.

The standard use of such a stub is to connect it in parallel with a transmission line, and tune it to some frequency that you need to reject. Signals at that frequency will be shorted out, others will not. The limitation of this technique is that the signal you need to reject must be greatly different from the signal that you want to receive. The major practical use of this technique is to reduce the strength of a harmonic produced by a transmitter, or to reduce the strength of a strong signal at some widely separated frequency that could overload your equipment.

A stub that is shorted at the far end works exactly the opposite – the short kills signals at the resonant frequency, but reflects all others back to the source. Thus a shorted stub acts like a parallel resonant circuit tuned to its resonant frequency.

Again, it isn't that simple, because REAL transmission lines have loss, so the reflected wave will be weaker than the signal from the generator, and the cancellation will be imperfect. The more loss in the line, the less cancellation there will be. Because of this loss issue, transmission line stubs are not very useful to us – their Q is too low. But one of their electromechanical cousins is – the resonant cavity!

A cavity is simply an electrical enclosure that can be made the equivalent of a stub, but with FAR less loss. A small inductive loop couples energy into the cavity. At it's resonant frequency all of the energy is absorbed by the resonance. Such a cavity is called a "band reject" cavity, and is used to reject a single very strong signal.

The most useful type of cavity is called a "band pass" cavity. It has two coupling loops, one at an input and one at an output. Signals at the resonant frequency are coupled through the cavity, others are rejected. Resonant cavities can be tuned for extremely narrow bandwidths, providing strong rejection of interference quite close in frequency to a desired signal. The cavity of Fig 3 is designed for use in two-way radio systems. It works in the 400-470 MHz spectrum, has a passband on the order of 1 MHz, and an insertion loss of less than 0.2 dB.

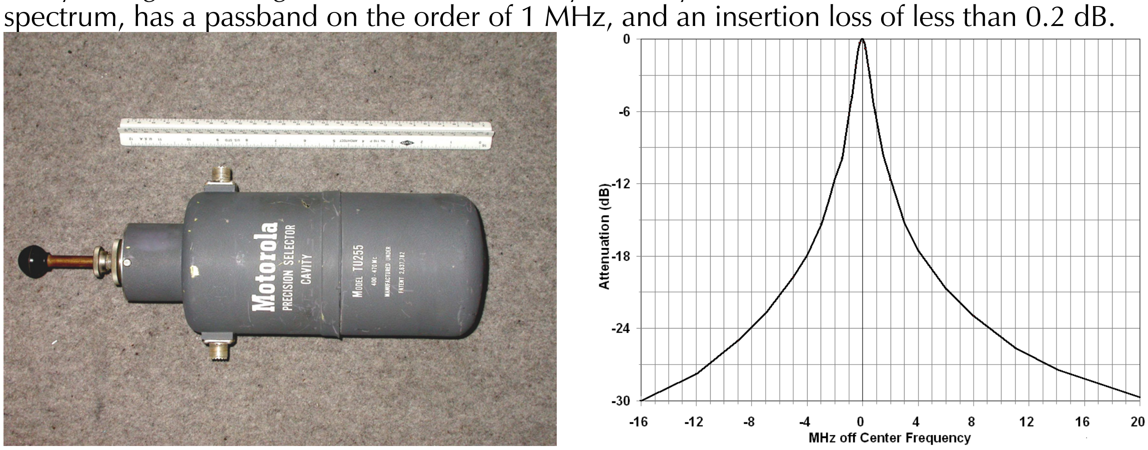

Fig 3 – A 450 MHz cavity and its response

Standing waves and loss In college transmission lines courses, engineers are taught that "standing waves increase the losses in the line," and the message is that they should be avoided at all costs. This is one of those "grain of truth lies" – losses <u>do</u> increase when standing waves are present, but unless the mismatch is really severe, losses don't increase enough to matter. The <u>worst</u> increase in loss that could be caused by the mismatch between 50 ohms and 75 ohms is 0.2 dB, for <u>any</u> length of line, no matter what the loss was without the mismatch! So from a <u>loss</u> point of view, such a small mismatch simply does not matter! The <u>only</u> time we need to be careful about getting the impedance right is when we are transmitting high speed data or high resolution video.

Twisting Cable pairs are twisted together for two very important reasons. First, bringing them more tightly together reduces the coupling of external magnetic fields (while increasing the coupling between the conductors) by reducing the loop area between them. Second, twisting them together in a very symmetrical fashion causes any noise coupled onto one conductor to be more perfectly cancelled (in the receiver) by noise coupled onto the other conductor. ***Twisting reduces both magnetic (inductive) and electric (capacitive) coupling.***

To understand how twisting does this, consider a magnetic field from a source that is closer to one side of the cable than the other. At any point along the cable, one conductor will be closer to the source than the other, so the induced voltage will be greater in that conductor than in the other. But one half twist along the cable in each direction, the other conductor will be closer to the source, and so will have the greater induced voltage, but the polarity will be opposite. The more symmetrical the twisting, and the "tighter" the twisting, the more perfectly the two induced voltages will match each other over the length of the cable, and thus be better cancelled by the receiver. The number of twists per unit length is called the "lay" of the cable.

Twisting also reduces capacitive coupling onto the cable, and for the same reasons. The ability of twisting to reduce coupling extends to very high frequencies. Ethernet networks run on high quality, unshielded, twisted pairs at frequencies in the hundreds of MHz, and require good crosstalk rejection to function well.

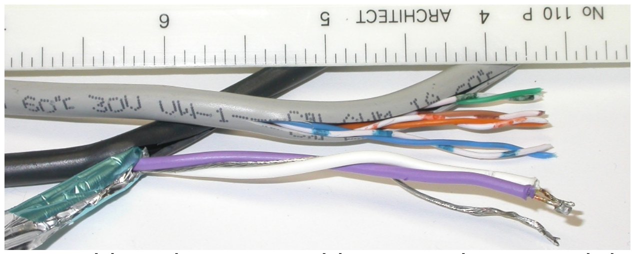

Fig 4 – Two twisted pair cables. The upper cable (CAT5) has a much higher twist ratio (lay)

Lumped Parameter Analysis

What if a cable is not a transmission line? If the cable is too short (or the frequency of the signal is too low), the cable will not act like a transmission line. Instead, it simply looks like the sum of all the series R, series L, and parallel C – the parallel loss component (G) is important only at radio frequencies. So, at audio frequencies, the tiny inductances, capacitances, and resistances add to form pair of series inductances and resistances, and a parallel capacitance. This simplified circuit that ignores transmission line effects is called a ***lumped parameter*** model. ***Lumped parameter*** models describe how a cable behaves at audio frequencies.

In audio systems, the inductance and resistance are too small to affect transmission of the signal in all but the longest cables, but capacitance is not – indeed, the capacitance can often be large enough to draw significant current from the audio source at high audio frequencies. To exhibit even the slightest transmission line effects, a cable would need to be longer than 1/20 wavelength at the frequency of a signal it carries. This works out to be nearly one mile (1.6 km) for a 20 kHz signal, but only 16 feet (5 m) for a 6 MHz signal.

Note: it is important to realize that the transition from simple lumped parameter to transmission line behavior is a gradual one. Transmission line effects <u>do</u> exist at all frequencies and for all cables, but for cables that are shorter than $\lambda/20$, they are simply too small to matter.

What Does the Cable Look Like At Audio Frequencies? Figs 5a-5d traces the development of the equivalent circuit for the cable connection between two pieces of equipment. Fig 5d is the circuit we will work with.

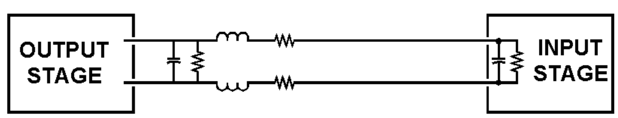

Fig 5a – An unshielded, balanced cable connecting two stages, ignoring stray C to "ground"

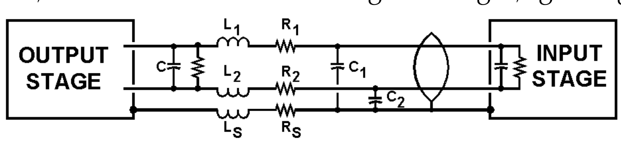

Fig 5b – A cable shield complicates things by adding more capacitors , R_S , and L_S .

Fig 5b shows that the shield complicates things considerably. First, there will be capacitance between the shield and each signal conductor. Second, the shield has both resistance and inductance along its length. Third, there will be <u>mutual</u> inductance between the three conductors. These components affect the behavior of the cable in several very important ways that we will study one at a time. We'll look at the effect of the capacitors first, and to do that, we'll pretend that the inductances are not there.

Capacitive imbalance C is the lumped capacitance between conductors – it is equal to the capacitance per foot (meters) multiplied by the cable length in feet (meters). C_1 and C_2 are the capacitances between the "red" wire and the shield, and between the "black" wire and the shield. In a perfect world, they would be equal, but they are not. It is common for C_1 and C_2 to differ by 4% for good cables, and by as much as 10% or more for poorly manufactured cables.

There are several reasons for this capacitance imbalance. First, the dyes used for different colors of plastic insulation can have different values of dielectric constant. Second, the insulations may have slightly different thicknesses. Third, poor control of tolerances in the manufacturing process can result in the pairs being off center, or for their centering to "wobble" from one part of the cable to another. "Wobble" can also occur due to mechanical deformation of the shield by a drain wire.

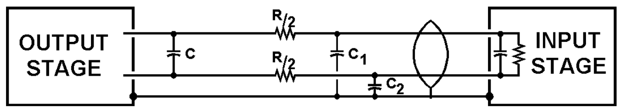

Fig 5c – For differential signals, inductance L is negligible.

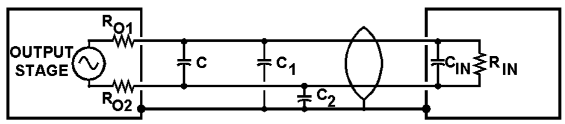

Fig 5d – The simplified circuit for differential signals – because (R_{IN}+R_O) is large, cable resistance R can be ignored.

Figs 5a-5d apply to paired cable whether it is shielded or not, and whether it is twisted or not. What changes from one type of construction to another are the relative sizes (and balances) of the component values. An un-shielded cable can have much smaller values of C_1 and C_2 than a shielded cable, but values of C_1 and C_2 will be affected by surrounding objects, and their imbalance can be much greater than for a shielded cable. And, of course, there will be no R_S and L_S in unshielded cables.

It always helps to "plug in" some circuit values that represent real world conditions. Let's say that we have 300 feet of Belden 8451 connecting a condenser microphone to a mix console. The output stages of condenser microphones and line level equipment have an output impedance on the order of 150-300 ohms, Belden's data sheet says that C is 34 pF/ft, and that C_1 and C_2 are about 33 pF/ft (the measurement for C_1 and C_2 is performed by connecting the "black" wire to the shield and measuring between "red" and shield). So for 300 ft, C, C_1, and C_2 are all about 10 nF, and they are in parallel with a typical R_{IN} of 1,500 ohms and C_{IN} of 1-10 nF.

If we replace that cable with a good AES3 digital audio cable like Gepco 5526EZ, C= 14 pF/ft, C_1 and C_2 are about 13 pF/ft. So for the 300 ft run, C, C_1 and C_2 are all about 4 nF. In addition to the capacitances being a lot smaller, cables designed for AES3 tend to be manufactured to closer tolerances, and Whitlock has found that they tend to have about half as much capacitive imbalance as analog cables. Also, 10 nF looks like 750 ohms at 20 kHz and 375 ohms at 40 kHz. This kind of load can cause some output stages to produce slew rate limiting distortion on strong high frequency transients. The much lower capacitance between the conductors of cables designed for AES3 makes them the best choice for balanced analog circuits too.

When we study balanced interfaces, we'll learn about how these differences affect the ability of a circuit to reject noise and RF. Suffice it to say at this point that at 20 kHz and above, the capacitive imbalance of these cables will dominate the total circuit impedance, and this difference can seriously limit the common mode rejection that can be achieved.

Inductive Imbalance – SCIN The second imperfection in shielded, balanced cable that we will study is the imbalance in the inductive coupling between the "red" wire and the shield as compared to the coupling between the "black" wire and the shield. This balance is important, because noise current often flows on the shield. This imbalance can occur in any shielded cable, but it is far greater in cables having a foil shield and a drain wire. The lower cable in Fig 4 and the cable in Fig 6a illustrate the problem.

Fig 6a – Foil/drain shielded cable

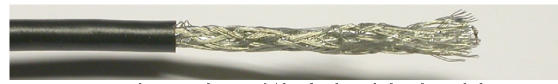

Fig 6b - A braid/foil shielded cable

The drain wire in Fig 4 is more closely coupled to the purple wire than the white one because it is physically closer to it along the length of the cable, and has the same twist ratio as the signal conductors. The resistance of the drain wire is typically an order of magnitude less than the resistance of the shield, so nearly all of the shield current flows on the drain wire. This characterization is true from dc to at least 5 MHz. Above that frequency, skin effect increasingly causes current to be equally distributed around the foil, and by 20 MHz, the inequality has disappeared.

Fig 5e is the equivalent circuit. M_{1-2}, M_{1-S}, and M_{2-S}, are the respective mutual inductances between the signal pair, the #1 conductor and the shield, and the #2 conductor and the shield. L_S and L/2 are approximately equal, but not precisely equal. So one of the effects is that for differential signals, the mutual inductance partially cancels the cable inductances, and causes the total inductance of the cable to be much smaller than for two wires in an open loop of the same length. We'll look at this in greater detail later on. But any imbalance in the mutual inductances between the shield and the two signal conductors will cause shield current to be converted to a differential voltage. Thanks to Neil Muncy, noise coupled by this mechanism is commonly known as ***"shield-current-induced noise" (SCIN)***.

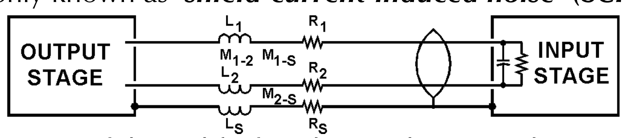

Fig 5e – A circuit model simplified to show only SCIN. The mutual inductances M_{1-2}, M_{1-S}, M_{2-S}, describe the coupling between the three conductors.

M_{1-2}, M_{1-S}, and M_{2-S} depend upon the physical construction of the cable. If M_{1-S} and M_{2-S} are equal (that is, the shield couples equally to the two signal conductors), and if L_1 and L_2 are equal, there will be no imbalance and thus no SCIN. For cables with a drain wire, M_{1-S} and M_{2-S} are not equal. The result of this imbalance is that any current flowing on the shield will be converted to a differential voltage on the signal pair. The strength of the conversion will be directly proportional to frequency up to about 5 MHz.

SCIN can still occur in cables that have only a braid shield, but the effect is far less pronounced, and is attributable to small variations in cable manufacturing tolerances. Below 5 MHz, this common-mode to differential-mode conversion is roughly 30 dB greater in foil/drain cables than in braid-shielded cables. SCIN is a major contributor to RF interference to audio systems, especially from AM broadcast stations and industrial noise. SCIN is also far less in a cable that has a foil/braid shield but no drain wire (Fig 6b, and cable BF in Fig 7a). This construction doesn't even require that the braid be very robust for the combination to be an effective shield.

Over the past decade, it has become fashionable to manufacture portable audio cables with a braid shield and a drain wire. The drain wire makes the cable easier to terminate, but it significantly increases SCIN below about 250 kHz as compared to a braid shield without the drain wire. ***In terms of EMC, drain wires are a really bad idea, and should be avoided!***

As part of some EMC research a few years ago, I evaluated the relative SCIN performance of foil/drain, foil/braid, braid, and braid/drain shields up to at least 30 MHz. In general, the braid shields were superior below about 15 MHz, the foil shields were superior above 30 MHz, and a foil/braid shield was effective throughout the RF spectrum. Unfortunately, no braid-shielded or foil/braid-shielded cables are currently available in North America that also conform to NEC fire safety codes for permanent installation. Figure 7a shows the SCIN performance of typical foil/drain shielded cables. Some of these cables are of significantly better manufacturing quality than others, but their SCIN performance is all equally poor. Cable BF has a foil/braid shield, and is of relatively primitive build quality, but it's SCIN performance is quite good!

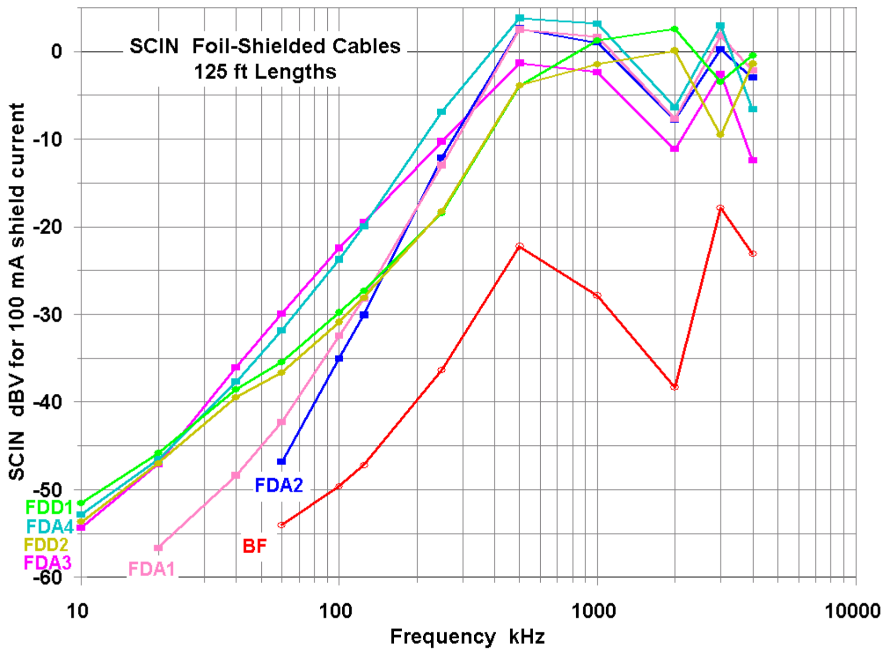

Fig 7a – SCIN in foil/drain shielded cables, and cable BF, which has a foil/braid shield.

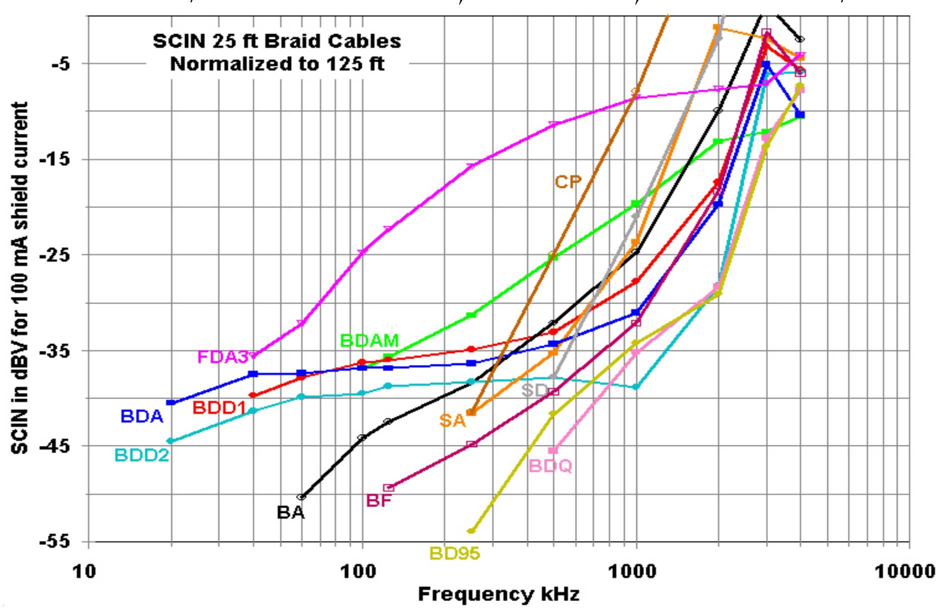

Fig 7b – SCIN performance of Braid-Shielded cables, except that FDA3has a foil/drain shield.

Fig 7b compares SCIN in cables having various shield constructions. Cables SA and SD have spiral shields. Cables BDxx have a braid shield and a drain wire, cable Cable CP has a conductive plastic shield and drain wire. BA is Belden 8412, BF has the foil/braid shield. FDA3, with a foil/drain shield, is shown for comparison. Note that below about 500 kHz the drain wire seriously degrades the SCIN performance of braid shields, while above that frequency it doesn't matter. BDQ has quad construction. BDAM is a high quality miniature cable used in snakes and is the same size as FDA3, but has a braid shield with a drain wire. Simply replacing the foil with the braid improves SCIN performance by 10-15 dB! Because nearly all RF detection follows square law, this results in RF interference being reduced by 20-30 dB!

Structured data cables CATn cables are four-pair cables, mostly unshielded, with each pair very tightly twisted together. To further minimize crosstalk, each pair has a different twist ratio (lay). Although capacitance between conductors of a pair is not a specified quantity, 12-14 pF/ft is typical. Cables of ordinary quality exhibit capacitance imbalances on the order of 350 pF/100 m (about 9% imbalance); premium cables, such as Belden's Mediatwist, are specified for 66 pF/100m (1.6% imbalance).

Fig 8 shows attenuation vs. frequency for three structured cables, two AES3 cables, and a low-loss coax optimized for MATV and CATV use. Cable AES3 #1 is a popular portable cable, while AES3#2 is a much larger cable optimized for low loss. The data shows that CAT5 and higher cables outperform the "standard" AES3 cable by a wide margin, while the premium, low loss AES3 cable is as good as the best structured cable (but at a considerable cost premium).

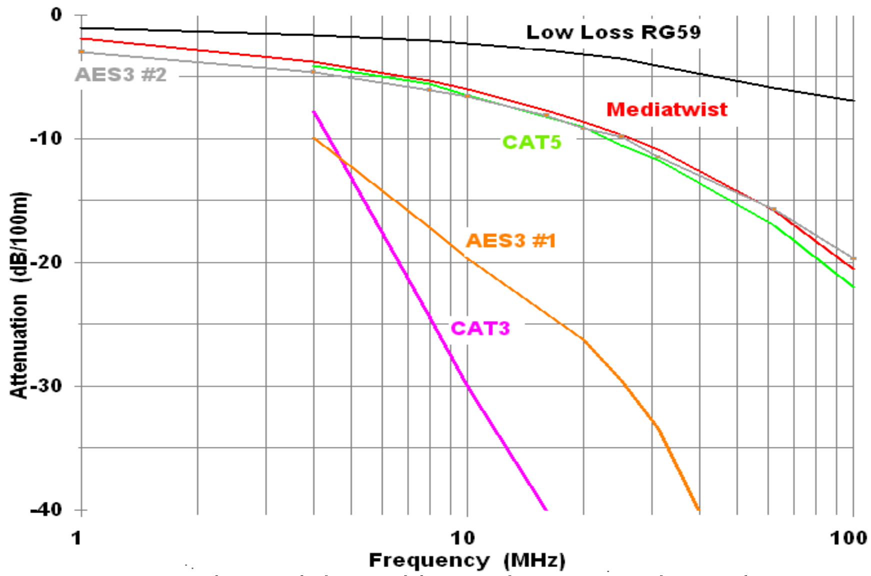

Fig 8 – Attenuation of typical data cables, and a very good coax for comparison

Notice also that in terms of ***attenuation***, there is little difference between the premium quality Mediatwist and a good CAT5 cable. Indeed, the key differences between Mediatwist and the CAT5 cable are the specifications for crosstalk, which are nearly 10 dB better, and relate directly to the improved symmetry of the cable. Whether handling data or analog audio, cable symmetry and reduced crosstalk both translate to greater rejection of noise!

Transmitting Digital Audio (AES3) Digital audio transported per AES3 includes components at frequencies up to at least 128 times the sampling frequency (for example, 6 MHz for 48 kHz sampling rate, 24 MHz for 192 kHz). At these frequencies, even a very short length of cable will act as a ***transmission line***. The longer the cable, and the higher the sampling frequency, the more pronounced those effects will be.

The AES3 standard for digital audio calls for cable that is balanced, shielded, and has a characteristic impedance of 110 ohms, ±20%. That impedance specification is really too broad, and no loss characteristic is specified for the cable. In 2003, informative notes were added to AES3 stating that holding tighter impedance tolerance on the cable, the driver stage, and the receiver, and using cable that had lower loss at high frequencies would improve the reliability of transmission over longer cables and would reduce both radiated and received noise.

Impedance matching In general, impedance matching is of critical importance in 1) the transmission of high speed data; and 2) the transmission of video. Reflections will be created by a mismatch, and will distort both digital and video waveforms. For there to be no reflections on a transmission line, there must be only one driver, that driver must have an output impedance equal to $\mathbf{Z_O}$, there must be only one load, also equal to $\mathbf{Z_O}$.

It is possible to have more than one ***receiver*** connected to a transmission line, and even to connect it anywhere along the line – ***the only requirement is that only the receiver at the end of the line can load the line***. Most of us are familiar with video systems where a single source drives multiple monitors with the feed to the second monitor "looping through" the first monitor. That first monitor must "bridge" the line – that is, there must be no terminating resistor on the line at that point. Most professional video monitors have 75Ω resistors that can be switched in and out, depending on whether the monitor is the first monitor or the last monitor on the line.

Exactly the same practice will work with digital audio, and although it is not provided for in the AES3 standard, several manufacturers make digital receivers with ***bridging*** inputs. A ***bridging*** load is one that does not draw enough current from the line to affect the line. In general, a ***bridging*** load should be at least ten times the impedance of the system that it bridges, and should have very little stray capacitance. With AES3 signals, the cable impedance should be 110 ohms, and so should the load impedance. The ***bridging*** impedance should be at least 1,100 ohms, and higher is better.

Older style cables designed for analog audio are generally unsuitable for the transport of digital audio. They may work for short distances, but fail when used for long runs. This is true for two reasons. First, the characteristic impedance of these older analog cables is generally much lower than 110 ohms, and is not well controlled. Second, analog cables tend to have excessive loss at high frequencies, causing the line to act as a low pass filter and round off the pulses that make up the digital signal.

It is also quite important that all cables carrying a digital audio signal have the same impedance. Any non-uniformity in the impedance of the transmission line will also distort the digital pulse. Maintaining a close tolerance on the 110 ohm cable impedance and using cable that has low loss at high frequencies will improve the reliability of digital audio transmission over greater distances and at higher data rates.

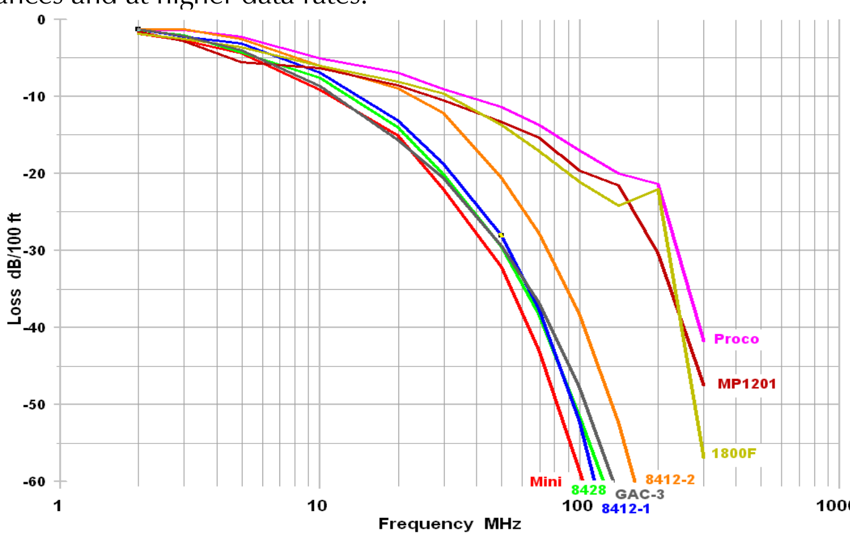

Fig 9 – Loss of typical digital (upper 3 curves) and analog (lower 5 curves) cables

75Ω or 50Ω BNC Connectors? This transmission line impedance issue continues to cause confusion among those who don't understand it, and insist on following simple rules like "always match impedance, no matter what." An example is a recent thread on the SynAudCon list concerning the need for special 75 ohm BNC connectors to match 75 ohm cables. Standard BNC connectors are designed with a 50 ohm impedance. At first glance it should be obvious that 75 ohm connectors should be used on 75 ohm cable, even if they cost four times as much – until you do the math that tells you that from a transmission line perspective, it doesn't matter! Let's do that.

Impedance mismatch matters ONLY if it causes a perturbation in the line that is 1) a significantly high VSWR and 2) is a significant fraction of a quarter wave at the frequency of the signal. A 75 ohm connector on a 50 ohm line represents a 1.5:1 VSWR, which is negligible.

So is a 50 ohm connector on a 75 ohm line – that's 1.33:1 VSWR. The "transmission line part" of a 75 ohm connector is on the order of 1/2" long. That's roughly 1/16 ns, or 1/20 wavelength at 600 MHz. At 1.2 GHz it's 1/10 wavelength.

Now, it <u>is</u> important to use a connector that is a good physical fit for the cable you're using, and that's the important issue. The "right" BNC connector is one that physically fits the cable, provides the greatest mechanical reliability, and can be quickly installed with the available tools!

Bottom line – if I were doing high speed data or SDI video, or if I were doing a <u>lot</u> of patching in a <u>very</u> high resolution video system (super high end commercial theater, very big screen), I would pay for 75 ohm connectors. For a less demanding application, I would be a fool to do so. Instead, I would spend my money on a connector that was physically more rugged, easier/faster to terminate reliably, and on any specialized tools needed to accomplish that.

Cable handling noise At least two mechanisms cause noise to be generated by the mechanical flexing of cable. Noise can be generated by the motion of, or mechanical stress on, a cable when phantom voltage is present. The phantom voltage charges the capacitance of the cable; if the conductors within the cable move or are stressed such that the capacitance changes, a noise current must flow to charge or discharge the change in capacitance. These voltages will be heard as impulse noise, and can be rather strong.

Triboelectric noise: Static electrical charges can be generated on insulation materials within cables as the cable is moved. A charge is built up by this motion, then discharged as the breakdown voltage is reached, and the process repeats. This is similar to the mechanism that creates a charge and an spark as you walk across a carpet, and is called the **triboelectric** effect. Triboelectric noise is minimized by careful selection of insulating materials, shield construction, the use of fillers, and the optimization of other elements of cable construction that minimizes the generation of static when the cable is flexed.

RS-232 Interconnections – A "band-aid" for an interesting special case This very old standard is not a very good one, and for several reasons. First, it is used for the transmission of data, sometimes at moderately high speed, but with no attention to the properties of the cables involved. Second, it fails to address any transmission line behavior of the cable – that is, it treats the interconnection as a lumped parameter circuit. Third, the circuits are unbalanced, so noise currents on the common conductor will pollute the data stream. Fourth, a pin 1 problem is often built into the interface!

The RS-232 standard provides for several protocols, but most equipment uses only one of those protocols. Thus, most applications of the RS-232 interface actually use only two or three circuits. In these situations, ordinary CAT5 cable can work extremely well. Here's how.

First, identify which circuits are actually used by the equipment you are interconnecting. Then, use one <u>pair</u> of the CAT5 cable for each one of those circuits. Let's say that our equipment uses DCD (pin 1), sends and receives data on pins 2 and 3, and uses the RTS line (pin 7). In this example, use the orange wire to connect pin 2 to pin 2, connecting the orange/white to the DB9 shell. Use the blue wire to connect pin 3 to pin 3, connecting the blue/white to the DB9 shell. Use the green wire to connect the pin 1's, connecting the green/white wire to the DB9 shell. Use the brown wire for pin 7, connecting the brown/white wire to the DB9 shell. And finally, before using the shell rather than pin 5, make sure that there is a shell at the equipment, and that it contacts the shielding enclosure!

What have we accomplished here? First, by using CAT5 cable, we have used the lowest capacitance cable available at relatively low cost. This minimizes the rolloff of the data pulses by cable capacitance, which is one of two major factors limiting the tradeoff between cable length and high data rates. In other words, we can use higher data rates for longer distances!

Second, each circuit is now carried on a tightly twisted pair. This minimizes the magnetic component of the pickup of high frequency noise, because each twisted pair begins to act as a common mode choke above about 10 kHz. It also reduces the magnetic component crosstalk between the RS-232 circuits, both because the pairs are twisted, and because each pair within a CAT5 cables is twisted at a different rate. Third, CAT5 cable has much lower loss at high frequencies, which further improves the distance vs. data rate compromise. Fourth, terminating signal returns to the DB9 shell works around the pin 1 problems built into many RS-232 interfaces. By using this scheme, an RS-232 interconnect can be made to work at significantly higher data rates, or on significantly longer cables, or both.

And finally, since four conductors are in parallel for the returns of the unbalanced circuits, power-related voltage drops in the return circuit due to ground loops are reduced by 12 dB. Note that the magnitude of these currents can be minimized by floating one end (as, for example, when a laptop running on battery power programs a DSP unit), and by the use of an isolated ground system. Either of these techniques will further reduce the noise on the return conductor(s).

Shielding

How shielding works A ***shield*** is defined as a conductive barrier or enclosure interposed between two regions of space, with the intent of preventing a field in one region from reaching the other region. A shield attenuates (weakens) the field by four fundamental mechanisms.

1) The field is **absorbed** within the conductive (resistive) shield as current flows through a resistance and is converted to heat.
2) The field is **reflected** by the discontinuity at the boundary between the impedance of the wave before it enters the shield, and the impedance of the shield itself. The field will also be reflected back into the shield when it travels through the shield and encounters the second boundary – that between the shield and the protected region.
3) The field is **diverted** around the protected region by a ***low reluctance*** *path. [Reluctance can be thought of as the equivalent of resistance for magnetic fields – in other words, is it's the reciprocal of permeability.]* ***Diversion is the magnetic field equivalent of a Faraday shield.*** A mu-metal shield around an audio transformer is an example of shielding by ***diversion***. To compute shielding effectiveness, ***diversion*** can be analyzed as ***absorption*** (that is, by calculating the skin depth of the path). The shielding provided by steel conduit can also be analyzed as ***diversion***.
4) A **shorted turn** around a source can "short circuit" a magnetic field, producing a large current in the shorted turn, which then produces flux of the opposite polarity that "bucks" the original flux. A large copper strap that surrounds the windings of a power transformer is an example of a shorted turn. Most of the flux should be contained within the transformer core, but any leakage flux will induce a current in the shorted turn (which is outside the core). That current produces a "bucking" field, canceling out the original field, and preventing if from causing interference to nearby equipment.

Absorption, ***diversion***, and ***reflection*** are completely independent mechanisms, and the total attenuation in any shield will be equal to the sum of the three. Depending on conditions (thickness and geometry of the shield(s) and the nature of the interfering signal), both ***absorption (or diversion)*** and ***reflection*** can be significant. Under other conditions, one or both may be of use, and under some conditions none may be of much help. Also, any attenuation of the fields by a shield will be additive to any rejection of the fields that result from twisting.

As we will soon learn, nearly all low frequency noise comes from <u>**near-field**</u> **magnetic sources** – power equipment, power wiring, and motors. We have four principal weapons against these sources – 1) **absorption** or **diversion** of the field by ferrous conduit and enclosures, 2) tightly twisting signal pairs, 3) reducing the loop area of the signal path, and 4) increasing the distance between the magnetic source and the vulnerable circuit. We will also learn that #2 and #3 are related.

Conduit and shielding The most effective way to provide **magnetic shielding** for audio cables is to enclose them within ferrous conduit. We'll begin our study of shielding with the **absorption (or diversion) loss** provided by conduit.

While **unbalanced** wiring inside ferrous conduit will be shielded, considerable shield current can still flow due to differing "ground" potentials at either end of the run. In addition, noise can be magnetically coupled into the circuit because part of the signal return path is through equipment grounds (large loop area). Both of these mechanisms will couple noise into the signal circuit by virtue of the IR/IZ drop in the shield!

Absorption loss occurs in the form of a simple exponential decay of field strength as the field passes through the shield – that is, for each mm of travel within the shield, some fraction of its total energy is lost in the form of heat. The equation for exponential decay is exactly the same as for the discharge of a capacitor, and for the reverberation in an ideal room. The **skin depth** of a material is defined as the thickness through which a field must travel to be attenuated by 1/e (where e is 2.718, the base of natural logs). 1/e is 0.37, or a loss of 8.7 dB. Thus, a convenient "rule of thumb" that **a shield absorbs 8.7 dB per skin depth**. In addition, the strength of a magnetic field that is being **diverted** by a higher permeability shield will also decay by 8.7 dB/skin depth within the cross-section of that field.

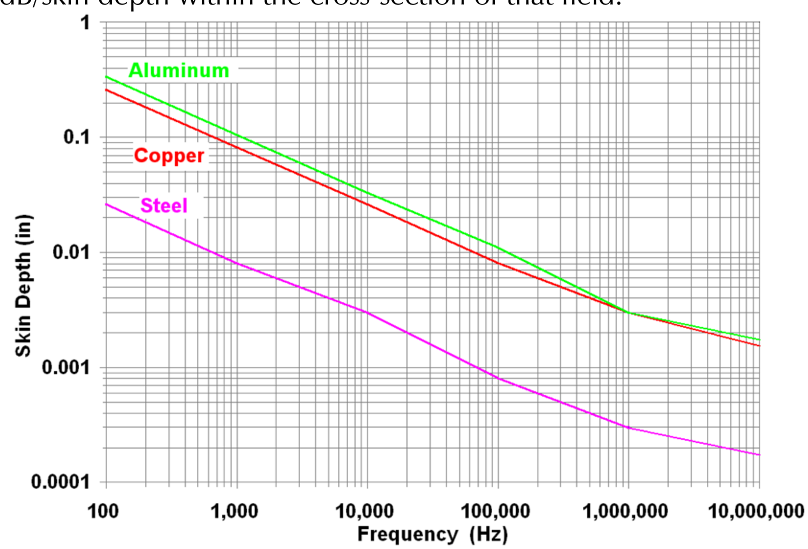

Fig 11 – Skin Depth Determines Shielding Effectiveness

Fig 11 shows the skin depth as a function of frequency for copper, aluminum, and steel, and Table 2 lists wall thickness for conduits of standard trade size. The thickness of 2" EMT is a bit less than 2 skin depths at 60 Hz, so its absorption loss is about 16 dB, while at 180 Hz, it is more than 2.5 skin depths, so its absorption loss approaches 22 dB. The thickness of 2" rigid steel conduit is about 4 skin depths at 60 Hz, so the absorption loss is about 33 dB, and about 52 dB at 180 Hz.

	Wall thickness (inch)	
Trade Size	**EMT**	**Rigid steel**
½ inch	.045	.11
¾ inch	.05	.115
1 inch	.055	.135
1¼ inch	.065	.14
1½ inch	.065	.145
2 inch	.065	.15

Table 2 – Wall thickness for conduits in North America

Fig 12, the predicted absorption losses (shielding) for these conduits, clearly shows the benefit of rigid steel. This can be a very critical factor if audio system wiring will be subjected to very strong magnetic fields, such as those generated by large motors, transformers, feeder cables (that is, the large cables feeding transformers and distribution panels), and variable speed motor controllers.

Saturation Magnetic materials are non-linear – that is, their permeability varies with the strength of the magnetizing field. At low field strengths, the materials are relatively linear – an increase in the field causes an approximately linear increase in the flux. But at very high field strengths, the material saturates – it is carrying all the flux that it can, and a further increase in the field won't increase the flux. In transformers, we are careful to operate magnetic materials in their linear region to avoid distortion.

While we can tolerate some distortion in shielding applications, it is critical that we avoid saturation. That's because flux that the shielding material can't carry (due to saturation) goes right through it to the protected region! In other words, when the shield is saturated, it provides far less shielding. A magnetic path with greater cross sectional area can carry more flux without saturation. This is another good reason for using rigid steel conduit (rather than EMT) near strong magnetic fields. Some materials, like mumetals, are far more prone to saturation than others. One solution, suggested by Manquen, is to nest a mumetal shield with a mild steel shield.

Variable speed motor controllers are a particularly nasty source of magnetic fields, producing square waves by chopping the 60 Hz sine wave at a frequency on the order of 10 kHz, with strong harmonic content well into the hundreds of kHz. If the motor controller is driven by an isolation transformer immediately adjacent to the motor, and with the secondary and Faraday shield(s) both grounded very close to the motor frame, these fields will be contained in the very small loop between the transformer and the motor. If, however, there is no isolation transformer, or if the secondary is grounded at some distance from the motor, these fields will cover a very large loop area. In the latter condition, the audio system and its wiring can require a very high degree of magnetic field shielding to prevent these fields from being detected. Currents are also established to ground via the capacitance between the motor windings and the motor frame.

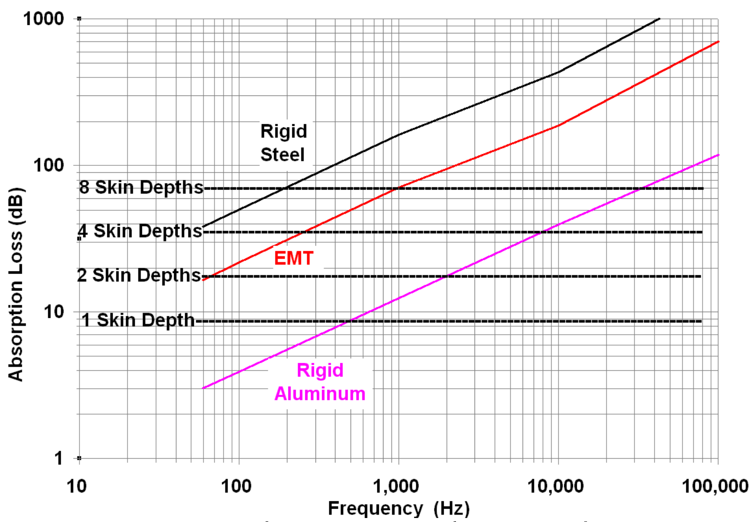

Fig 12 – Absorption Loss for 2" Conduit

Wave impedance The concept of the wave impedance is critical to understanding how and when ***reflection loss*** will be significant in a given situation. When the ***source*** of an interfering field is at considerable distance from our wiring (that is, when we are in the ***far field***), the wave impedance will be 377 ohms, the impedance of free space. In the far field, the magnetic field and electric field are of equal importance. In general, relatively thin copper and aluminum shields can be quite effective in the far field.

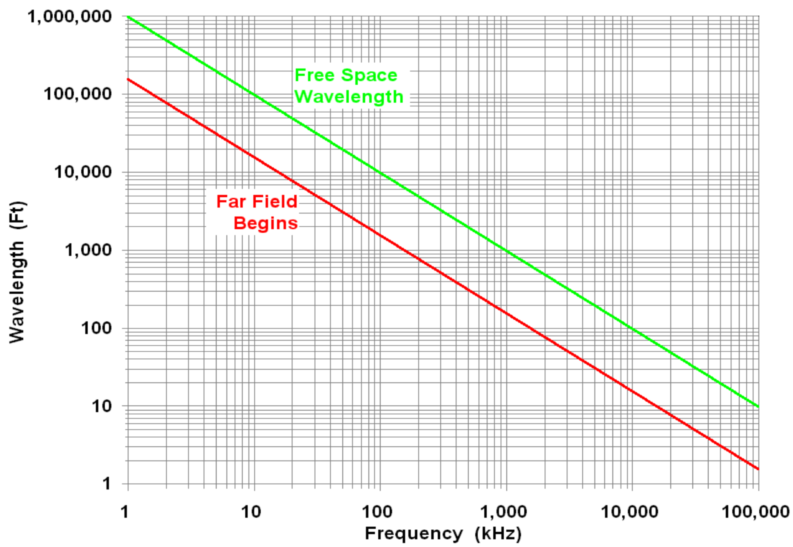

Fig 13 – Wavelength (λ) in Free Space; wavelength within a transmission line is 15-30% shorter than this length, but for the cable acting as an antenna it is only 3% shorter

Reflection Loss: When a wave enters a shield, the **reflection loss** R is equal to

$$R = 20 \log [(Z_W / 4 Z_S) \cos \phi] \text{ dB}$$

where Z_W is the impedance of the wave prior to entering the shield, Z_S is the impedance of the shield, and ϕ is the angle between the shield and the source of the field. Z_S is found by

$$Z_S = \sqrt{\frac{2\pi f \mu}{\sigma}}$$

where μ is the ***permeability*** of the shield and σ is its ***conductivity***.

The shield impedance for any conductor works out to be a tiny fraction of an ohm. Fig 13 is a plot of the Reflection Loss (R) for commonly used shielding materials.

Let's take the simple case where our interfering signal is an AM radio station a mile away transmitting on 1 MHz. We are in the far field (see Fig 13), so it is a plane wave, with the magnetic and electric field in balance, and the wave impedance is 377 ohms. Steel conduit would provide a reflection loss of 70 dB; copper or aluminum would provide about 110 dB of shielding.

When our wiring is in the near field of a current source, the magnetic field is much stronger than the electric field, so the wave impedance will be quite low and there is little if any reflection loss. A quick examination of Fig 13 shows that the far field begins at about 400 **miles** from a 60 Hz source, and about 1,500 ft from a 100 kHz source. Thus, in the real world, we are always in the very near field of the power system equipment, so the wave impedance for these fields is quite low!

On the other hand, when our audio system is in the near field of an interference source that has high voltage but low current, the wave impedance will be quite high. In this situation, the electric field predominates, and the dominant coupling mechanism will be capacitive. This allows a thin copper or aluminum shield to be quite effective.

Most fields start out with either the magnetic or electric field predominant – that is, they are established mainly by current or mainly by voltage. As the field extends outward through space, the dominant field will decay far more rapidly than the other, and eventually they will reach equilibrium at the 377 ohm impedance of free space. In the near field, the dominant field typically falls off with the cube of the distance (that is, $1/r^3$), while the other field falls off in proportion to the square of distance ($1/r^2$). In the far field, they both decay in proportion to the distance ($1/r$).

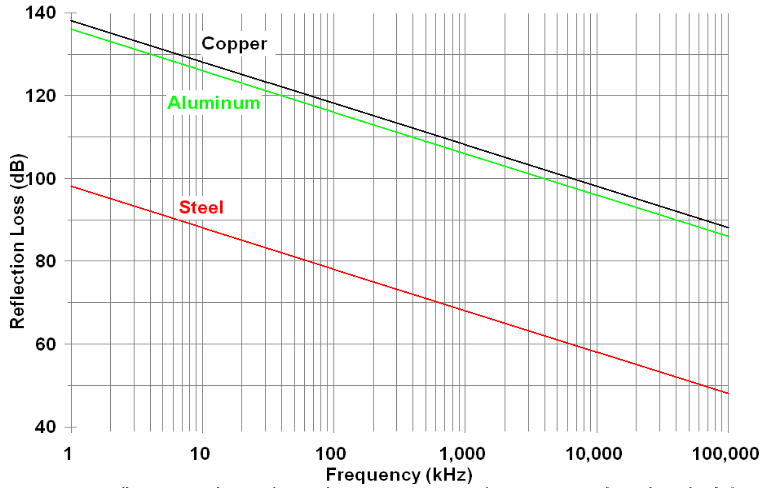

Fig 14 – Reflection loss for plane waves (that is, in the far field of the source). Reflection loss for near fields will be far less.

Total shielding effectiveness **S = A + R + B** where **A** is the absorption loss, **R** is the reflection loss, and **B** is a correction factor for the reflection loss for very thin shields (that is, much less than a skin depth). In this equation, **A** is used to account for both **absorption** and the decay of the field through a **diverting** barrier. The correction factor is a negative number of dB, and in general should be applied to near magnetic fields. The equation for **B** is

$$\mathbf{B} = 20 \log \left(1 - e^{-2t/\delta}\right)$$ where t/δ is the thickness in skin depths

The mechanism described by the "***B-factor***" is interesting. Consider a wave traversing the thickness of the shield. At the first boundary, some of it is reflected, while the remainder passes into the shield. Some of that will be absorbed, and some will be reflected by the second boundary – the other side of the shield and the air. That reflected energy will travel back through the shield and be absorbed, then reflected at the first boundary, and so on. And, each time the field hits the inner boundary, some will travel into the shielded enclosure. This effect happens with <u>all</u> shields, but in thick shields more of the wave is **absorbed** on each pass through the shield. The "***B-factor***" is large in shields that are a small fraction of a skin depth, because there is little attenuation of each reflection. The ***B-factor*** is small and can be ignored when the shield is a skin depth or greater.

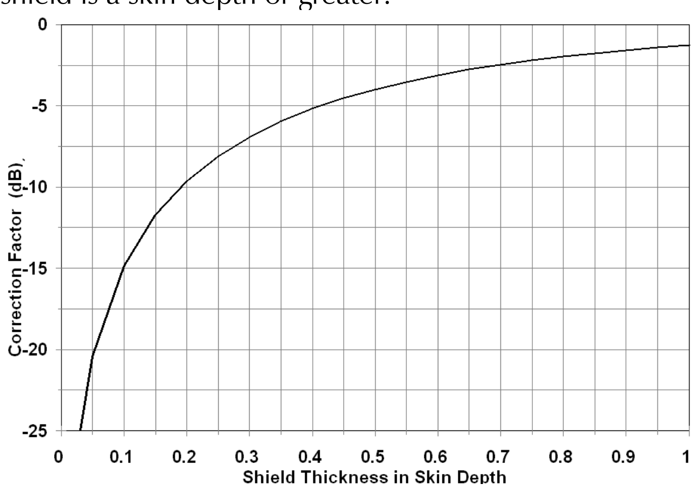

Fig 14 – The ***B-factor*** accounts for multiple reflections in thin shields

Cable shields are generally ineffective against magnetic fields, and for a combination of reasons. First, cable shields are far too thin to have measurable ***absorption loss***. Fig 11 shows that one ***skin depth*** in copper or aluminum is about 3/8 inch at 60 Hz, and about 1/32" at 10 kHz, while the thickness of a cable shield is rarely more than a few thousandths of an inch.

Because the shield is thin, the ***B-factor*** will be large and greatly reduce the ***reflection loss***. ***Reflection loss*** will be also be small if our wiring is in the ***near field*** because the ***wave impedance*** is low, since our equipment and wiring is nearly always in the very near field of the magnetic sources we are trying to shield against!

Loop Inductance and Loop Area in Cables A piece of wire has inductance along its length. Remember from Fig 5e that when two or more wires run in close proximity to each other, they will have approximately equal inductance along their length, but there will also be **mutual** inductance between them (see Fig 5e). If the wires are very close together, they will be very tightly coupled – in a tightly twisted pair, nearly 70% the magnetic flux produced by current in either wire will encircle the other wire. In this condition, the mutual inductance partially cancels the inductance of the wires themselves, typically reducing the **total loop** inductance ($L_1 + L_2 - M_{1\text{-}2} - M_{2\text{-}1}$) to about one-third of what it would be if wires of the same length were in an open loop.

To understand the practical importance of this fact, consider a tightly coupled cable pair (it could be either a twisted pair or a coaxial cable) running between two points. In the cable, one of the conductors carries the "high" side of an unbalanced signal, and the other conductor carries the return. Now, let's "ground" the return at each end. Fig 15 shows our circuit. At very low frequencies, inductive reactance is too small to matter, so the "signal return" current will divide between the return conductor of the cable and other "ground" paths – for example, the equipment ground wiring between the two ends of the cable – based on the resistances of the two paths. The lowest resistance path will carry the most current.

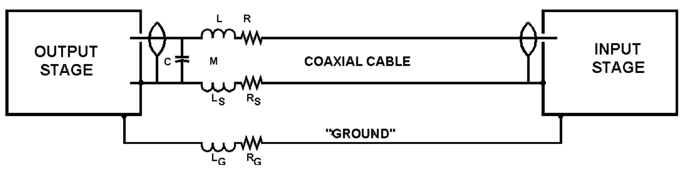

Fig 15 – Two tightly coupled conductors connecting "grounded" equipment

As frequency increases, the inductance of the path through the equipment grounds will become much larger than the resistances, but because the **loop inductance** is small, much more of the return current now flows through the cable pair. This happens whether the cable is a twisted pair, a parallel pair, or a coaxial cable, but it is most pronounced as the loop area, and thus the loop inductance, is reduced. And as we will learn, it is far more pronounced in coaxial cables.

Mutual inductance and noise rejection That mutual inductance is doing something else quite important – it makes the cable a tightly coupled transformer with an approximate turns ratio of 1:1 (1:1:1 if it is shielded pair) and a coupling coefficient of about 0.7. Transformer action will thus cause any noise current flowing in one conductor to induce a copy of the noise into the other conductor. A differential receiver connected across the signal pair will then see the difference between two nearly equal copies of the noise, and reject it. Thus, a paired cable tends to act as a common mode choke to external signals (noise)!

Coaxial cables We will study this effect first in coaxial cables. It can be shown (ref 1, 2) that for an ideal shield (that is, one that is cylindrical and allows current flow to be uniformly distributed around it), the mutual inductance will be equal to the shield inductance – that is, the coupling coefficient is unity – and that this mutual coupling will begin to work for us at a frequency $\mathbf{f_S}$, called the shield cutoff frequency. Below $\mathbf{f_S}$, the inductance is small, and most of the noise voltage appears across the resistance of the shield. Above $\mathbf{f_S}$, more of the noise voltage appears across the shield's inductance, and it is the voltage across the inductance that couples to the center conductor. The references show that $\mathbf{f_S}$ is $R_S / 2\pi L_S$, where R_S is the shield resistance and L_S is the inductance of the shield. The inductance of a straight wire is approximately $L_S = 6 \times 10^{-7}$ H/ft, so the equation simplifies to

$\mathbf{f_S = 0.265\ R_S\ kHz}$ where R_S is the shield resistance in ohms/thousand feet.

Typical values of R_S for a high quality braid shield are on the order of 5 ohms/thousand feet, which results in f_S = 1.3 kHz. Above the s***hield cutoff frequency***, more voltage is across the inductance than across the resistance, so the inductive coupling between the two conductors increases. In other words, the shield cutoff frequency $\mathbf{f_S}$, is where the cable just begins to start work as a common mode choke!

For the coaxial cable analysis to be correct, **shield current must be uniformly distributed on the shield.** If current is not uniform throughout the shield, the field inside cable will not be zero, and magnetic field rejection provided by reduced loop area will be further degraded.

It is clear from the equation that the effectiveness of the shield against magnetic fields is inversely proportional to the shield resistance – **to function as a common mode choke at low frequencies, the resistance of the shield must be small**. Foil/drain shielded cable has far greater resistance than braid-shielded cable, and, as we have learned in our study of SCIN, shield current is far from uniform. These two properties combine to make a foil shield even more ineffective against magnetic fields.

It is important to realize that the cable shield is not really providing magnetic <u>shielding</u> – rather, the cable is <u>rejecting</u> the magnetic field because the coupling between the shield and the center conductor causes any voltage induced in the center conductor to be cancelled by an equal voltage coupled from current induced in the shield! In effect, the cable acts as a common mode choke. As we have already learned, cable shields do not provide magnetic shielding!

Let's plug in some more numbers. The DC resistance of typical foil/drain cables is on the order of 20 ohms/1000 ft, and the drain wire carries nearly all of the current below about 250 kHz, because the resistance of the foil itself is typically 50-100 ohms/100 ft. In an ideal foil shielded cable (that is, the drain carries <u>no</u> current), $\mathbf{f_S}$ would be on the order of 12 – 25 kHz. The presence of the drain wire makes matters considerably worse because it makes the shield current non-uniform.

The same mechanism works for any transmission line, shielded or unshielded, balanced or unbalanced, but the numbers are different. In fact, a better name for $\mathbf{f_S}$ might be the ***<u>cable</u> cutoff frequency!*** Indeed, this is exactly what we're talking about with respect to the 20 kHz transition to common mode choke behavior for CAT5 cable (and other twisted pairs). In other words, it is the low frequency limit for the cable's ability to reject magnetic coupling by virtue of its parallel construction and small loop area.

One important difference between coaxial cables and unshielded twisted pair cables is that the mutual inductance between the coaxial shield and the center conductor is equal to the shield inductance, while the mutual inductance between conductors of a paired cable varies from about 50% of the inductance of one conductor for "zip cord" to about 70% of that inductance for the most tightly twisted pairs. A second important difference between coaxial cables and unshielded twisted pairs is that the cable cutoff frequency, $\mathbf{f_S}$, is much higher for twisted pairs because their series resistance is much greater than for a good coaxial shield. These differences combine to make a twisted pair somewhat less effective as a common mode choke than a coaxial cable – it begins working at a significantly higher frequency ($\mathbf{f_S}$), and the choking action is less when it does begin to "kick in" because of the 70% coupling coefficient.

Shields of balanced cables act very differently from shields of coaxial cables with respect to magnetic field rejection. First, it is the close coupling <u>between the two signal conductors</u> that make up a signal pair that reduces the loop area. Second, the <u>twisting</u> of the two signal conductors improves the balance of the coupling of external fields, and can further reduce the loop area. ***The cable shield is not a part of either mechanism!***

Remember also from our discussion of SCIN that the presence of a drain wire in many braid-shielded cables will also act as exactly the opposite of a shield – that is, rather than <u>reducing</u> the coupling of noise, it <u>causes</u> noise coupling. This is clearly seen in the SCIN data (Fig 7). In fact, in a cable with a drain wire, there are two shield conductors at low frequencies (below about 1 MHz) – the foil (or braid) is one of them, and the drain wire is the other! Shield current will divide between the two shields based upon the ratio of their resistance and influence of skin effect at the frequency(ies) of the current.

As frequency increases, three other effects show up. First, skin effect causes <u>noise</u> current to flow on the <u>outside</u> of the shield and <u>signal</u> current to flow on the <u>inside</u> of the shield. This tends to increase the effectiveness of the shield with increasing frequency. For typical coaxial cables, this phenomena begins between 500 kHz – 1 MHz. Second, the drain wire essentially disappears, and all the current flows on the foil or drain.

The third effect shows up as frequency continues to increase, and <u>degrades</u> the effectiveness of the shield against magnetic fields. This loss of effectiveness occurs when the frequency is high enough that the small holes in the braid distort the uniformity of shield current. For shields having as little as 85% coverage, the degradation begins as low as 500 kHz, canceling the improvement provided by skin effect. For shields with 98% coverage, the degradation sets in around 2 MHz. In these better-shielded cables, there is a narrow region of 2-3 octaves around 2 MHz where the shield has modest effect (~10 dB) against magnetic fields. The braid of typical portable microphone cables has 85% coverage.

This degradation due to non-uniformity of the braid can be reduced by the use of combination shields – that is, both foil and braid. This is a happy result – a combination foil/braid shield also minimizes SCIN! The unhappy part is that no foil/braid-shielded balanced audio cables are currently sold in North America.

Magnetic field rejection and shield current It is important to remember that the magnetic field rejection provided by the shield of a coaxial cable is the result of the ***mutual inductance between the shield and the center conductor(s)***, and the reduced loop area of the path through the coaxial cable as opposed to a path through "ground." In other words, magnetic field rejection occurs because the cable acts as a common mode choke above the shield cutoff frequency, $\mathbf{f_S}$.

For there to be any magnetic field rejection, shield current must flow ***at the frequency of the interfering signal***. For a shield that is less than about 1/8 wavelength, this requires that the shield be connected at both ends. For a shield that is more than a wavelength, antenna action can cause current with or without the connection.

Ott repeatedly takes pains to observe that the benefits of connecting a coaxial shield at both ends of the cable are often far outweighed by power-related noise currents coupled onto the shield by the resulting ground loop. A common solution is to isolate equipment from connections to random "grounds," powering them from isolated ground outlets. Even this measure may not eliminate noise that is magnetically coupled into a loop that includes the isolated ground conductor and the coaxial shield.

Even with this connection, magnetic shielding effectiveness is subject to the limitations already noted. And as we have learned, shield current is often quite problematic with balanced connections, exciting both SCIN and pin 1 problems, resulting in noise coupling that is often of far greater magnitude than the limited magnetic shielding that would be provided if the shield were connected. Thus, it is important to prevent the flow of shield current below the lowest frequency at which the shield can be effective. To prevent the flow of audio frequency shield current, the connection at the receiving end should be through a capacitor having a low value of stray inductance. A value on the order of 10-50 nF is generally suitable. Whitlock has shown that the shield should always have a dc connection at the sending end.

Electric Field Shielding The primary benefit of a cable shield is that it provides good shielding against electric fields. To be effective, the shield must be connected at one end to the signal reference. [Whitlock has shown that the best point of connection is to the shielding enclosure at the sending end – if the shield is connected only at the receiving end, common mode rejection will be degraded.] The shield prevents coupling of the electric field to the signal conductors by shorting the field to the signal reference. If the length of the cable is greater than $\lambda/20$ at the frequency of some interference, a connection at the receiving end is also required. The capacitor noted above provides that connection.

Ott also observes, "Even if the shield is grounded at only one end, shield currents may flow due to capacitive coupling to the shield. ***Therefore, for maximum noise protection at low frequencies, the shield should not be one of the signal conductors, and one end of the circuit must be isolated from ground.***" (Emphasis added).

Shield transfer impedance $\mathbf{Z}_T$ is defined by

$$Z_T = \frac{1}{I_S}\left(\frac{dV}{dl}\right)$$

where V is the voltage between the signal conductors and the shield, l is the length of the shield, and I_S is the shield current.

In other words, $\mathbf{Z}_T$ defines the voltage placed on the signal conductors as a result of shield current, and is a measure of the shielding effectiveness. A low value for $\mathbf{Z}_T$ indicates better shielding. In essence, SCIN is the result of the imbalance of $\mathbf{Z}_T$. At low frequencies (below about 1 MHz), $\mathbf{Z}_T$ is equal to the dc resistance of the shield. (Ott) For coaxial cables, $\mathbf{Z}_T$ can also be seen as the common impedance coupling between center conductor and the shield.

Openings in a shield All shield mechanisms – absorption, reflection, diversion, and eddy currents – depend upon the uniform flow of current in the shield. Any breaks (openings) in the shield will reduce that uniformity, because the current must flow around the opening. This distortion of the current path also increases the loop area for the current!

The loss of shielding caused by a break in the shield will be proportional to the wavelength of the opening that results – that is, there is a direct relationship between the loss of shielding effectiveness, the largest dimension of the opening, and the wavelength of the interfering signal. In general, many small openings do far less damage than one larger one of equal area. The tiny holes between the strands of a braided shield begin to reduce its effectiveness in the tens of MHz and above. An opening in an equipment housing or connector panel for XL connectors starts to become significant above about 100 MHz. If an opening is required for ventilation or to mount multiple connectors, it is far better to provide multiple small openings than one larger one. One of the worst possible types of openings is a long slot, because current must flow around it. If the opening is broken up into much smaller circular (or even rectangular) openings, shield current can still flow in the parts of the shield that remain between the openings. Ott provides graphs and equations to quantify these considerations.

Distance is your friend, especially in the near field! In the <u>very</u> near field, field strength of the dominant field falls off as the <u>**cube**</u> of the distance from a <u>**point source**</u>, and as the <u>**square**</u> of the distance for the other field. For example, doubling the distance between the noise source and victim circuit would divide the dominant field strength by 8. Since the induced voltage or current is proportional to the field strength, that's 20 log 8 = 18 dB for the dominant field, and 12 dB/doubling for the other field! In the <u>far</u> field, it's 6 dB per doubling. 6 dB/doubling also applies to coupling between parallel cables.

And finally, these words of wisdom from two authorities:

Clayton Paul: "***Absorption loss is the dominant shielding mechanism for near field magnetic sources at all frequencies.*** However, both reflection loss and absorption loss are quite small for near field magnetic sources at low frequencies, so that other more effective methods of shielding against low frequency magnetic sources must be used." [his emphasis]

"There are two basic methods for shielding against low-frequency magnetic sources: diversion of the magnetic flux with high-permeability materials and the generation of opposing flux via Faraday's law, commonly known as the 'shorted turn method.'" *[In these two statements, Paul is viewing diversion as the equivalent of absorption.]*

And from Henry Ott:

- ***Electric fields are much easier to guard against than magnetic fields.***
- ***The use of non-magnetic shields around conductors provides no magnetic shielding.***

- ***The key to magnetic shielding is to decrease the area of the loop. To do that, use a twisted pair or coaxial cable if the current return is through the shield instead of the ground plane.***
- ***Only a limited amount of magnetic shielding is possible in a receptor circuit that is grounded at both ends, due to the ground loop formed.***

Acknowledgements: Thanks are due Henry Ott, Clayton Paul, and Phil Giddings, for the clear and cogent analyses of the complex matter of shielding in their text books. All are worth owning. Bill Whitlock, Ray Rayburn, Dale Manquen, and Bruce Olson reviewed the manuscript and made useful suggestions.

References:

1. Henry Ott, "***Noise Reduction Techniques in Electronic Systems, Second Edition,***" Wiley, New York, 1988

2. Clayton R. Paul, "***Introduction to Electromagnetic Compatibility,***" Wiley, New York, 1992

3. Philip Giddings, "***Audio System Design and Installation,***" Focal Press, 1990

4. Edward Vance, "***Coupling to Shielded Cables,***" Wiley, New York, 1978

5. Philip Giddings, "***Shielding***," http://svconline.com/news/avinstall/shielding/

6. American Radio Relay League, ***ARRL Handbook***, updated annually http://www.arrl.org

7. Neil Muncy, "***Noise Susceptibility in Analog and Digital Signal Processing Systems***," *J. Audio Eng. Soc.,* vol 43, No. 6, pp 435-453, 1995, June

8. Jim Brown/Bill Whitlock, "***Common Mode to Differential Mode Conversion in Shielded Twisted Pair cables (Shield-Current-Induced Noise),***" AES Preprint 5747

9. Bill Whitlock, "***Balanced Lines in Audio Systems: Fact, Fiction, and Transformers,***" *J. Audio Eng. Soc.,* vol 43, No. 6, pp 454-464, 1995, June

10. Howard Johnson and Martin Graham, "***High-Speed Signal Propagation,***" Prentice Hall, Upper Saddle River, NY, 2003

Appendix – Transmission Lines at Audio Frequencies, and a Bit of History

If you study transmission lines in college or from the ARRL Handbook, you will learn the classic equation for characteristic impedance.

$$Z_O = [\ L/C)\ ]^{1/2}$$

What most of us have long forgotten (and that few universities teach) is that this is the simplified form of the equation. In our engineering classes, we make equations simpler by assuming certain conditions will be true for what we think will be the conditions when we use them. That's fine as long as we don't forget the assumption, but most of us have. The full equation for characteristic impedance is

$$Z_O = [\ (R+j2\pi fL)\ /\ (G+j2\pi fC)\ ]^{1/2}$$

At high frequencies (that is, **f** is large), (**2πfL)** is much larger than **R,** and **(2πfC)** is much larger than **G.** So at high frequencies, the equation becomes simply

$$Z_O = [\ j2\pi fL)\ /\ j2\pi fC)\ ]^{1/2} \qquad \text{and further simplifies to} \qquad Z_O = [\ L/C)\ ]^{1/2}$$

which is the familiar equation. But what if **f** is <u>not</u> large? Fig 1 shows the impedance of a typical coaxial cable at audio frequencies using the full equation. Why? Because at low frequencies, **R** is much greater than **2πfL,** and **2πfC** is much greater than **G.** Thus, at low audio frequencies, the equation simplifies to $Z_O = [R/j2\pi fC]^{1/2}$. This causes the impedance to be quite high and predominantly capacitive near zero frequency. As frequency increases, L becomes significant, and Z_O transitions to the familiar 50-100 ohm impedances at high frequencies. And the transition occurs right in the middle of the audio spectrum! Through the transition region, the "j" term causes Z_O to be a combination of resistance and capacitance. [Although the data presented here is for coaxial cables, virtually all commonly used cables, including twisted pairs, exhibit these characteristics!]

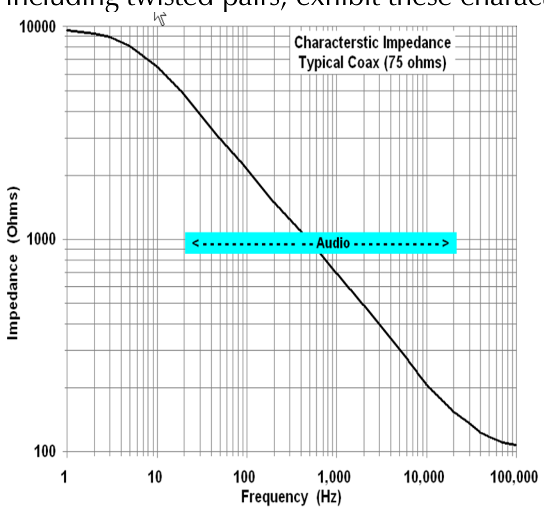

Fig. 1 Characteristic Impedance of Typical Cable at Audio Frequencies [6]

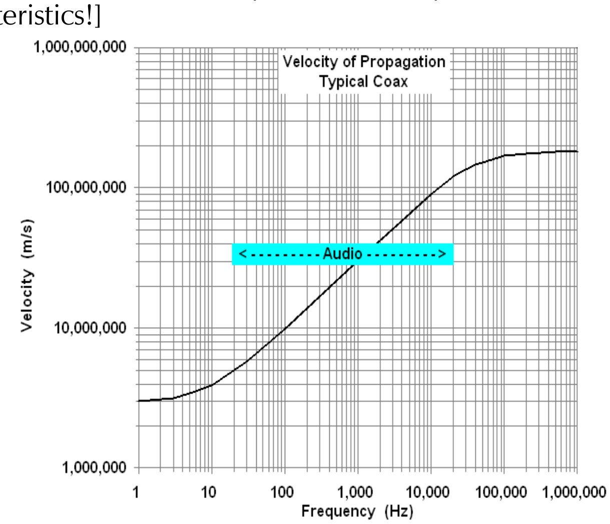

Fig. 2 Velocity of Propagation of Typical Cable at Audio Frequencies [6]

The graph for velocity of propagation (Fig 2) yields a similar surprise, going through the same sort of transition, and for the same reason. These data are computed for the theoretical case of a line terminated in its characteristic impedance, which, as we have just learned, varies over two orders of magnitude. Indeed, all of the characteristics of a cable vary by nearly two orders of magnitude through the audio spectrum – the lowest frequencies take 50 times longer to get to the other end of the cable than the high frequencies! These effects are too small to notice in a theater where cables are rarely more than 500 ft long, but they can be quite significant on a telephone line that is tens of miles long!

So we see that the behavior of transmission lines at low frequencies is, indeed, a very complex matter. Telegraph and telephone companies learned early on that very long lines (tens and hundreds of miles) need to be equalized – both for frequency response and to maintain velocity of propagation reasonably constant. Without that equalization, they could not even transmit Morse Code reliably on long circuits. Later, on the first transcontinental telephone circuits, voices were mangled beyond recognition. In both cases, the cause was the variation in $\mathbf{V_P}$ over the audio spectrum.

In 1893, the physicist Oliver Heavyside showed that if R/L could be made equal to G/C (or RC=GL), a constant velocity of propagation would result and the attenuation would be minimized. The problem was (and is) that in practical cables, L is much too small to achieve these ratios. In the earliest days of the telegraph (the latter half of the 19^{th} century), the signal was carried on a single iron wire strung on poles with the earth as a return. These circuits had more L, both because of their spacing and the use of the iron wire, and the variation in velocity was not too much of a problem for telegraphy (in those days, there was no AC power to induce hum). With the advent of the telephone (invented in the late 1870's), many more circuits and greater bandwidth without time distortion was needed To provide more circuits and reduce crosstalk, twisted pair copper cables were developed. This made L (the loop inductance) much smaller, which in combination with the transmission of audio motivated Heavyside's thinking.

The modern solution, based on Heavyside's work, developed by M. I. Pupin and G. A. Campbell, and patented by Pupin in 1900, was (and still is) to add inductors in series with a long line at intervals of several thousand feet. This serves to turn the line into a bandpass filter with a sharp cutoff between 3 kHz and 15 kHz, depending on how large the inductors are and how closely the inductors were spaced. Ordinary telephone voice circuits are equalized using 88 mH inductors at intervals of about 6,000 ft, resulting in an upper cutoff frequency of 3 kHz. Greater bandwidth could be achieved with smaller inductors and/or shorter spacing, and with the addition of repeater amplifiers along the way. The addition of the inductors did four important things – it extended the frequency response, it equalized the delay, it made the impedance more constant, and it imposed the low-pass filter characteristic. Eventually the repeaters themselves were designed to provide equalization. The work that made these equalizing repeaters possible was done by Harold Black (the inventor of negative feedback), Harry Nyquist, and Hendrik Bode, all of them engineers at Bell Labs! [3]

If cables could be constructed with the desired R, L, G, and C relationships for constant velocity of propagation, they would not have that extreme low pass characteristic. It is the addition of the discrete inductors that does that! Loading can be done continuously, and even the earliest transoceanic cables were continuously loaded. [1] To provide continuous loading, the center conductor of the more modern of these cables were wrapped with permalloy tape having a high permeability. This provided equalization of both time and amplitude response, but without the low pass filter. Around 1960, for example, one such trans-Atlantic cables operated as a carrier-based system (that is, audio was modulated onto radio carriers and sent over the cable), and was carrying 36 high grade circuits plus some utility grade circuits, and was planning an expansion to 48 circuits. Although the bandwidth of this cable or the circuits is not stated in the reference, the number of circuits implies bandwidth of at least several hundred kHz. [2] Note also that such cables will still exhibit the more gentle low pass characteristic due losses that increase with frequency (mostly due to skin effect).

Myth – 600 ohms and open wire line. Calvert computes Zo, Vp and attenuation for an open wire line that was commonly used in early telephony-- #12 AWG copper spaced 12" apart, and for which the basic parameters were well documented. For dry weather, G = 0.29 μS/mile, R = 17.1 Ω/mile, L = 3.73 mH/mile, and C = 7.83 nF/mile. At 300 Hz, $\mathbf{Z_O}$ is 1,345 $\Omega \angle$-36.8°, $\mathbf{V_P}$ = 2.23 m/s (VF=0.74); at 1,000 Hz, $\mathbf{Z_O}$ = 841 $\Omega \angle$-23.8°, $\mathbf{V_P}$ = 2.82 m/s (VF=0.94); and at 3 kHz, $\mathbf{Z_O}$ = 712 $\Omega \angle$ -10.2°, $\mathbf{V_P}$ = 2.95 m/s (VF=0.98). Attenuation is 0.08 dB/mile at 300 Hz, 0.102 dB/mile at 1 kHz, and 0.107 dB/mile at 3 kHz.

References:

1. J. B. Calvert, "***The Telegrapher's Equation***," http://www.du.edu/~jcalvert/tech/cable.htm, self published, 2003 (Dr. Calvert is associate professor emeritus at the University of Denver)

http://www.du.edu/~jcalvert/ His extensive website covers a broad range of topics and is quite interesting.)

2. R. K. Moore, "***Traveling Wave Engineering***," McGraw Hill, New York, 1960 (My college Transmission line text).

3. David A. Mindell, ***Opening Black's Box – Rethinking Feedback's Myth of Origin***," http://mit.edu/6.933/www/black.pdf

4. Richard Lao, ***The Twisted-Pair Telephone Transmission Line***, High Frequency Electronics, Nov, 2002

5. Belden Engineering Department, "***Characteristic Impedance of Cables at High and Low Frequencies***,"

6. James Hayward, "***Cable-itis Strikes Again***," Audio Ideas Guide, Summer/Fall 2000