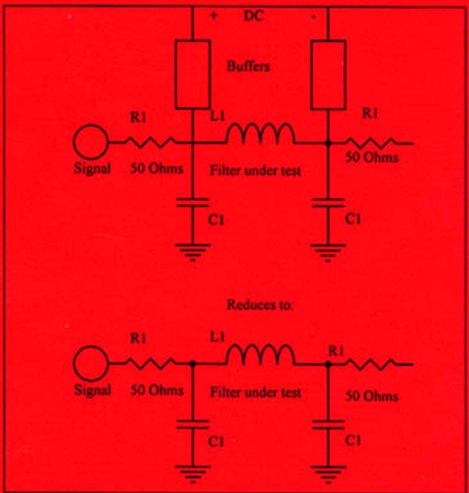
EMI Filter Design

Second Edition
Revised and Expanded



Richard Lee Ozenbaugh

EMI Filter Design

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This is dedicated to my wife, Pansy, for her moral support. Also, I wish to thank her for the computer time and for the seminar time she lost (and for not complaining when she had the right to), especially during this rewrite.

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Preface

Over the last twelve years, I have made friends with many EMI "gurus" across the United States, but primarily in the Southwest. Many have attended my EMI Design seminars and I have met others as an Applications engineer in the field. My desire was to increase my knowledge of EMI filter design by learning new and better techniques. I concluded, after some years, that very few of them had a concrete method for the design of these low pass filters (it is black magic, isn't it?). I got to know most of these engineers quite well. They would give me a very strange look when I would ask them this leading question: "What is your design technique?" I finally concluded that in most cases it depended on what was handy or readily available. They would have a certain capacitor or inductor on their bench or in stock (or readily available from a supplier). At best, they would calculate the needed inductor(s), or capacitor(s), to arrive at the desired loss. Some would give me profound statements such as, "We design for maximum loss!" or "We use [a certain program] to get the values needed." To accomplish this latter method, the "guru" would reach the desired loss by backward-engineering the network—modifying the inductor or capacitor values or changing the number of components to get the apparent desired loss. Some designs had so much capacitance across the line that the EMI filter inrush current, added to the power supply filter capacitors charging currents, would trip the circuit breaker. Others would spend days measuring the line and load impedance using network analyzers so the EMI filter could be matched to these two end sections. Unfortunately, the end sections would change from application to application so that the filter center piece would not function properly, because of the mismatch existing in the band pass.

Another technique I have found is for the engineer to grab any core and start winding wire around it for its inductor. Usually they could not even tell me the type of core—MPP, HF, tape wound ferrite? "Just an available core, something off the shelf." This does not give a reference point from which to start. Whether the filter works or fails, the filter inductor core must be known. The fallacious statement often made is that a given core was used by one of the filter manufacturers and therefore must be the correct type and about the correct number of turns.

A doctor from Sierra Vista, Arizona, wrote a very fine article in the late 1980s on commercial filters that included a good method for calculating the component values. The title of the article was something like "Gus, the Electrician." His filter component values came from "M-derived" filter design techniques

$$L = \frac{Rd}{\pi F}$$

$$C = \frac{1}{\pi F R d}$$

where L is the inductance, C is the capacitance, Rd is the design impedance, and F is the cutoff frequency. He felt that most of the original filter designers allowed their cutoff frequencies to fall much too close to power line frequency. This happens in the 461 specifications requiring high dB losses in the 10 to 14 kHz region. Otherwise, few EMI filter cutoff frequencies drop as low as indicated in the article. This is especially true for commercial filters. Gus would often switch the ground and return wire because he thought that ground was ground. This was a very good article and a very good design method for the component values. He did not offer a method to calculate the cutoff frequency.

Other engineers have correctly used the energy spectrum of the switcher frequency to determine the amount of loss required. There is trouble transforming this information when the engineer selects the filter component values. Many designs use very small capacitors between line and ground with very high values of resistors in series with them. Some have used capacitors and resistors in series such as 2000 pF in series with 470 ohms. The balanced circuit requires two of these networks wired from both inputs to ground. One network wired between the high, or hot, to ground and the other network wired between the neutral and ground. So at the higher frequencies (in the case stated, above 2 MHz), the line to line would be 940 ohms. Why? The line impedance would be somewhere between 50 and 130 ohms, depending on the line. The energy content of the switcher is low at these frequencies but, again, if this energy was high, why the high series resistance?* Is there a time when this is a good technique? Yes, but

*This was written when switchers worked at 80 kHz maximum but now some are in the MHz region.

the resistors are quite small in value. Usually, these values fall in the order of 10 ohms and the capacitors are in the area of 1 μF. This is done to reduce the full circuit Q to lower the tendency of the filter to oscillate at the filter's normal resonant rise frequency.

It has been very informative to be in the field, over the years, seeing these filters in operation, especially the higher wattage power line filters. Many of these filters were running very hot to the touch at much lower power ratings than the design specification. Others would have a heavy peak-to-peak voltage at some higher frequency, usually at some odd multiple of the line frequency, superimposed on the line voltage. Many of these troubles are caused by the filter cutoff frequencies too close to the line frequency, as mentioned earlier, and excessively high filter circuit "Q" allowing the filter to oscillate at the resonant rise frequency. Both of these conditions allow the filter to heat, and this in turn heats the capacitors. This will then blow the capacitors, destroying the filter. Some filter companies have rebuilt these filters using the same designed-in capacitor, only to have them fail again in a short period of time.

The impetus to develop simple design parameters and techniques came from the situations described above. Hopefully, the information supplied here will make life easier for those responsible for designing EMI filters. The general goal of this book is to provide proper design techniques that will help to nip all of the above practices in the bud. This must be done quickly because the EMI world is bad enough now and the worst is yet to come, because the specifications will become more involved and harder to meet as time moves on. The FCC will move to greater losses at lower frequencies, to follow VDE, and meanwhile VDE will move to tighter requirements. The same will hold for the MIL F 15733 moving to 28861, and 461 will move to rev D and then E. The 220A specification will also be updated. Other test methods will follow suit and be made harder to accomplish. The medical will move out of the industrial specifications to tighter requirements. The medical leakage current specification for equipment directly touching the patient will decrease below the present 100 μA—some are now asking for 50 μA.

The primary goal of this book is to provide a quick, but effective, design method for the filter design engineer, equipping the filter engineer to design the filter in minimal time and have the filter function with minimal adjusting of the prototype. Even if the filter can't be adjusted enough to meet the specification and requires major changes, the time lost will be minimal—not weeks or months.

The first chapters are basic but the EMI person should still scan them. This book is based on two main principles. The first is called the "engineer syndrome." At what point is a project or product engineered? In the EMI arena, it's a different core, different capacitor type or shape, different style filter (from L's to T's), or a different mechanical shape of the filter. This book advocates getting the filter designed, built, and tested as soon as possible.

The second is the KISS principle. The matrices of the various filter types are discussed in Chapters 13 and 14. Much of the material covered here is based on close approximations along with heuristic and empirical knowledge. The abbreviated method is as follows:

1. Decide the best filter type from Chapter 12.3.
2. Find the filter component values—both differential and common mode—either from a) the matrix equations of Chapter 13 or 14 or b) the section in 18.8, "F_0—The Easy Way." Method b will get the filter designer onto the proper ballpark property while a) will get the designer onto the ballpark playing field. Small adjustments to either of these will get the designer to home plate.
3. Design the components for the filter based on Chapter 6.
4. Design and build the case or container and have it silver plated using Chapter 17 to help—this assumes a military specification.
5. Install the components and test the filter in the open container.
6. Adjust the filter for the desired loss if needed by:
    a. Adding lossy components (4.4).
    b. Adding small line-to-line capacitors in parallel to the existing capacitors (keep the lead length as short as possible).
    c. Add ferrite beads if the current is low enough (typically 5 amp limit).
    d. Add several turns on the inductors (watch for saturation). (If this filter is a T, whatever turns are added to the central inductors of the multiple T, increase the turns by half on the two outer inductors because they are half the value of the inner inductors.)
    e. Add an RC shunt (a resistor and capacitor in series across the line) or a Cauer filter to tune one of the series inductor. See section 5.7.
7. Make sure that the product can be manufactured and is repeatable for production. Adjusting the filter as described above will move the designer from the ballpark property to home plate with 2a or 2b above.

Since the first edition of this book was published, I have determined that basic information and definitions of EMI are mixed in the field. Some have developed steadfast rules over time that other EMI people violate all the time.

One example is the common mode inductor. Some will insist that these two windings must be separated into sections so the two windings are apart on the core with gaps at each end. Others will wind these bifilar with good results.

RFI uses both techniques—sometimes one method is better than the other. The first method creates high leakage inductance while the other method does not. This leakage inductance may ring with a nearby circuit capacitor, causing problems. The second method adds capacitance between the two windings that may also cause other problems. So, both have their place and neither is absolutely correct—use the one best for the situation.

Richard Lee (Oz) Ozenbaugh

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Contents

1

Why Call EMI Filters Black Magic?

Most engineers, both designers of electromagnetic interference (EMI) devices and others, call EMI black magic. There are four main reasons for this. First, there has not been a well-defined design method. Second, the input and output impedances are not constant over the band of interest. The impedances are usually just good guesses, at best, because the measurements are costly to make and are rarely repeatable. Third, the filter insertion loss test method specifications often confuse or influence the design method. In the 220A specification, requiring 50-ohm source and load impedances, does the filter engineer design the filter for the 50 ohms or for real-world impedances? If the design passes the Military Standard (MIL STD) 220A, 50-ohm, test, what will happen in the real world? Is it the designer's responsibility to meet the specification or to stop the real-world emissions? Fourth, the design methods outlined in various publications are generally very complex. Most require measurements that are difficult to make and time consuming. Often expensive additional equipment is required to obtain the needed parameters.

1.1. WHAT IS EMI?

EMI is electromagnetic interference and is called conducted emissions or radiated emissions. This book covers mainly conducted emissions (CEs), which means any unwanted signal or noise on the wiring or copper. The reason for the reference to power cabling is that EMI filters are part of the power wiring and are designed to remove these unwanted properties from the copper wiring. Why is this from the wiring? What does this have to do with magnetic fields? The reason is that any current flow creates an associated magnetic field. You cannot have one without the other. Therefore, this high-frequency unwanted signal creates a magnetic field that can interfere with surrounding equipment. It is the filter's function to remove this current so that its associated magnetic field will not interfere. This noise can originate either from the line or from the associated equipment that the filter is built into. From the equipment side, the noise could be coming from computer clock frequencies, parasitic oscillations in the switcher power supply inductors or transformers, power supply diode noise, harmonics of the line frequencies due to the high peak current charging the power supply storage capacitor, and many other sources. From the line, the noise could be due to flattening of the sine wave voltage caused by the high peak currents slightly ahead of 90 and 270 degrees due to the total of the power supplies fed from the line without power factor correction circuitry. This generates odd harmonics that feed the EMI filter. Other sources of noise from the line are other equipment without any filtering and heavy surges of equipment being turned on and off. Lightning and EMPs (electromagnetic pulses, possibly from nuclear explosions) create other line problems for the filter.

To review, EMI is any unwanted signal from either the power line or the equipment and must be removed to prevent a magnetic field from interfering with closely associated equipment or to stop a malfunction of the equipment containing the filter. For example, if you or a loved one was in the hospital, you would not want the heart monitor to dip every time the x-ray machine was used just because the same copper connected them. Here, the heart monitor filter would remove the pulse from the x-ray machine. Or, better yet, the x-ray machine filter would stop the noise from leaving.

1.2. A STANDARD FILTER COMPANY CONTRASTED WITH AN EMI COMPANY

Most of the energy in the stop band (the frequency area to be attenuated) of the filter is reflected to its source. This fact is often overlooked in both standard filter and EMI technology. The remaining energy is expended in the inductors through the DC resistance of the coil, the core losses (eddy currents and hysterisis), and the equivalent series resistance of the capacitors. All engineers have learned this in the past but often forget it somewhere along the way. Whereas this is damaging to the wave filter people, it is an aid to the EMI group. The dissipative filter (discussed in Sec. 6.5) dissipates the energy in the stop band. Standard filter designers have several excellent filter design programs in their computers, such as Butterworth, elliptic, Chebyshev, and M derived. They know the input and output impedances of the source and load (usually the same), the allowed band-pass ripple, the 3-dB point, and the stop frequency (the first frequency with the required amount of loss). The wave filter designer is provided with passband ripple dB and other information that dictates a particular filter topology. When the computer design is complete, the designer will be very close to the results required. I am not saying that this is easy or that the standard filters never have to be reworked. The filter may have to be altered by adding stages or through the use of a different technique, or topology, to achieve the desired results.

Most EMI filter manufacturers design only the low-pass filters (all pole networks) needed for the required EMI attenuation. Rarely do they build band-pass or other conventional filters. The technology used in conventional, or standard, filters is truly different from that used in the EMI filter. The EMI filter design is very loose compared with that used by the conventional filter manufacturer, especially if the EMI filter designer uses the techniques mentioned in the preface of this book. These techniques would upset most of the normal filter, or wave filter, designers. The EMI filter component values are very flexible, so the engineer can use standard values. These filters are adjusted only to meet the required insertion loss specification assuming the rest of the specification is met.

The languages spoken by the two groups are also different. True filter houses often speak of poles, zeros, group delay, predistortion, attenuation, and terms such as the order of the filter. The EMI filter designer thinks in terms of attenuation, insertion loss, filter voltage drop, filter voltage rise, and the number of filter sections required to meet the insertion loss. Although the power source may have harmonics, the actual power supplied to the device through the filter is restricted to the fundamental frequency. Such a harmonic content is especially true for power supplied locally by shipboard generators and remote stations where the generator is near or well past peak power. So flat frequency response, low phase distortion, or low peak-to-peak ripple across the filter band pass is not an issue here. These power line harmonics furnish no power to the load, so the EMI filter designer is not concerned with them. As a result, terms such as group delay, ripple, and phase distortion are never heard.

To summarize, the requirements of the conventional, or wave, filter house are entirely different from EMI requirements. The technologies are completely different. The conventional, or wave, filter house component value is critical. This requires a better grade of components and often requires tuning. EMI houses often tune, such as with a Cauer filter, but the reasons are different.

1.3. POWER DENSITY SPECTRUM OR ENVELOPE

The EMI filter designer is not blessed with a density spectrum as described above. The designer knows the power line frequency and that it will have some harmonics. The lower frequency harmonics should be passed on to the load. Most are aware that the harmonic power provides no power to the load, but neither should the EMI filter attenuate them. The devastating effects of the magnetic fields are not as drastic as the higher frequencies, so this filtering is not as important. Such attenuation of these low frequencies would call for larger filter capacitors, and capacitor currents, and larger cores. These cores would be subject to greater eddy currents and hysterisis losses. Then each core would develop more heat within the filter. The filters must pass high power at the line frequency and also the power of some of the odd harmonics of this frequency. This demands larger components to handle the energy passed to the load, together with the larger wire diameter required for the current. This dictates that the filter will increase in cost, size, and weight. However, if it is a requirement to attenuate these frequencies, a silicone tape–wound toroid transformer could be added to the filter components and the low-frequency loss would be great.

Summarizing, for the reasons just given and for other reasons covered later in this book, the cutoff frequency usually starts around 4000 Hz for 60-Hz systems and around 8000 Hz, if possible, for 400-Hz systems to keep the components small for better performance (see Sec. 1.7).

1.4. POWER TRANSFER

The power to be attenuated is typically minimal compared with the power transmitted or throughput. Exceptions would occur when there are larger electromagnetic pulses from lightning (EMPs are discussed in Chapter 11). These conditions must also be treated or attenuated. This typical noise, called flea power, could be from any or all of the following: (1) the switcher frequency plus the odd harmonics, (2) the power supply and/or the switcher diode noise, and (3) the parasitic oscillations of the switcher inductor or transformer. This noise energy is usually much lower in level than the power supplied to the load. To move this power from the line to the load, larger filter components are needed.

The transfer of power creates (1) larger filters with more weight and cost and (2) much lower self-resonant frequency (SRF) values. The weight, size, and cost of the inductors and capacitors are directly proportional to the power. The SRF value of these filter components is indirectly proportional to the power demanded by the load. This limits the filter's usefulness at the higher frequencies. These conditions will dictate that other components must be added to the filter to compensate for this lack at higher frequencies. Unfortunately, this increases the weight, size, and cost of the filter. The assumption here is that the SRF values of the inductors and capacitors are well below the specified upper frequency limit that must be filtered. These additions to the filter could be handled by placing small ceramic or extended foil Mylar or polypropylene (depending on the power frequency) capacitors in parallel with the existing capacitors. The second capacitor will raise the SRF. Capcon and ferrite beads could also be used.

To review, smaller components have higher SRFs, making them perform better at higher frequencies. Several stages in tandem reduce the size of the components, resulting in higher SRFs and lower component costs.

1.5. SPECIFICATIONS: REAL OR IMAGINED

Specifications are another thing that makes EMI "black magic." Some test specifications cloud the design and make it overspecified. For example, one company had been using a particular filter for years without any problems meeting the EMI qualifications. Then the test specification was changed from the 220A to the current injection probe (CIP) method and the filter never passed. These specifications are such that they conflict with reality. The 220A specification calls for losses within the filter with a source and load impedance of 50 ohms. The filter will, in reality, feed a power supply that is rarely close to 50 ohms and work into a source of rarely 50 ohms. In the preceding example, the system was for aircraft where the power feed would be very short. For this line impedance to approach 50 ohms, the line frequency would be in the upper kHz range. What does the EMI filter designer do: match the 50 ohms or meet the real-world specification? What is meant here by the "real-world specification" is similar to the quality (qual) test that may follow. This test measures the noise out of the system through the filter. The line impedance simulation network (LISN) is often used as the source impedance for these tests and is closer to the real-world requirements.

Most specifications that call for an LISN require 50 ohms output impedance. Unfortunately, the 50 ohms is not reached until well above 100 kHz. If the only concern of the designer is to match this 50 ohms, the filter will be matched to this source impedance of 50 ohms. There are two concerns here. First, what is the lowest frequency and loss required? Is it below 100 kHz? The LISN output impedance drops rapidly from 50 ohms and the filter is then mismatched. Second, what happens in the real world when neither the source nor the load impedance is close to 50 ohms? Figure 1.1 shows a typical LISN where the line impedance, whatever it is, is shunted by the input network at frequencies above the point where the 22.5-µF capacitor's impedance is equal to 1 ohm. This is at 7000 Hz, so at frequencies above 10,000 Hz the line impedance is reduced to the 1-ohm value, while the load side of the LISN looks like 50 ohms, following the same logic, also above 10,000 Hz. The impedance of the inductor is low at 10,000 Hz (3.5 ohms). At 10 times these frequencies, the inductive reactance is only 35.2 ohms and reaches 50 ohms at 142,000 Hz. The entire network at 142 kHz generates a source impedance for the filter of 35 ohms at 44 degrees. It is obvious that the LISN will not look like 50 ohms until well above 150 kHz. The proposed LISN is shown in Fig. 1.2.

The real intent of the filter is to attenuate conducted emissions of differential and common mode origins from both the device and the line. The test specifications rarely prove that the filter will pass with any degree of satisfaction within the system specification or real-world specification. The filter can often pass the insertion loss in the test laboratory and fail when tested along with the

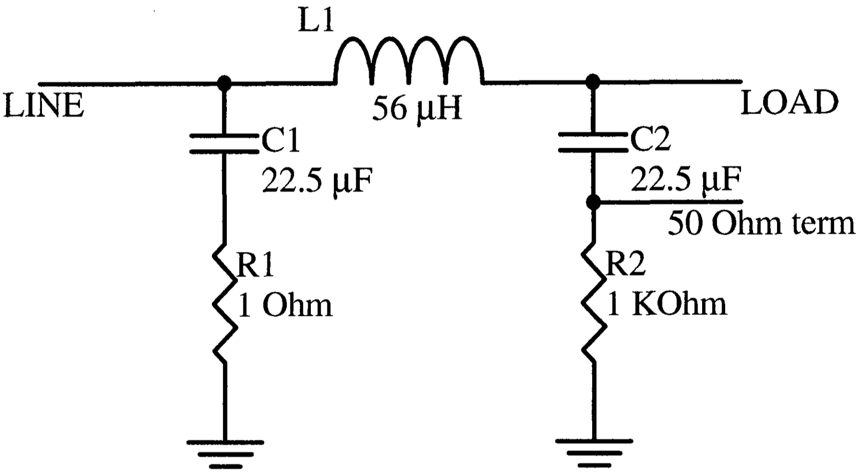

FIG. 1.1 Typical LISN.

system. Often, using a filter that passed the full test as a target, or bogie, gives disappointing end results. If the filter in question appears to pass the bogie, this filter is often been tagged as bad later by the system tester. The 461 specification (Fig. 1.7) is more realistic than the 220A, and Robert Hassett, the vice president of engineering at RFI Corp (now retired), has given several presentations for the Institute of Electrical and Electronics Engineers (IEEE) and other groups that show the advantage of moving away from the 220A test method to the CIP method. Mr. Hassett has tested many filters by both methods and shown the difference in the insertion loss curves (Fig. 1.3). The curves in Fig. 1.3 show the difference using an L filter with the capacitor facing the line as compared with the inductor facing the line (Fig. 1.4). The same is also true for the pi filter due to its input capacitor facing the line. Either filter will look good under the 220A tests. This is due to the 50-ohm source impedance. The CIP utilizes a 10-μF capacitor, and this will reduce the filter loss by 6 dB, especially at the lower frequency end.

In the case of Fig. 1.5, the signal generator—now normally from the spectrum analyzer tracking generator—has an output impedance of 50 ohms and feeds a coaxial switch (not shown) and the load impedance is the receiver's input impedance (also 50 ohms) fed through a second coaxial switch. A calibration path is provided between the two coaxial switches. This test method makes the pi filter function as a true three-pole filter, giving 18 dB per octave, 60 dB per decade. The real-world initial losses at the low-frequency end would shunt out the input filter capacitor. This would give initially 12 dB per octave of loss until well over

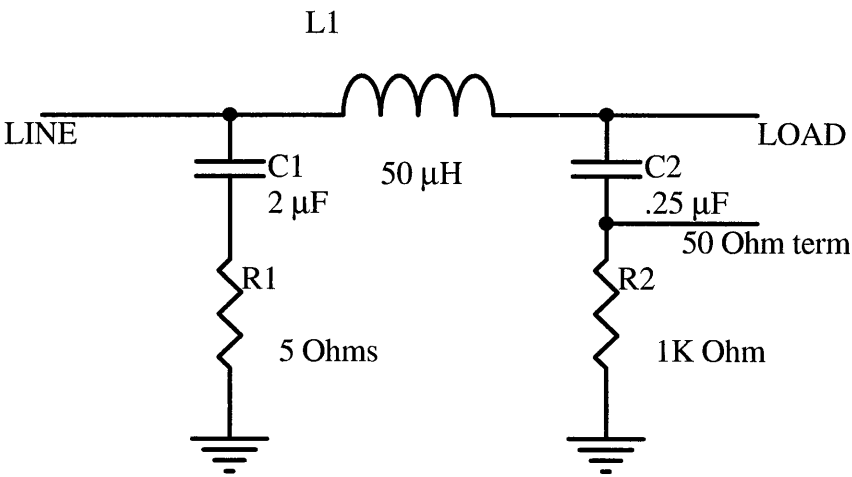

FIG. 1.2 The proposed 461 LISN.

100 kHz (depends on the line length). This test method masks this flaw of the pi filter or the L filter with the capacitor facing the low impedance.

Figure 1.6 shows the method used by Robert Hassett at RFI.

In the 461 specification, the current probes feed the measuring equipment to compare the two currents. This method shunts the input pi filter capacitor copying the real world up to the frequency where the 10-μF capacitor's SRF takes effect. This method makes the pi filter loss 12 dB per octave, 40 dB per decade, instead of 18 dB per octave, 60 dB per decade.

Other test methods have been suggested. Mitchell Popick, previously with Axel Corp. (now with The James Gerry Co. in Alhambra, California) and a member of SAE EMI group and the dB Bunch, recommends that the load side of the filter face a diode bridge that is properly loaded. This is much better for those that will feed a power supply, which is about 95% of the time. This also shows how the filter handles diode noise. The diode noise is the leading noise competing with the switcher noise. These are primarily the odd-order harmonics of the power line frequency and spikes during turn on and turn off. It works well in three phase systems as well, but make sure that the total inductive reactance of the filter inductors is much lower than the primary inductance of the transformer. This is especially true when multiphase transformers are to be used. These are often autotransformers and the primary inductance is much lower than for the isolation transformer type. The inductance of the filter and the primary inductance of the transformer can form a voltage divider reducing the voltage feeding the load.

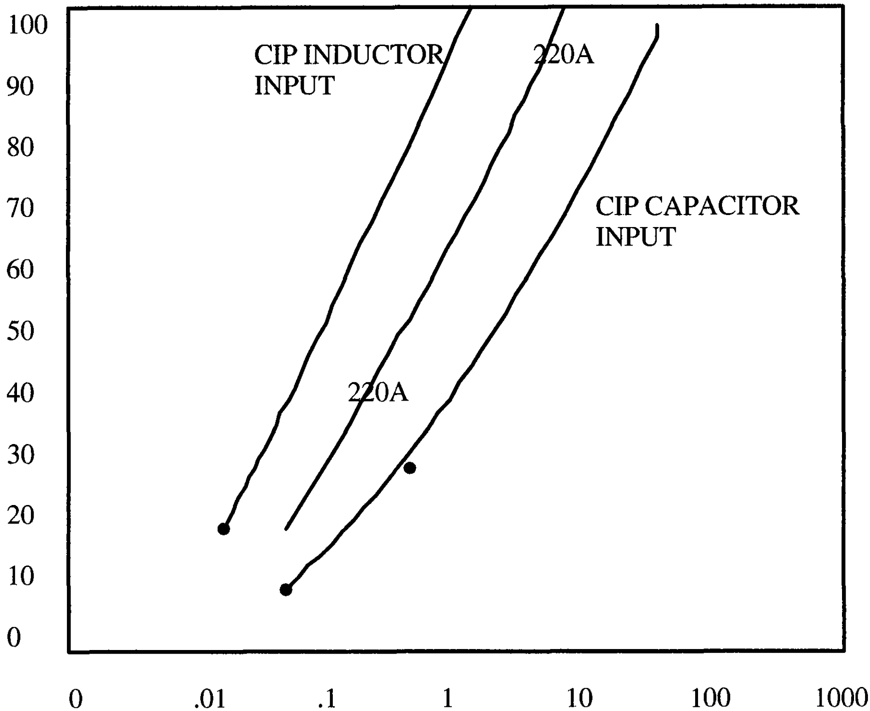

FIG. 1.3 Curves for L filter with capacitor facing low-impedance source.

This is another reason to avoid the commercial filters, where one filter fits all. These may work fine in some applications and fail in others.

In Fig. 1.7 the spectrum analyzer sees the diode, switcher, and parasitic noise that is allowed to pass through the filter under test. If the load is the system rather than the load resistor, as in the drawing, the analyzer will get the full noise signature of the equipment. This is much more of a real-world test because the filter must handle all the noise sources at the same time, which is what happens in the real world. In other words, the filter could saturate under this condition while the CIP and 220A are looking at a single frequency from the tracking generator. Of course, both 220A and CIP pass the full power from the line, but so does the test method above.

Reviewing, try not to use a capacitor input filter for the CIP test method because the loss is 0 dB per octave for this component whereas it should be 6 dB per octave. This component costs money, demands filter room, and adds weight without performing until the frequency is very high.

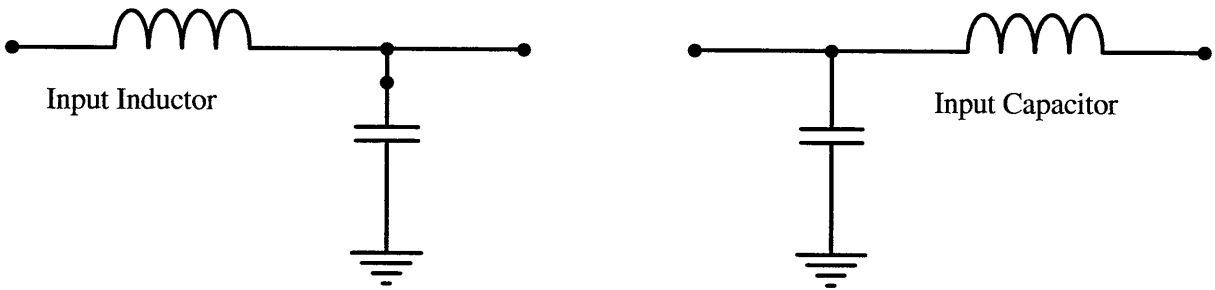

FIG. 1.4 L filter with capacitor facing the low-impedance source.

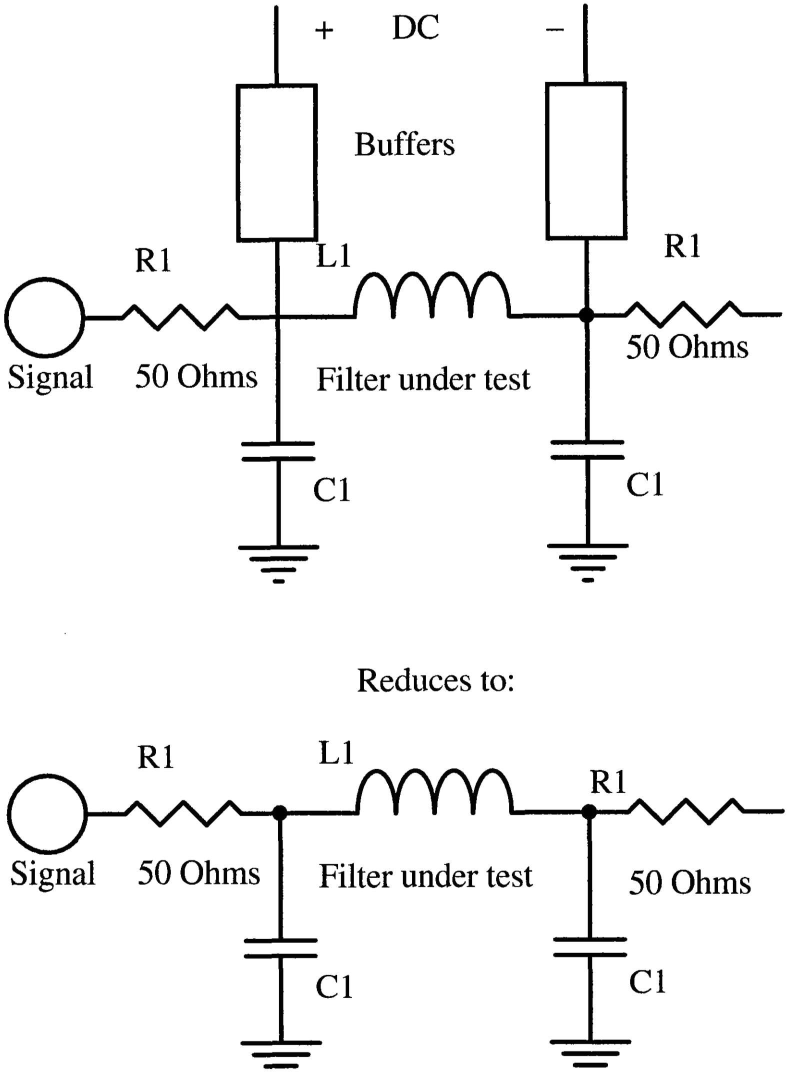

FIG. 1.5 220A test setup.

1.6. THE INDUCTIVE INPUT FOR THE 220A TEST METHOD

This is similar to the capacitor of the pi filter in the CIP method but not as severe. With the 50 ohms impedance in the 220A test system, what is the inductor impedance going to add? At least the 50 ohms is there and the inductive reactance adds to it at 90 degrees. We are speaking here of either an L or a T filter. They are not responsive until the impedance of the inductor reaches 50 ohms. Regardless, this takes effect orders of magnitude ahead as compared with the capacitance to ground in the CIP method. Both L and T perform very well in the CIP method but are somewhat limited in the 220A method. Calculate the frequency at which the inductor is 50 ohms. This will be the starting point where the inductance will start to function. This explains why most filters are the pi type required to pass the 220A specification.

Summarizing, watch the inductors of the L or T filters in the 220A test method. See the frequency at which they reach 50 ohms.

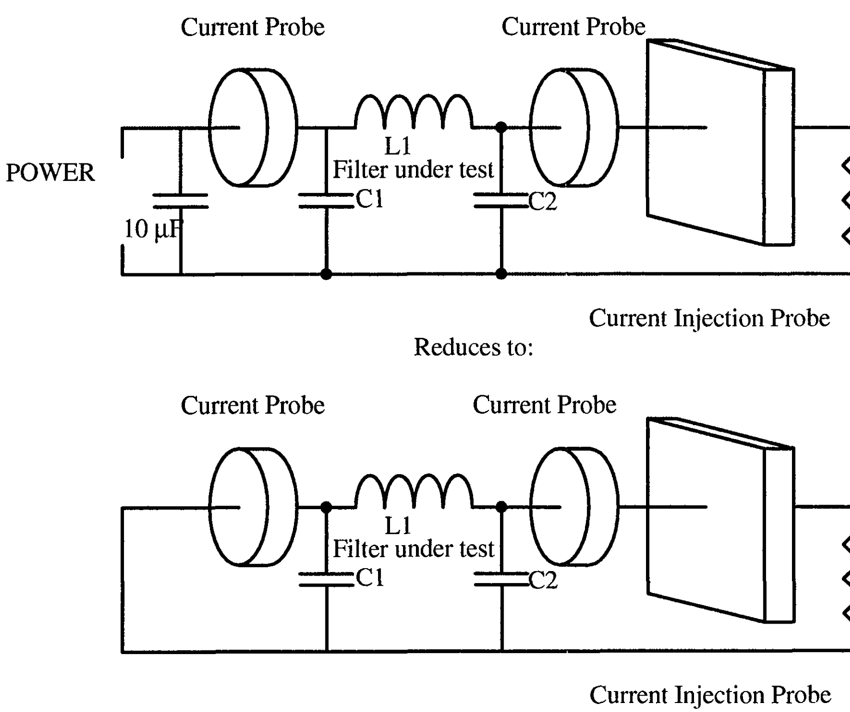

FIG. 1.6 CIP test setup.

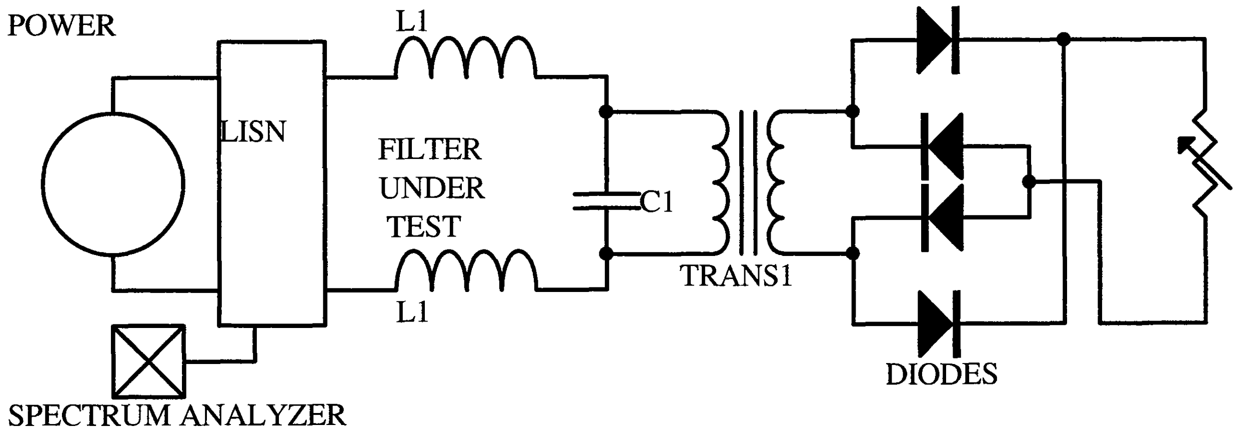

FIG. 1.7 The diode test method.

1.7. THE 400-HZ FILTER COMPARED WITH THE 50- OR 60-HZ FILTER

The problem with the 400-Hz power frequency is the voltage rise. Again, we are speaking of a system requiring substantial serious loss and load current. It is the cutoff frequency of the EMI filter that creates the problem. There is always a substantial voltage rise ahead of this cutoff frequency that pushes up the frequency spectrum at 400 Hz. This creates a severe voltage rise at 400 Hz. At 60 or 50 Hz, the rise is so much smaller that other factors will compensate for it, making the output voltage the same as the input or a slight voltage drop. This will be handled later in the book, but the main answer is getting the cutoff frequency as high as possible. This requires multistage filtering and impedance matching. For the same amount of loss (80 dB at 100 kHz), as the number of stages grows, the cutoff frequency increases, dropping the gain at 400 Hz. However, as the current level decreases through the filter, this often enhances the voltage rise problem. There is also a technique referred to as RC shunt that decreases resonant rises, say here at 6000 Hz. The resonant rise could be 10 to 15 dB. The resistor could be 10 ohms (covered later in the book) and the capacitive reactance at 6000 Hz would be 10 ohms. Would this lower the resonant rise? Of course it would, but now you have 5 dB at 4000 Hz. So what is the gain at 400 Hz? About the same as before, but I have seen cases where this fix made the condition worse. The resonant frequency dropped in dB but was also moved to a lower frequency, negating the fix.

Another method is the Cauer, or elliptic, filter. In a multistage filter one (or two) of the inductors is tuned to, in this example, 100 kHz. This adds many dB of loss, allowing all the filter component values to decrease until the filter loss is brought back to the needed 80 dB at 100 kHz. This pulls the resonant rise frequency farther away from the 400 Hz and reduces the voltage gain. Of course, each change in the inductor values requires a change in the tuning capacitor so that it still resonates at 100 kHz.

Reviewing, 400-Hz filters with high loss requirements at low frequencies demand special handling to get the loss low enough and still have little voltage gain at 400 Hz. The limit appears to be a 3% tolerance. The near-perfect design via computer will miss the mark but should fall easily within the 3% range. I have seen systems with 118 V in and 126 V out.

1.8. THREE-PHASE FILTERS

There are several styles of three-phase filters. For the larger current requirements, there is the "individual removal filter," which others call the "insert filter." Each filter of this type has large capacitance to ground. They are mounted directly to ground and are usually installed within an enclosure. These EMI filters are called three wire, four wire, and five wire—three for the delta, four for a "wye" with a neutral, and five with an added ground wire. All three-phase filters and, in case of a wye, a fourth filter of the same size for the neutral are included within this enclosure. Any leakage current, reactive current to ground, specification must be removed because these large capacitors are directly tied to ground. This type requires a very good grounding system, and if mounted off ground, the filter will not function properly. RFI Corp. has built these filters for as much as 1000 A per phase. Here, the individual filters must pass the specification—usually 220A—and are mostly of the multiple pi arrangement.

The next style is for the lower current types, typically below 100 A per phase. In some of these, leakage current specifications are given either by a current-to-ground limit or the maximum capacitance to ground. The three phases are in the same box. Here, the design is based mostly on how the filter is going to be tested. If the test is based on testing each leg to chassis (ground), the other filter legs will be grounded on both input and output terminals. Now the capacitors between the lines will work to ground, doubling the capacitance and making it easier to pass the tests. The feed-through capacitors to ground add to this capacity. In common mode (covered later), all the input terminals are tied together, as are the output terminals, so the three are working in parallel. Here, the line-to-line capacitors do not help the common mode loss.

A worse condition exists if the power line frequency also happens to be 400 Hz (see the preceding section). The best possible solution is a multistage in each leg with Cauer filters to bring the resonance rise as high in frequency as possible to reduce the voltage rise.

The common mode inductor should have the neutral (if required) wire wound on the core as well as legs A, B, and C. The unbalanced current will flow through the neutral wire and through the common mode inductor, canceling the unbalance within the common mode core.

To review, if this is a three-phase 400-Hz filter requiring substantial differential mode loss, the easiest to design is the individual removal filter or the insert filter. The types with all phases and the neutral in the same enclosure are the hardest to design. Virtual ground techniques can be used to reduce the leakage currents if the three voltages feeding the three wires are nearly equal. This technique greatly enhances the common mode loss. The virtual ground adds to the feed-through capacitors to ground, adding common mode loss. The question is how much capacitance you want between line to line and line to ground. Often these values are part of the specification. Again, use multistage sections to increase the filter resonant frequency. This will lower the 400-Hz voltage rise.

2

Common Mode and Differential Mode—Definition, Cause, and Elimination

There is a wide range of opinion about the definition, cause, and elimination of common mode noise and differential noise. This should cover most of these ideas.

2.1. COMMON AND DIFFERENTIAL MODE DEFINITIONS

A basic definition of common mode and differential mode is required. Differential mode means the normal transfer of energy down the line. In fact, this is also called normal mode. A voltage across the line with a current flowing in one direction in one wire and the opposite direction in the other wire is normal mode. In this case, the subject is differential mode noise. In other words, it flows just like normal power in the line energy.

Common mode means a voltage impressed across both, or all, lines. This voltage is between all these lines and ground. If there is only one line, then the pulse is still between this line and ground. In this unbalanced case, differential mode and common mode act the same—between line and ground.

A current flows in the same direction in all the lines and the return is ground. Again, the subject is common mode noise. I have found that opinions vary from EMI guru to guru with little agreement. I hope this section will, at least, achieve some agreement among the various groups. Many claim that if the common mode noise voltage impressed on these lines is not exactly equal, then it is not common mode.

Two signals cannot be equal on both lines because of differences between the lines, the different spacing between the lines and ground, different capacitance to ground between the lines, and so forth. Even the EMI filter input feed-through capacitors, MOVs, transzorbs, and the like upset the common mode according to some groups. Therefore, using that definition, common mode does not exist. But common mode does exist, so this definition must be lacking or faulty.

To sum up, differential mode noise voltage is impressed between the lines whereas the common mode noise is across the lines—typically two—and ground (Fig. 2.1).

2.2. WHAT CREATES COMMON MODE NOISE ON THE LINE SIDE?

The simple definition of common mode noise is a pulse of voltage on both power lines of equal value (Fig. 2.1). This pulse is between the power line wires and ground. The EMI filter should be designed to handle this energy. A lightning strike on the power line side will create a magnetic field that will cut the two, or more, power line wires. This voltage is impressed between the lines and ground. This strike will be several quick high-voltage pulses typically around 50 kHz. The spacing between the lines may be 3 or 4 feet, depending on the voltage and location, creating a slightly different voltage in the two power lines. This voltage will be added algebraically to the AC power line voltage on all the lines. All of this section assumes that the lines do not fuse and that transformers will take this pulse without failing. If any failure occurs on the line, this reduces the high-pulse problem at the filter and equipment following, but the power will be down. It will be the difference in the two line voltages feeding the transformer that is transformed to the secondary. This difference between the lines is now transferred to differential mode noise. This difference voltage will be transformed (stepped down) to the user side. There will be extra transformer losses due to the high-frequency core losses. These noise pulses are at higher frequencies, accentuating core losses. The skin effect within the transformer and on the lines where this high-frequency pulse is being conducted adds to the pulse losses. The primary

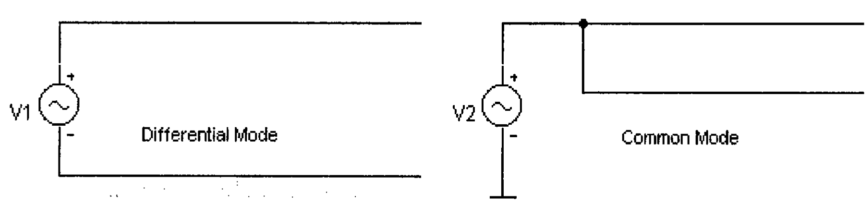

FIG 2.1 V1, differential mode, and V2, common mode.

to secondary capacitance of the transformer (Fig. 2.2) carries the common mode part of this pulse.

However, the secondary has many paths to ground through the winding capacity to the center taps at the transformer and each utility power service panel. The capacity between primary and secondary is further reduced if the transformer has a Faraday shield or screen. The high frequency of the pulse is further reduced by the interwinding secondary capacitance of the transformer.

Most of the voltage transferred from the primary to the secondary through the capacitance will be shunted to various service grounds (Fig. 2.3). It is primarily the difference voltage across the transformer that will carry the pulse to the utility service users. However, this voltage is differential mode and is carried by the two outer legs.

Often, this voltage from the power company is called two phase because it is 180 degrees out of phase. If the equipment works off both outside lines (220 V), the filter must handle this reduced unwanted common mode pulse. Both lines from the power panel are black (sometimes two different colors are used), and a safety green wire goes back to the service ground. Usually, the filter will have three transzorbs (see Chapter 9): one from line to line and the other two from each line to equipment ground and carried by the green wire. These transzorbs will have two different ratings. The line-to-line transzorbs will be rated at 250 or 275 V RMS and the two line-to-ground transzorbs will be rated at 150 V RMS. The transzorbs are rated, or listed, by their RMS value. A Harris transzorb V150LA20B has a rating of 150 V RMS and will fire at voltages around 212 to 240 V. The purpose of the MOVs, or transzorbs, is to eliminate these different noise pulses. The line-to-line MOV, or transzorb, eliminates differential mode and

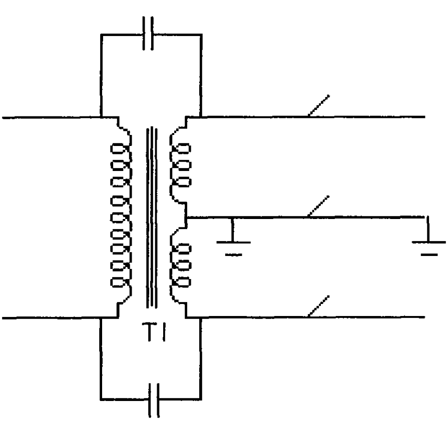

FIG. 2.2 The line transformer with the CM capacitors and center-tap grounds.

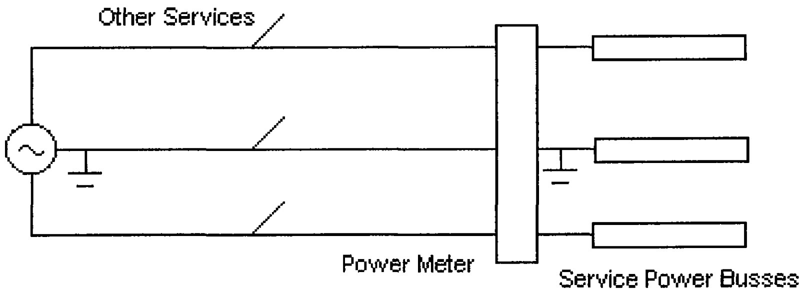

FIG. 2.3 Service, showing other services, power meter, and busses.

the line-to-ground MOVs or transzorbs eliminate common mode and help with the differential mode noise.

If the power for the system works off either side to the central ground, the pulse is differential mode. Most equipment has three lines to it from the service: the hot wire, typically a black wire, tied through the circuit breaker at the service; the neutral, typically white, tied directly to the service common ground; and the safety ground, typically green, also tied directly to the service common ground. The hot black wire carries the unwanted pulse to the equipment and the white common ground carries it back to the common ground. It is the filter's job to handle this pulse of noise that is now differential mode noise. Typically, the filter will have a transzorb from the hot to the filter case, or equipment, ground—the green wire—and another tied between the hot wire and return.

If the pulses are between the transformer and the user, the magnetic pulses cut the wires in the same way but the central wire is grounded repeatedly at all the services and transformer. Then the two outside lines are carrying the common mode pulses. If the equipment is tied to the outside lines, plus the grounded green wire, the filter must handle the common mode problem. The three transzorbs will be sized as before with two rated above 120 V RMS and the one from line to line rated above 220 V RMS, typically 250 V RMS. If the equipment is tied from one line to ground, the noise energy is differential mode carried by the hot and return wires, typically black and white leads. The transzorb would be rated above 120 V RMS.

2.3. WHAT CREATES COMMON MODE NOISE FROM THE EQUIPMENT SIDE?

Storage capacitors in most power supplies are hooked between the diode outputs and ground. This wire is the chassis green wire from the service ground. As the voltage on the storage capacitor raises and falls with respect to ground, the in-coming power lines (black and white wires) follow this with respect to the ground. This creates common mode from the equipment side back toward the line. Switchers do the same thing. An input transformer would eliminate this, and so would power factor correction circuits. Isolating the input supply from ground by placing the storage capacitor across the diode bridge and then following with an isolated switcher also works to remove the common mode noise.

Figure 2.4 shows the isolated supply with the storage capacitor (SC), the switch (Q1), and the load resistor (R1). Besides removing the common mode problem, the ground (green) wire has little current on it, referred to as leakage, or reactive, current. The EMI filter in front of this supply can be balanced with the differential capacitors from line to line, not to the green ground line, and leakage current will be minimal. Only small capacitors to ground, feed-through style capacitors, with a common mode choke will handle the common mode problem if the system is working on the 120-V side. More common mode loss is required if the system is off the 220-V side. However, larger capacitors to ground may be used because the system is balanced. See Secs 2.4 and 2.5.

2.4. WHAT ELIMINATES COMMON MODE NOISE FROM THE LINE AND EQUIPMENT?

Because common mode noise is between the lines and ground, capacitors to ground are required. Also, common mode inductors are used and will be discussed later. Here, the reactive capacitor current to ground, also called leakage current, will be the main subject. This ground current is the difference in current in the ground capacitors from both sides of the lines. A good isolation transformer eliminates both the leakage current and the common mode noise.

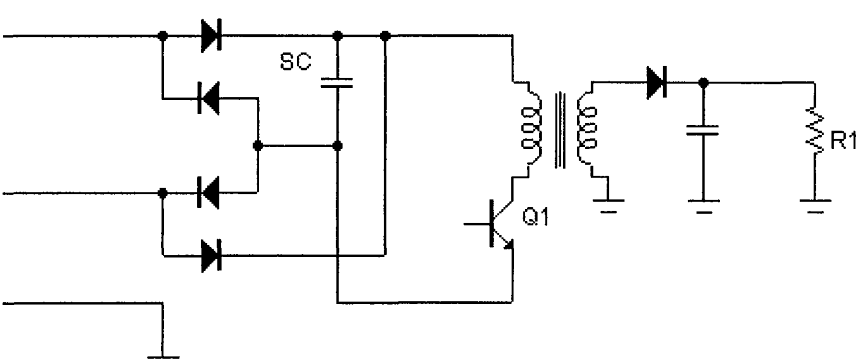

FIG. 2.4 Simplified isolated input section of the power supply.

Figure 2.5 shows two capacitors to ground for 220 V AC balanced lines. These capacitors could be feed-through type or leaded capacitors. If the voltage is equal and opposite (180 degrees out of phase) and the capacitors are equal, the ground current is zero at the line frequency. This works for capacitors with leads as well as for feed-through capacitors. If you have read the beginning of this chapter, you are aware that there is no such thing as 100% balance, so there will be some current to ground. An example would be 115 V from line to ground, 230 line to line, 5% capacitor tolerance, 5 mA allowed to ground and 60 Hz. Note that if the capacitance is a limit specified by the customer and the customer simply takes a capacitor measurement from either line to the common ground, this method will not pass. If the current limit is specified, isolate the filter and load from ground and measure the green wire current, and this method will pass. The current through either capacitor to ground is

$$I_1 = 2V\pi FC_1$$

$$(I_1 - I_2) = 0.005 = 2V\pi F(C_1 - C_2)$$

The difference in the capacitors is double the percentage, here 0.1C (one 5% high and one 5% low),

$$\frac{0.005 \times 10^6}{2V\pi F} = \frac{5000}{2V\pi F} = 0.1C$$

$$C = \frac{5000}{0.1 \times 2 \times 115 \times \pi 60} = 1.15\ \mu F$$

changing the capacitor value to μF and substituting the 0.1C.

Therefore, two feed-through or leaded capacitors of 1 μF will work well if the capacitor tolerance is 5% or less. Remember, the values used in this configuration works only in the line-to-line system and a much smaller single value of capacitance to ground would be required in the 120 V to ground arrangement to meet this low current value. Also, this should never be used for medical equip-

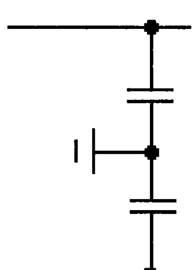

FIG. 2.5 Balanced line-to-line capacitors for 220-V balanced lines.

ment touching patients. If one of the wires opens, the full line-to-ground voltage (120 V) of the one remaining line is impressed across the one capacitor and the current to ground through it is 45 mA, well above the patient limit.

Another arrangement for the two-phase balanced system, which also eliminates using the feed-through capacitors, is to replace the ground at the common point with a capacitor to ground. This is used for common mode attenuation. Here V is the line-to-ground voltage, C_1 and C_2 are the line-to-common capacitors (assuming C_1 is the larger of the two), C_3 is the junction-to-ground capacitor, e is the junction voltage, and I_L is the maximum leakage current to ground requirement through C_3. The equations follow. If t is the tolerance, here from above 0.05, then the maximum difference in the capacitance is $2tC$, but the addition of these two is assumed to be $2C$. Assuming a common mode pulse of equal amplitude on both lines to ground, the two lines to junction add (in parallel), giving $2C$. The total capacitance to ground would be the two in parallel and then C_3 in series. Simplifying,

$$e = \frac{V(C_1 - C_2)}{(C_1 + C_2 + C_3)} = \frac{2VtC}{2C + C_3}$$

$$C_3 = \frac{I_L(C_1 - C_2)}{2\pi FV(C_1 + C_2) - I_L}$$

$$C_3 = \frac{2tI_LC}{4\pi FVtC - I_L}$$

$$C_T = \frac{I_L}{2\pi FVt}$$

The most practical solution along with the best overall performance is to have the three

$$C = \frac{3I_L}{4\pi FVt}$$

capacitors equal. Make C_3 equal to C in the preceding equation. Round down the three capacitors to a convenient standard value. Just make sure that the tolerance limit of the two line-to-junction capacitors is correct. The junction to ground is not as critical. Another method, but a costly one, is to sort, or grade, the capacitors into smaller difference percentages. Use the matched ones for the two line-to-junction capacitors and the oddballs for the junction to ground.

A difficulty arises if someone uses this 220 line-to-line filter for a 120-to-neutral filter. Now the junction-to-ground capacitor and the junction-to-neutral, or the return, capacitor are in parallel, and the voltage from junction to ground is V divided by 3. In this situation, the current through C_3 to ground is

$$I_G = \frac{3VI_L 2\pi F}{3 \times 4\pi FVt} = \frac{I_L}{2t}$$

where I_L is the original design leakage current and I_G is the new ground current and it is assumed that the actual calculated values of the capacitors are used. Otherwise, I_L is the resultant leakage current, which is lower than the specified value. In the preceding case, with I_L equal to 0.005 and t equal to 0.05, I_G is equal to 50 mA. This is well out of specification. Hopefully, the leakage current is not specified for this requirement or is a larger value for this application. A 220 V AC balanced filter, designed as above, should be marked on all documentation and the system should not be used for a 120 V AC line-to-ground application.

If the system is being built in house where the filter, power supply, and the rest of the system are under engineering control, build the filter in as part of the supply and design a current transformer for the ground lead of the capacitor. Use this to shut the system down if a voltage imbalance occurs or there is excessive current to ground for any reason. Put this network reasonably close to the line input with a current transformer on either side of the capacitor to ground. Design this to operate a relay that opens the system after the filter. I would suggest that the relay energize only in the excessive ground current.

Summarizing, the balanced line for 220 V AC for two phases with leakage current specifications can be met with the three-capacitor arrangement. This technique was developed for common mode noise. The normal voltages are 180 degrees out of phase whereas common mode is in phase. So the capacitors buck for normal mode and add for common mode. This adds to the loss of the common mode inductor. It is not advisable to use this in any medical equipment that would touch a patient. A way around this is a current probe that would shut down the equipment, and it would be better if this also opened the leads to the patient. If this cannot be done, mark all documentation never to use this system on a 120 system working between line and ground or for any medical application.

This application is similar to the virtual ground for three-phase systems.

2.5. WHAT CREATES DIFFERENTIAL MODE NOISE?

In Sec. 2.2, the common mode pulse was discussed. A transformer was placed between the lightning strike and the filter. For the most part, this created differential mode energy at the filter. Thus, it is recognized that common mode can create differential mode noise. I am not familiar with the reverse condition. On the line, differential mode noise is also created with inductive equipment on turn on and mainly turn off. This is a voltage pulse between the lines. Likewise, transformers between the output of the filter and equipment, or the isolated switcher transformer (see Fig. 2.4), produce differential mode for the filter to

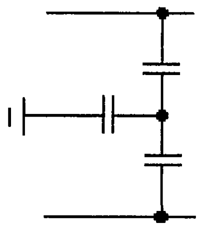

Fig. 2.6 Balanced line-to-line capacitors with extra capacitor.

handle. For the line side, MOVs, or transzorbs, help to clip the higher pulses from the line and equipment, and the differential filter section must handle the rest.

2.6. THREE-PHASE VIRTUAL GROUND

This technique can be used only for the three-phase type all within the same enclosure. Types requiring individual insert filters with all the capacitance to ground cannot use this technique.

This is very similar to the preceding two-phase application. A capacitor is tied from each phase to a common point, making a virtual ground. If the RMS of each phase voltage is the same (this is not a function of load current) and the three capacitors are the same, then the junction voltage is zero, or a virtual ground. If a fourth capacitor is tied between the junction and ground, the current through this capacitor is zero. If the unit is tested for ground current—not capacitance to ground—by isolating the equipment (or dummy load) and the filter from ground and measuring the current on the ground wire, typically green, the current should be well below the specification. Again, the capacitor values are equal for best overall results. Solve equations similar to the preceding equations. The voltage at the junction is equal to the line voltage if one phase fails and the phase angle will be between the two remaining phases—60 degrees from each. Therefore, this is risky for medical use involving application to a patient. However, usually three-phase high-power equipment never touches the patient. Use a current probe to monitor the current and open some relay to remove the power from the equipment or the patient or both. The latter assumes the engineer has control over this and not just design of the EMI filter. This situation is similar to that in Sec. 2.4, and it would be advisable also to read this section.

The best part of this is that the three-phase voltages give nearly zero current to ground, but for the common mode voltages, the three capacitors are in parallel with the fourth in series. This gives very low impedance to ground to eliminate the common mode noise.

This technique plus other small feed-through capacitors along with a properly designed common mode inductor will surely eliminate common mode problems. Remember, if there is a neutral wire, there should be equal windings for the three phases plus an additional winding for the neutral wound on the ferrite common mode inductor. Any unbalance in the phase currents is carried by the neutral, so the total magnetic flux will be zero in the common mode core. The theory is that the neutral current is low in a balanced system. Any multiple of the third harmonic that exists (3, 6, 9, etc.) in the system adds back in phase and adds on the neutral. So, the total third-order currents and the unbalanced current add on the neutral. Make sure the wire size can handle at least the peak phase currents. I have had many calls about neutral filters overheating well above the temperature of the other phase filters, and this is due to unbalance current and the third-order harmonics. Measure the current on the four wires to see if the neutral current is not well above the other three.

The equation for the leakage current follows. E_1 and C_1 are the voltage and capacitor associated with the first phase, and so on with E_2, C_2 and E_3, C_3, while C_4 is the capacitor between the junction and ground.

$$\frac{2\pi F C_4\left[E_1 C_1 \sin(2\pi F t) + E_2 C_2 \sin\left(2\pi F T + \frac{2\pi}{3}\right) + E_3 C_3 \sin\left(2\pi F t + \frac{4\pi}{3}\right)\right]}{C_1 + C_2 + C_3 + C_4}$$

To review, what removes common mode noise? The answer is common mode inductors, capacitors to ground, transformers, and transzorbs.

3

EMI Filter Source Impedance of Various Power Lines

One of the leading questions asked by people who have an EMI problem or are joining the EMI field is related to the source impedance. Several senior scientists and doctors of engineering have asked about the transfer function of a certain filter that RFI Corp. proposed for their application. They were going to measure the line impedance along with the load impedance and wanted to know the transfer function of the EMI filter that was designed in. The transfer function can be calculated, but most EMI filter manufacturers rarely measure them. If the filter manufacturer, here RFI Corp., could provide this information, this company was going to add another section between the filter and the load. The purpose would be to notch out resonant rises and other instabilities of the entire system. This procedure would work very well if they were assured that the line length, conductor spacing, or diameter of the conductor would not change much from installation to installation. These parameters require precious time to determine and could require expensive rental equipment for most companies. This would work well if their system, using the EMI filter device, were to be installed on one type of ship or aircraft, where these cable dimensions would be nearly the same. The technique would fail to work if the unit was installed on different vessels or used in various other applications. Also, they would need to have a look at all the possible load conditions, not just the peak load. If the device would go into some standby mode, the instabilities could shift and the filter along with the rest of the equipment could oscillate.

The harmonic content of the power line frequency varies from line to line. In the past, most commercial lines had little harmonic content because of the very low line impedance at these frequencies. This was discussed with a large power company in Southern California in 1985. They stated that 85% of the power was then used for "power and lights," that is, motors and lamps and, at that time, incandescent lighting. With the proliferation of computers, computer printers, scanners, copiers, fax machines, televisions, sound systems, and similar equipment now tied to these lines—and very few with power factor correction circuits—the voltage waveform is less sinusoidal. This is due to the voltage drop caused by the high current spikes that these machines demand a little ahead of 90 and 270 degrees. The power factor correction capacitors placed across the line by the power company do help to reduce the current gulp, but the voltage is still less sinusoidal. The harmonic content of generators at remote sites and shipboard installations is much greater because of the higher resistance of the generator and lines. The voltage supplied to the end users in these applications is less sinusoidal. A power consultant in Southern California found 100-A spikes above the nominal sinusoidal current in small office buildings that had power problems. Again, these spikes were a little ahead of 90 and 270 degrees, typically around 85 and 265 degrees.

3.1. SKIN EFFECT

As the frequency increases on the line, the depth of conduction is reduced. The wire cross-sectional area decreases because the radius of conduction decreases. The higher the AC resistance, the greater the dissipation of this unwanted energy. Skin effect can take its toll on the higher frequency energy on the power lines. This helps dissipate the Electromagnetic Pulse (EMP) and other higher frequency noise traveling on the power line in either direction. These power lines were constructed to handle power at very low line frequencies and not for the higher frequencies creating the loss. Although the characteristic impedance of the line may be 150 ohms, the loss of the line per unit length increases with frequency. For copper the equation of the skin effect depth in centimeters is

$$D = \text{depth (cm)} = \frac{6.61}{\sqrt{F}} \tag{3.1}$$

The cross section of the conducting area (CA) of the wire for frequencies above the skin depth is

$$\begin{aligned} \text{CA} &= |R^2 - (R - D)^2|\pi \\ &= D(2R - D)\pi \end{aligned} \tag{3.2}$$

where R is the radius of the wire in centimeters and D is the skin depth from above, also in centimeters. As the frequency increases, D decreases such that the D term is much smaller than 2R. Equation (3.2) at these upper frequencies is reduced to

$$\text{CA} = 2\pi RD \tag{3.3}$$

The original cross-sectional area, πR^2, compared to the value of CA, times the original DC resistance, will give an approximate value of the AC resistance at these upper frequencies due to the skin effect.

$$\begin{aligned} R_{ac} &= \frac{\pi R^2 R_{\text{dc}}}{2\pi RD} \\ &= \frac{RR_{\text{dc}}}{2D} \end{aligned} \tag{3.4}$$

the lowest frequency would be well above the value of

$$F = \frac{6.61^2}{R^2} \tag{3.5}$$

Below this frequency, the skin effect is the full radius of the wire. The frequencies we are discussing here should be several times this lower frequency. Replacing D with its equation above,

$$R_{\text{ac}} = \frac{RR_{\text{dc}}\sqrt{F}}{13.22} \tag{3.6}$$

If R_{dc} is the resistance in ohms for a small distance along the line, then R_{ac} will be the approximate AC resistance for this short section. R_{ac}, along with L, C and G, the conductance across the line, will form a short segment of this line, say a meter, Then the characteristic impedance of the line is

$$Z_0 = \sqrt{\frac{R_{\text{ac}} + \omega L}{G + \omega C}} \tag{3.7}$$

Substituting (3.6) into (3.7):

$$Z_0 = \sqrt{\frac{0.07564 RR_{\text{dc}}\sqrt{F} + \omega L}{G + \omega C}} \tag{3.8}$$

Simplifying yields:

$$Z_0 = \sqrt{\frac{0.07564 RR_{\text{dc}}\sqrt{F} + 2\pi FL}{G + 2\pi FC}}$$

deleting G because it is much smaller than $2\pi FC$ at the frequencies discussed in this section. The square root of F is removed in the first term in the numerator:

$$Z_0 = \sqrt{\frac{0.07564RR_{\mathrm{dc}} + 2\pi L\sqrt{F}}{2\pi C\sqrt{F}}} \tag{3.9}$$

To find the limit with respect to frequency (F), differentiate the numerator and the denominator separately:

$$\sqrt{\frac{0 + \dfrac{2\pi L}{2\sqrt{F}}}{\dfrac{2\pi C}{2\sqrt{F}}}} = \sqrt{\frac{2\pi L}{2\pi C}} = \sqrt{\frac{L}{C}}$$

The first term of the square root numerator vanishes and the F's in the last term of the numerator and denominator cancel.

$$Z_0 = \sqrt{\frac{L}{C}} \tag{3.10}$$

Therefore, the skin effect term has little or no effect on the characteristic impedance. The values of L and C are the dominant terms. This is the fundamental equation of the characteristic impedance of coaxial cables. This all shows that the normal characteristic impedance equation still dominates at the higher frequencies and this characteristic impedance does not change with skin effect, although the loss per unit length does.

This behavior is similar to that of coaxial cable except that coax is designed to handle higher frequencies. There are many different coaxial cables with the same characteristic impedance. These lines have different inside and outside diameters. Some coax lines have very small diameters and others have larger diameters. The dB loss per 100 feet varies from coax to coax and also varies with frequency. Think of it as a pad. The impedance of the pad may be 50 ohms but the loss of the pad varies from pad value to pad value. The main difference here is that this power line impedance also varies with frequency. It is this line impedance that dissipates the unwanted energies, not the characteristic impedance of the transmission line.

Skin effect also applies to the wire used for the inductors, the transformers (if used), and the rest of the EMI filter wiring. The purpose of the filter is to rid the system of unwanted signals or noise. What better way is there than to dissipate it? Use the skin effect to do this. For the filter inductor, never use Litz wire or strands for the turns. Allow the wire to help dissipate these upper frequencies. The same is true for the rest of the hookup wire. Is there an exception? Yes. Some larger current filters have capacitors in parallel to ground. Here, Litz or braided wire should be used so that the capacitors can shunt these signals to ground. Otherwise, the capacitors will be limited in performance and have a lower self-resonant frequency (SRF) due to the lead inductance.

It makes no difference how long the inductor leads are, and they are always the self-leads of the inductor. This lead is often a single strand as covered earlier. The lead's properties add to the loss of the inductor. These leads add to the inductance, and their skin effect also adds to the wire loss. The square blocks in Figs. 3.1 and 3.2 represent quality feed-through capacitors in the range 3.5 to 10 μF. The oval shapes stand for larger capacitors in the range 10 to 30 μF that provide low-frequency loss but for best results should be a braided wire. Figures 3.1 and 3.2 are top views.

The wires from the inductors are self-leads, but the wires to the oval capacitors should be better quality braids. If the inductor self-lead is attached to the capacitor, this is fine. But the capacitors to ground need the lowest impedance to work. Leads with inductance and skin effect between capacitors lower the SRF of the capacitor, degrading the performance.

Figure 3.3 shows a terminal hooked directly to an inductor. This lead could have been longer. The inductor's longer lead is tied to a feed-through that is part of the output terminal. However, there is a second capacitor tied in parallel to the feed-through. This last lead should be a braided lead that offers a very low skin effect and low inductance. Even with a quality lead, this lead should be as short as possible. There is no perfect lead here, and the cure is to make these leads as short as possible. This larger high-voltage style of capacitor is soldered to the container. Try to place this capacitor as close to the other component as possible to keep the tie leads short. See veeing the capacitor in Chapter 7.

3.2. APPLYING TRANSMISSION LINE CONCEPTS AND IMPEDANCES

Transmission line concepts (Fig. 3.4) may sound like a strange subject to introduce when discussing EMI filters, but the filter designer needs to have a basic understanding of this subject for several reasons. The first is to understand why high impedance is needed at the filter input end for EMP applications. The second is to understand that the power line energy loss is due not to the characteristic

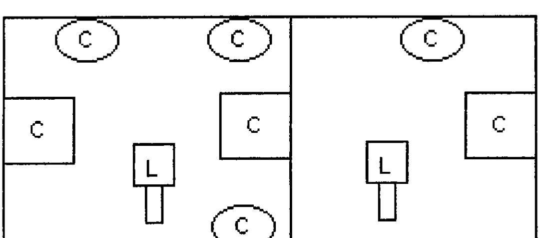

FIG. 3.1 Typical high-current power line filters.

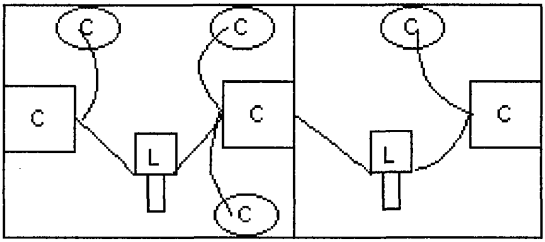

FIG. 3.2 Internal wiring of typical power line high-current filter.

impedance of the line but to the resistance elements along the line. These losses are due to elements such as skin effect, the DC resistance of the lines, and the conductance across the line. The third reason is that, especially in multiple-element filters, the filter has characteristics similar to those of transmission lines.

The shorter power lines start taking on their varying characteristic impedance at much higher frequencies. The characteristic impedance of these power lines varies all over the place; it varies with frequency and is not as constant as that of coax, twin lead, and twisted pair. The characteristic impedance of the open wire type normally varies typically between 90 and 180 ohms due to the spacing between the conductors and the diameter of the wire. The paired type, or twisted wires, enclosed in conduit is 50 to 90 ohms for the same reason as before with the added capacitance between the wires and the conduit. The conduit adds little shielding because of the thinness of the material. In about 95% of the cases, several different power line sections will be in tandem. These different power line impedances vie back and forth as the different power sections approach resonant lengths. The velocity of propagation is very low due to being constructed to carry only the power line frequency, not radio frequency. The electrical lengths of these lines appear to be about eight times their actual length. If the power line is struck with a pulse, the power line will dissipate some of this energy in the resistive

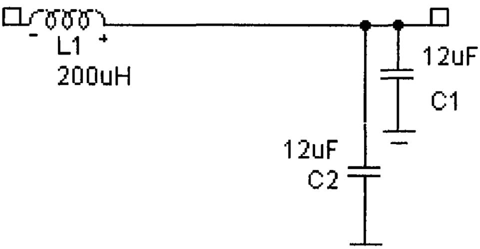

FIG. 3.3 Lead layout example.

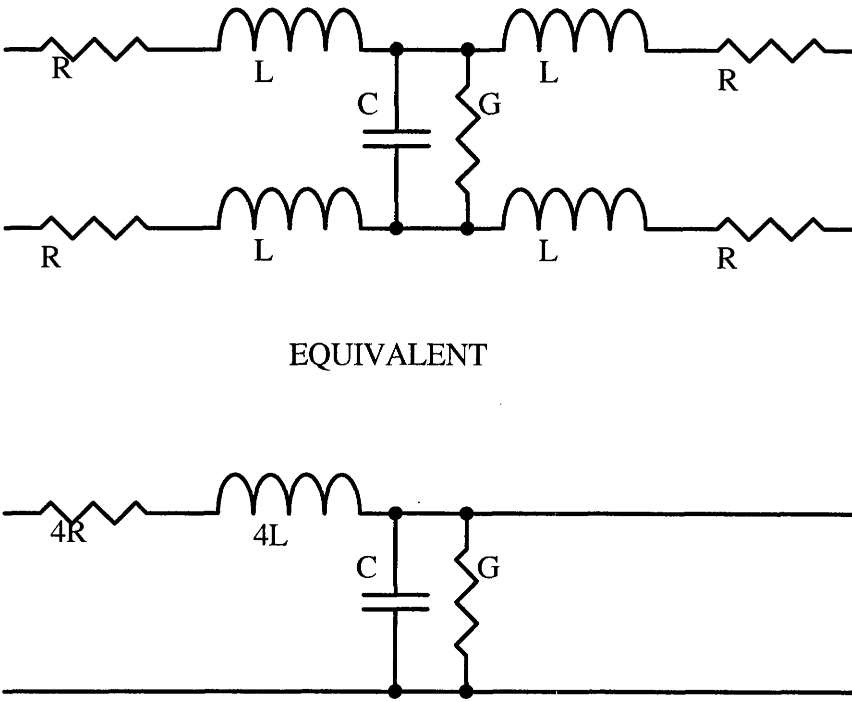

FIG. 3.4 The transmission line and differential mode equivalent.

elements mentioned earlier. The pulse will travel toward the filter end at the lower velocity of propagation of these cables. The fundamental frequency is around 50 kHz for lightning and soon assumes the characteristic impedance of the power line cable rather than the free-space impedance of 377 ohms. If the filter input impedance is high or looks like an open impedance to this pulse, the voltage soars (could double), aiding the arrester to function quickly. If the filter impedance is low, or a short to this pulse, the voltage drops to zero and the line current doubles. This slows or keeps the arrester (varistor or MOV or transzorb) from functioning. This is covered in Chapter 9.

I have seen graphs from the IEEE literature showing some line impedance spikes of almost 500 ohms at one frequency in the MHz range. Figure 3.5 shows the typical impedance range that the various types of power lines fall between for various twisted-pair cables. The graph is like an average of impedances, and the variations fall primarily between the two curves shown. A very minor percentage was above or below the curve.

The major part of these losses is due in part to the skin effect and the DC resistance, especially at the lower frequencies, for the longer lines. Short lines

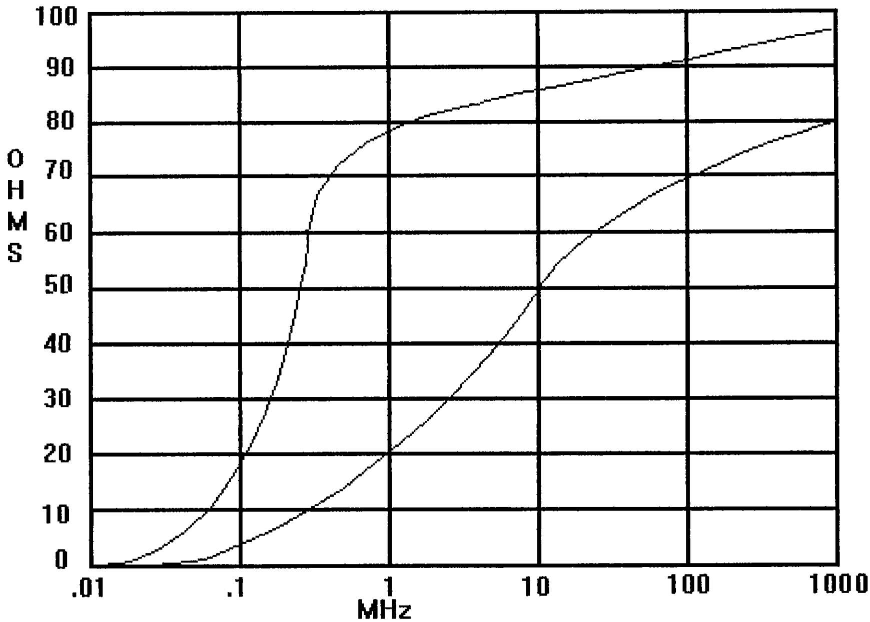

FIG. 3.5 Impedance range of a short line—twisted pair.

appear to form link coupling. The point here is that at frequencies below 10 kHz, both the longer and shorter lines look resistive and close to the DC resistance value. This value is near zero ohms for the better commercial power companies, but for most remote power systems, the resistance can be much higher. As the frequency increases, these lines have an output impedance that reaches 4 ohms at 10 kHz. The impedance then ripples its way to 50 ohms near 100 kHz for the longer lines and 250 kHz for the shorter lines. It should be obvious that the losses required for the Military Standard (MIL STD) 461 specification are such that the filter must be designed to have the proper losses with the line impedance in the very low or the zero ohms region. This should make it clear why the pi filter has trouble meeting the loss for the "real-world" and 461 requirements. The same is true using a 10-μF capacitor across the power line in the common current injection probe testing. The losses that the filter must meet are at frequencies in the region 10 to 14 kHz in the 461 specification and may be as much as 100 dB for power line filters. This is the type of filter used in secure room applications.

The skin effect applies within the filter. Regardless of filter type, often many feet of wires are used, so the skin effect will dissipate the higher frequencies within the filter body. The wires used for inductors are selected to be sufficient to carry the power line current at the power line frequency. The voltage drop and inductor temperature rise will dictate the wire gauge used. Stranded wire is used only to ease winding techniques rather than for high-frequency loss considerations. In fact, these high-frequency losses are beneficial to high-frequency filter loss requirements. In addition to the skin effect losses, core losses at the higher frequencies are beneficial in increasing the filter losses. These are dissipated and not reflected to the source.

3.3. APPLYING TRANSMISSION LINE IMPEDANCES TO DIFFERENTIAL AND COMMON MODE

The transmission line losses for the differential mode (Fig. 3.6) are all the resistive elements of the line including the reciprocal of G, the conductance across the line. These losses include the skin effect and some very minor losses due to the equivalent series resistance (ESR) of the capacitor. The DC resistance (DCR) of the inductor adds no loss because the DCR is already included in the wire—it is the wire. The impedances all along the line in each section will dissipate any differential mode pulse, motor, or inductive load switch, propagating down the line toward the line filter.

The difference in the requirement for common mode (Fig. 3.7) and differential mode attenuation is that G and the ESR of the capacitor have no effect. A pulse traveling down the line, lightning or EMP, toward the filter is partially consumed by the resistance of the line and the line's skin effect. Again, the DCR of the inductor(s) is included with the wire losses (this is the wire). The characteristic impedance also changes because the two lines are in parallel, so the capacitor across the line is out and so is G. The two inductors are in parallel, about 1.5 μH per meter on each line, or 0.75 μH, and the capacitance to ground determines the characteristic impedance. This makes the characteristic impedance quite high. This information explains why the EMI common mode filter section can be made very high.

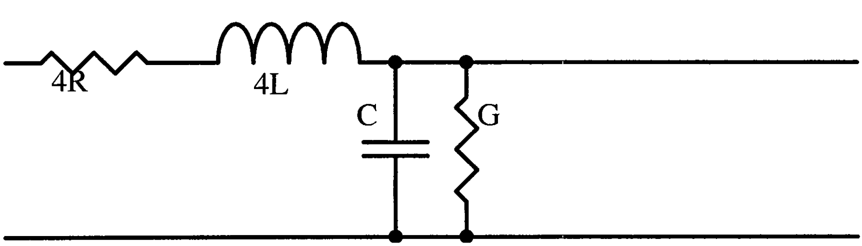

FIG. 3.6 Equivalent differential mode line.

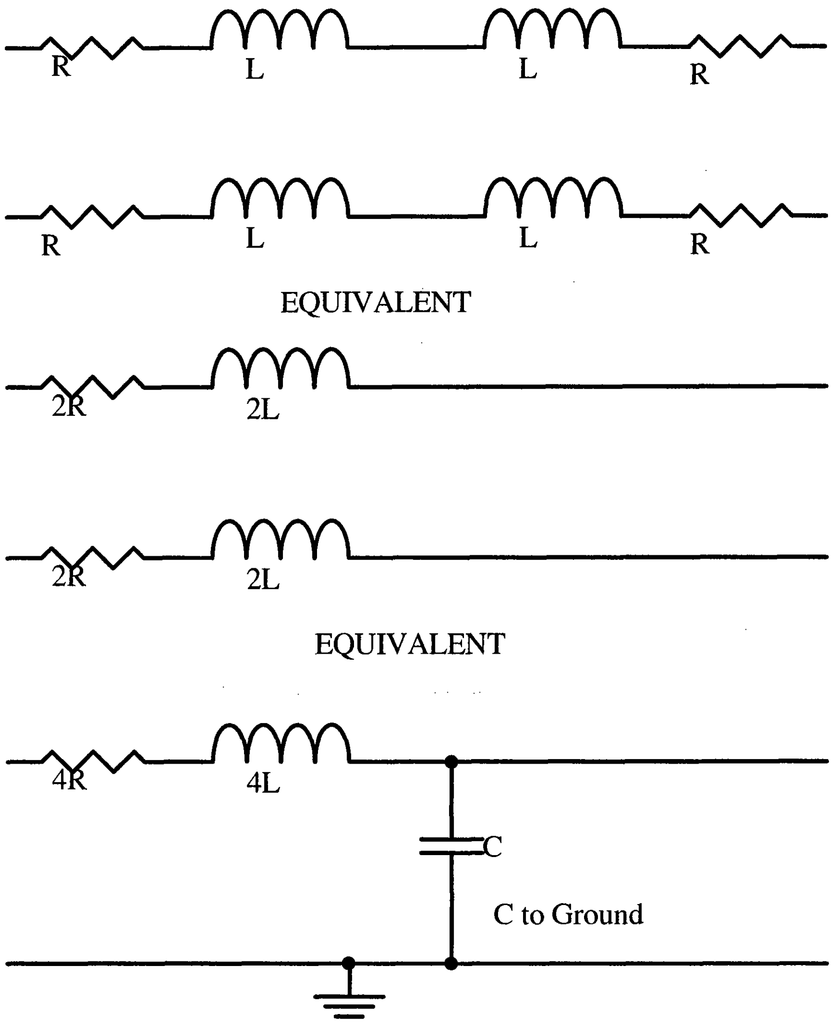

FIG. 3.7 Equivalent common mode transmission line.

3.4. DIFFERENCES AMONG POWER LINE MEASUREMENTS

Many groups have measured the line impedance and there has been little agreement when comparing their results. The reason is that the lines being compared vary so much. The velocity of propagation is quite slow in all these lines, making the lines appear much longer than they really are. In some lines, the ratio is eight times their actual length. Devices such as generators or transformers terminate the line at the higher frequencies. This is due to the capacity across these devices, which shunts the line at these higher frequencies. These lines were designed to carry DC, 50-, 60- or 400-Hz power and not these higher frequencies.

3.5. SIMPLE METHODS OF MEASURING AC AND DC POWER LINES

Power line impedance can be easily measured, with caution, if the designer wishes to do so. There should not be any need to do this, especially when using the techniques used in this book. This measurement can be accomplished with inexpensive test equipment on both AC and DC lines, but it would be much faster and more accurate using a network analyzer. Either line, AC or DC, should be resistive loaded and several runs will be needed. The first readings are with maximum loading, followed by medium loading and then low loading.

To measure the AC line, start well above the line frequency by at least a factor of 10 to keep as much of the AC harmonic power from the readings as possible. This is done to assure accurate readings and to protect the measuring equipment. The frequencies required will be over a wider frequency range well into the megahertz area. A frequency-selective level meter (FSLM) should be used along with a signal generator and a blocking capacitor. The capacitor should be high impedance to the line frequency and low to the frequencies to be measured. If the line frequency was 400 Hz, the lowest intended frequency to be measured should be well above 4000 Hz. If the impedance of the capacitor is 1000 ohms at 400 Hz, the impedance would be 100 ohms at the lowest frequency to be measured. The FSLM should be used in its narrowest input filter selection, if available. The generator frequency should still not be tuned to any multiple, especially the odd harmonics, of the line frequency until 20 times the line frequency is reached. The loss across the capacitor cancels except at the lowest frequencies, anyway, so that the 400-Hz loss could be still higher. The only component remaining is a resistor of some value, say 400 ohms, and this should be noninductive to well above the highest frequency reading. The resistor, capacitor, and signal generator—in that order—are tied across the line with the resistor at the hot end and the signal generator to the neutral. The signal generator is tuned to the desired frequency, the FSLM has its low side tied to the neutral, and two points of measurement are taken. First, the voltage reading is taken between the capacitor and resistor where the FSLM is tuned to peak at the generator's frequency. The second reading is taken at the high line without readjusting the FSLM frequency.

Make sure that the FSLM front end can withstand this AC line voltage without blowing the front end. Many of the frequency-selective voltmeters or level meters have precision input pads that would quickly generate some ugly blue, and expensive, smoke if the hot line is touched. A high-pass filter with a series capacitor input, not an inductive shunt input, could be built in a probe to be installed at the test lead to protect the FSLM's input (Fig. 3.8). The cutoff frequency of this filter must be at least half the lowest frequency to be measured.

This filter does not have to be special or does not require a flat Butterworth wave filter and can have ripple in the passband area of the high-pass filter. The filter errors are canceled as shown in the following equation and seen in the drawing later in this section. The ratio of the two readings will be the same because the errors cancel in the equation at the same frequency. As an example, at 400 Hz, the lowest reading should be 4000 Hz. The reason for the multiple of 10 is to avoid the potential high-level harmonic content on some power lines. The high-pass filter should have a cutoff frequency of 2000 Hz. Then the equations for the values of the inductors and capacitors in the high-pass filter above are

$$L = \frac{Z_{\text{fslm}}}{2\pi F_{\text{c0}}} = \frac{50}{2\pi 2000} = 0.004 \tag{3.11}$$

$$C = \frac{1}{2\pi F_{\text{c0}} Z_{\text{fslm}}} = \frac{1}{2\pi 2000 \times 50} = 1.592 \times 10^{-6}$$

This assumes that the input impedance of the frequency-selective level meter is 50 ohms. Record both of the readings and move on to the next frequency to be measured. The formula is

$$Z_0@F = \frac{V_b R_1}{V_a}$$

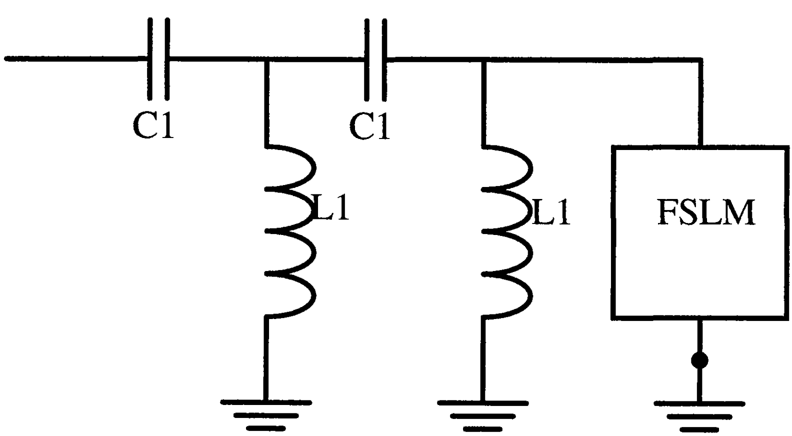

FIG. 3.8 High-pass filter.

The voltage level, V_a, is recorded at F_0 between the capacitor and resistor, and V_b is the high line voltage reading, also at F_0. R_1 is the value of the series resistor between the high side and the capacitor. F is the high-pass filter installed ahead of the FSLM to protect the input pads. The impedances at these various frequencies can be plotted for the various loads and compared (Fig. 3.9). It is often better to use a battery-operated FSLM and signal generator to avoid ground loops.

The DC measurements are almost the same, except that lower frequencies can be read and the highest frequency needed will be in the lower kHz range (compare Figs. 3.10 and 3.11). Obviously, the high-pass filter is not required. These resistive readings can again be plotted on the graph. The readings will level out (resistive) at a low frequency in the kilohertz range. Here, you could replace the FSLM with an AC voltmeter. This setup does not require as much caution as in the AC readings. Make sure, again, that your FSLM or AC voltmeter can handle the DC output or line voltage. If they cannot, simply use a capacitor (this replaces the high-pass filter) in series with the meter probe to take both readings. Figure 3.12 shows the changes for the DC line showing either a frequency-selective level meter or an AC voltmeter and the blocking capacitor, if required. The same equation holds. Again, the various impedances would be plotted on the

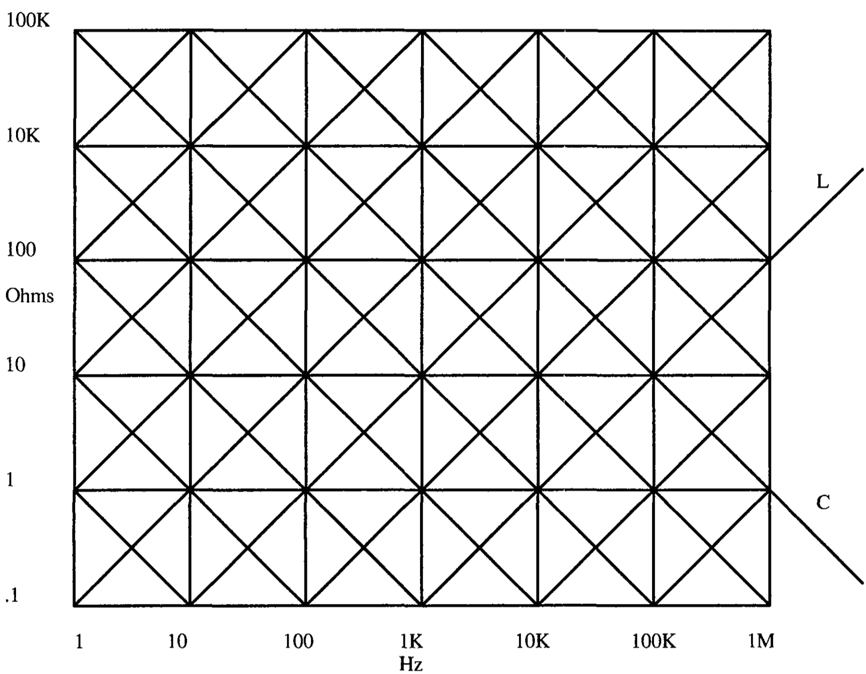

FIG. 3.9 Enlargement of impedance-frequency chart.

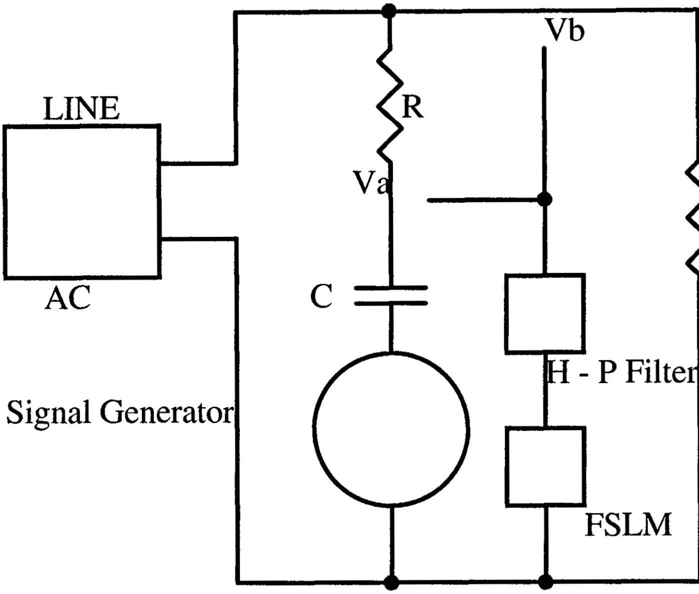

FIG. 3.10 AC line impedance measuring setup.

frequency graphs. These DC plots will normally start with readings in the low milliohm range at the lower frequencies up to around 10 Hz, followed by a slope of approximately 45 degrees upward (inductive). In some power supplies, the lower frequencies may have negative resistance readings. These plotted points will be on both sides of the best straight line, but the best straight (45 degree) line can be drawn and the inductance read directly off the frequency-impedance graph (Fig. 3.12).

$$Z_0 = \frac{V_b \times R}{V_a}$$

The DC power supply readings will be a smoother line, making it much easier to read directly off the graph, and flatten out at a much lower frequency. These points will then level out flat at the upper frequencies. The upper flat area, in ohms, is then the line output impedance of these upper frequencies. The operating frequencies of the switchers, clocks, or oscillators, which this power supply feeds, will fall well into this upper frequency range. This could starve some circuits being powered by this supply. A line simulation network (Fig. 3.13) can be developed using the value of the low ohms in series with the inductor read

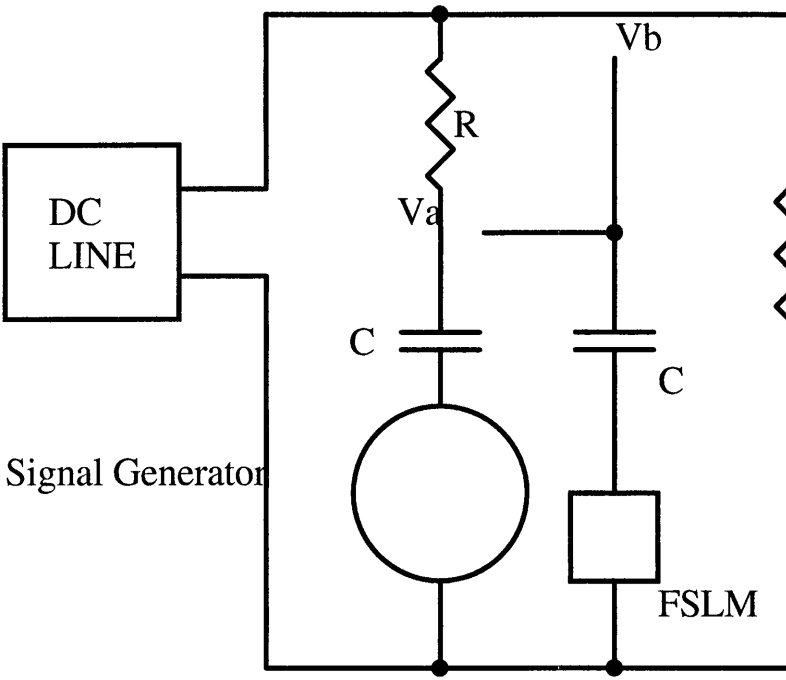

FIG. 3.11 DC line impedance measuring setup with equation.

from the graph. Then these two series elements will be in parallel with the higher upper frequency ohms.

Some power supply designers "cure" this problem with a capacitor tied across the output. This capacitor can oscillate with the inductance, read from the graph, causing ringing at the output of the supply. The best way is to calculate the maximum resistance that the following circuits can tolerate, say 2 ohms. Draw the 2-ohm line across the same graph used above, and read the frequency where the two lines cross. The first line is the 2-ohm line and the second is the inductive line from the graph, and say this cross point is 1800 Hz. Divide this by 2.5, giving 720 Hz here. Then calculate the capacitor needed to equal 2 ohms at this frequency. The answer for this problem is 110 μF in series with the 2-ohm resistor tied across the power supply output line. If the curve was plotted again using the same setup as above, the curve would start in the same way and head up the 45 degrees along the inductive direction to above 720 Hz, then head 45 degree downward in the capacitive direction and flatten out on the 2-ohm line. A good source for this is the catalog *Power Conversion Design Guide and Catalog,* Calex Manufacturing Co., Pleasant Hill, CA 94532.

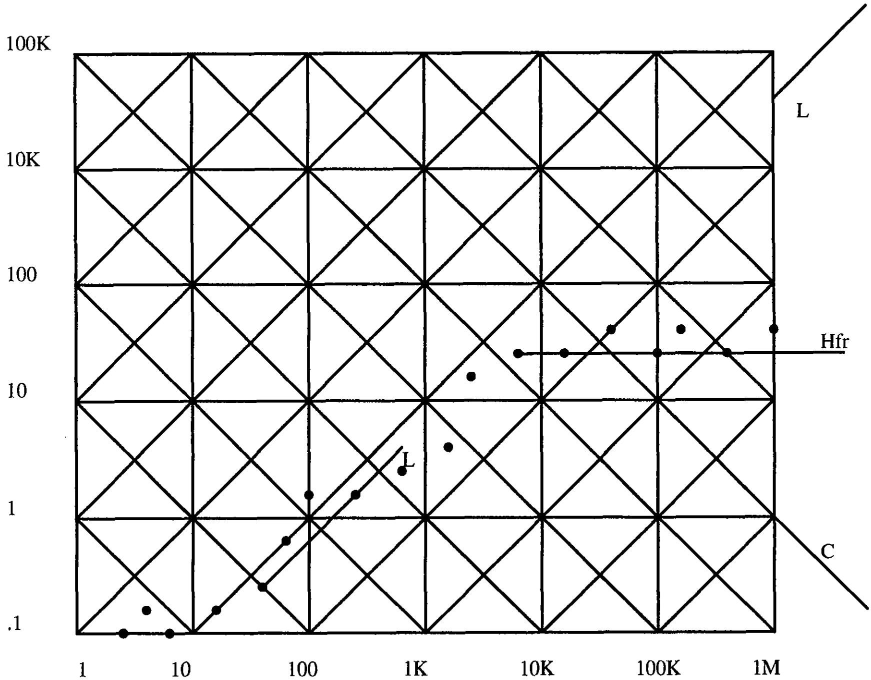

FIG. 3.12 Impedance-frequency chart with plots.

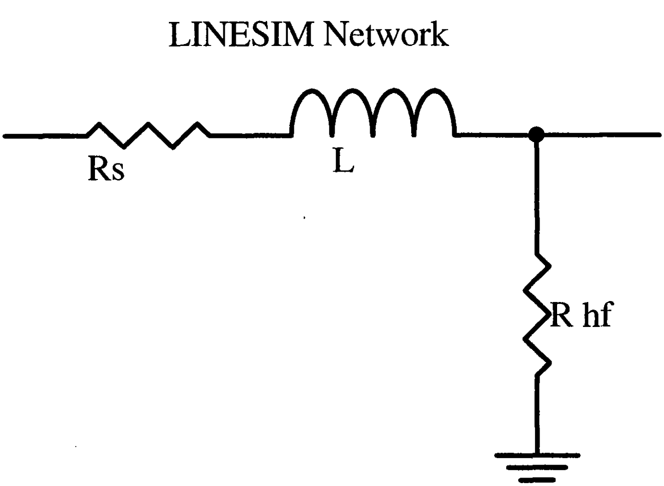

FIG. 3.13 Simulation network.

The output impedance of these supplies is important in EMI design because the wrong simulation impedance network could be used. Without the decoupling capacitor and resistor, the line simulation network looks like a small milliohm resistor in series with a small inductor. These two components are paralleled with a resistor value of the upper flattened out resistance. With the decoupling network used, the output could be replaced with a fixed resistor, as in our case above, using 2 ohms. The best way to measure the line impedance is still with a network analyzer if you have access to one.

3.6. OTHER SOURCE IMPEDANCES

A lightning strike will generate common mode energy between the line, or lines, and ground. Assuming this occurrs some distance down the line, the voltage divided by the current will be the characteristic impedance developed between the line and ground. This is the square root of the inductance divided by the capacitance between this line and ground. The inductance is around 1.5 μH per meter, and the capacitance per meter would be less than 1 pF. It should be obvious that this impedance will be quite high.

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4

The Various AC Load Impedances

The load impedance varies with different types of loads. Very few loads are truly resistive. Most are capacitive or inductive. Some loads generate high-current pulses at twice the line frequency, whereas others require high-frequency currents. All these statements, and the rest of the information in this section, assume that none of these loads have any input filtering or any other circuitry to offset the conditions mentioned here.

4.1. THE RESISTIVE LOAD

The resistive load is the easiest for the EMI filter to handle. This assumes that the storage capacitor is left out in the circuit in Fig. 4.1. The only error would be the crossover error. The diodes are missing in Fig. 4.2, although either circuit can be with or without the transformer. The power factor of the load is near unity and the inductors of the EMI filter can be designed to handle the normal RMS current. This means that the normal design method most magnetic engineers use to design the inductors will give the desired results. These inductors will not saturate at the peak current (see Appendix). If the load does use the diodes, this will add severe noise problems to the already severe switcher noise. The EMI filter must attenuate the entire load noise spectrum.

4.2. OFF-LINE REGULATOR WITH CAPACITIVE LOAD

The most common circuit used today is the off-line regulator (Fig. 4.3 with or without the transformer). The output of the diodes feeds a large storage capacitor

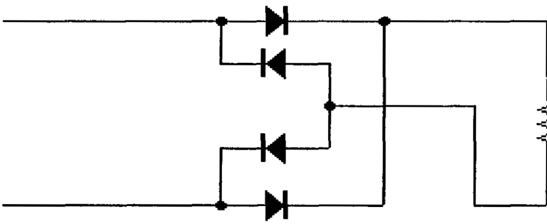

FIG. 4.1 Resistive load with diodes.

that in turn feeds a switcher(s). The load impedance also varies from nearly short to open, depending on whether the rectifier diodes are turned on or off. This depends on what instantaneous part of the sine wave voltage cycle is being fed to the diodes. The diodes are turned on if the sine wave voltage, plus the diode voltage drop plus *IR* losses, is greater than the capacitor stored voltage at that instant. High-current pulses charge the storage capacitor during turn-on (Fig. 4.4). The high-current pulses on the capacitor side of the diodes are often called sine wave pulses. But the curve shows that the main current pulse is well ahead of the 90-degree point (ahead of 1.57 radians on the plot) and the shape is not sinusoidal. On the diode side, this is rich in even-order harmonics, whereas the line side is rich in odd-order harmonics. Figure 4.4 starts at the angle of conduction of the diodes (in this case 0.8 radian), and a few degrees later shows the angle where the charge and discharge currents of the storage capacitor are equal, which is the lowest storage voltage point of the cycle feeding the load. Next follows the peak current angle that is well ahead of the 1.6-radian grid line. Past the 90-degree point where the charge and discharge are again equal is the maximum stored voltage point feeding the load. The cutoff point (in this example, 2 radians) follows this. The curve repeats in the next half-cycle. Just add 0.8 and 2 to π for the diode side and the second half is negative for the line side. If the load requirement drops, the start angle increases while the peak load current also drops. The stop angle stays about the same and may even increase slightly.

The curve was developed using the following values: $R_s = 1$ ohm, $R_l = 22$ ohms, C = 0.001 farad, $F = 60$ Hz, E_m is the peak line voltage fed to the diodes, and $N = 1$ for full-wave rectification.

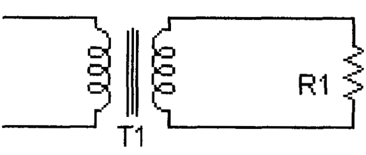

FIG. 4.2 Transformer with resistive load.

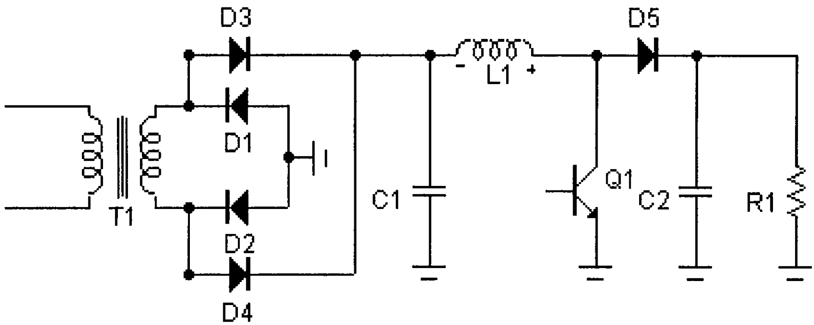

FIG. 4.3 Power supply with capacitive load.

The question at this time may be why this is being discussed here. It is to show what the true peak current is so that the inductor will not saturate and what the true RMS current is for sizing the wire gauge. This really is the actual requirement for the EMI filter. See the Appendix. The RMS current may be 8 A, while the peak current could be well above 25 A using this circuitry. The EMI filter must be designed to handle the peak current.

Figure 4.4 was developed by the following method (see Fig. 4.5).

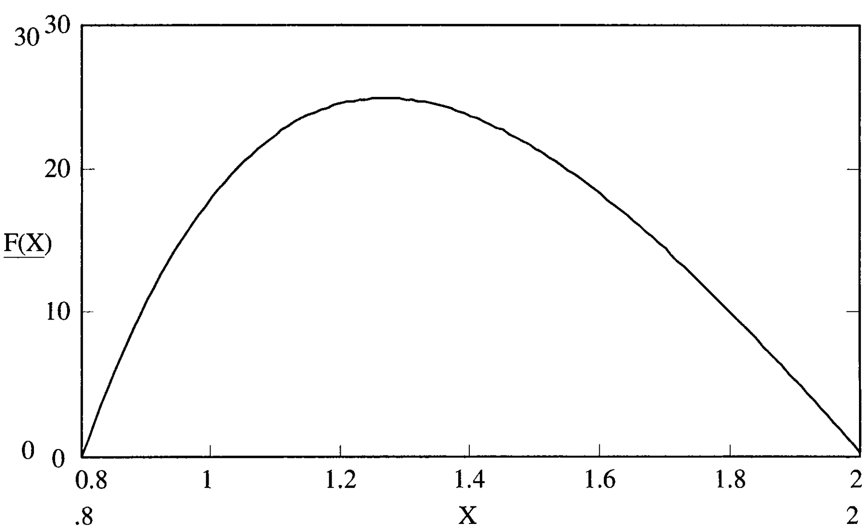

FIG. 4.4 Capacitor charging current through the diodes.

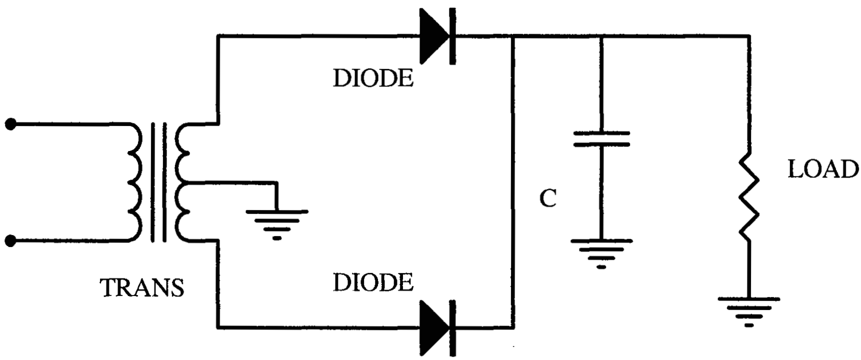

FIG. 4.5 The capacitive input filter.

*Equations for capacitor rectifier circuits.** In the equations, A is the start angle of conduction in radians where the diode is just turned on; B is the diode turn-off angle, also in radians; and N is 1 for full-wave and 2 for half-wave rectification.

$$\sin(A) = \sin(B)e^{\frac{(B - A - N\pi)}{\omega C R_L}} \tag{4.1}$$

where

$$Y = \pi - \tan^{-1}(\omega C R_L)\ X = \tan^{-1}\left(\frac{\omega C R_S R_L}{R_S + R_L}\right)$$

Equation (4.1) assumes the engineer knows the line frequency F, so $\omega(2\pi F)$ can be calculated, C is the storage capacitor in farads, R_s is the line resistance (typically 1 ohm), and R_L is the load resistance (usually the lowest value of resistance—highest current). The values of X and Y are then calculated along with $\tan(X)$. Substitute these into the first two equations. In one equation guess a value for A and solve for B. Insert B into the second equation and solve for A. Average the two values for A and start the process again by substituting this new value in the first equation. A spreadsheet is handy for this, and Keith Williams has a basic computer program for it. Once A, the start angle, and B, the stop angle, are known, U_1, the first minimum voltage angle guess, V_1, the first maximum voltage angle guess, and P_1, the first peak current angle guess, can be estimated.

*From Keith L. Williams, Grand Transformers, Grand Haven, MI, from O. H. Schade graphs developed in 1943.

$$
\begin{aligned}
U_1 &= \sin^{-1}\left|\frac{R_s + R_1}{R_1}\sin(A)\right| \\
V_1 &= \sin^{-1}\left|\frac{R_s + R_1}{R_1}\sin(B)\right| \\
P_1 &= \frac{(A + B)}{2}
\end{aligned}
\tag{4.2}
$$

Then U, V, and P can be solved through iteration. Substitute the known and the estimated U_1, V_1, and P_1 into one term and solve for the other U, V, and P. Average the two and resolve.

$$
\begin{aligned}
\sin(Y + X - B) &= \sin(Y + X - A)e^{\frac{(A - B)}{\tan X}} \\
\cos(X - U) &= \frac{(R_S + R_L)\sin(Y + X - A)e^{\frac{(A - U)}{\tan X}}}{\sin(Y)} \\
\cos(V - X) &= \frac{(R_S + R_L)\sin(Y + X - A)e^{\frac{(A - V)}{\tan(X)}}}{\sin(Y)} \\
\frac{\cos(Y + X - P)}{\cos(Y + X - A)} &= e^{\frac{(A - P)}{\tan(X)}}
\end{aligned}
\tag{4.3}
$$

Then E_u, the minimum voltage at angle U (in radians), and E_v, the maximum voltage at angle V (again in radians), can be found.

$$
E_U = \frac{E_M R_1 \sin(U)}{R_s + R_1} \qquad E_V = \frac{E_M R_1 \sin(V)}{R_s + R_1} \tag{4.4}
$$

where E_m is the peak line input voltage. Also, the peak current at angle P can be calculated.

$$
I_M = E_M \cos(X)\frac{\sin(P - Y - X) + \sin(Y + X - A)e^{\frac{(A - P)}{\tan(X)}}}{(R_S + R_L)\cos(Y)} \tag{4.5}
$$

Then the equation of the current during conduction is

$$
I = E_M \cos(X)\frac{\sin(\omega t - Y - X) + \sin(Y + X - A)e^{\frac{(A - \omega t)}{\tan(X)}}}{(R_S + R_L)\cos(Y)} \tag{4.6}
$$

Equation (4.6) is what is plotted in Fig. 4.4. Note here that the value of ωt lies between A and B. The voltage during conduction is

$$E = \frac{E_M R_L \cos(X)}{(R_S + R_L)}\left[\sin(\omega t - X) - \frac{R_S}{R_L}\,\frac{\sin(Y + X - A)e^{\frac{(A - \omega t)}{\tan(X)}}}{\cos(Y)}\right] \quad (4.7)$$

However, the requirement for the proper design of the EMI filter is the value of I_M. This is the current the inductor must handle without saturating. Then:

$$B_{KG} = \frac{1.55 L_{\mu H} I_M 10^{-2}}{N A_C} \quad (4.8)$$

Where I_M is the peak current just calculated, $L\mu H$ is the inductance in microhenrys, N is the turns, and A_C is the iron area, here in square inches, of the core.
Power factor:

$$\begin{aligned} P_F &= \frac{\text{true power}}{\text{apparent power}} = \frac{E_1 I_1 \cos(\phi)}{VA} \\ E_1 &\approx V \qquad A = \sqrt{\Sigma(I_N)^2} \\ P_F &= \frac{E_1 I_1 \cos(\phi)}{E_1\sqrt{\Sigma(I_N)^2}} = \frac{I_1 \cos(\phi)}{\sqrt{\Sigma(I_N)^2}} \end{aligned} \quad (4.9)$$

Once the peak current angle [P in Eq. (4.3)] is known from the equations, subtract P from $\pi/2$ (90 degrees) to get φ. The current fundamental frequency's peak must coincide with the peak draw of I_M in Eq. (4.5). This (φ) is the angle of lead. To finish the power factor, use the RMS value of the fundamental current (I_1) and the summation of the total current squared. This is continued in Sec. 7.5 of Chapter 7.

These approximate equations still hold true even if a transformer is inserted ahead of the off-line regulator. This current pulse is not truly a sine wave pulse, as most people describe it. These equations were developed for several reasons. The main reason was for the switcher noise to show the relationship between the total size and weight of the filter with and without an inductor of critical value (Figs. 4.6 and 4.7). This is discussed further in Chapter 12.

The total size and weight are smaller with the critical inductor for all serious specifications but might not hold for U.S. Federal Communications Commission specifications. The second reason is explained in Chapter 16 and the Appendix. This is also very important when discussing the three-phase filter in Chapter 16 for voltage rise. Finally, these equations were developed so that the full 3rd, 9th, 15th, and so on harmonic currents could be demonstrated for the neutral leg of the three-phase Y for capacitor design. See Chapter 7.

Stored energy feeds the switcher(s). The RMS value of the current is usually what the filter designer is given to design the filter inductors. Designers are not given this high peak current needed to design the inductors. The question is, what will this high-current pulse do to the filter inductors? Most inductor designers

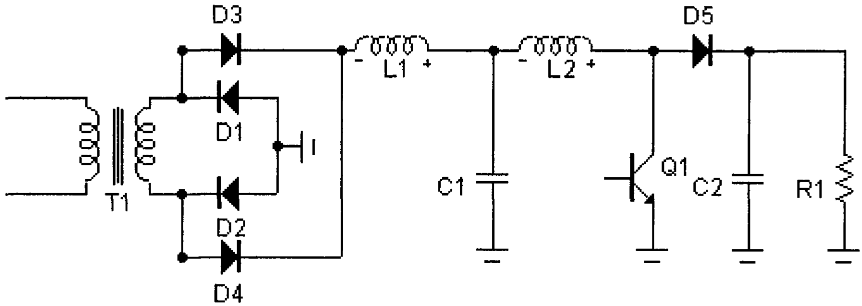

FIG. 4.6 Off-line regulator with the added critical inductance and transformer.

design the inductor somewhere around half-flux density at the RMS value of the current. This is the wrong approach for the design of this AC EMI filter inductor because the inductor will saturate. The equation is

$$H = \frac{0.4\pi N I_{\mathrm{p}}}{M_{\mathrm{pl}}} \tag{4.10}$$

where N is the number of turns, I_{p} is the peak current feeding the charging capacitor, M_{pl} is the magnetic path length in centimeters, and H is the magnetizing force. The maximum flux density, B_{m}, is a constant for the core material. The relationship between B and H is the permeability, μ. Thus, as H increases due to the large current pulse, I_{p}, the permeability drops. The permeability (μ) is a key player in determining the inductance, L. The core material is spoken of as "soft" because of the BH curve. These are S shaped, or sigmoid, as in Fig. 4.8 and are not made using square loop material. These cores require a strong magnetizing force, H, to drive the core into saturation. A "hard" core, or square loop core, is driven quickly into saturation. The soft core is the type of core material

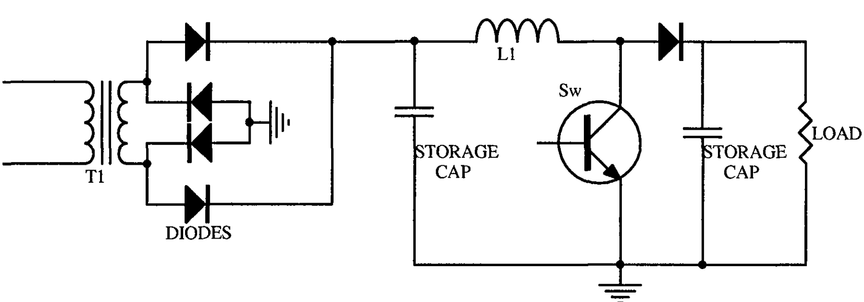

FIG. 4.7 Off-line regulator without the critical inductance capacitive load.

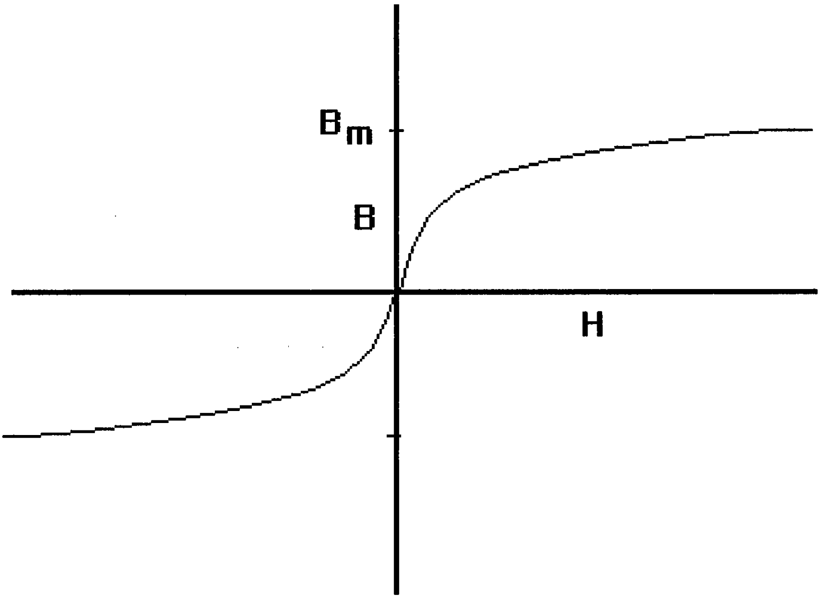

FIG. 4.8 *BH* curve of a "soft" (sigmoid) core.

needed for EMI filter inductors. Cores that have square loop characteristics are gapped, if they are to be used at all, reducing the hard magnetic core to a soft core (Fig. 4.9). Soft cores are also often gapped to make them even softer (less sigmoid) and for DC applications. The equation for the inductance is

$$L = \frac{0.4\pi\mu N^2 A_c 10^{-8}}{M_{pl}} \tag{4.11}$$

where the only new term is the cross-sectional core area, A_c, in square centimeters. So as the value of μ drops, the value of L drops and so does the inductive reactance.

What happens if the inductor saturates during this current pulse peak? The answer is that some switcher, diode, and other noise can ride through the saturated filter inductor during this time. These noise spikes can thus show up on the peak of the AC sine wave voltage. This is because these unknown high-current pulses saturate the inductor. The inductor design specification was underspecified. The only possible redeeming quality would be achieved if high-quality capacitors were used, producing a very low impedance to this noise as compared with the load and line impedance in parallel, but this is not likely at higher frequencies at the point where they are above their SRFs. This subject is continued in Chapter 11. Another way to avoid this is to design an inductor using a gapped core

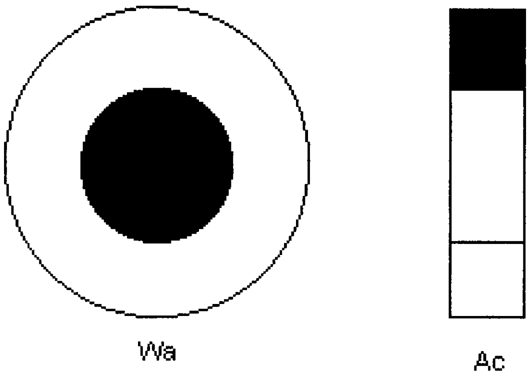

FIG. 4.9 Toroid showing W_a and A_c.

(Fig. 4.10). This tilts the *BH* curve, requiring a much higher magnetizing force to drive the inductor into saturation. It also makes the permeability much lower but more constant. This technique keeps the inductance more constant for low, medium, and high current demands throughout the conduction cycle.

The point is that the EMI filter will be bigger, it will weigh more, and it will cost more when using the off-line regulator because of the high-current pulses.

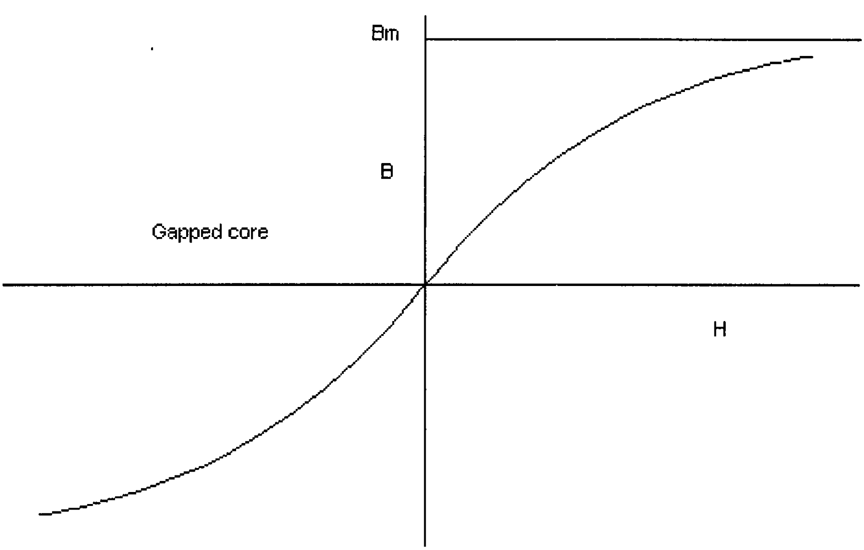

FIG. 4.10 Gapped core. The *BH* curve is more linear and less sigmoid.

4.3. OFF-LINE REGULATOR WITH INDUCTOR BEFORE THE CAPACITOR

Adding an inductor ahead of the power supply filter capacitor widens the current pulse width and lowers the peak current pulse (Fig. 4.11). If the value of the inductor is equal to or greater than the critical inductance, the current flows all the time and the current through the inductor is the average current. The normal inductance design method will work, again, in this application. The critical inductance is

$$L_c = \frac{R_0}{6\pi F} \tag{4.12}$$

where R_0 is the load resistance calculated using the lowest, worst-case, current and the highest line voltage and F is the line frequency. The disadvantage is that the stored voltage across the storage capacitor drops from the peak value of the AC voltage to the average voltage. This also adds to the weight of the power supply but cuts down on the EMI filter weight because it is not necessary to handle this high peak pulse current.

Some power supply company designers have been very proud of their light, compact power supplies. The power supply designers would ask the company to design a filter for this small power supply, but they would not understand why the filter is larger than their supply and weighs more. To keep their power supplies this small and light, they have eliminated all the normal internal EMI protection. Therefore, the output noise is high and the EMI filter is big! This is covered more fully in Chapter 11.

4.4. THE POWER FACTOR CORRECTION CIRCUIT

Today in Europe (Canada and the United States are following their lead) power factor correction circuits are a must (Fig. 4.12). I have seen power factors as low

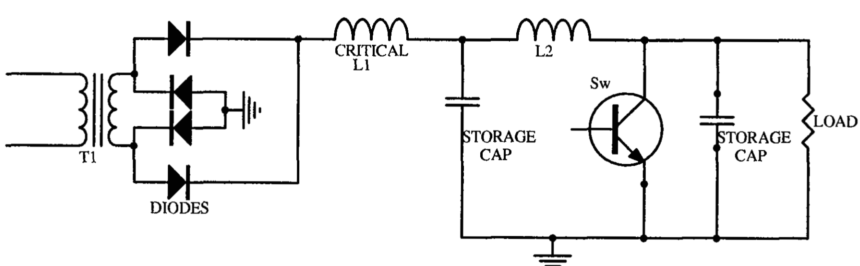

FIG. 4.11 Off-line regulator with critical inductance capacitive load, inductive input power supply.

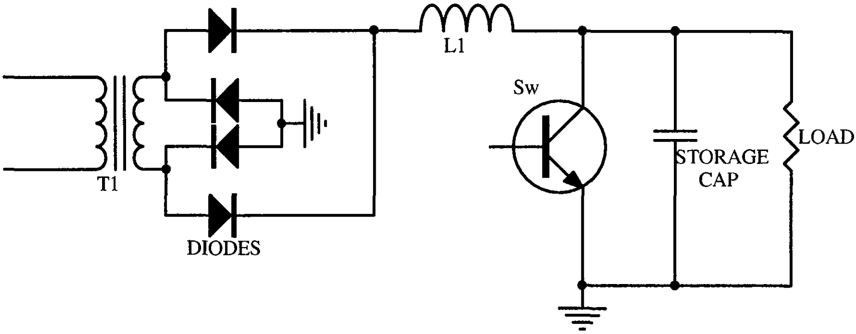

FIG. 4.12 The power factor correction circuit.

as 0.43 resulting from the power factor of the power supply and the EMI filter because of larger capacitors. Power supplies that use off-line regulators have power factors as low as 0.7 by themselves, and improper filtering only adds to the problem. The power factor correction circuits work by switching the diode output voltage without initially storing the energy in a large filter capacitor. Thus, the switcher current is in phase with, or follows, the line voltage. The switcher creates high-frequency pulses that follow the AC voltage sine wave returning the power factor back to near unity. This means that the output impedance of the EMI filter must be very low compared with the conducting load of the power factor correction circuit switcher frequency. The inductor conducts current in the same direction when the switch is open and closed, so the current through the diodes is a sine wave with a small triangular wave superimposed on top of it at the switcher frequency. It is the job of the EMI filter to attenuate this switcher frequency without starving the switcher. This is another reason why a filter designed for very similar specifications may work well for one group and fail for another. The first group could be using an off-line regulator and the second group a power factor correction circuit and the output impedance of the filter may be inductive (maybe a T). The inductive reactance would be low to the harmonics to the off-line regulator but be high to the power factor correction circuit. This would starve the power factor correction circuit.

The disadvantage of power factor correction circuits is that they are not 100% efficient. One of the main reasons for demanding power factor correction circuits was to allow more devices to be plugged into the wall sockets. These people did not account for the lower efficiencies of these circuits. So, little more equipment can be added to the wall outlets. The power required to drive the equipment using the power factor correction circuits is now greater, giving little gain for the wiring and circuit breaker. The initial efficiencies were around 70% but are close to 90% today.

There is also a power factor correction coil (Fig. 4.13). In most designs, the EMI filter looks very capacitive at the power line frequency, and some specifications demand a near-unity power factor for the filter for two reasons: obviously, for power factor correction, and also for leakage current problems. This technology is archaic, at best, and is mainly seen on 400-Hz power lines.

At the line frequency, the impedance of the inductors in the EMI filter is very low, so the capacitors add in parallel to a value, in Fig. 4.13, of $2C$. The inductive reactance of the power factor correction coil must be equal to the total capacitive reactance at the line frequency. This returns the power factor to near unity and the leakage current is reduced. If C_1 is equal to 3 μF, then Eq. (3.9) will yield the following.

$$L = \frac{1}{4\pi^2 F^2 C} = \frac{1}{4\pi^2 400^2 6 \times 10^{-6}} = 0.026 \tag{4.13}$$

where C is the total capacitance in farads of the filter to ground. This is a rather large inductor and the current through it is the same as the leakage current of the total capacitance (but 180 degrees out of phase). This technique converts leakage current of the capacitors to circulating current in this newly formed tank circuit. There are two conflicting factors to consider in dealing with the Q of this tank circuit. To reduce the leakage current as much as possible, the highest value of Q is required. On the other hand, due to aging of the filter components and working stresses on the filter over time, a lower Q is required. This is also true in installations where the power line frequency drifts over a wide range, such as remote power generators. This frequency drift would cause the network to be off-tuned from the center frequency and be operating on the side skirts of the

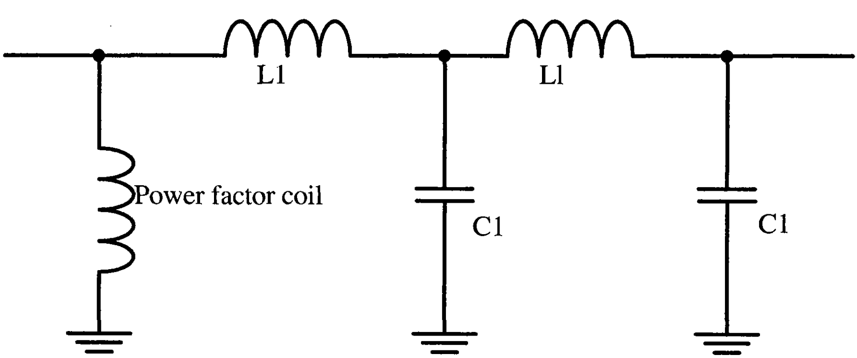

FIG. 4.13 Power factor correction coil.

impedance curve of the parallel tank circuit. Whatever the Q, high or low, the equation is

$$I_{circ} = Q \times I_{line} \tag{4.14}$$

where I_{circ} is the circulating current in this tank, the same as the leakage current prior to addition of the power factor correction coil, and I_{line} is the new leakage current. The reason for the concern about leakage current is the danger to anyone touching the filter, or the unit that the filter is mounted in, if the safety ground has been removed or broken off. The electric shock could be lethal when this ground is cut. Equation (4.13) shows that the leakage current has been reduced by a factor of Q.

RFI Corp., on Long Island, tunes the power factor correction coil for each filter before shipping. This technique allows them to ship these units as matched pairs with higher Q values. Otherwise, the Q should be limited to about 10. RFI attaches these power factor correction coils on the load side; others mount these power factor correction coils at the front end on the line side. I know of no functional electrical difference.

The specifications are leaning toward measuring the actual ground current rather than specifying the capacitance to ground. Using this trick would allow larger capacitors to ground and enhance the common mode action. Using a Q value above of 10, the 0.02 μF, for 400 Hz, could be changed to 0.2 μF. This would reduce the value and size of the common mode inductor, here called the Z (Zorro). This would enhance the common mode attenuation so that the filter easily passes the insertion loss requirement while reducing the value of the Zorro inductor. The only concern here is that the components and the line frequency must be stable to ensure that the units are not detuned or the leakage current will rise. This also returns the power factor back to near unity. The power factor correction coil becomes monstrous in size for 50 and 60 Hz and is rarely used.

4.5. TRANSFORMER LOAD

If the filter designer knows that the load is transformer fed, knowledge of the transformer is necessary. Try to get the customer or the transformer manufacturer to tell you the transformer primary inductance. The reason is that the total filter inductance in series across the filter must be much lower than the primary inductance. Otherwise, the filter and transformer inductance forms an inductive voltage divider.

Typically, the primary inductance is well into the millihenrys, possibly 50 mH or more, so this is not a problem, but it should be checked. Where the problem really shows up is with autotransformers and multiphase transformers that often employ autotransformers. Autotransformers typically have much lower primary inductance that often causes this problem. The autotransformer is smaller than the isolation transformer, and this allows the smaller inductance. The total EMI filter inductance should be less than 2% of the primary inductance to reduce the voltage divider effect.

4.6. THE UPS LOAD

Another load to consider is the UPS load. At least in the past, the typical UPS had zero crossing spikes that challenged the EMI filter. These spikes are very high frequencies and require quality capacitors of the feed-through type giving very low impedances to the 10th harmonic of the fundamental spike frequency. These spikes are typically at 25 kHz and higher, requiring a large capacitor with SRF to well above 250 kHz. I have known power specialists and consultants who analyzed power systems and concluded that a UPS was needed, only to find out that the UPS was creating the problem. Leaded capacitors cannot be used here, especially for the final capacitor facing the UPS because of the low SRF caused by the higher equivalent series inductance (ESL) and equivalent series resistance (ESR) of the leads. Often, a feed-through capacitor of 3.5 μF is enhanced with a leaded capacitor, often referred to as a "hang on," to bring up the total capacitance value. The leads must be very short.

To sum up this section, attempt to find out as much as possible about the load. Is the load a power supply and, if so, what type? If the power supply has a power factor correction circuit or an inductive input filter, the EMI filter must have low output impedance. The same is true if the power supply has a capacitive filter but the inductors of the EMI filter must be designed to handle high-current pulses. If the filter feeds a transformer, check the primary inductance and make sure the total EMI filter inductance is less than 2% of this primary inductance.

5

DC Circuit—Load and Source

There is a difference between AC and DC circuit designs for filters. It is understood that many off-the-shelf electromagnetic interference (EMI) filter manufacturers sell filters specified to work for both alternating and direct current. This adds to the difficulty in design because the sources are very different. Filter capacitors must be designed to handle the AC voltage and the total harmonic current. This calls for larger margins at 400 Hz. AC capacitors are typically designed to handle 4.2 times the RMS working voltage, whereas DC capacitors must handle 2.5 times the working voltage. Take a line with 250 V specified to ground. If the system is AC, this equates to 1050 V margin, which means the initial hipot will be 1800 V. This would increase the margins of the capacitor for the higher AC voltage. The inductor should be gapped to handle the direct current; otherwise the inductor will easily be driven into saturation. This dual requirement adds to the overall cost of the filter. In the preceding example, the DC 250-V line requires 625 V, calling for an initial hipot of 1000 V.

The filter manufacturer helps to cut the cost of these filters by building sizable in-house orders for shelf stock. These filters are usually grouped in families of different current values with the same specifications such as leakage current and dB attenuation. The same parts are used for all the different current values when possible, and this drives down the component costs. This makes the parts less expensive and reduces the cost to the customer. The filter customer can cut costs by having more than one application requiring the same filter. This increase in the number of filters required over a period of time cuts the filter price. This is true for all filter types.

The loads are also different. Some are resistive, as in heaters that maintain mountaintop repeaters above a certain temperature. Some equipment may use this supply voltage directly; others may be switchers, for which the inductors should be gapped for pot cores or C cores, but MPPs and powder cores are already gapped as a result of the manufacturing process. Gaps for the MPP, Hiflux (HF), powdered iron, ferrite, and now the new CMI cores have distributed gaps throughout the core.

This chapter discusses recommendations for elements within the circuit that reduce the EMI so that the insertion loss requirement of the filter is reduced, reducing the size, weight, and cost of the filter. Other techniques are also discussed.

5.1. VARIOUS SOURCE IMPEDANCES

The DC power line feed is very short, providing link coupling well up into the megahertz range. Often, it uses the normal chassis ground, making a balanced circuit impossible. This is where tubular filters play a role, using feed-through capacitors without line-to-line capacitors. Some have capacitance only; others make up the Ls and πs and the rest are Ts. This DC power is the type found in aircraft, shipboard, telephone company and mountain repeater sites. This power is furnished by various arranged systems such as battery racks, standby generators, solar panels, diesel generators, and wind generators. The battery feeds stored energy to the various systems and helps to regulate the voltage. These large batteries are called deep cycle batteries. Deep cycle batteries have more clearance between the inside battery bottom and the bottom of the battery plates. This allows a deeper discharge because of the increased clearance room for plate material and debris. These can feed the entire system from several hours to many days, depending on the system. They are designed in this way where the conditions warrant the higher costs or are referred to as *life threatening*, the highest cost system requiring more standby batteries, generator fuel storage, and the like. These types of conditions can be caused by any outside power failures, such as from a solar panel on a series of cloudy days, downed power lines, fuel supply outages, and bad weather conditions.

The plates of the battery rack act as a capacitor and shunt the middle frequencies of the unwanted conducted emission noises to ground, but the high frequencies are not attenuated because of the inductance of the cable feed and the battery plates. The radio frequency (RF) current on the power lead is radiated as an H field, and this is what should be avoided by filtering. The DC output normally feeds some switchers and these switchers create most of the RF noise, along with the diodes following the switcher and the parasitic oscillations of the switcher transformer(s).

The other type of DC system is from an AC power supply, and again the feed is very short. This DC normally feeds a switcher(s). The difference is that this output impedance is high at the switcher frequency even though the output impedance is only a few milliohms from DC to 10 Hz or so. The output impedance is inductive above this point and rolls off flat or resistive at, say, 5 kHz. This power supply looks like a milliohm resistor in series with an inductance, and this is shunted with the higher resistor value (see Sec. 3.5). Without being fixed, this situation starves the switcher because the inductive reactance of the inductor at the switcher frequency, and especially the harmonics, will be a high impedance equal to R_{hf} (Fig. 5.1). Is this true for all DC power supplies? No, in some rare cases the customer informs the designer long before the design is complete. Such specifications as the power supply must have low output impedance at a specific high frequency and other conditions that the supply must provide. The power supply people can make this output impedance low at switcher frequencies around 100 kHz. Does the filter designer need to know this output impedance of the DC supply feeding the filter before designing the new filter for the remote switcher?

What happens if the output impedance of the DC supply is low without the filter? The switcher will not be starved but the high RF current pulses will radiate, producing a very high H field. What happens if the output impedance of the DC supply is high without the filter? It has been stated repeatedly in this book that the switcher will then be starved but the weak RF current pulses will produce a very weak H field. The point here is that the filter fixes either condition, so the filter designer could care less about the output impedance of the power supply. This is easy to design and is treated in the next section.

5.2. SWITCHER LOAD

Most loads are of the switcher type. The output impedance of the filter at the load or switcher end must be such as not to starve the load. This assumes that the switcher does not have a capacitor at the switch to lower this impedance (or at least not the proper type of capacitor). The capacitors close to the switch are classed as part of the filter. If this capacitor is in the circuit without the filter

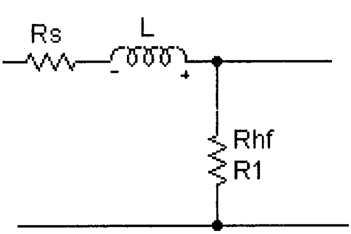

FIG. 5.1 Line simulation circuit.

designer's knowledge, it could detune the output network, lowering the cutoff frequency of this filter. The capacitor values add because they are in parallel. This assumes that the connecting wire is not a long lead. This would appear as an inductance tending to split the capacitors.

This DC application would require a single L filter so that the capacitor facing the load is as high as possible (Fig. 5.2). This is to reduce the drop in the DC voltage while the switcher is turned on. If double L filters are used, the value of the final capacitor is smaller than half the original. This could allow the voltage feeding the load to drop more than for the single L filter. This is still true with the double L even though the ripple voltage on the feed wire is the same (see Sec. 19.2). If the return is through the chassis, the capacitor facing the load should be a feed-through type. Otherwise, it should be a line-to-line capacitor in case the system has a return power lead. The input inductor of the filter has high impedance and reduces the H field no matter what the source impedance of the power supply happens to be. These are easy to design and the method is as follows. The switcher frequency is 60 kHz, the DC voltage is 60 V, and the *peak on current* is 1.2 A. Divide the voltage by the switcher on current to find the on impedance of the switcher, and divide this on impedance by 10 to minimize the voltage drop. Figure out the capacitor value by making the capacitive reactance equal to this impedance at the fundamental switcher frequency, 60 kHz. For the inductor, multiply the on impedance by 10 and then solve for the inductor reactance equal to this impedance at, as before, 60 kHz.

$$\frac{60}{1.2 \times 10} = 5\ \Omega \qquad C = \frac{1}{2\pi 60{,}000 \times 5} = 0.53\ \mu F$$

$$\frac{60 \times 10}{1.2} = 500\ \Omega \qquad L = \frac{500}{2\pi 60{,}000} = 1.326\ MH$$

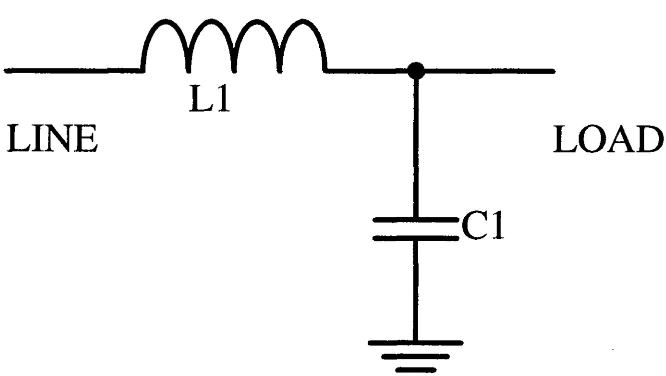

FIG. 5.2 Single unbalanced L filter.

This is rather a large inductor but should be easy to design because the main current is DC. If the duty cycle is 50%, the circuit carries only 0.6 A DC with a small ΔI AC current. However, if the circuit is balanced, then two inductors of 670 μH each are needed and must meet the same current requirement. The small capacitor of 0.6 μF is wired between the two inductors and faces the load. To calculate the actual inductor current, the pulse width is needed. Every pulse in this example is 4 μs times two for the two pulses per cycle and the 60-kHz cycle is 16.66 μs. So, 1.2 A times 2 times 2 divided by 16.666 is the average DC current through the inductor(s):

$$\frac{2 \times I_P T_{on}}{T_{total}} = \frac{2 \times 1.2 \times 4}{16.66} = 0.57 \text{ A}$$

The 1.4 MH would have to carry only about .57 A DC, making this possible even if the filter is unbalanced. This can be made into a simple formula:

$$C = \frac{10 I_P}{2\pi F_{sw} V_{dc}} = \frac{10 \times 1.2}{2\pi 60{,}000 \times 60} = 5.305 \times 10^{-7} = 0.5305\ \mu F$$

$$L = \frac{10 V_{dc}}{2\pi F_{sw} I_P} = \frac{10 \times 60}{2\pi 60{,}000 \times 1.2} = 1.326 \times 10^{-3} = 1.326\ MH \tag{5.1}$$

The equations give inductance in henries and capacitance in farads. The 0.6 μF capacitor may be paralleled with a smaller capacitor (0.01 μF) to increase the high-frequency response.

5.3. DC CIRCUIT EMI SOLUTIONS OR RECOMMENDATIONS

This is included here because most problem solutions deal with DC power feeds or are part of the power supply. Some of these may conflict, but use what works.

Most systems grow by these means: The inception of the idea is based on a customer request, product application, or new technology. The various concepts are listed, followed by choosing a method. The various design methods and design problem solutions are selected, and the various circuits or circuit boards are combined to test the system. This is followed by system corrections and then the power supply, oops, the input transformer, and double oops, the EMI filter (in the 2.25 ×1.6 × 0.75 inch or any other small leftover volume they could not fit anything else into). Usually, as described here, the EMI filter is thought of last. Then people wonder why the filter is so big and bulky and why it cannot fit in that little leftover volume next to the input transformer. The plan for EMI should begin almost at the inception of the idea and should be required on each board or circuit that is applicable. If this is done correctly, the EMI filter might fit the leftover volume and the cost, along with the weight of the filter, might be greatly reduced.

5.3.1. Some Ideas for the Initial Power Supply

Include an inductor of critical value in front of the storage capacitor to remove the high current peaks. Lack of this inductor increases the size of the EMI filter, it weighs more, and it costs more, especially for the military, whose specifications are harder to meet. Isolate the off-line regulator to reduce the common mode, and remember that a capacitor is not a capacitor at all frequencies. Power supply storage capacitors of larger size are inductive by 20 kHz. Parallel them with a good-quality extended foil or ceramic capacitor. Use the proper snubbers in the switchers to remove spikes and stop reverse currents through diodes, for example. This reduces diode noise. Keep the various supplies isolated, and use twisted pairs to power remote units rather than a ground return. Filter the isolated leads, or closely spaced traces on the printed circuit board, at the device end. In this way, the primary current in the traces or twisted pair is direct current rather than pulses at the switcher frequency. Lay out the power supply transformer, input, or switcher(s) for minimum magnetic coupling from the transformer to a susceptible receptor by increasing the distance between these devices or finding the best orientation of the transformer.

Sometimes the magnetic field radiating from or around the transformer is stronger in one direction than another—the lines of magnetic force may extend farther. Go to a toroid transformer, which is known to be quieter magnetically. Use single-ended converters rather than a flyback transformer, and do not use switchers known to be noisy, such as SCRs (this should be obvious). Place sections known to be noisy within a container or shield, which helps to stop the H and E fields. Cold-rolled steel must be quite thick to reduce the magnetic field generated by a transformer or any other device. These containers are often silver plated inside and out to enhance the surface conduction. The better the conduction, the better the H field is attenuated and the lower the current on the outer skin. This may require a mu metal can or foil which would be thinner and lighter for the same attenuation of the H field but would cost much more.

Moves to increase the self-resonant frequency of the inductor or transformer of the switcher also reduce the parasitic or at least raise the frequency of the parasitic. This parasitic frequency often feeds all the way back to the filter where it is attenuated. It can be attenuated with a smaller value of capacitance at the interface of the power supply and switcher. Filter these as close to their source as possible.

5.3.2. Other Parts of the System

Clocks and other sources generate more than their share of noise. Sometimes the noise does not start or originate in the clock but is created either in the power supply or in other circuits that the power supply feeds. Make sure each feed of the supply is filtered. Filter each feed such as the clock, DAC, or other device as close to the device as possible.

Separate ground systems for digital and analog. Separate grounds for the noisier systems so that the noise current on the ground lead does not induce that voltage in the quieter circuits and be amplified. Shield the noise generators so that magnetic radiation of the *H* field is a minimum.

Keep the noise within the EMI filter inductors and other inductors in nearby areas, even though the cores are toroids (which are quieter and radiate less), by mounting these inductors in quadrature to reduce their coupling. This is especially true if the cores are not toroids because they are known to have more flux outside the core material. This flux can induce currents in other susceptible nearby devices.

Figure 5.3 shows a balanced double "L" where the first two inductors feed the capacitor using the vee technique (see Chapter 7). The last two inductors in quadrature also use the vee technique. The lower left inductor could be magnetically coupled to the upper right inductor, but the increased distance and the capacitor between them reduce this.

5.4. LOSSY COMPONENTS

Another technique is to use lossy components. One such technique is from transmission lines where high frequencies are absorbed within the dielectric. Coaxial cable developers continue research to find dielectrics with lower and lower losses, whereas others have used this phenomenon to enhance the losses to make lossy systems. One such company, in Long Island, New York is Capcon, Inc. (Ed Reeves). Their material is tubular and can be threaded with the filter

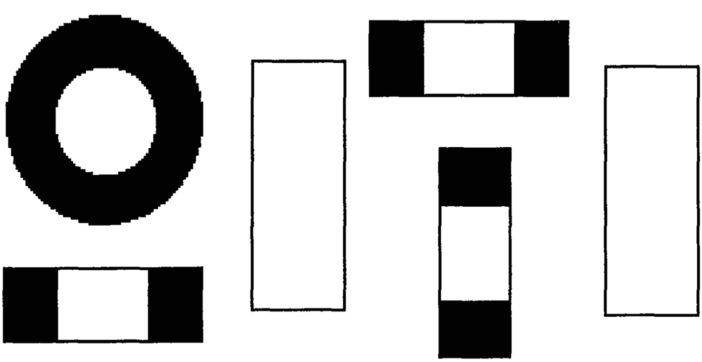

FIG. 5.3 Toroids in quadrature.

hookup wire; it provides substantial losses above 10 MHz and well into the GHz region. This depends on the total length of the material within the filter. Various inside diameters are available, allowing easy threading for capacitor, inductor, and other hookup leads. Capcon material functions better than most ferrite beads because most ferrite beads saturate around 5 A. These beads have a self-resonant frequency of about 50 MHz. The lossy component mentioned above does not fail either of these requirements. One foot of this lossy line material gives as much as 100 dB at 10 GHz, whereas the ferrite bead gives 20 dB loss at the peak of 50 MHz, and the loss falls to zero by 100 MHz. This material is also available in sheets with various thicknesses for enclosures and is very dissipative. Radiated energy can be reduced by covering the filter enclosure, or a noisy device, with this material.

Figure 5.4 shows the approximate loss of lossy suppressant tubing shielded (LSTS) and LST, which is the same tubing without shielding, measured per MIL STD 220A.

5.5. RADIATION EMISSIONS

The subject of this book is the design of the filter for conducted emissions. It is generally true for radiated emissions that if the conducted emissions are reduced,

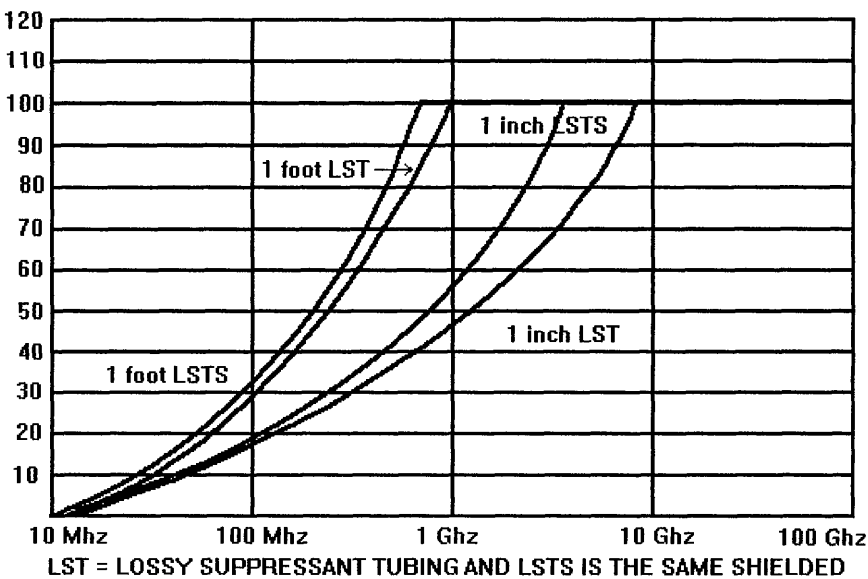

FIG. 5.4 Capcon lossy tubing and sheets.

the radiated emissions are also often reduced. If the conducted emissions are nearly eliminated, there is often little left to radiate. The opposite condition is even truer: if the conducted emissions are not eliminated, the device radiates. A condition that makes this statement erroneous is that the outer skin of the device is not a good conductor—plastic in the case discussed. Although the conducted emissions were greatly reduced, the radiated emission was still almost as offensive. If the case is to be plastic, there are conductive sprays that the enclosure can be coated with that help to reduce the emissions.

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6

Typical EMI Filters—Pros and Cons

Let us discuss the typical EMI filters that are used today. These are primarily the π, T, and L, with Cauer, RC shunt, and dissipative filters sometimes included. They include double and triple filters and sometimes even quadruple uses of the mentioned types of filters. Each EMI filter has some positive and negative attributes. Each has its best spot where it functions very well and other spots where it fails to the first degree. This chapter should clarify all of this. In addition, the transformer, if used, will add to the filter loss.

6.1. THE π FILTER

We start with a discussion of the pi filter (Fig. 6.1). This filter looks very good under the 220A 50-ohm test specification. The line-side capacitor will work into the 50-ohm line impedance at the low frequencies. If this is the only test requirement specified by the customer, then the π will test well with this test technique. The π filter will be passed with flying colors by the source inspector (if any). This is especially true for the three-phase type, for which the specification states "Measure one phase with the other two phases tied to ground." The input and output capacitors of each pi section are then doubled in value. This makes it easier for the filter designer to meet the specified loss within a given weight and volume. It means that the filter can easily pass the attenuation requirement with smaller values, package size, and weight.

Note that the center capacitor is twice the size of the end capacitors. All inboard capacitors are twice either of the two end capacitors for all multiple π filters (Fig. 6.2). Figure 6.3 shows the 220A test setup.

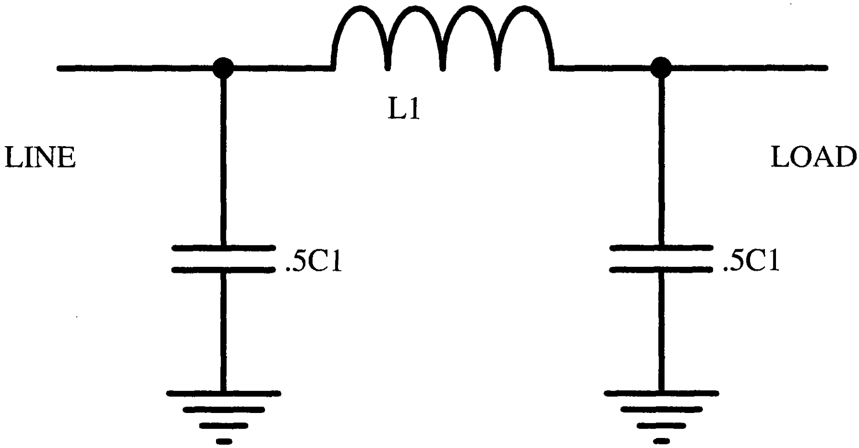

FIG. 6.1 The π filter.

The peak current is supplied to the filter through buffers or extended buffers. These are large values of inductance to keep the EMI test signal out of the DC power supply. The DC current supplied through the buffers is the peak AC current the filter must handle, not RMS. This is done to ensure that the filter inductors do not saturate at the full peak current. Two sets of 50-ohm matching pads from 6 to 10 dB each are omitted in Fig. 6.3.

On the other hand, this type of filter looks bad under the 461 or naval tests (Fig. 6.4). This is where a 10-μF capacitor shunted from line to line or to ground is required. The reason is that a filter capacitor facing the 10 μF is out of the

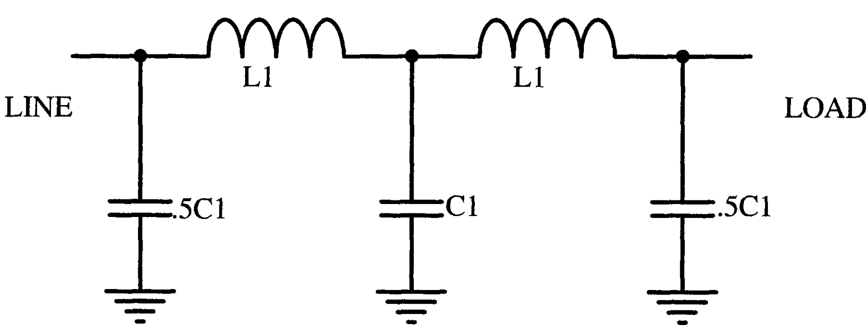

FIG. 6.2 Multiple π.

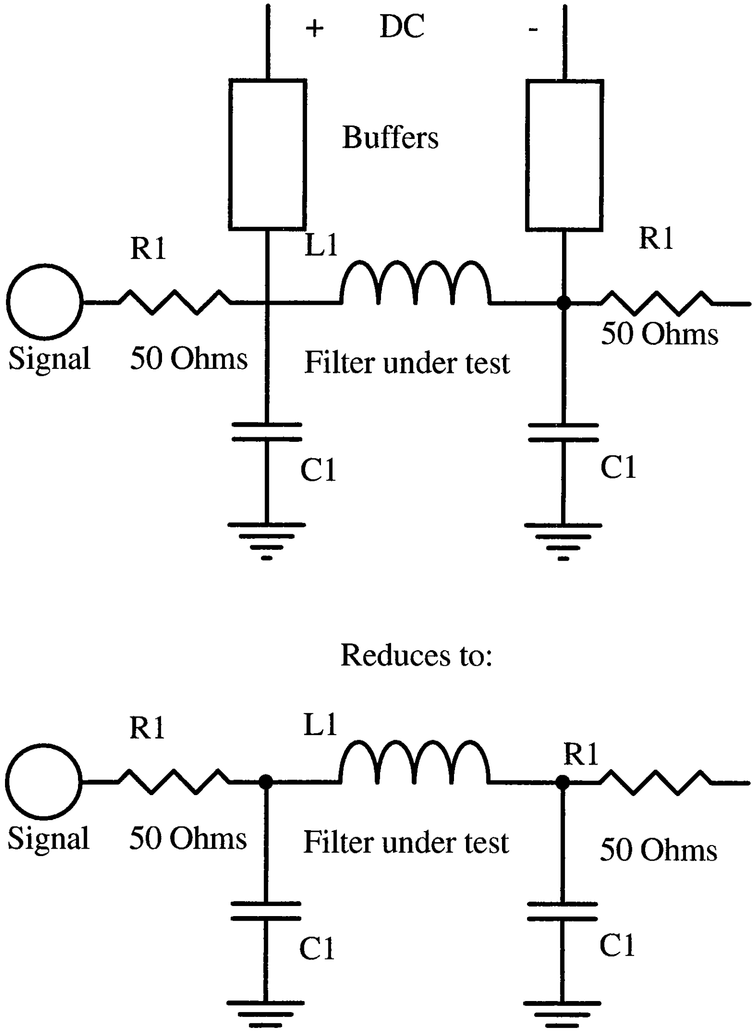

FIG. 6.3 The 220A test setup.

circuit unless the cap is an order of magnitude greater than the 10-μF capacitor. This value would be much too high for good filter design. Large capacitors have very low self-resonant frequency (SRF) and would have to be shunted with smaller capacitors of good quality to make up for this low SRF. The receiver or spectrum analyzer is switched back and forth between the two current probes to determine the filter loss.

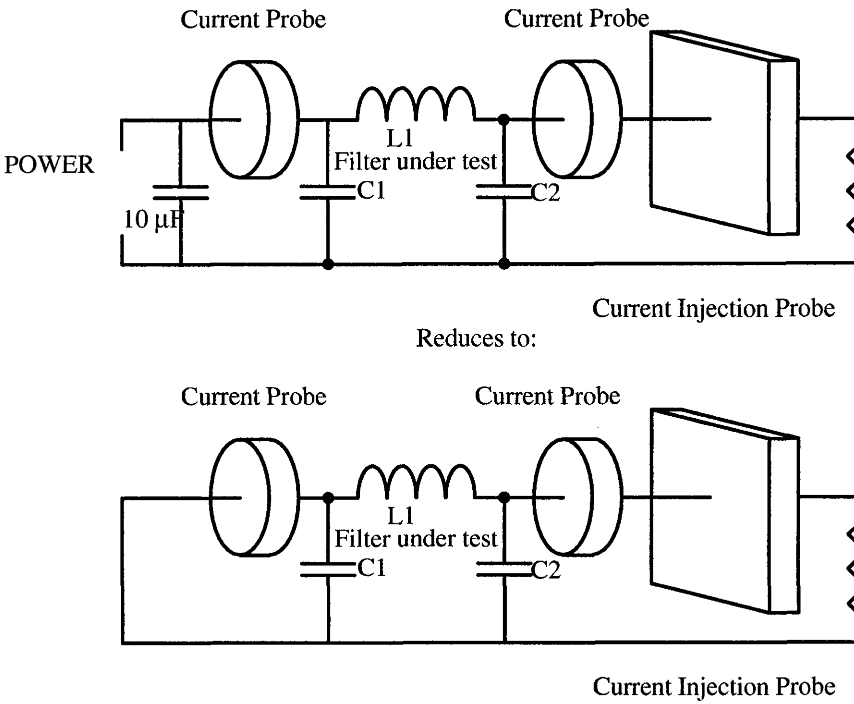

FIG. 6.4 The CIP test setup.

The three-element π looks like a two-element L filter at the lower frequencies in the 461 and naval tests. The loss is about 6 dB per octave for each element, so the 18-dB π filter is now a 12-dB loss per octave filter; 18 dB is for the single π and the double π is 30 dB, for example.

The π might also function well in some DC systems if the switcher frequency is high enough so that the capacitor impedance facing the load is small enough not to starve the switchers. Also, the capacitor impedance should not cause excessive voltage drop. This assumes that the switcher circuit has not handled this problem. The π filter is easily balanced by placing only half of the inductor needed in the high line and the other half in the neutral line (Fig. 6.5). This changes the shape from the π to a square or box shape. It does happen that the system must pass a specification after several prototypes are finished.

A qual test may have to be passed (Fig. 6.6). Now the earlier π, which passed all those costly tests, may not perform properly. This is because the filter loses the front, or line-side, capacitor at the lower frequencies. Some will wonder

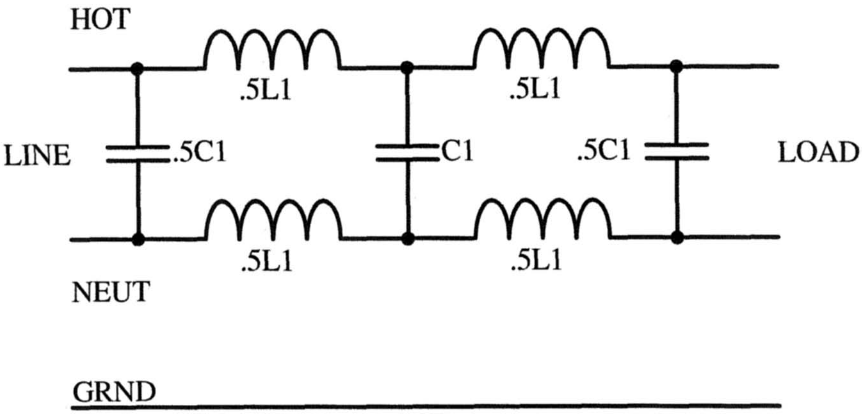

FIG. 6.5 Multiple balanced π.

why the filter is not doing the job when it properly passed the tests at the EMI test laboratory! The real world is not reflected by the 220A or 461 specifications, but the 461 specification is closer to reality.

6.2. THE T FILTER

The T filter gives 18 dB per octave, and the double T gives 30 dB per octave (6 dB per element) (Fig. 6.7). T filters work best in low-impedance lines (high current requiring small inductors). The line impedance is very low up to at least 100 kHz, but the 461 loss specifications start at either 10 or 14 kHz. The inductive input impedance of the T adds to the low line impedance. This gives the capacitor an input impedance to work into. These are also best in the higher current loads

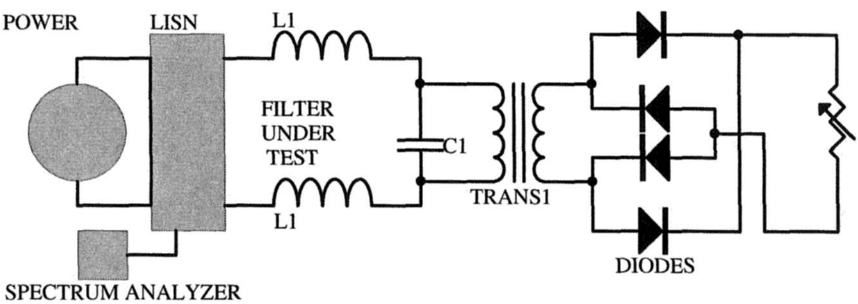

FIG. 6.6 System or diode test method.

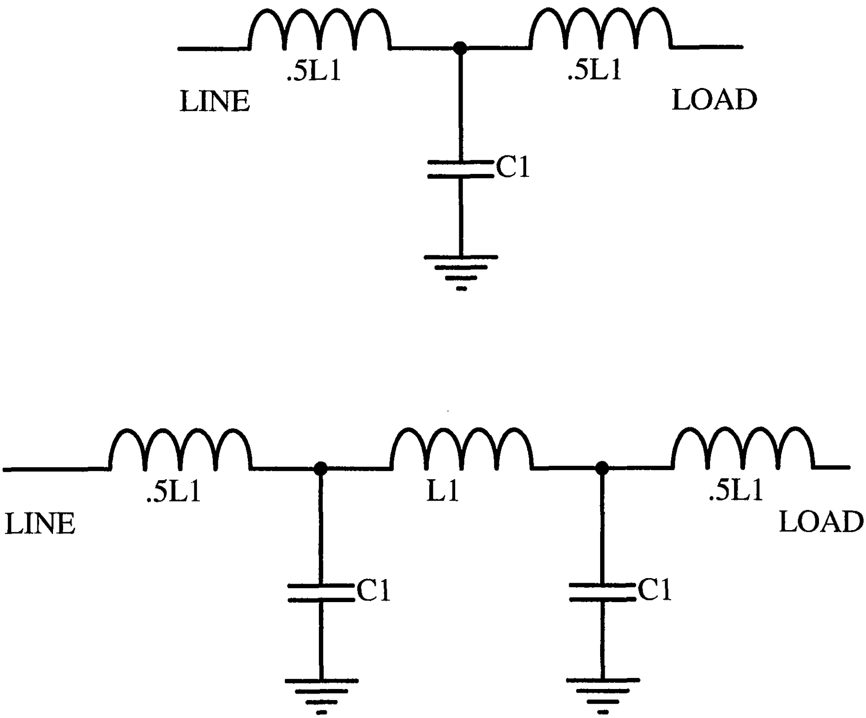

FIG. 6.7 The T and multiple T drawings.

if the design method does not call for too high values of these T inductors. This could result in the voltages soaring or dropping, feeding the load. I have seen 115-V AC 60-Hz lines as high as 132 V feeding a light load requirement. This happened because the resonant rise occurred at a very low frequency. This is usually caused by higher inductive values.

The T should never be used in the DC system if the load utilizes any switchers because the high impedance of the output inductor facing the load will starve the switchers. The switcher designer may have taken this into account by lowering the impedance with a capacitor at the switch input. This really makes the filter into a maladjusted double L because the capacitor shunts the inductor facing the load, but this is less troublesome in DC than AC.

Note here that the central inductor is twice the size of either of the two end inductors. All inboard inductors are twice the size of either of the end inductors. The T filter can be balanced by removing half of the inductor's values and placing this half in the neutral leg, forming an H pad (Fig. 6.8).

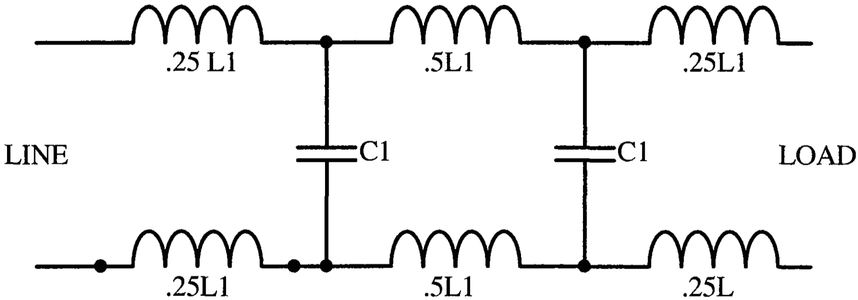

FIG. 6.8 Multiple balanced T filter.

6.3. THE L FILTER

The L type is the filter most often used. The π and the T are three element, or more if multiples are used, and should give 18 dB of loss per octave. The L, being only two elements, provides 12 dB of loss per octave. All of these loss figures refer to the loss starting above the cutoff frequency. A single L filter works best in the DC mode if the load has switchers. The inductor faces the DC source and the capacitor (of high quality with a high SRF) would provide a low impedance for the switcher frequency (Fig. 6.9). Why not a double L? The two inductors required for the same amount of loss would total less than the single inductor, and the same is true for the capacitor values. See Sec. 23.2. This smaller output capacitor, smaller than 0.5 of the original, may not furnish the needed energy storage, creating a larger peak-to-peak voltage drop feeding the switcher. This peak-to-peak voltage would be at the switcher frequency or twice the switcher frequency if this feeds the center tap of a class B amplifier or Royer. The double L can be used as long as the drop is not excessive or the switcher frequency is

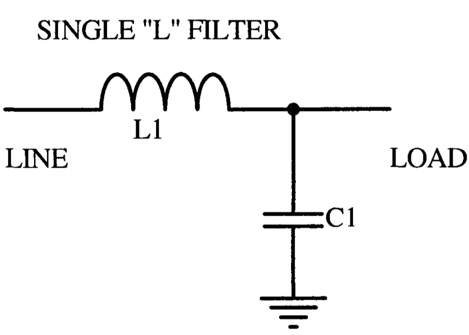

FIG. 6.9 Single L filter.

high enough. This statement is true even though the attenuation is improved for the DC source or power supply voltage. The L and multiple L work well in higher power applications (Fig. 6.10). Again, to balance, split the inductors and put the other half in the neutral (Fig. 6.11). The double L has 24 dB loss per octave.

6.4. THE TYPICAL COMMERCIAL FILTER

This is the type used in test equipment, computers, and other commercial electronic equipment. Here the manufacturer has to pass tests for UL, TUV, VDE, CSA, and the FCC. The tests are conducted by EMI test houses that help the manufacturer with all the documentation needed for the various agencies. These filters are the balanced π type and are often purchased from outside suppliers and often built offshore. The filters are mainly common mode in appearance, with a capacitor across the input and output from hot to neutral and two other capacitors to ground that must meet the leakage current specifications for whatever agency has the toughest requirement. The leakage inductance is often made high by adding washers to the center of the pot cores separating the two windings. This is also accomplished by winding the two windings as far apart as possible on a ferrite toroid core. The feed-through capacitors are grounded directly to the case (Fig. 6.12). See Chapter 7. Some of these techniques add differential mode to this common mode filter by increasing the leakage inductance or adding inductors to both lines. This then makes a balanced π type with both differential and common Mode. This all works because the losses specified for the FCC start at 450 kHz. The inductors and capacitors can be quite small to accomplish these tasks. Another point to remember is that the current through the leakage inductance

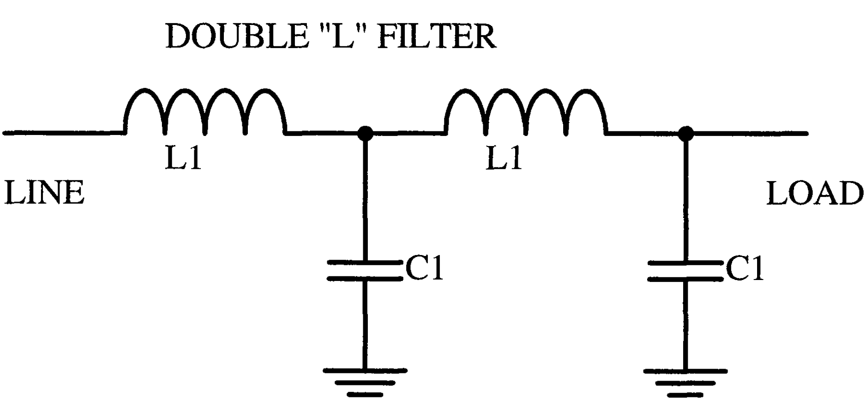

FIG. 6.10 Multiple L filter.

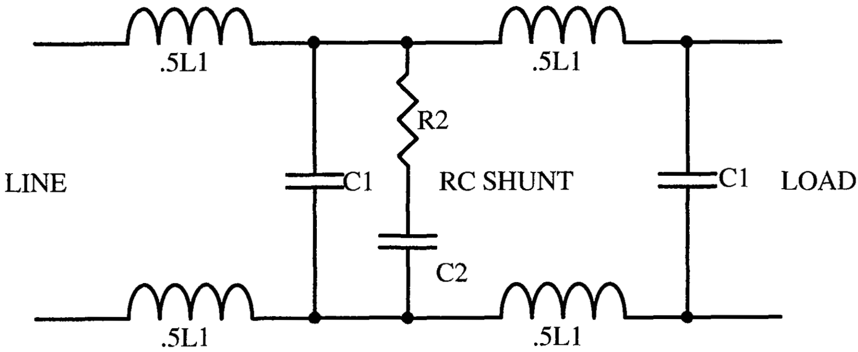

FIG. 6.11 Balanced multiple L with RC shunt filter.

cannot saturate this inductor because most of this is not through the core but in the surrounding air. This is because the common mode inductor is wound on ferrite cores with high A_l values. Some of these filters do not use the feed-through type of capacitor so the circuit changes to that in Fig. 6.13. These capacitors are less expensive but the self-resonant frequency is lowered by the added lead length.

6.5. THE DISSIPATIVE FILTER

This filter is rarely seen in the EMI arena today. It consists of one inductor and one capacitor along with two resistors (Fig. 6.14). The two resistors are tied in

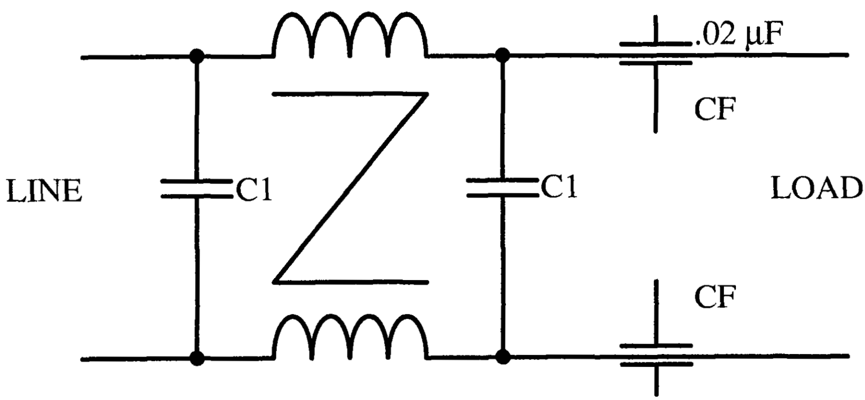

FIG. 6.12 Commercial filter with Z and feed-through capacitors.

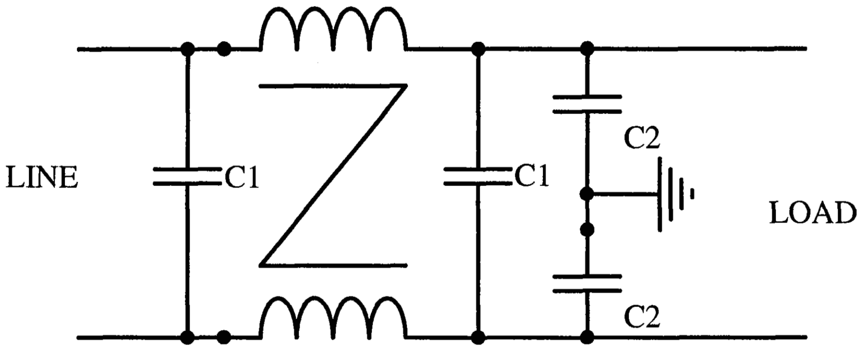

FIG. 6.13 Commercial filter with Z and leaded capacitors.

series across the inductor terminals, and the capacitor is tied to the center of the resistors and then either to the other line or to ground. These filters are similar to the line impedance stability networks (LISNs). This filter can be balanced by using half the inductor and resistor values to be split on both legs and the capacitor tied between the center points of the four resistors (Fig. 6.15).

The main disadvantages are that this filter appears to give 12 dB per octave but really gives only 6 dB. Another is that this adds cost and volume to the entire filter for only 6 dB loss for this section. The resistors should be noninductive and the resistor(s) facing the line should be at least 2 watts for electromagnetic pulse (EMP) purposes.

The advantages are as follows:

1. Using two of these at each end can match this filter to line and load

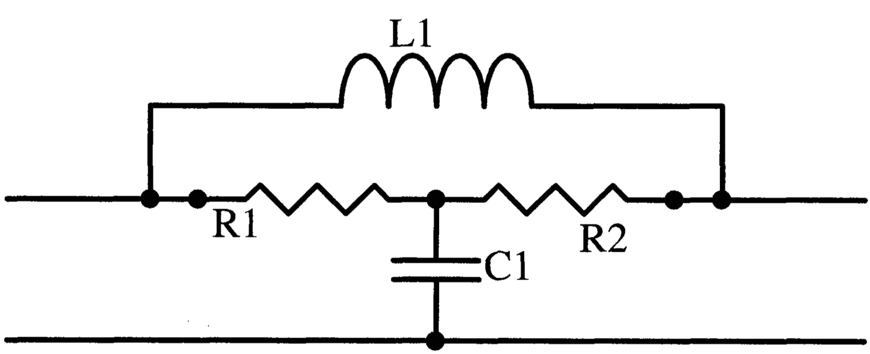

FIG. 6.14 Dissipative filter.

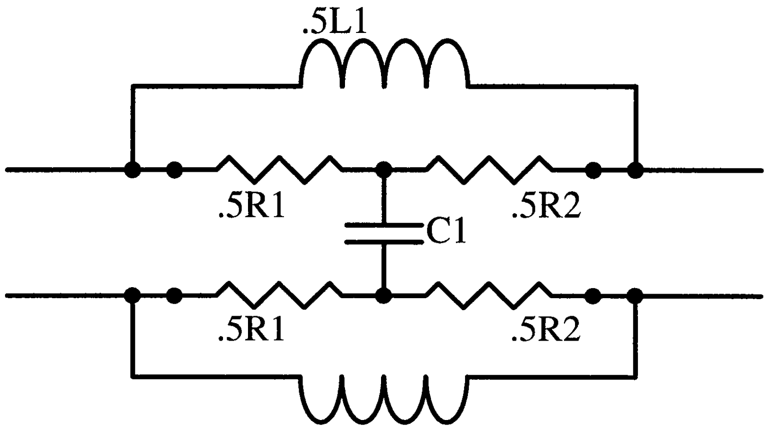

FIG. 6.15 Balanced dissipative filter.

impedance (Fig. 6.16). Any filter can be used between these two end units and designed to match the impedance of the inbound resistor values. This can work well to meet the 220A specification using four 50-ohm resistors and designing the middle unit for 12 dB less loss.

2. This filter dissipates the stop band energy. If any higher frequency comes down the line, most of the energy will be dissipated by the line-side resistor. The load-side resistor will dissipate most of the energy of the switcher or any other noise from the load. If this is the type described, the energy not dissipated would be reflected by the center filter and the opposite resistor of the same dissipative filter and most of this energy would be dissipated.
3. A filter would have a filter very similar to an inexpensive LISN would be included within the filter, and two if the unit is the type described and pictured in Fig. 6.16.

6.6. THE CAUER FILTER

The Cauer, or elliptic, filter is best used in very low impedance circuits (Fig. 6.17). These filters are usually used with multiple Ls and Ts. In any case, a capacitor is normally shunted across the center inductor. This is used to fix a problem frequency such as 14 kHz. The network is tuned to slightly above the problem frequency. Granted, the trouble may be fixed, but this center section will not be in the circuit above this problem frequency. The network will pass all the upper frequencies. Often, a resistor is placed in series with this capacitor and

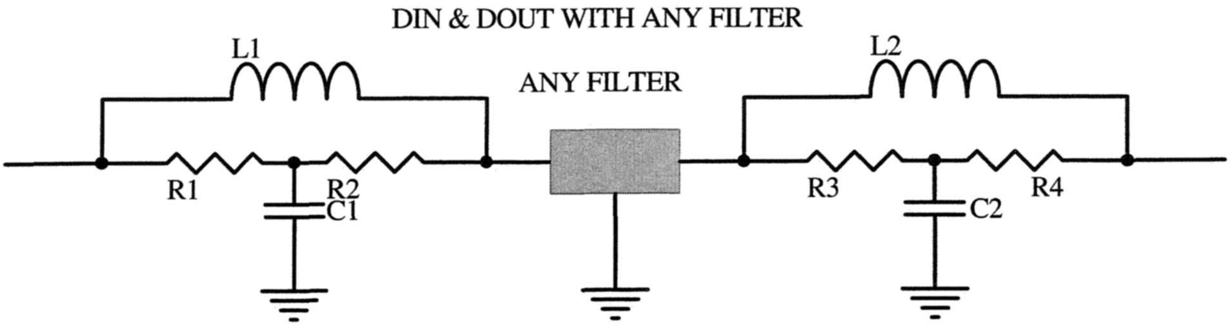

FIG. 6.16 The DIN and DOUT dissipative filter for both ends.

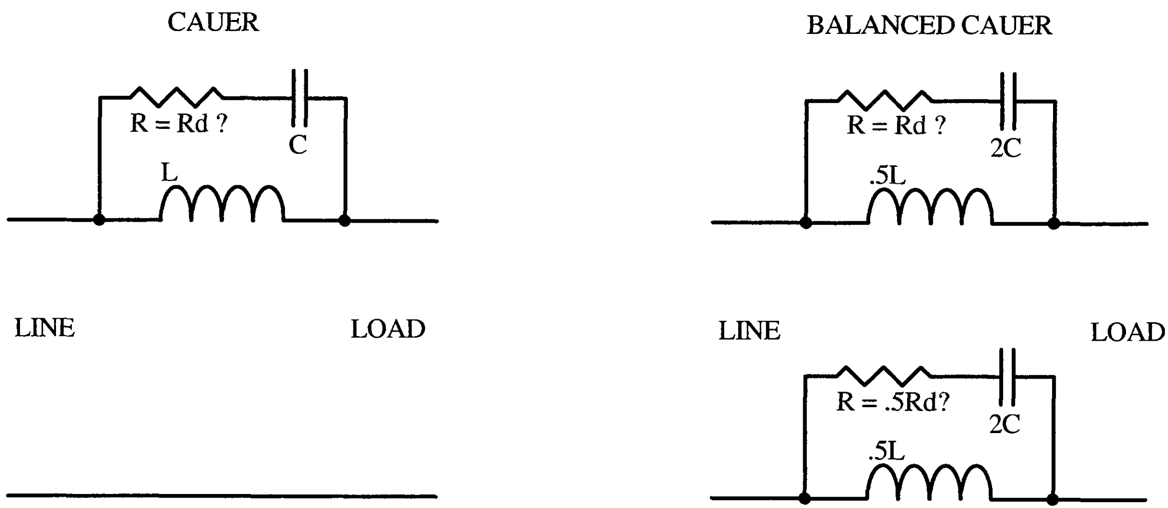

FIG. 6.17 Cauer and balanced Cauer.

the resistor limits the amount of bypass. The value of the resistors is typically around 10 ohms and is often the value of the design impedance of the filter. The design impedance is the lowest RMS line voltage divided by the highest RMS line current.

If this filter is to be balanced, use half the calculated value of inductance on the two lines, and parallel each inductor with two capacitor-resistor networks. The value of the two capacitor networks uses twice the capacitance and half the resistor values, and these would, again, be tied across the two inductors.

6.7. THE RC SHUNT

Another technique is preferred to the Cauer but is better when used in high-impedance, low-current circuits. This filter, called the RC shunt, uses fewer components and is automatically balanced across the line to start with, if required (Fig. 6.18). This is formed with a capacitor and a series resistor. Normally, the filter has a resonant rise lower in frequency than the trouble frequency. This is especially true if the filter is a multiple filter such as a double or triple L, π, or T. Usually, the number of resonant rises is one less than the multiple number, meaning that the single L, pi, or T would not have any resonant rise but the quad would have one less, equaling three. This holds true only if the circuit Q is low enough. The higher Q has a resonant rise for each network. Find the frequency of the lowest resonant rise and pick the capacitor value at this frequency that equals the filter design impedance. This will attenuate each resonant rise above the first and also the trouble frequency. If the resonant rise frequencies are of no concern—well above the fifth harmonic of the power line frequency and well

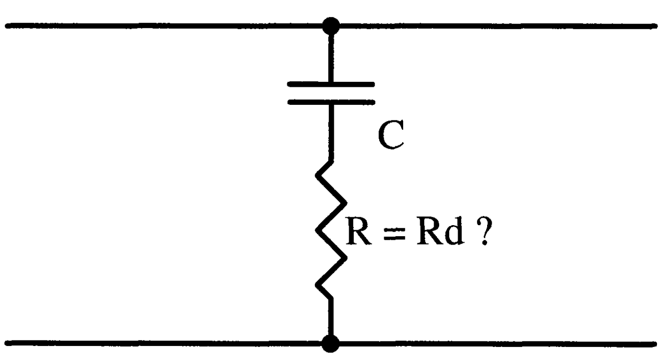

FIG. 6.18 Balanced and unbalanced RC shunt.

below 10 kHz—choose the capacitor to equal the design impedance at the trouble frequency.

As an example, the design impedance is calculated by dividing the highest current required by the load into the lowest anticipated line voltage. In this case, say, 100 V is the lowest line voltage at 10 A. This is the highest current at this lowest line voltage. The resonant rise frequency is 14 kHz. Then,

$$R_d = \frac{100}{10} = 10 \text{ ohms} \qquad C = \frac{1}{2\pi 4000 \times 10} = 3.979\ \mu F \tag{6.1}$$

The capacitor is 3.979 μF or 4 μF in series with the 10-ohm resistor tied across the line. This should remove the bump at 4 kHz and attenuate a problem frequency at 14 kHz. If the resonant rise is in an area of little concern, change the frequency to the problem frequency of 14 kHz. Then recalculate the value of capacitor needed at that frequency.

$$C = \frac{1}{2\pi \times 14{,}000 \times 10} = 1.137\ \mu F \tag{6.2}$$

The 10-ohm resistor is in series with a 1.2-μF capacitor tied across the filter. This lends itself to multiple L filters where the preceding network can be tied across any one of the capacitors. The closer it is to the load, the more it tends to minimize or reduce the impedance swings of the load.

6.8. THE CONVENTIONAL FILTERS

The filter houses rarely use conventional wave filters. Some gurus do use these filter types, but they require constant input and output impedance to work properly, especially for the low-frequency losses required by the military. The line and load impedances in EMI just do not provide this condition at these low frequencies. However, a filter would test well in the 220A test method if the filter was designed for 50 ohms. These filters would work very well between the two dissipative filters discussed in Sec. 6.5 (Fig. 6.19) because the noninductive resistors would give the proper match needed by these filters. This is especially true if the cutoff frequency of the dissipative filter is about half that of the regular filter.

6.9. MATRIX—TEST SPECIFICATION AND THE FILTER TO USE

This section covers the filter matrix to use for both the line and load conditions. This is mainly empirical, and these should be used as general guidelines. The first

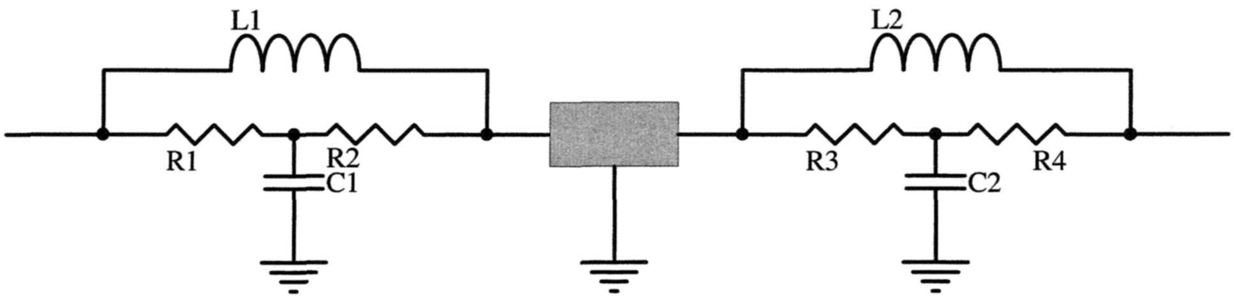

FIG. 6.19 Conventional filter used with DINs and DOUTs.

TABLE 6.1 The Line-Side Matrix

	DC to DC	Single phase	Three phase, 1 Can	Three phase, Inserts
FCC VDE	Single L, π	L, π, or T	L, π, or T	π
TUV				
220A	L, π, or T?	L, π, or T	L, π, or T	π
461	Single L, L, T?	L or T?	L or T	NA
System qual	Single L or L	L or T?	L or T	NA

is the line side covering DC to DC, single phase, and three phase (Table 6.1). The other dimension is FCC, VDE, 220A, 461 and the full system specification. The next part of the matrix is for the load side (Table 6.2). This is for 60 and 400 Hz. The other row is the off-line regulator, with inductor, the power factor correction circuit, and the power line filter. These are general recommendations only.

TABLE 6.2 The Load-Side Matrix

Source	DC	50/60 Hz	400 Hz
Off-line regulator, capacitive load	NA	Inductors must handle the peak current	Inductor must handle the peak current
Off-line regulator, inductive load	NA	Inductor must handle only RMS current	Inductor must handle only RMS current
PFCC	NA	Filter output impedance must be low at switcher *F*	Filter output impedance must be low at switcher *F*
Power line filter	NA	Inductor must handle the peak current	Inductor must handle the peak current
DC to DC	Filter output impedance must be low at switcher *F*	NA	NA
Transformer	NA	Watch primary inductance	Watch primary inductance

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7

Filter Components—The Capacitor

The differential mode components must be of the high-Q type. The individual component Q must be high while the circuit Q must be lowered to the point where the filter does not oscillate. Additional circuits are often added to lower the Q below 2.

7.1. CAPACITOR SPECIFICATIONS

The main specification for the capacitors is MIL STD 15573. The capacitor must meet various voltage ratings. For AC capacitors, the level must be 4.2 times the RMS voltage of the system. For example, in a 220-V RMS system, the capacitor must be designed to handle 924 volts, usually rounded up to 1000 V. For the DC capacitor, the multiplier is 2.5 times the system voltage. For example, a 50-V DC capacitor must be designed to handle 125 V DC, probably made up to 150 V. There may be specifications on certain creepage distances and corona specifications if the voltage is high enough. In the build to print type, the capacitor values may be specified as a minimum or something like ±10%. Regardless of how well the filter satisfies the insertion loss, the filter will be returned as being out of specification if these values are not adhered to. In most filters, if the capacitor value is reasonably higher, the filter will work better giving more insertion loss, but that will not suffice if the value is limited.

7.2. CAPACITOR CONSTRUCTION AND SELF-RESONANCE FREQUENCY

Figures 7.1 and 7.2 show the self-resonant frequency (SRF) of polyester (Mylar) and ceramic capacitors. Robert Hassett (now retired) of RFI Corp., Long Island, New York, provided both drawings. Designs are covered in a later section. Both are feed-through type capacitors giving higher SRF values. Figure 7.3 shows a method used to raise the SRF of either the wrap and fill type or any other style of capacitor. This is called "veeing the cap."

The advantages of the feed-through type are the low equivalent series resistance (ESR) and equivalent series inductance (ESL) due to short lead lengths. This, in turn, means a much higher SRF for the capacitor. The graphs show SRFs up to 1 GHz on the smaller values of ceramic caps. The problem is that the higher the power line RMS voltage becomes, the bigger the margin must be to eliminate creepage and corona. This requires a bigger capacitor and increases the cost and the container cost. As the line frequency increases, the line harmonic current content also increases. This also increases the loss due to *dv/dt*. If the margins were near the lower limits initially, now that the line frequency has increased, the initial margin may have to be increased. In other words, the margin size is not a function of frequency unless the capacitor is close to the margin limit.

Some filter companies use the large nonpolarized can-type filter capacitors. Some of these are oil impregnated. These may be very good capacitors for power supplies and other applications requiring nonpolarized capacitors, but watch out for their self-resonant frequencies. Most of these types have a very low SRF, of

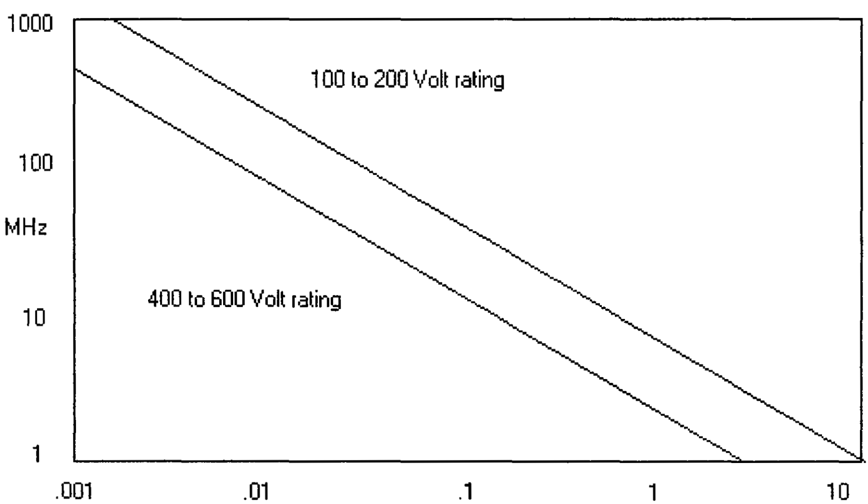

FIG. 7.1 Self-resonant frequency chart for Mylar.

the order of 50 kHz or less. This is where the old power supply trick may come in handy—the "Ye olde paralleling capacitor trick."

This calls for paralleling the capacitor with another capacitor of 0.05 the value of the original. This smaller capacitor would have a much higher SRF and lower ESR and ESL. An article in the IEEE magnetic manuals showed that only 6 dB of gain is realized by these two capacitors in parallel. This theory, though, assumed that the lead length would be close to the same. This would almost make the second capacitor have an ESR and ESL the same in proportion to the original. A feed-through type for the second capacitor would guarantee a workable system because of the very low ESR and ESL and much higher SRF of this high-quality type of feed-through capacitor. Experience shows that paralleling the capacitors often gives serious peaks because of the feed-through oscillating with the ESL of the original capacitor. The purpose of this nonpolarized style of capacitor is to handle the low-frequency requirement of the EMI filter and it should not be required over their SRF point. They often cost less for the capacity and working value; 10, 15, 20 and 30 μF at 480 V AC are available. Another method for lowering the ESL and raising the SRF of the capacitor is as follows.

7.3. VEEING THE CAPACITOR

In Fig. 7.3, there are four inductors that are often deemed insignificant. This is where the three Ls join at the top and represents some tie point or splice, with a

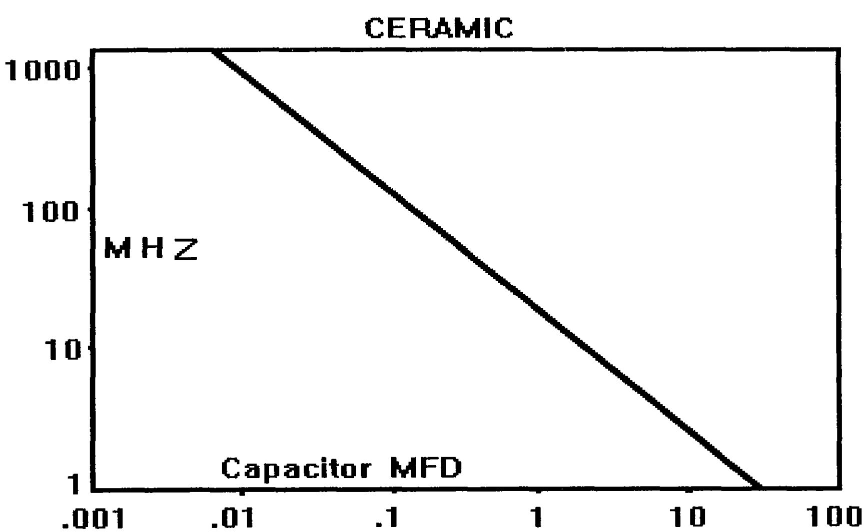

FIG. 7.2 Ceramic self-resonant frequency chart.

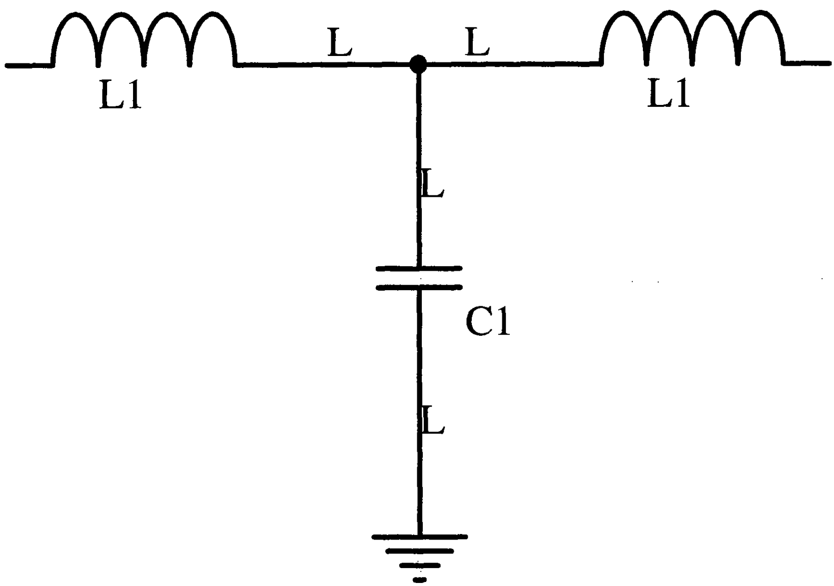

FIG. 7.3 The T filter used to demonstrate veeing.

fourth at the bottom. The two that make up the capacitor leads—the two top and bottom—add to the ESL and ESR, lowering the SRF of this capacitor. The two leads next to the inductors add to the inductor value. This increases both inductors, but they are orders of magnitude lower in value. This is similar to the concept of keeping the lead lengths as short as possible.

Alhough this technique has been around a long time, the new name for it is "veeing the capacitor." This is the old "keep the leads as short as possible" trick, especially on the capacitor side. The leads on the inductor just add a small amount of inductance to the two inductors. However, the leads facing the capacitor, the vertical leads in Fig. 7.3, increase the ESR and ESL of the capacitor, lowering the SRF.

Now, in Fig. 7.4, all the inductance in either leg of the vee adds to the inductors on both sides. The common tie point, or splice, is eliminated and the self-leads of the inductor are wired directly to the capacitor as close to the capacitor body as possible. The large nonpolarized capacitors discussed above, such as 20 μF at 480 V AC, have terminals on the top for wiring directly to. The capacitor ESL and ESR decrease, increasing the SRF. The same can be done to the bottom half if the circuit is balanced (Fig. 7.5).

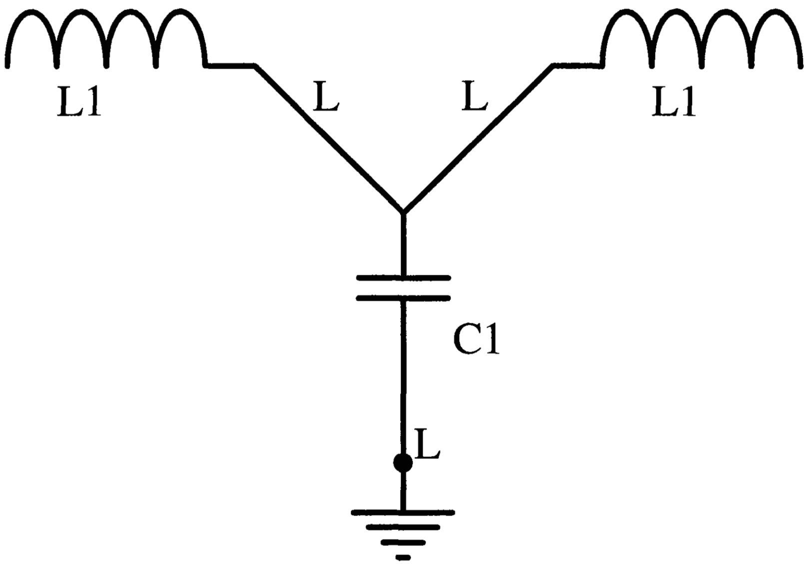

FIG. 7.4 Veeing the capacitor.

7.4. MARGIN, CREEPAGE, AND CORONA—SPLIT FOIL FOR HIGH VOLTAGE

First, margin, creepage, and corona must be understood. The margin requires a space calculated using 16 V per mil. If the working voltage of an AC capacitor is 220 V AC, the quality multiplier for most military specifications is 4.2 for AC; $220 \times 4.2 = 924$. Divide this by 16 = 57.75 mils or 0.05775 minimum margin. Use 3/32 of an inch. See Fig. 7.6. The two plates, or foils, extending to the left will be swedged together. The margin places a gap between the plates. If the margins are adequate, everything is okay.

If the capacitor is used at a voltage over its design rating, the voltage will creep over the dielectric slowly, carbonizing a path as it goes until it arcs to the swedged plates, ruining the capacitor. This will take some time to happen, and the carbonized path will grow slowly over time.

Corona, on the other hand, is high voltage related. To avoid corona, smooth surfaces are required, not sharp points. Capacitor winders require splitters, which spilt dielectric material to the proper width required for the capacitor. In this way, they buy only the wider widths and split the dielectric material for the desired width. The point here is that in the process of splitting, if a point is formed in a high-voltage section, corona will develop. One way to end this is to fold over the edge on the margin side forming a rounded smooth end facing the margin. Using

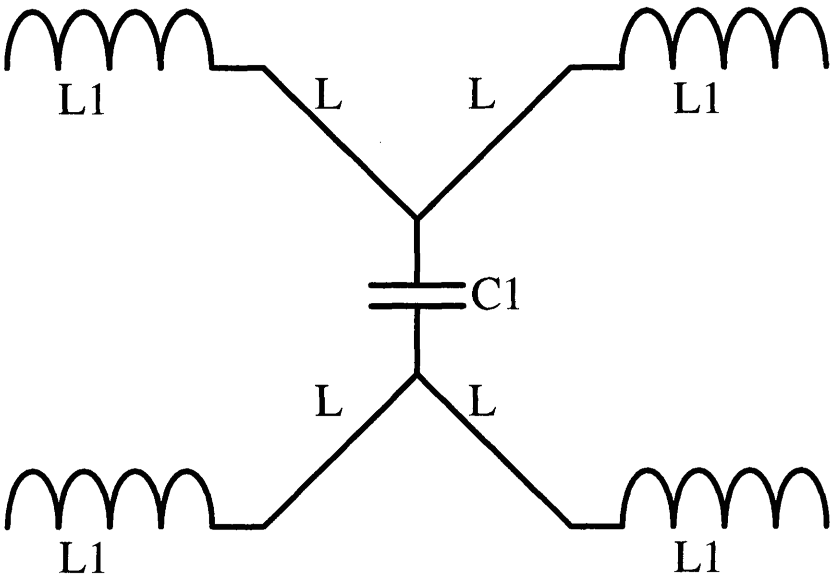

FIG. 7.5 Balanced vee.

this technique, the capacitor becomes thicker with an increased diameter. But this does eliminate all sharp points on the foil. See Fig. 7.7. In high-voltage capacitors, the plates, or foils, are wound in series to divide the voltage.

The voltage in Fig. 7.8 divides across the sections, but the active area is reduced. With two sections shown, the voltage is divided by 2, but the capacitor value must be multiplied by 2. The gap in the middle to separate the two foils must be greater than the value used for the margin. Use 32 mils per volt to calculate this distance. The right and left ends are swedged together. There is no connection to the center foil or plate. These types of capacitors are large and there may be several in series with more than one split or gap.

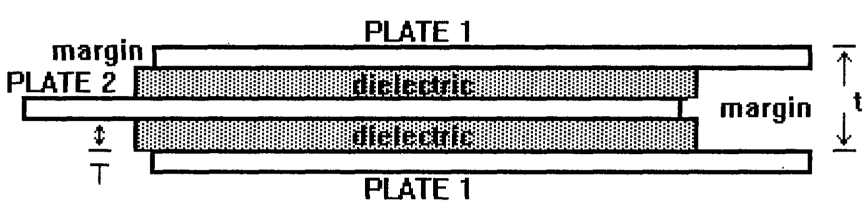

FIG. 7.6 Capacitor plates and margin. T is the dielectric thickness in mils, and t is the thickness of two dieletrics and two foil plate thicknesses in inches.

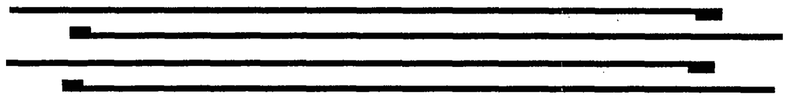

FIG. 7.7 Capacitor with rounded foil.

7.5. CAPACITOR DESIGN

7.5.1. Wrap and Fill Type

Three methods are used to build these capacitors and two subgroups. The initial method was to attach leads to the foil at the start before wrapping, or winding, the capacitor. These capacitors are now called inductive capacitors or chicklets because of the current flow through the foil to charge or discharge the entire length of the capacitor. The ESR and ESL are very high in these chicklets, so the SRF is quite low. As an example, for a foil 0.00023 inch thick and 0.5 inch wide, the cross-sectional area in square inches is 0.000115. Dividing this by 7.854×10^{-7} gives 146.42 circular mils. Dividing 17 by 146.42 gives 0.1161 ohm per foot. If the 0.23 mil aluminum foil is 100 feet long—an average size capacitor—the electron flow average length is half this, but there are two plates. The resistance is 11.61 ohms. The inductance would be of the order of 45 μH based on approximately 30 meters.

These should never be used for the main EMI filter capacitors. They can be used for RC shunts if the frequency is low enough. They can also be used for the Cauer types for low-frequency tuning capacitors, again depending on the frequency. The tuning frequency must be at least half the chicklet's SRF. The chicklets should not be used for the dissipative filter. If the filter has high-frequency problems that the RC shunt must also aid, the quality of the capacitor dictates the higher quality of the extended foil type.

The two subgroups are metallized and foil. For the metallized style, the dielectric is coated with a thin spray of metallized aluminum that becomes the plate of the capacitor, and the other is aluminum foil. The typical aluminum foil thickness is 0.00023 inch, or 0.23 mil. This foil is much thicker than the film,

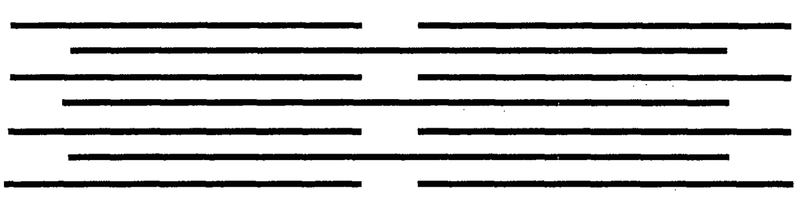

FIG. 7.8 Plates in series, high-voltage capacitor.

which is measured in microns. Foil will carry much more current and is therefore better for pulse applications and EMI filters where there are high harmonic currents from off-line regulators and similar harmonic generators. Most EMI filters are built with this type of construction and have a thicker aluminum foil if higher currents are expected. The metallized film has several advantages, however: this capacitor can be much smaller for the same capacitor value, and this type is self-healing. All dielectrics have small pinholes throughout their length. When the applied voltage stresses the film, the film often shorts out through one of the pinholes, causing the film to melt. The aluminum will reform, making the capacitor self-healing. Another advantage of this subgroup is that the metallized aluminum can be sprayed on both sides of the dielectric. This adds to the smaller size, promotes self-healing, and is better for extended life. To avoid the pinhole problem, several dielectrics are used. Typically, the thickness is 0.24 mil each, giving a total dielectric of 0.48 mil. The odds of two holes occurring together are remote. Film capacitors can be used for DC filters, but the capacitor facing the load must be foil.

The type used mostly in the EMI filter is the extended foil type. The foil extends beyond the winding arbor so that one plate, or foil, extends to the left and the opposite plate extends to the right. The area on the right of the first foil, or the left of the second foil, makes up the margins for both plate ends. It should be obvious why this type is called the extended foil type. The extension is typically 3/32 and each end is soldered (called swedging) for the contacts. The ESR and ESL of extended foil capacitors are both very low because the current flow travels only the average width of the capacitor foils, called the gauge, or height, of the capacitor. Also, the turns are in parallel, so the average diameter times the number of turns times the thickness gives the square inches of the aluminum. Convert this to circular mils (divide by 7.854×10^{-7}), and divide this into 17 for aluminum (10.374 for copper) to obtain the ohms per foot. Dividing this by 12 and multiplying by the gauge (the height of the capacitor) give the approximate resistance of the capacitor. This would be a close approximation of the ESR. Inductance is approximately 1.5 μH per meter. Divide 1.5 by 39.36 to get 38.1 nH per inch. This times the gauge gives a first-order approximation of the ESL. It should be apparent why a large capacitor diameter to length, or gauge, is desirable. A 2:1 ratio of the diameter to the height, or gauge, is the optimum ratio. If the ratio increases too much over this optimum, the capacitor starts to wobble on the winding arbor and the active plate area is diminished. The capacitance value drops but the gauge is also more than desired. The SRF is very high for this optimum type and should be used by the EMI filter designer.

The last method is similar to the extended foil type but is used for high capacitor current applications. This is called the tab type. The size and number of tabs depend on the current. These tabs are thin strips of conductor that are placed in the winding as the capacitor is being wound. These tabs are inserted—one for each plate—every so many turns and so many degrees are added to each tab so that the tabs end up uniformly spaced around the sides of the capacitor. The tabs extending out of each side of the capacitor are folded over and soldered together to form the contact.

Most new people in the EMI design arena are knowledgeable about power supply design, where the main concerns about the capacitors are their working voltage and capacitor value. Others are familiar with derating of the capacitors, and this quite helpful. The point is that AC capacitors must be designed to handle the total AC currents at the line frequency and the odd harmonics. If this is a single L or T, the capacitor should be of the foil type, not metallized film. In π and multiple filters, at least the last capacitor, the capacitor closest to the load side, should be of the foil type. See the next section.

Some capacitors are not designed properly to handle the full AC current flow or were designed for DC applications. This is especially true if the capacitors are the metallized film type. The capacitor designer may not have designed the capacitor to handle the total harmonic current. Capacitors must handle the harmonic current from either the line frequency side or the load. This is especially true for the harmonic current created by the off-line regulator or any power supply using a capacitor input filter. The foil making up the capacitor plates can be too thin, measured in microns, to handle this current. This raises the ESR losses, so the capacitor will heat and will fail in the months ahead. Many factories take the defunct unit apart and replace the blown capacitors with the same capacitor, stating, "We've always used this capacitor in these units." They are not aware that the capacitor has the wrong rating for this application. Commonly, the capacitor was designed for DC operation and not for AC by the capacitor manufacturer and was selected in error by the original filter designer. The filter capacitor must be selected to handle the harmonic currents of the off-line regulator and any other pulse type with high harmonic currents (Figs. 7.9 and 7.10).

If the line harmonic current is neglected, the line voltage is approximately sinusoidal and the load current equation, per Sec. 4.2, is

$$I = E_M \cos(X) \frac{\sin(\omega t - Y - X) + \sin(Y + X - A) e^{\frac{A - \omega t}{\tan(X)}}}{(R_S + R_L)\cos(Y)} \tag{7.1}$$

See Sec. 4.2 for the method of calculating the various constants. Even with all the constants inserted in the equation, further analysis is difficult. There is an easy method from days gone by that is not even taught now. This method is included on the disk provided with this book.

$$\frac{167\cos(0.346087)\left[\sin(\omega t - 2.03688) + \sin(1.235) e^{\left(\frac{0.801779 - \omega t}{0.361}\right)}\right]}{23\cos(1.69079)} \tag{7.2}$$

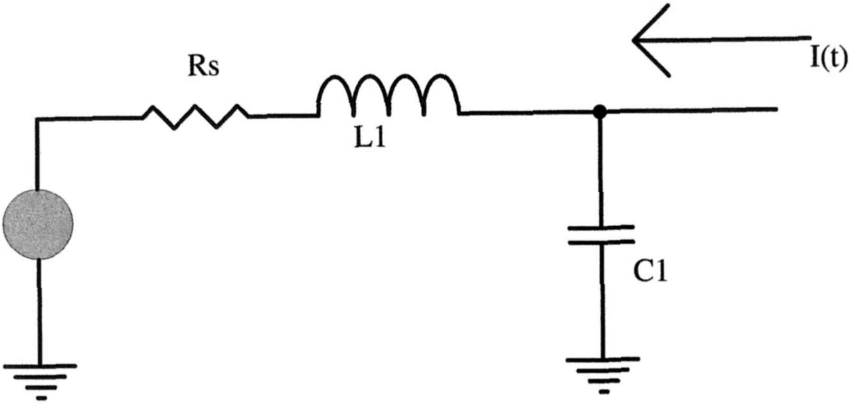

FIG. 7.9 The line voltage, E_S, the source impedance, R_S, filter, and $I(t)$.

Further reductions give

$$57.059\left[\sin(\omega t - 2.03688) + 0.944e^{\left(\frac{0.801779 - \omega t}{0.361}\right)}\right] \tag{7.3}$$

However, this is still no fun even with Fourier analysis. Here, I have left the work to a computer spreadsheet. With the speed of computers today, this is an easy task. I put all the constants at the top of the spreadsheet. In column A, I listed the angles from 0 to 2π in 360 steps. In column B, I listed the equation using IF statements—{IF (angle<A, 0, If (angle >B, O, Equation}. Here, the start angle A and stop angle

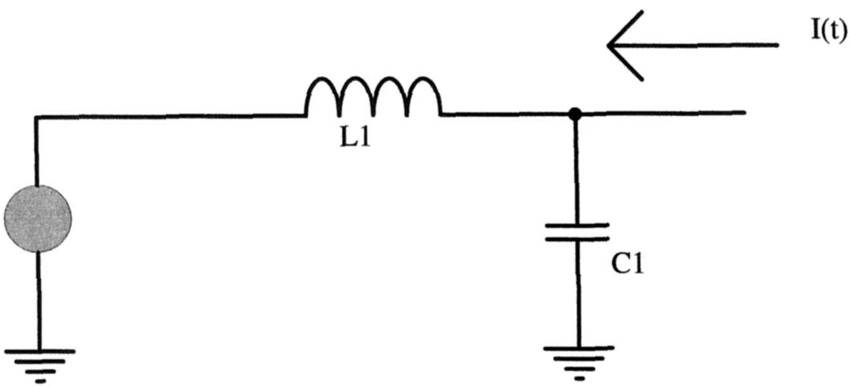

FIG. 7.10 Total harmonic current without the source resistance, R_S.

B must be known. The rest of the columns are A_n values followed by the B_n values and the calculation of C_n. See the Appendix and disk. The value of each harmonic and phase angle is given. The spacing can be changed so that only odds or evens or both odd and even can be looked at. Starting with a step of one, both the odd and even appear. If the evens are extremely small, eliminate them by changing the step from one to two (or vice versa). The fundamental current phase angle and the true to apparent current ratios and hence the power factor are also given. Two different spreadsheets are given—one for the line side, with all odds, and one for the storage capacitor side, with all evens. These Lotus programs can be used for any other wave if an equation can be developed.

The source resistance is low over the range of harmonics of interest and can be eliminated. The inductors and capacitors making up the filter can be lumped or added because the frequency of interest is well below the cutoff frequency of the filter for the most part. But, as the harmonic number increases, the capacitive reactance decreases and the capacitor current increases.

If this is a double L, the two inductors can be added (2L) and so can the capacitors (2C). (if they are the same—they should be).

The harmonic current for the line side—odd harmonics—first through 25 are as follows: 11.97, 8.56, 4.39, 1.06, 1.17, 0.86, 0.30, 0.5, 0.28, 0.2, 0.26, 0.11, 1.6. The square root of the sum of the squares is 15.65 and the RMS line current is 11.06. The difference is 4.59 or 4.6 A through the capacitors. An extended foil should be able to handle this in most cases, but can a metallized one? The metallized capacitor would depend on the size but probably not hold up in the long run. As for the chicklet capacitors, they would not stand the current for any length of time.

The solution here may be to replace this extended foil capacitor with the tab type if the total capacitor current is excessive. (See Table 7.1 through 7.4.)

TABLE 7.1 Chart of Voltage Ratings

Maximum AC rating	Maximum DC rating	Minimum margin (inch)
220	399	3/32
440	600	1/8
660	1,000	3/16
	1,500	3/16
	2,000	1/4
	2,500	1/4
	4,000	7/16
	6,000	1/2
	10,000	3/4
	15,000	1

TABLE 7.2 RFI Corp. Dielectric Rating

DC volts	AC volts	Mylar	Paper
100		1X 0.00025	2X 0.00025
400	120	2X 0.00025	3X 0.00025
600	250	2X 0.00035	3X 0.0003

Source: Robert Hassett, VP Engineering, retired.

For 3/32 inch and above use 1/8 if possible. For paper more than 4 inches wide, add 1/16 inch to all readings.*

Note that there are two foils and two dielectrics per thickness in the three columns. Also, some extra allowance for bends and so on is included in the preceding chart.

As the line frequency rises, the margins must be increased if the margins were on the ragged edge or if the loss is due to dv/dt increase. The margins also increase with the voltage as the information in the tables suggests. This must be done to handle creepage. These charts and drawings are included to help the designer. In the design, the first step is to determine the dielectric.

Polycarbonate is good without derating from –55 to +125°C. and has a low value of the dissipation factor, giving a higher Q capacitor. This type of dielectric is easily contaminated by impurities either in storage or during winding. However, the use of this material is greatly reduced because of the high cost.

Polyester (Mylar) is good for the same range if the voltage is derated by 50%. Polyester is also rated from –55 to +85°C without derating. This has a higher dissipation factor or the lowest Q. It is used more for 60 Hz and DC, especially the metallized type. For constant DC operation, there is no current flow though the capacitor, eliminating heat buildup in the metallized capacitor.

Polypropylene is also rated from –55 to +85 and to 105°C with a 25% voltage derating. The Q is higher, or there is a lower dissipation factor, than for Mylar (polyester). The cost of this material is now less than that of Mylar. This material is suggested for 400-Hz capacitors.

Paper in oil-filled caps is also good to +125°C, and the quality of the capacitor is similar to that of Mylar. However, with the costs being lower for polypropylene and Mylar, this is rarely used. This style must be

*All information in this section was provided by RFI Corp., which has been very helpful for many years. This is especially true of all the people in engineering. Without their help, this book, and much of the seminar material, would not have been available.

impregnated, whereas Mylar, polypropylene, and polycarbonate do not require impregnation.

The required thickness of the dielectric depends on the voltage. This is true with or without derating, and the dielectric material is most often double, triple, and quadruple lapped as the preceding tables suggest. This is to find the proper thickness for the voltage. The best way is to make sure that there is no more than 400 volts RMS per mil of dielectric material regardless of the dielectric material used. Each has a different value of K, the dielectric constant, needed to calculate the active length of the capacitor. Next, the full thickness of the dielectric plus the film must be determined based on the voltage derating listed in the tables.

If 200 V is needed at 105°C, the thickness must be good for 266 V because of the derating based on polypropylene and foil.

TABLE 7.3 RFI Corp. Oil-Impregnated Capacitors—Dielectric Paper to +125°C

Ply thickness X	DC volts	Test Flash	Test 1 min	AC volts	Test Flash	Test 1 min
2X 0.0002	65	160	130	35	90	70
2X 0.00025	100	250	200	55	140	110
2X 0.0003	160	400	320	85	215	170
3X 0.0002	200	500	400	95	240	190
2X 0.00035	200	500	400	120	300	240
1X (0.00035 + 0.0004)	215	540	430	130	315	260
2X 0.0004	240	600	480	175	440	350
3X 0.00025	400	1000	800	200	500	400
2X 0.0003 + 0.00025	500	1250	1000	240	600	480
3X 0.0003	600	1500	1200	300	750	600
2X 0.0003 + 0.0004	750	1875	1500	330	825	650
2X 0.0004 + 0.0003	800	2000	1600	340	850	680
4X 0.00025	850	2125	1700	360	900	720
3X 0.00035	850	2125	1700	360	875	700
3X 0.0004	1000	2500	2000	400	1000	800
4X 0.0003	1200	3000	2400	440	1100	880
3X 0.0005	1300	3250	2600	500	1250	1000
2X (0.00035 + 0.0004)	1500	3750	3000	550	1375	1100
4X 0.0004	1600	4000	3200	600	1500	1300
4X 0.00045	1800	4500	3600	660	1650	1320
4X 0.0005	2000	5000	4000	720	1800	1440
5X 0.0004	2200	5500	4400	880	2000	1600
5X 0.0005	2500	6250	5000	1000	3500	2000

Source: Robert Hassett.

TABLE 7.4 RFI Corp. Mylar with Aluminum Foil Thickness of 0.00025 – *t* in Equations

Mylar thickness	Plate pressed	Flat	Round
0.00040	0.00134	0.0014	0.00142
0.00045	0.00144	0.00150	0.00153
0.00050	0.00154	0.00161	0.00164
0.00055	0.00164	0.00172	0.00175
0.00060	0.00174	0.00182	0.00186
0.00065	0.00184	0.00193	0.00197
0.00070	0.00194	0.00204	0.00208
0.00075	0.00204	0.00215	0.00219
0.00080	0.00214	0.00225	0.00230
0.00085	0.00224	0.00236	0.00241
0.00090	0.00234	0.00247	0.00252
0.00095	0.00244	0.00258	0.00263
0.00100	0.00259	0.00268	0.00274
0.00105	0.00264	0.00279	0.00285
0.00110	0.00274	0.00289	0.00296
0.00120	0.00294	0.00311	0.00318
0.00130	0.00314	0.00332	0.00340
0.00140	0.00334	0.00354	0.00360
0.00150	0.00354	0.00375	0.00364
0.00160	0.00374	0.00396	0.00406
0.00180	0.00414	0.00439	0.00450
0.00200	0.00454	0.00482	0.00494
0,00210	0.00474	0.00503	0.00516
0.00220	0.00494	0.00523	0.00538
0.00230	0.00514	0.00546	0.00560
0.00240	0.00534	0.00568	0.00582
0.00250	0.00554	0.00589	0.00604
0.00280	0.00614	0.00653	0.00670
0.00300	0.00654	0.00696	0.00714
0.00350	0.00754	0.00803	0.00824
0.00400	0.00854	0.00910	0.00934
0.00450	0.00954	0.01017	0.01044
0.00500	0.01054	0.01124	0.01154
0.00550	0.01154	0.01231	0.01264
0.00600	0.01254	0.01338	0.01374

Source: Robert Hassett.

7.5.2. Round Capacitor Calculation

A 4-μF feed-through capacitor is required for 220 AC, 60 Hz, requiring an arbor 0.125 inch in diameter (see Figs. 7.11 and 7.12). First determine the material, here Mylar. The 4 μF almost eliminates metallized Mylar, but these are made by many companies and would be cheaper than building in house, unless the quantities are high enough. So extended foil is the choice. Divide the 220 RMS volts by 400 V per mil and get 0.55 mil. The available thicknesses are typically 0.24 and 0.32 mil. Two at 0.24 mil give 0.48 mil, which is too small. One 0.24 and one 0.32 give 0.56 mil and are usable if the capacitor designer is pressed for room. Here, two at 0.32 should be used, giving 0.64. This is the value of t in the formula, the total dielectric thickness.

Next, the foil thickness is determined and typically is 0.23 mil. The winding machine will then have six rolls of material—four Mylar and two foil. The value of the setup, in inches, is $2.2 \times t$ (in inches) $+ 2 \times 0.00023 = 2.2 \times 0.00064 + 0.00046 = 0.00187$. The 2.2 is to allow for the growth due to winding. The foil width is the dielectric width of the Mylar less one margin plus an extension. Here, $1 - 0.09375 + 0.09375 = 1$ inch. Solve for the active foil width:

$$A_{(FW)} = F_W - 2 \times E_{XT} = 1 - 2 \times 0.09375 = 0.8125 \tag{7.4}$$

where F_w is the foil width and E_{xt} is the extension, both in inches.

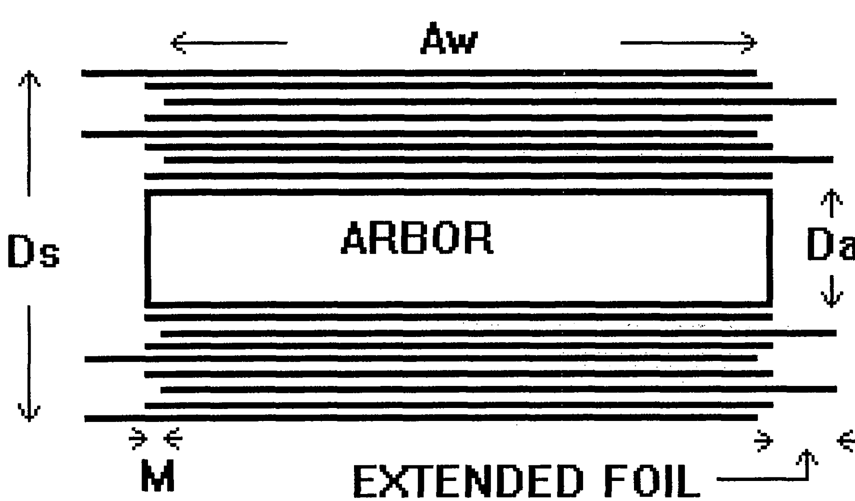

FIG. 7.11 Extended foil capacitors.

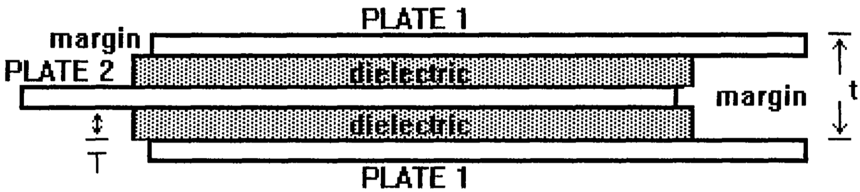

FIG. 7.12 Close-up view of extended foil plates with dielectric. T is the dielectric thickness in mils, and t is the thickness of two dielectrics and two foil plate thicknesses in inches.

Next, the active foil length is calculated, but first the capacitor constants, K, are needed for round feed-through capacitors. The typical K values are as follows:

Polyester (Mylar)	900
Polypropylene	1200
Polycarbonate	840
Paper (resin or PBT)	580 (wet), 550 (dry)

Solve for the active length of the foil:

$$\frac{KC \times t \times 1000}{A_W} = \frac{900 \times 4 \times 0.64}{0.8125} = 2835.7 \tag{7.5}$$

where K is the constant, C is the capacitance, and t is the total dielectric thickness. Next is the foil area. This is 1.27 × setup × the active length.

$$1.27 \times 0.00187 \times 2835.7 = 6.73 \tag{7.6}$$

From this value and the arbor diameter, the outer diameter (OD) of the capacitor is calculated. This is the square root of the foil area plus the arbor squared.

$$OD = \sqrt{F_a + A_{rbor}^2} = 2.597 = 2.60 \tag{7.7}$$

At this time, it is wise to check the gauge. This is the Mylar width plus two extensions, assuming there is no problem with the diameter for fit at 2.6 inches. Extra room is required for installation.

$$\text{Gauge} = D_W + 2 \times E_{xt} = 1 + 2 \times 0.09375 = 1.1875 \tag{7.8}$$

Compare the gauge with the diameter, here 2.6 /1.1875 = 2.2. This is close to the optimum ratio, giving close to the minimum ESR and ESL and resulting in the highest SRF. The winding is practical.

Next is the number of turns.

$$N = \frac{(\text{OD} - \text{ID})}{2 \times S_u} = \frac{(2.6 - 0.125)}{2 \times 0.00187} = \frac{2.475}{0.00374} = 662 \tag{7.9}$$

7.5.3. Pressed Capacitor

The arbor is removed after winding and the capacitor is then pressed flat with weights overnight. To calculate the flat sides from the round above, a stacking factor of 0.95 is needed. The small flat is

$$S_{(\text{flat})} = 2 \times (N + 2) \times S_U S_F = 2 \times 664 \times 0.00187 \times 0.95 = 2.36 \tag{7.10}$$

and the large flat is

$$L_F = 1.1 \times \left(\frac{S_{(\text{flat})}}{S_F} + \frac{\pi A_{(\text{rbor})}}{2}\right) = 1.1 \times \left(\frac{2.36}{0.95} + \frac{\pi \times 0.125}{2}\right) = 1.1 \times (2.48 + 0.196) = 2.94 \tag{7.11}$$

This calculation was based on the round capacitor. If a flattened capacitor is being designed, the K constants do change:

Polyester (Mylar)	800
Polycarbonate	750
Paper (resin or PBT)	515 (wet)

Other equations needed for the winding for either the flat or the round capacitors are:

Dielectric width = $Aw + 2M$
Foil width = $Aw + M + P$
Gauge = $Aw + 2M + 2P$

where P is the foil overhang that is sputtered over to make the lead connections on each end or form the end area for the feed-through type.

For the feed-through type (Fig. 7.13), the arbor tube is left in place and the maximum filter current, not the capacitor current, must be known. A conductive bolt, or a threaded rod, probably not copper because copper lacks the strength necessary to handle the torque required by the nuts, is silver plated by a military specification plating company (assuming a military contract). This bolt is designed to carry the full filter current and must fit through the center of the arbor. The diameter needed for this material to handle the current with low heat rise is often the factor determining the inside arbor diameter. This bolt is often plated to

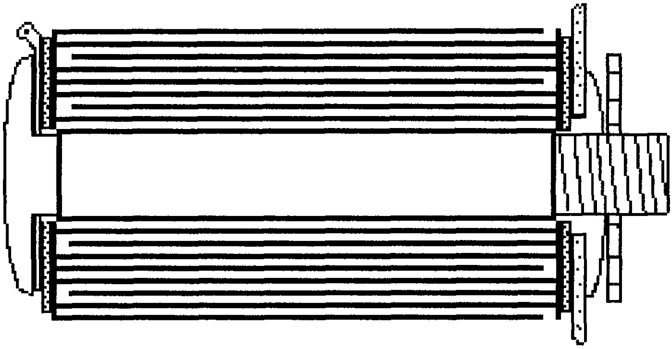

FIG. 7.13 Feed-through capacitor construction showing the case.

help handle the current and normally forms the outside stud for electrical connections for the output.

The construction is as follows. A plated conductive bolt is threaded through the contact washer or terminal, the plated conductive washer, and the body of the capacitor through the arbor hole. The washer presses against the inside sputtered end for good contact. The diameter of the conductive washer is the full diameter of the capacitor, so the silver-plated washer covers the entire extended foil area that has been swedged. Another conductive washer follows this full arrangement. It is then placed through the larger container hole. An insulator washer, usually made of ceramic, is slipped on the end bolt sticking through the can, followed by a nut tightened to a specific torque. This presses the last washer and the end of the capacitor to ground, forming one terminal. The hot side is formed by both the bolt on the outside of the can and the connection on the inside of the can. Some use a plate, called a ground cup and also ferrules, that has been stippled to press into the extended foil area to replace the two silver-plated washers. These numerous indentations or stipples help to maintain prolonged electrical contact.

8

Filter Components—The Inductor

The EMI filter requires what is termed a "soft core." This really means that the core is driven into saturation slowly rather than abruptly as required for pulse transformers and magnetic amplifiers. A hard core can be made soft by gapping the core. This technique tilts over the *BH* curve (Fig. 8.1), making the core a soft core—harder to drive into saturation.

8.1. INDUCTOR STYLES AND SPECIFICATIONS

Inductors for EMI filters come in three styles: tape-wound, toroids, C cores, and slugs. E cores, pot cores, and RM cores are rarely seen. There are several subgroups. For toroids, the styles are ferrite, powdered iron, MPP, and high-flux (HF). For C cores, or cut cores, the styles are various steel mil thicknesses as well as steel types. The same is true for the tape-wound toroid. For the slug, other than size, there are various mixes.

As far as specifications are concerned, some filter customers specify the core, the wire size, and the inductor value, usually as a range such as ±10%. Others specify creepage distance or list a specification that calls out the distance. In this case, the clearance listed is between the top and bottom wires to the core. For the tape-wound toroid this does not apply. This is more like a margin along the coil form. Sometimes a wire size is specified.

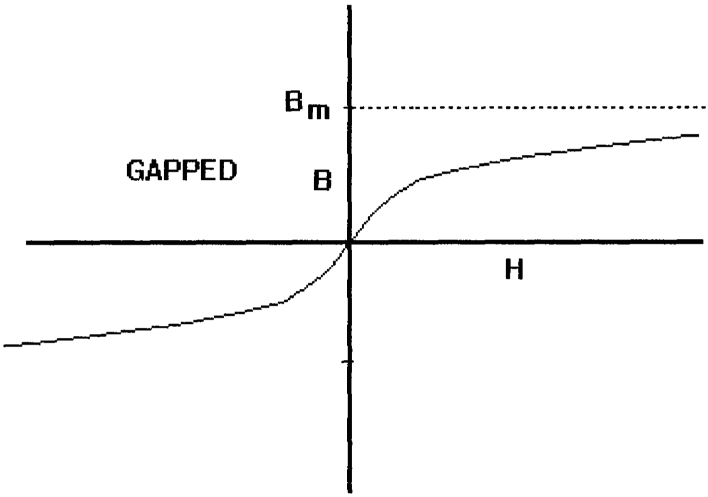

FIG. 8.1 *BH* curve for "soft" core material.

8.2. THE POWDER CORES

The core types used for EMI filters are often Molypermalloy toroids and sometimes the HF type. Powdered iron toroids are also sometimes used but often cause trouble at 400 Hz because they overheat. The main advantage is that all three types of cores can have very high Q values. This decreases the losses and also allows the coils to be "tuned" if necessary. The permeability of MPP or HF types should be limited to 125 and below at 400 Hz for the same reason. This is developed as follows using MPP cores from Magnetics, Inc. through Col. W. T. McLyman, Jet Propulsion Laboratory, Pasadena, California (now retired).

Core manufacturers provide important information about their cores such as outer diameter, inner diameter, height, window area (W_a), cross-sectional area (A_c), weight, and magnetic path length (M_{pl}). If the window area is given in circular mils, as Magnetics, Inc. does, the designer can choose as many circular mils per ampere as needed. The maximum number of turns that will fit in the core window can be calculated. As the current required increases, the turns that will fit in the window drop, and as the needed current falls, the number of turns increases.

For any circular mils per ampere that are chosen, NI is a constant, making H a constant for the core. From H, B can be found utilizing the information provided by the manufacturer and using the winding or fill factor of 0.4 for the window of the toroid. Please note here that this is not the proper way to design the inductor. This is done only to ensure that all the cores in the series are equally wound with the same fill factor.

$$N = \frac{0.4W_a}{C_M I_{rms}} NI_{rms} = \frac{0.4W_a}{C_m} H = \frac{0.4NI_{rms}}{M_{pl}} = \frac{0.16\pi W_a}{M_{pl}C_m}$$

where W_a is the window area in circular mils, C_m is the circular mils per ampere needed for proper wire sizing and low DC resistance (DCR), M_{pl} is the magnetic path length in centimeters, and 0.4 is the typical window fill factor for toroids.

Magnetics, Inc. lists core number 55438 for permeability 125. The window area in circular mils is 842,700, the magnetic path length (M_{pl}) in centimeters is 10.74, and the weight in pounds is 0.4. Using 600 circular mils per ampere, H follows.

$$\frac{0.16 \times 842{,}700 \times \pi}{10.74 \times 600} = 0.65734$$

for each permeability using the *BH* curves furnished by the manufacturer yields approximate values of B (Table 8.1). The given, or initial, permeability is listed as a reference and B was estimated from the *BH* curve. With this information, the approximate value of the actual permeability of the core was calculated. The watts per pound follow from these formulas, also provided by the manufacturer for MPP cores (Table 8.2). From the preceding equations the next column, watts/pound, was determined and the last column followed on multiplying the W/lb by the pounds furnished by the manufacturer, 0.4 here.

The watts per pound and total loss columns show a hefty jump from permeability 60 to 125 and continue hefty increases upward. Using permeability 125 is questionable. The same thing happens with powdered iron at even lower permeability. Reducing the frequency back to 60 Hz makes a large difference and

TABLE 8.1 The Permeability and Flux of a Core

Given *u*	Actual *u*	*B*	400 Hz (W/lb)	Loss (total)
14	13.69	900	0.0377	0.0151
26	25.86	1700	0.1827	0.0731
60	45.64	3000	0.3091	0.1236
125	83.67	5500	1.0576	0.4230
147	86.71	5700	1.3471	0.5388
160	95.84	6300	1.6806	0.6722
173	99.64	6550	1.8315	0.7326
200	103.45	6800	1.8448	0.7379
300	109.53	7200	2.3278	0.9379
550	117.14	7700	3.9327	1.5730

TABLE 8.2 Watts per Pound for MPP Cores

14	$0.312 \times 10^{-11} \times F^{1.32} \times B^{2.25}$
26	$0.360 \times 10^{-10} \times F^{1.16} \times B^{2.07}$
60	$0.828 \times 10^{-11} \times F^{1.23} \times B^{2.12}$
125	$0.489 \times 10^{-11} \times F^{1.28} \times B^{2.14}$
140, 160, 173	$0.425 \times 10^{-11} \times F^{1.23} \times B^{2.21}$
200	$0.788 \times 10^{-11} \times F^{1.35} \times B^{2.06}$
300	$0.890 \times 10^{-11} \times F^{1.26} \times B^{2.11}$

all the different permeabilities can be used. If this idea is violated, the core can overheat. This concept can be overlooked if the inductor design is very conservative and if the winding factor is low, with the turns well below the maximum or the window area of the core less filled. An example occurs where there is a single wire layer on the core. This is given as fair warning of things that have happened at several EMI filter companies. Refer again to Table 8.1.

Note that the flux density is too high at 125 and above anyway.

Another way to show the same thing is to require the same inductance at the same current. The following equation of A_l values for Table 8.3 was developed using the same core as above at 5 A RMS at 380 μH. See the next section for a discussion of A_l values.

$$I_p = \sqrt{2}I_{rms} \qquad N = 1000\sqrt{\frac{L}{A_l}}$$

$$B = \mu H \qquad H = \frac{0.4\pi N I_p}{M_{pl}}$$

The wire size is constant using the 600 circular mils per ampere, so No. 16 AWG is used. I_p is 7.07 A to determine H.

The B term was determined using the initial permeability, and the error increases with increased permeability. These could have been read off the BH curve, and the values would be less than those listed in Table 8.3. Regardless of how the value of B was determined, these should not be used much above 3000 gauss, so permeability 125 *might be okay*, but 147 and above should be avoided.

Another toroid is the ferrite one. This is used mainly for common mode. The ferrite core lacks the distributed gap so the BH curve is too square looped for a differential mode inductor. The ferrite toroids saturate at the coercive force level. However, they work well for common mode, where the sum total of currents through the center is zero and creates a balance of flux that cancels. Another disadvantage is that they are very noisy when driven into saturation. This shows that they should not be used as differential mode inductors.

TABLE 8.3 Constant Current at 5 A for MPP Core Showing A_l Values

Permeability	A_l	N	H	B
14	32	109	90	1260
26	59	81	67	1742
60	135	53	44	2640
125	281	37	31	3875
147	330	34	28	4116
160	360	32	26	4160
173	390	31	26	4498
200	450	29	24	4800
300	674	24	20	6000

C cores are used especially at the higher current lines, mainly the power line filters used for shelters, EMI test houses, screen rooms and secure communication applications. The disadvantage is their low Q values, and they are usually not effective much above 100 kHz. The thickness of the magnet core tape used to wind these C cores should be able to handle the fifth harmonic of the power line frequency. This dictates a thinner core material but increases the Q and gives further life at the higher frequencies. The potential harmonic current from off-line regulators is also improved using the thinner core material. This should be done to avoid heating the core. This is because of the harmonics on the line and from these off-line regulators. If the filter is to be used in areas where the power feed is more resistive with less harmonic content, this principle can be violated. Also, if the inductor was designed with a low temperature rise, then the core can take the extra heat. In this way, the wire and the core loss add to the filter via dissipation. Where the high current spikes can flatten the voltage, creating high levels of harmonic current and also some harmonic currents from the load, this principle, or the following option, cannot be ignored. The other option is to design the inductor using very low flux density and allow the line harmonics to heat the core.

The costs of the core and filter are increased by either method. The first allows a smaller core that is more expensive, and the second allows a less expensive type of core but is often offset by the larger size and weight. One other factor not to overlook is the smaller thickness of tape used in the first method, which has a lower stacking factor, decreasing A_C, the core cross-sectional area. The value of A_C, then, must be increased. The higher switcher frequencies and a parasitic oscillation from the load side should be allowed to dissipate their energy within this core material via eddy currents and hysteresis whatever the thickness of the tape. Additional losses for these higher frequencies can be gained in the skin effect of the wire. Allow both methods to dissipate this high-frequency energy.

The core material thickness that is normally applied is 11 or 12 mil for 60 Hz, 4 mil for 400 Hz, and 2 mil for DC for typical applications. Core thicknesses available are 1, 2, 4, 7, 11, and 12. These often drop down one size as described earlier. This reduces the core loss, lowers the temperature rise, raises the Q, and increases the self-resonant frequency (SRF). The tape-wound toroid of the same material has about the same constraints except that Q is higher and the turns are typically lower.

High currents demand large wire diameters. Welding cable is often employed because it has many strands such as 833 strands of No. 30. This makes the wire flexible, allowing it to be wrapped around the coil form. I have known many engineers who wanted to use Litz wire and high-frequency cores, which would increase the cost several orders of magnitude. But, for EMI filters, either Litz or welding cable should be used only for ease of winding the coil form. If the coil is to be tuned, the Q must be high at the tuned frequency. This calls for a core with thinner material and stranded wire, so the AC resistance will be low at and somewhat above the tuned frequency. For example, a 60-Hz inductor is to be tuned to 25 kHz. Change the core from a National Arnold CA to a CH. This moves the material from a 12- to a 4-mil core. The radius of conduction in centimeters is 6.62 divided by the square root of the frequency. Up the frequency to 30 kHz and take the square root of it, which gives 173.2. Multiply by 6.62 and divide by 2.54 times the 173.2, and the wire diameter in inches is now known. This is 0.030 inch. A No. 20 wire is 0.032, which is close enough, and this wire has 104.2 circular mils. If 10 A is to be carried, approximately 5000 circular mils is needed, which would require 48 strands of No. 20 wire. It should be apparent that Litz may be the best choice.

The C core is easy to gap using shims of half the gap value in each leg, showing that this would be good for DC operations and would make the core softer. They are easy and quick to wind on a bobbin or coil form, whereas toroids are much slower to wind. If many of these inductors are needed, several C core bobbins or coil forms may be wound at the same time, depending on the wire size, whereas a single toroid must be wound at a time. Usually, there is not enough demand for this inductor to warrant the time to set this production up. Therefore, C cores are quicker to manufacture.

Slugs, or plain cores, are also used for EMI filters but are harder to calculate because the flux path is not complete as in toroids or C cores. The trouble is that the flux can be routed through surrounding pieces or through the filter container. Give them extra clearance from the container and other surrounding components to reduce the flux through them. If another magnetic material is too close, the flux paths shift and also increase the inductance. This stray flux can heat the container and other objects, if violated. Usually, this type of inductor is in the low microhenry range. I have designed them to 158 μH and placed them literally hanging in the middle of the container and they have functioned well. The slug size was 1 inch in diameter and 2 inches long using Micrometals' P6464-140. The equations are:

$$L = \frac{(0.8U_{EFF}(RN)^2}{(6R + 9T + 10B)} \qquad N = \frac{1}{R}\left[\frac{L(6R + 9T + 10B)}{0.8U_{EFF}}\right]^{\frac{1}{2}} \tag{8.1}$$

This is for multilayer windings, where U_{EFF} is the effective permeability, R is the radius, T is the length, D is the core diameter, and B is the coil build (Fig. 8.2). The value of U_{EFF} is a low percentage of the value of because the T-to-D ratio used in most EMI filters is so low, typically 1:1 or 2:1. The estimation in most cases is a U_{EFF} of 5.

The SRF of the inductor is a function of the size and number of turns. There are several ways to raise the SRF for the toroid. The first is to use the progressive winding technique, such as six winds forward around the core followed by five turns back over the previously wound turns, then another six turns forward over the previously counterwound turns followed by another five turns back (Fig. 8.3). Continue until all the turns are on the core and the number of turns can be placed around the core without making much more than 320 degrees of the circumference of the core window. The 6/5 ratio can change to place the last turn at approximately 20 degrees separation from the start lead. Place a barrier or space between the start point and the finish point. This barrier is used so that the two ends cannot slide together and could be anything such as tape. This technique decreases the turn-to-turn capacitance that is then in series across the inductor.

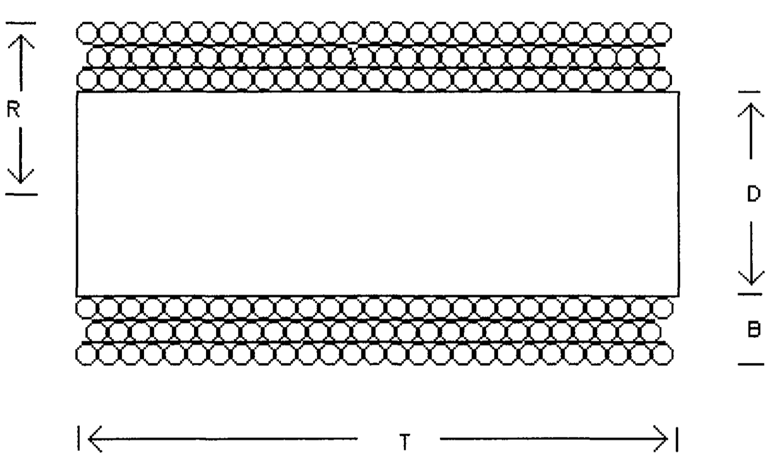

FIG. 8.2 Slug.

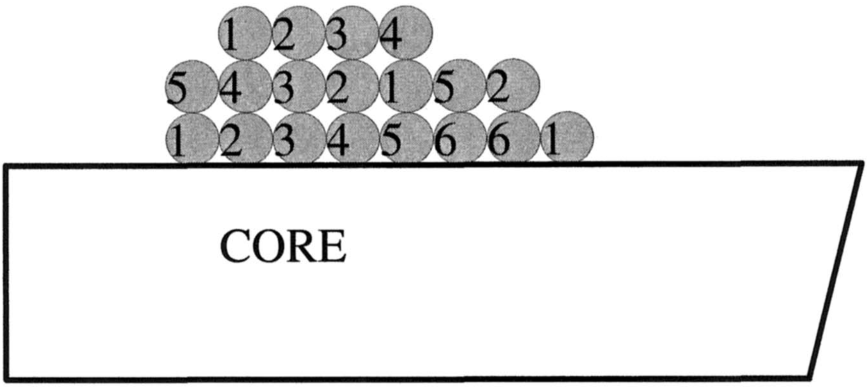

FIG. 8.3 The progressive wind, a higher SRF.

This works because the capacitance varies with the voltage difference between the turns and the voltage difference is small for the nearby turns using this technique. The disadvantage is that this looks as if a 5-year-old child wound it. People normally cover it with tape to help hide this because if uninformed customers saw it they might question the ability of the company.

Windings that are side by side where there is less than a full 360 degrees around the core are also satisfactory, giving lower capacitance from turn to turn for the same reason. The voltage from turn to turn is very low. Where multiple layers are needed, it is obvious that layer 2 to layers 1 and 3 has much higher turns, making a higher voltage gradient and giving much higher capacitance from turn to turn. This guarantees low SRF, which compromises the filter action. The second way is to keep the winding off the core. Tape the core with several thicknesses to keep the turn-to-core capacity as low as possible. Toroids that are already coated can be purchased. Often a layer of tape is needed the raise the wire farther off this core. This capacitance is from turn to core and is somewhat in parallel, so the capacitances add and are more directly in parallel with the inductor. Figure 8.4 shows the core with tape and a single layer of turns. If multiple layers are needed, use the progressive technique just described, along with this technique. If a single layer is adequate, try to separate the turns slightly to lower the turn-to-turn capacitance. Do not allow the two ends to come too close together; try to keep them at least 20 degrees apart.

Another way to raise SRF is to use heavier insulation. This forces higher spacing from turn to turn and helps raise the copper farther off the core and decrease the turn-to-turn capacity (Fig. 8.5).

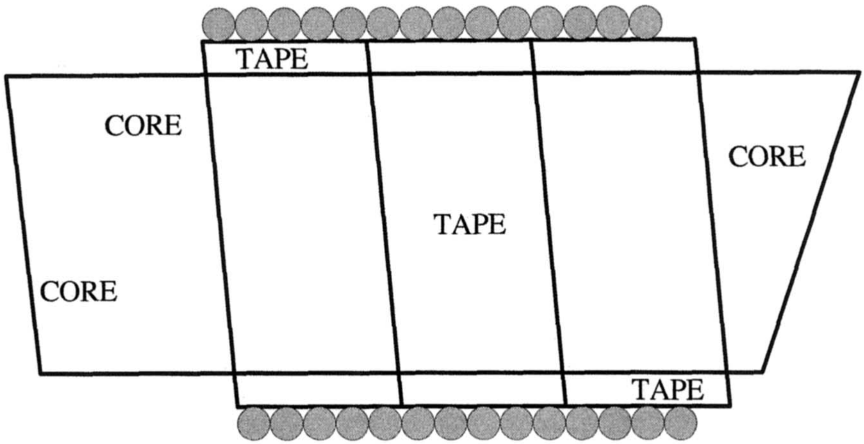

FIG. 8.4 Taped core for better SRF.

8.3. INDUCTOR DESIGN

The design of the inductor has been covered for the most part in Chapter 3 (see Sec. 3.2). Another common way is to design with A_l values. These are listed as millihenrys per 1000 turns and sometimes as microhenrys per 100-turns. Nanohenrys per turn is also used but rarely for the type of filters designed in this book. This technique can be used for all inductors including common mode as long as the A_l value is listed by manufacturer or known by some other means. This was developed from the following:

$$L = \frac{0.4\pi\mu N^2 A_c \times 10^{-8}}{M_{pl}} \tag{8.2}$$

In this equation, the needed inductance, L, is known along with 0.4, π, A_c, M_{pl}, and the permeability for the core that the designer would like to use. What is not known is N or N squared. These known core quantities make up A_l.

$$A_l = \frac{0.4\pi\mu A_c \times 10^{-8}}{M_{pl}} \tag{8.3}$$

The manufacturer provides the A_l values.

$$L = A_l N^2 \tag{8.4}$$

The needed inductance is known and the turns need to be found. The A_l value is given, so

$$N_2 = N_1\sqrt{\frac{L_1}{A_L}} \tag{8.5}$$

where N_2 is the needed turns, N_1 is the known turns (1000), L_1 is the needed inductance in the same units as A_l (millihenrys), and, again, A_l is the known millihenrys associated with the turns given (1000).

For example, the Magnetics, Inc. 55351 MPP core is listed as having an A_l value of 51 mH per 1000 turns. The designer needs 0.8 mH (800 μH). L_1 is equal to 0.8 and A_l is equal to 51. Then 0.8 divided by 51 is equal to 0.01569 and the square root of this is 0.1252. This value times 1000 equals 125 turns.

$$N_2 = 1000\sqrt{\frac{0.8}{51}} = 125.245 \tag{8.6}$$

The current is multiplied by the circular mils to determine whether the window area is adequate, that is, to see if the turns will fit in the core window area. This is a small core but we need 1 A, and using 600 circular mils per ampere would lead to 23 AWG at 650 circular mils per ampere. The 650 times 125, the number of turns just calculated, equals 81,250 circular mils, but the winding

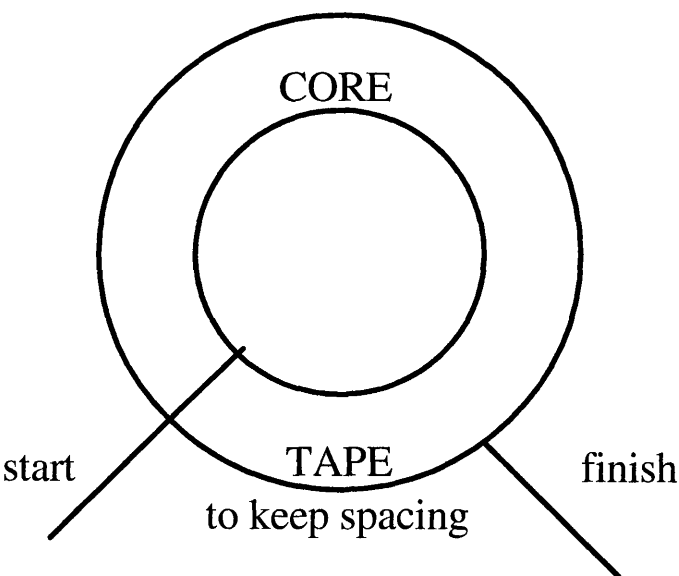

FIG. 8.5 Core showing the start-to-finish spacing.

factor for this toroid-type core is only 0.4. The 81,250 divided by 0.4 yields 203,125 circular mils.

$$\frac{650 \times 125}{0.4} = 203{,}125$$

Magnetics, Inc. lists the window area as 293,800 circular mils, so the windings will fit. The wire size could be changed to 22 AWG, giving 253,125, and still fit. This would reduce the copper losses and lower the temperature rise.

I recommend Col. W. T. McLyman's programs for selecting the proper inductor core and turns. The heat rise is a common filter problem and the filter design should be rather conservative, and McLyman's programs meet these requirements.

8.4. CONVERTING FROM BALANCED TO UNBALANCED OR THE REVERSE

Conversion from balanced to unbalanced, or the reverse, is required because the equations are all based on unbalanced circuits. It is better to balance the filter if the supply or equipment has not already been grounded. This happened to a customer who specified the complete filter design. We tried to tell the customer that the bottom half of his filter was not working, but (alas poor customer, I knew you well!) it was never changed because it went through qual that way and would have cost almost $100,000.00 to change it.

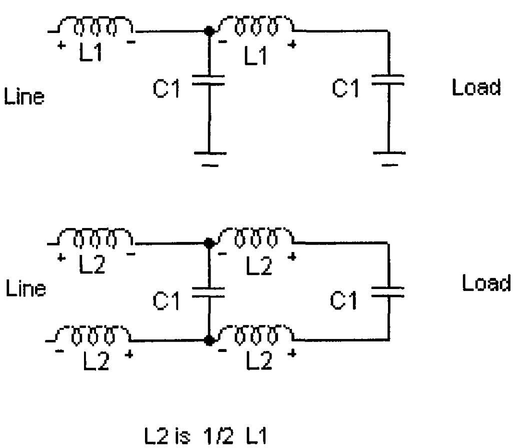

FIG. 8.6 Multiple L network converting from unbalanced to balanced.

To balance the filter, as in the 800 μH discussed in the last section, use a double L with some value of capacitor—say 1 μF—with a limit of 0.1 μF to ground. The picture would be 800 μH series, 1 μF Y capacitor shunt, followed by two just like the first two, then finish with the two feed-throughs on both lines. Simply divide both of the 800-μH inductors in half so that the circuit ends up with 400 μH on both the hot and neutral with the *X* capacitors of the same values across the lines. The inductors are smaller and the SRF should be higher. Mount the four inductors in quadrature (the two inductors that would be farthest apart would have to be mounted in the same way, hoping that the extra distance would reduce the coupling), and keep the capacitor leads as short as possible. Figure 8.6 shows the before and after drawings.

9

Common Mode Components

The causes of spikes or electromagnetic pulses (EMPs) on the line are lightning, large inductive equipment shutting off, and magnetic pulses created by nuclear activity. Lightning and nuclear activity create common mode pulses between the lines and ground, whereas the equipment type creates a differential mode type of pulse between the lines. On the load side, the switcher parasitic and the diodes are the leading culprits generating noise. Any current pulse, on the load side of the off-line regulator, to ground appears as common mode to the line. Adding a transformer or keeping the power supply isolated reduces these common mode conducted emission pulses. The one exception is the primary to secondary capacitance of the transformer, but this is so small that it can be neglected. Power supply filter capacitors pumping current to ground should be eliminated. To eliminate the common mode, the EMI filter employs common mode chokes and feed-through capacitors to ground. The problem is another specification limiting the leakage current. This is the capacitive reactance current flow through the capacitor between line and ground. This is another misnomer; it is really reactive current.

9.1. CAPACITOR TO GROUND

For 400 Hz, the limit of the capacitor to ground is 0.02 μF for Military Standard 461. Worse yet, the medical leakage current specification is harder to meet. If the device touches a patient, the total system leakage is limited to 100 μA. This means that most of the power supply people want the filter restricted to 20 to 40 μA. It is difficult to have the common mode losses meet the common mode loss specifications with capacitors to ground this small. A transformer would help. See Chapter 10. There are two schools of thought on this because the specifications governing this, as usual, are not clearly written. The first group thinks that this is the total capacitor to ground. In a three-phase, four-wire circuits, the capacitor limit value for 400 Hz is 0.02 μF. This capacitor value would then be shared, or

$$\frac{0.02}{4} = 0.005\ \mu\text{F}$$

This means 0.005 μF for each of the four legs to ground. The second group thinks that this is the maximum per line whatever the number of lines. If the system is well balanced, the current on each leg would nearly cancel through these capacitors at the ground point. (See the next section on virtual ground.) I hold with the latter group because it makes my job as a filter designer easier. However, ground fault equipment would not allow any capacitance to ground that would produce a current above the current threshold of the ground fault device, or it would have to be in the circuit after the filter. Again, a transformer would circumvent this.

9.2. VIRTUAL GROUND

In a two-phase system, where the two lines are 180 degrees out of phase, a virtual ground can be employed. This is the common method for power. Either leg to the common ground gives 120 V RMS and the two outer legs 230 V line to line and is really two phase. If the two line voltages are nearly equal and opposite, then two capacitors of nearly equal value from each line to ground have nearly equal, but opposite, currents. Whatever current flows in one capacitor to ground, the other capacitor has that much current flowing from ground. The two currents almost cancel. The small ground current is due to the slight differences in the two voltages and the differences in the capacitor values.

The same is true for three-phase systems. If the three RMS voltages are nearly equal and the capacitors to ground are nearly equal, the current to ground will be small. If the line voltages are the same and the capacitors are the same, the ground current is zero.

One way to help remove some of the difference currents is to employ the virtual ground technique. Tie the junctions of the capacitors together to form a virtual ground. Under ideal situations, the junction voltage to ground would be zero. Tie a capacitor of equal value from the junction to ground. Ground current will flow through the added capacitor based on the junction voltage. This technique further reduces the current on the ground lead.

Why are capacitors to ground necessary? Common mode reduction most likely requires them even with a transformer. The three capacitors in two-phase systems and the four in three-phase systems have a reasonably high impedance to ground at 50, 60, and 400 Hz, but what about at 14 kHz and above? The two, or three, capacitors from the lines to the junction are in parallel and are in series with the capacitor from the junction to ground.

9.3. Z FOR ZORRO

Ferrite toroid cores are often used for common mode inductors because they have the very high A_l values required for these common mode filters. For common mode testing, all lines are tied together in parallel and all the differential inductors are in parallel in the balanced design. Capacitors between the lines are of no value for common mode, but the capacitors from line to ground add in parallel. In the preceding three-phase, four-wire case, the total capacitor to ground (my method) would be

$$0.02 \times 4 = 0.08\ \mu\text{F}$$

To get the needed common mode loss, a common mode choke usually must be added. The total inductance and the total capacitance to ground normally do not give the required loss. This problem is often solved by placing a well-grounded barrier or shield across the filter center (Fig. 9.1). Split the value of the feed-through capacitor limit to ground by two. Two of these smaller feed-through capacitors are then installed in this shield. For best results with this method, put the Zorro inductor at the low-impedance end, or the line side, and the feed-through capacitors on the high-impedance end, or load side. In this case, where they are being split, the first Zorro is placed at the line and the first feed-throughs

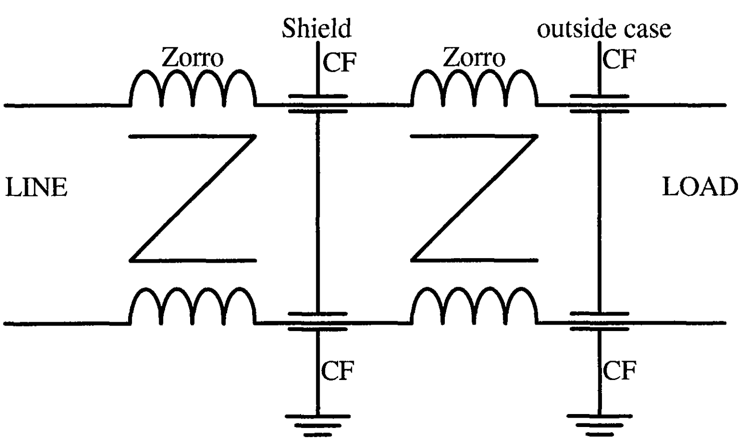

FIG. 9.1 The double-Zorro common mode filter.

are the last in the section—mounted in the shield. The next section starts with the Zorro and ends with the feed-throughs at the end. Try to use an even number of differential mode filters so that they can be split evenly in the two cavities. Say two L filters are required so that the first cavity would start with the Zorro, followed by the first L, then followed by the feed-through capacitors. The second section would be the same.

Of course, the line-to-line capacitors are not shown because they contribute nothing to common mode. However, they would follow the differential mode inductors. Put a second common mode inductor in the filter. This puts one common mode inductor, Zorro, in the front half and another one in the second half. This technique forms a double L in the common mode filter section, reducing the common mode inductor greatly in size, and now the specification can be reached most of the time. Remember that the inductor's magnetic fields buck or cancel for the differential mode and they have a high magnetic gain for the common mode. Figure 9.2 shows a three-phase delta with spacing between the windings.

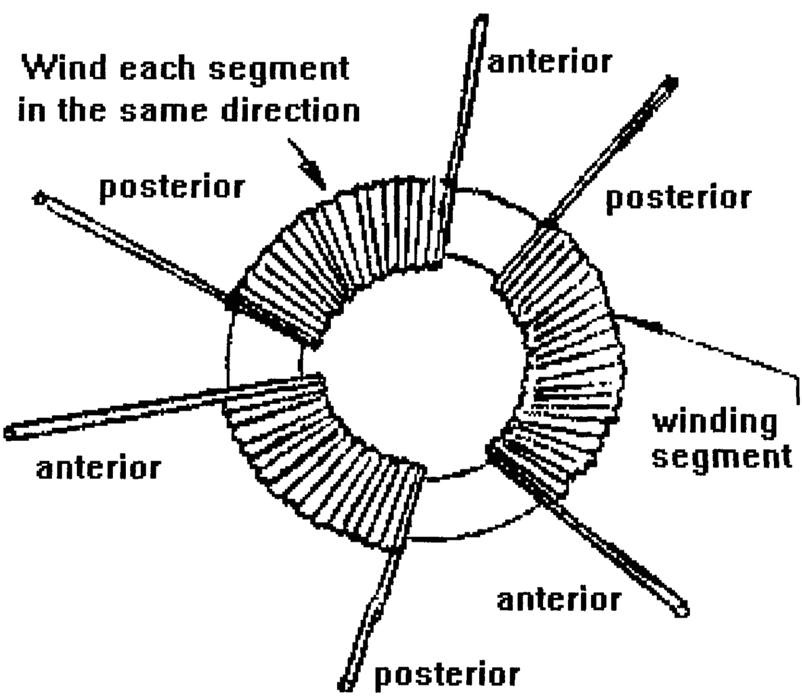

FIG. 9.2 The RFI Corp. three-phase, three-wire common mode inductor. Note: (1) Three windings on one core wound as shown. (2) Wire to be tightly wound and evenly distributed over segment.

For the three-phase wye filter, another set of windings is required, demanding a larger core for the same form. These are excellent for creepage, but the leakage inductance is greater, giving a differential inductance. These separate windings create 0.5 to 1% leakage or differential inductance. Figure 9.3 shows them wound together, reducing the leakage inductance, but this could be subject to creepage and corona problems.

Some designers have gone from the separate core windings to the quadfilar type to eliminate ringing. Leakage inductance and the stray capacitance in the inductor and other wiring caused this.

9.4. CONVERTING COMMON MODE TO A DIFFERENTIAL MODE FILTER

The common mode inductor has one value assigned to it. The inductance value is written above the Z for Zorro, say 10 MH, which is a typical value. Either winding should read the indicated inductance if the measurement is made with a good inductance bridge. The reason comes from the inductance formula:

$$L = \frac{0.4\pi\mu N^2 A_c \times 10^{-8}}{M_{pl}} \tag{9.1}$$

If an inductance bridge is used to read each winding is this example, the reading of both windings is 10 MH. If the inductance bridge is used to measure the two

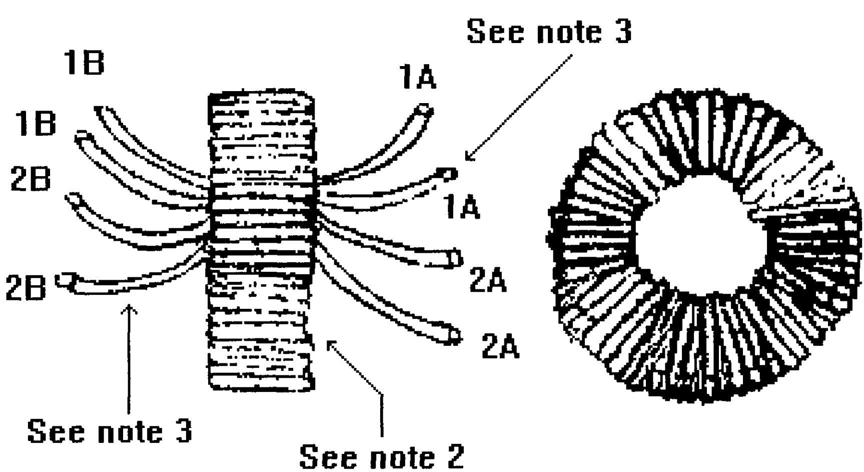

FIG. 9.3 The RFI Corp. tightly coupled—low leakage—common mode inductor. Note: (1) Quadfilar winding on one core as shown. (2) Wire to be tightly wound and evenly distributed. (3) Lead length to be 4 inches minimum.

windings, aiding, what would the aiding inductance be? From Eq. (9.)1, the turns are squared, so twice the turns gives four times the inductance, or 40 MH. However, the windings are split, half on each line, so each half is 20 MH. These two windings are now measured in parallel, so we are back to 10 MH! In other words, either winding would measure X and both measured in parallel—aiding—would measure X. If this was measured with the windings opposing, it would read the leakage inductance.

A common mode inductor using a ferrite toroid core can be designed using the A_l value of the core. This would be used, though, only for bifilar types requiring very low leakage inductance. The only difference between designing the differential mode and the common mode is that the winding or window fill factor is no longer 0.4 but now is reduced to 0.2 to fit the two windings. Divide the number of leads into 0.4 to get the winding factor.

The single-layer ferrite toroid winding can also be found when the core I_d and the wire size (American wire gauge, AWG) needed to handle the current are known. Get the wire diameter, W_d, in the same units as the I_d of the core. The total turns, N_t, is equal to

$$N_t = \frac{\pi(I_d - W_d)}{W_d} \tag{9.2}$$

If the wire diameter is much smaller than the core diameter (and it always is except for the type discussed in Chapter 19), this approaches

$$N_t = \pi\frac{I_d}{W_d} \tag{9.3}$$

Divide the turns by two, or whatever the number of wires is, and use the integer value to solve for the inductance. Knowing the A_l value of the core, the A_l number of turns (1000 typically), and the turns that the core can support, the inductance value can be calculated. This would be in a single layer wound less than halfway around. If this is not greater than the needed inductance, pick another core, usually the next bigger, with a larger I_d. Once a core is found for which the inductance is larger than needed, resolve the number of turns required by using the normal A_l equation. Find the number of turns necessary, and use the higher integer.

For example, 8 MH is needed with a current of 2 A peak. The AWG is No. 18, picked for the current specified. The diameter of the wire in inches is 0.0429. Now, pick the core. Here, 42915-TC is selected from Magnetics, Inc. The I_d of this core in inches (same units) is 1.142

$$N_t = \pi\frac{1.142}{0.0429} = 83.629$$

Divide this figure (83.629) by 2, obtaining 41.81, and use the integer. Here the integer is 41 turns. The A_l value of this core is 3868 MH per 1000 turns using F material. Here, the backward formula for A_l is used for 41 turns.

$$L = \frac{A_l 41^2}{1000^2} = 6.502 \text{ MH}$$

This is a little low, so a larger core or a material with a higher A_l must be selected.

The diameter of the wire is still 0.0429 inch. Pick a new larger core, here 43615-TC from Magnetics, Inc. The I_d of this core, in inches (same units), is 1.417.

$$N_t = \frac{\pi \times 1.417}{0.0429} = 103.768$$

Again, divide this by 2, obtaining 51.88 turns, and use the integer.

$$L = \frac{4040 \times 51^2}{1000^2} = 10.508 \text{ MH}$$

The integer is 51 turns. The A_l value is 4040 using F material as before. This is 2.5 MH more than our goal of 8 MH. Solve for the turns needed with the normal A_l formula.

$$N_t = 1000\sqrt{\frac{8}{4040}} = 44.499$$

Use the upper integer or 45 turns for each half of the windings. Keep the end gaps between the two halves as far apart as possible. This creates a visible winding gap on the core and makes this gap as large as can be. The difference between 51.88 and 45 gives this gap spacing between both ends of the two windings; 6.88 times the wire diameter of 0.0429 inch gives approximately this spacing. This is on an arc and the whole turn will not touch, eating some of this circumference, so the value will be less than the 0.295 inch calculated. Again, I would tape, or fill, these spaces so that both gaps between the two windings are secure (Fig. 9.4). Now the next question: will it fit the required box? Add 2.2 times the wire diameter to the O_D to get the outside diameter, which may not fit the box. Try two smaller cores stacked together; the A_l value doubles.

The problem is, now that the method of obtaining the proper core size is known, how was the value of 10 MH determined for the Zorro? Two things must be resolved. The first is how to convert from the common mode to the differential mode—really, from balanced (as the common mode is) to unbalanced, often called normal mode. If several balanced normal mode networks follow the

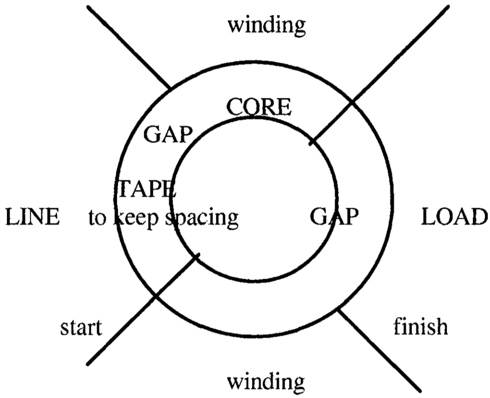

FIG. 9.4 Single-layer common mode inductor showing the spacing between windings.

common mode inductor, this must be converted back to unbalanced differential mode. All this is done in order to ease the calculations.

In Fig. 9.5, the *Z* is first followed by two L filters in turn, followed by feed-through capacitors. The common mode has been discussed and converts to a single inductor of 10 MH value. The following two inductors are in parallel with a value of 0.5*L* each, giving an effective value of 0.25*L*. The following capacitor, C1, is not in the circuit of the common mode, so the capacitor equals zero. The following two inductors give the same result as the last, 0.25*L*. C1 again equals zero, leaving the final component of C2, which is doubled in value. The results are as shown in Fig. 9.6.

The total for the two differential mode coils is 0.5*L* but often ends up as part of the overall headroom for the common mode part of the filter. The reason

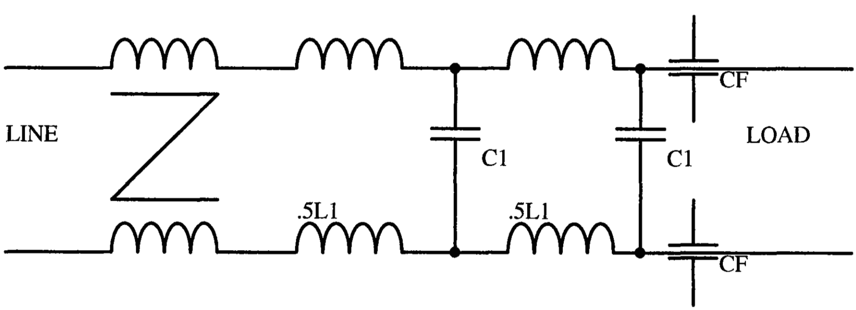

FIG. 9.5 The common mode and differential mode filter.

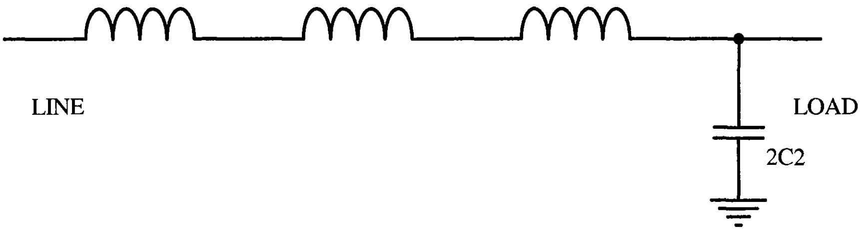

FIG. 9.6 The common mode converted to the unbalanced filter of Fig. 9.5.

is that the total value of these two coils may be around 400 μH, which is an order of magnitude lower than for the Zorro inductor. C2 may be limited in size by the specification, such as MIL STD 461, where the maximum for 400 Hz is 0.02 μF but now totals 0.04 μF. This is now in a form where the actual value of the common mode inductor can be solved using techniques in Chapters 16 and 17 (Fig. 9.7).

The second point to discuss is the test setup normally used to test the common mode inductor (Fig. 9.8). The two inputs are shorted together. The input is fed from a generator with 50 ohms output impedance. The two filter outputs, also shorted together, feed the load. The load impedance is also 50 ohms. This is the input impedance of the spectrum analyzer. This was another reason to convert from the balanced to the unbalanced filter as before.

This all works out to the circuit in Fig. 9.9, and the equations are in the next section.

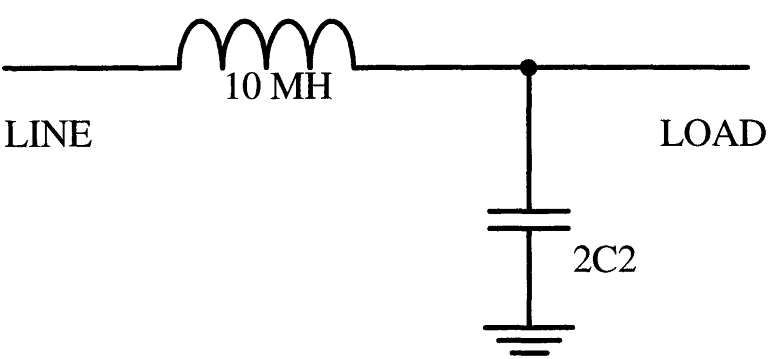

FIG. 9.7 Final conversion of the common mode to the unbalanced filter of Fig. 9.6.

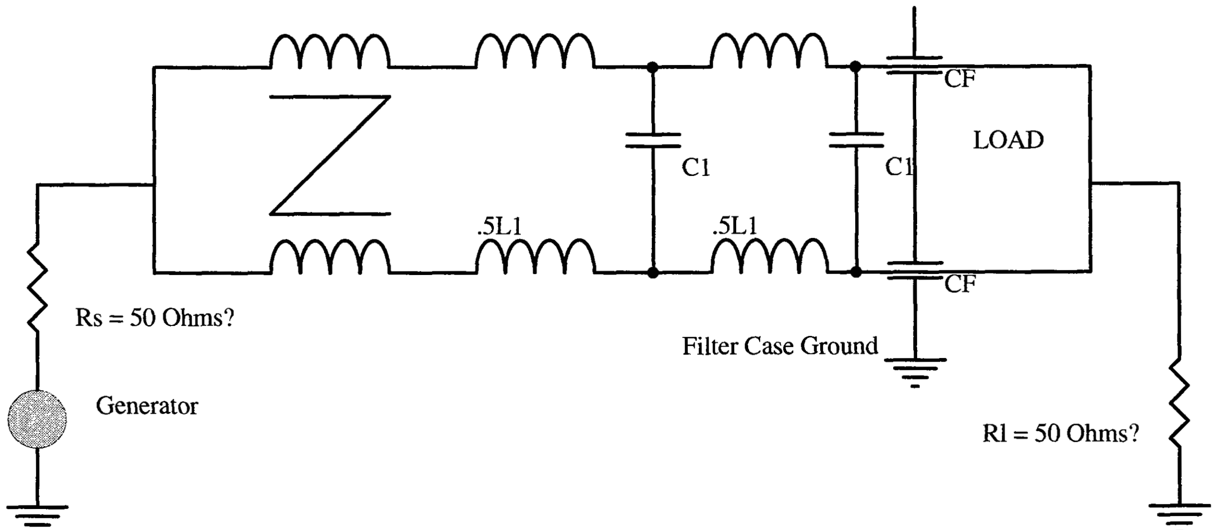

FIG. 9.8 The 220A test method for common mode section of the filter.

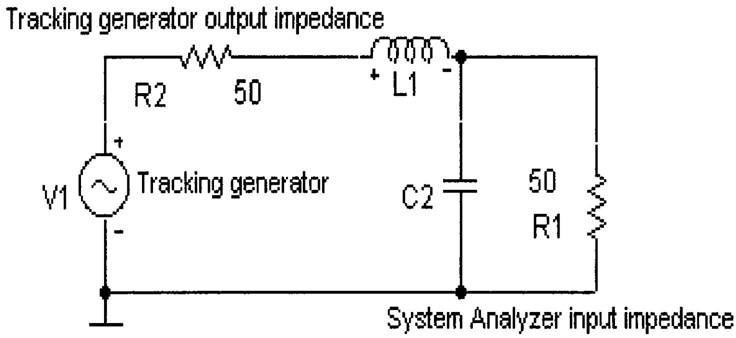

FIG. 9.9 The converted test method to calculate the common mode values.

9.5. EQUATIONS FOR THE COMMON MODE VIA THE DIFFERENTIAL MODE

The method used to arrive at Fig. 9.9 can be converted to the circuit here, where C is equal to two times C2, L is the common mode inductor needed for the proper insertion loss, R_s is the source impedance, and R_l is the load impedance. Because the equation includes R, L, and C, equations based on Q (charge) rather than I (current) of the two networks are the easiest to work with, and this generates the matrix in Eq. (9.4). This is from impedance matrix equations.

$$\begin{pmatrix} V_{in} \\ 0 \end{pmatrix} = \begin{bmatrix} LS^2 + R_sS + \dfrac{1}{C} & -\dfrac{1}{C} \\ -\dfrac{1}{C} & R_lS + \dfrac{1}{C} \end{bmatrix} \begin{pmatrix} Q_1 \\ Q_2 \end{pmatrix} \tag{9.4}$$

The delta, determinant, of the matrix is next:

$$\Delta = \frac{LCR_lS}{C} \left| S^2 + \frac{(R_sR_lC + L)S}{LCR_l} + \frac{(R_s + R_l)}{LCR_l} \right| \tag{9.5}$$

Substitute in the initial requirements of V_{in} and 0 and solve for Q_2:

$$\begin{vmatrix} \left(LS^2 + R_sS + \dfrac{1}{C}\right) & V_{ins} \\ -\dfrac{1}{C} & 0 \end{vmatrix} \tag{9.6}$$

and solve for

$$Q_{2s} = \frac{V_{ins}}{C\Delta} = \frac{V_{ins}}{LCR_l S\left|S^2 + \frac{(R_s R_l C + L)}{LCR_l} + \frac{(R_s + R_l)}{LCR_l}\right|} \tag{9.7}$$

but Q_2 is not the goal, V_0 is. In reality, the goal is the ratio

$$\frac{V_{0s}}{V_{ins}}$$

$$Q_{2s}S = I_{2s} \qquad I_{2s}R_l = V_{0s} \tag{9.8}$$

$$\frac{V_{os}}{V_{ins}} = \frac{1}{LC\left|S^2 + \frac{(R_s R_l C + L)}{LCR_l} + \frac{(R_s + R_l)}{LCR_l}\right|}$$

This equation has been published in many articles, but most often they do not include R_s, the source impedance. Most do not use the two feed-through capacitors in parallel, which are now doubled in value.

Most solve this by completing the square. This means, in the case of Fig. 9.5,

$$\left|\frac{(R_s R_L C + L)}{2LCR_l}\right|^2 \tag{9.9}$$

must be added to complete the square of the first two terms in the main denominator and subtracted from the last term. This term is always much greater in the common mode application than the last term in the main denominator, making the new last term, for omega squared, negative. This makes the solution a hyperbolic function and very lossy, as suggested by the test setup. Most leave out R_s, making the value of a reduce to $1/(2RC)$, the dampening factor. This also reduces the last term of the denominator to $1/(LC) - \omega^2$.

$$a = \left|\frac{(R_s R_L C + L)}{2LCR_l}\right|$$

$$\omega = -\sqrt{\left|\frac{(R_s R_L C + L)}{2LCR_l}\right|^2 - \frac{(R_s + R_l)}{LCR_l}} \tag{9.10}$$

giving the following answer where a is the dampening factor.

$$\frac{1}{LC}\left|e^{-at}\sinh(\omega t)\right| \tag{9.11}$$

This may now look easy, but there is another way, knowing that the common mode will always be lossy: forget the sin or sinh solution and set the main denominator of Eq. 9.8 to

$$(S + a)(S + b) \tag{9.12}$$

So that $a+b$ is equal to the middle term and ab is equal to the end term of the main denominator of Eq. (9.8) repeated here.

$$\frac{V_{os}}{V_{ins}} = \frac{1}{LC\left|S^2 + \dfrac{(R_sR_lC + L)S}{LCR_l} + \dfrac{(R_s + R_l)}{LCR_l}\right|} \tag{9.13}$$

This is a simple solution, and both a and b are included within the same quadratic. Either a or b can be assigned the positive square root, but the solution is better with b being the more positive

$$a = \frac{(R_sR_lC + L) - \sqrt{(R_sR_lC + L)^2 - 4LCR_l(R_s + R_L)}}{2LCR_l}$$

$$b = \frac{(R_sR_lC + L) + \sqrt{(R_sR_lC + L)^2 - 4LCR_l(R_s + R_L)}}{2LCR_l} \tag{9.14}$$

and takes the form

$$\frac{e^{-at} - e^{-bt}}{(b - a)} \tag{9.15}$$

The term $b - a$ reduces to

$$\frac{\sqrt{(R_sR_lC + L)^2 - 4LCR_l(R_s + R_L)}}{LCR_l} \tag{9.16}$$

This cancels LC, so the final answer is

$$\frac{R_l\left|e^{-at} - e^{-bt}\right|}{\sqrt{(R_sR_lC + L)^2 - 4LCR_l(R_s + R_L)}} \tag{9.17}$$

In the normal test arrangement, R_s and R_l are both 50 ohms and C, because of leakage current, is whatever the specification requires. Remember, the value of the capacitor is doubled here because the two feed-throughs are in parallel and, therefore, add. Now that the common mode is reduced to a simple single L filter, the required value of the common mode inductor can be easily solved. See Chapters 17 and 18.

9.6. COMMON MODE INDUCTOR USED FOR DIFFERENTIAL MODE

Some filter manufacturers have designed common mode inductors that also function partially as differential mode inductors. This is done as before with very wide winding spacing that generates the leakage inductance needed to provide this differential inductance. Some flux of one coil fails to cut the other creating this leakage inductance. Another way this is accomplished is with pot cores (Fig. 9.10). They use a split bobbin so that some flux in one half of the bobbin fails to cut the other half. This has been expanded to finding two separate bobbins that fit in the core with additional room to place a washer between the two bobbins. This washer is cut, or split, in order to avoid the washer acting as a shorted turn. The material of the washer has little to do with the differential inductance created. It is the separation of the two windings that causes the leakage inductance.

The leakage inductance is easy to measure with an inductance bridge. Shunt one winding of the common mode inductor, and read the inductance of the other winding. If all the flux lines cross or cut the other turns, the reading is zero. This

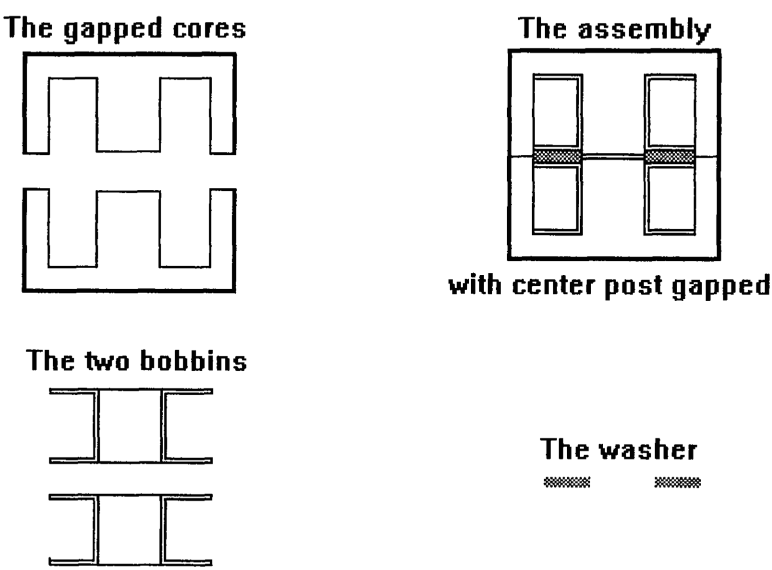

FIG. 9.10 Gapped pot core and assembly.

is truly impossible to accomplish because the turns cannot all be so tightly coupled. The difference is the leakage inductance measured by the inductance bridge. Another way is to measure both legs together, opposing, and the inductance bridge reads the leakage inductance. Some state that the leakage inductance is in the air, not the core, so it cannot saturate. This is true to some degree, but it saturates a minor amount because not all is in the air.

9.7. OTHER WAVE SHAPES BESIDES SINE WAVES WORK

Some engineers to whom I have spoken think that the common mode inductor requires a balanced sine wave for the common mode to work properly. In the single-phase common mode inductor, the sum of the magnetic fields caused by the currents will still cancel no matter what the wave shape. Because the current is equal and opposite, these two flux fields still cancel. If the currents are not equal—for example, a current difference created by the capacitors to ground—the two fields almost cancel and give some differential mode inductance. This is the problem in speaking to purists: most of them claim that the feed-through caps unbalance the common mode because they are not matched. All of this is true, but to what degree? This is also true of any arresters from line to ground because they have different values of capacitance across them and leakage currents through them.

This statement is very true in three-phase circuitry if the phases are exactly 120 degrees apart. Assuming this is not happening, the wave shape still need not be sinusoidal. Any current of any shape gives a magnetic field that cancels in the single-phase, three-phase, or DC system. If the reverse was true, common mode would not work for three phases. Because of the imbalance of the phases and the harmonic content, the voltage is not very sinusoidal; ergo, the common mode will not work. The three-phase Zorro requires three windings for the delta and four wires for the added neutral in the wye. If the three legs are not balanced, the neutral carries the difference current and magnetic flux will still cancel because of the fourth neutral winding.

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10

The Transformer's Addition to the Filter

The transformer is overlooked as a filter element. This is because of the obvious disadvantages of weight, size, and cost. Using off-line regulators and switchers can often eliminate transformers. On the other hand, switchers often complicate EMI issues because of the high-frequency switching noise they generate. This is worse as the switcher frequency increases. Also, switchers, because of their present-day glut of components, defeat the MTBF requirements. The question for the system designer is, do the disadvantages of transformer overshadow the advantages?

10.1. TRANSFORMER ADVANTAGES

The main advantages of the transformer, without considering their EMI advantages, are isolation, voltage translation, common mode rejection, and the potential of low-leakage current. Another advantage that is overlooked is the ruggedness of this device. Transformers can handle voltage spikes without difficulty.

If the transformer is an autotransformer, isolation, common mode rejection, and leakage current advantages are done away with because of electrical cross-coupling of the secondary to the primary. This chapter assumes that a standard isolation transformer is used.

10.2. ISOLATION

Isolation is accomplished because the primary and secondary are coupled magnetically rather than physically or tied together electrically. Therefore, whatever the wiring arrangement of the secondary is, such as one output leg tied to chassis ground, it is not coupled electrically to the primary. The green safety chassis ground lead does not carry any current back to the service ground from the secondary even thought the secondary is tied to this chassis ground.

10.3. LEAKAGE CURRENT

There are various specifications regarding leakage current. In commercial applications, the leakage current specification is typically 5 mA for the system. The medical requirement is 300 μA (0.3 mA) for the system, but the transformer is usually specified as 100 μA. Powertronix, a transformer house in Foster City, California, can deliver a transformer with less than 50 μA. Leakage current through the transformer is, in theory, zero but in reality is a function of spacing and shielding between the primary and secondary. The leakage current reduction is also a function of a Faraday screen, if used. For transformers with bifilar winding, where the primary and secondary are wound together at the same time, the leakage current is drastically higher. The reason for all this is that the leakage current is a direct function of the primary-to-secondary capacitance. If the windings are bifilar, there is little spacing other than the wire insulation. Obviously, no Faraday screen or shield can be used, so the leakage current is due to the capacity from wire to wire and is high.

10.4. COMMON MODE

Common mode tracks directly with the leakage current of the transformer. Common mode—equal voltage on both lines of either the primary or the secondary—does not create a magnetic field across the primary or secondary of the transformer. The only coupling is through the capacitance from primary to secondary. Any reduction in the leakage current by reducing the capacitance also reduces the common mode.

10.5. VOLTAGE TRANSLATION—STEP UP OR DOWN

Step-up and step-down requirements were the main reason for their development years ago and so are not discussed here.

10.6. THE TRANSFORMER AS PART OF AN EMI PACKAGE

The problem with EMI filter design is opposite stated requirements. One of these is a heavy common mode requirement specified along with an impossible leakage current requirement. These two items conflict because, without a transformer, the filter requires sizable capacitors to ground on both the hot and the return wires and a common mode inductor to remove the common mode noise, but the leakage current requires little or no capacity to ground. A quality transformer can come to the rescue. The transformer gives the filter designer common mode loss and also large differential mode loss.

If the EMI filter is first in the power stream, care must still be taken in the capacitors-to-ground values for leakage requirements. However, the transformer should have removed most of the common mode noise anyway, easing the filter part of the noise that it must reduce. So small capacitors and common mode inductors should be sufficient to do the job. If the EMI filter follows the transformer, any reasonable capacitor values could be used because the transformer will eliminate this ground current from the primary side because this is common mode.

Is this all the transformer would do for the filter designer? No. Power transformers using laminations, C cores, and tape-wound toroids with steel thicknesses of 12, 11, 7, and 4 mils exhibit very high wattage losses in watts per pound at higher frequencies. This enhances the differential mode section of the transformer. These graphs are hard to read, but National Arnold 12 Mil Selectron C Core lists the following approximate loss for 900 gauss. Then the loss per octave and decade can be calculated.

Hertz	Watts per pound
1,000	0.27
2,000	1.1
10,000	24
20,000	92

$$10 \log\frac{0.27}{1.1} = 6.1$$

$$10 \log\frac{24}{93} = 5.8$$

$$10 \log\frac{0.27}{24} = 19.5$$

$$10 \log\frac{1.1}{92} = 19.22$$

Here, the weight in pounds cancels, leaving watts divided by watts. The first two are for the octave (frequency doubling) loss using the data from the table. Both are near 6 dB per octave, and the next two are for the decade (10 times the frequency) loss, which gives 20 dB per decade. This loss, or cutoff frequency, starts near the third harmonic for the steel type. Twelve-mil steel is proper for 60

Hz, and this octave, or decade, loss should start by 180 Hz. This means that the tester should expect to see something close to 6 dB by 360 Hz and close to 20 dB by 1800 Hz. However, the cutoff frequency would vary from transformer to transformer. This differential mode loss is dissipated, not attenuated. Another way to evaluate this is through the core manufacture estimated loss equations. These have the form

$$\text{Watt/lb} = CF^A B^E$$

where C is a constant (possibly to a power), F is the frequency at some power A, and B is the flux density at some power E. Because the weight (lb), flux density (B), and constant C of the core remain the same, they cancel. Therefore the standard dB equation can be used.

$$\frac{\text{Watt(1)/lb}}{\text{Watt(2)/lb}} = \frac{C}{C}\frac{F_1^A}{F_2^A}\frac{B^E}{B^E}$$

$$\frac{\text{Watt(1)}}{\text{Watt(2)}} = \frac{F_1^A}{F_2^A} = \left(\frac{F_1}{F_2}\right)^A$$

$$dB = 10 \times A \log\frac{F_1}{F_2}$$

Because the engineer is interested in the loss per octave or decade, the ratio of F is either 0.5 or 0.1 and this is to the A power. Any frequency ratio can be calculated, however. Armco 14-mil steel is listed as 1.68 for A. This yields 5.05 dB per octave and 16.8 dB for the decade. The problem is that with these equations it is difficult at best to get a near fit to the listed data, but this loss for Armco is in the ballpark. Nevertheless, this core loss adds greatly to the differential mode loss of the filter.

The last equation can be worked backward.

$$dB = 10 \times A \log\frac{F_1}{F_2} = 10A \log(F_R)$$

$$A = \frac{dB}{10 \log(F_R)}$$

where F_R is the frequency ratio, which is 0.5 for 6 dB. Back-solving for A, knowing dB is 6 and F_R is 0.5, gives 1.993. As an example, say a transformer has 6 dB of loss at 600 Hz. The designer needs 60 dB at 20 kHz. The frequency ratio 20,000/600 equals 33.33. The logarithm of 33.33 to the 1.993 power is 3.035, and this times 10 is 30.35 dB. Adding back the 6 gives 36 dB. The designer needs 60 dB at 20 kHz, so 24 dB more is needed. A double-L filter added to the system would do nicely. Four elements at 6 dB each give 24 dB. The cutoff frequency for the double L would, in theory, be 10 kHz. These would prove to be reasonably small components giving very high self-resonant frequencies (SRFs). Is there any other action within the transformer to aid in the loss?

10.7. SKIN EFFECT

The higher frequencies of either common or differential mode are also dissipated within the high-frequency resistance of the wires. However, this does not come into play until 30 kHz and above. The radius of conduction, in centimeters, is

$$R_C = \frac{6.62}{\sqrt{F}}$$

10.8. REVIEW

In other words, the isolation transformer adds the same as any other EMI filter element—6 dB per octave or 20 dB per decade. A further advantage is that the frequency cutoff point is so much lower than for the normal EMI filter components. Heavy common and differential mode loss is realized by using the isolation transformer. Again, disadvantages are the added weight, size, and possibly cost. If the transformer eliminates one or more filter sections, the increase in cost may be compensated by eliminating the costs of these components. For example, if a 12-mil (for 60 Hz) steel core gives only 6 dB at 600 Hz, an additional 20 dB by 6000 Hz, and another 6 dB by 12,000 Hz, the total loss is 32 dB at 12,000 Hz, and we can add a few for 14,000 Hz. This may be more than enough loss so that no other filter elements may be needed.

On the other hand, the transformer's effectiveness diminishes as the primary-to-secondary equivalent capacitance comes into play. The approximate primary-to-secondary equivalent capacitance is

$$C = \frac{1}{2E\pi F} = \frac{100 \times 10^{-6}}{2 \times 120 \times \pi \times 60} = 2210\,\text{pF}$$

where E is the input volts at 120, F is 60 Hz, and I is 100 μA. The maximum value would be 2210 pF, and 50 μA would be half that value. In a 220A test setup, the transformer would be effective to above 700 kHz. This is where the capacitor would be equal to the source plus load impedance, here 100 ohms. However, the transformer effectiveness would be much lower than this frequency because of the SRF caused by leakage inductance. But then all filter components suffer from this dilemma, not just the transformer.

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11

Electromagnetic Pulse and Voltage Transients

This chapter develops a ratio according to the ideas stated in the following. The circuit discussed here consists of a battery, a transmission line, and a switch. The end piece is different, either a short or an open. The battery has an output impedance of Z that equals the impedance of the transmission line. When the switch is closed, a step function travels down the line at the velocity of propagation. The voltage divides between the characteristic impedance of the line and the source impedance. The main interest is in what is at the far end of the line. If the line is open at the far end, the voltage doubles. The wave travels back, at the same velocity, to elevate each segment of the transmission line to full voltage. At the time the wave reaches the source, or battery, end, the current drops to zero and the full battery voltage is impressed across the transmission line (Fig. 11.1).

If a short is at the far end instead of an open, the current doubles when the wave reaches the shorted end. One half of this current depletes the initial stored voltage, segment by segment, while the other half continues to flow through the line from the battery. This discharges each segment. When the pulse reaches the battery end, the line is fully discharged and the double current flows from the battery, or source. The full voltage is dropped across the internal source impedance, Z. If the line and battery Zs are not equal, the pulses iterate back and forth until equality is reached. The main point here is to see the difference between the two far end conditions—one near open and one near short. In the link, or short line, the source impedance is thought to be very low and, again, the question is what is at the filter end of the line: a shorted condition or nearly open impedance condition. If the initial condition is high impedance for the load, the full strike

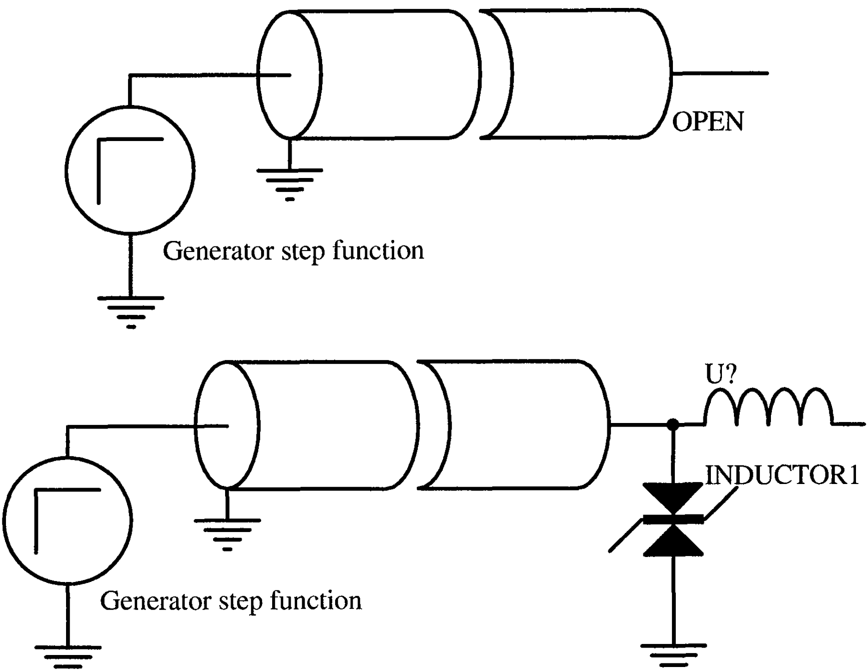

FIG. 11.1 Open-ended transmission line applied to the transzorb.

voltage is impressed across the suppressor, aiding turn-on or firing. The device then drops to low impedance and a high current is carried through the suppressor. The suppressor and the line impedance will dissipate the energy—we hope. If the energy is too great for the suppressor, the suppressor normally first shorts and then blows open.

The power in the pulse is dissipated in the MOV, line, and source impedance. If the condition is near short, the voltage divides, according to the ratio, and delays the firing of the protector (Fig. 11.2).

Some engineers place capacitors across the transzorb, or MOV, making the condition similar to the preceding shorted condition. Transzorbs and MOVs are much faster on turn on, these days, but the capacitor must charge to well past the turn-on voltage before the arrester can act. Another point is that the current through the capacitor is of the order of twice the initial line current or several times 100 A. All of this action slows, or delays, the turn-on of the transzorb. This can also blow the capacitor, especially if the capacitor is the metallized film type, if it was not rated for this amount of current regardless of its construction, and again if the capacitor voltage rating was exceeded. All of this can stress not only

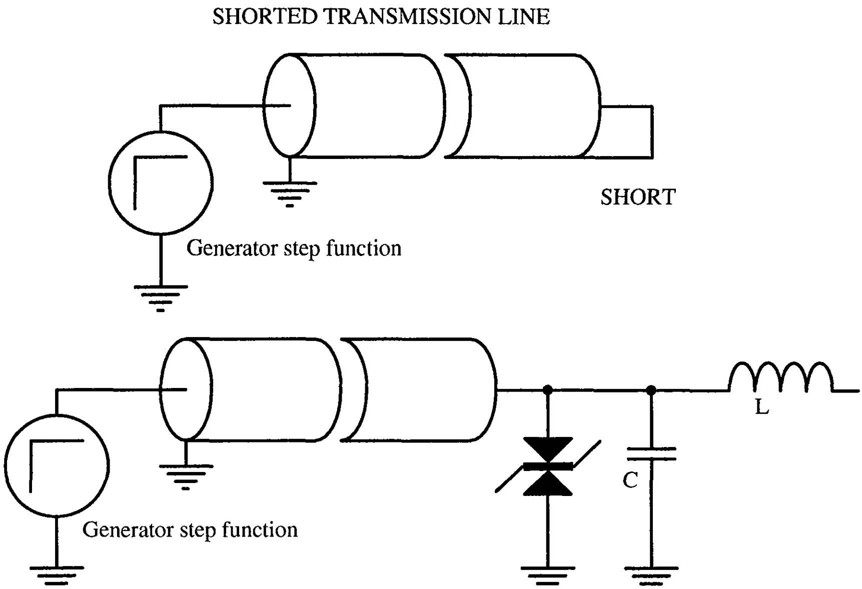

FIG. 11.2 Shorted transmission line applied to the transzorb.

the capacitor but also all the remaining filter components. These engineers argue that all they are trying to do is protect the arrester, but when a high-energy pulse occurs, both the capacitor and arrester can be blown. If the filter is designed to withstand this stress, it may well hold up, but this also adds to the cost of the filter. If the far end (filter input end) is open—transzorb directly across the line followed by an input series inductor—the voltage will rise quickly. The voltage will not necessarily double; the impedance ratio is not known. The voltage will help the arrester to fire, speeding up turn-on. The transzorb may blow, but the filter and the following equipment will be protected. Most of these arresters conduct in less than a microsecond under these conditions, so the peak current through the inductor and the voltage stored across the following filter capacitor are reduced.

11.1. THE THREE THEORIES

The first theory is from the purists. They say that the arrester is not necessary and it is the job of the filter to handle these pulses of energy. This might be true if the components were designed to handle them. In this case, the inductor must be designed to handle the full pulse voltage and pulse duration without arcing. The following capacitor must withstand twice the pulse voltage without blowing.

The inductor must be wound with spacing between the turns or with very heavy insulation. To calculate this, the total number of turns and the maximum spike voltage must be known.

$$\text{Maximum spike} = 10{,}000\text{V} = V_{\text{max}} \quad \text{and} \quad \text{Turns} = 50 = N \qquad (10.1)$$

The volts per turn is

$$\frac{V_{\text{max}}}{N} = \frac{10{,}000}{50} = 200$$

Most wire insulation withstands 500 V and should not require special insulation, and the turns, here, could touch each other. This is easy to measure. Take the length required for the inductor and strip one end. Bury this length of wire, with both ends exposed, in an aluminum basket of shot pellets or any other small conductive container. Apply the test potential to the stripped wire end and the other high-voltage lead to the shot container. What would happen if there were only 10 turns? The volts per turn is

$$\frac{10{,}000}{10} = 1000$$

This is marginal, so special insulation should be used and, as an extra margin of safety, the turns should not touch. Granted, the wire should withstand 1000 V from wire to wire, or turn to turn, but this is too risky. This spike inductor should also have only one layer of wire and be raised off the core to protect the windings from arcing to the core. Many are wound on forms without cores and wound in or on a solenoid form. In the 10-turn case, these techniques must be employed. In the 50-turn case, the same extra measures with respect to the core should be employed: use better wire insulation, keep the turns apart so that they cannot touch, and keep the turns as far as possible away from the core, if any (Fig. 11.3). Also, watch the tape. Problems have been caused by tape that had too low an isolation voltage. There is an additional theory about this. The inductor acts as a transmission line, and the full spike voltage is impressed across the first turn, one at a time. This charges the capacity to ground for each turn and then moves on to the next turn. Many inductors impressed with high voltage seem to burn out on the first few turns.

The second group wants to protect the equipment and not the filter, so the arresters are placed at the equipment end of the filter. They also argue that they do not want to exercise the arresters on every pulse that comes down the line. The questions asked are, “What if the pulse blows the filter? Can the equipment still operate?” These engineers answer the questions with a question, “If the transzorb blows, can the equipment still operate?” The answer to their question could be yes for two reasons.

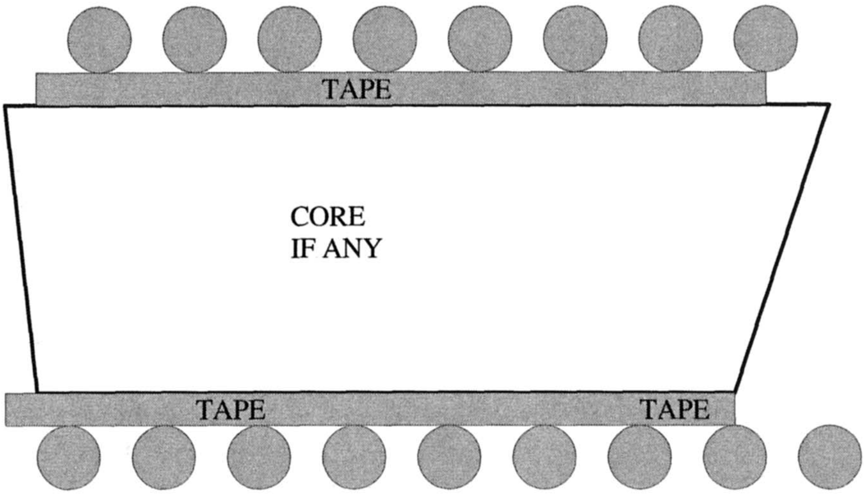

FIG. 11.3 The solenoid, or core with tape, and insulated wire with spacing.

1. The arresters often blow shorted; then, with the high current through them, they blow open. If the arrester is then open, the equipment can still function.
2. I like placing the arresters on the outside of the filter at the input. With proper access to the filter, these blown units, shorted or open, can be cut out and the equipment can operate until new arresters can be installed.

The third group locates the arresters at the front end of the filter to protect the entire equipment. The diatribe of the opening comments also pertains to this solution. I like the MOV at the front end with an inductive input to follow as part of the filter (Fig. 11.1).

When the pulse reaches the arrester and inductor, the pulse sees high impedance from the inductor; the voltage rises rapidly, firing the MOV. The initial line current continues through the arrester, and the pulse is dissipated in the MOV and the line impedance. The voltage that the filter sees is the MOV clamping voltage. Also, the following capacitor should withstand at least twice the MOV or transzorb clamping voltage.

There is a fourth group that is a combination of groups one and three. They want the input inductor split with the arrester tied to the junction of the inductors (Fig. 11.4). The first half limits the arrester current but must be able to withstand the pulse. This first inductor is often wound on a bobbin without a core to eliminate the potential for any arcing to the core.

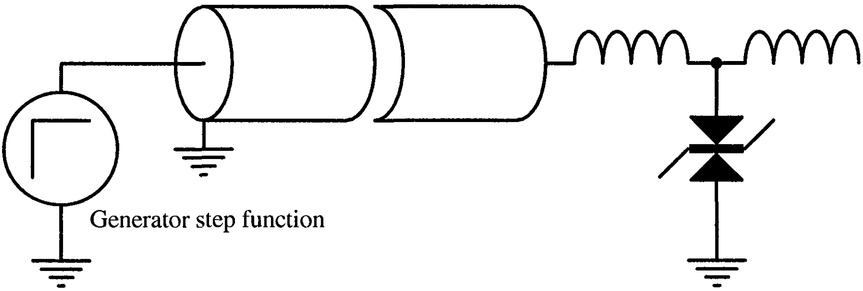

FIG.11.4 Split input inductor to limit the transzorb current.

There is a fifth group that uses two arresters, one on each end. This is for very high pulses on the order of 100 kV. Often the arrester is sandwiched between two input inductors and the last is across the output capacitor. The filter will take the remaining energy not handled by the initial MOV and spread it over time, lengthening the pulse width but reducing the peak energy. The last MOV, or whatever, will handle the remaining energy.

11.2. THE LOCATION OF THE ARRESTER

From the preceding section, it should be clear that I prefer the MOV at the head end of the filter with the filter having a series inductor at the filter input (Fig. 11.5). I also prefer them to be mounted outside the filter, where there should be access to change them if the need ever arises.

In the single-phase balanced circuit, three transzorbs are required: one from hot to ground, one from neutral to ground, and one from line to line. The first two protect the equipment from common mode pulses—pulses from both lines to ground. The last is any pulse between the lines. In the three-phase filter, where

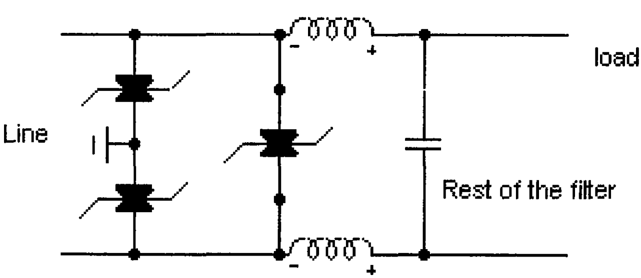

FIG.11.5 Single phase with arresters.

all three lines are treated within the one filter enclosure, six transzorbs are required. Three are wired from the three lines to ground, and three are wired from line to line. If the units have internal arresters, the entire filter must be replaced after an EMP or other pulse that blows the transzorbs. This can create an inventory problem requiring expensive filters in stock. In the other method, where the arresters are located outside, the blown arresters can be cut out. The equipment can be operated without protection in an emergency until the transzorbs can be replaced.

11.3. HOW TO CALCULATE THE ARRESTER

The specification of the shape and size of the strike is usually furnished by the end user of the device. The job of the filter designer is to calculate the joules needed. The style of transzorb is selected, and the clamp voltage for the peak current is found from the transzorb specification sheet. From this and the various shapes, the joules are figured out. Once the joules are known, check that the transzorb initially selected will handle this amount of joules. The method follows. The peak current is 50 A, and the circuit is three phase with 208 V line to line and 120 V line to ground. Six transzorbs are needed, with three from line to line and three from lines A, B, and C to ground. The latter three can be changed to fit the 120 V line, say 150 V, but I recommend keeping all the same because production or installation people do not know the difference. The radial lead is selected because of ease of installation. The voltage is specified for the RMS voltage, so 230 is the lowest we can use and 250 is the next. With the 230 V transzorb, the clamp voltage according to the catalog is 700 V at 50 A. Next is the shape of the wave, and each shape has a *K* factor. This shape factor, *K*, is needed next (Fig. 11.6).

The *K* factors for the pulses in Fig. 11.7 are as follows: ramp = 0.5, a constant height = 1.00, and a sine pulse = 0.637. The *K* factors based on the pulses in Fig. 11.7 are as follows: dampened sine wave = 0.86; dampened exponential = 1.4. If the drawing or specification has a ramp followed by the dampened exponential with the time of 5 μs for the full ramp time and a total of 50 μs to the 50% current point of the exponential (Fig. 11.8), the calculation is as follows.

$$\text{Energy of the ramped wave} = KV_c I_p \tau = 0.5 \times 700 \times 50 \times 10^{-6} = 0.087$$

$$\text{Energy of the dampened wave} = KV_c I_p \tau = 1.4 \times 700 \times 50 \times (50 - 5) \times 10^{-6} = 1.4 \times 700 \times 50 \times (50 - 5) \times 10^{-6} = 2.205$$

where the time is the difference. This gives

$$\text{Total} = 2.292$$

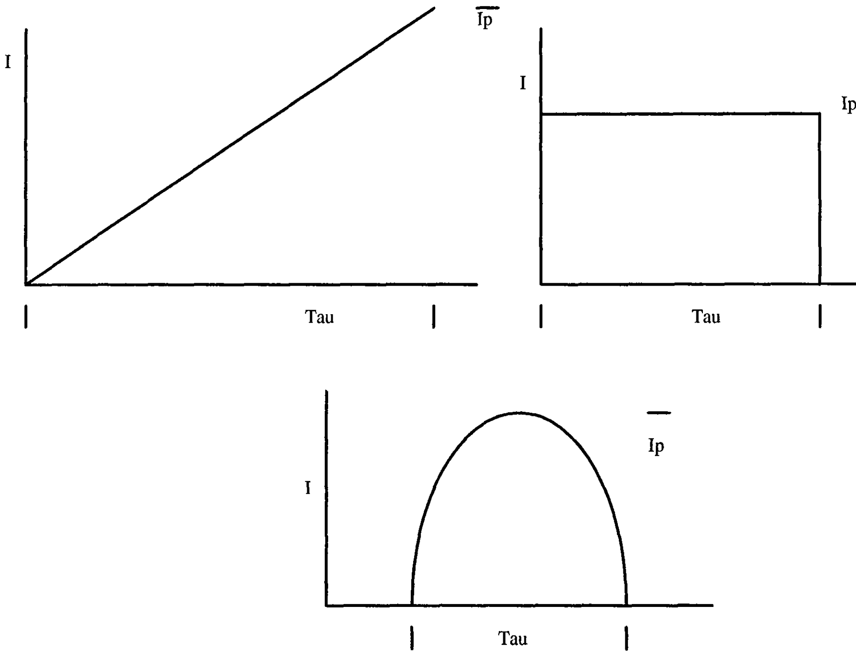

FIG. 11.6 The *K* factor calculated using the full time of the pulse.

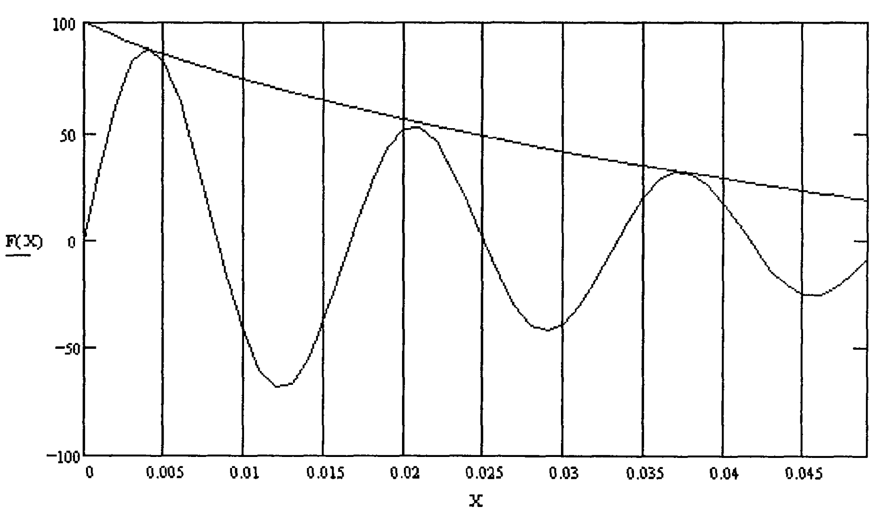

FIG. 11.7 Dampened sine wave based on the time to 50% of peak current.

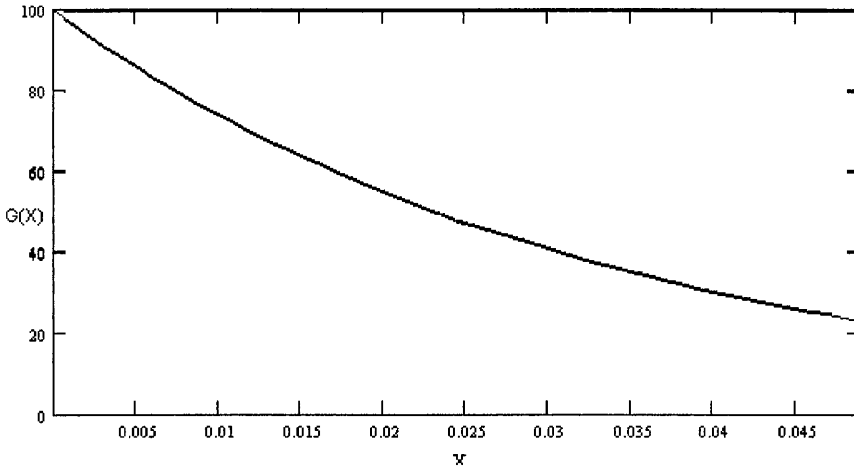

FIG. 11.8 Exponential based on the time to 50% of peak current.

Multiply this by 1.2 for a safety factor:

$$J = 2.2925 \times 1.2 = 2.751$$

and round this to 2.75 joules. The joules are listed as 20 W-s or J, giving good headroom.

This was calculated using a Harris V230LA4. The typical pulse calculation is shown in Fig. 11.9.

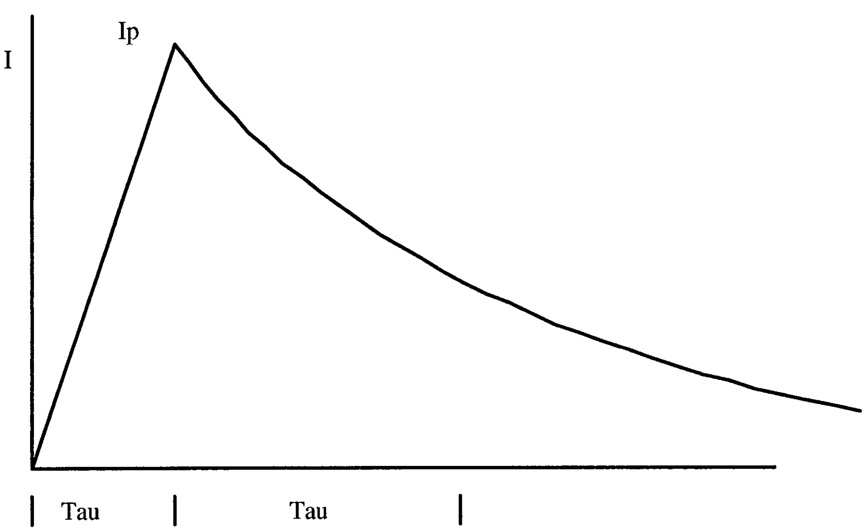

FIG. 11.9 Typical pulse calculation.

11.4. THE GAS TUBE

Besides transzorbs and MOVs, there are gas tubes. These handle 25,000 A over short periods of time. One of the leading companies is Joslyn Electronics, located in Goleta, California, near Santa Barbara. These are best used in DC systems, where the DC voltage will stabilize back to a constant voltage soon after the pulse. The tube will then deionize quickly. In AC systems, because of the continuous change sinusoidal voltage, the gas tube does not fully deionize. In some cases, the system must be shut down for a short time to rid the tube of the ionization.

12

What Will Compromise the Filter?

The filter may test well using the specified test method and still fail to work as designed for many reasons, the most prevalent of which are discussed here. As mentioned in an earlier section, some filters are compromised by ground faults. The case discussed was a balanced filter with the bottom half shunted around. This reduced the loss by 50%. It was caused by lack of communication between two different engineering groups, resulting in double the weight, size, and cost for half the performance. More candidates for failure are discussed here.

12.1. SPECIFICATIONS—TESTING

Filters are usually designed to pass the more prominent 220A test with 50 ohms load and source. An EMI filter was designed for the 220A specification but the new customer wanted to test using the current injection method described by Military Standard (MIL STD) 461. The filter failed. This test setup requires two 10-μF capacitors to ground on both the supply and return power leads. The filter was unbalanced with all the components on the supply, or hot, side. The test setup was detecting common mode noise, and even though common mode was added, the filter failed the tests, so the filter had to be redesigned.

12.2. POWER SUPPLIES EITHER AS SOURCE OR LOAD

Power supplies are tricky either as the source of supply or as the load. As the source, most designs are inductive output, not at the DC output, but become inductive as the frequency increases. From 0 to 10 or 50 Hz, depending on the supply, the output impedance is in the milliohm range, and it starts climbing after this and looks inductive. If the EMI filter following the DC supply is capacitive input, the inductance of the supply and the input filter capacitor can ring. However, if the EMI filter following the DC supply is inductive input, the stored energy of the inductor has been known to blow the supply on turn-off. The best arrangement for this application is to make the first stage of the EMI filter a T. The inductor facing the supply is half the value of the inductor in the L. A resistor at the input to ground furnishes a discharge path for the storage capacitor on the supply side and reduces the voltage rise of the filter inductor. The resistor value would be of the order of 25 to 100 K ohms.

If the EMI filter is to feed a power supply, low output impedance is required. This means the preceding solution with the T, cannot be the only filter component. This really requires an L with a quality capacitor, preferably the feed-through type, for the output element. The central inductor can be the total of the T inductor plus the L inductor. This is assuming the input in both cases is DC. In Fig. 12.1, the central inductor is three times the line-side inductor because the central inductor is really two inductors. The T has two inductors equal to half the value of the L inductor each. So the central design value would call for 1.5 the calculated value.

What about AC power supplies? Many companies use AC switcher supplies to generate AC at other frequencies such as 50 Hz for Europe and 400 Hz. Some of these generators provide three-phase power outputs. Watch these closely if an EMI filter design is required; they are usually rich in harmonics. One such 400-Hz supply had a strong 2400-Hz component and the filter had a resonant rise close to this. The filter overheated with little or no load current. The filter was not designed to handle the 2400 Hz because no one thought to tell the filter designer this fact. However, to be fair, the customer was probably not aware of this either.

12.3. TRANSFORMERS: 9- AND 15-PHASE AUTOTRANSFORMERS

Autotransformers can cause problems by themselves if the EMI filter requires substantial low-frequency loss. This occurs when the ratio of the total filter

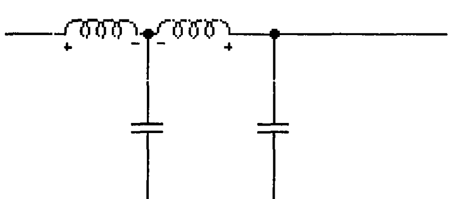

FIG. 12.1 The input T with the output L.

inductance even approaches as little as 2% of the primary inductance. This forms a voltage divider, so the output voltage is below requirement. Again, the customer was either not aware the following device was an autotransformer or not aware of the problem. The 9 and 15 phase often start with an autotransformer yielding low primary inductance.

The cure for this is to add a filter stage or two to reduce the required total inductance of the filter. For example, in going from a single stage to two stages, the values of both the total inductance and capacitance drop drastically. If this was a single L section with 12 dB per octave, it would change to 24 dB per octave, reducing all the component values greatly. In addition, raising the capacitance, allowing lower total inductance in series across the line, may be required. However, this can play havoc with leakage current specifications.

12.4. NEUTRAL WIRE NOT PART OF THE COMMON MODE FILTER

In three-phase systems, the currents are always unbalanced. The neutral wire carries this unbalanced current. This neutral must be part of the common mode choke. This means that there are four equal windings on the common mode core. Without this, the ferrite core is driven into saturation because the difference current is not present to balance the core. In this mode, the ferrite core is heated and generates noise that masks the actual noise of the load. Therefore, the noise is sometimes worse with the filter.

The same is true in single-phase systems. Both the hot and return legs are wound on the common mode core. But in many cases there are grounds ahead of the core, compromising equal current flow through this core. Now the core is unbalanced and in saturation, again generating more noise than the load. If this condition exists and cannot be avoided, remove the ferrite core. It will create more problems than it can cure.

12.5. TWO OR MORE FILTERS IN CASCADE—THE UNKNOWN CAPACITOR

This happens when more than one filter type follows another (Fig. 12.2). For example, in a secure room or screen room, a large three-phase filter powers the entire room. Each power leg and neutral has an insert rated to carry 100 A or more, and these inserts in turn are enclosed in a larger cabinet. This could be fed to cabinets in the room to power a full 19-inch rack of equipment through a filter in each rack, and the individual equipment mounted in the rack is also filtered. This situation places three filters in tandem or in cascade. If the power line filter insert feeding the rack required 100 dB at 10 kHz, this may have forced the designer to cut well below the normal recommended cutoff frequency. This would

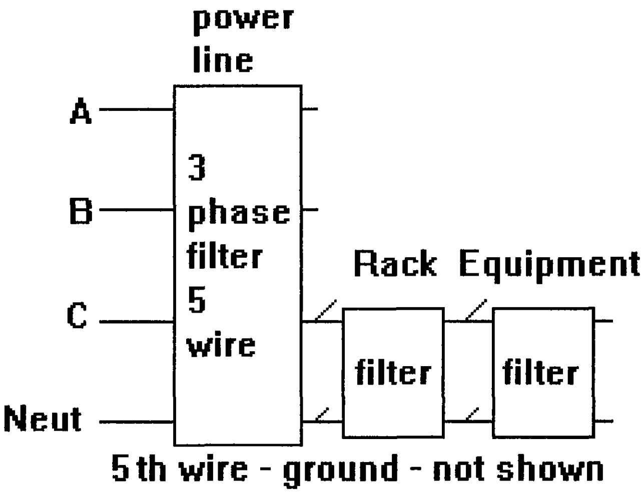

FIG. 12.2 Three filters in cascade or tandem.

be done to get the proper attenuation or insertion loss for reasonable cost, size, and weight. The filter in the bottom of the rack may be a single π filter and be chased by another π in the equipment. These filters detune each other, especially if any, or all, have higher circuit Q values of 2 or better. The higher circuit Q increases the potential of these filters to oscillate, and this would move the cutoff frequency farther into the normal passband. The latter problem has been known to reduce the line voltage to the point that the rack equipment failed to work.

If any of the filters were designed incorrectly initially, this would accentuate the problem. It would be worse if there were multiple double feeds with other filters in cascade off the same power line filter. The cascaded capacitors would total and cause higher line and harmonic currents that would add heat to the filters. This would increase the number of resonant rises within the filter chain along with added suck outs or resonant drops. The problems, as mentioned before, that this situation could create should now be obvious.

12.6. POOR FILTER GROUNDING

A properly designed filter may check out well in the EMI test laboratory or at the EMI filter design house. The reader may have been in these laboratories and have seen all the grounding techniques that are necessary to test the filter and system equipment (Fig. 12.3). The test bench is covered with a sheet of copper that is well grounded. The equipment, or filter, under test is often C clamped tightly to the copper sheet. Most filters are designed to be mounted directly to a ground

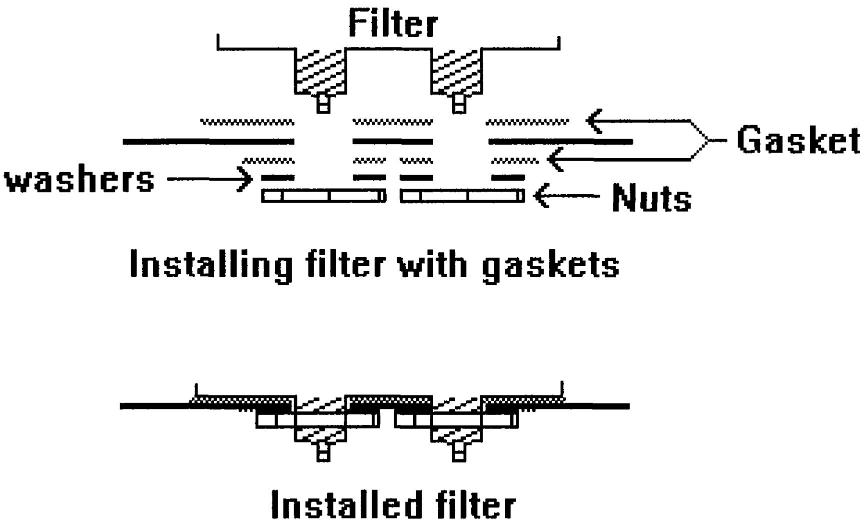

FIG. 12.3 Filter with feed-throughs showing the grounding through the mounting.

plate through input feed-through studs or through the connectors. The filter is mounted through chassis holes with EMI gaskets used on both sides of the chassis. The filter is tightened down to ground with the proper nuts and washers. The gaskets give thousands of ground points with this technique. This also carries the chassis ground plane through the cutout holes in the chassis. If this ground is not provided, the filter fails to live up to its dB rating. Some people scream that the filter is no good even though they were at the laboratory when the filter was checked. Without a good ground, the filter's feed-through capacitors and other components to ground cannot work (Fig. 12.4). The transzorbs and MOVs tied to ground cannot function.

It is easy to see that the filter in Fig. 12.4 does not function properly if the filter case is not grounded or is poorly grounded. The two load-side feed-throughs would be out of the circuit, along with the two line-side transzorbs to ground. The two similar capacitors on the input are noncapacitive input terminals or connectors.

12.7. THE "FLOATING" FILTER

This follows from the topic in Sec. 12.6. The filter was designed according to the method just described. This filter was to be mounted to ground as described, and the users complained that the filter was not functioning. Sure enough, the filter was mounted or hung in air through a plastic hanger. A nice 6-inch green wire (normal hookup wire and not Litz) ran from one lug on the filter through a cable

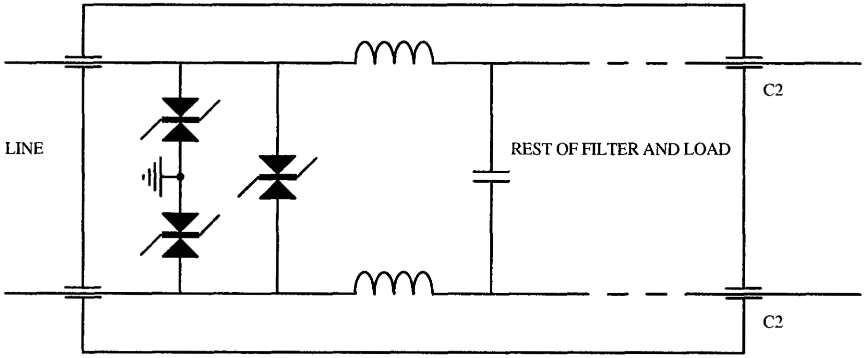

FIG. 12.4 Filter showing lack of a good ground.

harness to a ground. This was not even Litz wire, which could carry most of the upper frequencies. It was common hookup wire, which acts as a good point to radiate an *H* field to the surrounding wires in the same cable form. I was never able to have these users make a change, although I was able to show some of them the folly of their ways. They were way behind schedule, with their customer pressing them to ship, and could not find time to make changes. I think they must have gotten a release from the EMI specification from their customer.

The "green wire" just mentioned has the following properties (Fig. 12.5):

1. A skin effect adds to the AC, or RF, resistance, making the ground more resistive.

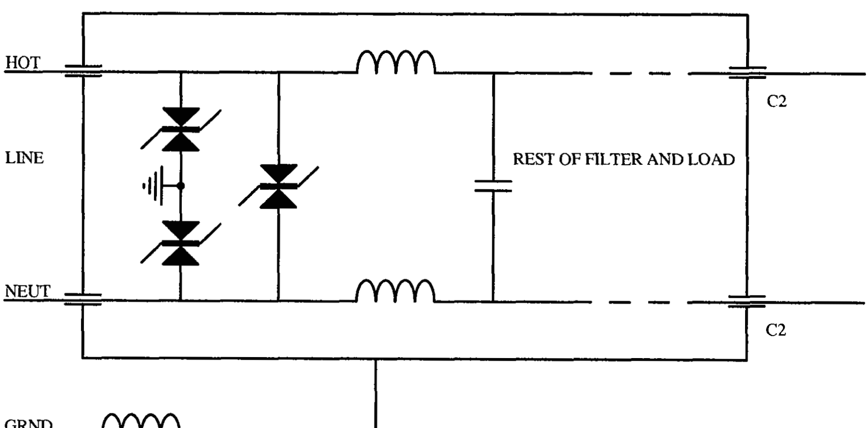

FIG. 12.5 Drawing of the "little green wire."

2. A slow velocity of propagation, making the apparent length about eight times longer, adds inductance to the lead, further impeding the RF current.
3. From 1 and 2, a series RL circuit is formed.
4. Whatever the RF current in the lead, this would radiate to the surrounding areas and wires.

The green wire's inductance would be close to 1.5 μH because of the slow velocity of propagation. This would make the 6 inches almost eight times the length, or close to 48 inches. The full length would equate to at least a meter, and the typical inductance is 1.5 μH/m.

12.8. UNKNOWN CAPACITOR IN THE FOLLOWING EQUIPMENT

This is similar to Sec. of 12.5 discussing filters in tandem and applies only to DC filters. This is also another reason why a filter may be praised by one group and cursed by another. Granted, this double capacitor may not compromise every situation in DC filtering, but the additional component adds cost, volume, and weight (Fig. 12.6). The added capacitor detunes the filter.

12.9. INPUT AND OUTPUT TOO CLOSE TOGETHER

I have seen filters designed by in-house people who put the input and output studs of the filter on the same front face. Try not to do this because the inputs and outputs are too close together. One in question was a face that measured 2 × 3 inches. This means that the filter components that make up the input and output sections of the filter are too closely spaced. This makes a filter designed for 40 dB a 26 dB or so filter because it is easier for the RF to radiate from input to output

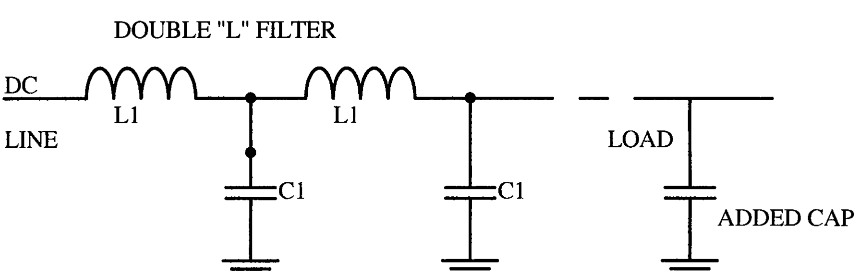

FIG. 12.6 The added unknown or hidden capacitor.

or vice versa. I have seen designs where the input and output are in the same connector and the input and output wires are in the same cable.

The best design is a long filter body with the terminals on the opposite smaller faces. In this way, the unwanted energy is cleaned up as the signals travel through the filter sections toward the opposite end. Figure 12.7 shows an input connector with threads so that EMI gaskets can be used on both sides of the enclosure material. The internal gasket should be the full width and height, and the outside gasket should be the same diameter as the mounting washer. The mounting nut follows this. The output terminals and feed-through capacitors are on the far end.

If you cannot escape this condition, say in the 2 × 3 inch section just mentioned, place a shield between the two halves. Do not allow the harnesses feeding the filter to carry both feeds. There are two ways to divide the 2 × 3 area, and the choice is determined by the component size. The first is to divide the 3-inch width by two and install a shield that runs well to the other end of the filter. This makes the shield 2 inches high (Fig. 12.8). The input or output section is now 2 × 1.5 inches. The input studs and components should be installed on one side of the filter, and the additional components should be installed toward the rear and then turn the corner and head toward the front again. If some components are larger than the compartments formed by the shield, the second method of using the shield to split the 2-inch height can be used. Continue with the same technique.

If the components still do not fit, it is best to change the layout so that the filter can run from end to end without doubling back as in Fig. 12.7.

12.10. GASKETS

EMI filters are normally mounted through small holes in the case, and gaskets are required to give a very good ground to the filter case (Fig. 12.9). Often, the gasket is omitted, and the filter has a poor ground return. Any feed-through capacitors

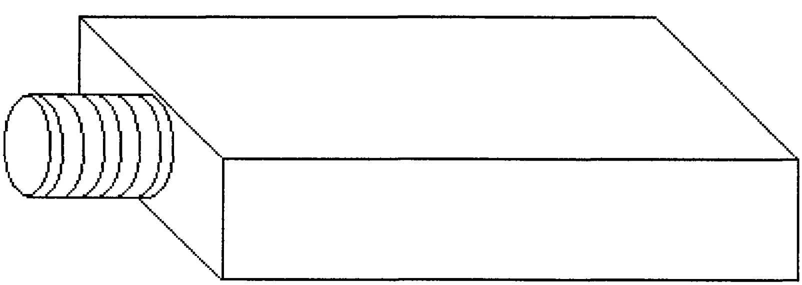

FIG. 12.7 Filter using a connector: length compared to width.

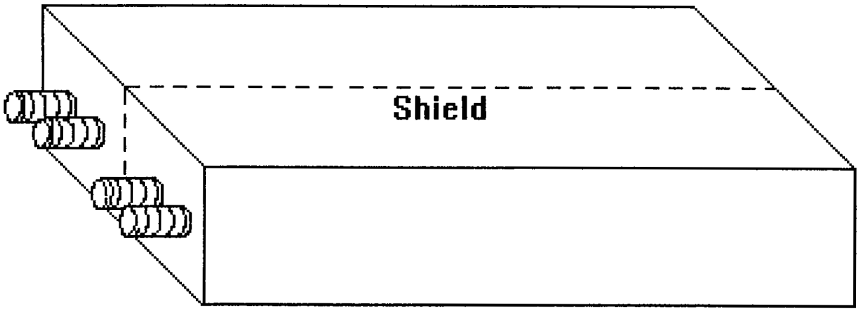

FIG. 12.8 Input and output on same filter face.

and MOVs to ground or case within the filter are compromised. Sometimes, the filter is removed for various reasons or replaced and the same gaskets are reused. This again reduces the effectiveness of the ground, and the components to case ground of the filter are less effective. On the other hand, it is better to reuse the gasket than to be without any gasket.

Another reason to use gaskets is to complete the missing ground path through the holes through which the filter is to be inserted. The proper way to mount the filter requires two gaskets: a gasket between the case and filter and the second set between the outside case and washer(s).

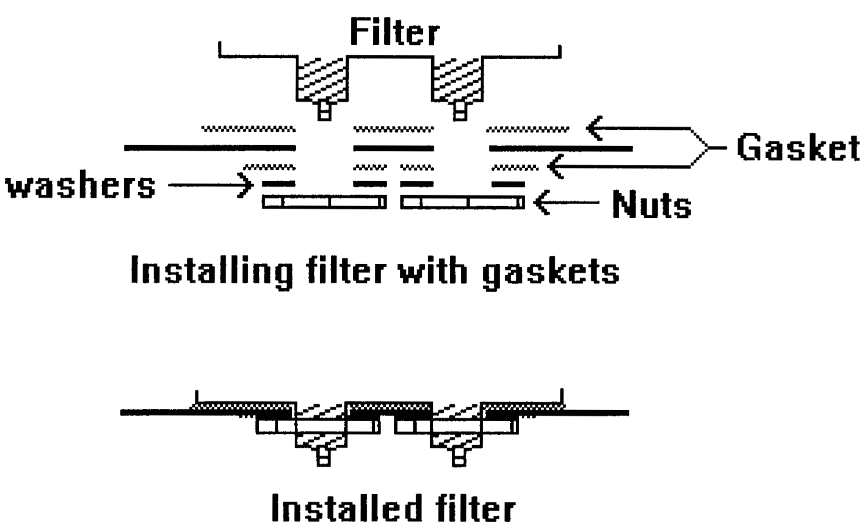

FIG. 12.9 EMI gasket used to install a feed-through input filter.

There is a proper torque for an EMI gasket, supplied by the gasket manufacturer, that ensures excellent grounding without collapsing the gasket. This torque gives about the minimum DC and AC resistance between the filter and ground.

When two objects are tightened down without gaskets or with old, reused gaskets, few points touch, giving a resistive path. When new gaskets are used, thousands of points touch, giving a low-impedance path.

13

Waves as Noise Sources

The waves more commonly encountered within systems are discussed here. These waves are not as pure as drawn here; some parasitic oscillations from the transformer and other components will be superimposed on the waveform. These waveforms also have added rise and fall times, some of which are shown as pure step functions that are impossible to achieve. Each voltage, or current, waveform is from Fourier equations and is listed along with its differential.

13.1. THE SPIKE

Spikes are one of the common waves, or noise sources, encountered (Fig. 13.1). They are seen when double-ended, or Royer, circuits are used, with one switcher turning off while the other is turning on. The two currents add and, as far as the switchers are concerned, the current nearly doubles. Other elements, such as diodes, also add to this current, and often the total spike current is many times the average switcher current. This spike occurs twice per cycle of the switcher frequency, so the frequency of the spike is double the switcher frequency. Without some correction in the switcher circuit in a DC system, this high spike current can greatly reduce the DC voltage. This impairs switcher operation, and most of this spike energy is passed back to the EMI filter.

With an off-line regulator, the storage capacitor is often very large, in the 500 to 2000 μF range or more. If the spike frequency is 60 kHz, twice the 30 kHz switcher frequency, this large storage capacitor may be well above its self-resonant frequency (SRF) and be either open or inductive. If this section is not isolated at this point, a good-quality feed-through capacitor should be tied in

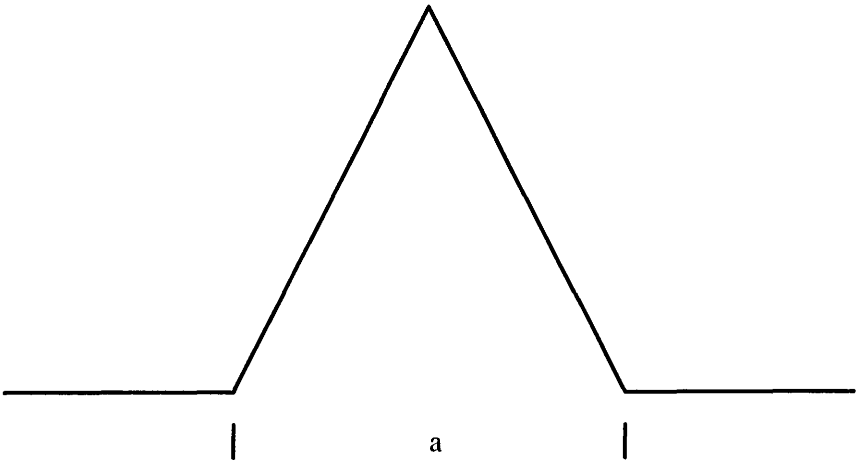

Fig. 13.1 The spike.

parallel here. If this section is isolated, a good line-to-line capacitor (X capacitor) should be tied across the storage capacitor. Another well-placed capacitor should be used at the junction of the B+ and the switcher transformer. This should be a good quality feed-through if the switcher is not isolated or a good-quality line to line if it is isolated.

An easy way to calculate this feed-through or line-to-line capacitor is as follows. Estimate the lower DC storage voltage, and divide this by twice the normal switcher current (about the peak of the spike current) for the worst-case, or lowest, working impedance. The line-to-line added capacitor impedance should be 0.1 this impedance at the spike frequency, 60 kHz here. The leads of the line-to-line capacitor should be as short as possible. Keep this capacitor's SRF well above 10 times the spike frequency—again 60 kHz here—or 600 kHz. Round the calculated value up to the next higher standard value of capacitance.

This is a pulse, and it requires the ability to handle the higher pulse currents. Therefore foil, not metallized film, should work with the proper derating. Ceramic capacitors work very well here. The leads must be short because articles from the Institute of Electrical and Electronics Engineers (IEEE) state that a capacitor with the same lead length will not give the higher SRF required. The equation of the spike is as follows:

$$\frac{Ea}{T} + \frac{Ea}{T} \sum_{N=2,4,6,8}^{\infty} \left| \frac{\sin(X)}{X} \right|^2 \cos\left(\frac{2\pi N t}{T}\right) \tag{13.1}$$

where a is the pulse width and

$$X = \frac{2\pi Na}{T}$$

and the differential is:

$$-\left[\frac{E}{2\pi a}\sum_{n=\text{odd}}^{\infty}\frac{1}{n}\left(\frac{\sin(X)}{X}\right)^2 \sin\left(\frac{2\pi nt}{T}\right)\right] \tag{13.2}$$

13.2. THE PULSE

The pulse is similar to the quasi-square but pulses in the same direction, like the spike (Fig. 13.2). This is similar to the Royer, where the dwell time for both halves is not on for the full half-period. Therefore, this generates pulses of current twice per period, once for each half, and is again at twice the switcher frequency. The design and considerations are the same as for the spike.

The equation of the pulse is:

$$\frac{Ea}{T} + \frac{2E}{\pi}\sum_{N=2,4,6,8}^{\infty}\frac{1}{N}\cos(\pi Nt)\sin\left(\frac{\pi Na}{T}\right) \tag{13.3}$$

where a is the pulse width. The differential is

$$\frac{-4E}{T}\sum_{N=2,4,6,8}^{\infty}\sin(\pi Nt)\sin\left(\frac{\pi Na}{T}\right) \tag{13.4}$$

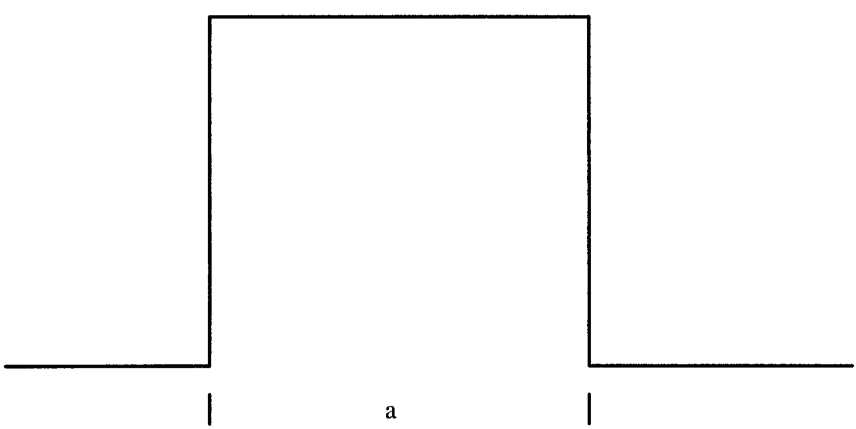

FIG. 13.2 The pulse.

13.3. THE TRAPEZOID

The trapezoid (Fig. 13.3) is more realistic than the pulse or the quasi-square because both have a step function or zero rise time and fall time. The trapezoid has a rise and fall time like that in the real world. Also, the energy in the EMI spectrum is less with the trapezoid. The disadvantage of the trapezoid is that the efficiency of the power supply is reduced. More power is dissipated in the switch during the rise and fall time.

A capacitor, found by the method shown for the spike, can be paralleled with the main storage capacitor. Follow the same method; T_0 is the pulse width and T_r is the rise and fall time. The equation of the trapezoid is

$$\frac{2E(T_0 + T_R)}{T} \sum_{n=\text{odd}}^{\infty} \left[\frac{\sin\,[\pi n(T_0 + T_R/T]}{\pi n(T_0 + T_R)/T}\right] \sin\left[\frac{\pi n T_R/T}{\pi n T_R/T}\right] \cos\left(\frac{2\pi n t}{T}\right) \quad (13.5)$$

The differential is

$$-4E \sum_{n=\text{odd}}^{\infty} \left[\sin\left[\frac{\pi n(T_0 + T_R}{T}\right]\right] \sin\left[\frac{\pi n T_R/T}{(\pi n T_R/T)}\right] \sin\left(\frac{2\pi n t}{T}\right) \quad (13.6)$$

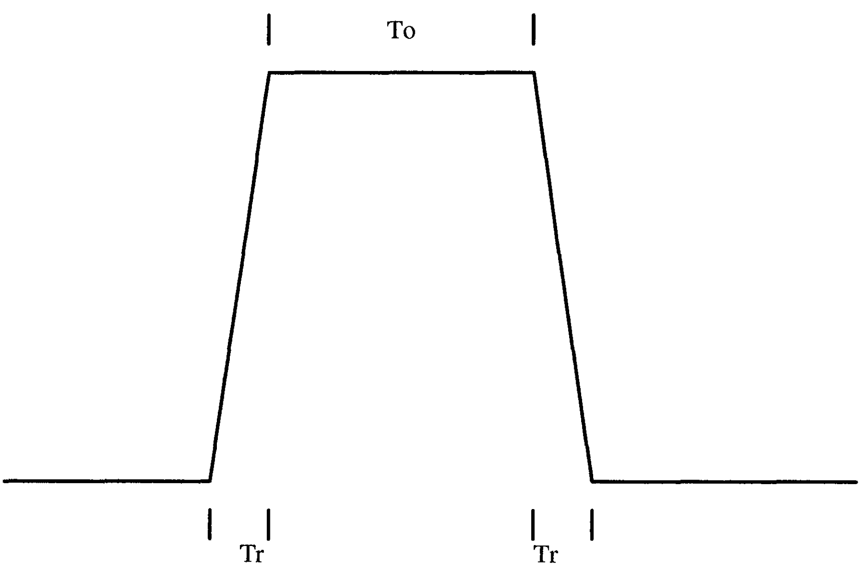

FIG. 13.3 The trapezoid.

13.4. THE QUASI-SQUARE

The quasi-square is the wave applied to the gate, or control, of the switch, and the reciprocal is applied to the opposite control (Fig. 13.4). The pulse width is a and the dwell angle is one half the angle between the pulses. The dwell angle occurs before and after each pulse, or there are four dwells per period. This turns on the opposite device every half-cycle, and each is turned on for less than a half-cycle. The output is the current pulse, and the pulse is twice the quasi-square frequency. The pulse is filtered by the technique described in Sec. 13.1 for the spike.

This quasi-square wave and its output wave, the pulse, are not real-world waves. Both lack the rise and fall time that any wave has. This is due to small capacitors and inductors in the path. All devices such as circuit boards, the natural capacitance of wiring, and the input capacitance of the switch have properties that require time to discharge or change. This results in ramps in the wave shape, changing the so-called square, quasi-square, and pulse into trapezoids that show the rise and fall time. The quasi-square wave should look more like Fig. 13.5.

The equation of the quasi-square wave is

$$\frac{4E}{\pi} \sum_{N=1,3,5,7}^{\infty} \frac{1}{N} \sin(n\omega t) \cos(N\phi) \tag{13.7}$$

The differential is

$$\frac{8E}{T} \sum_{N=1,3,5,7}^{\infty} \cos(N\omega t) \cos(N\phi) \tag{13.8}$$

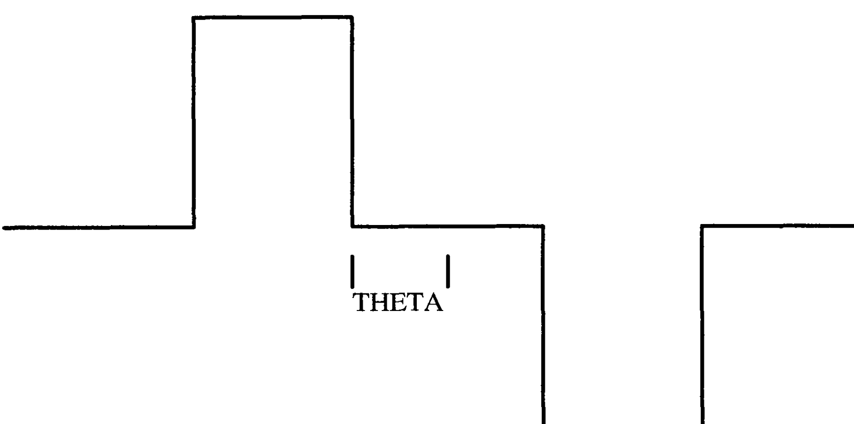

FIG. 13.4 The quasi-square.

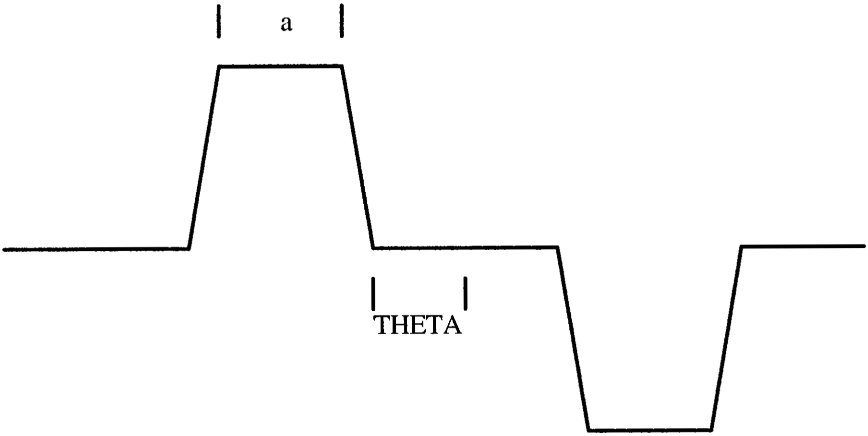

FIG. 13.5 The quasi-square with rise and fall times.

13.5. WHY DIFFERENTIATE?

Each pulse listed up to now has been differentiated for good reason. If each of these waves represents current pulses, the H field needs to be calculated. If each wave is only a small change in the DC current state, the radiated H field is small. A steady-state DC current generates only a steady-state H field. This H field cuts the surrounding surfaces only during turn-on and turn-off. These waves are often pulses with peak amplitudes of many amperes. They generate strong varying H fields, leading to radiated energy (Fig. 13.6). The formula is

$$\frac{0.4\pi N I_P}{M_{pl}} \tag{13.9}$$

where N is the number of turns, I_p is the peak current, and M_{pl} is the magnetic path length. If this is a wire carrying this I_p current, the turns are then equal to one and M_{pl} is the circumference. The radius is from the wire to the surface in question (usually to the closest outer metal cover or another wire). If the length between the wire and cover, or the part to which this is radiating is R_r, then this equation is reduced to

$$\frac{0.2 I_P}{R_R} \tag{13.10}$$

What is needed is the differential of H with respect to t, and the only variable that depends on t in Eq. (13.10) is I_p. Therefore, dH/dt is equal to

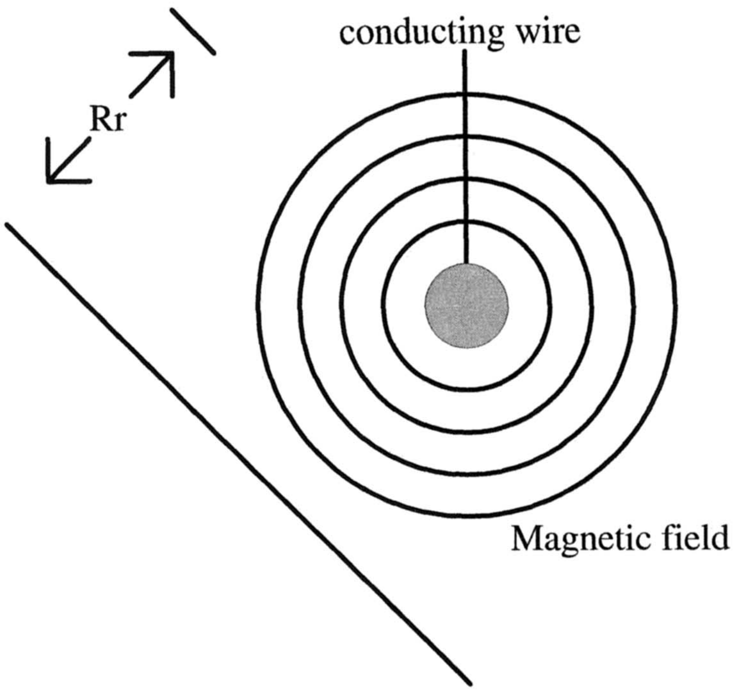

FIG. 13.6 The radiated field.

$$\frac{0.2}{R_R}\frac{dI_P}{dt} \qquad (13.11)$$

So, 0.2 divided by R_R, $n = 1$, times the differential from the proper wave in question should give the approximate magnetic field intensity in amperes per meter if R_R is in meters.

13.6. THE POWER SPECTRUM—dB μA/MHz

How much power is in the various waves discussed in the earlier sections? These form envelopes that provide the peak power, which varies with the amplitude of the current, I_p. If the current pulse width is a and T is the period, the equation for the pulse power dB/MHz is as follows (Fig. 13.7):

$$20 \log\left(\frac{2Ea \times 10^6}{T}\right) \qquad (13.12)$$

where E is the amplitude. This gives a flat line across the frequency spectrum to

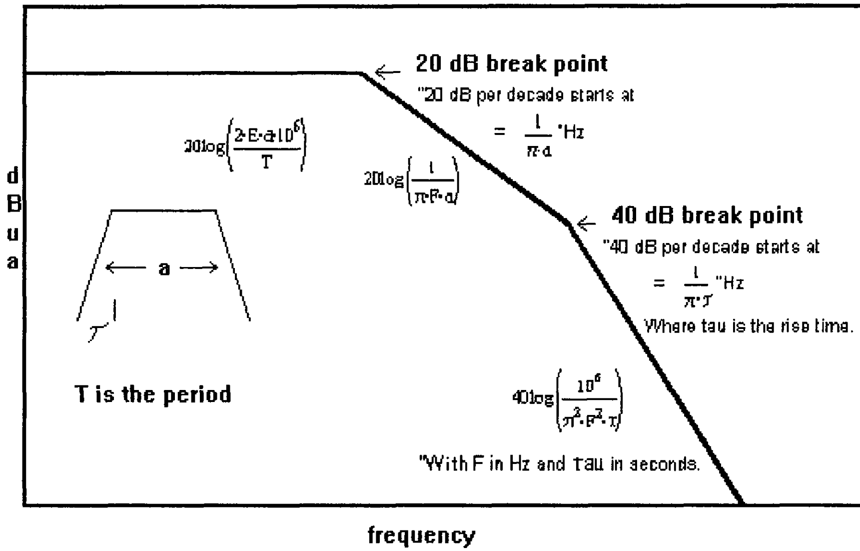

FIG. 13.7 The dB μA per megacycle.

the 20 dB per decade, or 6 dB per octave, breakpoint. This point starts at the frequency

$$\frac{1}{\pi a}\ \text{Hz} \tag{13.13}$$

where a is the pulse width in seconds and the decibel level after the breakpoint is

$$20\log\left(\frac{1}{\pi F a}\right)\text{dB} \tag{13.14}$$

There is also a 40 dB per decade breakpoint, 12 dB per octave, which depends on the rise time. The larger the rise time, the sooner the breakpoint occurs. The point here is that EMI energy can be greatly reduced by using this principle. The efficiency of the power supply decreases but the EMI energy is less. This is usually a better trade-off for smaller and lighter units and should reduce the total cost. If the rise time is 10% or better, this becomes, in essence, an L filter added to the existing EMI filter at frequencies above the 40 dB breakpoint.

$$\frac{1}{\pi\tau}\ \text{Hz} \tag{13.15}$$

The added loss after the 40 dB per decade breakpoint is

$$40 \log \left(\frac{1}{\pi^2 F^2 \tau} \right) \text{dB} \tag{13.16}$$

where F is in hertz and τ is in seconds.

13.7. MIL STD 461 CURVE

This specification is in dB μA/MHz rather than insertion loss (Fig. 13.8). It is rather difficult to convert from this to insertion loss. If the impedances were equal, the conversion would be the obvious 120 dB. Insertion loss is stated in terms of how much loss is required, whereas dB μA/MHz, or dB μV/MHz, is in terms of how much noise is allowed. There is a 13 dB conversion factor due to changing from 50 ohms to a probe estimated to be 1 ohm. This subtracts from the 120, giving 107 dB, and then the probe correction factor is needed. This is a function of the probe, the frequency bandwidth. This last correction is added back to the 107 dB. Usually, this method comes with the probe.

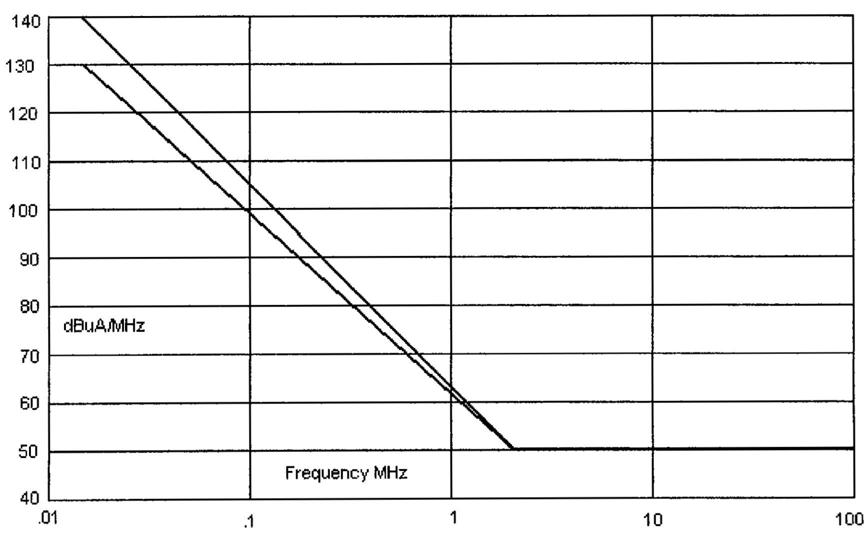

FIG. 13.8 MIL STD 461 C specification. Upper curve, Navy, Air Force; bottom curve, Army.

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14

Study of the Off-Line Regulator

Col. W. T. McLyman (Jet Propulsion Laboratory, National Aeronautics and Space Administration; now retired) has worked out the approximate weight and volume relationships using the area product, A_p. These depend on the type of core and the winding technique. The purpose here is to show that the entire weight and volume are reduced if an inductor of critical value is included in the linear power supply. This inductor is placed between the diodes and storage capacitor. This mainly applies to filters that have serious loss requirements at very low frequencies such as specified in Military Standard (MIL STD) 461. Years ago, these were called inductive input power supplies. This also applies to power factor correction circuits that remove the high current spikes required to charge the storage capacitors twice per cycle. Regardless of which technique is used, the size and weight of the EMI filter are reduced, especially for 461 and similar specifications.

Also discussed in this chapter is what the high harmonic content does to the line voltage, which affects other equipment operating on the line. In the last section, the Keith Williams method is discussed, which is used primarily throughout this chapter and the book. The following section uses the same equations to show that the added harmonic content on the power lines is increasing. This happens today because most of the electronic equipment making up the load does not employ power factor correction circuits or any other technology to remedy this problem. Another reason for the higher harmonic content is the lack of the inductive input technology discussed in Sec. 14.1. This condition will improve as the loads meet the newer standards requiring power factor correction circuits, but the equipment making up the load will take years to replace.

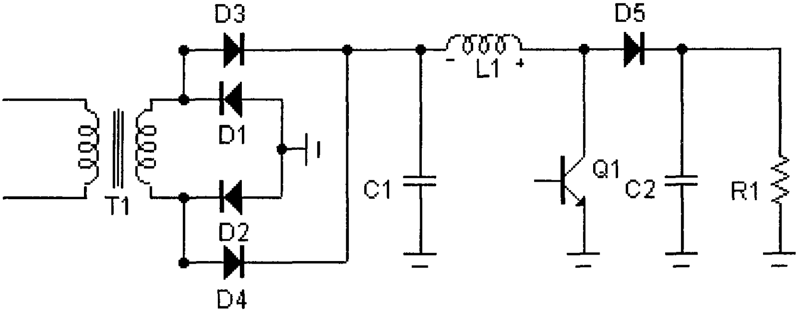

FIG. 14.1 Power supply with capacitive load without critical inductor.

14.1. WITH OR WITHOUT CRITICAL VALUE OF INDUCTANCE—SIZE AND WEIGHT DIFFERENCE OF THE FILTER

In referring to this section, many have said that they know that the critical inductor will make the EMI filter smaller, yet rarely is this application seen in practice. In other words, the weight and volume of the EMI filter without the inductor are much greater than those of the filter and inductor together using the critical inductor (Figs. 14.1 and 14.2). This is especially true if the line frequency is 400 Hz. The reason is seen in Eq. (14.1), where F is in the denominator. The inductor L_C is much smaller. From another standpoint, this is true if the EMI filter specification demands high attenuation at very low frequencies, as required by the 461 specifications (Fig. 14.3). This may not be a good trade-off if the only goal is to meet Federal Communications Commission (FCC) and similar specifications, especially at 50 or 60 Hz. If the dB loss of the filter must be 20 dB, or greater at 10 kHz, this should be looked into regardless of the line frequency.

The formula for the critical inductance, L_C, is

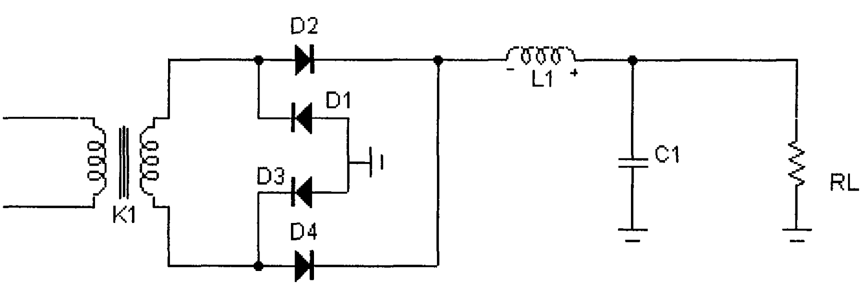

FIG. 14.2 Power supply with critical inductor, L1.

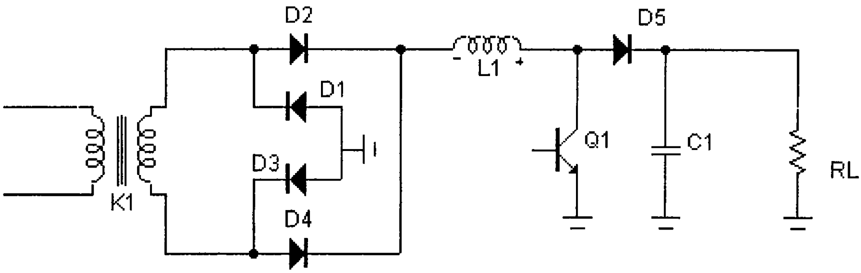

FIG. 14.3 Simplified power factor correction circuit.

$$L_c = \frac{R_0}{6\pi F} \tag{14.1}$$

The value of R_0 is the highest DC supply voltage divided by the lowest load current, and F is the line frequency, not the ripple frequency. The addition of the inductor lowers the storage capacitor average voltage, but the power supplied to the switcher(s) must be the same. This raises the current minimum, decreasing the value of R_0 and, therefore, L_c. This subject is covered in W. T. McLyman, *Transformer and Inductor Design Handbook* (Marcel Dekker). This is handy if the cores for the L_c inductor and the filter are not known. Otherwise, the weight and size are known.

If the area product is known in centimeters to the fourth power, the *approximate* weight, in grams, and the volume, in cubic centimeters, can be estimated from Figs. 14.4 and 14.5. These have been assembled from McLyman's book. The K_v factor multiplied by the $A_P$value gives cubic centimeters, and K_w multiplied by A_P gives grams. This is useful if the inductor has not been fully designed. If the inductance is known, the energy can be calculated from Eq. (14.2). From this, A_P can be found from Eq. (14.3) knowing the core type. From this and Figs. 14.4 and 14.5, the volume and weight can be estimated.

Laminations and tape-wound cores are rarely used for EMI filters but are included in Figs. 14.4 and 14.5. The lamination information is needed for the critical inductance in some applications. The area product is the cross-sectional area of the core in square centimeters times the wire area or window area, also in square centimeters. The A_P must be multiplied by the stacking factor, if applicable (laminations and C cores). Continuing with McLyman's book, the value of the critical inductance must be found from Eq. (14.1). The next step is to find the energy stored in the inductor:

$$\frac{L_c(I_P)^2}{2} \tag{14.2}$$

CORE TYPE	K_v Cm^3	K_w Grams	x
POT CORE	14.5	48.0	1.20
POWDER CORE	13.1	58.8	1.14
LAMINATIONS	19.7	68.2	1.14
C CORE	17.9	66.6	1.16
SINGLE-COIL	25.6	76.6	1.16
TAPE WOUND	25.0	82.3	1.15

FIG. 14.4 Ratios of area product (A_p) to weight and volume.

CORE TYPE	K_j 25^0·C	K_j 50^0·C	x
POT CORE	433	632	1.20
POWDER CORE	403	590	1.14
LAMINATIONS	366	534	1.14
C CORE	323	468	1.16
SINGLE-COIL	395	569	1.16
TAPE WOUND	250	365	1.15

FIG. 14.5 Value of K_j for the various core types.

where I_p is the peak current the inductor must carry. The smallest load current determines the value of L_c, but the peak of the largest charging current of the storage capacitor, I_p, through the inductor determines the energy that the inductor must handle. The minimum area product, A_P, required to handle the energy is then, according to McLyman,

$$\frac{[2(\text{energy}) \times 10^4]^X}{B_M K_u K_j} \tag{14.3}$$

Here B_m is the maximum flux density in teslas (gauss/10,000), K_u is the winding factor for the type of core configuration chosen (typically 0.4 for most configurations), and K_j (°C) comes from Fig. 14.5 (from McLyman's core configuration table on page 106 of the book cited earlier).

As an example from McLyman, figure the powder cores at 20 A with 100 μH. For the powder core, K_v is 13.1, K_w is 58.8, and X is 1.14. The energy is 0.02. From the energy and the K_j value of 403, A_P can be calculated as 19.13. Multiplying by K_v and K_w yields 2.48 pounds and 15.29 cubic inches. The design should be for 25°C to reduce the temperature rise of the entire filter, and this may be good advice for the critical inductor too. The saving of volume and weight would be greater if the 50°C temperature rise can be tolerated for the critical inductor because K_j is in the denominator, that is, if the surrounding circuits can withstand it. X is repeated here for convenience.

If the current pulse equations, from Sec. 4.2, for charging the storage capacitor, without the critical inductor, are solved for the approximate peak current, the energy can be found. These are repeated here.

$$\begin{aligned} \sin(Y + X - B) &= \sin(Y + X - A)e^{\frac{(A-B)}{\tan X}} \\ \sin(A) &= \sin(B)e^{\frac{(B-A-N\pi)}{\omega C R_L}} \end{aligned} \tag{14.4}$$

where

$$Y = \pi - \tan^{-1}(\omega C R_L) \qquad X = \tan^{-1}\left(\frac{\omega C R_S R_L}{R_S + R_L}\right) \tag{14.5}$$

Equations (14.4) and (14.5) assume the engineer knows the following: the line frequency, F, so that ω ($2\pi F$) can be calculated; C, the storage capacitor, in farads; R_S, the line resistance (typically 1 ohm); and R_L, the load resistance (the lowest value of resistance, which is the highest current). The angles are in radians, but degrees are given here for the most part. The values of X and Y are then calculated, along with tan(X). Substitute these into the first two equations. In one equation, guess a value for A and solve for B. Insert B into the second and solve for A. Average the two values for A, and restart the process by substituting this new value into the first equation again. A spreadsheet is handy for this, with each row being a new estimate. If the engineer is going to write a spreadsheet of this type, it is advisable to multiply the initial guess by 49, add the calculated value to it, and then divide by 50. Do this until the values are similar, and then multiply by 9 and divide by 10. The reason is that the changes may be too abrupt and throw the value of A into the wrong court. See the Appendix and the disk. Once A, the start angle, and B, the stop angle, are known, U_1, the first minimum voltage angle guess; V_1, the first maximum voltage angle guess; and P_1, the first peak current angle guess, can be estimated.

$$\begin{aligned} U_1 &= \sin^{-1}\left|\frac{(R_s+R_l)}{R_l}\sin(A)\right| \\ V_1 &= \sin^{-1}\left|\frac{(R_s+R_l)}{R_l}\sin(B)\right| \\ P_1 &= \frac{(A+B)}{2} \end{aligned} \tag{14.6}$$

Then U, V, and P can be solved through iteration. Use similar techniques for the averaging, as before, otherwise the changes may be too great. Again, a spreadsheet is handy and all can be estimated within the same spreadsheet. Substitute the known values of A and B and then the estimated U_1, V_1, and P_1 into one term and solve for the other U, V, and P. Average the two and solve again. A spreadsheet is included on the disk.

$$\begin{aligned} \cos(X-U) &= \frac{(R_s+R_l)\sin(Y+X-A)e^{\frac{(A-U)}{\tan X}}}{\sin(Y)} \\ \cos(V-X) &= \frac{(R_s+R_l)\sin(Y+X-A)e^{\frac{(A-V)}{\tan(X)}}}{\sin(Y)} \\ \frac{\cos(Y+X-P)}{\cos(Y+X-A)} &= e^{\frac{(A-P)}{\tan(X)}} \end{aligned} \tag{14.7}$$

Then E_U, the minimum voltage at angle U (in radians), and E_V, the maximum voltage at angle V (again in radians), can be found.

$$E_U = \frac{E_M R_l \sin(U)}{R_s+R_l} \qquad E_V = \frac{E_M R_l \sin(V)}{R_s+R_l} \tag{14.8}$$

where E_M is the peak line input voltage (or the stepped-up or stepped-down peak voltage from a transformer). Also, the peak current, at angle P, can be calculated. This peak current is the value required for this section. The beauty of this is that the appendix program will solve any of these for you with other useful information such as the power factor and peak-to-peak ripple.

As an example, E_M is 167 V, F is 400 Hz, R_s is 1 ohm, and R_l is 10 ohms. The storage capacitor is 0.0002 F (200 μF), and N is 1 for a full-wave rectifier. It follows that ω is 2513.3. So both X and Y can be solved; X is 0.428626 and Y is 1.767176, and their sum is 2.195802. The initial guesses for A and B are π/3 and π/1.6, respectfully. Through iteration, A and B can be determined using Eqs. (14.4). The start angle, A, is 36.98775 degrees and the stop angle, B, is 123.7759 degrees. This accuracy is not required but is given here to show the values calculated. The initial guess for the peak current angle, P, is the average of A and B, which is 80.35581 degrees. Of course, these are all calculated with radians, but I give the terms in degrees for clarity. The final value for the peak current angle is 71.55862 degrees. With A, B, Y, P, X, E_M, R_s, and R_l known, substitute into Eq. (14.9) for the value of I_M. From the appendix equations and the disk provided, the peak-to-peak ripple is 42.2 V, the valley voltage is 99.78, the fundamental current is 16 A and the leading angle is 13.33 degrees. However, the total harmonic RMS current is 19.26 A. The power factor is 0.81854.

$$I_M = E_M \cos(X) \frac{\sin(P - Y - X) + \sin(Y + X - A)e^{\frac{(A-P)}{\tan(X)}}}{(R_s + R_l)\cos(Y)} \tag{14.9}$$

The peak current (P) is 38.54 A. Again, this is the value needed to design the filter inductor. The inductor must not saturate at this peak current of 38 A. This is not the worst situation, where the RMS current of 16 A is compared with the peak of 38 A. The reason is that this system is approximately 1500 W. The conduction angle is 86.73 degrees, so the peak current is low. If R_s was 20 ohms, the conduction angle would be less, the peak would be 26.93 A, the peak-to-peak ripple would have been less, and so on. But the ratio of the peak current to the RMS would be much greater. In the initial problem, the EMI filter designer would be given the 16 A value and the inductors would saturate.

If, in our example, the filter required several 100-μF inductors for the 16 A, two MPP 55716s taped together with 27 turns with three strands of AWG 15 un2 would give the desired results. The flux density would be 3348 gauss with an approximate temperature rise of 16.2°C. It would be well below saturation if this was a true sinusoidal wave at 400 Hz. This wave is neither sinusoidal nor pulse but has properties of each. However, for the total harmonic current of 19 A, the flux density is still only 3976 gauss with 21.6°C. The flux density is a little high but is acceptable for a 50, 60, or 400 Hz sinusoidal wave. However, 38 A is well above saturation at 7951 gauss with a rise of 68.7°C and would be calculated if the wave was a pulse. This temperature rise is not realistic because this is a peak current, not RMS. The core would saturate but not overheat. The proper way to avoid this is to determine the true total RMS, using the equations provided here or on the CD. Add a little headroom; here, go to 20 A. Keep the flux density well below 50% saturation. For an MPP, saturation is 7000 gauss. Strive for a flux density of 3000 gauss or below. Going to the next size bigger core, MPP 55111, and using two cores taped together, as before, with three strands of AWG 14 with 39 turns will meet the inductance requirement. The flux density is 2509 and the temperature rise is 24°C. Now the 38 A peak should not drive the core into saturation. However, the wound diameter is 2.5 and the height is 2 inches. The volume consumed is $2.5 \times 2.5 \times 2 = 12.5$, which is not too far from the McLyman estimate of 15.29. These cores would be expensive, especially if several of these inductors were needed. The powder core program on the CD calculated the MPP core used here.

It should be apparent that a different core must be used. This calls for a switch to a C core or tape wound using the same steel. Again, this program is on the CD. The C core style is a 4 mil, as normally specified for 400 Hz. This core gives adequate Q and lower core losses. Better yet, use a 2-mil core for better Q and better tuning, if required. But this change adds to the cost. The core chosen is CH-44. Two coil forms are wound with two AWG 15 un2 wires with 32 turns. The coils are in parallel on each leg, and there are three layers on each coil form. The required gap is 40 mils or 20 mils per leg. The flux density is less than 10.4 kG and the temperature rise is 30°C. This is based on the necessary 20 A. The approximate size is $2.3 \times 1 \times 1.75$, and with one core per inductor it should be less expensive than the two stacked MPP cores.

If the designer placed an inductor (L_C) between the diodes and the storage capacitor of 1.33 MH, the filter inductors would drop in value and could go to almost 4000 gauss. However, the 1.33 MH inductor requires some headroom, especially if the load current will drop below the value needed or if the value of R_0 increases well above 10 ohms. If the load is reasonably constant, 2 MH should do the job. See Eq. (14.1) and Fig. 14.2. This would create a constant current of 12.13 A, further reducing the size and cost of the filter cores. L_C would require a 4-mil CH-61 core with two coils in parallel with 75 turns each of AWG 14. The gap is 36 mils with 18 mils in each leg and the temperature rise is 25°C. This change allowed one core for each filter inductor instead of two as before. Also, the core was changed to MPP 55110, which has a higher A_l value and is less expensive. There are 37 turns with three strands of AWG 14. The temperature rise is 10°C.

With L_C being primarily a DC inductor and a constant current of 12.13 A, the storage capacitor can also be reduced to at least half value. The current leaving the capacitor is the same as that charging the capacitor. This gives a further reduction of cost, size, and weight of the entire power supply and EMI filter. The current in the filter is a square wave with a peak of 12.13 A.

14.2. THE ADDED POWER LINE HARMONIC CONTENT CAUSED BY THE OFF-LINE REGULATOR

Continuing with the system in Sec. 14.1, 1500 W can be powered by normal wall plugs (i.e., a regular wall outlet) and does not require special wiring from the service. However, the total harmonic current is over 19 A. Now the system leans toward a dedicated line. Adding the fact that the peaks go to over 38 A, the line is definitely a dedicated line with its own circuit breaker. If the equipment is listed as 120 V, 12.5 A and is now wall fed, the odds are that the breaker will trip. The question is, what is the total resistance from the input of the power supply, diodes, and transformer (if any) through the filter, the line, and the breaker? In the power supply equation, R_S was 1 ohm. At the peak current instant, the voltage drop is 38.54 V (1×38.54), and this is at 71 degrees. The sin of 71 degrees is 0.9455. This times the peak of 167 is 157.90. Now subtracting the drop at this angle gives 119.36 V.

The point here is that this type of equipment is on the increase. Not all gulp a peak current of 38 A and not all are at 71 degrees. In 1985 and before, 85% of the total power was for lights and power for motors for refrigerators, lathes, and so on. Only 15% was for radios, televisions, copiers, and some computers. Now homes have fax machines, computers, several TV sets, and other devices requiring some peak current well above RMS. The medical field with magnetic resonance imaging (MRI) equipment, x-ray equipment, heart monitors, and so on all adds to the voltage distortion ahead of 90 and 270 degrees. So the voltage is no longer just the fundamental frequency. The same is happening to the three-phase systems with so much odd-order harmonics; the third harmonics, and their multiples, are heating the neutral and the neutral filters because of the unbalance currents plus the third order. These add back in phase (120 degrees times 3, 6, 9, 12, 15, etc.; all are multiples of 360 degrees or 0 degrees) on the neutral. The people at RFI have received many calls asking why the neutral filter is so hot. They are asked to measure the currents on the legs. The current is then reported to be greater than the phase currents. The next question asked is whether RFI can design a filter to correct this to remove the third-order currents. When the customers are told the size, weight, and approximate cost, they look for other solutions. Some of these solutions are to rebalance the loads, change to power factor correction type power supplies where possible, and increase the size of the neutral to handle the current required.

The best solution is to use power factor type front ends, but this is hard to do at higher powers. The critical inductor (L_C) also helps, especially at nearly constant current requirements. The inductor is the smallest possible with this arrangement.

14.3. KEITH WILLIAMS' METHOD

Keith Williams works for Grand Transformers in Grand Haven, Michigan, and well known throughout the magnetic community. He developed these equations from O. H. Schade graphs developed in the mid-1940s. These equations are complex trigonometric transcendental functions and can be solved only through iteration. In other words, these equations cannot be solved by algebraic means. Keith Williams has a program that solves these complex equations.

15

Initial Filter Design Requirements

The goals of this chapter are as follows:

1. Differential mode design goals
2. Common mode design goals
3. Methods for reducing the size of the inductor related to inductor current
4. Estimate of common mode load impedance

This chapter discusses these goals and what can happen if they are not met.

15.1. DIFFERENTIAL MODE DESIGN GOALS

The idea is to make the differential mode filter transparent to the line. This is ideal for both DC and AC systems and easy to accomplish at DC, 50 and 60 Hz. However, it is a demanding task at 400 Hz in any system requiring substantial loss at low kilohertz frequencies. The requirement stated really means that the load impedance is transferred to the input of the filter at the line frequency and most of its lower harmonics. The harmonic content depends on the quality of the line and the load. The higher the line impedance, the more effect the odd-order harmonic content has in distorting the sine wave voltage shape. These harmonics are odd harmonics. The filter cutoff frequency to accomplish this goal should be above the 15th harmonic because the level of any harmonic above the 15th, even for the poorest quality line and load, is insignificant. This is why the rule is set at this harmonic of the line frequency. This is easy to do for 50 and 60 Hz, and the cutoff will be well above this goal. The problem is 400 Hz. Actually, the 15th harmonic is not high enough for 400 Hz because of the resonant voltage rise at

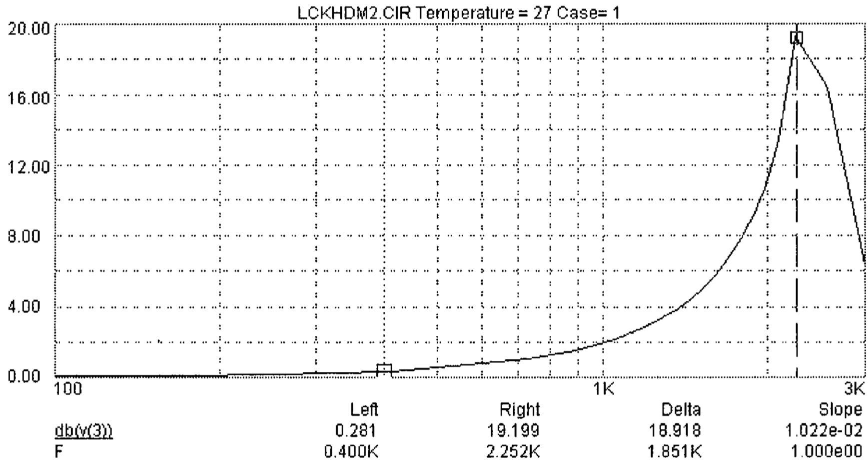

FIG. 15.1 Resonant rise at 2252 Hz: effect at 400 Hz.

400 Hz. Try to get the cutoff as close as possible to 8 kHz. However, this requires too many stages. In the past, the cutoff was formulated at the 10th harmonic, but this did not take 400 Hz into account.

In Fig. 15.1, the cutoff frequency is well above 4000 Hz but has a serious resonant rise at 2252 Hz. This gives a 0.281 dB gain at 400 Hz, which translates to a 3.3% voltage rise. But this is calculated for ideal conditions, and the rise is much more than 3.3% in operation. Note that this condition is not a problem at 50 or 60 Hz. The dB gain is zero. Actually, this filter was designed for 60 Hz, so there would be no problems. There is insufficient power at these 60-Hz harmonics to cause problems in the 2250 area.

15.2. INPUT IMPEDANCE OF THE DIFFERENTIAL MODE FILTER

The equation is simply

$$Z_{\text{in}} \text{ at } N_{\text{H}} = R_{\text{load}} \tag{15.1}$$

where Z_{in} is the input impedance of the filter, N_{H} is the harmonic number and is set equal to 15, and R_{load} is the load impedance at the same frequency. Practically speaking, the load impedance is constant over the frequency range of interest in this discussion.

This goal is hard to reach, especially for the high-current filters with which the required loss is heavy at a low-frequency end. The filter designer should strive to reach it. If the cutoff frequency allows the filter to attenuate the lower harmonics, higher capacitor currents result. These heat the capacitors because of the current through the equivalent series resistance (ESR). The low-frequency cutoff also increases the harmonic currents through the inductors and increases their heat through the inductor DC resistance (DCR) and higher core losses. This raises the operating temperature of the filter. These filters are normally power line type and are often too hot now without this aspect. This low cutoff frequency would lower the resonant rise frequency and raise the circuit Q. Either one of these facts would increase the odds that the filter will oscillate as well as overheat.

15.3. OUTPUT IMPEDANCE OF THE DIFFERENTIAL MODE FILTER

This section follows the same logic as for the input impedance of the filter. The filter should be transparent to the load. If the input impedance goal is met, the output impedance goal is normally met. Meeting these two goals makes for better filter operation.

$$Z_0 \text{ at } N_H = R_s \tag{15.2}$$

where Z_0 is the filter output impedance and R_s is the line impedance at the same frequency. The other terms are the same. The line impedance here is the basic DC resistance of the line. This holds true for most lines to 5 kHz before any rapid increase to higher line impedances is reached. At 10 kHz the impedance is about 4 ohms on most lines.

15.4. INPUT AND OUTPUT IMPEDANCE FOR A DC FILTER

Both requirements of the two preceding sections are easily met for a DC system unless the load is a switcher. Here, the output impedance of the filter must be very low at and above the switcher frequency. This statement rests on the premise that the switcher designer has not anticipated this and corrected for it up front, but it is still something that should be done.

$$Z_0 \text{ at } F_{sw} \ll R_l \text{ at } F_{sw} \tag{15.3}$$

where F_{sw} is the switch frequency and the rest of the terms are the same. The same holds true at the 10th harmonic of the switcher frequency. The switch may not be all that starved at the fundamental and yet be starved at the 9th or 11th harmonic if the output impedance is slightly inductive or the output capacitor is above its self-resonant frequency (SRF). Little power is carried by the higher frequencies above the ninth harmonic, so these may be attenuated.

If this goal is not met, the switch is starved. The output impedance of the filter should be of the order of 0.1 of the load impedance *during conduction or on time*. This is also a function of the pulse width. The main goal is to make sure that the drop is not excessive so that the switcher can function properly. As an example, $F_{SW} = 80$ kHz, $V_s = 28$ V, and $I_{on} = 8$ A during the on time.

The on impedance followed by the required capacitor impedance and then followed by the capacitor value is

$$\frac{28}{8} = 3.5 \qquad \frac{3.5}{10} = 0.35 \qquad C = \frac{1}{2\pi 80{,}000 \times 0.35} = 5.68\ \mu\text{F} \tag{15.4}$$

The output impedance of the filter at 80 kHz to meet the requirement is, from Eq. (15.4), 0.35, and the capacitor value is 5.68 μF. Use a high-quality capacitor close to this value. If the pulse duration is 5 μs and the period, for 80 kHz, is 12.5 with one pulse per cycle, the value of the capacitor can be lowered. In other words, the capacitor must support the pulse for 5 μs. Dividing 5 by 12.5 gives 0.4. Multiplying 0.4 by the capacitor value of 5.68 μF gives 2.272 μF, which is rounded up to 2.5 μF. Again, this must be a good quality wrap and fill, ceramic, and would be better if a feed-through type. Also, this would require a very good quality capacitor with a self-resonant frequency above at least the 10th harmonic of the switcher frequency, 800 kHz in the example.

15.5. COMMON MODE DESIGN GOALS

The common mode does not have to meet any of the requirements discussed for the differential mode. The cutoff frequency can be as low as desired, cutting well into power harmonic frequencies. Watch for leakage inductance in the Zorro inductor(s). This generates a high amount of differential mode inductance and can saturate the inductors at the current peaks. Some state that because the leakage inductance is in air, it cannot saturate. The answer is that the leakage is not all just in air.

The disadvantage is that the common mode grows to very large sizes as the cutoff frequency is lowered, but there is no lower frequency limit for band-pass or other reasons. The real limit is set by the current rating and size of the common mode inductor. These inductors should be designed to have little effect on power factor correction circuits, switchers, or any other load. Again, this assumes little differential mode inductance within the common mode inductor. The typical amount is 1 to 2% and can be of the order of the differential inductance. Common mode inductors are often in the range of 10 MH. At 2%, this equates to 200 μH, which is often more than the differential inductor value. Therefore, the common mode often helps with the differential losses, but the reverse is also true. Some companies do this on purpose by separating the windings and thereby reducing the coupling. The ferrite toroid, in which the common mode windings are each distributed over half the core, have this characteristic.

15.6. ESTIMATE OF COMMON MODE LOAD IMPEDANCE

If the DC system is balanced, using the hot wire and a return wire, common mode inductors can be used without hindering the DC load. This assumes that the differential mode properties are low.

The common mode impedance of the typical nonisolated circuit in Fig. 15.2, using the switcher as the common mode noise source, is as follows: G (the switcher etc.) is now the noise source, and the circuit is drawn backward from that in Fig. 15.2. This keeps the format used throughout this book of having the source, or line, to the left and the load to the right in each drawing.

The storage capacitor is out of the circuit by now (Fig. 15.3), depending on the size and the frequency of interest. The source could be the parasitic oscillation from an inductor at some frequency above 300 kHz. The diodes and the transformer look capacitive, and the switcher inductor remains the same. The current probe measures the noise. The circuit looks something like that shown in Fig. 15.4.

Figure 15.4 shows the four capacitors across the diodes and the output transformer capacitor. If the transformer has a Faraday screen, the final capacitor is even smaller. All of this makes the common mode circuit impedance much higher than the differential mode impedance (Fig. 15.5).

Also, the circuit impedance is greater than the current probe until the upper frequencies are reached. If the circuit lacks the transformer, the diode capacitance to ground is still much higher. Figure 15.6 shows that the common mode design impedance is higher than the differential mode impedance in most applications. This information allows the following technique. Calculate the cutoff frequency as before, and then determine the inductor and capacitor values with the same equations.

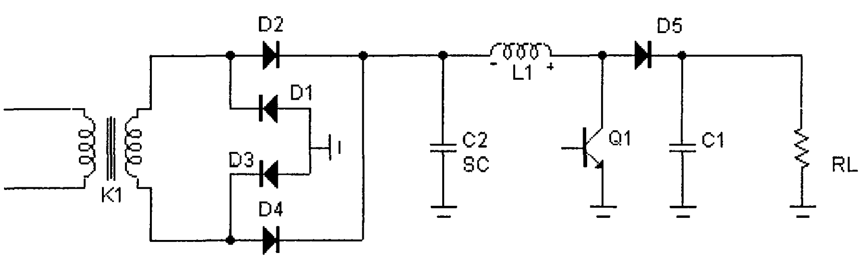

FIG. 15.2 The switcher as the noise source.

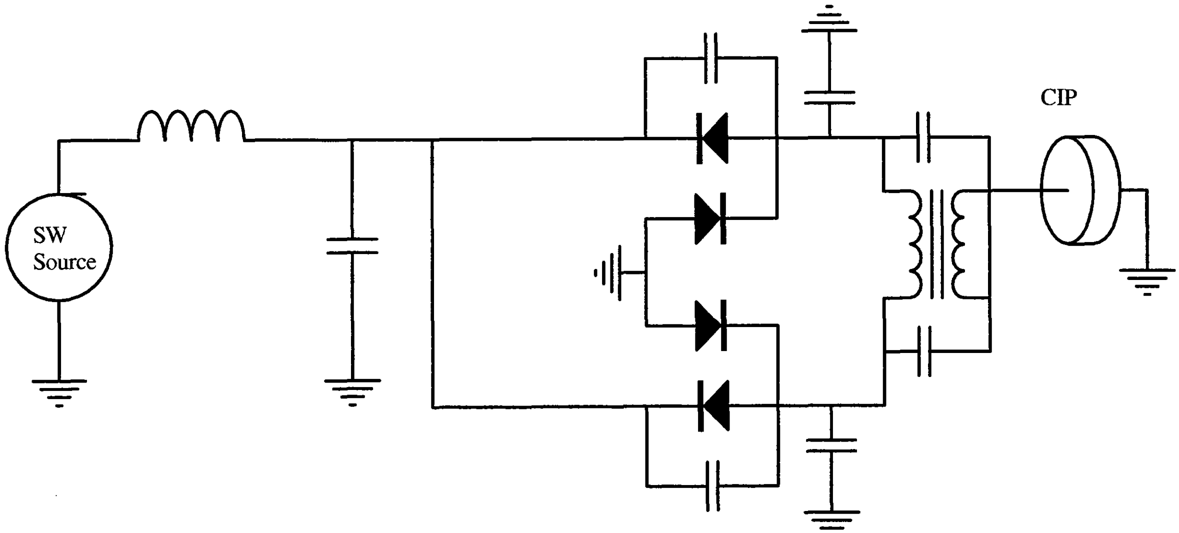

FIG. 15.3 Capacitors around the power supply components.

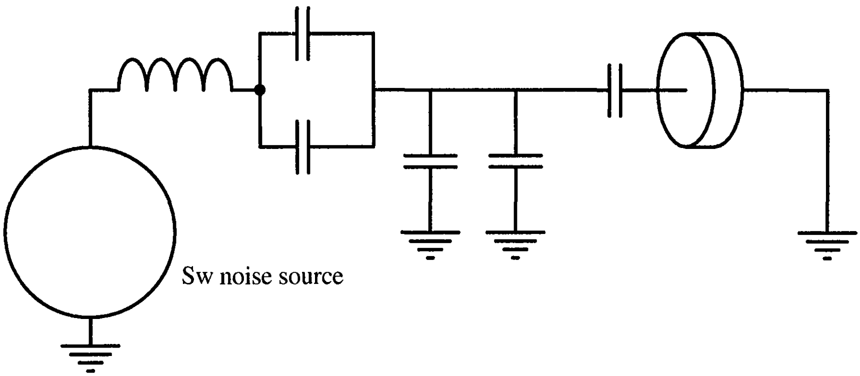

FIG. 15.4 Power supply components reduced to capacitors.

$$L = \frac{R_d}{2\pi F_0} \qquad C = \frac{1}{2\pi F_0 R_d}$$

where R_d is the design impedance, which is the same as for the differential mode, and F_0 is the cutoff frequency. If there is a leakage current specification, calculate the capacitor value of the capacitor to ground, or the specification may state the maximum value of capacitance to ground. Divide the needed value of capacitance by the maximum value, and multiply the inductor by this value. The cutoff frequency remains the same, but the impedance grows by the multiplier.

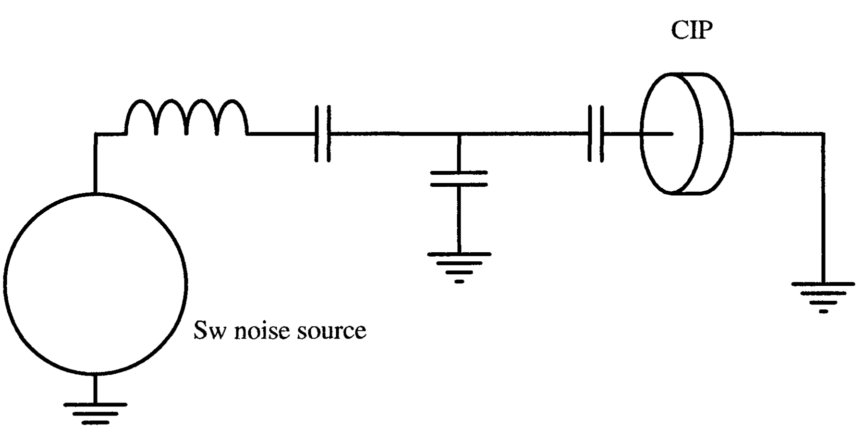

FIG. 15.5 Final switcher reduced to capacitors.

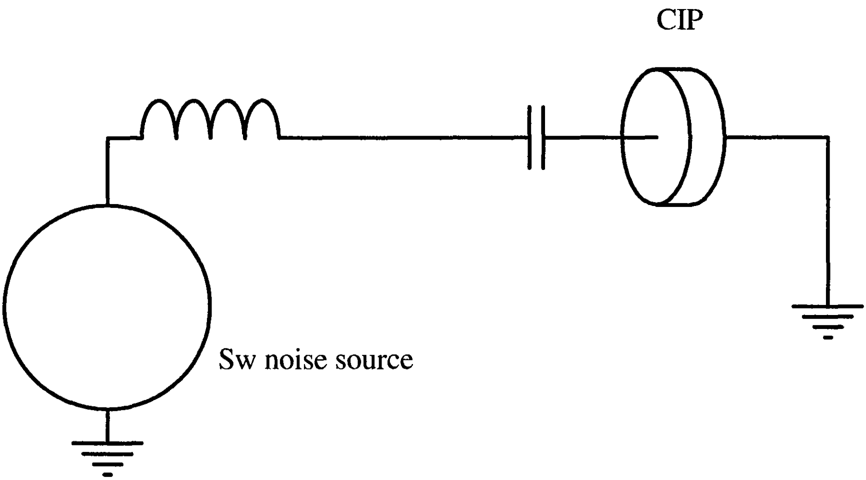

FIG. 15.6 Simplified common mode load impedance.

As an example, the 400 Hz maximum to ground per line is 0.02 μF. In a single-phase circuit, two of these would be in parallel for the common mode application. With the design impedance of 10 ohms and a cutoff frequency of 3350 Hz, the common mode inductor would be 475 μH and the capacitor 4.75 μF using the preceding equations. This is much too big to meet the leakage current specification! An easier way to calculate the capacitor would be to divide the inductance value calculated by the square of the impedance, 10, or 100 here. Divide this value of capacitor by the maximum value of 0.04 μF, the maximum allowed value of the capacitor to ground—the total of 0.02 times the two lines. This gives 118.8, so the inductor is multiplied by this value, which equals 56.4 MH. The value of inductance is excessive at almost any current, and the common mode should be split into two sections. The value of F_0 jumps and makes the two common mode inductors fall to reasonable values of inductance. This reduces the size, weight, and cost and results in a better self-resonant frequency. The maximum capacitance to ground for each line section drops to 0.01, making the two in parallel 0.02 μF. *This technique of changing the values should not be used in differential mode.*

15.7. METHODS OF REDUCING THE SIZE OF THE INDUCTOR DUE TO INDUCTOR CURRENT

Some of these filters may require high current. One way to help the design is to balance the circuit. Here, half of the inductor is in the hot and the other half is in the return. Cutting the value in half drastically improves the ability to design this inductor with a low temperature rise, reasonable flux levels, and, possibly, a reasonable cost. Other approaches with C cores use parallel windings. Each arm, or side, is wound to carry half the current. This reduces the wire size, so the wire is easier to wind. Where the system allows, using larger capacitors lowers the inductance, which also helps to ease the inductor current problem. Make sure the proper ratios are not seriously violated, otherwise a resonant rise will occur at the line frequency.

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16

Matrices—Review of *A* Matrices

Matrices of various forms exist. Besides the A matrix, Z, Y, H, G, B, and now the scatter parameter (S) are used. The advantage of the A, or chain, matrix is that each element of a circuit can be chained, or in tandem, as the elements of the circuit appear, as in the circuit in Fig. 16.1. R_1 and R_3 are series elements, and R_2 and R_4 are shunt elements. Each of these resistors can make up one complete matrix and each matrix can be multiplied, or chained, by the following element. Continue through each following element as they appear in the circuit, such as here.

$$\begin{bmatrix} R_1 \end{bmatrix}\begin{bmatrix} R_2 \end{bmatrix}\begin{bmatrix} R_3 \end{bmatrix}\begin{bmatrix} R_4 \end{bmatrix}$$

Each matrix is formed by four elements, such as

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}$$

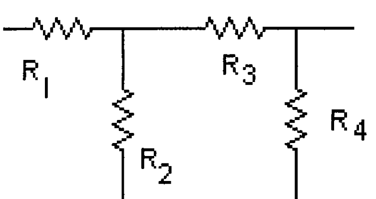

FIG. 16.1 Matrix series and shunt elements.

The series elements, R_1 and R_3, make up the matrix element B:

$$\begin{bmatrix} 1 & R_1 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & R_3 \\ 0 & 1 \end{bmatrix}$$

and the shunt elements make up the matrix element C but are the reciprocals of their values:

$$\begin{bmatrix} 1 & 0 \\ \dfrac{1}{R_2} & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{1}{R_4} & 1 \end{bmatrix}$$

These four matrices make up the following matrix equation:

$$\begin{bmatrix} V_{\text{in}} \\ I_{\text{in}} \end{bmatrix} = \begin{bmatrix} 1 & R_1 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{1}{R_2} & 1 \end{bmatrix} \begin{bmatrix} 1 & R_3 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{1}{R_4} & 1 \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix}$$

The matrix multiplication of the first two matrices and the last two together yields (for matrix multiplication, see Sec. 16.1)

$$\begin{bmatrix} V_{\text{in}} \\ I_{\text{in}} \end{bmatrix} = \begin{bmatrix} 1 + \dfrac{R_1}{R_2} & R_1 \\ \dfrac{1}{R_2} & 1 \end{bmatrix} \begin{bmatrix} 1 + \dfrac{R_3}{R_4} & R_3 \\ \dfrac{1}{R_4} & 1 \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix}$$

If all the Rs are equal, the two 2×2 matrices reduce to

$$\begin{bmatrix} V_{\text{in}} \\ I_{\text{in}} \end{bmatrix} = \begin{bmatrix} 2 & R \\ \dfrac{1}{R} & 1 \end{bmatrix} \begin{bmatrix} 2 & R \\ \dfrac{1}{R} & 1 \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix}$$

and the final matrix multiplication reduces to

$$\begin{bmatrix} V_{\text{in}} \\ I_{\text{in}} \end{bmatrix} = \begin{bmatrix} 5 & 3R \\ \dfrac{3}{R} & 2 \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix}$$

If the load resistor is also R, then I_o is equal to V_o/R and the voltage and current follow.

$$V_{in} = 5V_o + 3\frac{RV_o}{R} = 8V_o$$

$$I_{in} = \frac{3V_o}{R} + \frac{2V_o}{R} = \frac{5V_o}{R}$$

The input impedance of the resistive pad is then $8R/5$ or $1.6R$. The only difference between the preceding text and the text to follow is that the EMI filters use reactive components, adding J factors, or imaginary components to the system.

16.1. CHAIN MATRIX *A*: TRANSFER FUNCTIONS

The advantage of the A matrix is that it transfers the output to the input. If there are five A matrices in tandem, each transfers the quantity from the right to the left. This was the style of matrices used years ago for analog servo systems. A water load was transferred to a pump, the pump transferred the load to the electrical motor, which transferred the load to a control panel and feedback system. The feedback was from the water system. What this system does is to compare the output level to the input level, which is equated to so much dB loss.

16.2. REVIEW OF *A* MATRICES

This section reviews the chain matrix to take advantage of this tool to calculate the insertion loss and other filter properties. These can be chained to form filter modules such as T, π, L, Cauer, dissipative, and other filter elements along with their multiples. The matrix includes the load and source impedances, allowing direct calculation of the insertion loss.

Knowledge of this section should allow easy additions by the reader as needed. All the unit filters, except the dissipative and Cauer filters, can be handled with four elements. These two would also have four elements, but each term would have two terms, one real and one imaginary, for a total of eight. Therefore, all the filter equations must be equally complete so that each matrix solution can use the same form, or template, for each solution. This is like an overlay, and because each matrix fits the overlay, continuous solutions can be calculated. Because any of these units are in the matrix equation form, these elements can be cascaded or chained to form the entire new matrix transfer function. The user can place all of the elements as needed in tandem, and as long as the chain matrix does not run out of room (in terms of the computer program), the matrices can be chained ad infinitum.

The filter designer must know the voltage, current, required insertion loss, operating frequency of the power line, and other information about the load. In these equations our only concern is the filter loss at a known frequency, such as –30 dB at 14 kHz. Often, the required filter loss is specified by an EMI test laboratory after running the proper tests on the unit or system. In other cases, it is just a good educated guess by the equipment or power supply designer using equations given in various chapters earlier. The filter designer's goal is to meet the specified needed loss with some added headroom. From the final matrix of all the combined elements, the loss at a frequency can be calculated with an estimated load and source impedance. The government normally specifies the insertion loss of a filter at 50 ohms load and source impedance in the 220A specification. The designer can insert these values for any load and source needed. In our case, the designer would pick a filter arrangement, such as a double T, and use some program to find the cutoff frequency needed to get the loss. Programs such as GOAL-SEEKER, an old shareware program, or the newer Lotus What If and Backsolver programs can be used. Again, these programs are used to detect the cutoff frequency giving the needed loss with about 6 dB headroom. If the cutoff frequency, F_0, is too low, cutting too deep into the frequency passband (15 times the line frequency is suggested for 400 Hz), another stage must be added (here another T). GOAL-SEEKER, or whatever method you use to iterate between the needed loss and the cutoff frequency, is then used again to find the new higher cutoff frequency to make sure that the filter is still not too close to the passband as before. Usually, the passband should be at least 10 times the line frequency for 400 Hz for higher loss requirements.

All the matrices listed here are 2×2 square matrices giving four elements. As the number of elements or stages grows, the complexity of the terms grows, not the number of elements. Thus, a simple single element has four terms but the terms are easily calculated. A four-element L section also has four terms, but the terms may be algebraically complex. Also, the terms of the Cauer and dissipative filters are made up of both real and imaginary terms. Thus, the matrix may have four elements but can have eight terms. All the layouts should be the same to use the same matrix equation solutions, such as

$$\begin{bmatrix} A & JB & C & JD \\ E & JF & G & JH \end{bmatrix} \tag{16.1}$$

where A, C, E, and G are real and B, D, F, and H are imaginary and where A through H may be quite complex algebraically. The determinant of the chain matrix, called delta (δ), must always equal one, and the imaginary term is reduced to zero. Close counts here, though, as in the imaginary reading of 3.8E-7, are near enough to zero to qualify. Changing from one style of matrix to another leads to other cases creating answers close to one or zero. Some networks have parallel branches, such as the dissipative filter. The two branches were each initially set up using the A matrix and then each was converted to a Y matrix (admittance matrix). These were then added and converted back to the A matrix form. This often creates round-off errors. The δ of the determinant is slightly off from one and leads to an imaginary term different from zero. That is,

$$(A + JB)(G + JH) - (C + JD)(E + JF) = 1 \tag{16.2}$$

This is true for all the matrices in the chain, as well as the resultant matrix, except for the dissipative, Cauer, and other filters because of the matrix conversions, adding some ambiguity to some elements. The order of progression of the matrices is the same as that in which the whole filter types are laid out. The order cannot be changed without arriving at an incorrect answer, but combining them can be worked in any order as long as you are combining or reducing adjacent matrices.

The solution of a two-element column matrix composed of the two matrices as before is

$$\begin{pmatrix} V_{\text{in}} \\ I_{\text{in}} \end{pmatrix} = \begin{pmatrix} A+JB & C+JD \\ E+JF & G+JH \end{pmatrix} \cdot \begin{pmatrix} K+JL & M+JN \\ P+JQ & R+JS \end{pmatrix} \begin{pmatrix} V_{\text{o}} \\ I_{\text{o}} \end{pmatrix} \tag{16.3}$$

$$\begin{pmatrix} V_{\text{in}} \\ I_{\text{in}} \end{pmatrix} = \begin{bmatrix} (A+JB)(K+JL) + (C+JD)(P+JQ) & (A+JB)(M+JN) + (C+JD)(R+JS) \\ (E+JF)(K+JL) + (G+JH)(P+JQ) & (E+JF)(M+JN) + (G+JH)(R+JS) \end{bmatrix} \times \begin{pmatrix} V_{\text{o}} \\ I_{\text{o}} \end{pmatrix} \tag{16.4}$$

The first term (the A, real, and B, imaginary, position) is equal to

$$AK + JAL + JBK - BL + CP + JCQ + JDP - DQ \tag{16.5}$$

These can be separated into real and imaginary components.

$$(AK - BL + CP - DQ) + J(AL + BK + CQ + DP) \tag{16.6}$$

The same follows for the rest of the terms. All the values of A through S reduce either to numbers or to basic algebraic equations. These filters are based on two terms, K and R_{d}, where K is the normalized frequency, so that these can be plotted, and R_{d} is the design impedance. The first element just rendered reduces to a new $A' + JB'$ and so do the rest of the elements. The new combined matrix is

$$\begin{bmatrix} V_{\text{i}} \\ I_{\text{i}} \end{bmatrix} = \begin{bmatrix} A'+JB' & C'+JD' \\ E'+JF' & G'+JH' \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix} \tag{16.7}$$

Again, the determinant of the matrix equals 1, which means that the J terms vanish for those derived purely via A matrix components.

$$\begin{aligned} A'G' - B'H' - C'E' + D'F' &= 1 \\ J * (A'H' + B'G' - C'F' - D'E') &= 0 \end{aligned} \tag{16.8}$$

Note, as stated before, that close counts for the J term, equaling zero on some of the derived matrices because of round-off errors, are created especially when moving between different matrices such as converting to Y matrices—doing some addition—and then converting back to the A matrix form. Also note that $V_o/R_L = I_o$. If this is the final form after multiplying all the matrices, it can be transformed to complex algebra by means of the following:

$$V_i = \frac{V_o(R_l(A' + JB') + C' + JD')}{R_l}$$

$$\frac{V_i}{V_o} = \frac{(R_l(A' + JB') + C' + JD')}{R_l} \tag{16.9}$$

$$\frac{V_i}{V_o} = \frac{R_l A' + C' + J(B'R_l + D')}{R_l}$$

So the voltage ratio

$$\frac{V_o}{V_i} = \frac{R_l}{R_l A' + C' + J(B'R_l + D')} \tag{16.10}$$

Dropping the primes for clarity,

$$\frac{V_o}{V_i} = \frac{R_l}{R_l A + C + J(BR_l + D)}$$

$$\frac{V_o}{V_i} = \frac{R_l}{\sqrt{(AR_l + C)^2 + (BR_l + D)^2}}$$

This voltage ratio is compared with the voltage ratio of the source and load impedances without the filter (Fig. 16.2). The Insertion loss is

$$\frac{R_s + R_l}{R_l} \cdot \frac{R_l}{\sqrt{(AR_l + C)^2 + (BR_l + D)^2}} \tag{16.11}$$

$$\frac{R_s + R_l}{\sqrt{(AR_l + C)^2 + (BR_l + D)^2}} = \text{insertion loss} \tag{16.12}$$

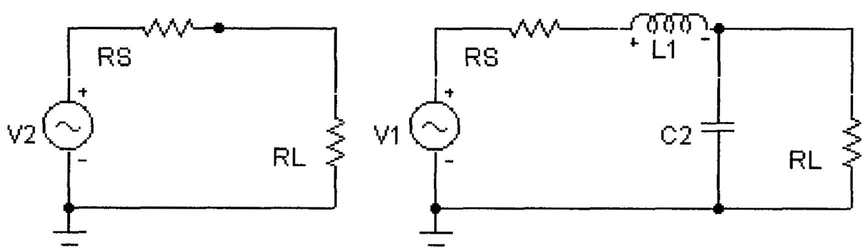

FIG. 16.2 Insertion loss with and without the filter.

where R_s and R_l are the source and load impedances. The dB loss is given by

$$20 \log \left[\frac{R_s + R_l}{\sqrt{(AR_l + C)^2 + (BR_l + D)^2}} \right] \tag{16.13}$$

The angle of lag is

$$\tan^{-1} \frac{BR_l + D}{AR_l + C} \tag{16.14}$$

Engineering manuals and the IEEE literature confirm that there are many reasonably priced programs (others not so reasonable) available that list these components. These plot, list dB losses, give phase angles, etc., but I do not know any that give the values of the components needed for a specific loss. Of course, there are programs for conventional filter design techniques in which known input and output impedances are mandatory. Some use these programs to back-engineer an EMI filter by changing the component values until the goal of so many dB loss at a particular frequency is reached. This and most other similar methods give random values of components that yield the needed loss for the 50 ohm test setup but can do strange things to the passband in the real world. This method can also give a resonant rise at a low harmonic of the line frequency, oscillate, heat the filter, or present such a low input impedance that the circuit breaker may trip on turn on. Also, for 400 Hz, this technique can give a serious voltage rise at 400 Hz. There have been cases in which 120 V in produced 126 V out at 400 Hz.

If the inductors and capacitors are pure, the series element of inductive reactance is placed in $C + JD$ and any shunt or parallel element's reciprocal impedance is placed in $E + JF$. In this case, being pure, the C and E terms are zero. If the inductance has significant resistance, or DC resistance (DCR), this resistance value is placed in term C. If the shunt capacitor has added resistance, as with an RC shunt, It is 1 over $R + JX_c$. This gives the following for the E and F terms:

$$\frac{R}{\left[R^2 + (X_c)^2\right]} + \frac{JX_c}{\left[R^2 + (X_c)^2\right]} \tag{16.15}$$

A simple L formed with a series inductor followed by a parallel, or shunt, capacitor would form a small chain matrix such as this. Note that C and E are both 0 because their small values of equivalent series resistance (ESR) and DCR are neglected—or treated as pure—and the inductor faces the line while the capacitor faces the load. But the system is missing the zeros or the unused terms are gone This deletes the following terms: B, C, F, and H in the inductor section and B, D, E, and H in the capacitor section.

$$\begin{pmatrix} 1 & JX_l \\ 0 & 1 \end{pmatrix} \cdot \begin{bmatrix} 1 & 0 \\ \dfrac{J}{X_c} & 1 \end{bmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.16}$$

The new matrix becomes

$$\begin{bmatrix} 1 - \dfrac{X_l}{X_c} & JX_l \\ \dfrac{J}{X_c} & 1 \end{bmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.17}$$

This might look bad, but the first term reduces to $1 - K^2$ as follows. The inductance and capacitance equations are

$$L = \frac{R_d}{2\pi F_0} \qquad C = \frac{1}{2\pi F_0 R_d} \tag{16.18}$$

The inductive reactance using Eq. (16.18) is

$$X_l = 2\pi F_l = \frac{2\pi F R_d}{2\pi F_0} = \frac{F R_d}{F_0} \tag{16.19}$$

The capacitive reactance, also using Eq. (16.18), is

$$X_c = \frac{2\pi F_0 R_d}{2\pi F} = \frac{F_0 R_d}{F} \tag{16.20}$$

But

$$F = K F_0 \tag{16.21}$$

The inductive and capacitive reactance becomes

$$X_l = K R_d \qquad X_c = \frac{R_d}{K} \tag{16.22}$$

where K is the normalized frequency, R_d is the design impedance, and F_0 is the cutoff frequency. The matrix of (16.17) is

$$\begin{bmatrix} 1 - K^2 & JKR_d \\ \dfrac{JK}{R_d} & 1 \end{bmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.23}$$

Note that terms 1 and 4 are real, resistive, and 2 and 3 are imaginary. Also note that the preceding matrix yields a delta value of 1. What happens if the order is changed? This places the shunt capacitor on the line side and the series inductor facing the load.

$$\begin{bmatrix} 1 & 0 \\ \dfrac{J}{X_c} & 1 \end{bmatrix} \begin{pmatrix} 1 & JX_l \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.24}$$

Substituting these values for X_l and X_c:

$$\begin{bmatrix} 1 & 0 \\ \dfrac{JK}{R_d} & 1 \end{bmatrix} \begin{pmatrix} 1 & JKR_d \\ 0 & 1 \end{pmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.25}$$

Multiply the two matrices:

$$\begin{bmatrix} 1 & JKR_d \\ \dfrac{JK}{R_d} & 1 - K^2 \end{bmatrix} \begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.26}$$

Note the difference in the two answers, showing what happens when elements are processed in the wrong order. Compare (16.22) and (16.26) (Fig. 16.3). Either answer is correct, depending on what element the designer wants facing the line or the load. Most EMI filter designers want the inductor on the line side,

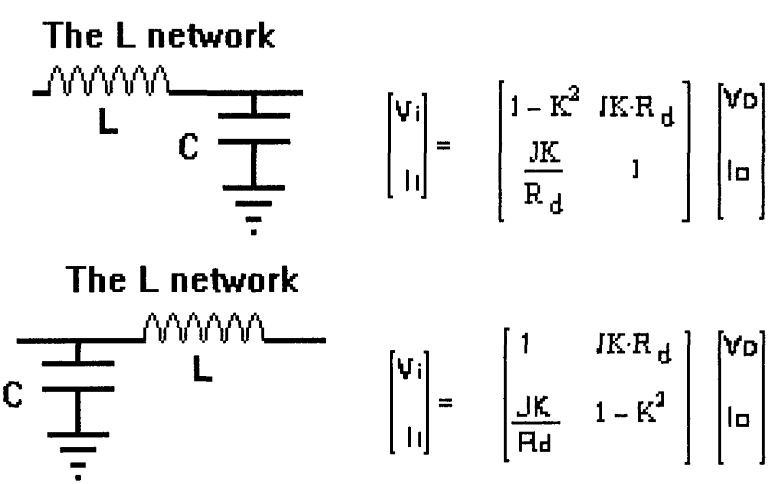

FIG. 16.3 Comparison of the LC and CL filter matrices.

especially for 461 specifications or any specification requiring high insertion losses at frequencies near 14 kHz.

A multiple L of the type in Eq. (16.23) with the inductor on the line side can be formed by

$$\begin{bmatrix} 1-K^2 & JKR_d \\ \dfrac{JK}{R_d} & 1 \end{bmatrix}\begin{bmatrix} 1-K^2 & JKR_d \\ \dfrac{JK}{R_d} & 1 \end{bmatrix}\begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.27}$$

This reduces to

$$\begin{bmatrix} 1-2K^2+K^4+J^2K^2 & 2JKR_d - JK^3R_d \\ -JK\dfrac{(-2+K^2)}{R_d} & J^2K^2+1 \end{bmatrix}\begin{pmatrix} V_0 \\ I_o \end{pmatrix} \tag{16.28}$$

Substituting for J squared (–1) gives

$$\begin{bmatrix} 1-3K^2+K^4 & 2JKR_d - JK^3R_d \\ -JK\dfrac{(-2+K^2)}{R_d} & -K^2+1 \end{bmatrix}\begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.29}$$

The delta of the matrix is 1. If the cutoff frequency, F_0, is 4000 and the frequency of interest is 400, then K is equal to 400/4000, which is 0.1, and if R_d is 20, the values in Eq. (16.29) would be

$$\begin{pmatrix} 0.9701 & 3.98\,\mathrm{J} \\ 9.95\times10^{-3}J & 0.99 \end{pmatrix}\begin{pmatrix} V_o \\ I_o \end{pmatrix} \tag{16.30}$$

If the load resistance is also 20, then $I_o = V_o/R_L$ and V_i is equal to

$$0.9701V_o + 0.199\,JV_o$$

and I_i is equal to

$$9.95 \times 10^{-3}J\,V_o + 4.95 \times 10^{-2}V_o$$

Then Z_{in} is equal to

$$19.614 + 7.7672 \times 10^{-2}J$$

Z_{in} equals 19.614 at 13.613 minutes.

The value of the impedance is not too far from the expected 20 and there is, in essence, no phase angle.

Various values of K can be plotted, or any one frequency can be quickly checked. The main advantage is that the filter element values are given for the required insertion loss. The various filter sections needed are listed for each term of the matrix, and the method for designing other units is included in the next chapter. All the equations are listed to follow the format of (16.7) without the primes.

$$\begin{bmatrix} V_{\text{in}} \\ I_{\text{in}} \end{bmatrix} = \begin{bmatrix} A & JB & C & JD \\ E & JF & G & JH \end{bmatrix} \begin{bmatrix} V_{\text{o}} \\ I_{\text{o}} \end{bmatrix} \tag{16.31}$$

17

The Filter Design Technique

Here we discuss how to build the matrices for each filter element in the following list. The names can be changed if the designer wishes to build these filters into a computer program. These names are only those I used initially in my programming years ago with Lotus.

1. Unit matrix
2. R_S matrix
3. LINESIM matrix
4. LISN matrix
5. DIN and DOUT matrices
6. RCSHU matrix
7. LSER matrix and CSHU matrix
8. The L
9. Pi filter
10. T filter
11. Cauer matrix

17.1. THE UNIT MATRIX

The simplest matrix is the unit matrix. This is the equivalent of scalar multiplication by one. The unit matrix requires no input. Make sure that all the unused elements of the matrix field have unit elements in them and not other filter elements. These other elements in these sections give erroneous higher losses, making the filter appear to have the required specified loss but be well out of the required specification when it is built and tested.

The values are as follows. Unit matrix:

$$\begin{vmatrix} 1+J0 & 0+J0 \\ 0+J0 & 1+J0 \end{vmatrix} \tag{17.1}$$

In reality, the unit matrix is a square diagonal matrix as follows.

$$\begin{vmatrix} 1 & 0 \\ 0 & 1 \end{vmatrix}$$

The off-diagonals are all zero, and again, this is like multiplying the matrix field by one. In the beginning, the Lotus program had seven matrix multipliers. The first element was the R_S matrix, followed by the rest of the elements, and the rest of the multipliers were filled with the unit matrix. For example, here seven multipliers started with three L matrices, R_S, three Ls, and four units in this order. The cutoff frequency and component values were calculated from the bottom. If the cutoff frequency was too low, one of the units was replaced with another L.

17.2. THE R_S MATRIX

The value needed comes from the "Common Terms" section and is used for the input-side line impedance R_S (Fig. 17.1). This term is the lower frequency line resistance or source impedance. In the 220A tests, this would be 50 ohms. This is fine for the frequencies under 100 kHz, where skin effect has not yet taken over and the filter is being designed for a real-world insertion loss around 14 kHz.

Most of the various groups who have tested power lines have agreed that

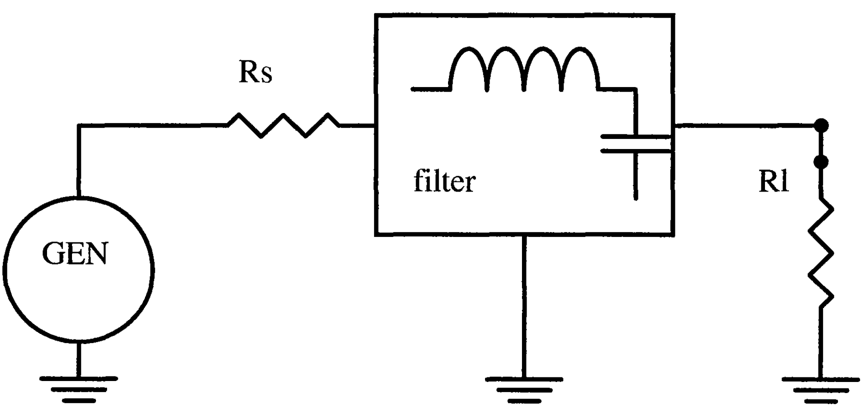

FIG. 17.1 The R_S element.

the line impedance is around 4 ohms at 10 kHz. Here, the value of 4 ohms is assigned.

$$\begin{vmatrix} 1+J0 & 4+J0 \\ 0+J0 & 1+J0 \end{vmatrix} \tag{17.2}$$

which is reduced to the following (but all the matrices must have the eight terms so that they can overlay the equation format and the solutions may be repeated whatever the matrix placed at the multiplication spot):

$$\begin{vmatrix} 1 & 4 \\ 0 & 1 \end{vmatrix} \tag{17.3}$$

No further information is needed for this unit. This is part of the line, not the filter.

17.3. THE LINESIM MATRIX

This matrix is used to match a known line or the output impedance of a DC power supply (Fig. 17.2). The small series resistor is the value of R_s from the "Common Terms," but the values of the shunt resistor R_{hf} and the series inductor L must be given. R_s, K, and F_0 should be available from the "Common Terms" area and are needed here. R_{hf}, the high shunt resistance, and L must be known. R_s acts as the low-value series impedance of the source. A different symbol can be used. This value must be provided if R_s is not used.

The frequency of interest is

$$F = KF_0 \tag{17.4}$$

This makes X_l equal to

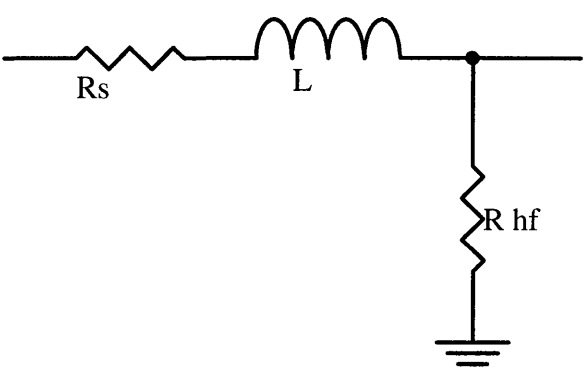

FIG. 17.2 The LINESIM network.

$$X_l = 2\pi FL = 2\pi KF_0L$$

R_s, R_{hf}, and L_1 are given so that values can be inserted directly into the matrix. This is so that the line simulation network impedance follows the normalized frequency. Thus, as K varies, the inductive reactance of this network varies along with the filter losses.

The LINESIM is

$$\begin{vmatrix} V_i \\ I_i \end{vmatrix} = \begin{vmatrix} \dfrac{R_{hf}+R_s}{R_{hf}} + \dfrac{JS_l}{R_{hf}} & R_s + JX_l \\ \dfrac{1}{R_{hf}} + J0 & 1 + J0 \end{vmatrix} \begin{vmatrix} V_o \\ I_o \end{vmatrix} \tag{17.5}$$

As an example, F_0 is 4000, R_s is 4, R_{hf} is 50, L is 90 µH, and K floats, so it can be plotted. $X_l = 2.262K$, so the matrix becomes

$$\begin{vmatrix} 1.08 + 0.0452JK & 4 + 2.262JK \\ 0.02 + 0J & 1 + 0J \end{vmatrix} \tag{17.6}$$

When this matrix is multiplied by the rest in the series, the various values of K can be plotted. The LINESIM network is not part of the filter and accounts for the line impedance.

17.4. THE LISN MATRIX

The input section, formed by R_{in} and C_{in}, is in series shunt across the line. The midsection is made up with one inductor in series, L_1. The output section, with R_o and C_o in series shunt, is also across the line. All these values must be known and should be provided by the manufacturer. The initial values are from a standard LISN but can be changed. The default values in any of these sections can always be used. The engineer can input the main values LISN used repeatedly and save these values so that this information need not be reentered each time this LISN is used. If more than one LISN is used, several LISNs can be added by copying the equations into each new section and changing the known values. These would be called LISN1, LISN2, and so forth.

These units are used by the EMI test laboratories and are often required for the various test specifications (Fig. 17.3). If the test specification calls for a LISN, this matrix can be handy for the design of the EMI filter. Here, the LISN is used as the input feed to replace the R_s matrix. These are also used to evaluate the full system to determine whether the system meets the specification. These units (Fig. 17.4) can be purchased from companies such as Solar Electronics, Hollywood, CA 90038 (Attn. Al Parker).

The LISN matrix is the hardest to format and is formed by the following multiplication of the three matrices.

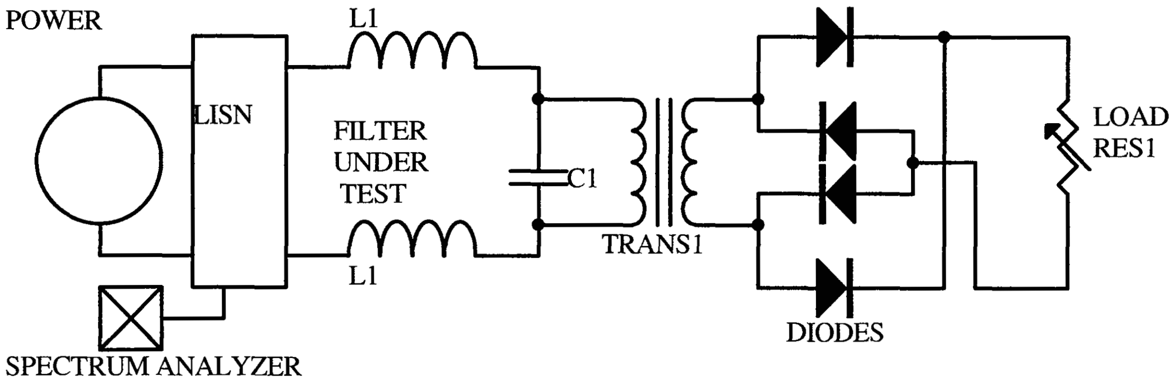

FIG. 17.3 The LISN used for testing.

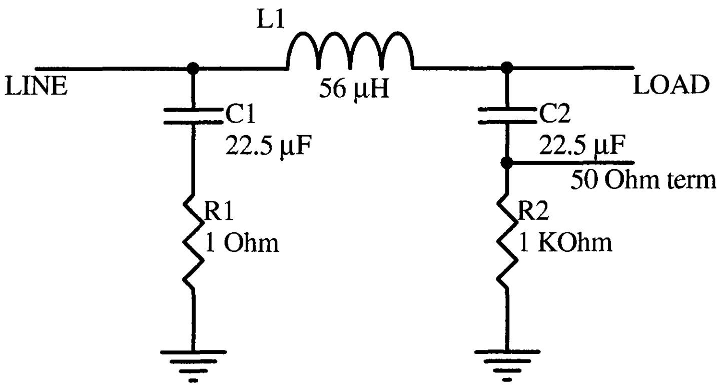

FIG. 17.4 The common LISN.

$$\begin{vmatrix} 1 & 0 \\ \dfrac{R_1+JX_{c_1}}{R_1^2+X_{c_1}^2} & 1 \end{vmatrix} \begin{vmatrix} 1 & JX_{1_1} \\ 0 & 1 \end{vmatrix} \begin{vmatrix} 1 & 0 \\ \dfrac{R_2+JX_{c_2}}{R_2^2+X_{c_2}^2} & 1 \end{vmatrix} \tag{17.7}$$

Solving the first two,

$$\begin{vmatrix} 1 & JX_{1_1} \\ \dfrac{R_1+JX_{c_1}}{R_1^2+X_{c_1}^2} & 1+\dfrac{JX_{1_1}(R_1+JX_{c_1})}{R_1^2+X_{c_1}^2} \end{vmatrix} \begin{vmatrix} 1 & 0 \\ \dfrac{R_2+JX_{c_2}}{R_2^2+X_{c_2}^2} & 1 \end{vmatrix} \tag{17.8}$$

The full solution is

$$\begin{vmatrix} 1+JX_{1_1}\dfrac{R_2+JX_{c_2}}{R_2^2+X_{c_2}^2} & JX_{1_1} \\ \dfrac{R_1+JX_{c_1}}{R_1^2+X_{c_2}^2}+\dfrac{R_2+JX_{c_2}}{R_2^2+X_{c_2}^2}+\dfrac{JX_{1_1}(R_1+JX_{c_1})(R_2+JX_{c_2})}{(R_1^2+X_{c_1}^2)(R_2^2+X_{c_2}^2)} & 1+\dfrac{JX_{1_1}(R_1+JX_{c_1})}{R_1^2+X_{c_1}^2} \end{vmatrix} \tag{17.9}$$

Typical LISN values with F_0 at 4000 are given in Table 17.1.

$$X_{C_1} = \frac{1}{2\pi K F_0 C_1} = \frac{1.768}{K}$$

$$X_1 = 2\pi F_0 L_1 = 1.407K$$

$$X_{C_2} = \frac{1}{2\pi K F_0 C_2} = \frac{1.768}{K}$$

TABLE 17.1 Typical LISN Values with F_0 at 4000

Line section	Midsection	Load section
R_1 = 1 ohm	L_1 = 56 μH	R_2 = 50 ohms
C_1 = 22.5 μf		C_2 = 22.5 μF

Using the standard outline of the 2 × 2 matrix with eight terms, the equations of the individual terms follow.

$$\begin{vmatrix} A+JB & C+JD \\ E+JF & G+JH \end{vmatrix}$$

The terms are

$$A = 1 - \frac{X_{L_1}X_{C_2}}{R_2^2 + X_{C_1}} = 1 - \frac{1.407 \times 1.768}{50^2 + 1.768^2 / K^2} = 1 - \frac{1.497 \times 1.768K^2}{50^2K^2 + 1.768^2}$$

$$B = \frac{JR_2X_{L_1}}{R_2^2 + X_{C_2}^2} = \frac{J50 \times 1.407 \times K}{50^2 + 1.768^2 / K^2} = \frac{J \times 50 \times 1.407 \times K^3}{50^2K^2 + 1.768^2}$$

$$C = 0 \tag{17.10}$$

$$D = JX_{l_1} = J1.407K$$

$$E = \frac{R_1R_2(R_1 + R_2) + R_1X_{C_2}^2 + R_2X_{C_1}^2 - X_{L_1}(X_{C_2}R_2 + X_{C_1}R_1)}{(R_1^2 + X_{C_1}^2)(R_2 + X_{C_2}^2)}$$

The two capacitors are often the same values as in Fig. 17.5. This reduces to

$$E = \frac{(R_1 + R_2)\,|R_1R_2 + X_{c_2}^2 - X_{1_1}X_{c_2}|}{(R_1^2 + X_{c_2}^2)(R_2^2 + X_{c_0}^2)}$$

where R_1 and R_{in} are the same, which also follows for C_1 and C_{in}, R_2 and R_{out} and C_2 and C_{out}. These values are interchangeable as in the LISN in Fig. 17.5. From the preceding values,

$$E = \frac{51K^2(47.51K^2 + 1.768^2)}{(K^2 + 1.768^2)(50^2K^2 + 1.768^2)}$$

$$F = J\frac{X_{C_1}(R_2^2 + X_{C_2}^2 - X_LX_{C_2}) + X_{C_2}(R_1^2 + X_{C_1}^2) + R_1R_2X_L}{(R_1^2 + X_{C_1}^2)(R_2^2 + X_{C_2}^2)}$$

This reduces to

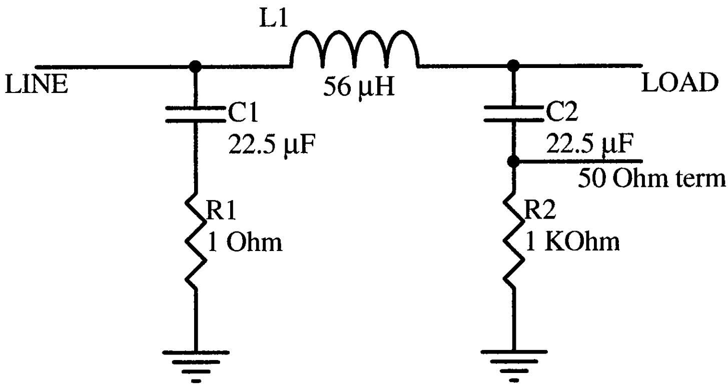

FIG. 17.5 The common LISN with equal capacitors.

$$F = J\frac{|70.35K^5 + 4417.34K^3 + 11.053K|}{(K^2 + 3.126)(2500K^2 + 3.126)}$$

$$G = \frac{R_1^2 - X_{L_1}X_{C_1} + X_{C_1}^2}{R_1^2 + X_{C_1}^2}$$

$$G = \frac{3.126 - 1.4876K^2}{K^2 + 3.126}$$

$$H = \frac{JR_1X_{L_1}}{R_1^2 + X_{C_1}^2} = \frac{J \times 1.407K^2}{K^2 + 3.126}$$

In the computer program, these are already provided, but the component values must be entered if they are different from these default values. This shows where they come from.

Finally, the component values are entered. These are (A through H) based on the following simpler equations:

$$P = 2\pi F_0 K$$

$$A = \frac{P^2C_2^2R_2^2 - P^2C_2L_1 + 1}{1 + P^2C_2^2R_2^2}$$

$$B = \frac{JP^3C_2^2L_1R_2}{1 + P^2C_2^2R_2}$$

$$C = 0$$

$$D = JPL$$

$$E = P^2\frac{R_1C_1[1 + P^2C_0(C_0R_0 - L)] + R_0C_0^2[1 + P^2C_1(C_1(C_1R_1^2 - L)]}{(1 + P^2C_1^2R_1^2)(1 + P^2C_0^2R_0^2)}$$

$$F = JP\frac{C_1(1 + P^2C_0^2R_0^2) + C_0(1 + P^2C_1^2R^2) + P^2C_1C_0L(P^2R_1R_0C_1C_0 - 1)}{(1 + P^2C_0^2R_0^2)(1 + P^2C_1^2R^2)}$$

$$G = \frac{1 + P^2C_1(C_1R_1^2 - L)}{1 + P^2C_1R^2}$$

$$H = J\frac{|P^3C_1^2R_1L_1|}{(1 + P^2C_1^2R_1)}$$

With this, the only variable is P, and this term varies with K. Again, this allows all these equations to be plotted as a function of K. K is the normalized frequency, which has the value one at the cutoff frequency.

17.5. THE D_{IN} AND D_{OUT} MATRICES

This is the dissipative filter and gives a loss of 6 dB per octave. D_{IN} faces the line and D_{OUT} faces the load if both filters are used, one at each end. Several terms must be specified, including R_{in} and R_{out}. The value of K is the same as K in the other filters. R_{in} in Fig. 17.6 (bottom), should be the wanted filter input impedance, such as 50 ohms, for the D_{IN} filter. R_l, the output impedance needed, should match the load for the D_{OUT}, which can also be 50 ohms. This would be a nice feature for the 220A specification. Here, the term is R_{out} but R_4 is used in Fig. 17.6 (bottom). The opposite two resistors, R_{out} (R_2 in Fig. 17.6, bottom) for D_{IN} and R_{in} (R_3 in Fig. 17.6, bottom) for D_{OUT}, can be used to match any filter wanted—Butterworth, π, or other—and these can also be, here, designed for 50 ohms.

The delta is not always equal to one because of the conversion back and forth between the Y and A matrices. This leads to minor errors, or round-off errors, within the terms. The values of K and F_0 could come from the "Common Term" area and the values of R_d, R_{in}, and R_o should be listed with this matrix. Both D_{IN} and D_{OUT} are the same, but R_{in} and R_o should switch values.

As F_0 decreases in value, both L and C increase in value. If the line frequency is 400 Hz, the central filter cutoff should be 4000 Hz and above, say 4000 Hz. Then if the K ratio is 0.5, or making its cutoff 8000 Hz and still having the full filter give a good impedance match at 10,000 or 14,000 Hz, say 50 ohms, the basic D_{IN} and D_{OUT} equations are

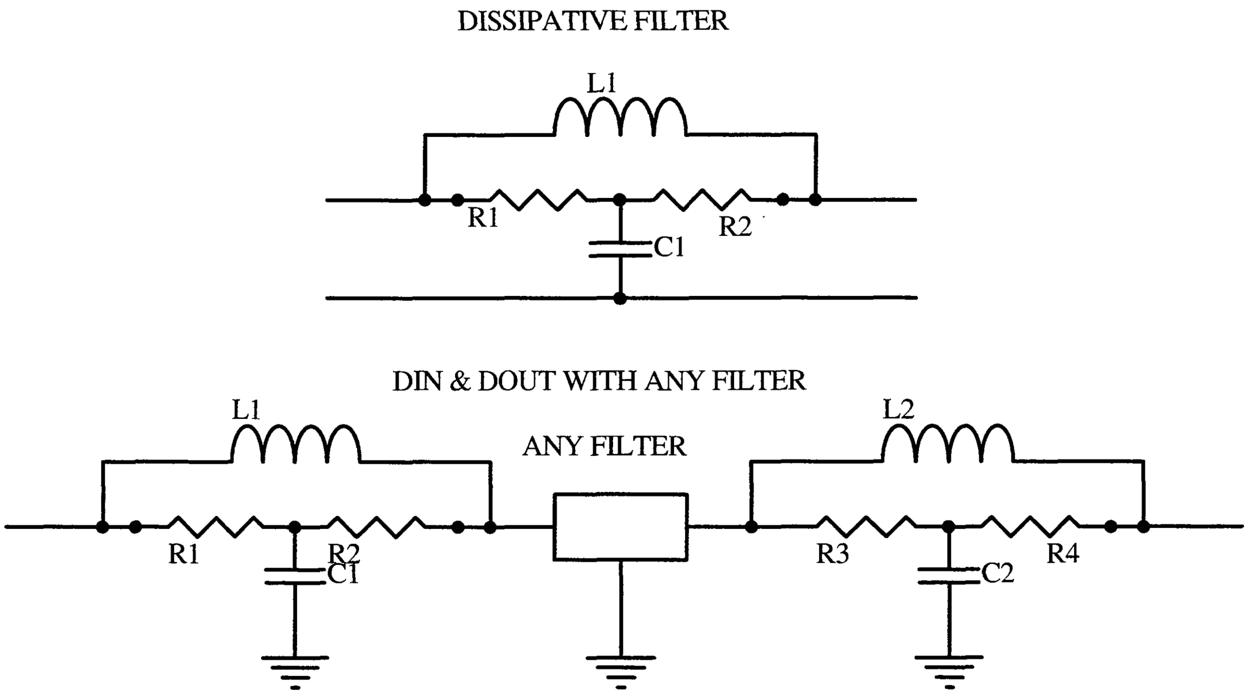

FIG. 17.6 The D_{IN} and D_{OUT} with any filter.

$$K_d = KK_{ratio} \quad R_d = \sqrt{R_{in}R_o}$$
$$L = \frac{R_d K_{ratio}}{2\pi F_0} \quad C = \frac{K_{ratio}}{2\pi F_0 R_d} \tag{17.11}$$

where K_d is the equivalent of the value of K in the other filters, R_d is the design impedance for the D_{IN} and D_{OUT} only, and F_0 is the cutoff frequency for all the filters, including the D_{IN} and D_{OUT}.

The format of the D_{IN} or D_{OUT} matrix follows the same as that of the others given here:

$$\begin{vmatrix} A+JB & C+JD \\ E+JF & G+JH \end{vmatrix} \tag{17.12}$$

If the design calls for all the Rs to be 50 ohms and K_d is equal to 1, "K ratio" = 1, the various elements, A through H, equal the end values given after each element:

$$\begin{aligned}
A &= \frac{4K_d^2R_d^2 - K_d^2R_{in}(R_{in}+R_o) + (R_{in}+R_o)^2}{K_d^2R_d^2 + (R_{in}+R_o)^2} \\
A &= 0.75 \\
B &= J\frac{2K_d^3R_dR_{in}}{4K_d^2R_d^2 + (R_{in}+R_o)^2} \\
B &= 0.25J \\
C &= \frac{K_d^2R_d^2(R_{in}+R_o)}{4K_d^2R_d^2 + (R_{in}+R_o)^2} \\
C &= 12.5 \\
D &= J\frac{2K_d^3R_d^3 + K_dR_d(R_{in}+R_o)}{4K_d^2R_d^2 + (R_{in}+R_o)^2} \\
D &= 37.5J \\
E &= \frac{K_d^2(R_{in}+R_o)}{4K_d^2R_d^2 + (R_{in}+R_o)^2} \\
E &= 0.005 \\
F &= J\frac{2K_d^3R_d^3 + K_d(R_{in}+R_o)^2}{R_d(4K_d^2R_d^2 + (R_{in}+R_o)^2)} \\
F &= 0.015J \\
G &= \frac{4k_d^2R_d^2 - K_d^2R_o(R_{in}+R_o) + (R_{in}+R_o)^2}{4K_d^2R_d^2 + (R_{in}+R_o)^2}
\end{aligned} \tag{17.13}$$

$$G = 0.75$$

$$H = J\frac{2K_d^3R_dR_o}{4K_d^2R_d^2 + (R_{in} + R_o)^2}$$

$$H = 0.25J$$

Typical values are $F_0 = 4000$, K ratio = 1, $R_{in} = 50$, $R_o = 50$. Then $K_d = K$, $R_{in} + R_o = 100$, and

$$\begin{vmatrix} 0.75 + J\,0.25 & 12.5 + J\,37.7 \\ 0.005 + J\,0.015 & 0.75 + J\,0.25 \end{vmatrix} \tag{17.14}$$

as shown by the preceding solutions of A through H. Note that the delta is equal to 1. These are the values at 4000 Hz.

D_{IN} and D_{OUT} are the same but R_{in} and R_o, if the two values are different, exchange values so that R_{in} of D_{IN} faces the line and R_o of D_{OUT} faces the load. The K ratio is usually 1 but must be the same if both D_{IN} and D_{OUT} are used. This means that two matrices are required: one matrix for D_{IN} and one matrix for D_{OUT}.

17.6. THE RCSHU MATRIX

This matrix is used to lower the Q of the filter and correct for a resonant rise and/or a problem related to frequency, such as insufficient loss at a particular frequency. The design method is to calculate the needed capacitor value to equal the design impedance at the problem frequency and then place this combination in series across the line. These are normally mounted inboard between two other filter sections, such as between two Ls, and would be tied across the first L's capacitor. Often, this filter section can cure several problems at once. If, for example, a three-stage L is needed to obtain the required insertion loss at 10 kHz, two resonant rises appear in the graph. As the load impedance deviates from the design impedance, the peak value of each resonant rise changes. Also, the loss near 10 kHz may be at, or somewhat over, the insertion loss limit. Determine the lower resonant rise frequency and tie the resultant network across either the first or the second L's capacitor. The result should be that the first resonant rise is reduced and the second resonant rise is gone. The loss around 10 kHz, assuming that the first resonant rise is well below 10 kHz, is. Circuit Q is reduced to impede any oscillations by the addition of an RC shunt (Figs. 17.7 and 17.8). This is easier to design than the Cauer and is automatically balanced.

The matrix is simple to form: R_c is the series resistance and should be equal to R_d, the design impedance from the "Common values to all filters," if any such area exists, or listed here. C is calculated as before, and $X_c = R_d$ at the problem frequency. The proper values are placed in the $E + JF$ area of the matrix

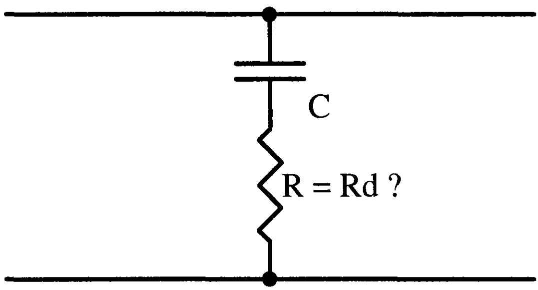

FIG. 17.7 The RC shunt.

as listed here. Once the value of the capacitor is known, the capacitive reactance for the matrix must be a function of K, so the matrix varies with frequency and can be plotted.

$$X_c = \frac{1}{2\pi F_0 KC} \tag{17.15}$$

$$E = \frac{R_c}{R_c^2 + X_c^2} \qquad F = \frac{JX_c}{R_c^2 + X_c^2}$$

RCSHU:

$$\begin{vmatrix} 1 + J0 & 0 + J0 \\ E + F & 1 + J0 \end{vmatrix} \tag{17.16}$$

An RCSHU example is as follows: A triple L was designed with two resonant rises. The first fell near 3800 Hz and the second near 4600 Hz, and the design impedance was 10 ohms. This filter was also out of limits at 14 kHz. Design the capacitor impedance to equal the 10 ohms at 3800 Hz for the first, lowest, resonant rise. This gives

$$C = \frac{1}{2\pi 3800 \times 10} = 4.188\ \mu F = 4.2\ \mu F$$

The self-resonant frequency (SRF) of this capacitor need not be as good as that of the rest of the filter capacitors unless there are also higher frequency problems. This is being used primarily to solve some low-frequency problems, and the highest frequency is 14 kHz in the case just stated. If the SFR is better than the fifth harmonic, 70 kHz here, of this upper frequency, the job will get done. This statement is being made so that the filter designer does not select a capacitor that costs many times more than one that will do the job. Note that X_c is a function of K and not single valued as shown before. Tie the 10 ohm resistor and the 4.2 μF in series across one of the inbound capacitors that make up the first or second

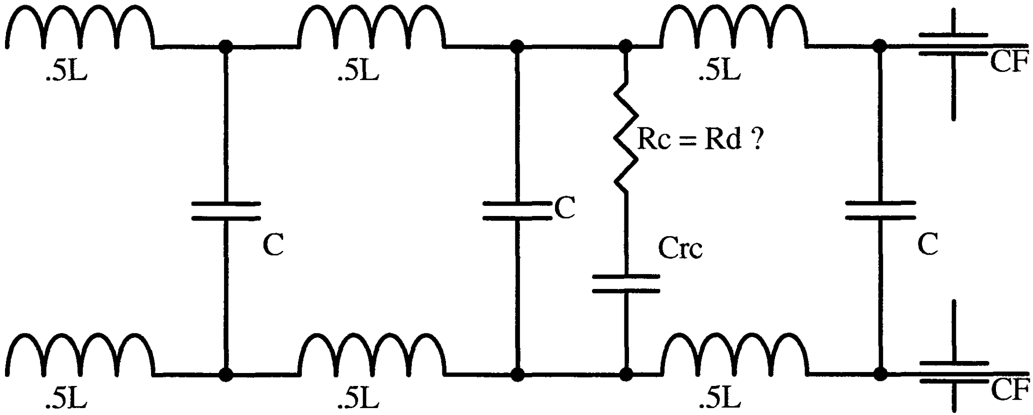

FIG. 17.8 Multiple L with RC shunt.

L counting from the load end of the filter. The closer to the load, the better this network works, because the load is the higher impedance.

17.7. THE SERIES INDUCTOR, LSER, AND THE SHUNT CAPACITOR, CSHU

These should be understood from the various earlier descriptions but are reviewed here anyway. There are many uses for each of these, and in some locations they can be used together. For example, these two together can make up Ls, Ts, πs, and so forth that are not functions of R_d. CSHU, alone, can be used to duplicate the value of a feed-through capacitor. The LSER, the series inductor, can be the common mode inductor. LSER and CSHU can work together to form part of the common mode filter (Fig. 17.9).

The LSER is

$$X_l = 2\pi F_0 KL \tag{17.17}$$

It is obvious that the value of X_l is a function of K and is placed in the matrix at D.
LSER:

$$\begin{vmatrix} 1+J0 & 0+JX_l \\ 0+J0 & 1+J0 \end{vmatrix} \tag{17.18}$$

CSHU:

$$X_c = \frac{1}{2\pi F_0 KC} \tag{17.19}$$

This is the same as RCSHU except that R_c is zero, making the E value zero and F the reciprocal of X_c.

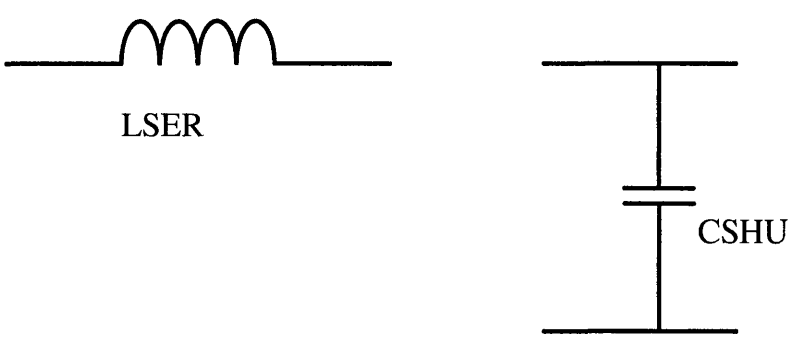

FIG. 17.9 The LSER and CSHU.

CSHU:

$$X_c = \frac{1}{2\pi F_0 KC} \tag{17.20}$$

17.8. THE L MATRIX

This is a review of Sec. 11.2 showing the origin of the values that make up the L filter, where K is the normalized frequency and R_d is the design impedance (Fig. 17.10).

$$F = KF_0$$

$$L = \frac{R_d}{2\pi F_0}$$

$$X_l = \frac{2\pi F R_d}{2\pi F_0} = KR_d$$

$$C = \frac{1}{2\pi F_0 R_d}$$

$$X_c = \frac{2\pi F_0 R_d}{2\pi F} = \frac{R_d}{K}$$

Here, the inductors and capacitors as treated as pure. The series element of inductive reactance is placed in the JD term, and the reciprocal of this impedance of the shunt capacitor is placed in the JF term. The inductor faces the line, and the capacitor faces the load. The L matrix is formed as follows:

$$\begin{vmatrix} V_{in} \\ I_{in} \end{vmatrix} = \begin{vmatrix} 1+J0 & 0+JKR_d \\ 0+J0 & 1+J0 \end{vmatrix} \begin{vmatrix} 1+J0 & 0+J0 \\ 0+\dfrac{JK}{R_d} & 1+J0 \end{vmatrix} \begin{vmatrix} V_o \\ I_o \end{vmatrix} \tag{17.21}$$

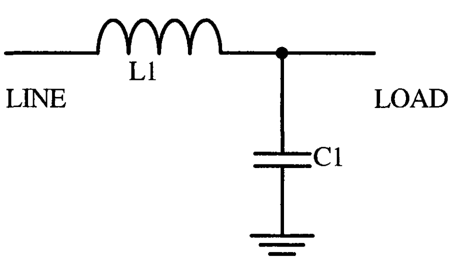

FIG. 17.10 The L filter.

Get rid of the unnecessary zeros, which are here only because the eight positions must be filled so that the computer program sees the full eight units. In this way the same solution formula can be reused repeatedly like a template.

$$\begin{vmatrix} V_{\text{in}} \\ I_{\text{in}} \end{vmatrix} = \begin{vmatrix} 1 & 0 + JKR_{\text{d}} \\ J0 & 1 \end{vmatrix} \begin{vmatrix} 1 & J0 \\ \dfrac{JK}{R_{\text{d}}} & 1 \end{vmatrix} \begin{vmatrix} V_{\text{o}} \\ I_{\text{o}} \end{vmatrix} \tag{17.22}$$

The new matrix becomes

$$\begin{vmatrix} V_{\text{in}} \\ I_{\text{in}} \end{vmatrix} = \begin{vmatrix} 1 - K^2 & JKR_{\text{d}} \\ \dfrac{JK}{R_{\text{d}}} & 1 \end{vmatrix} \begin{vmatrix} V_{\text{o}} \\ I_{\text{o}} \end{vmatrix} \tag{17.23}$$

Then, inserting the other zeros so that the L has the same form as the rest,

$$\begin{vmatrix} V_{\text{in}} \\ I_{\text{in}} \end{vmatrix} = \begin{vmatrix} 1 - k^2 + J0 & 0 + JKR_{\text{d}} \\ 0 + \dfrac{JK}{R_{\text{d}}} & 1 + J0 \end{vmatrix} \begin{vmatrix} V_{\text{o}} \\ I_{\text{o}} \end{vmatrix} \tag{17.24}$$

17.9. THE π MATRIX

In this matrix, the capacitor value is split. Half is placed in front of the inductor in parallel with the line, and the other half is in parallel with the load (Figs. 17.11 and 17.12). The capacitive reactance of both of these halves is then doubled. Yet, it is the reciprocal of this impedance placed in term JF.

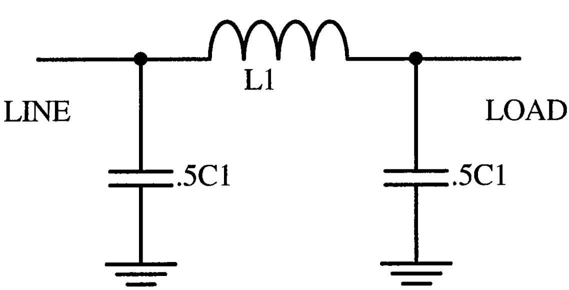

FIG. 17.11 The π filter.

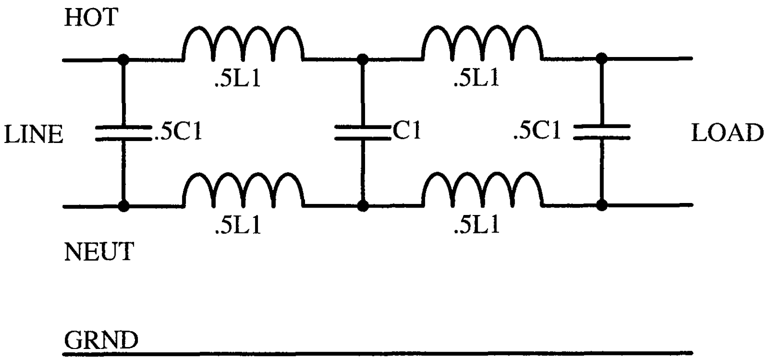

FIG. 17.12 Multiple balanced π filter.

$$\begin{vmatrix} V_{in} \\ I_{in} \end{vmatrix} = \begin{vmatrix} 1 & 0 \\ \dfrac{JK}{2R_d} & 1 \end{vmatrix} \begin{vmatrix} 1 & JKR_d \\ 0 & 1 \end{vmatrix} \begin{vmatrix} 1 & 0 \\ \dfrac{JK}{2R_d} & 1 \end{vmatrix} \begin{vmatrix} V_o \\ I_o \end{vmatrix} \tag{17.25}$$

or half the value then used to make up the L matrix in term JF.
The π matrix is

$$\begin{vmatrix} V_{in} \\ I_{in} \end{vmatrix} = \begin{vmatrix} 1 - \dfrac{K^2}{2} + J0 & 0 + JKR_d \\ 0 + \dfrac{JK(4 - K^2)}{4R_d} & 1 - \dfrac{K^2}{2} \end{vmatrix} \begin{vmatrix} V_o \\ I_o \end{vmatrix} \tag{17.26}$$

with the same form as the rest.

17.10. THE T MATRIX

In this case, the inductor is split with half of the inductor in series with the line and the other half in series with the load (Figs. 17.13 and 17.14). This makes the JD term half the value of the π or L. The full matrix of the T is

$$\begin{vmatrix} V_{in} \\ I_{in} \end{vmatrix} = \begin{vmatrix} 1 & \dfrac{JKR_d}{2} \\ 0 & 1 \end{vmatrix} \begin{vmatrix} 1 & 0 \\ \dfrac{JK}{R_d} & 1 \end{vmatrix} \begin{vmatrix} 1 & \dfrac{JKR_d}{2} \\ 0 & 1 \end{vmatrix} \begin{vmatrix} V_o \\ I_o \end{vmatrix} \tag{17.27}$$

The full T matrix becomes

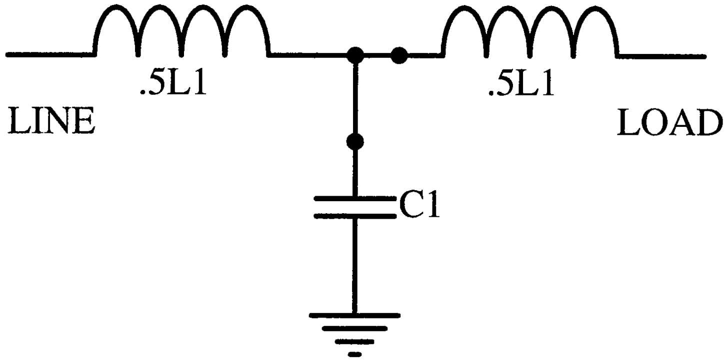

FIG. 17.13 The T filter.

$$\begin{vmatrix} V_{\text{in}} \\ I_{\text{in}} \end{vmatrix} = \begin{vmatrix} 1 - \dfrac{K^2}{2} + J0 & 0 + \dfrac{JKR_{\text{d}}}{2}\left(2 - \dfrac{K^2}{2}\right) \\ 0 + \dfrac{JK}{R_{\text{d}}} & 1 - \dfrac{K^2}{2} + J0 \end{vmatrix} \begin{vmatrix} V_{\text{o}} \\ I_{\text{o}} \end{vmatrix} \tag{17.28}$$

and better

$$\begin{vmatrix} V_{\text{in}} \\ I_{\text{in}} \end{vmatrix} = \begin{vmatrix} \dfrac{2 - K^2}{2} + J0 & 0 + \dfrac{JKR_{\text{d}}(4 - K^2)}{4} \\ 0 + \dfrac{JK}{R_{\text{d}}} & \dfrac{(2 - K^2)}{2} + J0 \end{vmatrix} \begin{vmatrix} V_{\text{o}} \\ I_{\text{o}} \end{vmatrix}$$

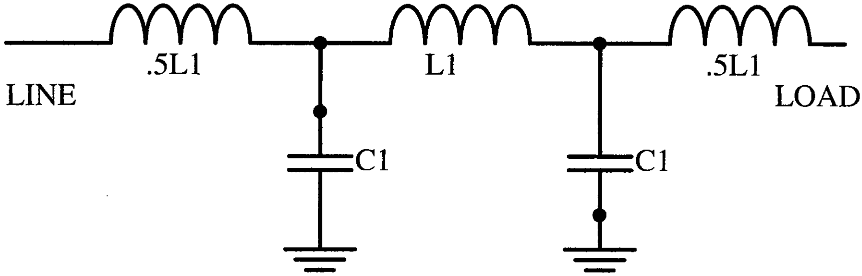

FIG. 17.14 The multiple T.

17.11. THE CAUER MATRIX OR ELLIPTIC FILTER

The Cauer filter (Fig. 17.15) gets rid of a problem frequency. In the 461 specifications in which 100 dB or so is needed at 10 to 14 kHz, a Cauer is sometimes used. F_m is the problem frequency and R_c is the resistance in series with the capacitor. Above F_m, the Cauer looks like a capacitor and the resistor limits the minimum impedance in this capacitor leg. The impedance of R_c is usually assigned the value of R_d, the design impedance of the filter. This is usually employed on the center inductor of a triple-L network. The Cauer is often associated with a triple-L filter (Fig. 17.16). The inductor value is the same as before:

$$L = \frac{R_d}{2\pi F_0} \tag{17.29}$$

and the Cauer capacitor must resonate with L at the problem frequency, F_m.

$$C = \frac{F_0}{2\pi F_m^2 R_d}$$

The impedances are

$$X_l = KR_d \qquad X_c = \frac{2\pi F_m^2 R_d}{2\pi F_0 F} = \frac{M^2 R_d}{K} \tag{17.30}$$

where M is the multiplier to convert from F_0 to F_m.

$$F_m = MF_0$$

The C term is

$$C = \frac{K^4 R_c R_d^2}{K^2 R_c^2 + R_d^2(K^2 - M^2)}$$

and the D term is

$$D = J\frac{(K^3 R_c^2 R_d - K^3 M^2 R_d^3 + KM^4 R_d^3}{K^2 R_c^2 + R_d^2(K^2 - M^2)}$$

R_c and R_d should be equal but are listed here separately if for any reason they should be different for a special problem that any engineer may encounter. If the two terms are equal, C and D reduce to

$$C = \frac{K^4 R_d}{K^2 + (K^2 - M^2)} \tag{17.31}$$

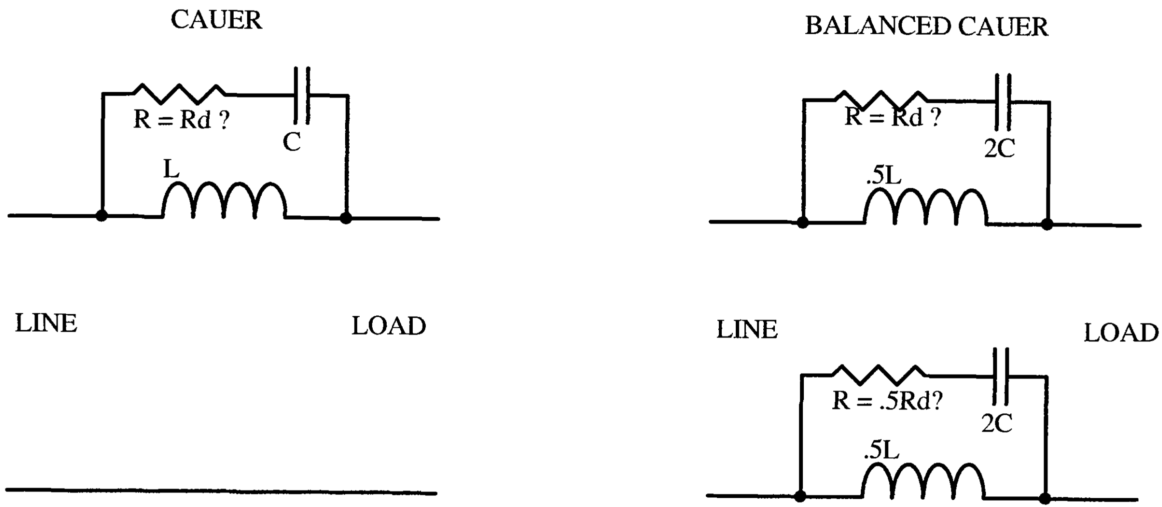

FIG. 17.15 The Cauer matrix.

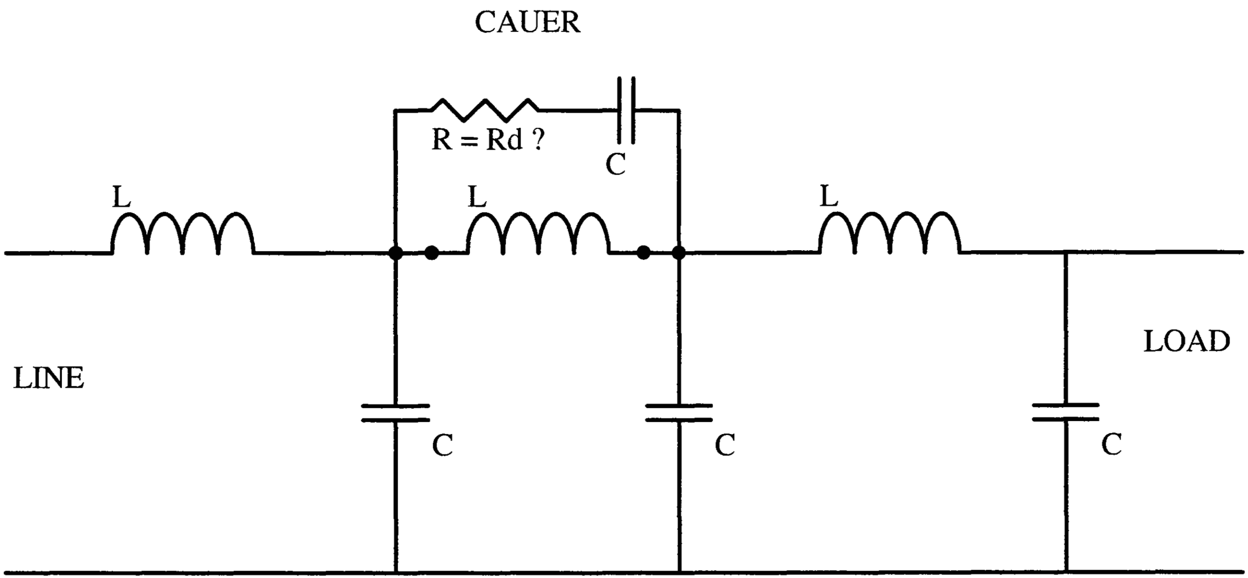

FIG. 17.16 The Cauer or elliptic filter.

which approaches R_d at the higher k values, and

$$D = J\frac{(K^3R_d - K^3M^2R_d + KM^4R_d)}{K^2R_c^2 + R_d^2(K^2 - M^2)}$$

which approaches

$$-J\frac{R_dM^2}{K}$$

which is the capacitive reactance, and this approaches zero as the frequency increases. The matrix for the Cauer is

$$\begin{vmatrix} 1 + J0 & C + JD \\ 0 + J0 & 1 + J0 \end{vmatrix} \tag{17.32}$$

and without the zeros is

$$\begin{vmatrix} 1 & C + JD \\ 0 & 1 \end{vmatrix}$$

The conclusion of this chapter is that the matrix terms are difficult to evaluate, and these are needed if you wish to form these transfer functions on your own. The hope here is that this chapter is clear enough for the reader to evaluate.

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18

Matrix Applications

The first step is to find the filter type needed for the application. The listings in Sec. 5.8 should help. Either through documentation from the EMI test house or mathematically from sections of Chapter 11, the required insertion loss must be known. Often, this is given as a list of frequencies and the required loss at each. The filter must be checked at each frequency to make sure the loss is met at each frequency. The load current input voltage and the line frequency must be known to get the design impedance, R_d, and to find the minimum cutoff frequency. Figure out the value of R_d by dividing the voltage by the expected highest load current, and the cutoff frequency should be 10 times the line frequency.

The first matrix in the chain, or cascade, should be a matrix that represents the line. This could be the R_s, Linesim, LISN, or any other line unit added by the user. These are not part of the filter, or filter construction, but aid in the loss. Therefore, these values should be as accurate as possible. The second, third, fourth, and so on are parts of the filter. These are matrices multiplied together, and the final matrix is solved for the full loss of the filter. As K varies, the loss changes and these can be plotted. This continues into the next section.

One condition discussed here is the low-current applications in which operational amplifiers (OpAmps) are used. Many design engineers make some errors when using this application. These engineers substitute active filters to replace passive components in the lower current applications. These filters must have *very clean power* or they cross talk, generating the same noise, and in some cases even more noise, that the active filter was meant to cure.

18.1. SINGLE-PHASE AC FILTER

As an example, an L is chosen by the filter designer. The design impedance is found to be 10 ohms, and the line frequency is 400 Hz. The loss needed is 40 dB at 20 kHz. This is a low frequency, and the R_s matrix is chosen for the first matrix to represent the line. R_s is thought to be around 4 ohms at 20 kHz. The second matrix is the L matrix chosen by the EMI design engineer. If these are automatically multiplied as a system in the computer, all the other matrices in the chain have unit matrices in them. The final matrix is the result of only the R_s and the L filter because only they control the answer. The value of F_0 is changed to meet this 40 dB loss by varying the value of F_0 or by using a program such as GOAL-SEEKER (an old shareware program). In the newest Lotus program, this feature is built in. The engineer should add 6 dB for headroom to the loss. This gives the filter designer some latitude in component values without fear of harming the performance of the filter. It also gives some additional attenuation to some unknown frequency spike that could push the equipment using the filter over the limit. The value of F_0 here is 1200 Hz to get the desired loss of 46 dB (40 + 6 headroom), and this is sitting on the third harmonic and is far too low. Another L is added. Skip by one matrix, and leave it as a unit matrix. This allows changing it later to an RCSHU, or whatever, if needed. To review:

Matrix 1: R_s
Matrix 2: L
Matrix 3: left as a unit
Matrix 4: changed to L
All remaining matrices: units

This is followed by the column matrix with elements V_0 and I_0 as shown in Chapter 17. I_0 can be replaced by V_0/R_d, where R_d is the design impedance and is the lowest load resistance seen by the filter. Proceed by adding the next L filter. This changes the cutoff frequency to 4600 Hz, which is well above the 4000 Hz recommended limit. The initial inductance was 1.3 MH, and now the requirement is two at 345 μH. The same is true for the capacitor. It was 13.3 μF and is now two at 3.46 μF. This can be balanced and look as in Fig. 18.1.

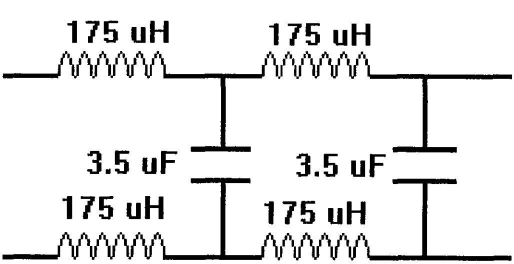


FIG. 18.1 The balanced L.

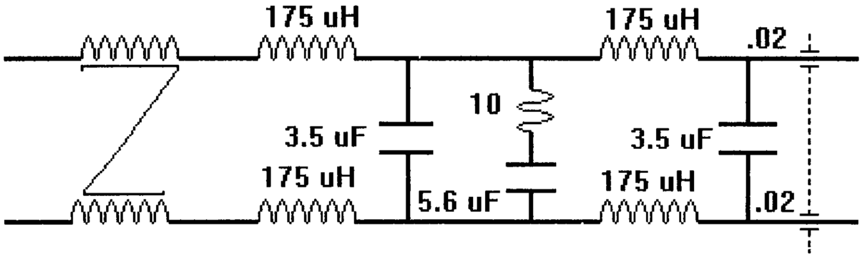


FIG. 18.2 The full balanced double L with Zorro, RCSHU, and feed-throughs.

The load impedance was changed to see how the filter responded in the standby mode, and a resonant rise was found at 2760 Hz. This required an RCSHU to be added to the third matrix, and the common mode components are added in Fig. 18.2. The new listing for the matrix layout is

Matrix 1: R_s
Matrix 2: L
Matrix 3: now an RCSHU
Matrix 4: L
All remaining matrices: units

This matrix group is still followed by the column matrix with elements of V_o and V_o/R_d. Check that all dB losses at the other frequencies listed in the specification meet their requirements for this filter. Most expect all the losses at the higher frequencies to pass if the lowest passes. This is usually true, but there are exceptions, so all must be checked. This is treated later, but the frequency that requires the lowest cutoff frequency determines the cutoff frequency used.

The remaining job is to determine the *Z* (Zorro). The two feed-through capacitors are added, giving 0.04 µF (Fig. 18.3). The line-to-line capacitors are out of the circuit along with the RCSHU. The 175 µH inductors are divided by 2, giving 87.5 and a total of only 175 µH for the four, adding little to the common mode insertion loss. Give all this up to headroom. This is reduced to a single L.

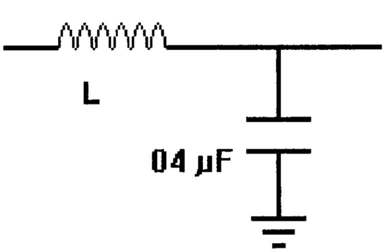


FIG. 18.3 The common mode part of the filter in Fig. 18.2.

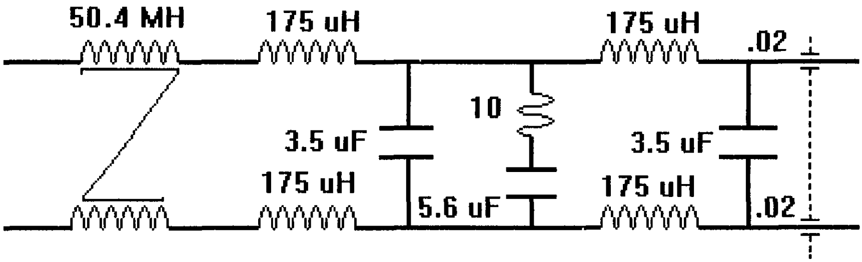


FIG. 18.4 The value of Zorro for Figs. 18.1 and 18.2.

Remove or replace these elements with the unit matrix, except the R_S and the L matrix. The loss for the common mode is 46 dB at the switcher frequency of 50 kHz. The 0.04 μF capacitor must be maintained. This requires an impedance of 1125 ohms and a cutoff frequency of 3550 Hz. The Zorro is 50.4 MH, which is ridiculous. The filter has been designed to component level, but the Zorro should be fixed. Some would substitute any reasonable lower value, hope for the best, and pray for an exemption from the specification if the unit fails. The circuit is shown in Fig. 18.4.

All that remains is to design the two different inductors and the capacitors or choose a supplier. This filter may be out slightly when tested but should require only minor changes. High-frequency problems? Add Capcon (see Chapter 4) or ferrite beads. Differential problems at the switcher? Improving the quality of the RCSHU capacitor or increasing its value and/or raising the inductors to 180 μH, or 200 if needed, may solve the problem. Common mode? Increase the Zorro somewhat. Too high a value for the common mode inductor, as before? Change the feed-through values to 0.01, and add a second section. This is now a double L, as shown in Fig. 18.5.

The new value of each Zorro is 7.9 MH. Now, the case could still be presented again, examining the contribution of the differential mode to the

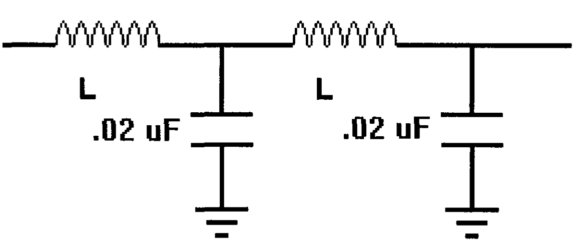


FIG. 18.5 Calculation of the double Zorro.

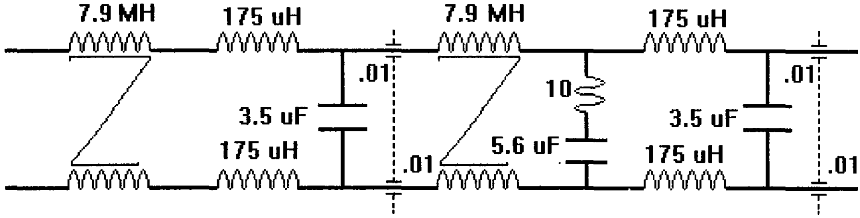


FIG. 18.6 Balanced single-phase filter with two double Ls.

common mode filter. One set of parallel differential inductors aids each common mode inductor. The parallel value of these inductors is 87.5 μH. The Zorro is still an order of magnitude greater than the 87.5 μH and is still not subtracted to lower the size of the Zorro. This difference is left to increase the headroom. Note that this filter (Fig. 18.6) has an internal shield to ground (the case). Note also that the size of the Zorro has dropped. This is also a much more practical inductor with this much current.

18.2. THREE-PHASE FILTER

There are two types of three-phase filters: the wye (or Y) and the delta. These two types are then divided into two groups. The first is the higher current type, in which each leg is a separate insert, with each insert the same type and size. This is very important because of the possibility of installing the smaller neutral filter in any of the main legs.

Also, the third, sixth, ninth, and so on currents add in phase on the neutral wire along with the unbalanced current of the three legs. This harmonic content would be the odd multiples of 3 and some of the even multiples such as 6 or 12. Therefore, the wire size and filter size are the same for the neutral as required for the other three legs.

$$\begin{array}{c} A\,sin(\omega Nt + N0) \\ A\sin(\omega Nt + N120) \\ A\sin(\omega Nt + N240) \\ 3A\sin(\omega Nt) \end{array} \tag{18.1}$$

This assumes that the peak harmonic current on all three legs is the same. If not, the term $3A$ is replaced by $(A+B+C)$. The three phases add for all multiples of 3, if that harmonic exists. The sixth should be a very low level, if it exists at all, but the ninth is strong, especially where the off-line regulator is used as part of the power supply. Beware of multiphase transformers that reduce these harmonics. This is a function of the number of phases used, but most eliminate the lower harmonics such as the third harmonic. This eliminates that problem but can form a inductive voltage divider. The following approximate equation for the off-line regulator current is normalized to one. Refer to Eq. (4.1).

$$\frac{4\tau T}{\pi}\sum_{N=1,3,5}^{\infty}\frac{1}{T^2 - N^2\tau^2}Cos\frac{\pi N\tau}{2T}Sin\frac{\pi N}{2}Sin\frac{2\pi N\tau}{T} \tag{18.2}$$

If the 3, 9, 15, 21, and 27 are added, the peak reaches almost to 0.7. This current, plus the unbalanced current, should prove that the neutral filter must be the same size as the other three legs. The last three-phase type is the lower current mode.

To design the three-phase filter, the voltage and current seen by each filter leg must be known. Find the maximum power required by the load. Divide this by 3 to obtain the power per leg. Divide this answer by the line-to-ground voltage for a wye or the line-to-line voltage for a delta. Multiply, for delta only, the last answer by the square root of 3.

$$\frac{12{,}000}{3} = 4000 \qquad \frac{4000}{208} = 19.231$$

$$19.231\sqrt{3} = 33.309 \qquad 33.309\sqrt{2} = 47.106$$

If the total load power, taking all the inefficiencies into consideration, is 12 kVA, the power supplied to each line is 4 kVA. If the line-to-line voltage is 208 V, the current is 19.231 times the square root of 3. This equals 33.309 line A. Assuming that no off-line regulators are used, the current peak is 47.106 A. The inductors must not saturate at this current, here 47.1 A. The design impedance is 208 V divided by 33.309, giving 6.24 ohms. Use 6 ohms for the design impedance.

18.2.1. Low-Current Wye

The power is from each leg to neutral, and the neutral current is zero if the currents in the three legs are well balanced. But, as mentioned before, the odd third-order harmonics are still present. This is true only for the fundamental of the power line frequency as discussed earlier. In the low-current filter, all of the components are in the same container and the capacitors are wired from the leg to neutral. This makes for smaller capacitors because of the charge from 208 to 120 V. This saves money, weight, and volume.

These filters are often called five wire because of the three legs plus the neutral and the ground. The ground must be left intact and unfiltered. I have seen many filters with components in the ground leg from various filter manufacturers. This is a violation of the electrical code. The ground lead must be intact and not broken. The only exception would be if ferrite beads or toroids were slipped over the wire, leaving the ground lead a solid wire.

The output feed-through capacitors in Fig. 18.7 are to case ground, and the ground wire is wired to the case. Also, the Zorro common mode inductor can be used for common mode rejection because all parts are within the same enclosure.

The common mode inductor cancellation of the magnetic field of the differential mode power current works as shown in Fig. 18.8. Discussing the balanced fundamental power frequency first and ignoring the harmonics, the A, B, and C currents generate magnetic fields that cancel. Here, a magnetic field cannot be generated with no current flow in the neutral common mode leg. If the system is unbalanced, (we know that it is), the difference current flows in the neutral leg of the common mode inductor and still brings the common mode back to balance. As far as the harmonics are concerned, whatever harmonic current flows in any leg also flows in the neutral leg and still cancels. So, as far as the power delivered to the load is concerned, the magnetic field of the common mode is neutralized. Any common mode pulse or signal, either from the load or the line, is attenuated by upsetting this balance. The design technique is the same. R_d is the lowest line voltage divided by the highest line current required by the equipment—not inrush. Find the cutoff frequency, and the values of the inductors and capacitors follow.

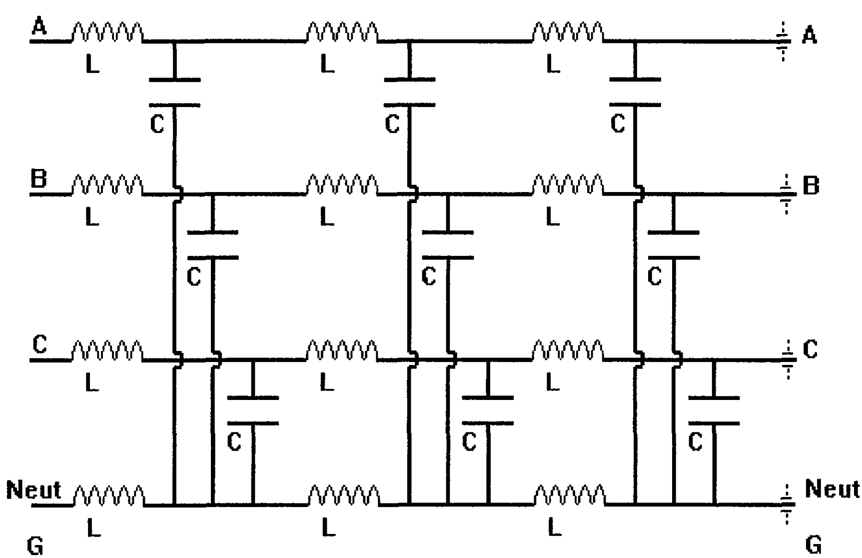


FIG. 18.7 The low-current wye triple-L filter.

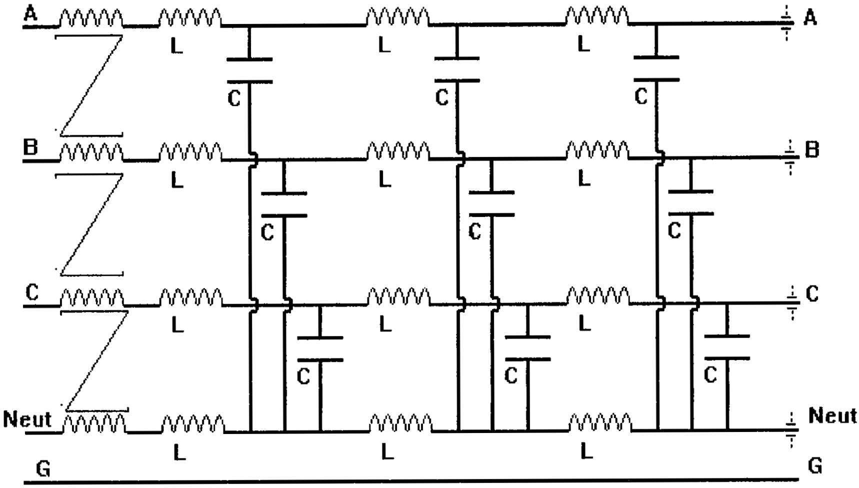


FIG. 18.8 The low-current wye triple-L filter with Zorro.

18.2.2. High-Current Wye

These filters become so large and heavy that they must be built using the insert technique. All three legs and the neutral are the same electrically and physically, and all inserts are installed in a larger enclosure. In this way, if any leg blows, only the failed filter of that leg needs to be replaced. This technique also eases the installation by allowing the outer enclosure to be installed and then the inserts placed in this enclosure one at a time. This is better than hoisting the entire enclosure and contents, which has been done by various contractors. They installed all the inserts, hoisted the full weighted unit, and then dropped it. The entire filter had to be replaced at a premium cost because of time constraints. This was expensive, to say the least.

18.2.2.1. *A Single Insert*

All the capacitors in Fig. 18.9 are to case ground. In this arrangement, there are no components from line to line, such as capacitors or MOVs. The leakage specification, even with power factor correction coils, is rarely met. There really should not be a leakage current specification for this high-current type of filter. This should be known before the specification is written and often is not known or heeded by some specification writers. It has been asked after the filter is shipped that some leakage specification also be met that was not provided to the manufacturer initially, and power factor correction coils had to be added. This improved the situation, but the filter still did not pass.

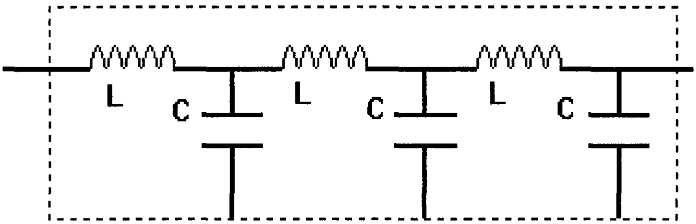


FIG. 18.9 The single high-current insert.

There is some limited common mode in the differential inductors, which are in parallel, and the capacitors, which are also in parallel. Each inductor is divided by the number in parallel, and the capacitors multiply in parallel. Also, this assembly lacks the ability to use the Zorro common mode inductor for common mode rejection. In the following case, the equivalent circuit is composed of $L/4$ and $4C$ as shown in Fig. 18.10.

The common mode equivalent is as shown in Fig. 18.11. The working impedance is now one fourth the design impedance, but the cutoff frequency is the same.

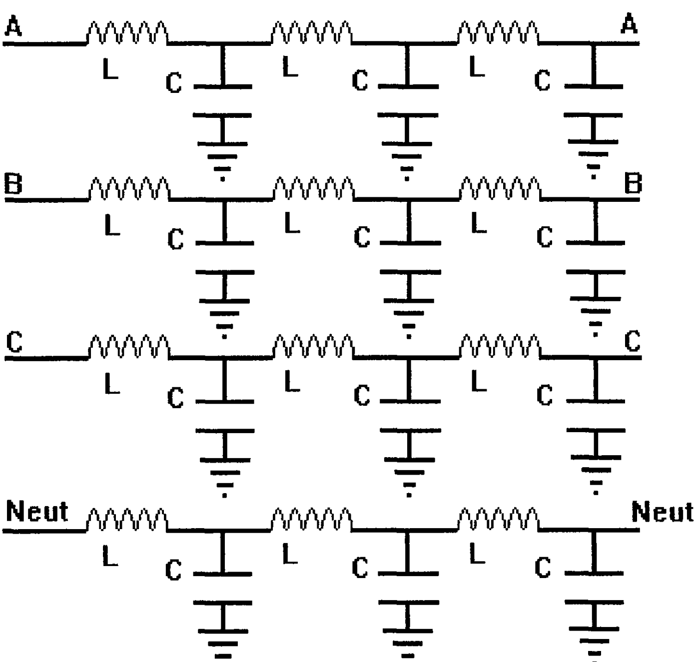


FIG. 18.10 The three-phase four-wire filter.

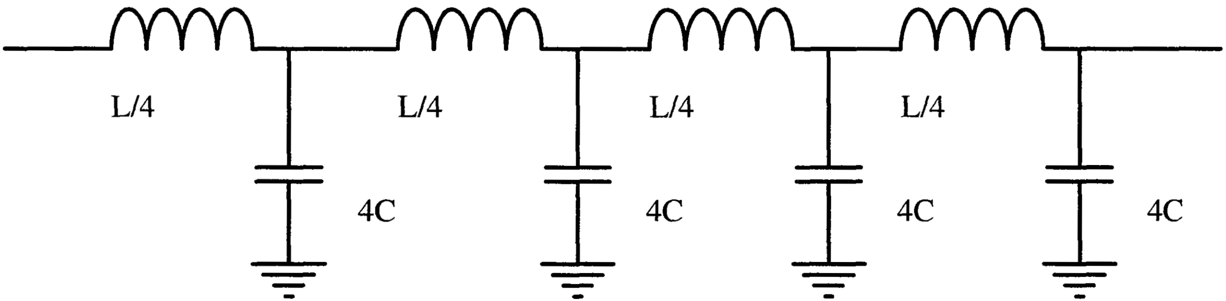


FIG. 18.11 The common mode portion of the four-wire filter.

$$R_d = \sqrt{\frac{L/4}{4C}} = \sqrt{\frac{L}{16C}} = \frac{1}{4}\sqrt{\frac{L}{C}}$$

$$F_0 = \frac{1}{2\pi\sqrt{(L/4)4C}} = \frac{1}{2\pi\sqrt{LC}}$$

Follow the single-phase design method. R_d is the line voltage divided by the maximum line current in any one leg.

18.2.3. Low-Current Delta

This lower current type is often specified to pass 220A with the stipulation "to test any one line with the other two legs grounded." This is a good spot for the π type because the line and load impedances are both 50 ohms. The line-to-line capacitors can be shared, giving twice the capacitance to ground (Fig. 18.12). This makes our job easier.

If the A leg is under test in the 220A system, legs B and C are grounded. Then the capacitors from A to B and A to C are in parallel on both sides of the differential filter inductor. The impedance is decided in the same way using the

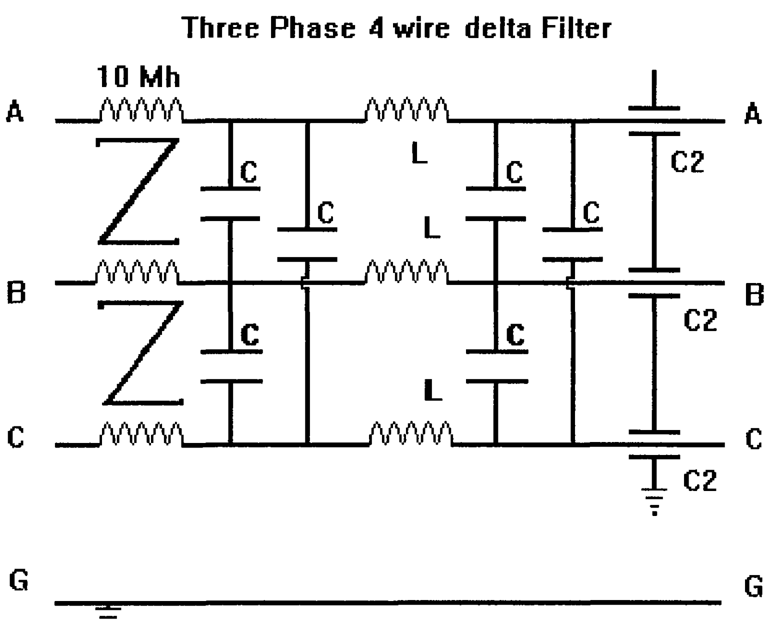


FIG. 18.12 The four-wire delta showing case ground for feed-through operation.

line-to-line voltage and current. Follow the same procedure for the single phase. If the 220A specification is all that has to be met, the capacitor values can be divided by 2, making for smaller capacitors and saving money, weight, and volume. Pick the R_S for matrix 1 and place the π in matrix 2; the rest should be unit matrices.

Matrix 1: R_S
Matrix 2: π
Matrix 3: unit
Matrix 4: second π if needed
All remaining matrices: unit matrices

Remember, the column matrix follows this. If the single π meets all the requirements, the design is finished. If not, add a second π for matrix 4, leaving matrix 3 a unit matrix. This is done for the same reason, to be able to add another network for matrix 3. Once the capacitor value is established, split the value between the lines. If the π is a multiple, check for a resonant rise and fix as before. Continue with the design as in the single phase. Check that each frequency listed in the specification meets the required loss of the list. If not, the filter may have to be redesigned for more loss or any of the other solutions depending on the frequency of the problem.

18.2.4. High-Current Delta

This is the same as the high current wye except that the capacitors are again tied to ground. These should be from leg to leg as in the low-current delta. There is no convenient way to do this using the insert method. The value of the required capacitance should be doubled to give half the required reactance to ground and the other half back to the other leg.

18.3. TELEPHONE AND DATA FILTERS

These are easy EMI filters to design, and the design program provided works well. The reason is that the input and output impedance is known, typically 50, 75, 135, and 600 ohms, and the two impedances are always the same. The filter is always balanced; the two inputs are called tip and ring, and so are the two outputs. Most telephone filters are 300 ohms from line to ground and 600 ohms line to line—actually tip to ring. These are typically πs or Ts. The currents are low, and the filter resistance is usually not critical because it is such a small part of the 300 ohms. That is the way to design it. Use the 300 ohms for the source and load, and this will give the filter for the tip and the same for the ring (Fig. 18.13). As an example, a filter requiring 60 dB of loss at 20 kHz matched to 300 ohms would consist of four π filters (Fig. 18.14).

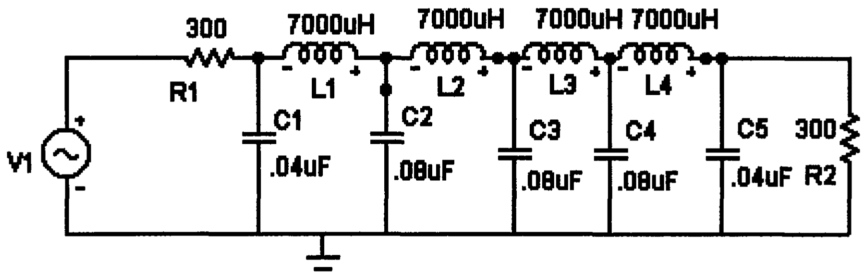


FIG. 18.13 Telco filter.

Data filters are not much different; different impedance and the amount of loss required would be about all. However, remember that Fig. 18.13 is only half the filter. If this is the filter for the tip input and output, there is another for the ring side.

18.4. IMPEDANCE-MATCHED FILTERS—WHAT IS THE IMPEDANCE LIMIT?

This follows from Sec. 18.3. As the impedances of the source and load grow, so does the inductance, and the capacitance drops in value. It comes to a point where the inductor becomes too large and the total DC resistance (DCR) is now a good portion of the specified design impedance. The capacitors become too small. For example, in a 1000-ohm impedance filter, the inductor is 1 million times the capacitor value. I would suggest this as the upper value.

18.5. PULSE REQUIREMENTS—HOW TO PASS THE PULSE

This is a continuation of Secs. 18.3 and 18.4. Pulses are made up primarily of odd harmonics. The quickest way is to obtain the pulse width, take the reciprocal to get the frequency, and multiply this frequency by 10. The EMI filter should then have a cutoff frequency above this frequency. If this is impedance matched as in Secs. 18.3 and 18.4, the band pass will be very flat. This will pass the fundamental, 3rd, 5th, 7th, and 9th along with some of the 11th. The pulse distortion should be minimal.

18.6. THE DC-TO-DC FILTER

The DC-to-DC type is often a tubular type and is unbalanced. Figure 18.15 shows the proper hardware for installing the tubular through a wall or equipment

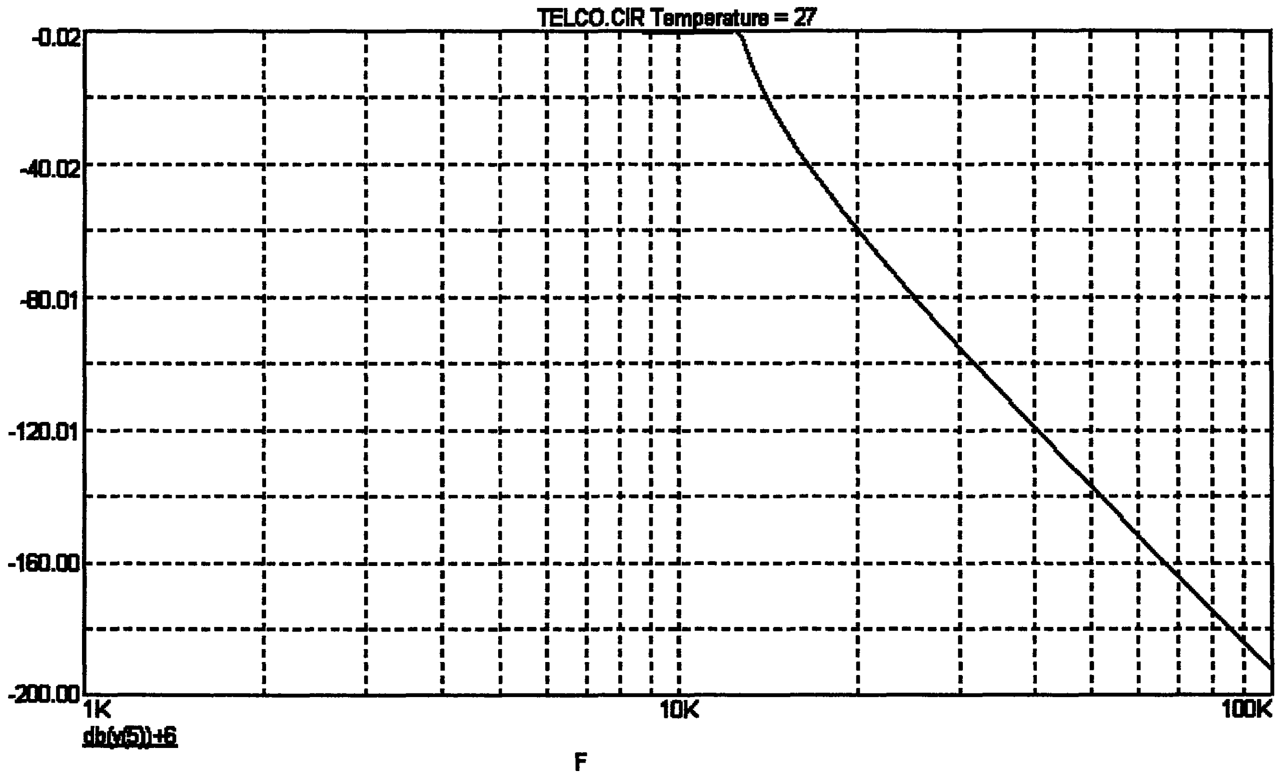


FIG. 18.14 The Telco loss.

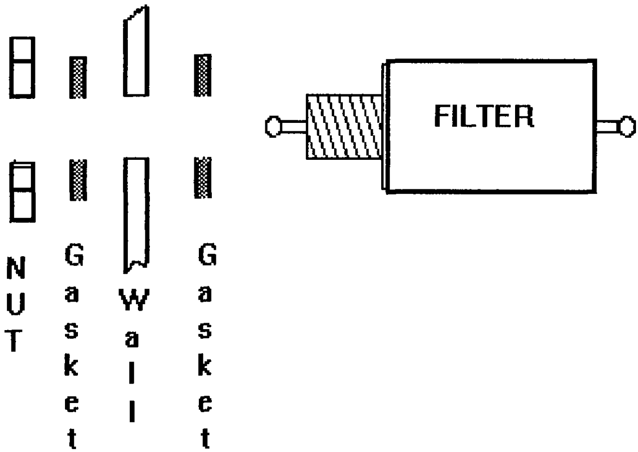


FIG. 18.15 The unbalanced feed-through filter.

housing. Figure 18.16 shows the unit installed. The filter is often a single feed-through capacitor giving just 6 dB per octave loss. In some other filters, inductor(s) are included, making up the L or T type. These last two give 12 or 18 dB per octave. The filter requires a good ground, or the filter is out of business. If this is an L or T, then only the inductor(s) is in the circuit. This reduces the loss to 6 dB rather than the 12 or 18 that the L or T would give. But what if the filter is a single feed-through type? This would net zero loss, all because the filter is not grounded.

The construction is as follows (Fig. 18.17). The output pin is a snug fit through the capacitor center arbor, and this pin is isolated from the outer wall. The capacitor is soldered to this outer wall washer (ground). The inductor and

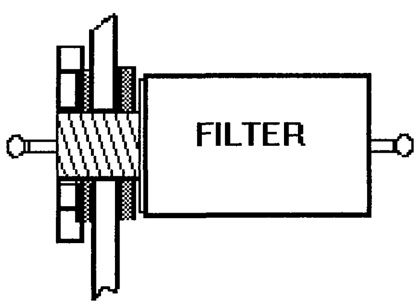


FIG. 18.16 The installed feed-through filter.

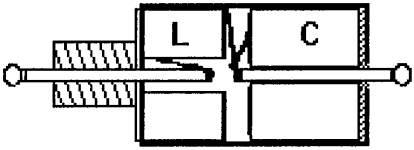


FIG. 18.17 Construction of the single-phase feed-through.

capacitor are soldered to the output pin. The outer tube is placed over the inductor and capacitor and soldered to the end plate. The inductor is soldered to the input pin pressed through the threaded area and extends into the I_d of the inductor. This top capacitor and threads are soldered to the tube, completing the unit.

18.7. LOW-CURRENT FILTERS

The low-current filter suffers from conditions opposite to those that affect the high-current filter. Here, the inductors get bigger while the capacitors get smaller. One way is to employ RC filters in which the value of R should be less than 10% of the minimum load resistance. The disadvantage is that the circuit gives only 6 dB per octave. A better method for low-current filters is to employ active filters (Fig. 18.18). They are small and light and can be designed with many poles, but sometimes the higher frequencies suffer because of the open-loop gain of the OpAmp. It is often overlooked here that the DC feeding OpAmps must be very clean. Therefore, the filter(s) needs a filter (Fig. 18.19). There have been cases in which a number of high-impedance lines must be filtered. The plus and minus

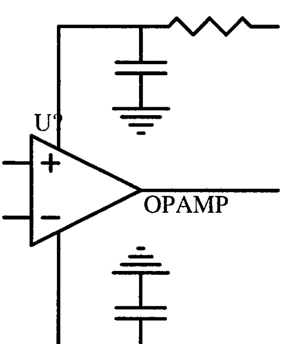
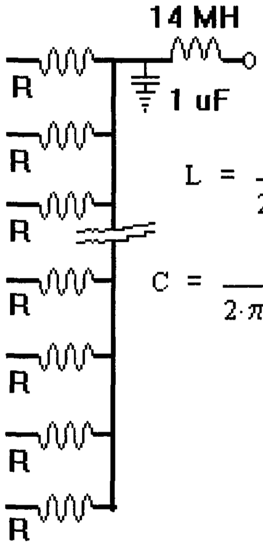


FIG. 18.18 The active filter.



$$L = \frac{10 \cdot 120}{2 \cdot \pi \cdot 14000} = 0.014$$

$$C = \frac{10}{2 \cdot \pi \cdot 14000 \cdot 120} = 9.474 \cdot 10^{-7}$$

FIG. 18.19 The filter required to filter the active filters.

voltage for each OpAmp must be filtered with an RC filter or a conventional passive filter, with the capacitor facing the load. The combined voltage feed must be filtered with a passive filter.

$$R >> \frac{B^+}{10 I_{max}}$$

If the voltage is 12 V and the maximum current is 10 mA, the maximum value of R is 120 ohms. Make R 100 ohms, and the capacitive reactance value is also 100 ohms at half the needed cutoff frequency. Note that this gives a 1 V drop, and if this is excessive, either the resistance must be lowered or the voltage supply must be increased to allow for it. If 14 kHz is the OpAmp cutoff frequency, then at 7 kHz, C is equal to 0.22 μF; and round this off to a standard value. The same is true for the other B, supply. If there is a group of these, say 10, then the total B^+ current is 100 mA. This gives an impedance of 120 ohms.

The impedance of the inductor should be 10 times the 120 ohms just calculated, and the impedance of the capacitor should be one tenth this impedance of 120 ohms. A good quality 1 μF capacitor should remove the noise so that one OpAmp does not add noise to the next OpAmp. The inductor removes any of these signals from the main supply.

18.8. F_0—THE EASY WAY

The filter can almost be designed without any program—truly, on the beach of a desert island with wet sand and a sharp stick! In other words, filters can almost be designed without a calculator if the design is simple. This means that the design should be without RCSHU or Cauer circuits. Let us look again at the single-phase section. The goal was 46 dB, with headroom, at 20 kHz. If the engineer selected the L, as before, this would give 12 dB per octave or 40 dB per decade of attenuation. The beached engineer would use his sharp stick to write the values in Fig. 18.20 in the wet sand.

The designer divides the frequency by 2 and subtracts 12 dB for each row, opting for the single L. It is obvious at the 5 kHz at 22 dB level that a single L would not work for this 400 Hz power system. This is because the next row is half this value of 5 kHz. This figure is below the minimum of 4 kHz that should be followed for better design that keeps the filter transparent to both the line and load. Look how close this lonely filter engineer was to our reading of 1260 Hz for 46 dB of loss!

The engineer has many miles of wet sand and starts over a few feet away in new clean wet sand with a double L at 24 dB per octave (Fig. 18.21). The engineer concludes that the double L will work and that the cutoff frequency is 5000 Hz. The double-L value of F_0 was 4600 Hz, which is close. The 10 ohm design impedance must also be known, so the value of the inductors and capacitors can be calculated. For the inductor, the engineer scratches the following:

$$\frac{10}{2\pi 5000} = 318\ \mu\text{H}$$

and obtains 318 μH. For a balanced filter, this is divided by 2, giving 160 μH—not all that far away from the 175. Actually, the value of 3 for π would be

Frequency	dB
20 kHz	46
10 kHz	34
5 kHz	22
2.5 kHz	10
1.25 kHz	- 2

FIG. 18.20 F_0, the easy way, chart 1.

Frequency	dB
20 kHz	46
10 kHz	22
5 kHz	- 2

FIG. 18.21 F_0, the easy way, chart 2.

close enough for beach work. This would have given the value of 333 μH. The capacitor followed on more fresh sand a few feet over (I hope the tide does not come in)!

$$\frac{1}{2\pi 500{,}010} = 3.18\ \mu\text{F}$$

The designer knows that dividing the inductance value by the square of the impedance can solve the capacitor value. This is 10 ohms and squared is 100 (much easier to do in the wet sand); again, this is close to the 3.5 used before. The designer would round this up to 3.2 μF anyway. The Zorro remains to be calculated, and the designer knows that the feed-through limit is 0.02 μF, or paralleled giving 0.04 μF to ground (Fig. 18.22). Note how close this is to the 3550 earlier. The Zorro inductor value follows:

$$\frac{10}{2\pi 3125} = 509.3\ \mu\text{H}$$

Frequency	dB
50 kHz	46
25 kHz	34
12.5 kHz	22
6.25 kHz	10
3125 Hz	- 2

FIG. 18.22 F_0, the easy way, chart 3.

and the capacitor follows but should be held to two times the maximum to ground, or 2×0.02 or 0.04. Again, the value of π, for beach work, could be rounded to 3, but the value is 5.09 μF. Dividing the 5.093 by 0.04 tells the engineer that the capacitor is 127.32 times too big! So the beached designer multiplies the inductor by this figure, 127.32, and obtains 64.8 MH. This value is somewhat close to the 50.4 MH from the earlier design. This value of inductance is also too big as was concluded in the prior design. This engineer, similarly, adds another L filter, making the common mode a double L. The double L has 24 dB of loss per octave (Fig. 18.23).

The calculation of the inductor follows:

$$\frac{10}{2\pi 12{,}500} = 127.3\ \mu\text{H}$$

and the capacitor:

$$\frac{1}{2\pi 1{,}250{,}010} = 1.273\ \mu\text{F}$$

which is 63.65 times too big as compared with the total of 0.02 μF (four feed-throughs of 0.01 each), so the inductor is multiplied by 63.65, giving 8.1 MH, which is close to the earlier 7.9. Note that the impedance of the common mode filter section is no longer the 10 ohms.

$$\sqrt{\frac{8100}{0.02}} = 636.396 \qquad \text{or} \qquad 10 \times 63.65$$

and the cutoff frequency is still 12,500 Hz:

$$\frac{1}{2\pi\sqrt{0.0081 \times 0.02 \times 10^{-6}}} = 12{,}5000\ \text{Hz}$$

The final design is shown in Fig. 18.24.

Frequency	dB
50	46
25	22
12.5	- 2

FIG. 18.23 F_0, the easy way, chart 4.

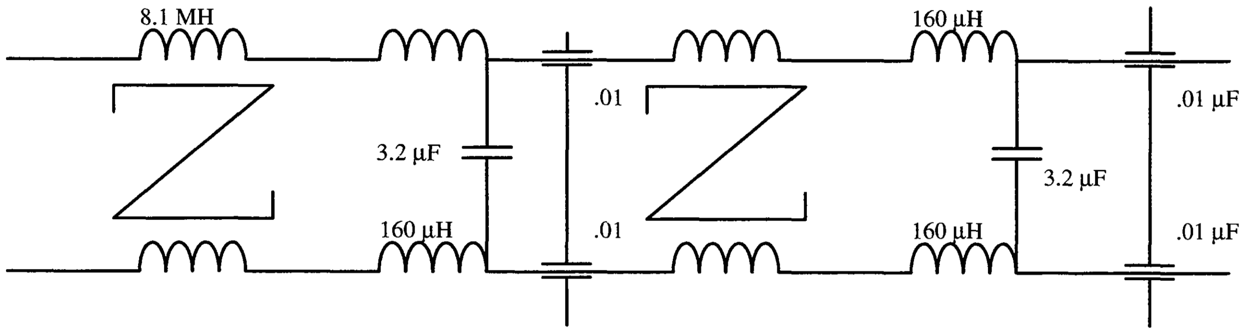


FIG. 18.24 The fully designed filter using "F_0, the easy way."

Frequency	dB
160 kHz	0
16 kHz	60
8 kHz	78
4 kHz	96

FIG. 18.25 Using the dB loss per decade and also the dB octave loss.

This technique of trading inductance for capacitance can be used in common mode filter design but not in differential, or normal, mode design because the design impedance changes. There is no need to match the impedance in common mode. Also, the two Zorros lowers the circuit Q because the inductance is so much lower than with the single Zorro. The Q was already very low in common mode. Thus, the increase does not enhance oscillation. This technique, if used in the differential mode, would raise the Q in the differential mode, which is already too close to the edge of oscillation. Also, the common mode load impedance of the actual equipment, in most applications, is much higher than the differential mode impedance. This fights against the common mode Q increase and reduces the chances of oscillation. The same technique works on Ts and πs except that 18 dB is used, replacing 12 dB for the L with

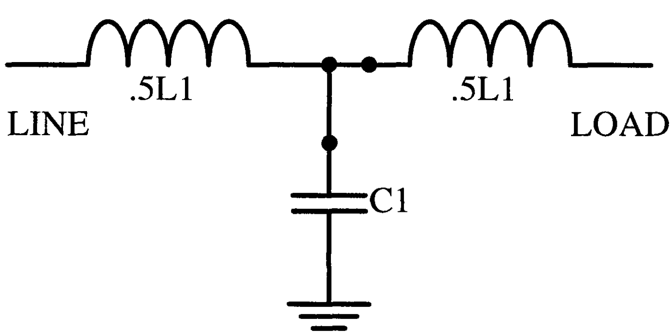


FIG. 18.26 The T used to show the element count.

36 dB Loss @ 100 kHz		86 dB Loss @ 800 kHz	
Frequency	Loss	Frequency	Loss
100 kHz	36	800 kHz	86
50 kHz	24	80 kHz	46
25 kHz	12	40 kHz	34
12.5 kHz	0	20 kHz	22
		10 kHz	10
		5 kHz	- 2

FIG. 18.27 The dB loss at two different frequencies.

18 dB for the Ts and πs as the frequency in the chart is divided by 2. Use 30 dB per octave for two or multiple networks, adding 12 dB for each network added. Three networks would give 42 dB per octave. Count the elements of the entire network: the double T has five elements at 6 dB per octave each, giving 30 dB each octave. If the frequency is high enough, the decade system can be used and the decade and octave can be used together. As an example, a T has 18 dB per octave and 60 dB per decade. If 90 dB is required at 160 kHz, will the filter be okay for a 400 Hz line frequency? Add 6 dB headroom for a required loss of 96 dB. The first step was the decade followed by two steps of octaves. The filter should be okay (Figs. 18.25 and 18.26).

Again, the filter must be checked at each frequency listed in the specification to see that the filter meets the assigned loss at that frequency. One specification was as follows. 30 dB at 100 kHz and 80 dB at 800 kHz. Adding 6 dB for headroom, 36 dB loss is required at 100 kHz and 86 dB at 800 kHz. Using the single L, the loss chart is shown in Fig. 18.27. Most would think that a filter designed for the lowest problem frequency would be good enough to handle all the rest of the list—good odds but not always true as in the preceding example. The single L's cutoff frequency must be 5 kHz, not 12.5 kHz

The easy method for the computer or calculator is as shown in Fig. 18.28.

Required Loss	dB	
The Loss per Octave for filter	L	
Number of Octaves required	X	$X = \frac{dB}{L}$
Frequency @ needed Loss	F	
The Estimated Cutoff Frequency	F_O	$F_O = \frac{F}{2^X}$
	$F_O = \frac{F}{2^{\frac{dB}{L}}}$	

FIG. 18.28 F_0, the easy way with calculator.

18.9. REMOTE HIGH-VOLTAGE SUPPLY FED FROM A LOCAL DC POWER SUPPLY

A very capable power supply company had a contract to design a power supply with several output voltages. One was 60 V at 0.5 A. After the design, they built and tested three and sent two to the customer for approval. After several weeks, the approval was received and the well-designed EMI filter at the power input also passed the 461 specification. Two days later, a panic call came from the customer. The request was for them also to design a remote power supply utilizing the 60 V at 0.5 A for a built-in remote monitor. This monitor power supply requires 16 inches of cable. A key player had died, so the other group could not finish the design on time. The output was 20 kV at 0.8 mA. The power supply company designed a double-ended forward converter with each half conducting with a duty cycle of 0.8 of its conduction time with a switching frequency of 50 kHz. This makes the current pulses 100 kHz at an on-time current of 0.625 A. The questions are:

1. With a proper filter at the front end of the main supply, is another filter needed?
2. Where would the filter, if needed, be located: at the output of the original supply, somewhere in the cable, or at the input of the new supply?

3. Besides the obvious snubbers etc., what else could be done to reduce the size and cost or to eliminate the filter?
4. Do we care what the 60 V power supply's output impedance is at 100 kHz?

The answer to the first question is yes, and the reason will be apparent as the other questions are answered. The second answer is, if needed, at the input of the remote supply because the current on the cable between the two supplies needs to be steady DC with as little ripple as possible. This is to eliminate radiation to other wires within the harness and also to the surrounding areas. For number 3, if the input section of the remote supply was isolated, requiring a return lead to the first supply, a twisted pair would help eliminate the radiated emissions. Although the filter is designed properly, there will be some ripple on the cable after filtering. A shielded pair would also be good but ground the shield only at one end. The best end would be the 60 V end, to reduce ground currents. Also, Capcon could be used to cover the wires between the two supplies and dissipate the energy.

Before the last question is answered, the following should be understood. What if the original designer knew that the remote switcher frequency was 100 kHz and designed the 60 V output impedance to be milliohms around this frequency? What would the results be? The path of least resistance for the 100 kHz current pulses is through the cable, creating high-level radiation to the surrounding media. Discounting radiation, the rest of this circuit should work fine. A filter at the input to the remote should reduce this radiation.

Now the opposite condition: the original engineering group were unaware of the purpose of the 60 V, and the output impedance is 24 ohms in the 100 kHz region. What would the results be now? The radiation would be reduced, but the switcher may not function properly at the reduced voltage. The peak on current is 0.625 A, reducing the supply voltage to 45 V. A filter at the input to the remote should reduce this switcher drop or starvation.

Note that the solution to both problems is the same: a filter. A good filter reduces the peak-to-peak ripple level, reducing the radiation. This filter also reduces the output impedance to the switchers following the filter located in the remote power supply. The answer to question number 4 is no. The system needs a filter whatever the output impedance and will fix either problem.

To design the filter, all the processes covered in this book could be gone through—the dBuA/MHz, the conducted emissions, and the radiated emission—but this takes time and involves testing costs. The filter can be designed without all this and may require some final adjusting, but so would the system after all the time and expense.

The easiest way is to divide the 60 V by the 0.625 A *on current* and get the on impedance of 96 ohms. Divide this by 10 to get the filter capacitor impedance at 100 kHz of 9.6 ohms. The value of the capacitor is

$$\frac{1}{2\pi \times 100{,}000 \times 9.6} = 0.166\ \mu\text{F} = 0.2\ \mu\text{F}$$

or round this up to 0.2 μF and pick a type having a self-resonant frequency (SRF) at least 10 times the switcher frequency. This is 100 kHz, in this case, giving 1 MHz SRF. The inductor should have 10 times the 96 ohms, for this example, or 960 at 100 kHz.

$$\frac{960}{2\pi 100{,}000} = 0.002\ \text{H} = 2\ \text{MH}$$

This is split with half in the hot side and half in the return side of 1 MH each (Fig. 18.29). This is a large value of inductance but easy to design for the DC current of 0.5 A, although the SRF may fight. Make it as high as possible.

The filter impedance is

$$\sqrt{\frac{0.002}{0.2 \times 10^{-6}}} = 100$$

which is close to the 96 ohms, and the cutoff frequency is

$$\frac{1}{2\pi\sqrt{0.002 \times 0.2 \times 10^{-6}}} = 7958\ \text{Hz}$$

The approximate loss is shown in Fig. 18.30.

This should knock the socks off the radiated emissions and give a low-impedance source to the switcher. The pulse power is

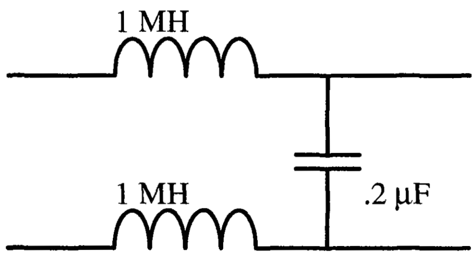


FIG. 18.29 The remote filter in the monitor power supply.

Frequency	dB Loss
8 kHz	9
16 kHz	12
32 kHz	24
64 kHz	36
100 kHz	42

FIG. 18.30 The monitor filter loss.

$$20 \log \left| \frac{2Ea \times 10^6}{T} \right|$$

which equals

$$20 \log \left| \frac{2 \times 0.625 \times 8 \times 10^6}{10} \right| = 120 \text{ dB}$$

The 20 dB per decade starts at

$$\frac{1}{\pi a} = \frac{1}{\pi \times 8 \times 10^{-6}} = 39{,}789 \text{ Hz}$$

which is 39,789 Hz, and the level at 100 kHz is

$$20 \log \left| \frac{1}{\pi F a} \right| = 20 \log \left| \frac{1}{\pi \times 10^5 \times 8 \times 10^{-6}} \right| = -8 \text{ dB}$$

where a is in seconds and F is in Hz.

With –42 approximate loss in the filter and –8 from the wave, the full loss would be –50, and subtracting this from 120 would leave 70 dBua. This should be adequate because this is not on the power line and what does radiate is reduced by distance. However, this may still be out of specification, depending on the requirements in the specification. If this is out of specification, add the shielded balanced cable first, with the ground at the initial supply end, first. If the system still fails, add another, identical filter. This should give another –42 dB, so now the level would be around 30 dBua and must be adequate. Do not try to solve this problem by going to a double L with smaller values. This could work, but the DC voltage drop feeding the switcher could be excessive.

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19

Applications Using Round or Square Conducting Rods

Two similar conditions are treated in this section. The first is the very high current filters based on toroids and round conducting rods through them. The capacitors are the feed-through type, also threaded onto the same conductive rod. The second method uses square conductive rods, but the capacitors are not centered on the rods as in the first method. They are 90 degrees to the rod, and the feed-through bolts are threaded through holes in the conductive bar. In some cases, the conductive rod, or bar, is made from material other than copper because of the strength needed to support the torque of the nuts. This is especially true in the first case following requiring aluminum. Make sure to use a large enough bus or bar so that the heat rise is low. In some, heat sinks have been inserted between the cores making up the inductor. The capacitors either surround this bus or are tightened against this bus, in the second system, and so the bus temperature must be low. This section discusses their major disadvantage. The best solution is not to require them! Several thoughts are discussed here, with some ideas of how to design around them if the need comes up.

19.1. VERY HIGH CURRENT FILTERS

First, three ideas are reviewed here. The first is the number of filters in cascade or tandem. If the Q is low enough, the number of resonant rises is one less than the tandem number in most applications. The maximum is normally four multiple filters in tandem, which should yield three resonant rises. This holds only if the Q is low enough; otherwise, there would be four resonant rises. In high-current filters, the rule of a maximum number in cascade, or tandem, is broken for several reasons. The new maximum number is seven filters in tandem. For example, there could be seven L filters one after another. This gives six potential resonant rises if the Q is low enough. One reason for the exception of a maximum of four in cascade is the high capacitor values required for proper design. They become very unrealistic, and the self-resonant frequency (SRF) drops to low values for the two, three, or four cascaded filters. As the number in tandem increases, the capacitor value decreases. The inductor values have the opposite problem. They shrink in value into the low μH range. To fix this low value problem of the inductors and to reduce the size of the capacitors, the second idea must be discussed. This is paralleling the filters. If the number in parallel is N, the impedance of each unit in parallel is multiplied by N. The value of the inductors is N times larger, and the value of each capacitor is divided by N. The impedance of each filter unit in parallel is

$$\sqrt{\frac{NL}{\frac{C}{N}}} = \sqrt{\frac{N^2L}{C}} = N\sqrt{\frac{L}{C}} \tag{19.1}$$

and the parallel filter impedance is the same as the initial impedance because of N in parallel.

Another solution to the high-current filter is not to filter it! Filter the lower current areas. The Army Corps of Engineers consistently ignores this idea. They wanted an entire power line, with currents of 600, 800, and 1000 A per phase, filtered rather than to filter the distribution legs, which would be easier. It is likely that the sum total of the branch filters would be smaller in volume and weight than the one big filter. These branch filters would probably cost less, also. The filter's ability to operate at higher frequencies, because of the improved SRFs, would also greatly increase.

Another costly solution is to use transformers to step the voltage up and the current down (Fig. 19.1). This raises the design impedance, making the components much more realistic. The higher design impedance increases the value of the inductors, making them more realistic. This also holds true for the capacitors and makes for smaller, more realistic capacitor values. Both components would have better SRFs. Notice the second word, cost. These filters are as expensive as gold anyway, so two transformers on opposite ends may not affect the price too much. Also, the transformer adds to the differential and common mode loss, cutting the cost of the filter. See Chapter 10.

Another way may be to filter the power before the step-down transformer to save the cost of one transformer. The input transformer is either a wye to delta or a delta to delta. Another advantage of adding the two transformers is that the neutral current caused by unbalanced currents on the neutral is eliminated. This

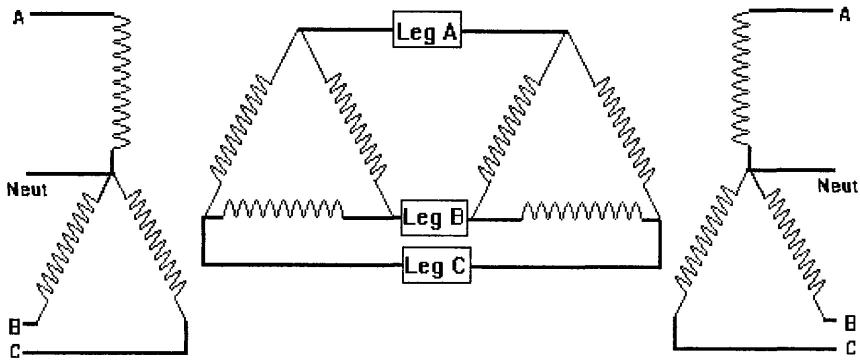


FIG. 19.1 The step-up and step-down transformer for lower current filtering.

is true of all the harmonics that are multiples of 3 and especially the third harmonic current. These add in phase on the neutral along with the neutral unbalanced current and are eliminated. Filtering this as a three-phase three-wire DELTA would also end the requirement to filter the neutral, which must be as big as the filters in the other legs. (See Sec. 18.2 for more information.) Carry the ground wire through to these filters because they must be well grounded to function properly. The cost saved by eliminating the fourth neutral filter helps offset the transformer cost if this technique is used. The current in each leg is still high and may require several filters in parallel besides using the transformer(s) technique.

The following is based on approximations that are useful for designing the filter quickly. The values can be moved later when they are at least on the ballpark property if not on the ballpark playing field.

The cutoff frequency, F_0, the value of the inductance in microhenrys, and the value of capacitance in microfarads are listed in Table 19.1. This is for 440 V, 400 A at 400 Hz requiring 100 dB at 14 kHz. These values were calculated by using the methods discussed earlier through matrix algebra. The design impedance is 440 divided by 400, or 1.1, and multiple Ls were chosen.

TABLE 19.1 High-Current Unparalleled Filter Values

Number of elements	Cutoff frequency F_0	Inductance	Capacitance
5	3750	46	38.5
6	4500	39	32.5

The cutoff frequency of the five L filters in tandem is a little low for 400 Hz. This makes the six-element filter the better choice. But look at the value of the capacitors required for either the five- or six-element filters. This would require five or six capacitors this size and the inductors to make up the filter. These values for the capacitors are much too high. Even if two capacitors were paralleled, 20 μF for the five elements and 16 μF for the six elements, the SRF would be too low. However, many EMI filter manufacturers use the GE-type capacitors but the droves to make these filters. In some filters a total of 220 μF is common. The SRF of each capacitor would be in the low-frequency area around 60 kHz or lower, depending on the type of capacitor chosen. Paralleling the entire filter would aid in solving this problem, as suggested in Table 19.2. The design impedance is now 4.4 ohms, with four in parallel, and the current is now 100 A. This would form a two by two, giving four filters in parallel. With this low current (100 A per filter in parallel), conventional design could be used.

Four elements are much too low to consider for 400 Hz; five elements are okay. The advantage of six elements is that the capacitance has been reduced from 9.2 to 8 μF, giving a little higher SRF, and the same is true for the inductance. It is obvious that this solution makes for a large number of elements. If the filter engineer opts for the six-element filter, this equates to six by four inductors at 152 μH and six by four capacitors at 8 μF, or a total of 48 components. Big? Yes! Heavy? Yes! Costly? Yes! (I hope you have to build more than one to help offset the design costs!)

This is the disadvantage of the high-current filters. The best solution would be to bring this power into the service, then filter the individual feeds after each breaker. The current falls, requiring less elements in each filter and ending the possible need for the two transformers and the need to parallel the filters. Some prefer the opposite technique, to filter feed the breaker so that the filter is always alive. This means that the distribution bus feeds the filters and the filters feed the breakers. The reason is that most designs have too large capacitors in the filters, and this creates difficulty in turning the power on because of the added inrush of filter current. Look at Table 19.2, where six 32.5-μF capacitors are in parallel. The inductor furnishes low inductive reactance at the power frequency, so these

TABLE 19.2 High-Current Four-in-Parallel Filter Values

Number of elements	Cutoff frequency F_0	Inductance	Capacitance
4	3000	233	12.0
5	3950	177	9.2
6	4600	152	8.0

give little isolation between the capacitors. This gives a total of approximately 195 μF. This capacitance plus the normal inrush may cause the breaker problem, but many designs have more capacitance, adding to the problem.

If the engineer is stuck and has to design this system, some ideas follow. The current-carrying element is of bronze, brass, or other material, rather than a wire or wires. The conductor rod should be round. Copper comes to mind but may not support the end nuts, putting too much tension on each threaded end. This stronger conducting material is often be plated with silver. This slightly improves the current-handling ability of the rod. The bar or rod has a diameter specified to handle the current flow with little heat rise. The conductor is not insulated in most designs. Both the inductor and capacitor inner diameters should be slip fits over this conductor without being too loose. The capacitors should be the feed-through type with the arbor inside diameter a tad larger than the rod. The inductor is also a toroid with the inside diameter slightly larger than the rod. These are slipped onto the rod to make the filter elements. N, the number of turn(s) of the inductor, is equal to one. This is the copper or silver-plated rod, with a 95% or better fill factor compared with the typical 40% for toroids. This bus runs through the inside diameter of the toroid as well as the capacitor. Later in this section, the bus bar type of copper rather than the rod is discussed.

The approximate design equations for these high-current filters follow. The inductance is

$$L = \frac{0.4\pi U_e A_c N^2 \times 10^{-8}}{M_{pl}} \tag{19.2}$$

where N is now just one turn, A_c is the cross section of the core in centimeters squared (times a stacking factor, S_f), U_e is the effective permeability, and M_{pl} is the magnetic path length, also in centimeters. The core is tape-wound silicon steel that has a "soft core" for its BH curve (Figs. 19.2 and 19.3). This inductor should not be made using a square loop type of material. The tape thickness should be a

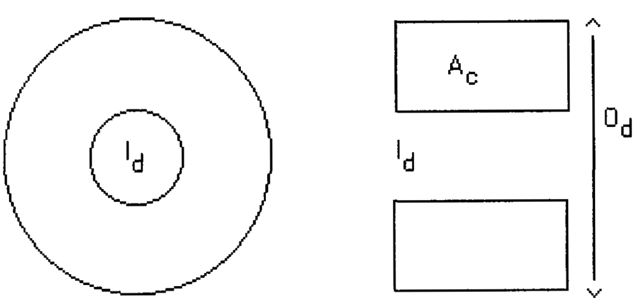


FIG. 19.2 Tape-wound toroid showing I_d, O_d, and A_c.

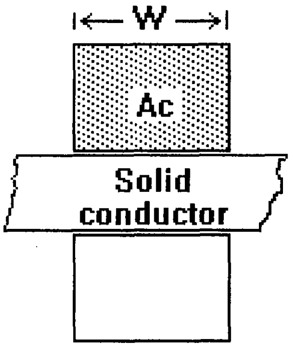


FIG. 19.3 Toroid inserted on the copper bar.

size smaller than the size required for the line frequency, especially if the inductor must be tuned as part of a Cauer. This is to prevent core losses at the harmonics of the line frequency and harmonics from any off-line regulators. Otherwise, the high harmonic current heats the core, requiring a larger core. Trade-offs of this type are discussed later. The core manufacturer should also gap this toroid so that the peak current does not saturate the core. This can be determined from the BH curve, where H is

$$H = \frac{0.4\pi N I_p}{M_{pl}} = \frac{0.4\pi\sqrt{2} \times I_{rms}}{M_{pl}} = \frac{1.7772 \times I_{rms}}{M_{pl}} \tag{19.3}$$

because N is equal to one. The last two equations imply that the majority of the current passing through this filter must be close to unity power factor and the off-line regulators are in the minority. If this is not true, the value of H is higher. This core should be insulated, but some houses use quality tape. The inside diameter must be maintained so that the toroid slips over the rod easily without being too loose.

Then A_c and M_{pl} are

$$A_c = \frac{S_f W(O_d - I_d)}{2} \qquad M_{pl} = \frac{\pi(O_d + I_d)}{2} \tag{19.4}$$

where the units are in centimeters. L is

$$L = \frac{0.4U_e(O_d - I_d)WS_f \times 10^{-8}}{(O_d + I_d)} = 0.4U_e W S_f \times 10^{-8} \tag{19.5}$$

where $O_d >> I_d$. However, the M_{PL} in Eq. (19.4) is often given as

$$M_{PL} = \pi(O_d \times I_d)^{0.5}$$

Watch this: the poorer the conductive material is, the larger I_d must be to handle the current so that the bus does not have excessive heat rise. This heat rise could destroy the capacitors. The dimensions are still in centimeters. The I_d is the diameter + of the conductor, and O_d is the maximum allowable diameter based on the size of the insert or filter box. The amount of inductance needed is known because it was previously calculated. The effective permeability, U_e, approaches

$$U_e = \frac{U_m M_{pl}}{M_{pl} + gU_m} = \frac{\pi O_d}{2g} \tag{19.6}$$

where g is the gap, also in centimeters, and U_m is the core permeability. This equation is based on U_m being in the tens of thousands range. The right side of the denominator of the center equation is much larger than the left-hand term. Replace M_{pl} with Eq. (19.4) and substitute U_e into Eq. (19.5):

$$L = \frac{0.4\pi O_d W S_f \times 10^{-8}}{2g} \tag{19.7}$$

This is the individual value of each inductor used to make up the filter (six of them in the preceding case before paralleling). Changing O_d, W, and g to inches,

$$L = \frac{0.2474 O_d W S_f \times 10^{-8}}{g} \tag{19.8}$$

Solve for the width, in inches, of the inductor. For the toroid specification used here, the toroid dimensions given are O_d, I_d, and height. The W used here is the height dimension.

$$W = \frac{4.043 L g 10^8}{O_d S_f} \tag{19.9}$$

In this way, the width of the inductor core in inches (here, the width is the height of the core; there is only the conductor rod through the I_d) can be found to meet the required inductance. Make g vary with O_d so that g/O_d is a constant for each filter in parallel. In this way, W varies with L. The stacking factor, S_f, varies with the tape thickness only.

A good source for the toroid inductor, and also the C core discussed later in this section, is National-Arnold (Adelanto, CA 92301; Richard H. Wood, Engineering). Richard Wood has spent many hours doing research on pulse transformers. He has concluded that the width of the toroid and C core should be 1 inch maximum to handle pulses properly. Therefore, stack the toroids and C cores on the round rod and the bus to reach the desired width if the main feed is to be pulses, for example, the current pulses from off-line regulators or power factor correction circuits. The pulse permeability is lower than that specified, and the 1 inch width supports the pulses better, so use this technique if a high percentage of pulses are expected. Otherwise, National-Arnold makes toroids of almost any size needed up to 3 feet in diameter and width. Their loss graph in watts per pound is shown in Fig. 19.4. For better information, obtain their catalog for their tape-wound toroids and C cores.

Each toroid, or C core, is insulated from the rod or bus and from each other. In Fig. 19.5, the spacing between the cores could be filled with a heat sink. National-Arnold gaps the toroid, as in Fig. 19.6, by slitting the toroid, filling the gap, and coating the toroid so that the gap defies detection. In some cases, the toroid is cut halfway and then again cut fully across the core.

The I_d and O_d of the feed-through capacitor (Figs. 19.7 and 19.8) are the same as for the inductor just calculated. The width must also be found. (See Sec. 7.2, which is mostly repeated here.) D_s is the outside diameter, D_a is the diameter of the arbor, and t is the thickness, all in inches. D_s is the outside diameter, O_d, and D_a is the inside diameter, I_d. Make D_a, or I_d, slightly larger than the conductor rod and D_s, or O_d, the same diameter as the high-current inductor calculated before.

Solve for the number of turns, N:

$$N = \frac{D_s - D_a}{2t} \tag{19.10}$$

Solve for the mean diameter (D_m)

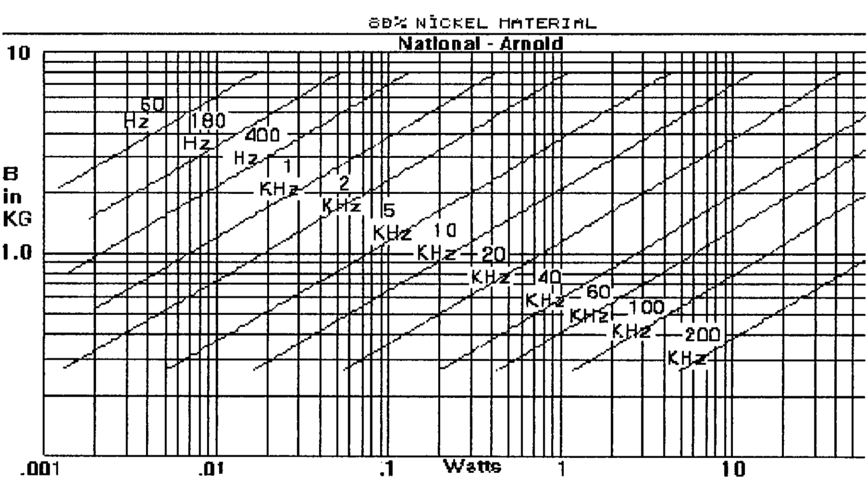


FIG. 19.4 The National-Arnold core losses in watts per pound.

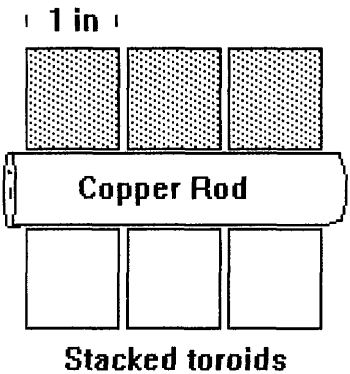


FIG. 19.5 Stacking of tape-wound toroids on the copper rod.

$$D_m = \frac{D_s + D_a}{2} \tag{19.11}$$

Solve for the mean length of turn (L_m)

$$L_m = \pi D_m \tag{19.12}$$

Solve for the active length (L_t)

$$L_t = N L_m \tag{19.13}$$

Substituting Eqs. (19.10), (19.11), and (19.12) into Eq. (19.13),

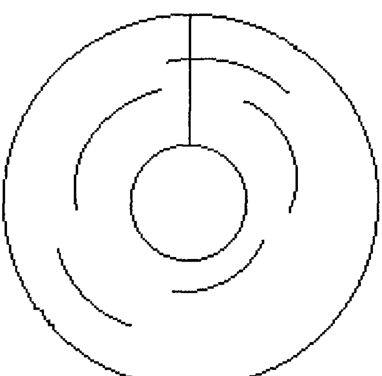


FIG. 19.6 Tape-wound core gapped by National-Arnold.

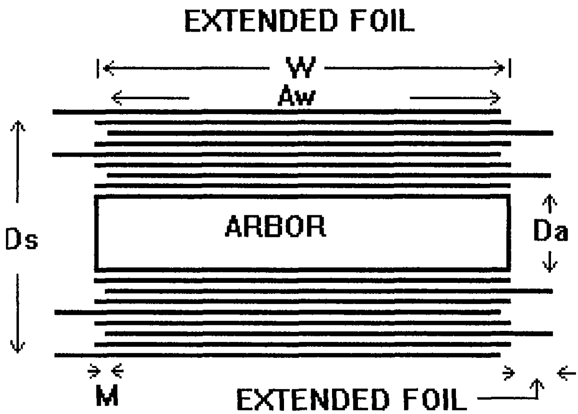


FIG. 19.7 The extended foil capacitor.

$$L_t = \pi \frac{D_s^2 - D_a^2}{4t} \tag{19.14}$$

To continue, we need to know the K value of the chosen material and find the dielectric thickness in mils. Values of K are also a function of shape, and the dielectric is often chosen on the basis of the dissipation factor and size. Mylar gives the smallest dissipation factor within its normal temperature range.

Polyester (Mylar), 900
Polycarbonate, 840
Paper (Resin or PBT), 580 (wet)

Solve for the active width (A_w)

$$A_w = \frac{CKT}{L_t} \tag{19.15}$$

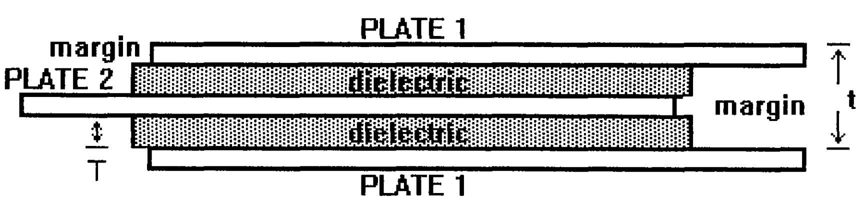


FIG. 19.8 Blowup of capacitor plates and dielectrics. T is the dielectric thickness in mils and t is the thickness of two dielectrics and two foil plate thicknesses in inches.

where T is the thickness of the material in mils, C is the capacitor value in μF, and K is the dielectric constant from before.

Substituting Eq. (19.14) into Eq. (19.15),

$$A_W = \frac{4CKT}{\pi(D_s^2 - D_a^2)} \tag{19.16}$$

If D_s, the O_d of the inductor, and D_a, the I_d of the inductor, have been decided, the width to be allotted for this capacitor can be found. This equation helps to find the width required for the capacitor. The final width is slightly larger than $A_W + 2M$, where M is the margin for corona. The extra width is to allow for the extended film foldover and sputtering. This conductive tab should be soldered to the conductive bus bar, and the same style of tab should be soldered to the ground end of the capacitor. This should be soldered to a ground lug or to the enclosure's wall. Some extra width should be allowed for these two tabs. The tabs must carry the capacitor current, not the full current of the filter section.

Figure 19.9 shows how the capacitor was slipped onto the conductor (following an inductor) and soldered to several ground points, first to the ground lugs and then the other side soldered to the center conductor. A toroid is slipped onto the conductor afterward. The inductor is coated, so voltage is not present on the surface of the core.

This continues until all the components are slipped over the conductor. Another way to mount the capacitor is to solder or sweat the capacitor to a partition that helps to isolate the compartments and forms a screen and shield. These are slipped down the conductor with the capacitor first, and the shield is soldered in place to the container. Be careful not to heat this to the point where

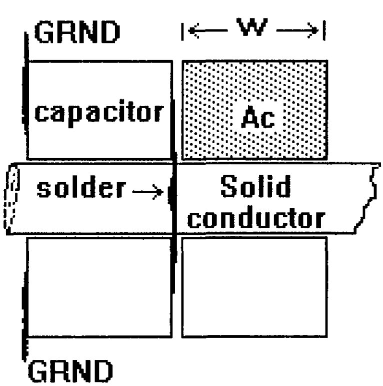


FIG. 19.9 The stacked capacitor and inductor on the copper conductor rod.

the capacitor solder melts, disconnecting the capacitor from the shield; spot welding may be better (Fig. 19.10).

The conductor rod is threaded on both ends. The rod is slipped into its container box with the top and front cover removed and slipped through a centered hole in the back cover. A ceramic insulator is slipped on over the threads and centered in the hole of the back cover. A washer, smaller in diameter than the insulator, is slipped on with the nut. This forms the output terminal. The components are slipped on one at a time over the long exposed conductor via the open front and top of the filter container. Each capacitor is soldered before feeding on the next inductor (toroid) until all the components are installed. The front cover is then installed and soldered, followed by the ceramic insulator and then the washer and nut. This forms the input side of the filter. The filter is tested and cured in an oven with a vacuum and then tested again. Often the feed-through capacitor requires additional torque to ensure proper ground and terminal contact. Plated brass rods are often used because they take the required torque. Then the top cover is soldered in place. A potting compound is often used to hold the elements in place. This unit should not require any if the components are snug to the conductor (this assumes the rod is rigid). If the rod is loose or if the specification calls for potting, use just enough in the bottom to hold the components in place or stationary. These high-current filters are heavy enough without the potting compound. To keep the filter as light as possible, use microballoons mixed with the potting compound.

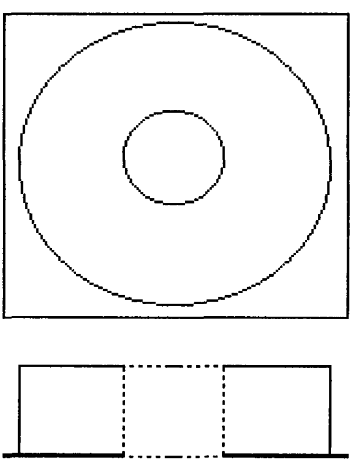


FIG. 19.10 The capacitor with ground plate.

The advantages of filters in parallel are as follows (Fig. 19.11):

1. The current reduction drops the silver-plated rod diameter by approximately

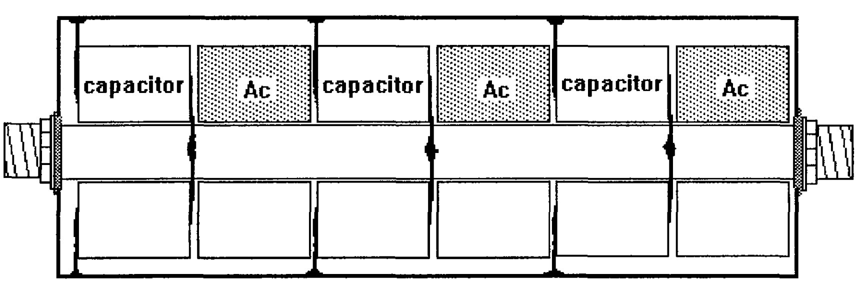
$$O_d = \frac{I_d}{N^{0.5}} \tag{19.17}$$

   where N is the number of filters in parallel, not the turns. This reduces the I_d, giving more inductance and capacitance for the O_d.
2. The inductor value becomes reasonable by being multiplied by N. The capacitor value is divided by N, making the capacitor smaller and more reasonable. This improves the capacitor's SRF and the inductor's efficiency.
3. H is reduced to H/N, reducing the size of the gap needed, where N is the number of the filters in parallel, not the number of turns. The value for turns is still equal to one.
4. The lower current filters may now be conventionally designed filters in parallel rather than the type discussed here.

The equation for the inductor width [Eq. (19.9)] repeated here is

$$W_l = \frac{4.043Lg10^8}{O_dS_f} = \frac{404.3gL_{uh}}{O_dS_f} \tag{19.18}$$

This is the initial full-sized filter, before paralleling. The value of the paralleled inductance is equal to the individual inductor of Eq. (19.18) times N, the number in parallel.



Top view of high current filter showing three sections

FIG. 19.11 The high-current filter showing three sections.

$$W_{\mathrm{ln}} = \frac{403.3NgL\mathrm{uh}}{O_{\mathrm{d}}S_{\mathrm{f}}} \tag{19.19}$$

Returning to the initial full-sized filter, before paralleling, the value of paralleled capacitance, then, is equal to the individual capacitor divided by N. This is the capacitor width divided by N:

$$A_{\mathrm{w}} = \frac{4CKTt}{\pi(D_{\mathrm{s}}^2 - D_{\mathrm{a}}^2)} \qquad R = \frac{4CKTt}{\pi} \qquad W_{\mathrm{cn}} = \frac{R}{N(O_{\mathrm{d}}^2 - I_{\mathrm{d}}^2)} \tag{19.20}$$

W_{cn} is the full capacitor width, so this is again the initial value of the capacitor before paralleling the filters. This means that W_{c}, must be divided by N. The total length of the filter—back cover to front cover—is approximately

$$W_{\mathrm{lc}} = 1.1S(W_{\mathrm{cn}} + W_{\mathrm{ln}}) = 1.1S\left|\frac{R}{N(O_{\mathrm{d}}^2 - I_{\mathrm{d}}^2)} + \frac{404.3gNL_{\mu\mathrm{H}}}{O_{\mathrm{d}}S_{\mathrm{f}}}\right| \tag{19.21}$$

where S is the number of filter elements (the number of L filters in tandem here, six) and N is the number of filters in parallel. The multiplier, 1.1, is a factor allowing for the margins, the capacitor lead thickness, the shield (if used), and some for soldering. R is found from the initial capacitance needed, in µF, before paralleling. $L\mu\mathrm{H}$ is the inductance in µH before paralleling. The inductance and capacitance were determined initially from the number of filter elements in tandem (S). S must be an even number for those folded back and even or odd for those in line.

The volume for the full filter, including all those in parallel, is the full filter length W_{lc}, from Eq. (19.21), times the O_{d} squared times the number of filters in parallel, N.

$$V_{\mathrm{tot}} = W_{\mathrm{LC}}NO_{\mathrm{d}}^2 = 1.1S\left|\frac{R}{N(O_{\mathrm{d}}^2 - I_{\mathrm{d}}^2)} + \frac{404.3gNL_{\mu\mathrm{H}}}{O_{\mathrm{d}}S_{\mathrm{f}}}\right|NO_{\mathrm{d}}^2 \tag{19.22}$$

If O_{d} is again >> I_{d}, the denominator of the first term cancels, reducing the first term to R. The second term is directly multiplied by N times O_{d} squared.

$$V_{\mathrm{tot}} = 1.1S\left|R + \frac{404.3gN^2O_{\mathrm{d}}L_{\mu\mathrm{H}}}{S_{\mathrm{f}}}\right| \tag{19.23}$$

Whatever the method, this volume remains the same for all the different numbers in parallel. Pick a combination that gives the filter a nice form factor, where the filter is not a cube or 60 feet long, 4 inches wide, and 8 inches deep. These can be folded back on themselves, dividing the length by two and twice the depth. Figure 19.12 is a nice form factor in which the length is twice the diagonal.

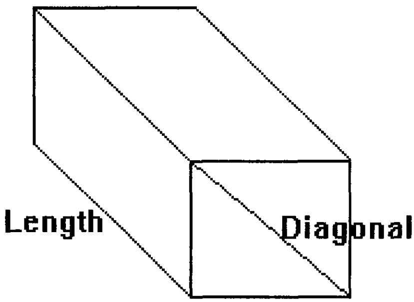


FIG. 19.12 Diagonal-to-length ratio.

The easiest method is to calculate the diagonal of the end piece compared with the length. Twice the diagonal is normal and about the minimum, and three times the diagonal is about maximum. There are three equations. The first equation is for straight, where the O_d is the depth and N times O_d the width. The second has the same width, but the length is half and the depth is twice as large or twice the O_d. This is folded back, so that the input and output terminals are on the same face (see Figs. 19.11 and 19.12). The last is the even count type, such as 4, 6, 8, and 9, which gives 2 by 2, 3 by 2, and 4 by 2, which requires the full length, and the width and depth are multiples of O_d as before. These equations cannot be differentiated to solve for a minimum because all that I have tried give points of inflection.

The three equations for the length to equal twice the diagonal follow with the assumption that $O_d >> I_d$. Any other ratios for the equations are easily established. Again, watch for higher I_d dimensions because other materials require a higher cross-sectional area than copper and the O_d may not be greater than the I_d:

$$\begin{aligned}
1.1S\left|\frac{R}{N(O_d^2 - I_d^2)} + 404.3gN^2O_dL_{\mu H}\right| &= 2O_d^3N\sqrt{N^2 + 1} \\
0.55S\left|\frac{R}{N(O_d^2 - I_d^2)} + \frac{404.3gN^2O_dL_{\mu H}}{S_f}\right| &= 2O_d^3N\sqrt{N^2 + 4} \qquad (19.24) \\
1.1S\left|\frac{R}{N(O_d^2 - I_d^2)} + \frac{404.3gN^2O_dL_{\mu H}}{S_f}\right| &= O_d^3N\sqrt{N^2 + 16}
\end{aligned}$$

The first equation is for one filter deep, or the depth is O_d, and N times the O_d wide. The second equation is for the folded-back type, so the depth is again twice the O_d, the width is still N times the O_d, and the length is half. The last is for the even count 4, 6, 8, and 9. Again, parallel filters are rarely more than six. Surely by this count the conventional-type filter design could handle this reduced current. Here, four in parallel would give 100 A each for the case listed in Table 19.3, and a conventional filter could handle this current. These form cubic equations. For the current of 400 A, the values are as in Table 19.3 for the filters in parallel. Each listing is for one through nine filters in parallel. Rarely are more than four in parallel ever used.

The single filter giving an impedance of 1.1 ohms is ridiculous. Look at the diameter of the inductors and capacitors. I am not sure that the capacitor can be wound, because most extended foil winding machines can handle about 6 to 8 inches in diameter. The inductor can be as big as 3 feet in diameter and the same width at National-Arnold in Adelanto, California. The volume is good, and so is the width (this should be length now). Two in parallel, giving 2.2 ohms each, are better, giving a smaller diameter and less width. All these are O_d depth elements and N times the O_d wide.

Figures 19.13 and 19.14 are for the folded-back type.

The rear terminals are all shorted together, and the normal metal protective shroud, which covers the entire rear, is not shown. The same is true for the front, where there is also a division between the top and bottom halves. The same idea is used for three, five, and seven (rarely used). These are all folded back as in these figures. The number of stages (S) must be even to split between the top and bottom halves of the two folded-back designs. The length is cut in half (plus some for the shrouds covering the front and rear connections). Four is a two by two, six is a three by two, and eight is a four by two, requiring the full length. Nine is a three by three, also running the full distance. The approximate sizes for the five folded-back designs are listed here. Again, long before this many filters are in parallel, the design can revert to conventional filter types.

For the five in parallel and folded back for 400 A (now 80), R is 601.38 and Q is equal to 0.3558 with each inductor = 195 μH. Each capacitor is equal to

TABLE 19.3 Chart for One to Nine Filters in Parallel

Z	*S*	F_0	*L*	*C*	I_{rms}	*dIA*	*Width*
1.1	6	4400	39	32.8	400	11.5	32.5
2.2	6	4475	78	16.2	200	8.0	35.8
3.3	6	4530	115	10.6	133	6.4	40.0
4.4	5	3850	181	9.4	100	5.8	52.7
5.5	5	3900	224	7.4	80	4.9	57.5
6.6	5	3900	269	6.2	66.6	4.4	62.0
7.7	5	3930	311	5.3	57.2	4.0	66.3
8.8	5	3930	356	4.6	50	3.6	70.3
9.9	5	3960	397	4.0	44.5	3.4	74.1

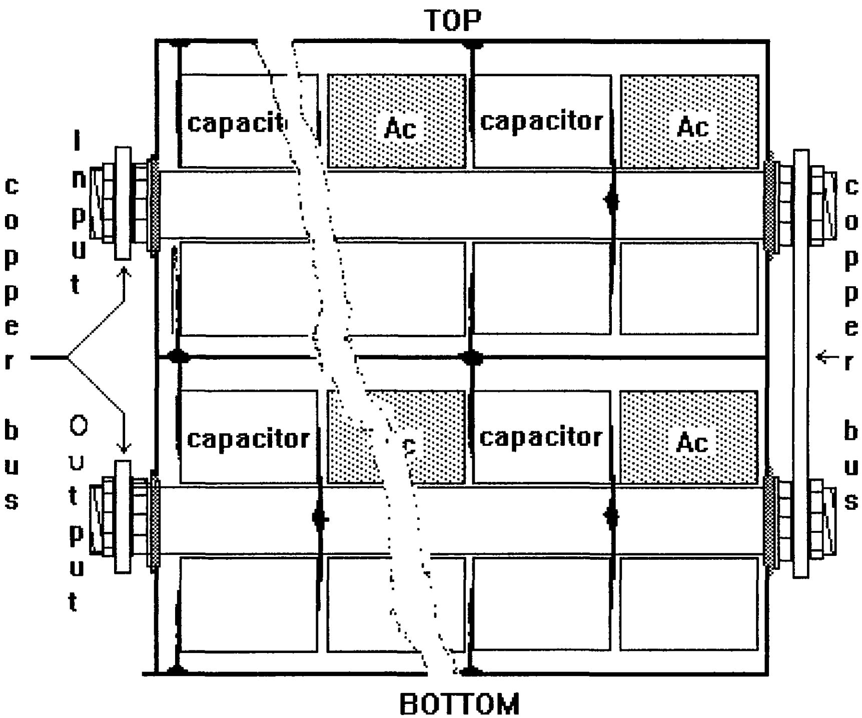


FIG. 19.13 Side view of two filters in parallel folded back.

6.6 μF. The required diameter is 4.5 inches. F_0 is 4436, which is very good. The folded length is 28.2 inches. Z is 5.5 and S is equal to 6, so the bottom and top component count is even. The volume is 2.9 ft^3. Granted, this is only an approximation, but if a 400 A filter can be built in less than 3 ft^3, this is very good. The inductors would be 1.8 inches wide, and the capacitors would be 6.5 inches wide. The capacitors may have to be split, requiring 3.27 inches each. Winding machine can handle various sizes. Extra room is allowed for soldering, width of contacts, and margins. It should be realized that the use of more than four in parallel is very rare but is shown here for comparison. Also, the filter design could revert to the conventional design techniques with this lower current of just 80 A. This high-current type of design really starts at 100 A or so, depending on the line voltage.

The same technique can be used on rectangular buses using C cores of the same material instead of toroids. The capacitors must be wound on a special rectangular arbor and so require more care while winding. It takes four to five

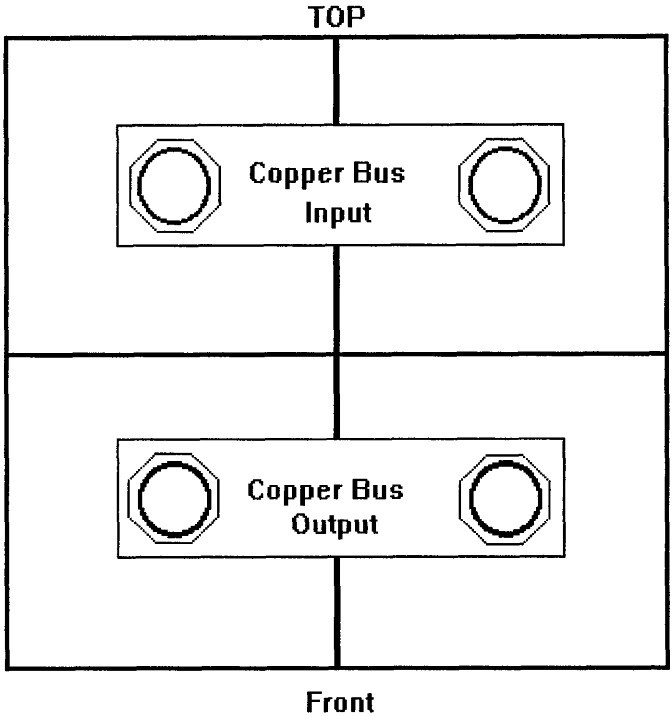


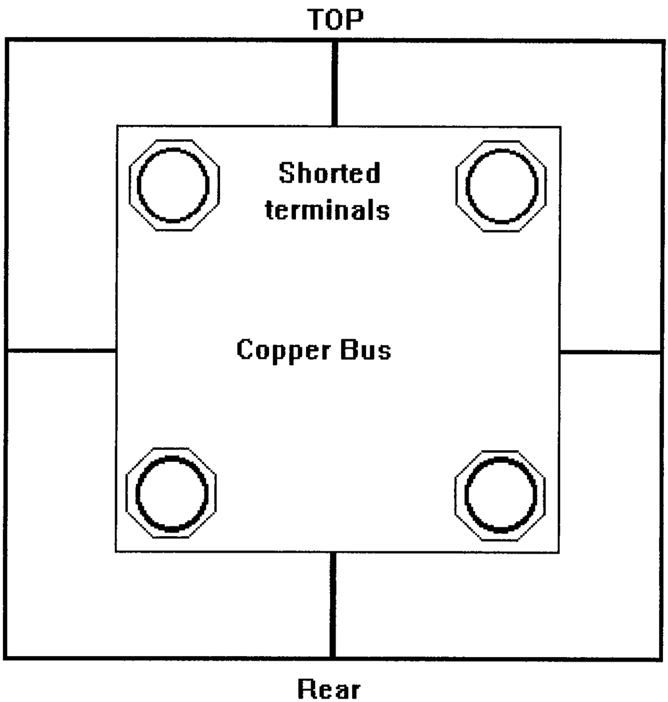
Fig. 19.14 Front of two filters in parallel folded back with rear view below.

times as long to wind them, making them much more expensive. Often the dielectric is composed entirely, or partly, of reconstituted mica because of the abrupt turns required for Mylar. This overstresses some of the dielectrics.

The strap in Fig. 19.15 is stainless steel, not strap iron as used by a designer some years ago when he ran out of stainless steel. The strap iron carried almost all of the flux around the two gaps and melted the strap in a few seconds. These cores are also available from National-Arnold.

Mount the capacitor as in Fig. 19.16 with either a shield or leads. This design technique is summarized in Sec. 19.3.

The approach just discussed should be avoided. It costs much more than breaking the filters down into smaller currents after the service panel. In this way, several filters of the same size can be used for different feeds rather than filtering the entire line. If three or four EMI filters of the same size are used, a spare is not unreasonable. The sum total of the filters plus spares may still be less than the one single filter.



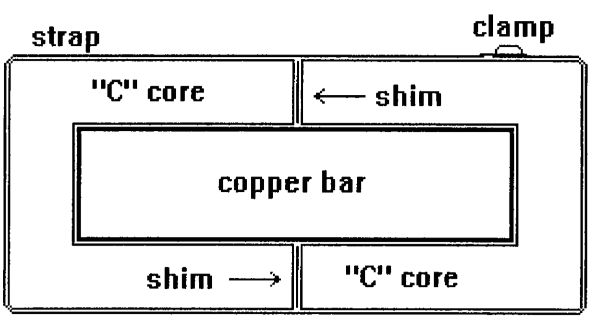


FIG. 19.15 The C core showing shims and clamp.

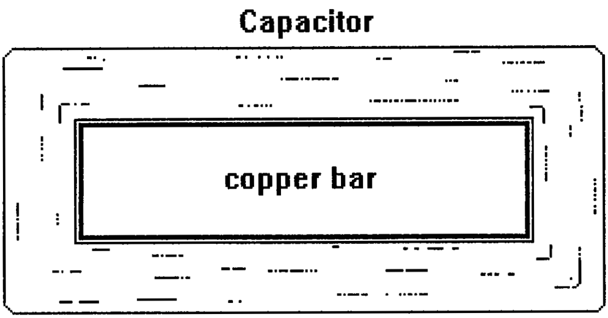


FIG. 19.16 The "square capacitor" used for copper bar.

19.2. HIGH-CURRENT SECOND METHOD

Each inductor assembly (C core) surrounds the bus, each capacitor assembly has its own rod or bus, and these are tied together at the main bus (Figs. 19.17 and 19.18).

The three capacitors are the feed-through type, but the bolt carries no current as the normal feed-through would and as the last feed-through capacitor in Fig. 19.17 would. The capacitors carry only the normal capacitor "leakage current"—really reactance current to ground—but this is not through the bolt. These three feed-through capacitor bolts are "hot." The plate that the feed-through is bolted to is ground, and the bolt is isolated from the plate. The three capacitor foils are pressed to the ground plate, giving good SRF, equivalent series inductance (ESL), and equivalent series resistance (ESR). The extra ground plate can be avoided if these three bolts are nonconductive and do not need to be isolated from the plate or the outside filter wall. The nonconducting bolts, here, would

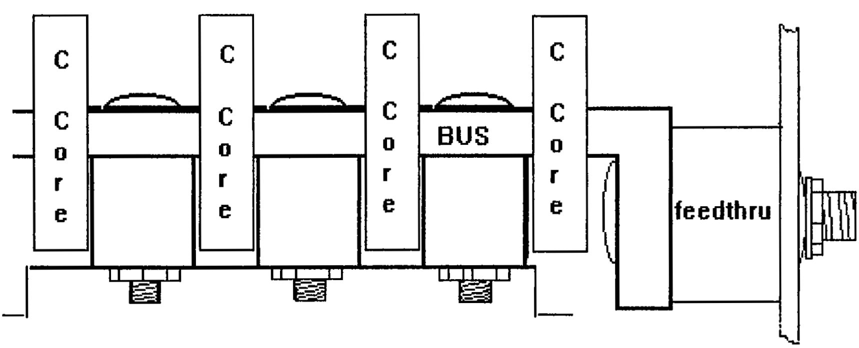


FIG. 19.17 Second method for high-current design, drawing A.

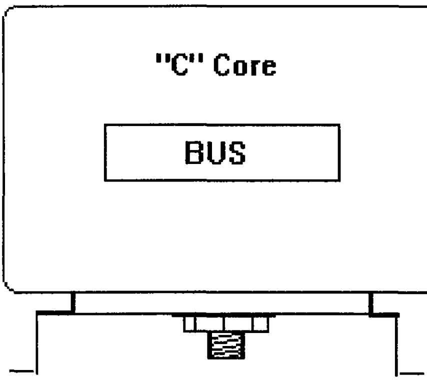


Fig. 19.18 Second method for high-current design, drawing B.

protrude through the wall of the filter and could be touched without being shocked. The advantage of this style over the style using the ground plate is that the three capacitors, along with the last feed-through, could be torqued again after the heat cure from outside the filter body. In the first style, the cover over these three bolts must be left off through the heat cure for access to these bolts. This torque is required for proper pressure for both capacitor foils of each capacitor to function properly. There are several advantages of this style over the type in Sec. 19.1:

1. In the first style, the I_d of all the capacitors had to be the same as that of the inductors. In the second style, the I_d can be very small for the three capacitors to ground. In other words, this capacitor can also be crushed or flattened in some designs. This is not true for the output feed-through capacitor. This output capacitor bolt carries the full filter current to the outside world. The bolts through the capacitors to ground carry no current (unless it is conductive and shorted) to the outside world. The output feed-through capacitor bolt must be able to carry the full load current to the outside world, but this is a much smaller capacitor value.
2. The opposite is true of the inductor. The bus can be as big as needed to carry the full filter current through the center of the C core because this second style, the inductor bus, does not pass through the I_d of the capacitor.
3. With the second style it is easier to adjust the tension required for proper capacitor action.

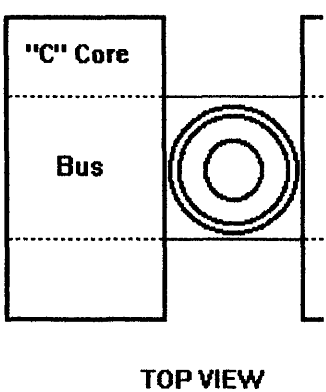
All the methods that can be developed to design these high-current filters require iterations. As the O_d grows, with inductance and capacitance values held constant for the proper design, the width of the capacitor decreases. This increase of the value of O_d increases the M_{pl}, which decreases the H—the magnetomotive force—which moves the flux density away from saturation—and increases the A_c value of the inductance. If the A_c and M_{pl} are growing at the same rate, this holds the inductance constant. This increase of M_{pl} requires a larger gap to hold U_e constant; otherwise, the inductance increases. So, around and around it goes.

The disadvantage is that the bus must be oversized to allow room for the capacitor feed. A copper bus is not very big to handle the current, 400 A here. Copper can be used here because no torque is required, but often poorer conductors with very conservative current densities are used. This gives the needed width to support the capacitors to ground but also increases the filter volume. It is a trade-off. C_m is the circular mils per ampere required and W_a is equal to F times G of the C core.

$$W_a = 7.854\, C_m I_{rms} 10^{-7} \text{ square inches}$$

The question is, can the capacitor fit on the bus? A top view of the capacitor and inductor assembly along with the equations is shown in Fig. 19.19.

The approximate value of O_d can be calculated from Fig. 19.19 because all the quantities are known. A_w should be at least the C core dimension E



$$L_t = \frac{C \cdot K \cdot T}{A_w} \quad \textbf{Active length}$$

$$N = \frac{O_d - I_d}{2 \cdot t} \quad \textbf{Turns}$$

$$D_m = \frac{O_d + I_d}{2} \quad \textbf{Mean Diameter}$$

$$L_m = \pi \cdot D_m \quad \textbf{Length per turn}$$

$$L_t = N \cdot L_m = \frac{\pi \cdot \left(O_d^2 - I_d^2\right)}{4 \cdot t}$$

$$O_d = 2 \cdot \sqrt{\frac{C \cdot K \cdot T \cdot t}{\pi \cdot A_w}} \quad \textbf{Approximate Od}$$

FIG. 19.19 Top view of C core, bus, and capacitor with equations.

(Fig. 19.20). The O_d should be, at maximum, the C core dimension G. If thin shims or washers are added to the bus end of the capacitor, the O_d can extend beyond the length of G to a maximum of G plus twice E. If the O_d of the capacitor is greater than this, the volume of the filter increases some.

For the inductor, the D dimension can be calculated from the equations just listed and knowing the gap, M_{pl}, A_c, and W_a:

$$A_c = 6.432EdS_f$$

$$M_{pl} = 5.08(G + F + 2E)$$

$$W_a = 6.452FG$$

$$A_p = W_aA_c = 41.62DEFGS_f$$

where D, E, F, and G are in inches and A_c, M_{pl}, W_a, and A_p are in centimeters. Knowing the value of the inductor, the needed width D can be found. The core manufacturer can make these cores to the needed width, D, or several cores can be stacked to reach this dimension. If the capacitor O_d dimension is less than G, the capacitors can be paralleled (Fig. 19.21).

The advantage of this technique is that the capacitors can be paralleled across the bus (Figs. 19.22 and 19.23). This reduces the volume required for the capacitors, bringing the C cores closer together and cutting the volume of the filter. The bus area of the cross section is constant for the current required to be

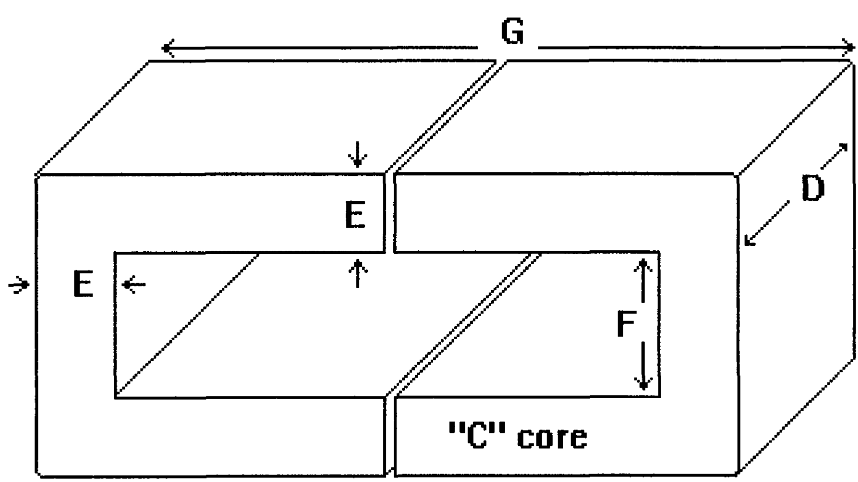


FIG. 19.20 C core with dimensions.

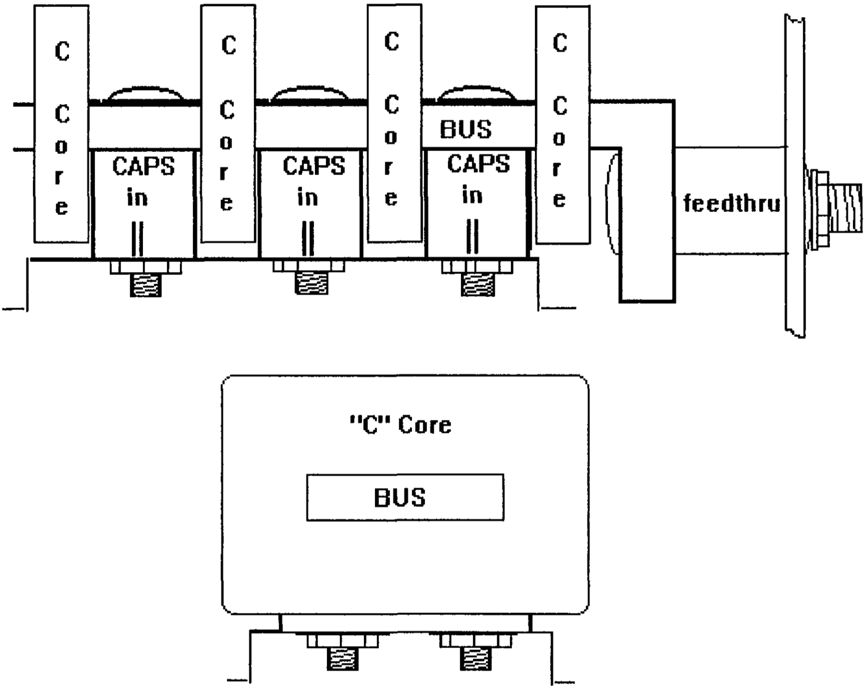
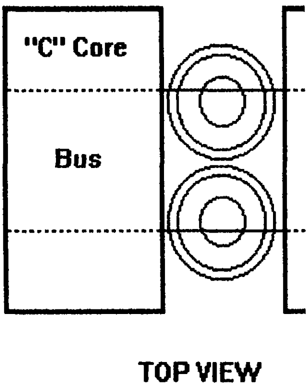


FIG. 19.21 Second method showing capacitors in parallel.

handled by the bus bar, so as the width increases for the parallel capacitors, the thickness drops. There is a limit to this because the bus must fill the C core window area (W_a). Therefore, there is a maximum ratio of width to thickness, which is limited to about 5:1. Otherwise, the C core cannot be manufactured. The cross-sectional area is a constant that varies with the current that the bus must carry. This main bus can be copper because the bus is not threaded and is not torqued as before.

Therefore, the bus is not wide enough to hold more than two capacitors across it. Otherwise, the capacitors across the bus will extend beyond the C core G dimension (window length). Also, this could extend beyond the G plus twice the E dimension ($G + 2E$ = the full length of the C cores) at both edges, increasing the volume if shims or washers are added between the capacitors and the bus. For conditions when copper must be used, some other ideas are given later.

The only difference here is that C is half the value. There is also a practical limit, and keep in mind that each hole in the bus restricts the current. Sometimes,



$$L_t = \frac{C \cdot K \cdot T}{A_w} \qquad \textbf{Active length}$$

$$N = \frac{O_d - I_d}{2 \cdot t} \qquad \textbf{Turns}$$

$$D_m = \frac{O_d + I_d}{2} \qquad \textbf{Mean Diameter}$$

$$L_m = \pi \cdot D_m \qquad \textbf{Length per turn}$$

$$L_t = N \cdot L_m = \frac{\pi \cdot \left(O_d^2 - I_d^2\right)}{4 \cdot t}$$

$$O_d = 2 \cdot \sqrt{\frac{C \cdot K \cdot T \cdot t}{\pi \cdot A_w}} \qquad \textbf{Approximate Od}$$

FIG. 19.22 Top view showing parallel capacitors and equations.

thin copper busing is placed across the bus to make up for the cross-sectional area hogged out for the capacitor bolts. This technique adds to the length of the bolt. It is done to maintain the cross-sectional area of the bus at the point where the bolt holes are drilled, keeping the bus temperature lower to lengthen the life of the capacitor.

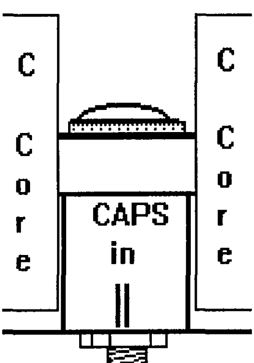


FIG. 19.23 Second method for high-current filter showing added bus.

19.3. HIGH-CURRENT METHOD THREE

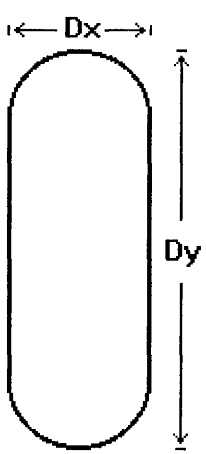
This technique gets rid of the bolt through the center I_d by using Y capacitors with very short leads. The self-resonant frequency approaches that of the feed-through type and has little added equivalent series resistance or inductance over the feed-through type. The advantages are as follows:

1. Very little bus bar area is drilled out compared with the bolts in method one or two.
2. The torque required for the feed-through type of capacitor is eliminated because the capacitor is fully assembled and stressed prior to installation in the filter.
3. The leads are soldered, removing the necessary good contact with the bus bar or ground plane.

The disadvantage is that the capacitors can rarely be reused because of the difficulty of getting the short leads back in the holes in the ground plane.

The approximate equations needed to fit the capacitor are listed in Fig. 19.24 and continue in Fig. 19.25. A_W is the length of the capacitor in the E direction and is longer than E to keep the ground plane away from the C core inductors. The C cores are insulated, but it is still not a good idea to be pressing or touching ground. Thus, A_W is equal to E plus 1/8 inch or longer. This eats up volume but reduces the D_X dimension. The values of C, K, T, and t are known,

The active length Lt of pressed capacitor



$$D_a = \frac{2\cdot(D_y - D_x)}{\pi} = \text{the Arbor Diameter}$$

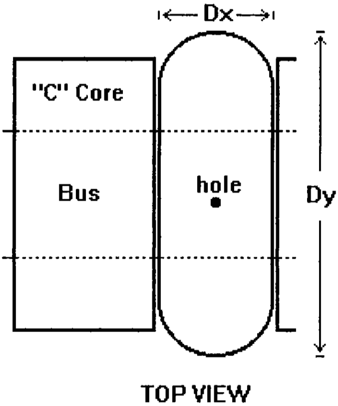
$$N = \frac{D_x}{2\cdot t} = \text{Number of turns}$$

$$D_m = \frac{D_x}{2} + D_a = \text{the Mean diameter}$$

$$L_m = \pi\cdot D_m = \text{The mean length per turn}$$

$$L_t = L_m\cdot N = \frac{1}{4}\cdot(D_x\cdot\pi + 4\cdot D_y - 4\cdot D_x)\cdot\frac{D_x}{t}$$

FIG. 19.24 The pressed capacitor and equations.



$$\frac{C\cdot K\cdot T}{A_w} = \frac{D_x}{4\cdot t}\cdot\left[4\cdot D_y-(4-\pi)\cdot D_x\right]$$

$$D_x = \frac{2.33\cdot A_w\cdot D_y - 2.33\cdot\sqrt{A_w^{2}\cdot D_y^{2} - .8584\cdot C\cdot K\cdot T\cdot t\cdot A_w}}{A_w}$$

$$D_a = \frac{2\cdot\left(D_y-D_x\right)}{\pi} = I_d \qquad O_d = D_a + D_x$$

FIG. 19.25 The pressed capacitor. Top view of C core and bus with equations.

so L_t can be solved from the equation in Fig. 19.24. Remember, t is the thickness of two dielectrics and two foils in inches; see Fig. 19.8.

Also, T is the dielectric in mils. Then D_x can be solved from the quadratic equation given in Fig. 19.25. The last half must be smaller than the first half under the square root. This also ensures that the last half of the quadratic is smaller than the first term. The negative solution is the only valid solution; otherwise, D_x would be larger than D_y. Once D_x is known, D_a, the arbor diameter, is solved and O_d also follows. In Fig. 19.25, the value is shown to be larger than $G + 2E$ of the C core. This is okay if D_y is smaller than the container with extra clearance. Again, none of these Y capacitors should touch the container wall.

19.4. REVIEW OF HIGH-CURRENT FILTERS

Most high-current filters are made with pieces of these various technologies. Some extend the width of the conductor bar for the capacitor(s). Others use the vee technology for the capacitors with shunts across the vee legs to carry the main current. Others mount the inductor sideways and reverse each section so that the capacitors are on opposite ends (Fig. 19.26).

The two wires to and from the capacitor (Fig. 19.27) across the two buses are in parallel with the single wire directly between the two buses. These wires must carry all the current between the two buses. Otherwise, the diameter of the wires must increase. This is veeing the capacitor. The inductance in both the capacitor legs adds to the inductance in the C cores if the two vee wires can handle the current. This keeps the SRF of the capacitor high. Otherwise, the shunt wire inductance is in parallel with these two vee wires (Fig. 19.28). With the

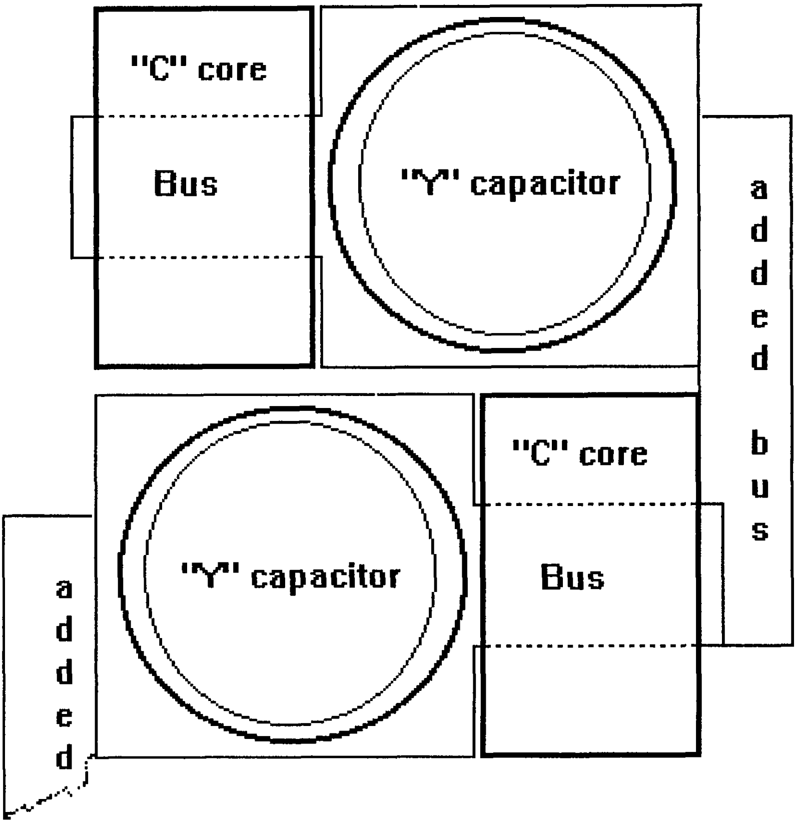


FIG. 19.26 High-current filter, alternative method.

parallel shunt bus wire, The SRF of the vee capacitor is still higher than in the situation in which the capacitor has single long leads.

The main thought is to avoid these high-current filters if at all possible. If the designer cannot avoid them, parallel the filters so that conventional lower current filters can be used, avoiding the high-current design. This section is intended to give the designer some thoughts on the design that will help if this situation cannot be avoided. The equations listed earlier are marginal but should get the designer close so that, with minor modifications, the filter will be complete.

Another alternative method is to replace the feed-through capacitor with pressed or flattened Y capacitors mounted on a bus connected between the two buses running through the C cores (Fig. 19.29). The width of the capacitors should be about the same as that of the bus, and the buses carry all the current except for

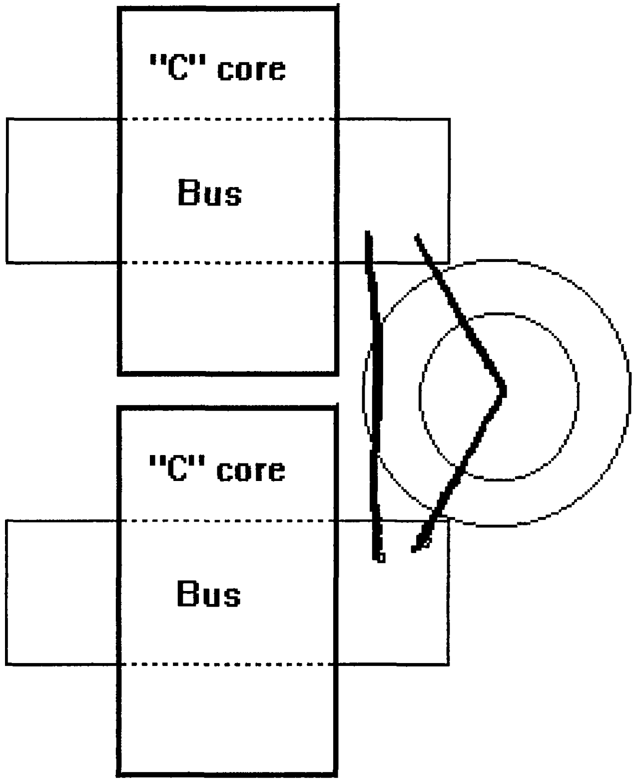


FIG. 19.27 High-current filter, alternative method two.

the leakage current through the capacitors. Several capacitors can be mounted on the bus, giving each capacitor a higher self-resonant frequency. Other advantages are that the overall volume is smaller, the total capacitance is the sum of those along the bus, and the Y capacitors do not require torquing after the filter is heat cured. The output feed-through requires torquing after the filter curing for best capacitor results. If the capacitors along the bus can be spaced a little wider, allowing ample room to solder the leads, capacitors may be on both sides on the bus. The upper capacitor bus hole would be between two capacitors on the bottom, for example. The D_x dimension of the capacitor would be narrower, allowing the bus also to be narrower. The cross section of the bus must be the same to carry the full current. The bus can be cut out of thick copper sheet or bolted together. In this way, the bus between the C cores may be a different size than the bus through the C cores. Various alternative methods may be developed using many of these ideas in combination depending on the situation.

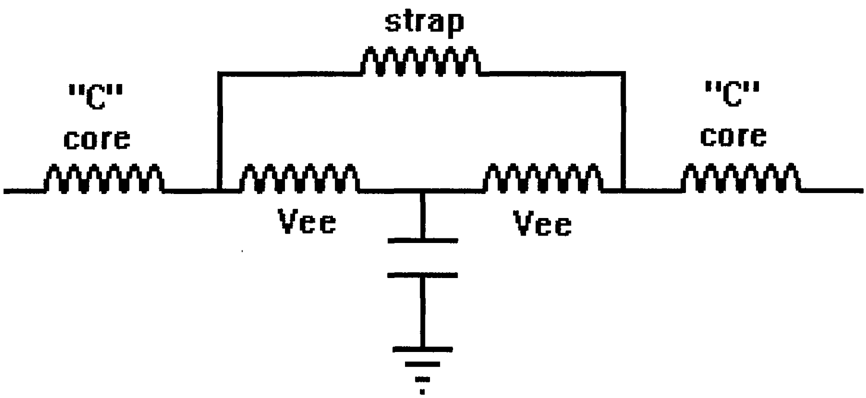


Fig. 19.28 The vee capacitor in parallel with shunt wire.

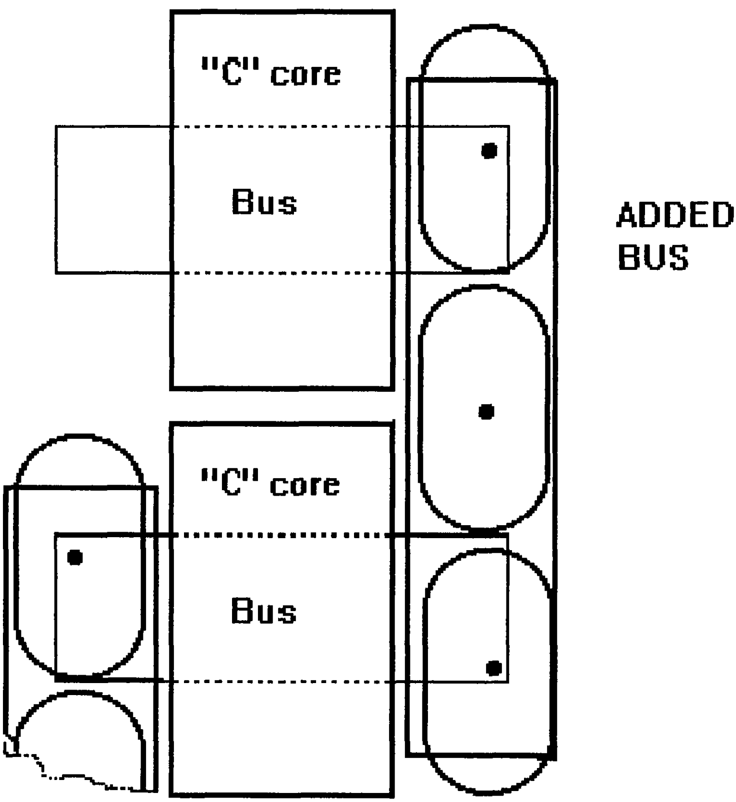


Fig. 19.29 The last alternative high-current filter.

Using the preceding suggestion, the inductance along the bus adds to the overall inductance and is similar to veeing. The capacitor lead length is low, making ESL and ESR very low and keeping the SRF high. The disadvantage is that all the capacitors are lost during repair or when adjusting for insertion loss if components must be added.

19.5. THREE IN PARALLEL

In some cases, three filters have been paralleled. In one case, 1000 A was required and three 333 A filters were paralleled to accomplish this (Fig. 19.30). High loss was required, and four π sections were needed for the loss. Several inductors were tuned to meet the low-frequency loss (Cauers), and large capacitor values were stacked between the inductors—tuned or otherwise. Tuning works best when low-impedance capacitors to ground are located on both sides of the tuned inductors. Large copper buses shunted the three sections together on the input and output terminals.

The capacitors were GE 20 μF, 480 V AC, and eight of these were used along with three feed-through capacitors at 5 μF. These feed-through capacitors were used to connect the input and output terminals along with the central shielded sections. This was required because 100 + dB was required at 14 kHz and above. To get 100 dB and maintain it through 1 GHz requires shielded, or isolated, sections, and feed-through capacitors do this job well.

The input and output section is one 5 μF feed-through and one 20 μF GE. The 40 μF sections are two 20 μF capacitors, and the 45 μF is two 20 μF plus a 5μF feed-through capacitor. This last section is where the shield appears across the enclosure. This separates the two filter sections. Dirt in one chamber cannot easily be transferred into the next. This technique is necessary to maintain 100 dB and below. This filter holds the 100 dB well through 1 gHz.

Sum the total capacitance to ground. Here, this is 175 μF. At the line frequency, the inductors are out of the picture. Divide the total capacitive

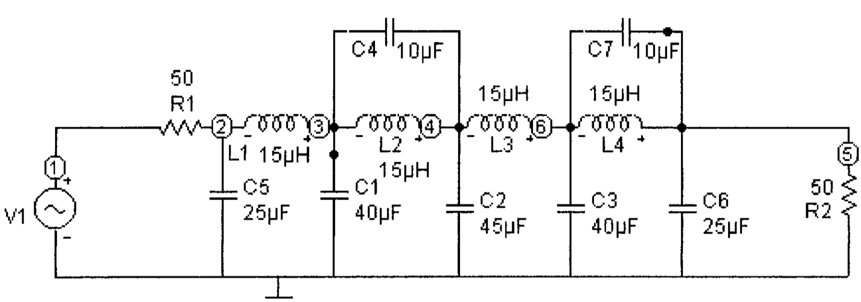


FIG. 19.30 One of the three filters.

reactance into the line voltage to get the reactive current, which is 90 degrees out of phase with the line current. The angle of lag can be determined, and the square root of the sum of the squares gives the hypotenuse and it is surprising, even with all this capacitance.

$$I_C = 2 \times E_{RMS}\pi FC = 2 \times 480\pi 60 \times 0.000175 = 31.66$$

So the leakage current is 31.66 A. The full current is

$$I_T = (333^2 + 31.66^2) = 334.5$$

The angle of lag, in degrees, is

$$\Theta = \tan^{-1}\left(\frac{31.66}{333}\right) = \tan^{-1}(0.095) = 5.431$$

The leakage current cannot be specified in this type of filter.

19.6. CONCLUSION

The conclusion is obvious! Try to avoid these high-current types if at all possible. These types are most often required for secure rooms, vaults, and EMI test laboratories. In the latter, high current may be required to drive equipment under test so one cannot use the redistributing principle as described throughout this chapter. For vaults and secure rooms, redistributing principle can be used. Bring in the power, apply it to the breakers, and filter each leg following the circuit breakers. Granted, this area must be isolated from the rest of the secured area via copper walls similar to those in the screen room.

20

Packaging Information

Packaging is a very important subject and is as critical in the EMI world as it is in the rest of the electronics arena. Component layout is important in all electronics. Components that are susceptible to H fields must be moved away from these fields, yet wiring or printed circuit board traces must be short as possible.

20.1. THE LAYOUT

The physical layout of the EMI filter for best performance is long and thin. From Fig. 20.1, it is easy to see that the full length is much greater than the height and width. In the EMI filters shown in Figs. 20.4 and 20.5, the toroids are spaced apart by the value of the capacitor. This removes the tendency toward cross talk or mutual inductance by increasing the distance between the inductive components. Also, note the distance between the dirty or clean end and the clean or dirty end. The output terminals are on the far end, away from the input connector end. In Fig. 20.2, the output terminals are not shown. They are on the blind side.

If the input and output must be on the same face, a shield should run almost the full length of the filter as shown in Fig. 20.3. The components are divided between the two halves. Half run from the front to the back half and then double back to the front on the other side of the shield. The shield must be fully grounded for the entire length.

In the lower current mode, these parameters still hold true. The filter's aspect ratio should be long in length compared with the height and width. The components should run from one end to the other, as in a transmission line, rather

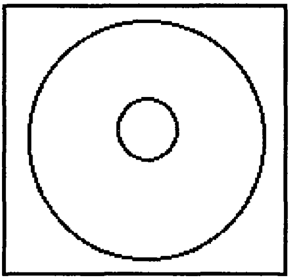


Fig. 20.1 The high-current insert showing front cover.

than hop back and fourth within the filter body. This reduces the cross talk effect, and Capcon should be used to cover lead wires to help attenuate the upper frequencies. The inductors should be mounted in quadrature—yes, even toroids—as shown in Fig. 20.4 and the alternative Fig. 20.5. The alternative requires more room, but all the toroids are in quadrature; two toroids in Fig. 20.4 are not, but they are separated by additional distance. The upper left and the lower right are in the same plane, with a capacitor between, but the distance is farther, which reduces the magnetic coupling.

The top two inductors would be wired directly to the line-to-line capacitor and the bottom two also wired directly to the capacitor using the vee technique. See Sec. 7.3. This continues to the next section, still maintaining the quadrature of the inductors. These are not usually mounted to a printed circuit board because of the current and the DC resistance of the boards. Most EMI houses do not design their filters to utilize printed circuit boards.

Some have designed the EMI filter as part of the power input, and this is proper if the filter is mounted in a container that is a good conductor and is grounded to the third green wire so that the feed-through capacitors can

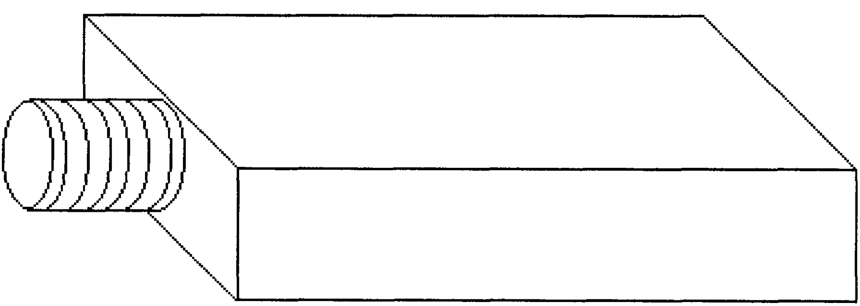


Fig. 20.2 Filter showing good length-to-width ratio.

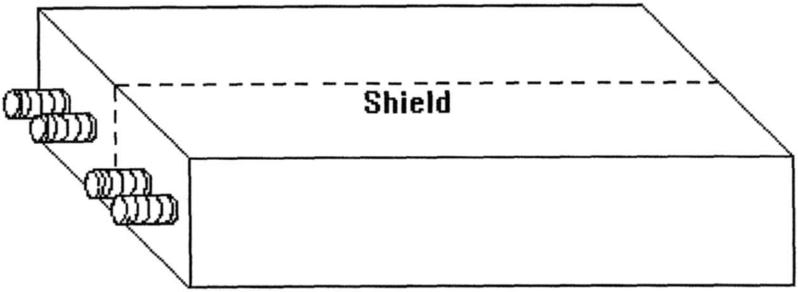


FIG. 20.3 Filter showing shield to separate clean and dirty areas.

function properly. This works well if the needed loss is for Federal Communications Commission (FCC) standards or if the required loss is low. Other designers have placed the filter within the power supply using open, or exposed, components. This technique rarely works. The filter must have shielded components to function properly or other magnetic fields either are influenced by the filter or couple their magnetic field to the inductors of the filter. This way, a 60 dB filter is now only 24 dB and the filter designer does not know what went wrong.

The case or container of the filter must be a good conductor. The better this surface conducts, the lower the magnetic field is on the outside of the case. Even though cold-rolled steel is often used for the filter body, the container is often silver plated to enhance the conduction in filters for military or other groups requiring severe loss. This attenuates the *H* field and improves the radiated

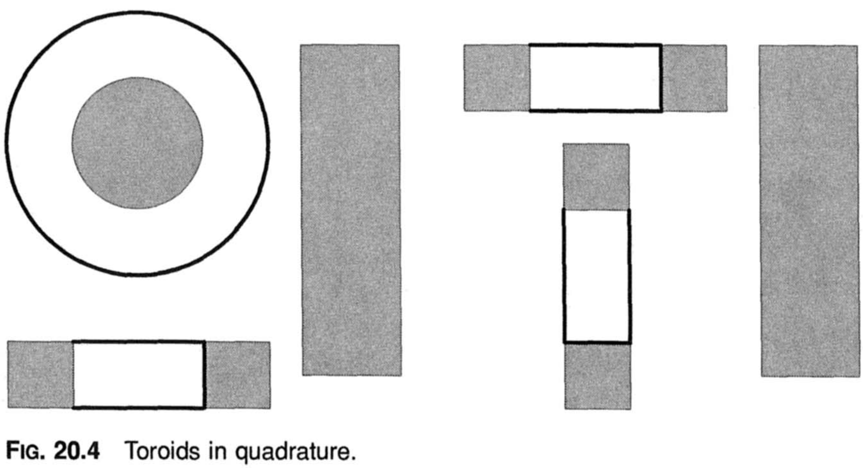


FIG. 20.4 Toroids in quadrature.

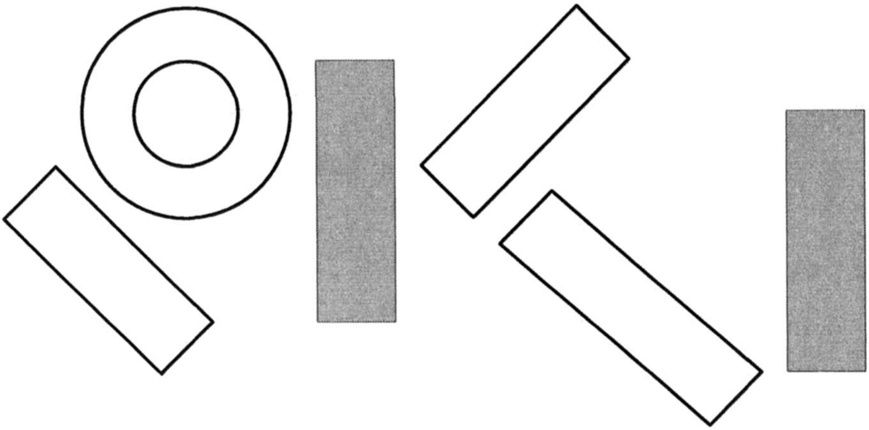


FIG. 20.5 Alternative method—balanced quadrature.

emissions. The *H* field establishes a current on the surface of the filter body container. The better this surface conducts and the thicker the case material, the weaker the current on the other side of the case. This reduces the *H* field departing from the case wall and is true for *H* fields either out of the filter or into the filter. The container must also be a good conductor so that the feed-through capacitors can function. The same would be true if the good conducting case was not wired to ground or if the ground wire was loose or missing.

In the circuit of Fig. 20.6, neither the feed-throughs, C2, nor the two common mode arresters work properly if the case ground is resistive or if the

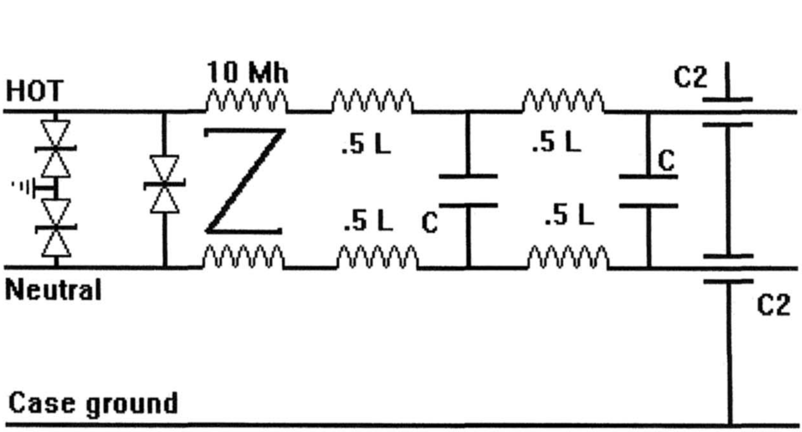


FIG. 20.6 Arresters and feed-throughs to ground.

container has a high resistance. The one thing that we can say about ground is that it isn't ground, or as most say, *ground isn't.* So, grounding of these filters cannot be marginal or overlooked.

20.2. ESTIMATED VOLUME

The approximate volume of the higher current filter was discussed in Chapter 19. The equation is repeated here. S is the number of filter elements (the number of Ls, Ts, or πs). The equation was based on Ls. O_d and I_d are the outside and inside diameters in inches.

$$L = \frac{0.4U_e(O_d - I_d)WS_f10^{-8}}{(O_d + I_d)} = 0.4U_eWS_f10^{-8}$$

where L is the inductance and S_f is the stacking factor. For the capacitor, R is

$$R = \frac{4CKTt}{\pi}$$

where C is the capacitance in μF, K is the dielectric constant, and T and t are both thicknesses but T is in mils. This was calculated on the basis of Ls, but Ts and pi filters should still be close. These are approximations anyway.

$$V_{tot} = = \left| \frac{R}{N(O_d^2 - I_d^2)} + \frac{404.3gNL_{uh}}{O_dS_f} \right| NO_d^2$$

If O_d is much bigger than I_d, V_{tot} approaches

$$V_{tot} = 1.1S \left| R + \frac{404.3gN^2O_dL_{uh}}{S_f} \right|$$

On the other hand, Col. W. T. McLyman has provided the following data, combined in Table 20.1, to calculate many different magnetic properties from his book *Transformer and Inductor Design Handbook* (Marcel Dekker, New York). Our goal here is volume and in the next section volume to weight. See also Chapter 14.

The main idea is to find the area product, A_p, in centimeters to the fourth power; from this, and knowing the core, the approximate volume can be found (Fig. 20.7). According to McLyman, in the method listed in his book, the energy of the inductor must be determined. The equation is

$$\text{ENG} = \frac{LI_p^2}{2}$$

TABLE 20.1 McLyman Magnetic Factors

Core type	K_j25°C	K_j50°C	x	$K_v C_m^3$	K_w (g)
Pot core	433	632	1.20	14.5	48.0
Powder core	403	590	1.14	13.1	58.8
Laminations	366	534	1.14	19.7	68.2
C core	323	468	1.16	17.9	66.6
Single coil	395	569	1.16	25.6	76.6
Tape wound	250	365	1.15	25.0	82.3

The type of core to be used must be known to use the Table 20.1. K_j and X come from Table 20.1. K_u is the winding factor—0.4 for a toroid—and B_m (in teslas) must be known for the core type. The A_p is

$$A_p = \frac{|2(\mathrm{ENG}) \times 10^4|^x}{B_m K_u K_j}$$

Find the A_p for the different sizes of inductors, and add the different A_p values for all the inductors for the total A_p. For example, a three-stage balanced L filter would require six inductors, all of the same value. Find the energy based on the peak current, and from knowledge of the core type, oabtain the components for the A_p.

If the peak current is 5 A and the inductors are 250 μH, ENG is

$$\mathrm{ENG} = \frac{250 \times 10^{-6} \times 5^2}{2} = 0.0031$$

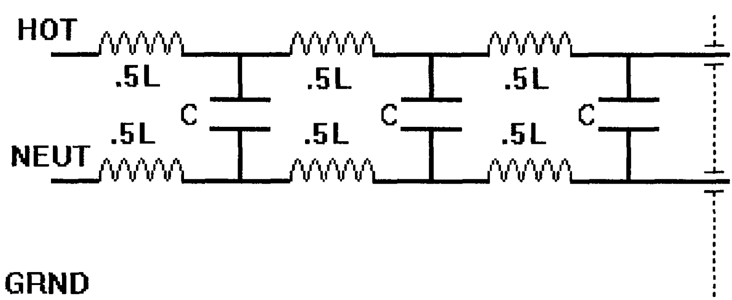


FIG. 20.7 Using the McLyman parameters for size and weight.

If MPP powder cores are the choice, then $K_u = 0.4$, K_j at 25°C = 403, $B_m = 0.7$ tesla, and $X = 1.14$. The A_p follows:

$$A_p = \frac{|2 \times 0.0031 \times 10^4|^{1.14}}{0.7 \times 0.4 \times 403} = 0.9792$$

Round this up to 1 and find the total A_p and then the volume:

$$V_{L_{tot}} = 1\times\ 13.1 \times 6 = 78.6C_m^3$$

where 13.1 comes from Table 20.1 for the MPP core for cubic centimeters. The capacitors are not included, but the weight ratio of the capacitors is of the order of that of the inductors. The total would be 160 cm^3, and this volume is only 60% utilized. This gives 267 cm^3, but allow for the container, feed-throughs, and wiring and round this up to 280 cm^3. This equates to 17 $inches^3$. These are rough estimates that should get the design engineer in the ballpark.

20.3. VOLUME-TO-WEIGHT RATIO

Bob Hassett at RFI Corp. (now retired) has done some research on the size-to-weight ratio for EMI filters. This was primarily done on the tubular types of filters mentioned in an earlier chapter. The ratio is 1.5 ounces per cubic inch. This equates to 1.6 pounds for the filter discussed in the preceding section, which needed 17 $inches^3$.

From McLyman, again, use his area product and use the same A_p as in the preceding section for one inductor. Multiplying this value of 58.8 for each inductor by the six inductors, as listed for the powder cores in grams, gives

$$58.8 \times 6 = 352.8 \text{ g}$$

Doubling this for the capacitors, wiring, feed-throughs, container, and input terminals,

$$352.8 \times 2 = 705.6 \text{ g}$$

The weight in pounds is

$$\frac{705.6}{459.53} = 1.535 \text{ pounds}$$

This looks like the same ballpark that RFI Corp. attends, even though McLyman is at NASA Jet Propulsion Laboratory in Southern California and Bob Hassett is in Bay Shore on Long Island, New York. Anyway, their two methods are close to agreement.

20.4. POTTING COMPOUNDS

Potting compounds add substantial weight and so should be used sparingly. Most think that these enhance cooling, but most potting compounds do not aid this function. In fact, they often hinder heat transfer. This could be a desirable feature to avoid heating sensitive components. If heat transfer is the goal, some compounds have this feature, but they are often too expensive or not readably available. Often, fine granules of aluminum are added to the epoxy to enhance the heat transfer. Most often, potting is added to the filter to tie down the components. Do not fill the enclosure with the potting material. Cover the bottom to a height necessary to support the components.

21

Design Examples

21.1. SOUTHEAST ASIA FILTER FOR THE NAVY

A company in the Philippines wrote a marginal specification for a 400 Hz EMI filter for the navy. Then they started the typical requests for changes, expecting all the rest of the specification to remain the same. Can you make it smaller and lighter etcetera? I do not know how anyone can think that weight and size can be cut and still keep the same features. So some of the initial things were trimmed.

The filter had capacitors tied from line to line, doubling the capacitor values to those shown in Fig. 21.1. This gave a resonant rise (Fig. 21.2) of over 11 dB at 1.56 kHz and a 0.58 dB rise at 400 Hz. A 0.58 dB rise does not sound like much, but the equation is

$$10^{0.58/20} - 1 = 1.069 - 1 = 0.069 = 6.9\%$$

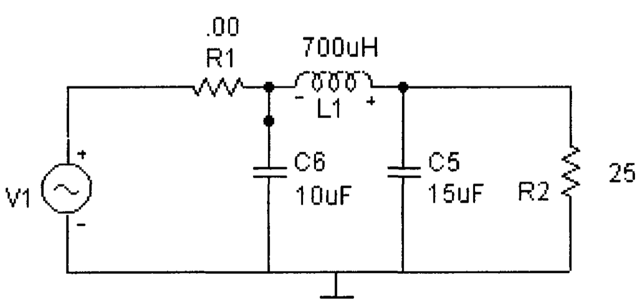


FIG. 21.1 The simplified 400 Hz Asian filter.

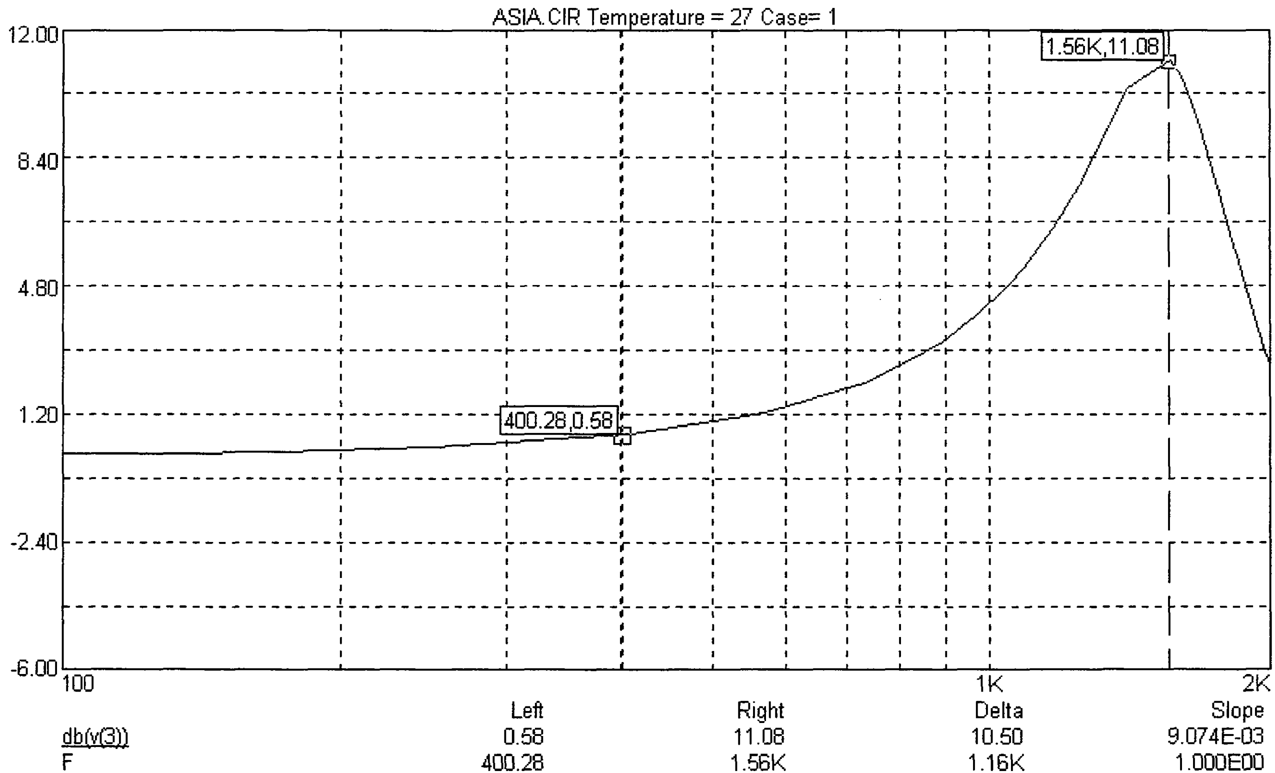


FIG. 21.2 The resonant rise.

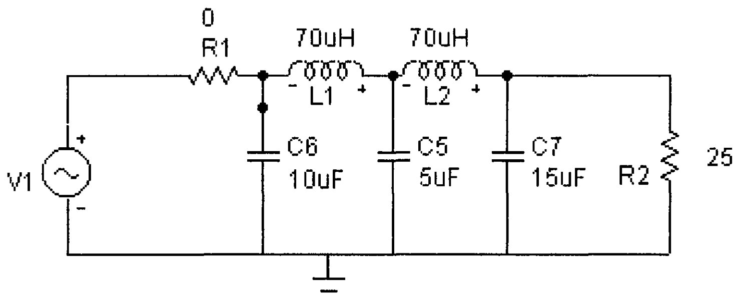


FIG. 21.3 The Asian navy fix.

This is a heavy percentage increase. The value would be larger in the real world. The fix is shown in Fig. 21.3, but the capacitors had to remain the same. Note that the center capacitors do not add because of being isolated by the inductor in the other two legs. Both filters have 60 dB at 20 kHz using the 220A specification.

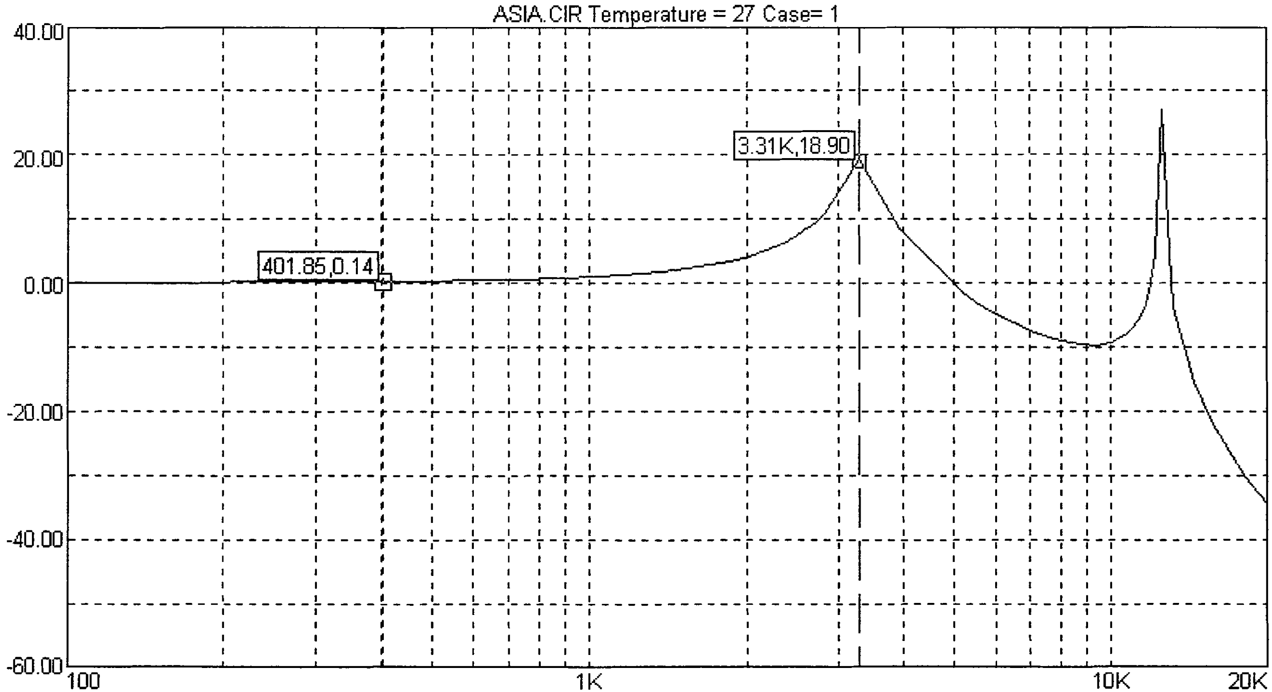
The resonant rize has increased to 18.9 (Fig. 21.4), but so has the frequency. The dB increase at 400 Hz is now 0.14. Using the same equation,

$$10^{0.14/20} - 1 = 1.0162 - 1 = 0.0162 = 1.62\%$$

The 1.62% is ideal and the real rise percentage would be more but should be less than 3%. This was more of a repair than a new design, but it could have been improved greatly with a completely new design. This is the big story regarding 400 Hz filters, especially three-phase 400 Hz EMI filters. Note the two resonant rises in Fig. 21.4. This shows that the Q is still too high.

21.2. THE FAULTY 400 HZ SOURCE

Some years ago, a company on Long Island required a 400 Hz three-phase EMI filter. Some time later, they stated that the filters were overheating. The conclusion was that the capacitor current to ground—leakage current—was excessive. The line frequency was 400 Hz, so power factor correction coils were suggested. A doghouse was added to the filter body, and the three inductors were wired in and returned to the customer. Some weeks later, the customer was livid because these filters were heating with minimal load current. It was so bad the doghouse cover bowed and broke the solder loose. The customer's test people were complaining, but there was no complaint from the field. These filters were working fine. When we loaded the filter, it became hot but nothing like their report. We were using a 400 Hz motor generator to drive our filter and a resistor load bank The customer was using an electronic three-phase 400 Hz source to power our filter and their load.



	Left	Right	Delta	Slope
db(v(4))	0.14	18.90	18.77	6.459E-03
F	401.85	3.31K	2.91K	1.000E00

Fig. 21.4 Improved frequency plot.

There were two problems, and neither group knew of either problem. Their 400 Hz power source was rich in 2400 Hz. This was used to derive their 400 Hz. Our filter had a resonant rise at 2400 Hz. Who knows the current we were pushing through our capacitors to ground?

This is the problem of a few components to make the filter, especially for 400 Hz and, again, especially for three phase. This filter was a single π. Granted, this filter met the 220A test specification, but it should have been at least a double π. The resonant rise of the previous section (21.1) is still not high enough at 3.31 kHz after the fix, but 2.4 kHz in this section is even worse. Both of these sections must be pulled up to over 4 kHz to avoid these problems.

21.3. THE ROUND ROD FILTER IN CHAPTER 19

The filter (Fig. 21.5) was specified for 60 A with loss at 1 MHz of 80 dB. This was a good spot for Chapter 19–type technology. This was also a 400 Hz three-phase four-wire design. It required only a single π with two feed-throughs, one on each end. A bolt ran through the two capacitors with space in the middle. These were two MPP toroids where the threaded rod just fit through the I_d. The outer diameter of the MPP core had to be less than the two feed-through capacitors' outer diameter. The values needed are two 0.6 μF feed-through capacitors and two MPP cores totaling 5 μH.

The 220A test results showed this filter to be "flat lined" at 80 dB well into the GHz range (Fig. 21.6). These small feed-throughs and the MPPs are high-frequency devices, which accounts for flat lined loss. It remained flat because it was single chambered.

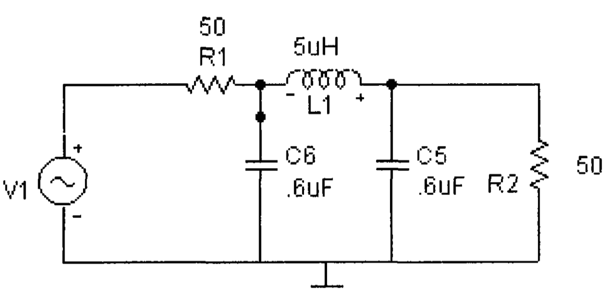


FIG. 21.5 The high-frequency rod filter.

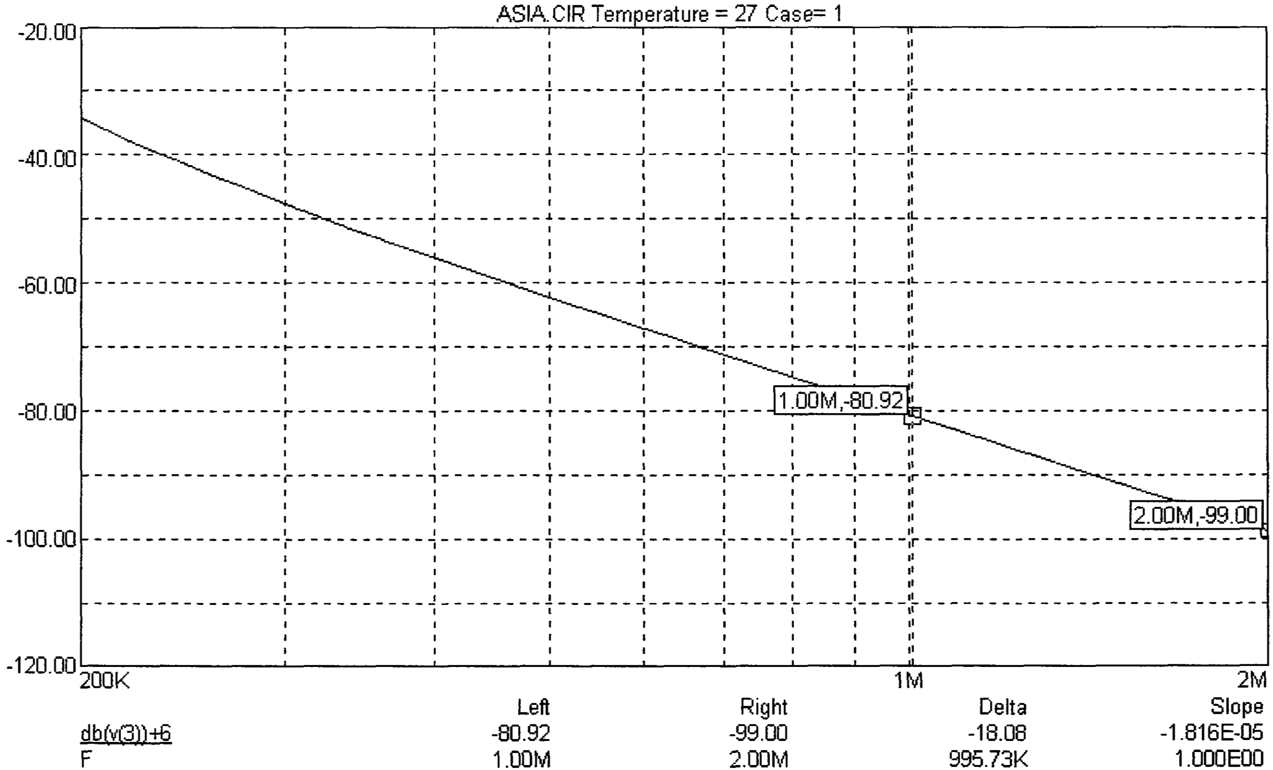


FIG. 21.6 Loss of the rod filter.

22

Questionable Designs

The drawings in this chapter have been around for years, and I have no idea of the initial group or company that designed them. The reason that they are here is for your critique, and the hope is that you come to the same viewpoint that most filter people agree with.

22.1. 28 VOLTS AT 35 AMPS

The high current in Fig. 22.1 does suggest low values of inductance and high values of capacitance. C1 at 88 μF is far too high and would be, at best, resistive, inductive at any reasonable switcher frequency, and yet this filter has four of them. They add weight, volume, and cost with little EMI benefit except at the very low-end frequencies. No doubt this is electrolytic and resistive by 30 kHz.

The RC shunt, composed of the 10 ohm resistor and the 8.1 μF capacitor, kicks in at 2 kHz, showing that the switcher frequency (or the current pulses) is well above this frequency, as are most nowadays. The resistor should be nearer the design impedance of 0.8 ohm. This would drop the capacitor value to 1 μF, again reducing the weight, volume, and cost but also raising the self-resonant frequency (SRF). The 10 ohm resistor and the 8.1 μF capacitor along with the 0.8 ohm and the 1 μF would give little help to the 6.8 μF feed-throughs. Although the 0.8 ohm and the 1.0 μF would offer higher impedance, it is not that much more. The 6.8 μF almost stands alone.

The inductors add little common mode loss to the filter. In most applications, the Zorro would be at least an order of magnitude above the 15 μH inductor in the circuit in Fig. 22.1. In the figure, the two differential inductors contribute

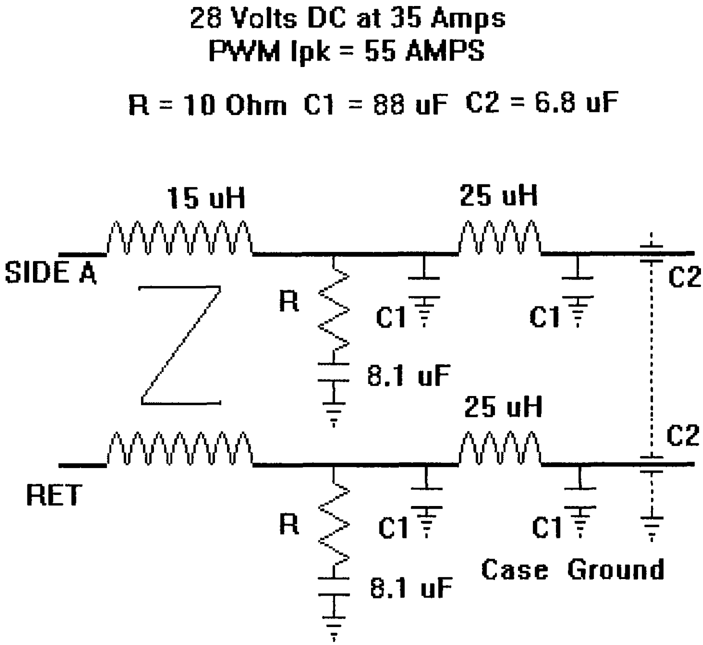


FIG. 22.1 DC filter, 28 V at 35 A.

12.5 μH to the common mode—the same order of magnitude—giving a total of 27.5 μH. This makes for a rare design.

There is some differential, with the two balanced inductors totaling 50 μH and the two feed-throughs in series giving 3.4 μF. The impedance is 3.8 ohms, and the cutoff is 12.2 kHz.

22.2. 120 VOLTS, 60 HZ, WITH TRANSZORBS

Figure 22.2 shows another case in which the differential inductor is larger by two orders of magnitude than the common mode inductor. The two differential inductors give 1.2 MH of common mode, so the 15 μH adds no real support for this situation. The 15 μH is smaller than the error of these two differential components. It seams that the two inductors are misnamed. The two 2.4 MH should have been the common mode and the two 15 μH inductors the differential mode inductors.

The two *RC* shunts are not correct, either. The 47 ohm may be high unless the design impedance is about 47 ohms and the capacitor is too low. The

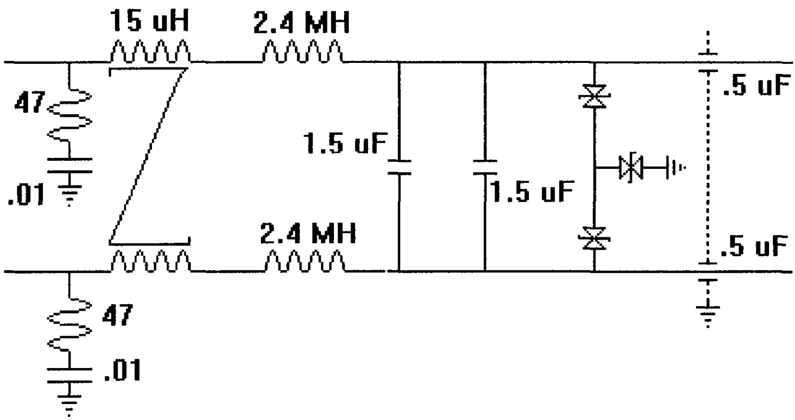


FIG. 22.2 60 Hz with 120 V and transzorbs.

attenuation of this *RC* shunt starts at 340 kHz, so it will not help with any switcher noise and most of the parasitic noise.

The transzorbs are wired wrong. The third transzorb in the ground return should be wired line to line, making two from line to ground for common mode pulses. The third should be from line to line for differential mode pulses. These should be moved to the input to protect the filter and placed outside the filter container for ease of replacement.

The feed-throughs may be high for 60 Hz, giving a higher leakage current to ground, more than what is specified.

22.3. THE 28 V DC FILTER

The question about the circuit of Fig. 22.3 is, can the equipment that this filter is used in even be turned on? Discounting the 0.5 ohm leg capacitor and the other smaller capacitors, the total capacitance is 2520 μF across the line in a circuit that draws only 6.2 A. This filter current plus the equipment inrush plus the other capacitor current will be more than the circuit breaker can handle.

From an EMI standpoint, these three large capacitors are useless anyway. They are inductive long before any EMI requirement must be met. This is the disadvantage of using most analysis programs because most treat these as pure capacitors well into the GHz realm. Some programs allow the designer to add this information, but most enter the component information as pure. Those that would do so are often not aware of the true property of the component.

The two *RC* shunts are almost useless. They do not come into play until 720 kHz, which is far above most switcher frequencies and their harmonics and

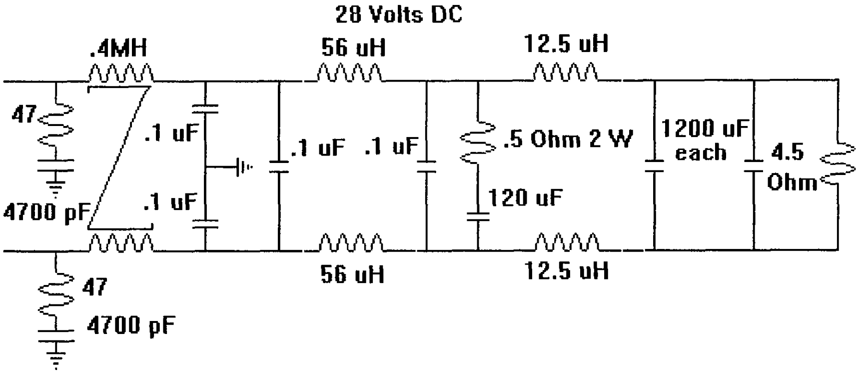


FIG. 22.3 The 28 V DC filter.

above most parasitics of the switcher transformers and/or inductors. If there is a high-frequency problem with this filter, the 47 ohm resistor does not help. The value should be closer to the load impedance, which should be near the design impedance of the filter. This should be closer to 4.5 ohms.

The 0.1 μF capacitors to ground should be feed-throughs for better SRF and should be located on the load side. These two capacitors are in series across the line. This gives a line-to-line value of 0.05 μF and adds to the input capacitor of the balanced pi for a total of 0.15 μF. If the leakage inductance of the Zorro is small and if the line impedance is small, this renders the 0.15 μF useless well into the upper kHz range. This is not a VDE, FCC, or CSA filter because the component count is too high for these types. If losses are required in the lower kHz range, this technique is not valid. If this filter was tested using the 220A or some of the better LISNs, creating a line impedance in the lower kHz range, the filter could look good. This is, again, why the pi filter should be limited in use.

If the 0.5 ohm and 120 μF are there to lower the DC supply impedance, this network should be moved to the load side—why lower the impedance and then raise it by the two 12.5 μH inductors? However, the 120 μF is too big to reduce the switcher drop, and so is the 2400 μF. The best solution would be to remove these four components along with the two *RC* shunts. Move the capacitors to ground, as stated earlier, and then double the size of the output pi capacitor. This makes this an L, which is all that is there anyway at a cutoff frequency 1.414 times higher. The circuit now looks as shown in Fig. 22.4.

With all the space saved, the 0.1 μF could become a 6.8 feed-through, aiding the 0.4 MH Zorro and raising the 12.5 to 56 μH to make this a balanced T. Careful here, though, this could add to the switcher drop. Add another 0.2 μF and make this a double balanced L. More must be known to design this properly, such as the switcher frequency, the on time of the pulse, and the current peak

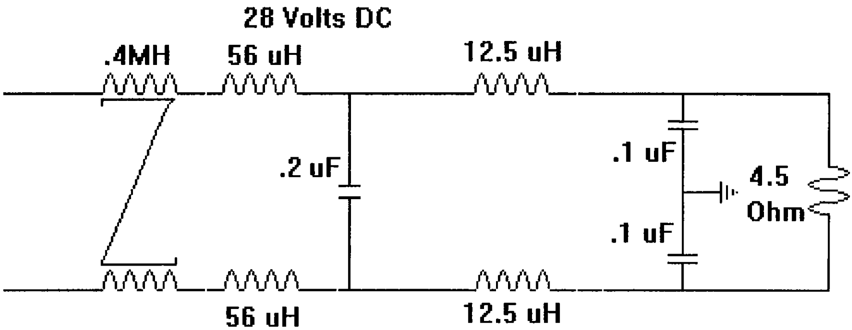


FIG. 22.4 The modified 28 V DC filter.

during this time. In other words, Figs. 22.3 and 22.4 will not function properly. This EMI filter needs to be fully redesigned.

22.4. 120 V AC 400 HZ

The main complaint is that there is no real common mode or differential mode in the filter in Fig. 22.5. The *RC* shunt's minimum impedance is 453 ohms, giving little aid to the common mode or differential, and the same is true for the 4700 pF to ground. This may have been a medical application requiring low capacitance to ground for low patient current. The current of nearly 1 A agrees with this. This

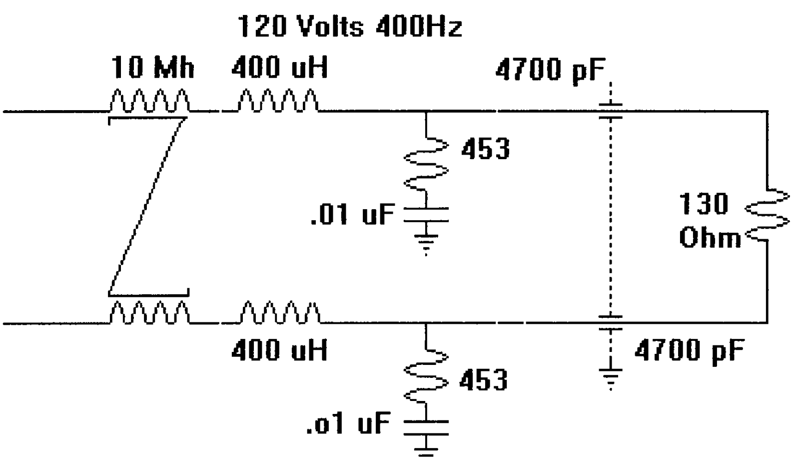


FIG. 22.5 The 120 V 400 Hz filter.

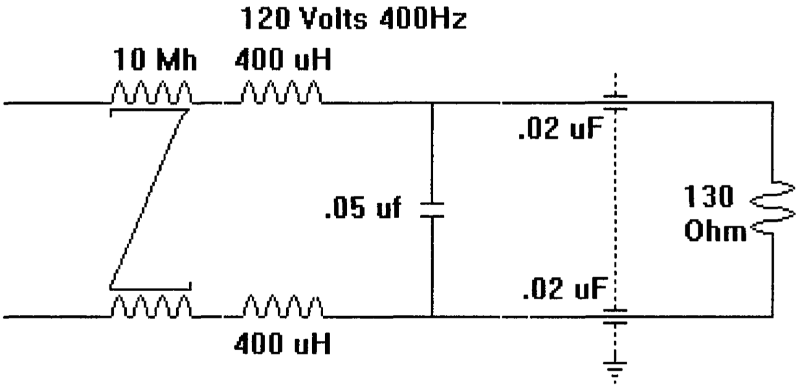


FIG. 22.6 The modified 115 V 400 Hz filter.

filter gives 6 dB per octave for both common and differential mode. If the 800 μH—the bottom and top 400 μH to convert to unbalanced—is correct, a 0.05 μF tied line to line after the inductors would give over 40 dB of loss at 450 kHz for FCC.

If this is not for medical use, change the 4700 pF for the limit of 0.02 μF, the legal limit for 400 Hz. This is an order of magnitude larger. Now, with the added 0.05 μF from line to line, there are both common mode and differential mode losses. If the two *RC* shunts are there for a problem frequency, the 453 ohm resistor is too high. Change these two resistors to 130 ohms each. One of these *RC* networks, wired from line to line, may replace both networks to ground. Otherwise, leave the *RC* shunts out. The changes are shown in Fig. 22.6.

Now, the filter has 10 MH and 0.04 μF making up the common mode half of the filter and 800 μH and 0.05 μF making up the differential mode half. The common mode design impedance is 500 ohms and the cutoff is 8 kHz; the differential mode impedance is 130 ohms with a cutoff frequency of 25 kHz.

22.5. REVIEW

It is impossible to look at a filter after the fact and judge it fairly. We do not know all the required parameters of the filter: medical, 461, 220A only, resonant rises at some unknown frequencies. Information may have been provided to the filter designer either from the EMI test laboratory or related to some other aspect that we are not aware of. I hope the reader understands that I am not trying to find fault with any group.

23

Review of Filter Design

This chapter is a catchall for discussing some thoughts mentioned in the book but not clarified. First, the full design technique is reviewed. If no other chapter of this book is read, at least read this chapter!

23.1. FILTER DESIGN REVIEW

This is an overview of all the design information in this book with references back to the chapter that discusses the subject. Much additional information has also been stated repeatedly without being documented.

23.1.1. Steps in Designing a Filter

Before designing the filter, the following should be known: the power line frequency if AC, minimal RMS line voltage, maximum RMS load current, type of load, and lowest switcher frequency if more than one. Some of this can be guessed at. If this is a single-phase 60 Hz line, a minimum of 100 V RMS should do and the RMS line current could be raised by 10%. Attempt to determine the type of load this filter must feed power to: off-line regulator, power factor correction circuit, inductive input filter, or resistive. For a DC system, this may well be a switcher. If the filter is being designed for multiple use for off-the-shelf filter sales, worst cases from previous chapters should be accounted for in the design.

1. Find the best filter type from Chapter 15 by knowing the type or need of the load. Also, the specification will help here. Knowledge of the specification and losses required will help in selecting the line impedance, also from Chapter 15. Federal Communications Commission (FCC) requirements give a higher line impedance because the loss requirements are at frequencies that give a higher line impedance. The 50 ohm line impedance is valid at these frequencies.

2. Calculate the design impedance from the lowest voltage divided by the highest current.

3. Determine the filter cutoff frequency, both differential and common mode. Use either the matrix equations of Chapter 16 or 17 or Sec. 18.8 on F_0 the easy way. The second approach will get the filter designer onto the proper ballpark property and the first will get the designer onto the ballpark playing field. This can be a computer program written by the reader based on Chapters 16 and 17 or from the provided disk. Equations for the cutoff frequency are

$$L = \text{loss for } \pi \text{ or T} = (2N + 1) \times 6$$
$$L = \text{loss for an L} = 12N$$

The first equation is for the T or π filter and the second equation is for the L. Either of these equations provides the value of L for Fig. 23.1. N is the number of filters in tandem.

Loss required	dB		
Loss per Octave (Table 23.1)	L		
Number of Octaves	X	=	dB/L
Frequency requiring the loss	F		
Cutoff frequency	F_O	=	f/2X

$$F_O = \frac{F}{2^{\frac{dB}{L}}}$$

FIG. 23.1 F_0 the easy way equations.

Another way to select the value for L is as follows. Choose the loss based on the filter type. The number on top in Table 23.1 is the number of filters in tandem. Table 23.1 gives the loss per octave for the filter type listed on the left, the value of L in Fig. 23.1.

Once 10 times the power line frequency or more is reached, the filter type and the cutoff frequency are known. As an example, three frequencies are provided by the EMI test house. They list the approximate dB losses required to meet the specification and the associated peak frequencies. Table 23.2 is given.

Table 23.2 gives the test house readings followed by the loss added by the filter designer for the required headroom. The last two rows show the cutoff frequency obtained by using the preceding methods and equations for the single L filter and the readings for a double L filter obtained by the same techniques.

If the power line frequency is 60 Hz, the single L will work and the cutoff frequency is 1.4 kHz, but if the power line frequency is 400 Hz, the double L must replace the single L at 12.98 kHz. Round either of these to a convenient lower frequency. Often, the filter cutoff frequency is determined by the lowest problem frequency. Be careful, because exceptions do occur, as in the preceding example, so all listed trouble frequencies must be checked. Do not assume that it is only the lowest frequency listed that will give the proper cutoff frequency.

4. Equate the component values from steps 2 and 3 using the following equations. R_d is the design impedance. This is the proper equation for all filters, that is, π, T, and L. The difference is that the value of L is divided or split for the T and the value of C is divided or split for the π.

$$L = \frac{R_d}{2\pi F_0} \qquad C = \frac{1}{2\pi F_0 R_d}$$

In a multiple T filter, the first and the last L are half the preceding value but the values of the central inductors are the full value. Remember also that the value

TABLE 23.1 F_0 the easy way chart

Filter	1	2	3	4	5	6	7
L	12	24	36	48	60	72	84
π	18	30	42	54	66	80	92
T	18	30	42	54	66	80	92

TABLE 23.2 Chart of F_0 for L and Double L Example

Test house trouble frequencies	F_1	F_2	F_3
Frequencies	36 kHz	80 kHz	120 kHz
Test house measured loss	36 dB	58 dB	71 dB
Add 6 dB headroom	42 dB	64 dB	77 dB
Single L F_0	4.5 kHz	2.8 kHz	1.4 kHz
Double L F_0	17.2 kHz	14.98 kHz	12.98 kHz

of C above will be divided for the π. Again, the central values of C for a multiple π will be the whole value and the first and last C values will be split. The T and the π are exactly opposite each other. Choose the filter layout based on Chapter 6 and also some information in Chapter 3. Balance the filter if possible.

5. Find the size of the components. Use McLyman's data in Chapter 20.

The component volume comes from the outside diameter of the toroid, connector, or round capacitor and O_d is squared times the length. The pressed capacitor volume is D_x times D_y times A_W. Total all the components, and divide by 0.6. The aspect ratio should be a minimum of 2 to a maximum of 5 for the length to end diagonal, but this is also a function of component size. For example, a $9 \times 3 \times 2$ may give the calculated needed volume and a good aspect ratio, but the wound inductors may be 3 inches in diameter. Either all the inductors will be aligned along the 9×3 inch direction or the height must change. This example may work, but quadrature is violated and cross talk may be a problem even though these are toroids.

6. Design and build the case, or container, on the basis of the components and volume just determined, and make sure there are no solder or weld voids, which can allow radiation or destroy the potential of passing any environmental tests, such as for humidity and salt spray. Have the case silver plated for better surface conduction if the loss is substantial. Also, aluminum is often used for better conduction giving lower radiation. This is plated for still better conduction for lower radiation and for ease of soldering.

7. Install the components, and test the filter in the open container. Tack the lid down only for easy alteration or adjustment.

8. Adjust the filter for the desired loss, if needed, by the following steps:

Add lossy components (Sec. 5.4).

Add small line-to-line (X) capacitors in parallel with the existing capacitors. Keep the lead length as short as possible.

Add ferrite beads if the current is low enough (typically 5-Amp limit).

Add several turns on the inductors (watch for saturation). If this filter is a T, whatever turns added to the central inductors of the multiple T, increase the turns by half on the two outer inductors. They are half the value of the inner inductors.

Add an RCSHU (a resistor and capacitor in series across the line). See Sec. 5.7.

Adjustments will move the designer from either the ballpark property or the playing field to home plate with one of the approaches in step 3 above.

9. Make sure that the end product is buildable and repeatable for production.

23.2. FILTERS IN TANDEM

This section discusses the filter growing from one section to two sections to three sections and so on. This book states repeatedly (as when going from a single to a double L in Table 23.2) that as the sections grow, the size of the components falls. The test problem is as follows:

Line frequency, 400 Hz
Line voltage, 120 V
Maximum line current, 6 A
Needed loss, 60 dB at 20 kHz

This will be solved first for the L then the T and followed by the π, only to prove the same holds true for all three types (Tables 23.3, 23.4, and 23.5). The program is the one used in Chapters 16 and 17. This is a low current, and the maximum should be four sections but six are given. The value of L is in MH and C in μH. The first L and C are the individual component values, and the final values of L and C are the total component values for the filter.

TABLE 23.3 The L Filter Showing the Change as Sections Are Added

Number in tandem	F_0	Inductance	Capacitance	Extension Inductance	Extension Capacitance
1	405	7.860	19.6	7.860	19.6
2	2890	1.130	2.82	2.26	5.64
3	5170	0.615	1.54	1.85	4.62
4	6760	0.470	1.18	1.88	4.72
5	7760	0.410	1.02	2.05	5.10
6	8400	0.378	0.95	2.26	5.70

TABLE 23.4 The T Filter Showing the Change as Sections Are Added

Number in tandem	F_0	Inductance	Capacitance	Extension Inductance	Extension Capacitance
1	940	1.70	8.64	3.40	8.64
2	3130	.508	2.54	2.032	5.08
3	5075	0.314	1.56	1.884	4.68
4	6460	0.246	1.23	1.968	4.92
5	7500	0.215	1.07	2.15	5.35
6	8435	0.198	0.99	2.37	5.94

Note how the total L and C drop in value from one to three sections, but little is gained past three sections. Little self-resonant frequency (SRF) is gained in moving the capacitor value from 1.54 (1.6 μF) to 1.181 (1.2 μF). The circuit can be balanced, with better SRF, moving the 620 μH to 310 μH.

The T filter in Table 23.4 is the same as the L in Table 23.3. It is ridiculous to go beyond three sections or four at the most. The inductance 314 μH is the value in each arm, or the total inductance for this one T is 628 μH. The central inductors are twice this initial value or 628 μH each. Round these to even values, and the three-section T circuit is shown in Fig. 23.2. The balanced circuit is shown in Fig. 23.3.

Note that the π filter values of the cutoff frequency, F_0, are so much lower for the various sections as compared with the T filter. Both the T and the π are said to have 18 dB of loss per octave. But the T has this loss, which is missing in the π. This is due to the low line impedance at 20 kHz, estimated at 4 ohms.

TABLE 23.5 The π Filter Showing the Changes as Sections Are Added

Number in tandem	F_0	Inductance	Capacitance	Extension Inductance	Extension Capacitance
1	555	5.735	7.164	5.735	14.32
2	2520	1.263	1.578	2.526	6.312
3	4730	0.672	.841	2.016	5.064
4	6360	0.500	.626	2.000	5.008
5	7450	0.427	.534	2.135	5.340
6	8155	0.390	0.488	2.340	5.850

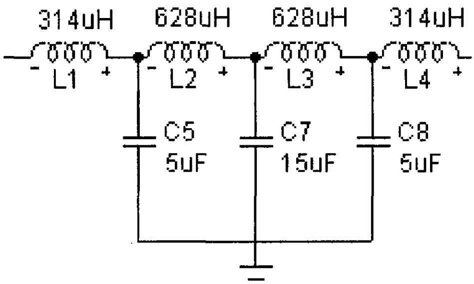


FIG. 23.2 The multiple T.

If this had been equated with 220A, the cutoff frequencies would have been similar to those for the T. Under these conditions, three sections may be too low and a fourth section may be required. This is caused by the value of F_0. F_0 is cutting into the band pass, which is decided by the 10th harmonic of the power line frequency. The value listed for C is the value of the outside capacitors, and all inboard capacitors are twice this value. The four-section π is shown in Fig. 23.4.

In any of these applications, using 6 dB headroom, the values are not critical. Round them to reasonable values. For the two end capacitors in the balanced π in Fig. 23.5, use 0.6 μF rather than 0.63 μF.

Comparing the L, T, and π sections, the total inductance and capacitance drop greatly from one section to two. A reasonable drop repeats going from two sections to three sections. A small drop occurs from three to four sections. This is followed by a modest increase from four to five sections and again from five sections to six. If the problem is being caused by a low SRF, adding a section may help. Going to a balanced arrangement almost doubles the inductor SRF, and

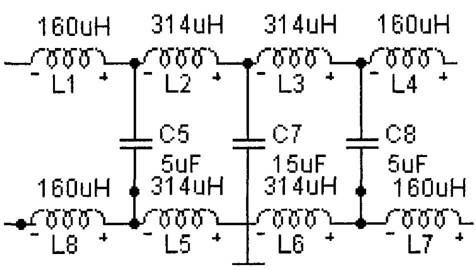


FIG. 23.3 The balanced multiple T.

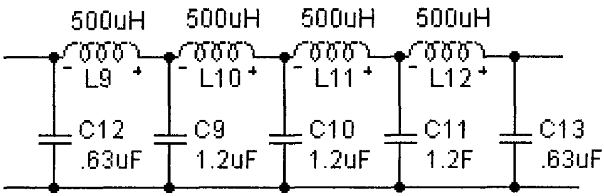


FIG. 23.4 The multiple π.

this should help. If the problem is the filter size, going up a section also helps because the component size is reduced, giving more room.

Looking again at the three types discussed, the π is the least reasonable and should be used only when a 220A test specification is called for. The reason this happens is that the front-end capacitor is out of the circuit at the low frequencies because of the low line impedance. This capacitor aids the filtering when the line impedance rises to a reasonable value. An easy way to calculate this improvement of lower total inductance and capacitance is through the equations listed earlier. These are repeated in Fig. 23.6; they cannot be differentiated because they are not continuous functions.

As the value of L doubles and triples because of a higher number of filters in tandem, the denominator decreases and the cutoff frequency rises. The value of the loss, L, can be calculated from

$$L = \text{loss for } \pi \text{ or T} = (2N + 1) \times 6$$

$$L = \text{loss for an L} = 12N$$

The upper value of L is for the π and T types, and the lower value of L is for the L filter (see Table 23.1).

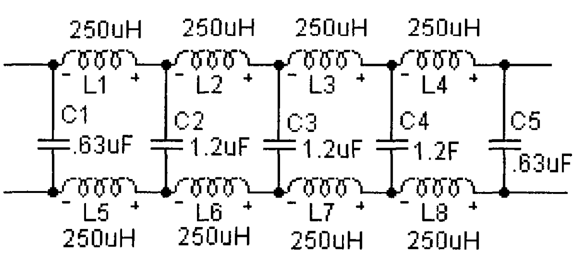


FIG. 23.5 The balanced multiple π.

Loss required	dB		
Loss per Octave (Table 23.1)	L		
Number of Octaves	X	=	dB/L
Frequency requiring the loss	F		
Cutoff frequency	Fo	=	f/2X

$$F_o = \frac{F}{2^{\frac{dB}{L}}}$$

FIG. 23.6 F_0, the easy way equations.

23.3. *Q*

The value of the circuit Q of the filter must be low, as stated, without explaining the reason, often in this book. If the Q is low enough, the number of resonant rises is one less than the filter sections. The resonant rise level is also a function of Q. The input voltage level of the resonant rise frequency times Q of the circuit at that frequency equals the voltage output at this frequency. If high-quality capacitors are used, the Q of the circuit will fall to the inductors because capacitors normally carry a much higher Q rating, typically 10 times greater than the inductors. For a single L circuit, the equation for Q is

$$L \equiv \frac{R_d}{2\pi F_0} \qquad X_l = \frac{2\pi F R_d}{2\pi F_0} = \frac{F R_d}{F_0} = K R_d$$

$$Q = \frac{X_l}{R_d} = K$$

This value of Q is for one inductor. The value of the Q will be much lower than K because the design impedance, R_d, is lower than the operating impedance of the load. These simple equations do not figure in the DC resistance of the inductor, the line impedance, and the equivalent series resistance of the capacitors. This is another reason the design criterion sets the cutoff at least 10 times the line frequency and the headroom of 6 dB. The line harmonic content should be low enough by the time the circuit Q is high enough to cause ringing or a resonant rise. In multiple circuits, the Q is a function of the total inductance. The value of Q must be multiplied by the number of inductors, but as the cascaded sections increase, the value of F_0 rises, lowering the individual inductance, and also the total inductance (up to four) decreases.

For the T and π filters, the equation for the value of L in the loss equation is $(2N+1)$ and the total loss per octave is $6(2N+1)$ dB. For the L, the loss per octave is $(12N)$ dB. Either of these two values replaces the value of L in the previous section. The first equation is for either T or π and the last is for L. N is the number in tandem (Fig. 23.7).

The needed loss at the trouble frequency (dB), the tandem number (N), the trouble frequency (F_t), and the needed Q are known. Substitute 1 for Q and solve for F. This is the approximate frequency where Q is one (because of losses not accounted for, the actual value of F will be higher). Is this near any power line harmonic, especially the odd harmonics, or near any known problem frequency? If yes, or just to be safe, add an RC shunt to this filter where X_c equals R_d at this frequency F. If another RC shunt is required for other reasons, only one is needed.

$$F_O = \frac{F_T}{2^{\frac{dB}{(2N+1)6}}}$$

$$L = \frac{R_d}{2\pi F_0} = \frac{R_d 2^{\frac{dB}{(2N+1)6}}}{2\pi F_T}$$

$$X_L = \frac{2\pi F R_d 2^{\frac{dB}{(2N+1)6}}}{2\pi F_T} = \frac{F R_d 2^{\frac{dB}{(2N+1)6}}}{F_T}$$

$$Q = \frac{N X_L}{R_d} \quad \frac{F N 2^{\frac{dB}{(2N+1)6}}}{F_T}$$

FIG. 23.7 Equations for Q.

Opt for the lower frequency *RC* shunt. If there are higher frequency *Q* problems, opt for a higher quality capacitor that has a higher *Q* and a higher SRF. Also, in this case, make sure the resistor used in the *RC* shunt is noninductive.

23.4. TESTING THE FILTERS

Throughout this book, the test specifications have been mentioned, but not the testing itself. The test equipment input and output impedance is 50 ohms unbalanced. The connectors are BNC and the cable is a coaxial cable (coax). This means one side is ground. So, how are the various EMI filters checked?

In Fig. 23.8, the first ground to the left is the tracking generator coax ground at the filter. The central ground represents the filter ground, and the last ground on the right is the spectrum analyzer ground. The filter is pressed to ground with clamps, vices, or other means. As stated throughout the book the, this filter must be grounded to work properly. If it is not grounded, the capacitors within the tube are off ground and cannot function. Here, common mode and differential mode are not really treated differently.

In the balanced filter in Fig. 23.9, the hookup is the same. Will this check the filter? Note that the bottom row of inductors is out of the picture, shorted out. The inner capacitors are also out of the picture because of being in series with the inductors. The loss pattern would not meet any specification. The question is, what will fix the problem?

The answer is that the unbalanced output and input of the test system must be converted to balanced. Matching transformers must be used on both ends (Fig. 23.10). These transformers are 1:1, 50 ohms in and out high-frequency devices. They are small high-frequency balun wound in ferrite balun cores. Not all the internal ground components are shown in Fig. 23.10. This is measuring the

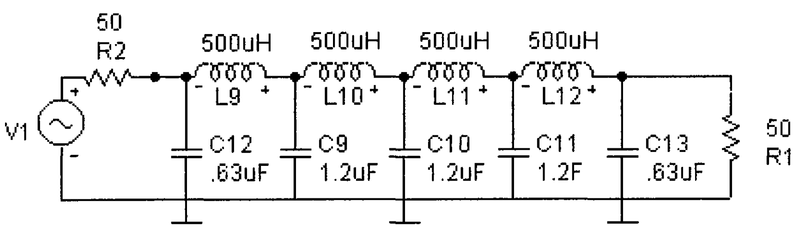


FIG. 23.8 Test setup for an unbalanced filter.

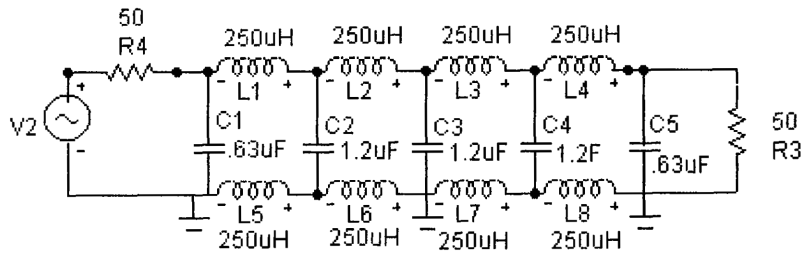


FIG. 23.9 The balanced filter.

differential mode only. All capacitors and inductors add loss to the filter. What about the common mode loss? How is it measured?

In Fig. 23.11, the differential mode components are removed. The balanced inputs are shorted together and the output is also shorted together. The figure shows the capacitance to ground. The Zorro, or common mode choke, is included. The test equipment is grounded at the filter. A common mode loss pattern should be obtained. If you test the filter as in Fig. 23.9, only the common mode inductor and the one feed-through are being tested.

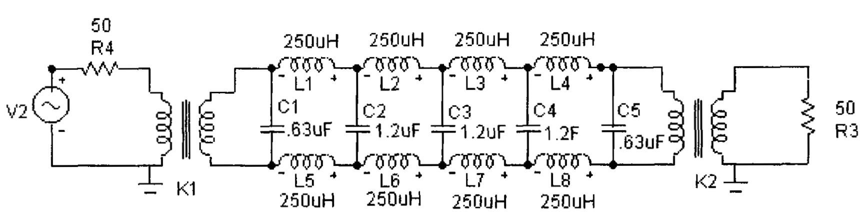


FIG. 23.10 The balanced filter with matching transformers.

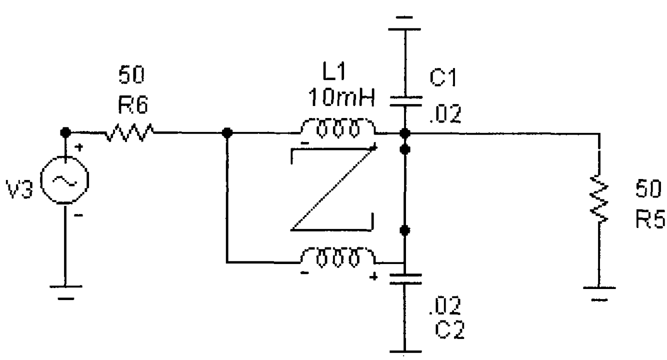


FIG. 23.11 The common mode part of Fig. 23.10.

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