MILITARY COLLEGE OF ELECTRONICS AND MECHANICAL ENGINEERING

Faculty of Electronics
Secunderabad - 500 015

DISSERTATION

ON

101 p + fig.

EMI FILTER DESIGN

Guide :

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1997

Military College of Electronics and Mechanical Enginrring

FACULTY OF ELECTRONICS

SECUNDERABAD

CERTIFICATE

Certified that this is a bonafide report of the dissertation work done by *Maj Raghunandan P Nair* during the year 1997 in partial fulfilment of the requirement for the award of the Degree of Master of Technology in Electrical Engineering by the Jawaharlal Nehru University, New Delhi.

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ACKNOWLEDGEMENT

I wish to express my gratitude to Lt Col AK Lal who has supervised and assisted in the completion of the dessertation.

I am indebted to the Faculty of Electronics MCEME Secunderabad for extending the complete cooperation and facilities for the long hours of work.

I would like to thank my department for extending the facilities and resources to me.

Major Raghunandan P Nair

CHAPTER - I

INTRODUCTION

1.1 BACKGROUND

ELECTRO-MAGNETIC-INTERFERENCE (EMI) has gained importance over the years since the importance of secrecy and the traffic of information has prolifered. Today, with the proliferation of electrical and electronic gadgets and equipments the Electromagnetic Pollution has come to a level where organisations and committees have had to bring about restrictions and lay down specifications for manufacturers and industries.

As a result of the awareness of this fast growing pollution the ELECTRO MAGNETIC INTERFERENCE FILTER Industry has taken birth. This has been a malignant growth that has percolated to the design stage of every industry.

1.2 OBJECTIVE

The objective of the Thesis is to study the various aspects of design of EMI FILTERS.

1.3 ORGANISATION OF THE THESIS

Chapter I is the introduction and out line of the Thesis.

Chapter II gives an introduction of EMI filter design. The comparison between a normal filter company and an EMI company is given. It further brings out the lack of a power density spectrum, the aspects of power transfer and specifications.

Chapter III gives details of the various types of filters (π, T, L). These include circuit descriptions upto component level. Mathematical model is not brought out at this stage. A description of the commercial filter, Cauer filter, conventional filter follows. Finally, giving the filter matrix for line and load conditions.

Chapter IV describes the differential mode components. The design and construction details in respect of capacitors, and inductors have been brought out. Certain details in terms of actual dielectric ratings that are being followed in certain firms have been illustrated.

Chapter V brings out the details of the common mode components. Conversion of common mode to differential mode filters .

Chapter VI describes the compromises that a filter designer has to make under due to unavoidable factors which are beyond the scope of EMI filters.

Chapter VII gives the various types waves as noise sources.

Chapter VIII gives the initial filter design requirements.

Chapter IX brings out the filter design techniques using the various matrices.

Chapter X shows the matrix applications.

Chapter XI gives a review of filter design.

CHAPTER - II

WHY CALL EMI FILTERS BLACK MAGIC

Most engineers, both designers of electromagnetic interference (EMI) devices and others, call EMI black magic. There are three main reason 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 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 Ohms source and load impedances, does the filter engineer design the filter for 50 Ohms or for real-world impedances? Even if the design passes the 50 Ohms tests, what will happen later in the real world ? There are practical problems due to the wide variation in the real world. The design methods outline in various publication are generally very complex. Most require measurements that are difficult to determine and time-consuming. Often, expensive additional equipment is required to obtain the necessary parameters.

2.1 DESIGN CONCEPTS OF A STANDARD FILTER CONTRASTED WITH AN EMI FILTER

Most EMI filter manufactures design and build only the lowpass filters needed for EMI attenuation. Rarely do they build bandpass or other conventional filters. The technology used in conventional filters is truly different from that used in the EMI filter, whose design is very loose compared to that used by the conventional filter manufacture. The filter component values are very flexible, so the engineer can use standard values. These are adjusted only to meet the insertion loss specifications.

While, True filter houses speak of poles, zero, group delay, predisposition, attenuation, and the order of the filter. The EMI filter designer thinks in terms of attenuation, insertion loss, voltage drop, 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. So flat frequency response, low phase distortion, and low peak-to-peak ripple across the filter bandpass are not issues here. These power-line harmonics furnish little 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 not often heard.

2.2 POWER DENSITY SPECTRUM OR ENVELOPE

The EMI designer is not blessed with density spectrum as described above. The designer knows the power-line frequency and its harmonics, which should be passed on to the load. The harmonic power provides little power to the load, but neither should the EMI filter attenuate them. Such attenuation would call for larger filter capacitor currents along with eddy currents and hysteresis losses in the core which develop more heat within the filter. The filters must pass high power levels at 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 to handle the current. This dictates that the filter will increase in cost, size, and weight.

2.3 POWER TRANSFER

The power to be attenuated is minimal compared to the power transmitted if the subject is not on the much larger electromagnetic pulses. These conditions must also be treated or attenuated. This typical noise could be any, or all, of the following:

(a) The switcher frequency pulse the odd harmonics.

(b) The initial power supply and/or switcher diode noises.

(c) The parasitic oscillations of the switcher inductor or transformer. This noise energy is at a much lower level than the power supplied to the load. To move this power to the load, larger filter components are needed.

The transfer of power creates the following:

(a) Larger fitter with more weight and cost.

(b) Much lower self-resonant frequency (SRF) values.

The SRF value of these filter components is indirectly proportional to the power demanded by the load. This limits the filter's usefulness at higher frequencies. The condition will normally dictate that other components to be added to the filter to compensate for this lack at higher frequencies. These additions to the filter could be handled by placing small ceramic or extended foil Mylar capacitors in parallel with the existing capacitors. The second capacitor will raise the SRF value.

2.4 SPECIFICATIONS; REAL WORLD OR IMAGINED

Another thing that makes EMI "black magic" are the specifications. Some test specifications cloud the design. These specifications are stated in such a way 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 closed to 50 Ohms and work into a source of rarely 50 Ohms. The line impedance simulation network ratio (LISN) is often used as the source impedance for these tests and is closer to real-world requirements.

Some specifications call for an LISN value that gives an output impedance of 50 Ohms starting well above 100kHz. 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 concern here. The first: What is the lower frequency of loss required? Is it below 100 kHz? The LISN output impedance drops rapidly from 50 Ohms and the filter is then mismatched. The second is: What happens in the real world when neither the source nor load impedances are close to 50 Ohms?

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

CHAPTER - III

TYPICAL EMI FILTERS

The typical electromagnetic interference (EMI) filters used today. Are primarily the pie, T and L with the Cauer, resistance-capacitance (R-C) shunt, and dissipative sometimes included. They include double and triple filter and sometimes even quadruple use of the mentioned types of filters. Each EMI filter has some positive and negative attributes. Each has its best spot in which it functions very well and other spots in which it will fail to the first degree.

3.1 THE PIE FILTER

This filter looks very good under the Military Standard 220A,50 Ohms-test specification (Figures 3.1, 3.2 & 3.3). The low frequencies still work in the 50 Ohms load and line impedances. If this is the only test requirement specified by the customer, then the pie tests well using this test technique. The pie filter will be passed with flying colors. 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 pie section are then doubled in value. This eases the job of the filter designer to meet the specified loss within a given weight and volume. This means that the filter can easily pass the attenuation requirement with smaller values, and package size and weight. 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 pie filters.

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. This is done to ensure that the filter inductors do not saturate at the peak current. There are two sets of 50 Ohms matching pads, from 6 to 10 dB each, that

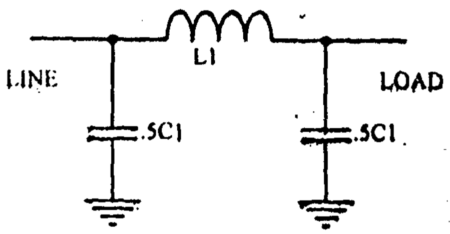


Figure 3.1 The π filter.

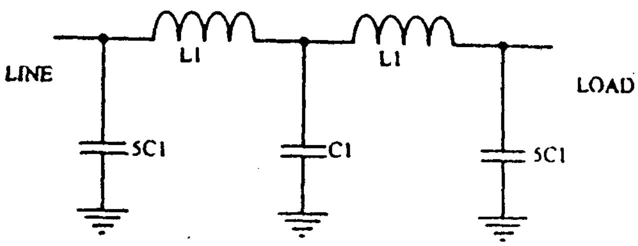


Figure 3.2 The multiple π.

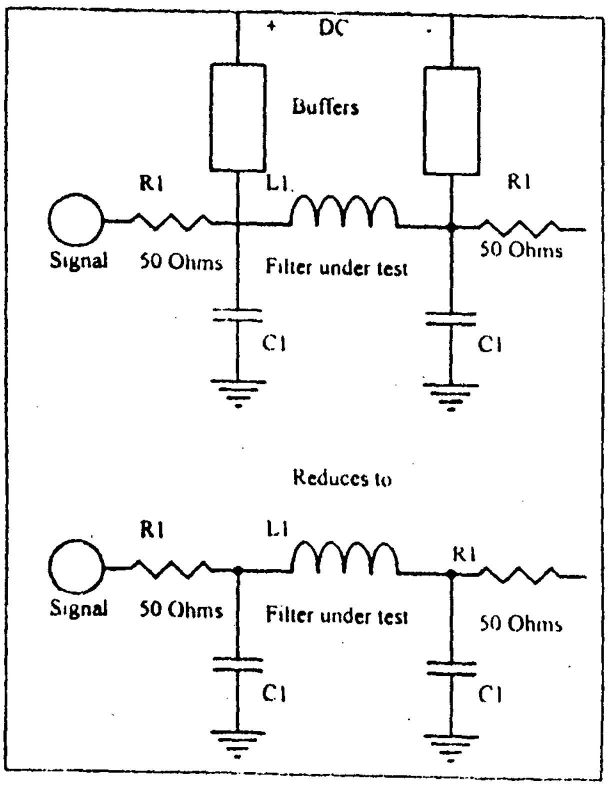


Figure 3.3 Military Standard 220A test setup.

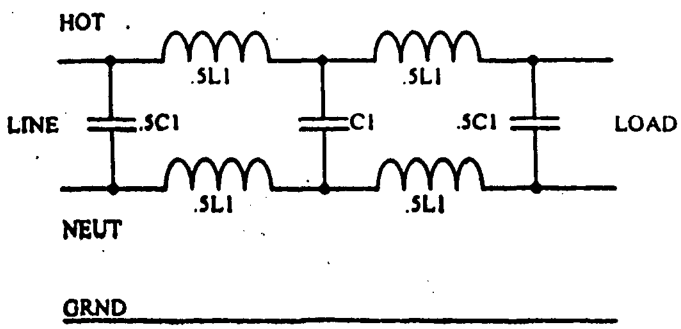


Figure 3.4 The multiple balanced π.

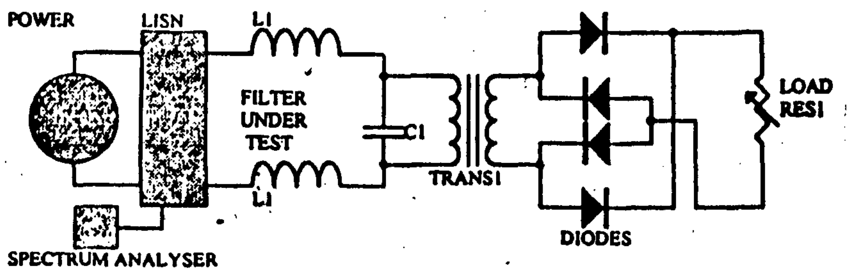


Figure 3.5 The system or diode test method.

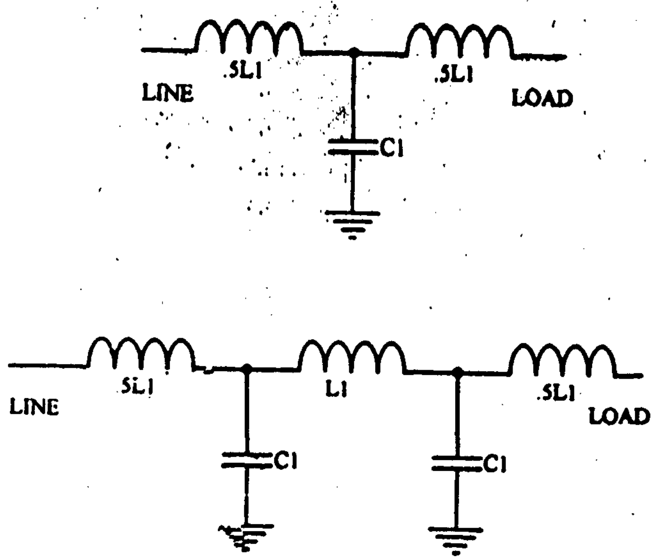


Figure 3.6 The T and multiple T.

are omitted from these figures.

The pie might also function well in some dc systems if the switcher frequency is high enough so that the capacitor impedance facing the load does not starve the switcher or cause excessive voltage drop. This assumes that the switcher circuit has not handled this problem. The pie filter is easily balanced by placing only half of the inductor needed in the high line and the other half in the neutral line (Figure 3.4). This changes the shape from the pie to a square or box shape. It does happen that the system must pass a specification after several prototypes are finished. A quality test may have to be passed (Figure 3.5). Now the earlier pie that passed all those tests may not perform properly. This is because the filter loses the front, or line-side, capacitor at the lower frequencies. Some will wonder 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.

3.2 T FILTER

These give 18 dB per octave, and the double T give 30 dB per octave (6 dB per element, Figure 3.6). T filter work best in low-impedance lines (high current). The line impedance is very low up to at least 100 kHz, but 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 impedance to work into. These are also best at the higher current load if the design method does not call for values for these T inductors that are too high. This could result in soaring or dropping voltages, feeding the load. The T should never be used in a dc system if the load utilizes mainly switchers because the high impedance of the output inductor facing the load will starve the switcher. 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. The

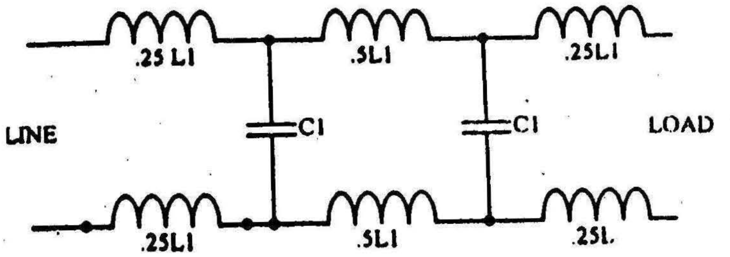


Figure 3·7 **The multiple balanced T filter.**

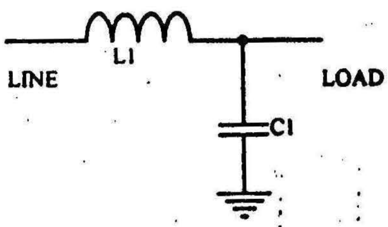


Figure 3·8 **Single L filter.**

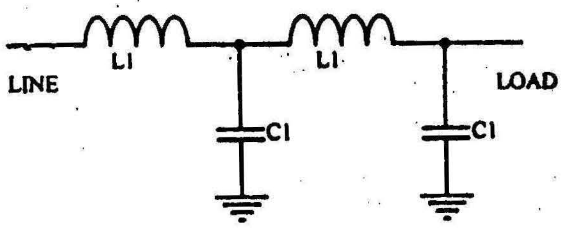


Figure 3-9 **Multiple L filter.**

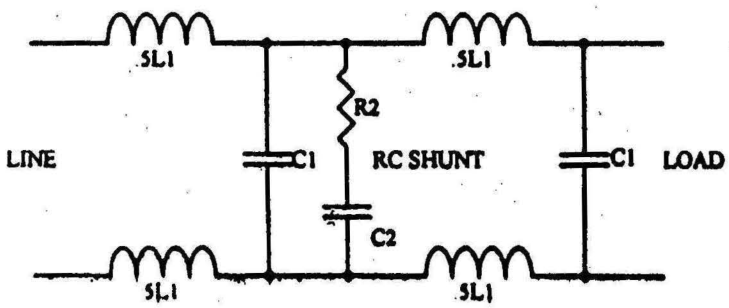


Figure 3·10 **Balanced multiple L with R-C shunt filter.**

T filter can be balanced by removing one-half of the inductor's values and placing this half in the neutral leg, forming an H pad. (Figure 3.7).

3.3 L FILTER

The L type is the most often used filter. Both the pie and the T are three element, or more if multiples are used, and should given 18 dB loss per octave. The L, because it is only two element, provides 12 dB loss per octave. All these loss figure refer to the loss starting above the cutoff frequency. A single L often work best in the dc mode if the load has switchers, because a large L would face the dc supply and the large capacitor (or high quality) would provide a low impedance for the switcher frequency (Figure 3.8). Why not a double L? The two inductors required for the same amount of loss would total less than the single inductor; the same is true for the capacitor values. 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 at twice the switcher frequency if this feeds the center tap of a class B amplifier or Royer (the higher the switcher frequency, the lower the drop). The double L can be used as long as the drop is not excessive or the switcher frequency is high enough. This is true even though the ripple is the same, or better, to the dc source or supply voltage. The L and multiple L work well in higher power applications (Figure 3.9). Again to balance, split the inductors and place the other half in the neutral leg (Figure 3.10). The double L has 24 dB loss per octave.

3.4 TYPICAL COMMERCIAL FILTER

This is the type used in test equipment, computers, and other commercial electronic equipment. The tests are conducted by EMI test houses that help the manufacture with all the documentation. These filter are mainly common mode in appearance, with a capacitor across the input and output from live to neutral and two

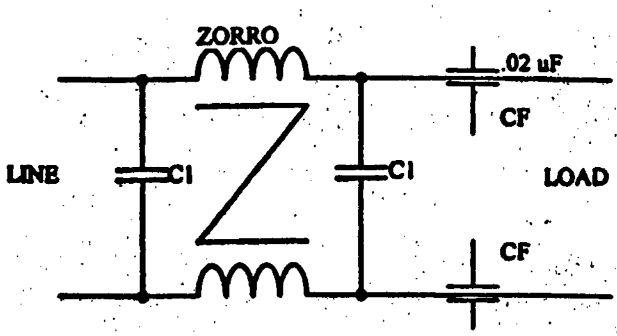


Figure 3-11 Commercial filter with Z and feedthrough.

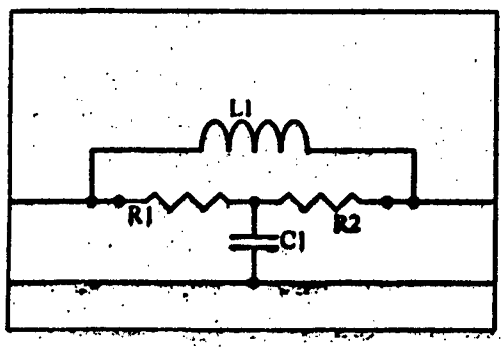


Figure 3-12a Dissipative filter.

other capacitors to ground that must meet leakage current specifications. The leakage inductance is often made high by adding washers to the center of the pot cores separating the two windings. The feed through capacitors(CF) are case grounded (Figure 3.11). Some of these techniques add a differential mode to this common mode filter. This then makes a balanced pie type of differential mode. This all work because the losses specified by the FCC start at 450 kHz. The inductors and capacitors can be quite small to accomplish these tasks. Further, the current through the leakage inductance can saturate this inductor. This is because the common mode inductor is wound on ferrite cores with high A1 values.

3.5 DISSIPATIVE FILTER

18. This filter is rarely seen in the EMI arena today. It consists of one inductor and one capacitor along with two resistors (Figure 3.12a). The two resistors are tied in series across the inductor terminals, and the capacitor is tied to the center of the resistor and then to either the other line or to ground. These filter are similar to the line impedance stability networks (LISN). This filter can be balanced using half the inductor and resistor values to be split on both legs and the capacitor tied between the center point of the four resistors (Figure 3.12b).

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 W for electromagnetic pulse purpose. The advantages are as follows :

(a) This filter can be used to match impedances by using two of these at each end (Figure 3.12c). 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 Ohms resistors and designing

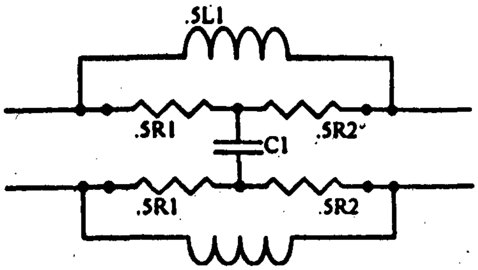


Figure 3-12b Balanced dissipative filter.

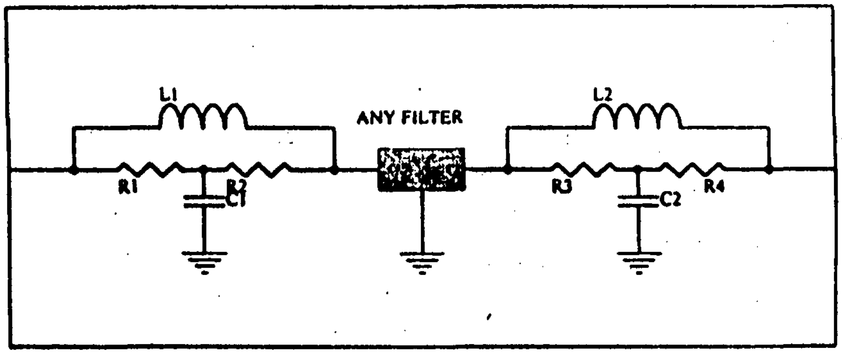


Figure 3-12c DIN and DOUT dissipative filters for both ends.

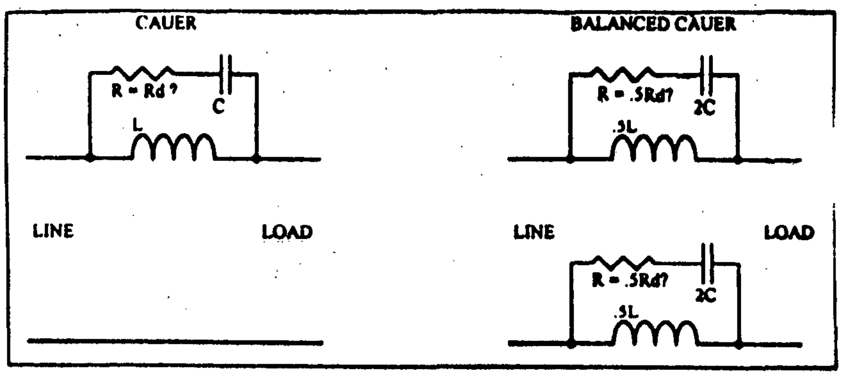


Figure 3-13 Cauer and balanced Cauer.

the middle unit for 12 dB less loss.

(b) This filter dissipates the stop band energy. If a higher frequency comes down the line, most of the energy is dissipated by the line-side resistor. The load-side resistor dissipates most of the energy of the switcher or any other noise from the load. If this is the type just described, the energy not dissipated is reflected by the center filter and the opposite resistor of the same dissipative filter, and most of this energy, would be dissipated.

(c) A filter very similar to an inexpensive LISN would be included within the filter, and two if the unit is the type described by Figure 3.12c.

3.6 CAUER FILTER

The Cauer, or elliptical, filter is best used in very low impedance circuit (Figure 3.13), usually with multiple Ls and Ts. In any case, normally a capacitor is shunted across the center inductor. This is used to fix a problem frequency, such as 14 kHz. This 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 much above this problem frequency. The network will pass all the upper frequencies. A resistor is often placed in series with this capacitor so that the amount of bypass is limited by the resistor. 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, half the calculated value of inductance is used on the two lines, and each inductor is shunted 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.

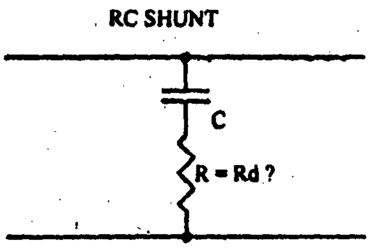


Figure 3·14. Balanced and unbalanced R-C shunt.

3.7 THE R-C SHUNT

Another technique is preferred to the Cauer but is better when used in high impedance, low-current circuits. This filter, called the R-C shunt, uses fewer components and is automatically balance across the line to start with, (Figure 3.14). This is formed by 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, pie, or T. Usually the number of resonant rises is 1 less than the multiple number meaning that the single L, pie, or T would not have resonant rise but the quadruple would have 1 less, equaling 3. This holds true if the Q of the circuit is low enough. The higher Q has a resonant rise for each network. The frequency of the lowest resonant rise is found, and the capacitor value is choosen at this frequency that equals the design impedance of the filter. This attenuates all the resonant rises and also the trouble frequency-if any. If the resonant rise frequencies are of no concern (well above the fifth harmonic of the power line frequency and well below 10 kHz); the capacitor is chosen to equal the design impedances at the trouble frequency.

For 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 4 kHz.

$$R_d = 100/10 = 10\Omega \qquad C = 1/\{2\pi\ 4000 \times 10\} = 3.979\mu F$$

The capacitor is 3.979μF, in series, with the 10 Ohms 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, then, the frequency is changed to the problem frequency of 14 kHz. Then, the value of capacitor is recalculated which is needed at that frequency. The 10 Ohms resistor is in series with a 1.2µF capacitor tied across the filter. This lends itself to multiple L filter in which the network can be tied across any one of the capacitors. The closer the R-C shunt is to the load, the more it tends to minimize or reduce the impedance swings of the load. This is because the load impedance is the highest impedance and the R-C shunt works better at higher impedances.

3.8 CONVENTIONAL FILTERS

Conventional filters are rarely used by filter houses. Some use these filter types, but they require constant input and output impedances 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. These filters would work very well between the two dissipative filters, because the noninductive resistors would give the proper match needed by the these filters, especially if the cutoff frequency of the dissipative filter is about half that of the regular filter.

CHAPTER - IV

DIFFERENTIAL-MODE-COMPONENTS

4.1 CAPACITOR CONSTRUCTION AND SELF-RESONANT FREQUENCY

The differential mode components must be of the high-Q type. The individual component Q must be high, and the current Q must be lowered to the point at which the filter does not oscillate. Figures 4.1a and 4.1b show the self resonant frequency (SRF) for polyester (Mylar) and ceramic capacitors. Both are the feed through type, giving a higher SRF. The next drawing shows the construction and simple equation to calculate the number of wraps to obtain the capacitance needed for the extended foil type. The advantage of the feed through type is the low ESR and ESL (Equivalent series resistance and inductance) as a result of the short lead lengths. This in turn means a much higher SRF. Figure 4.1a and 4.1b show SRFs to 1 GHz for the smaller ceramic caps. The problem is that the higher the power line root-mean-square (RMS) voltage, the more margin is needed to eliminate creepage and corona. This necessitates a larger capacitor and increases the cost. As the line frequency increases, the line frequency harmonic current increases. This also increases the loss as a result of dv/dt. If the margins are near the lower limits initially, when the line frequency increases, the initial margin may need be increased. In other words, the margin is not a function of frequency unless the lower frequency margin is initially small. Many filter builders use the large can type of nonpolarized filter capacitors. Some of these are oil impregnated.

These oil-filled and large-can type capacitors may be very good for power supplies and other applications requiring nonpolarized capacitors. Most of these types have very low SRFs of the order of 20 kHz. If the capacitor is paralleled with another capacitor 0.05 the value of the original. This smaller capacitor should have a much

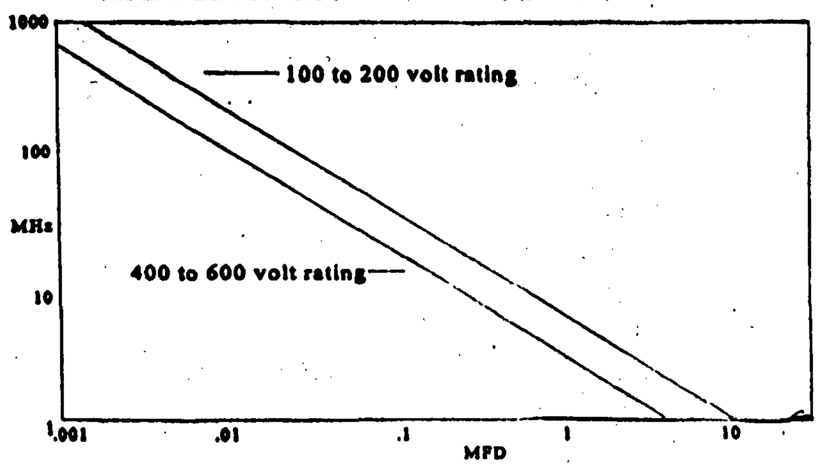


Figure 4.1a. Mylar self-resonant frequency chart.

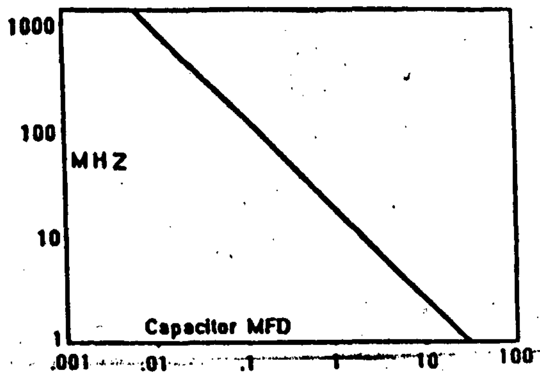


Figure 4.1b Ceramic self-resonant frequency chart.

Veeing the capacitor
Wiring inductance

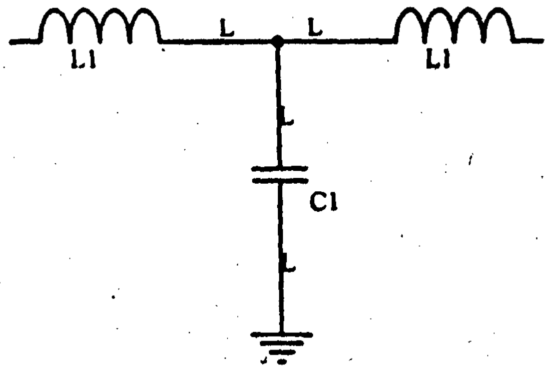


Figure 4.2a The T filter used to demonstrate veeing.

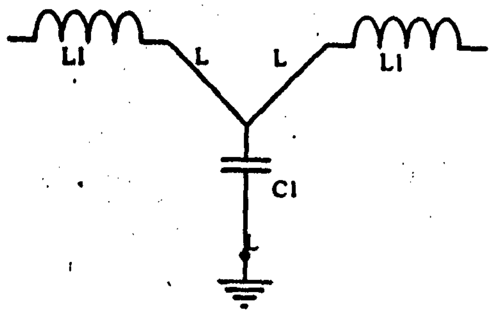


Figure 4.2b Veeing the capacitor.

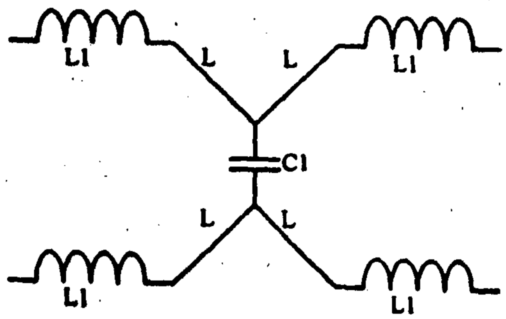


Figure 4.2c The balanced vee.

higher SRF and lower ESR and ESL. An article in the Institute of Electrical and Electronics Engineers magnetic manuals showed that the gain of only 6 dB is realized by these two capacitors in parallel. This theory, though assumes that the lead length would be the same. This would almost make the second capacitor have an ESR and ESL the same as 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 feedthrough capacitor. Another way to lower the ESR and raise the SRF of the capacitor is as follows. In Figure 4.2a, there are four inductors that are often deemed insignificant. The two that make up the capacitor leads add to the ESL and ESR, lowering the SRF. The two in the inductor leads add to the inductor value, aiding the inductance, but are orders of magnitude lower in value. This is the same concept as keeping the lead lengths as short as possible. Although this technique has been around a long a time, the new phrase for this technique is "veeing the capacitor", the old "keep the leads short as possible" trick-especially for the capacitor (Figure 4.2b, 4.2c). The leads on the inductor add only a small amount of inductance to the inductor.

4.2 CAPACITOR DESIGN

4.2.1 WRAP AND FILL TYPE

There are three methods used to build capacitors. The initial method was to attach leads to the foil before wrapping, twisting, 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 equivalent series resistance and equivalent series inductance are very high in these chicklets, so the self-resonant frequency is quite low. These should never be used for an electromagnetic interference (EMI) filter except in the resistance-capacitance R-C shunt or Cauer type for low-frequency problems. If the filter also has high -frequency problems that the R-C shunt must also aid, the quality of the capacitor dictates the higher quality of the extended foil type. The two subgroup are metallized or foil. The film-dielectric-is coated with a thin spray of metallized aluminum that becomes the plate of the capacitor; the foil is similar to aluminum foil. The foil is thicker which carries more current and is therefore better for pulse applications and EMI filter in which there are high harmonic current from off-line regulators. Most filter are built with this type of construction. The metallized film has two advantages: this capacitor can be much smaller for the same capacitor value, and it is self-healing. All dielectrics have small pinholes throughout their length. Thus, when the film sprayed on the dielectric is stressed by the applied voltage, the film often shorts out, causing the film to melt.

The aluminum re-forms, 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 pulse if the total capacitor current is low enough.

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 areas to the right of the first foil and to the left of the second foil make up the margins for both plates. This is then called the extended foil type. The two ends are sputtered over to make the two contacts. The ESR and ESL of the capacitor type are both very low because the current flow travels only the width of the capacitor, giving low resistive losses and little inductance. Each of these is in parallel with the other turns giving even lower ESL and ESR. The SRF is very high. This is the type required by filter designers. The last method is similar to the extended foil type but is used for high capacitor current applications; the tab type. The size and number of tabs depend on the current. These tabs are thin strips of copper that are solder dipped. As the capacitor is wound these tabs are inserted-one for each plate-every so many degrees are added to each tab so that the tabs are

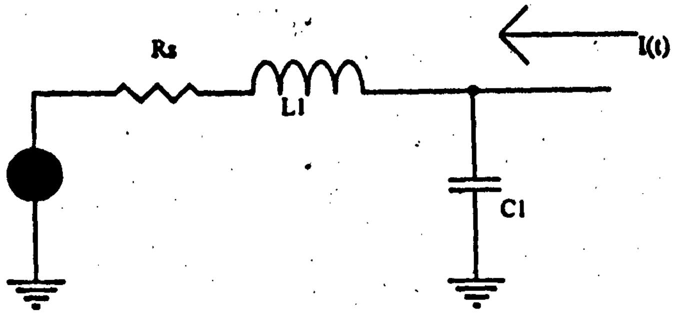


Figure 4·3a Line voltage source E_s, source impedance R_s, a single filter, and I_{tu}.

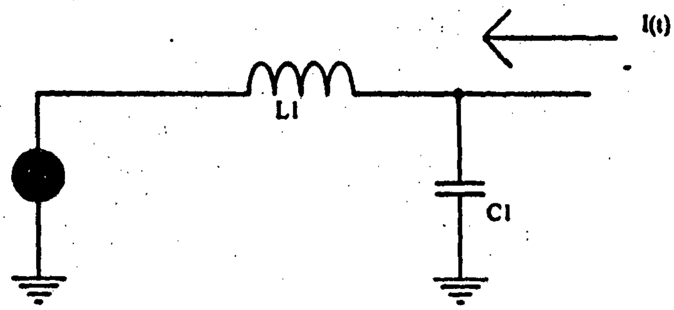


Figure 4·3b Normalized harmonic current without source resistance R_s.

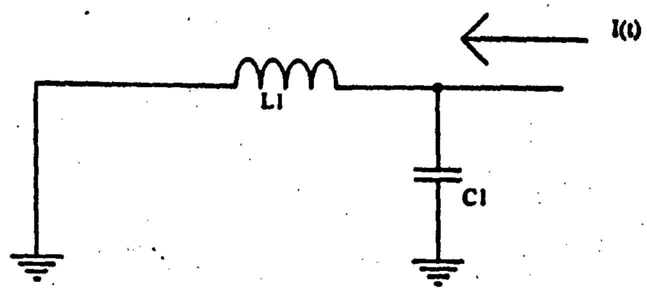


Figure 4·3c Normalized harmonic current without the source voltage.

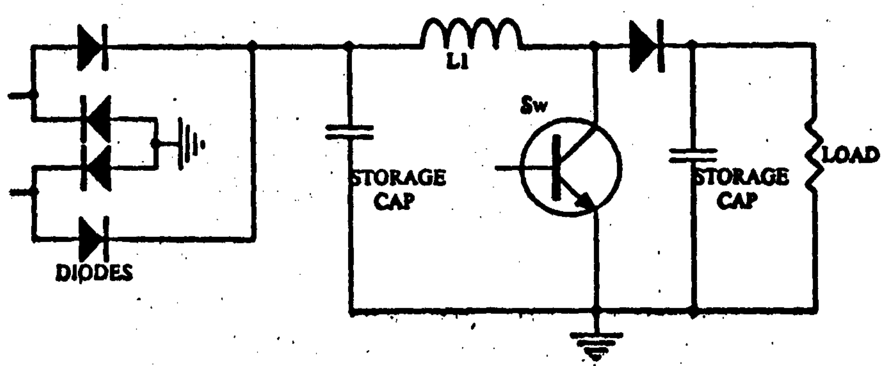


Figure 4·4. Off-line regulator with capacitive load.

uniformly spaced around the side of the capacitor. These tabs, extending from each side of the capacitor, are folded over soldered together to form the contact.

The point is that ac capacitors must be designed to handle ac current at the line frequency and odd harmonics. If this is a single L or T, the capacitor should be of the foil type, not metallized film. In pie and multiple filters, at least the last capacitor, the closest to the load side, should be of the foil type. Some capacitor are not designed properly to handle the full ac current flow or were designed for dc applications. This is especially true if the plates are the metallized film type of capacitor. The capacitor designer may not have designed the capacitor to handle the total harmonic current. Capacitors must handle the harmonic current either from the line frequency side or from the load. This is especially true of the harmonic current created by an off-line regulator, because the foil making up the capacitor plates can be too thin to handle this current. This raises the ESR losses so that the capacitor heats and will fail in the months ahead. Commonly, the capacitor was designed for dc operation, not for ac. The filter capacitor must be selected to handle the harmonic currents of the off-line regulator and any other pulse type of harmonic current (Figure 4.3a, 4.3b, and 4.3c).

4.2.2 OFF-LINE REGULATOR WITH CAPACITIVE LOAD

The most common circuit used today is the off-line regulator (Figure 4.4). The output of the diodes feeds a large storage capacitor that in turn feeds switcher(s). The load impedance also varies from nearly short to open depending on whether the rectifier diodes are turned off. This depends on what part of the sine wave voltage cycle is applied to the diodes. The diodes are turned on if the sine wave voltage, plus the diode voltage drop, is greater than the capacitor's stored voltage at that instant. High current pulses charge the storage capacitor during turn-on (Figure 4.5a). The high- current pulses on the capacitor side of the diodes are often called sine wave

Figure 4·5 a Capacitor pulse charging current through the diodes.

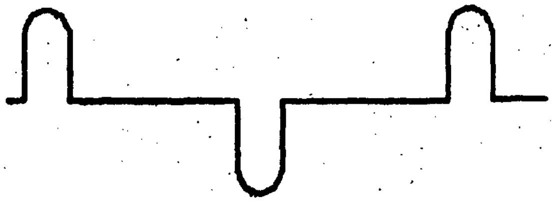


Figure 4·5 b Capacitor pulse charging current on the line side of the diodes.

pulses.

The Fourier equation for this is:

$$\frac{2\tau}{\pi T} + \frac{4\tau T}{\pi} \sum_{N=2,4,6,8}^{\infty} \frac{1}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \cos\frac{\pi N}{2} \cos\frac{2\pi Nt}{T} \quad (4.1a)$$

or

$$\frac{2\tau}{\pi T} + \frac{4\tau T}{\pi} \sum_{N=2,4,6,8}^{\infty} \frac{(1)^{N/2}}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \cos\frac{2\pi Nt}{T} \quad (4.1b)$$

where $\tau/2$ is the pulse width time, T is the time of the full wave (two pulses), and N is the harmonic number. These equations are normalized for a peak of 1A. This is all even harmonics and includes the dc value in the first term. The second, sixth, tenth, and so on, are 180^0 out of phase; the fourth, eighth, twelfth, and so on, are in phase.

On the line side of the diodes (Figures 4.5b), the equation becomes,

$$\frac{4\tau T}{\pi} \sum_{N=1,3,5,7}^{\infty} \frac{1}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \sin\frac{\pi N}{2} \sin\frac{2\pi Nt}{T} \quad . \quad (4.2)$$

and the first sine term can be replaced by

$$\sin\frac{\pi N}{2} = (-1)^{(N - 1)/2}$$

This lacks the dc portion, as it should, and the harmonic component is all odd. The first, fifth, ninth, and so on, harmonics are in phase; the third, seventh, eleventh, and so on, harmonics are out of phase by 180@ Here the last term states that all the terms pass through zero at t=0. This places the fundamental in phase with the line voltage, giving rise to a unity power factor.

If the line harmonic current is neglected, the line voltage is approximately sinusoidal and the normalized load current equation is, as per Equation (4.2) above.

$$I_{(t)} = \frac{4\tau T}{\pi} \sum_{N=1,3,5,7}^{\infty} \frac{1}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \sin\frac{\pi N}{2} \sin\frac{2\pi Nt}{T}$$

or

$$I_{(t)} = \frac{4\tau T}{\pi} \sum_{N=1,3,5,7}^{\infty} \frac{1}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \sin\frac{2\pi Nt}{T} \tag{4.3}$$

The source resistance is low over the range of harmonics of interest and can be eliminated. The inductors and capacitor making up the filter can be acceptable because the frequency of interest is below the cutoff frequency, for the most part. If this is a double L, the two inductors can be added (2L) and so can the capacitors (2C), if they are the same-and they should be). This places the inductor and capacitor in parallel for $I_{(t)}$.

The voltage across the reactive pair, E_c is

$$E_c = \frac{-JX_c\, JX_l\, I_{(t)}}{J(X_l - X_c)} \tag{4.4}$$

The current through the capacitor follows.

$$I_{(c)} = \frac{E_c}{-JX_c} = \frac{X_l\, I_{(t)}}{X_l - X_c} = \frac{4\pi^2N^2LCI_{(t)}}{4\pi^2N^2LC - T^2}$$

Separating the section that belong inside the summation and outside :

$$I(c) \quad = \quad 4\pi^2 LC \frac{N^2}{4\pi^2 N^2 LC\text{-}T} \; I_{(t)} \tag{4.5}$$

and substituting in the I(t) equation (4.3) for the normalized harmonic current,

$$16\pi\tau LCT \sum_{N=2,4,6,8}^{\infty} \frac{N^2}{4\pi^2 N^2 LC\text{-}T^2} \; \frac{(-1)^{(N-1)/2}}{(T^2 - N^2\tau^2)} \cos\frac{\pi N\tau}{2T} \sin\frac{2\pi Nt}{T} \tag{4.6}$$

where $\tau/2$ is the width of one pulse with time in seconds, T is the time of the full wave (two pulses), also in seconds, C is the capacitance, farads, L is the inductance, henries and N is the harmonic number. These equations are normalized for a peak of 1 A. The multiplier (M_{ip}) Equation is

$$M_{ip} = \frac{\pi}{\cos^{-1} [(2V_p\ FC - I_0)/2V_p\ FC] + \sin^{-1} [I_0/\pi\ (2V_pFC - I_0)]} \tag{4.7}$$

where Vp is the power line voltage peak, C is the value of the storage capacitor, farads, Io is the dc load current, and F is the line frequency. The denominator is the value of pie needed in Equation (4.6).

The table below showing the approximate current of all the odd harmonics from I(c). If these are squared and added, and the square root then taken, the total normalized odd current is about 2.713 A upto the 19th harmonic. Column I is the harmonic number, column 2 is the harmonic current, and column 3 is the square root of the sum of all the previous harmonics. This is for a normalized current pf I A. It is high because fo is 4448.5 Hz, which is close to the 11th harmonic of the line frequency. The 400 Hz current through the 2 μF is 0.58 A. Dividing the 2.713

by 0.58 shows that the current through the capacitor(s) is 4.67 times greater in this example. **This all shows that the harmonic current must be allowed for in the design of the filter and filter cutoff frequencies should not fall near any of the odd harmonics of the line frequency. This dictates the use of foil, not metallized film.**

Table 4.1 Harmonic Current Content

Harmonic No	Harmonic Current	$\sqrt{\sum_{1}^{N} I^2_{(c)}}$
1	0.00059	0.00059
3	0.00558	0.00566
5	0.01757	0.01844
7	0.04388	0.04763
9	0.12084	0.12987
11	2.69788	2.7010
13	0.20669	2.7080
15	0.11234	2.7110
17	0.07941	2.7120
19	0.06095	2.7130

The solution here may be to replace this extended foil capacitor with the tab type if the total capacitor current is excessive (Table 4.2, 4.3, 4.4 and 4.5). Here there are two foils and two dielectrics per thickness in the three columns in Table 4.2. Also, some extra allowance for bends. As the line frequency rises, the margins must be increased if the margins are on the ragged edge or if the losses as a result of dv/dt increase. The margins also increase with the voltage, as suggested by the data in

Table 4.2 through 4.5. This must be done to handle creepage. In the design, the first step is to determine the dielectric.

Table 4.2 Voltage Rating

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 4.3 RFI Dielectric Rating

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

Source : From Robert Hassett, VP Engineering.

Table 4.4 RFI Corp. Oil-Impregnated Capacitors: Dielectric

Paper to 125 Degree C

Ply thickness X	DC Volts	Test Flash	Test 1min	AC Volts	Test Flash	Test 1min
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
X(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	2200	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	600	1650	1320
4X 0.0005	2000	5000	4000	720	1800	1440
5X 0.0004	2200	5500	4000	880	2000	1600
5X 0.0005	2500	6250	5000	1000	3500	2000

The required thickness of the dielectric depends on the voltage. This holds true with or without derating, and the dielectric material is most often doubly, triply, or quadruply-lapped as the table suggest to find the proper thickness for the voltage.

Table 4.5 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

Mylar thickness	Plate Pressed	Flat	Round
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

Each has a different value of K, the dielectric constant, needed to calculate the active width of the capacitor Aw. Next, the full thickness of the dielectric plus the film must be determined based on the voltage derating and the preceding tables. If 200 V is needed at 105 degree C, the thickness must be good for 266 V because of the derating based on polypropylene and foil.

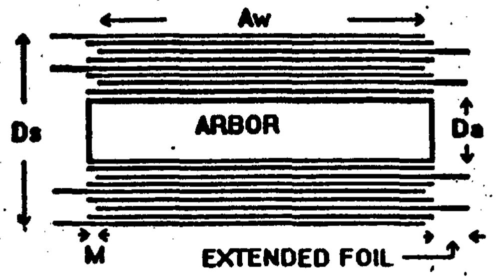


Figure 4-6a Extended foil capacitors.

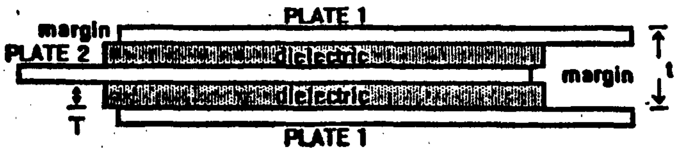


Figure 4-6b Close-up view of extended foil plates with dielectric. T is the dielectric thickness in Mils while t is the thickness of two dielectrics and two foil plate thicknesses

4.2.3 Round - Capacitor : Extended - Foil

Solve for the number of turns, N(Figure 4.6a and 4.6b)

$$N = \frac{D_s - D_a}{2t} \tag{4.8}$$

Where t is the total thickness of the two dielectrics and the two foils, as in Figure 20b. The value of t is in inches. T is the value of the thickness, mils for one dielectric only.

Solve for the mean diameter D_m :

$$D_m = \frac{D_s - D_a}{t} \tag{4.9}$$

Solve for the mean length of turn L_m :

$$L_m = \pi D_m \tag{4.10}$$

Solve for the active length L_t :

$$L_t = NL_m \tag{4.11}$$

Substituting Equations (4.8), (4.9), and (4.10) into (4.11),

$$L_t = \frac{\pi D_s^2 - D_a^2}{4t} \tag{4.12}$$

Now, to know the K value of the chosen material, the dielectric thickness is found in mils (also listed in the charts), and the decimal is moved over three places. The values of K is also a function of shape, and the dielectric is often chosen based on the dissipation factor and size. Mylar has the smallest within its normal temperature range. Typical K values are polyester(Mylar), 900; polycarbonate, 840; and paper (resin or PBT), 580 (wet). Solving for the active width (A_w:)

$$A_\omega = \frac{CKT}{L_t} \tag{4.13}$$

where T is the thickness of the dielectric only and the value of the material is in mils, C is the capacitance value, μF, and K is the dielectric constant. Substituting Equation (4.12) into Equation (4.13),

$$A_\omega = \frac{4CKTt}{\pi(D_s^2 - D_a^2)} \tag{4.14}$$

If the space allotted for this capacitor is known, this equation can help to figure out the width needed to fit the capacitor. The final width will be slightly larger than Aw+2M, where M is the margin for the corona.

4.2.4 Pressed - Capacitor

The arbor is removed after winding, and the capacitor is then pressed flat, If the height (D_y) and the pressed, or flattened, width (D_x) are known, the required diameter D_a of the arbor on which to wind the capacitor can be determined :

$$D_a = \frac{2(D_y - D_x)}{\pi} \tag{4.15}$$

Solving for the number of turns :

$$N = \frac{D_x}{2t} \tag{4.16}$$

where t is again in inches, along with D_x and D_y.

The mean diameter is

$$D_m = \frac{2\,(D_y - 0.215\ D_x)}{\pi} \tag{4.17}$$

The mean length is

$$L_t = \pi D_m = 2\,(D_y - 0.215\ D_x) \tag{4.18}$$

The active length is

$$L_t = L_m N \tag{4.19}$$

Substituting Equations (4.15) through (4.18) into (4.19),

$$L_t = \frac{D_x\,(D_y - 0.215\ D_x)}{t} \tag{4.20}$$

$$A_\omega = \frac{CKTt}{D_x\,(D_y - 0.215\ D_x)} \tag{4.21}$$

With the active width and knowing the needed margins, the full length can be determined for the space needed to fit the capacitor. The value of T is the thickness in mils of one dielectric; the value of t is in inches and is the total thickness of both foils and dielectrics.

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

(a) Dielectric width = $A\omega + 2M$

(b) Foil width = $A_\omega + M + P$

(c) Gage = $A_\omega + 2M + 2P$

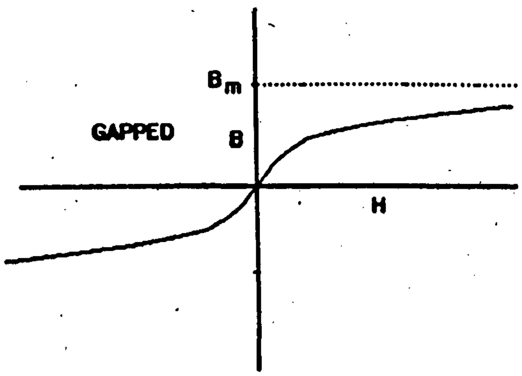


Figure 4.7 *BH* curve for "soft" core material.

Where P is the foil overhang that is sputtered over to make the lead connections on each end or to form the end area for the feethrough type.

4.3 INDUCTOR CONSTRUCTION AND SRF

The EMI filter requires what are termed as soft cores. This means that the core is driven into saturation slowly rather than abruptly as required for pulse transforms and magnetic amplifiers. A hard core can be made soft by gapping it. This technique tilts over the BH curve, making the core soft-harder to drive into saturation (Figure 4.7). The core types used for EMI filters are often Molypermalloy toroids and sometimes the high flux type. Powdered iron are also sometimes used but often cause problems at 400 Hz because they often overheat.

For the same reason, the permeability of MPP or HF types should be limited to 125 and below at 400 Hz line frequency. The core manufacture provides important information about its cores, such as outer diameter, inner diameter, 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, 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 drop, and as the needed current falls, the number of turns increases. For any circular mils per ampere that is chosen, 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 inductor of 0.4 for the window of the toroid. This is not the 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 factors.

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

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 DCR, M_{pl} is the magnetic path length, centimeters, and 0.4 is the typical window fill factor for toroids.

The Magnetics, Inc, core for 125 permeability is 55438, with window area, 842,700 circular mils; magnetic path length (M_{pl}), 10.47 cm; and weight, 0.4 pounds. Using the circular mils per ampere as 600, H follows.

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

Using the BH curves furnished by the manufacturer yields the approximate values of B for each permeability; (Table 4.6). The given, or initial, permeability is listed as a reference, and B estimated from the BH curve. With this information, the approximate value of the actual permeability of the core calculated. The watts/pound follows from these formulas, also provided by the manufacturer for MPP cores (Table 4.7). From these equations W/pound was determined, (the last column in Table 4.7) followed by multiplying the w/pound by the pounds furnished by the manufacturer, here 0.4. The watts/pound and the total loss columns in Table 4.7 indicate a hefty jump from 60 to 125 permeability and continue hefty increases upward. Using a permeability of 125 is questionable.

Table 4.6 Permeability and Flux of a Core

Give u	Actual u	B	400 Hz (W/Pound)	Total loss
14	13.69	900	0.0377	0.0151
26	25.86	1700	0.1827	0.1236
60	45.64	3000	0.3091	0.1236

Give	Actual		400 Hz	
u	u	B	(W/Pound)	Total loss
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 4.7 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	$0.890 \times 10^{-11} \times F^{1.26} \times B^{2.11}$

Reducing the frequency back to 60 Hz makes a large difference so that all 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 turns well below the maximum. This is when the window area of the core is less filled.

Table 4.8 Constant Current at 5 for a MPP Core Showing Al Values

Permeability	Al	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

The C core are used at higher current lines, mainly the power line filters used for shelters, EMI test houses, screen rooms and secure communication application.The thickness of the magnet core tape to wind these C cores should be able to handle the fifth harmonic of the power line frequency and the potential harmonic current from off-line regulators to avoid heating of the core. This is because of the harmonics on the line and from these off-line regulators. If the filter is used in areas where the power feed is more resistive with less harmonic content, this principle can be violated. When 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 a very low flux density and allow the line harmonics to heat the core. The first allows a smaller core that is more expensive, and the last allows for a less expensive type of core, but this is often offset by the larger size and weight. One other factor not to overlook is the smaller thickness of the tape used in the first method, which has a

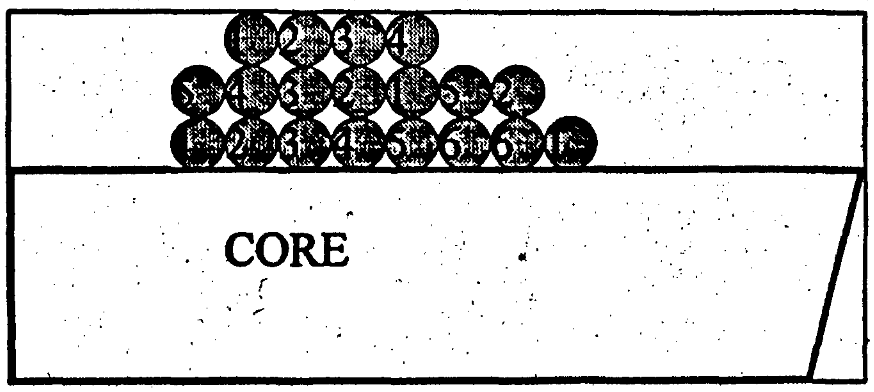


Figure 4·8 Progressive wind: a better SRF.

lower stacking factor, decreasing the A_c, the core cross-sectional area. The size of the value of A_c, must then be increased. The higher switcher frequencies and parasitics 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 by the skin effect of the wire. Both methods can dissipate this high-frequency energy.

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". If many of these inductors are needed, several C cores may be wound at the same time, depending on the wire size, but toroids must be wound one at a time. Therefore, C cores are quicker to manufacture.

The SRF of the inductor is a function of the size and the number of turns. There are several ways to raise the SRF for the toroid. The first is to use a progressive winding technique, such as six winds forward around the core followed by five turns back over the previously wound turns, then another six forward over the previously counterwound turns, followed by another five back (Figure 4.8). This is continued 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 degree of the circumference of the core window. A barrier or space is placed between the point of start to the point of finish. This barrier is used so that the two ends do not slide together, and can be of any material, even tape. This technique decreases the turn-to-turn capacitance in series across the inductor. The disadvantage is that it looks like a 5-year-old child wound it.

Side-by-side windings in which there is less than a full 360 degree around the core are also adequate, giving a lower capacitance from turn to turn for the same reason. The voltage from turn to turn is very low. When many layers are needed, it is obvious that layer 2 compare with both layers 1 and 3 has much higher turns, creating a higher voltage gradient and giving much higher capacitance from turn to turn.

This guarantees, a very low SRF, which compromises, the filter action. The second method is to keep the winding off the core. The core is tapped with several thicknesses to keep the turn-to-core capacity as low as possible (Figure 4.8). Toroids can be purchased that are already coated. This capacitance is from turn to core and is somewhat in parallel so add and are more so directly in parallel with the inductor. Figure shows tape with a single layer of turns.

4.4 INDUCTOR DESIGN

The RMS value of the current is usually what the filter designer is given design the filter inductors. Designers are not given this high peak current needed to design the inductors. The question is, what will these high current pulses do to the filter inductors? Most inductor designers design the inductor somewhere around half flux density at the RMS value of the current. This is the wrong approach to the design of this AC EMI filter inductor because the inductor will saturate. The equation is

$$H = \frac{0.4\pi N I_p}{M_{pl}} \tag{4.23}$$

where N is the number of turns, T_p is the peak current feeding the charging capacitor, and M_{pl} is the magnetic path length. H is the magnetizing force. The maximum flux density B_m is a constant for the core type and material. The relationship between B and H is the permeability μ. Thus, as H increases because of the large current pulse I_p, the permemining the inductance L. The core material is spoken of as "soft" because of the BH curve. These are S shaped, or sigmoid, as

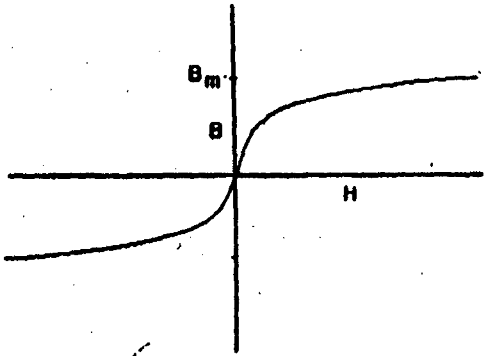


Figure 4-9 *BH* curve of a "soft" (sigmoid) core.

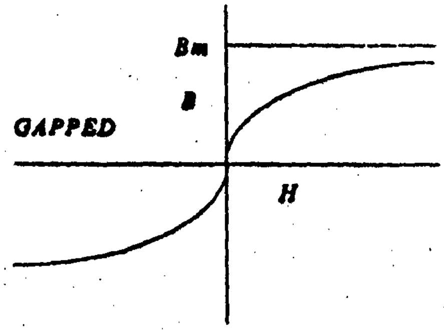


Figure 4-10 Gapped core. The *BH* curve is more linear and less sigmoid.

in Figure 4.9 and are not made using square loop material. These cores require a strong magnetizing force, H to drive the core into situation. A "hard" core, or square loop core, is driven quickly into saturation. The soft core is the type of core material needed for EMI filters. Cores that posses square loop characteristics are gapped, if they are used at all, reducing the hard magnetic core to a soft core. Soft cores are also often gapped to make them softer (less sigmoid) and for dc applications.

$$L = \frac{0.4\pi\mu\ N\ A_c \times 10^{-8}}{M_{pl}} \tag{4.24}$$

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

If the inductor saturates during this current pulse peak. Then 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 top 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 if high-quality capacitors are used. Producing a very low impedance to this noise compared with the load and line impedance in parallel.

$$\Delta V_{rms} = \frac{I_0}{2\sqrt{3}\ FC} \quad \text{or} \quad \Delta V_{rms} = \frac{0.2887\ I_0}{FC} \tag{4.25}$$

where I_0 is the dc load current, F is the ripple frequency (twice the line frequency in single phase), and C is the capacitor in farad. Substituting in equation (4.25),

$$I_{rms} = 2\pi FC\ \Delta V_{rms} = 1.814 I_0 \tag{4.26}$$

I_{rms} is a straight function of I_o, the dc load current, not a function of the storage capacitor value. In the past, many thought that as the size of the capacitor increased, the conduction width decreased and the current pulse height increased. All this affects the I_{rms} of the line: the higher current peak is offset by the narrower width. The point is that the RMS line current can be calculated and given to the EMI filter designer. This may allow inductors to be undersized and to saturate on the peaks of the current pulse.

Another equation relating the capacitance to the peak-to-peak ripple is

$$C = \frac{I_0 \times 10^{-6}}{2F \ (P\text{-}P)} \ \mu F \tag{4.27}$$

where P-P is the peak to peak voltage and F is the line frequency. Another way to avoid this problem is to design and inductor using a gapped core (Figure 4.10). This tilts the BH curves, 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.

4.3.1 OFF-LINE REGULATOR WITH INDUCTOR AHEAD OF THE CAPACITOR

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

The critical inductance is :

$$Lc = \frac{R_o}{6\pi F} \tag{4.28}$$

where R_o is the load resistance calculated using the lowest, worst case current and the highest line voltage. 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 reduces the EMI filter weight because the filters need not handle this high peak pulse current.

Another common method is to design with A_l values. This is normally listed as millihenries per 1000 turns and sometimes as microhenries per 100 turns. This technique can be used for all inductances, including common mode, as long as the A_l value is listed by manufacturer or known by some other means. This is developed from the following

$$L = \frac{0.4\pi \ \mu N^2 \ A_c \times 10^{-8}}{M_{pl}} \tag{4.29}$$

$$A_l = \frac{0.4\pi \ \mu A_c \times 10^{-8}}{M_{pl}} \tag{4.30}$$

All these A_l terms are either constants or are common to the core selected, where A_c is the iron cross-sectional area, μ is the permeability, and N is the number of turns, the variable. The Ac value is normally provided by the manufacturer.

$$L = A_l N^2$$

The inductance is known, and the number of turns must be found. The A_l value is

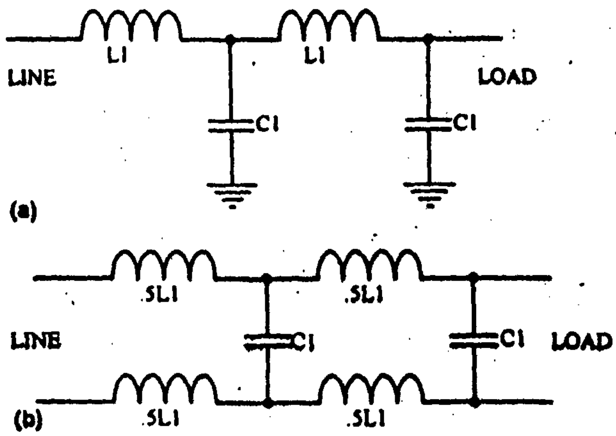


Figure 4.11 The multiple L network converting from unbalanced (a) to balanced (b).

given, so

$$N_2 = N_1 (L_1/A_l)^{1/2}$$

where N_2 is the needed number of turns, N_1 is the known number of turns , L_1 is the needed inductance in the same units as A_l (millihenries), again, A_l is the known millihenries associated with the turns given.

4.5 CONVERT FROM BALANCED TO UNBALANCED OR THE REVERSE

Conversion from balanced to ubalanced circuits, 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.

To balance the filter, Simply divide inductors in half,both the live 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 the farthest apart would have to be mounted in the same way in the hope that the extra distance would reduce the coupling), and keep the capacitor leads as short as possible. Figure 4.11 illustrates before and after.

CHAPTER - V

COMMON MODE COMPONENTS

The causes of spikes or electromagnetic pulses on the line include lightning, shutting off of large inductive equipment, and magnetic pulses created by nuclear activity. Lightning and nuclear activity create common mode pulses between the lines and ground. The equipment type creates a differential mode type of pulse between the lines. On the load side, the switcher is a leading culprit 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 isolating the power supply reduces these common mode conducted emission pulses. The one exception is the primary to secondary capacitance of the transformer. Power supply filter capacitors pumping current to, or from, ground should be eliminated. To eliminate the common mode, the electromagnetic interference filter employs common mode chokes and feedthrough capacitors to ground.

5.1 CAPACITOR TO GROUND

For 400 Hz, the limit of the capacitor to ground is 0.02 µF for Military Standard 461. The medical leakage current specification is even harder to meet. If the device touches a patient, the total system leakage is limited to 100 µA. This means that most power supply people want the filter restricted to 20-40 µA. It is difficult to have common mode losses common mode loss specifications with capacitors to ground this small. There are two schools of thought on this problem because the specifications governing it, are not clearly written. The first group thinks that this is the total capacitor to ground. In threephase four-wire circuits, the capacitor limit value for 400 Hz is 0.02 µF. This capacitor value is then shared. The second group thinks that this is the maximum per line whatever the number of lines. If the system is well

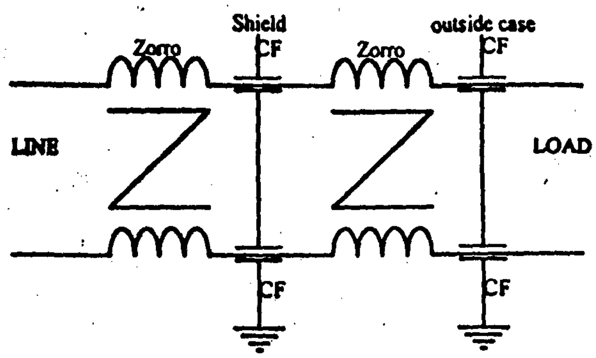


Figure 5-1 Double-Zorro common mode filter.

Notes
1. 3 windings on one core wound as shown.
2. Wire to be tightly wound and evenly distributed over segment.

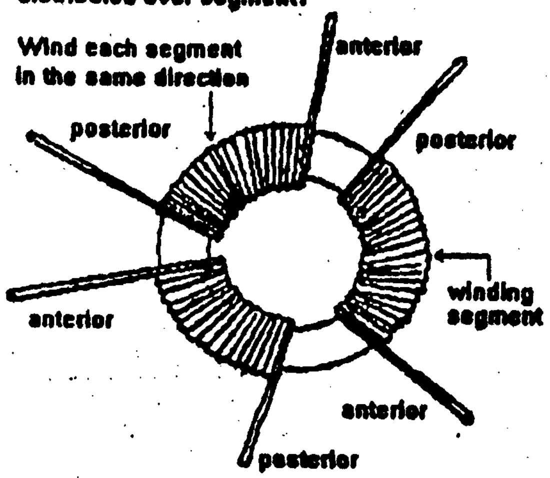


Figure 5-2 RFI Corp. three-phase three-wire common mode induc

balanced, the currents nearly cancel through these capacitors.

5.2 Z FOR ZORRO

Ferrites are often used for common mode inductors because they have the very high Al values required for these common mode filters. In the common mode case, all lines are tied together so that all the differential inductors are in parallel in the balanced design. To obtain 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 (Figure 5.1). For best results using this method, the Zorro inductor is placed at the low-impedance end, or the line side, and the feedthrough capacitors on the high-impedance end, or load side. Here, where they are being split, the first Zorro is placed at the line and the first feedthroughs are the last in the section-mounted in the shield. The next section starts with the Zorro and ends with the feedthroughs at the end. Try to use an even number of differential mode filters so 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, and then followed by the feedthrough capacitors. The second section would be the same.

A second common mode inductor is placed in the filter, one in the front half and one in the second half. This forms a double L in the common mode, greatly reducing the size of the common mode inductor, and now the specification can be reached most of the time. The inductor's magnetic fields buck or cancel for the differential mode and have a high magnetic gain for the common mode. Figure 5.2 shows a three-phase delta with spacing between the windings.

5.3 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,10 MH, for example. Either winding should read the indicated inductance if the measurement employed a good inductance bridge. The reason is from the inductance formula no 4.29 above.

A toroid ferrite common mode inductor can be designed using the A_l value of the core. This would only be used, though, 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 dill factor is no longer 0.4 but now is reduced to 0.2 to fit the two windings.

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

$$N_t = \frac{\pi(I_d - W_d)}{W_d} \tag{5.1}$$

If the wire diameter is much smaller than the core diameter this approaches:

$$N_t = \frac{\pi I_d}{W_d} \tag{5.2}$$

Dividing the turns by 2, and using the integer to solve for the inductance when the A_l, the A_l turns and the turns that the core can support in a single layer wound only less than halfway around are known. If this is no greater than the inductance needed, choose another core with a larger I_d.

Once a core is found in which the inductance is greater than needed, the number

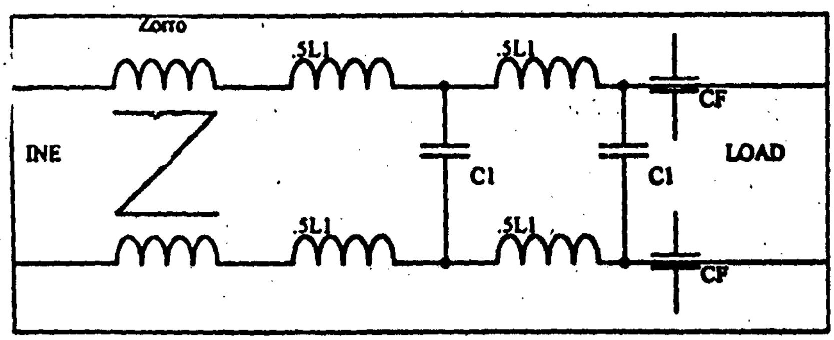


Figure 5.3 The common mode and differential mode filter.

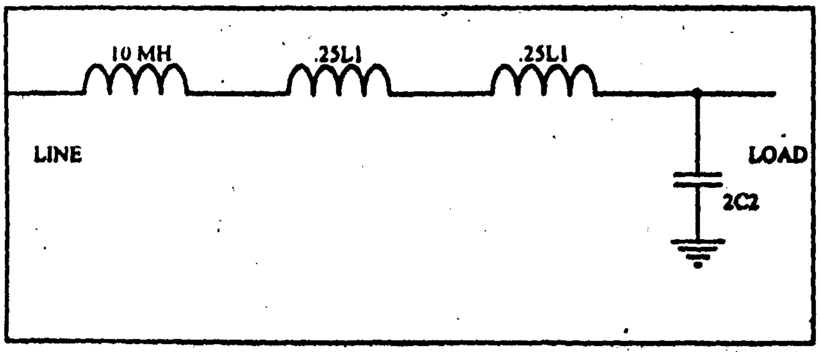


Figure 5.4 The common mode converted to the unbalanced filter of Figure 7.5.

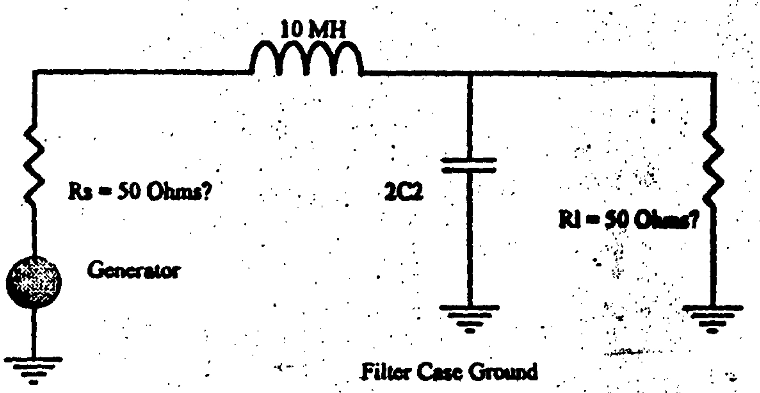


Figure 5.5 The converted test method to calculate the common mode values.

of turns required using the A_l equation is resolved . The number of turns necessary is found and the higher integer is used.

The problem is now that the method of sizing the proper core size is known, how was the value of 10 MH determined for the Zorro? Two issues must first 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 the normal mode. If the common mode inductor is followed by several balanced normal mode networks, this must be converted back to unbalanced differential to ease the calculations.

In Figure 5.3, the Z is first followed by two L filters in turn, followed by feedthroughs. 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 0.5L each, giving an effective value of 0.25L. The following capacitor C1 is not in the circuit of the common mode, that is, equals zero. The following two inductors give the same result as the last, 0.25L. C1 again equals zero, leaving the final component of C2, which is doubled in value. The results are shown in Figure 5.4. The two differential mode coil total is 0.5L but often end as part of the overall headroom for the common mode part of the filter. The reason is that the total value of these two coils may be around 400 μH, which is an order of magnitude lower than the Zorro inductor. C2 may be limited in size by the specification, as in Military Standard 461, in which the maximum for 400 Hz is 0.02 μF but now totals 0.04 μF.

5.4 EQUATIONS FOR THE COMMON MODE FROM DIFFERENTIAL MODE

The method used to arrive at Figure 5.5 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, equation based on Q (charge) rather than

I (current) of the two network are the easiest to work with, generating the matrix in Figure 5.5. This is from impedance matrix equations.

$$Vin_0 = \begin{bmatrix} LS^2 + RsS + \dfrac{1}{C} & \dfrac{-1}{C} \\ \dfrac{-1}{C} & R_lS + \dfrac{1}{C} \end{bmatrix} \begin{pmatrix} Q1 \\ Q2 \end{pmatrix}$$

Δ The determinant of the matrix is

$$\Delta = \frac{LCR_lS}{C} \left[S^2 + \frac{(R_sR_lC + L)S}{LCR_l} + \frac{R_s + R_l}{LCR_l} \right] \qquad (5.4)$$

Substituting in the initial requirements of V_{in} and 0, and solving for Q_2;

$$\begin{bmatrix} (LS^2 + R_sS + 1/C) & V_{ins} \\ -1/C & 0 \end{bmatrix}$$

and

$$Q2S = \frac{V_{ins}}{C\Delta} = \frac{V_{ins}}{LCR_lS\ [S^2 + (R_sR_lC + L)\ S/LCR_l + (R_s + R_l)/LCR_l]} \qquad (5.5)$$

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

V_{os}/V_{ins}, where $Q_{2s}S = I_{2s}$ and $I_{2s}R_l = V_{os}$:

$$\frac{V_{os}}{V_{ins}} = \frac{1}{LC\ [S^2 + (R_sR_lC + L)\ S/LCR_l + (R_s + R_l)\ /LCR_l} \qquad (5.6)$$

This equation has been published in many articles, but most often the authors do not include R_s, the source impedance, and often do no allow for the two feedthrough capacitors, in parallel, which are now doubled in value. Most solve this by completing the square. This means, Equation (5.5),

$$\left(\frac{R_s R_L C + L}{2LCR_L}\right)^2$$

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 ω^2, negative. This makes the solution a hyperbolic function and very lossey, as suggested by the test setup. Most omit the Rs, making the value of a reduce to 1/(2RC), the dampening factor. This also reduces the last term of the denominator to $1/(LC) - \omega^2$.

$$a = \frac{R_s\, R_L\, C_{ll} + L}{2LCR_l}$$

$$\omega = \sqrt{\left(\frac{R_s R_L C + L}{2LCR_l}\right)^2 - \frac{R_s + R_l}{LCR_1}} \qquad (7.10)$$

$$\frac{1}{LC[(S + a)^2 - \omega^2]}$$

giving the following answer, where a is the dampening factor:

$$\frac{1}{LC}\left(e^{-at} \sin h(\omega t)\right)$$

Another way to know that the common mode will always be lossey is by setting the main denominator Equation (5.4) to:

$(S + a)(S + b)$

So that a + b is equal to the middle term and ab is equal to the end term of the main denominator Equation (5.4), repeated here:

$$\frac{V_{os}}{V_{ins}} = \frac{1}{LC \quad [S^2 + (R_sR_lC + L)S\ /LCR_l + (R_s + R_l)\ /LCR_l]} \tag{7.8}$$

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

$$a = \frac{(R_sR_LC + L) - \sqrt{[(R_sR_LC+L)\text{-}4LCR_L(R_s + R_L)]}}{2LCR_l}$$

$$b = \frac{(R_sR_lC + L) + \sqrt{[(R_sR_LC+L)\text{-}4LCR_L(R_s + R_L)]}}{2LCR_l}$$

and takes the form of

$$\frac{e^{-at} - e^{-bt}}{b - a}$$

The term b - a reduces to

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

This cancels LC, so the final answer is

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

In the normal test arrangement, Rs and Rl are both 50 ohms, and C, because of leakage current, is whatever the specification requires. The value of the capacitor is doubled here because the two feedthroughs are in parallel and, therefore, add.

CHAPTER - VI

COMPROMISES OF THE FILTER

The filter may test well using the specified test method and still fail to work as designed for all too many reasons, and the most prevalent are discussed here.

6.1 FILTERS IN CASCADE

This happens when more than one filter type follows another (Figure 6.1). These filters detune each other, especially if any, or all, have higher circuit Q of 2 or better. The higher circuit Q increases the potential of these filters to oscillate, and this moves the cutoff frequency farther into the normal band pass. This problem has been known to reduce line voltage to the point that the rack equipment fails to work. If any of the filters are initially designed incorrectly, this accentuates the problem, even more so if there are multiple double feeds with other filters in cascade from the same power line filter. The cascaded capacitors total and cause higher line and harmonic currents that add heat to the filters. This also adds some additional resonant rises within the filter chain, along with added drains or resonant drops.

6.2 POOR FILTER GROUNDING:-

A proper designed filter may check out fine in the electromagnetic interference (EMI) test laboratory. 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 ground through input feedthrough 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 to ground with the proper nuts and washers. The gaskets give thousands of grounds by this technique. This also carries the chassis ground plane through the

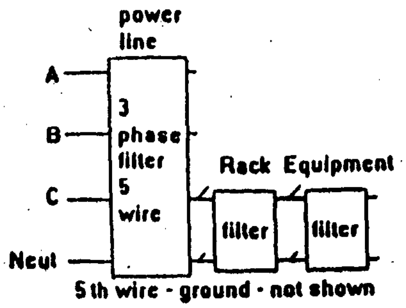


Figure 6.1 Three filters in cascade (in tandem).

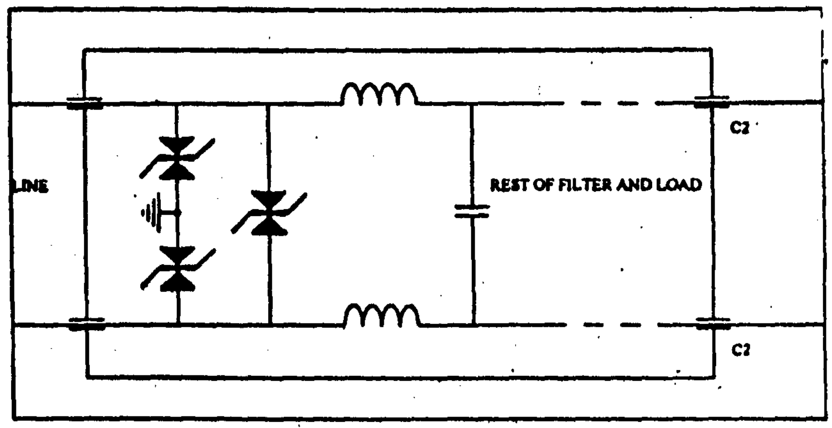


Figure 6-2 The π filter showing the lack of a good ground.

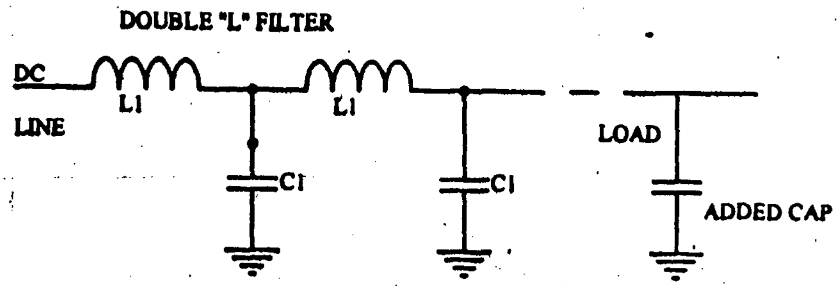


Figure 6:3 The added unknown or hidden capacitor.

cutout holes in the chassis. If this ground is not provided, the filter fails to live up to its decibel rating. Without a good ground, the filter@ feedthrough capacitors and other components to ground cannot work (Figure 6.2).

6.3 UNKNOWN CAPACITOR IN THE FOLLOWING EQUIPMENT

This applies only to dc filter. Double capacitor may not compromise every situation in dc filtering, but the additional component adds cost, volume, and weight (Figure 6.3). The added capacitor detunes the filter.

6.4 THE INPUT AND OUTPUT TOO CLOSE TOGETHER

The input and output studs of a filter should not be placed on the same front face. The inputs and outputs are too close together. 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, or vice versa. The best design is a long filter body with the terminals on the opposite smaller faces. This way the unwanted energy is cleaned up as the signals travel through the filter sections toward the opposite end.

6.5 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 . The gasket is often omitted, and the filter has a poor ground return. Any feedthrough capacitors and MOVs to ground or case within the filter are thus compromised. Sometimes, the filter is removed for various reasons or replaced, and the same gaskets are reused. This again educes 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 do without a gasket. Another reason to use gaskets is to complete the missing ground path through the holes through which the filter is 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). 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 approximately the minimum dc and ac resistance between the filter and ground.

CHAPTER - VII

WAVES AS NOISE SOURCES

The waves discussed here are the more commonly encountered within electrical systems. These waves are not as simple as drawn here: some parasitic oscillation from the transformer will be superimposed on the waveform. These waveforms also have added rise and fall times, and some of these 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.

7.1 THE SPIKE WAVE

Spikes are a common waveform, often seen when double-ended, or Royer, circuits are used (Figure 7.1), that is, one switcher is 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, etc. 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 that 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 drop the dc voltage. This impairs switcher operation, and most of this spike energy is returned to the electromagnetic interference (EMI) filter. With an off-line regulator, the storage capacitor is often very large, in the 500-2000 μF range. 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 feedthrough capacitor should be tied in 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

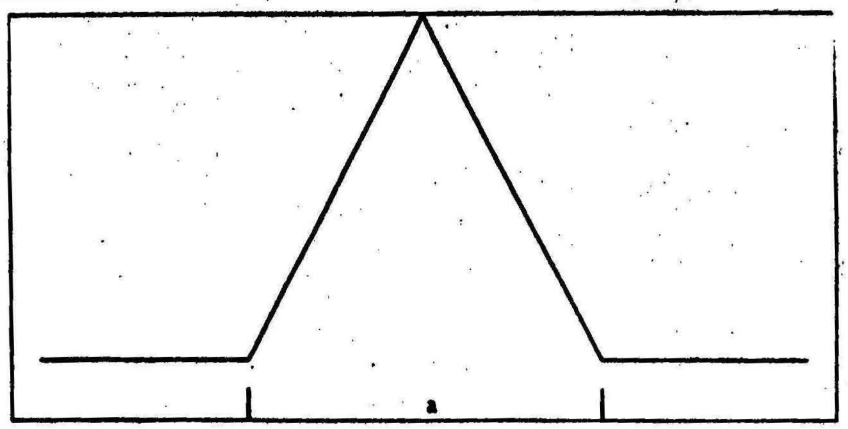


Figure 7-1 The spike wave.

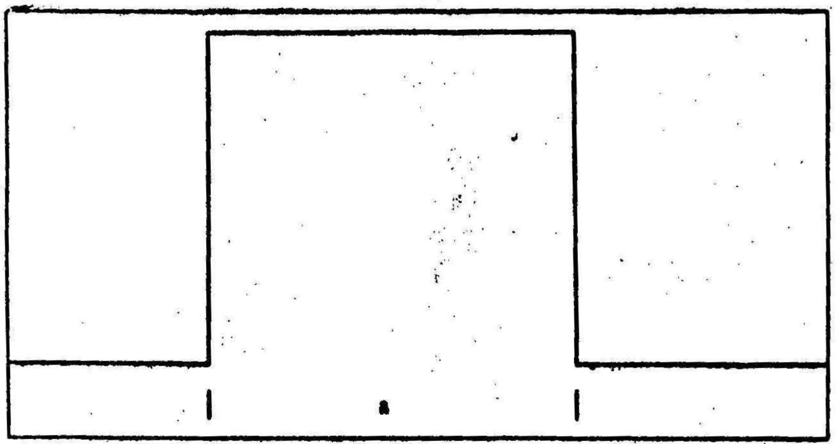


Figure 7-2 The pulse wave.

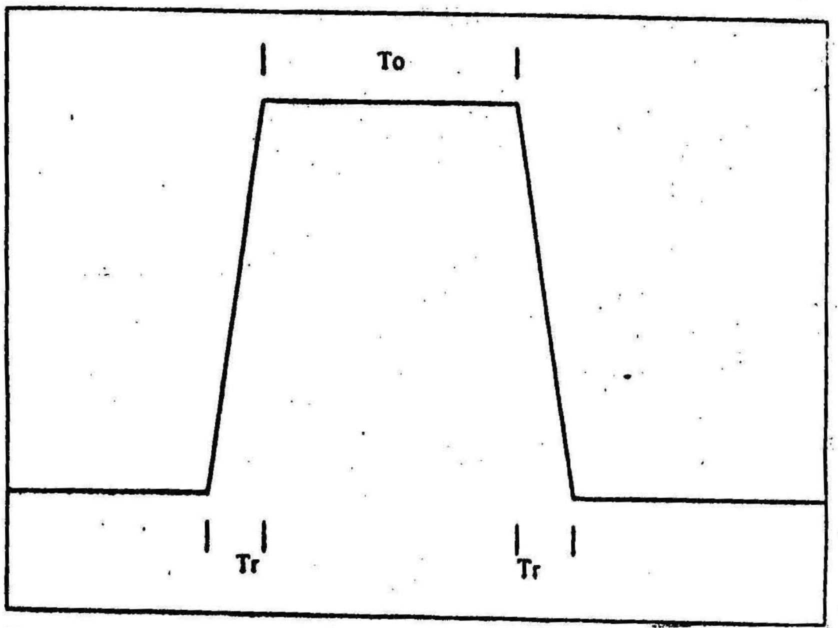


Figure 7-3 The trapezoid wave form.

power feed and the switcher transformer. This should be a good-quality feedthrough, if the switcher is not isolated, or a good-quality line to line capacitor if it is isolated.

The spick is a pulse, and it requires the ability to handle higher pulse currents. Therefore foil, not metallized film, should work with the proper derating. Ceramic capacitors work very well. The leads must be short as capacitor with the same lead length will not give the higher SRF required. The equation for the spike is as follows:

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

where a is the pulse width and

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

and the differential is

$$\frac{Ea}{2\pi a} \sum_{N=2,4,6,8}^{\infty} \frac{1}{N} \left(\frac{\sin X}{X}\right)^2 \cos \frac{2\pi N t}{T}$$

7.2 THE PULSE WAVE

The pulse is similar to the quasi-square waveform but pulses in the same direction, like the spike (Figure 7.2). This is similar to the Royer, in which 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 for the pulse is :

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

where a is the pulse width. The differential is as follows :

$$= \frac{4E}{T} \sum_{N=2,4,6,8}^{\infty} \sin \pi\, Nt \quad \sin \frac{2\pi Na}{T}$$

7.3 THE TRAPEZOID WAVE

The trapezoid waveform is more realistic than the pulse or the quasi-square because they have a step function or a zero rise and fall time. The trapezoid has a rise and fall time like that in the real world (Figure 7.3), 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 chosen using the spike method 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.

$$\frac{2E\,(T_0 + T_r)}{T} \sum_{N=2,4,6,8}^{\infty} \frac{\sin \pi\, N(T_0 + T_r)\,/T}{\pi\, N(T_0 + T_r)} \; \frac{\sin \pi\, NT_r/T}{\pi\, NT_r\,/T} \cos \frac{2\pi Nt}{T}$$

The differential is as follows :

$$4E \sum_{N=2,4,6,8}^{\infty} \sin \frac{\pi\, N(T_0 + T_r)\,/T}{T} \; \frac{\sin \pi\, NT_r\,/T}{\pi\, NT_r} \sin \frac{2\pi Nt}{T}$$

7.4 THE QUASI-SQUARE WAVE

The quasi-square is the wave applied to the gate, or control, of the switch, and

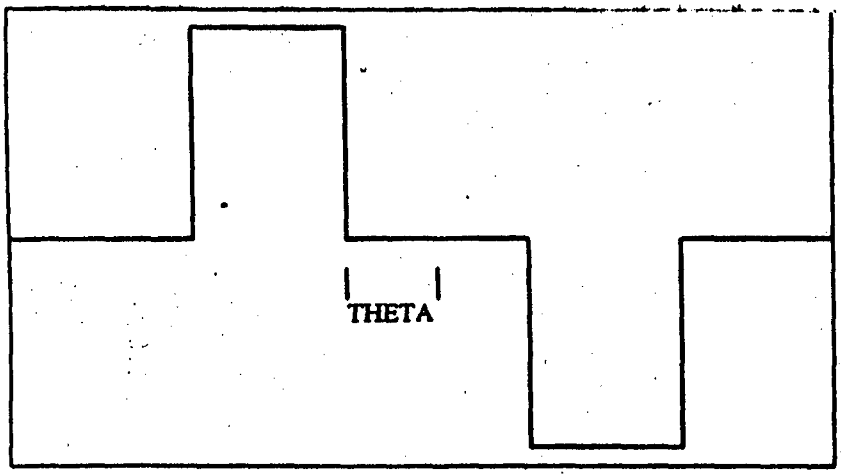


Figure 7·4 The quasi-square.

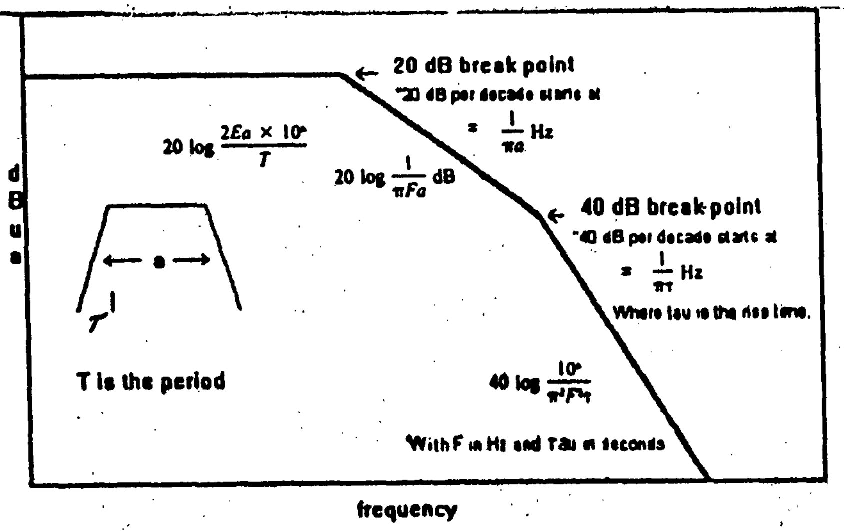


Figure 7·5 The dBuA per megahertz.

the reciprocal is applied to the opposite control (Figure 7.4). 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 four dwell angles 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.

This quasi-square wave, along with its output wave, the pulse are not real world. Both lack the rise and fall time of any wave. This is a result of 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 change, discharge, or change. This produces 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 Figure 7.5.

The equation for the quasi-square wave is

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

The differential is as follows :

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

7.5 REASON TO DIFFERENTIAL

Each pulse just listed has been differentiated for good reason. If each of these waves represents current pulses, the H field must be calculated. If each waves 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 cuts the surrounding surfaces only during turn-on and turn -off. These waves are often pulses with a peak amplitude of many amperes. They generate strong varying H field, leading to radiated energy . The formula is as follows:

$$H = \frac{0.4\ \pi\ NI_p}{M_{pl}} \tag{7.1}$$

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 I 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 :

$$H = \frac{0.4\ \pi\ NI_p}{2\ \pi\ R_r} = \frac{0.2\ NI_p}{R_r} = \frac{0.2\ I_p}{R_r}$$

What is needed is the differential of H with respect to t, and the only variable that depends on t in the right-hand equation is I_p.

$$\frac{dH}{d_t} = \frac{0.2}{R_r}\ \frac{dI_p}{d_t}$$

So, 0.2 divided by R_r, n = 1, multiplied by 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.

7.6 THE POWER SPECTRUM

How much power is in the various waves discussed in the earlier section? These form envelopes that provide the peak power, which varies with the amplitude of the current I. If the current pulse width is a and T is the period, the equation is as follows (Figure 7.5):

$$\text{Pulse power} = 20 \log \frac{2Ea \times 10^6}{T} \text{ dB/MHz}$$

where a is the pulse width and E is the amplitude. This gives a flat line across the frequency spectrum to the 20 dB per decade, or 6 dB per octave, breakpoint. This point starts at the frequency:

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

where a is the pulse width in seconds and the decibel level after the break point is:

$$20 \log \frac{1}{\pi F a} \text{ d8}$$

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 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.

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

The added loss after the 40 dB per decade breakpoint is:

$$40 \log \frac{10^6}{\pi^2 F^2 \tau} \text{ d8}$$

where F is in hertz and τ is in seconds.

CHAPTER - VIII

INITIAL FILTER DESIGN REQUIREMENTS

8.1 DIFFERENTIAL MODE DESIGN GOALS

8.1.1 Input Impedance of the Filter

The idea is to make the filter transparent to the line. This is ideal for both ac and dc systems and means that the load impedance is transferred to the input of the filter at the line frequency, along with most of its harmonics. The harmonic content depends on the quality of the line. These harmonics are odd harmonics. The filter cutoff frequency to accomplish this goal should be above the ninth harmonic. This is because the level of any harmonic above the ninth, even for the poorest quality line, is insignificant. This is why the rule is set at the tenth harmonic of the line frequency.

$$Z_{if} \text{ at } N_{ft} = R_l$$

Where Z_{if} is the input impedance of the filter, N is the harmonic number and is set equal to 10, and R_l is the load impedance at the same frequency. Practically speaking, the load impedance is constant over the frequency range of interest here.

This goal is hard to reach, especially for high-current filters in which the required losses are heavy at the low-frequency end. If the cutoff frequency allows the filter to attenuate the fifth harmonic and some of the third harmonic, higher capacitor current result. These heat the capacitors because of the current through the ESR. The higher harmonic currents through the inductors also heat the capacitors because of the current through the dc resistance and higher core losses. This raises the operating temperature of the filter. The low cutoff frequency lowers the resonant rise frequency and raises the circuit Q. Either of these increases the odds that the filter will oscillate.

8.1.2 Output Impedance of the Filter

Here again the filter should be transparent to the load. If the input impedance goal is met, the output impedance goal is also normally met. Meeting these two goals makes for better filter operation.

$$Z_{of} \text{ at } N_{fi} = R_s$$

where Z_{of} 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 is true for most lines to 5 kHz before any rapid increase to higher line impedances is reached. At 10 kHz the impedance is about 4Ω on most lines.

8.1.3 Input and Output Impedance for a DC Filter

Both of the two preceding requirements are easily met for a dc system unless the load is a switcher. Here, the output impedance of the filter must be very low. This statement rest on the premise that the switcher designer has not anticipated this

$$Z_{of} \text{ at } F_{sw} << R_l \text{ at } F_{sw}$$

where F_{sw} is the switch frequency and the rest of the terms are the same. The same holds true at the tenth harmonic of the switcher frequency. The switch may not be all that starved at the fundamental any yet be starved at the ninth or eleventh harmonic, if the output impedance is slightly inductive or the output capacitor is above its self-resonant frequency. 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 ensure that the drop

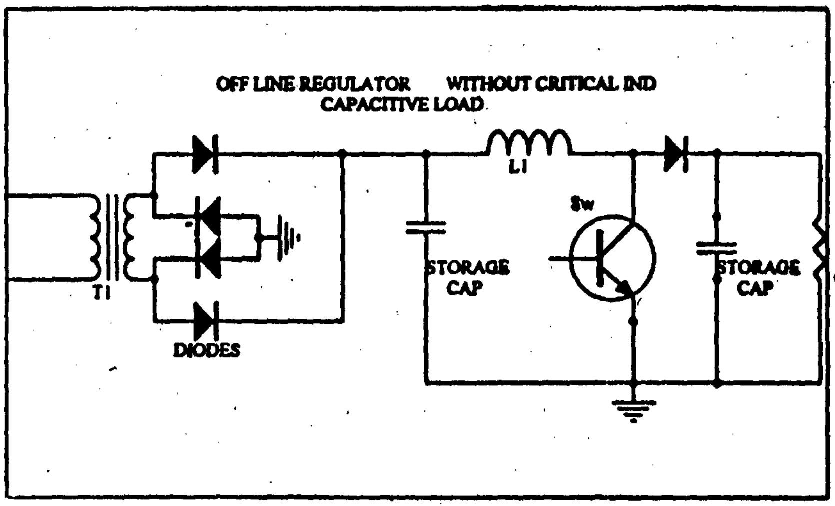


Figure 8.1 The switcher as the noise source in an off-line regulator without critical inductance anu with capacitive load.

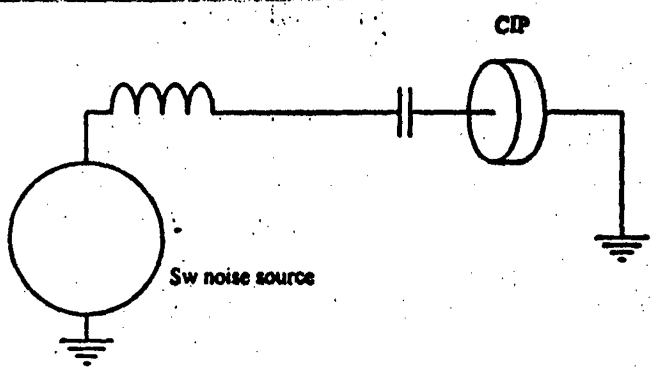


Figure 8.2 Simplified common mode load impedance.

is not excessive so that the switcher can function properly.

8.2 COMMON MODE DESIGN GOALS

The common mode need not meet any of these requirements for the differential mode. The cutoff frequency can be as low as desired, cutting well into power harmonic frequencies. One has to be cautious of the leakage inductance in Zorro inductor(s). This generates a high amount of differential mode inductance and can saturate the inductors at the current peaks. The disadvantage is that the size of the common mode becomes very large as the cutoff frequency is lowered, but there is no lower frequency limit caused by band pass or other factors. These inductors should 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.

8.3 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 Figure 8.1, using the switcher as the common mode noise source, is as follows: G (the switcher and other elements) is now the noise source, and the circuit is drawn backward from the figure. This keeps the format used throughout this book of keeping the source, or line, to the left and the load to the right of each diagram.

In the Figure 8.2 if the circuit lacks a transformer, the diode capacitance to ground has a higher impedance than the current probe until the upper frequencies are reached. The Figure 8.2 shows that the common mode design impedance is higher than the differential mode impedance in most applications. This information allows the following technique. The cutoff frequency is calculated as before, and then the inductor and capacitor values are calculated using the equations:

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

Where R_d is the design impedance and is the same as for the differential mode and F_o is the cutoff frequency.

If there is a leakage current specification, the capacitor value of the capacitor to ground is calculated, or the specification may state the maximum value of capacitance to ground. The needed value of capacitance is divided by the maximum value, and the inductor is multiplied by this value. The cutoff frequency remains the same, but the impedance grows by the multiplier.

CHAPTER IX

FILTER DESIGN TECHNIQUES

To build the filter elements together the element matrices have to be put together. To discuss how to build the matrices for all the filter elements the following list may be visualised. The names are chosen at random and may be changed .

(a) Unit matrix.

(b) R_s matrix.

(c) LINESIM matrix.

(d) LISN matrix.

(e) DIN and DOUT matrices.

(f) RCSHU matrix

(g) LSER matrix and CSHU matrix.

(h) L filter.

(j) π filter.

(k) T filter.

(l) Cauer filter and matrix.

The matrix is a tool whose advantage can be taken to calculate insertion loss such as Ts, πs, Ls, Cauer, and dissipative filters, along with their multiples. The matrix includes the load and sourse impedances, allowing direct calculation of the insertion loss. All unit filters, except dissipative and Cauer filters, can be handled with four elements. These two also have four elements, but each term has two terms one real and one imaginary,for a total of eight. Therefore, all filter equations must be equally complete so that each matrix solution can use the same form for each solution. This is like an overlay, and because each matrix fits the overlay, continuous solutions can be calculated. Any of these units can be described by the matrix equation, and thus these elements can be cascaded or chained to form an entirely new matrix transfer function. The user can place all of the elements as needed in tandem.

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 the 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 electromagnetic interference (EMI) test laboratory after running the proper tests on the unit. In other cases it is only a good educated guess by the equipment or power supply designer using equations given earlier. The filter designers 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 given frequency can be calculated with an estimated load and source impedance. The government normally specifies the insertion loss of a filter at 50Ω load and source impedance in the 220A specification. The designer can insert these values for any load and source needed.

9.1 THE UNIT MATRIX

The simplest matrix is the unit matrix. This is the equivalent of scalar multiplication by 1. The unit requires no input. It is first verified that all the unused elements of the matrix field contain unit elements, not other filter elements, because these other elements in these sections will give erroneous higher losses, making the filter appear to have the required specified losses even though it is well out of the required specification.

The values are

$$\begin{bmatrix} I + J0 & 0 + J0 \\ 0 + J0 & I + J0 \end{bmatrix}$$

In reality, the unit matrix is a square diagonal matrix, as follows :

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

The off-diagonals are all zero, and again, this is like multiplying the matrix field by 1.

9.2 THE R_s MATRIX

The value needed for an R_s filter (Figure 9.1) comes from the "Common Terms" section, and is used for the input-side line impedance R_s. This term is the lower frequency line resistance or source impedance. This is fine for frequencies under 100 kHz, at which a skin effect is not operative and the filter is being designed for insertion losses around 14 kHz. Most of the various groups who have tested power lines have agreed that the line impedances are around 4Ω at 10 kHz. Here, the value of 4Ω is assigned.

$$R_s = 4\pi = \begin{bmatrix} 1 + J0 & 4 + J0 \\ 0 + J0 & 1 + J0 \end{bmatrix}$$

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

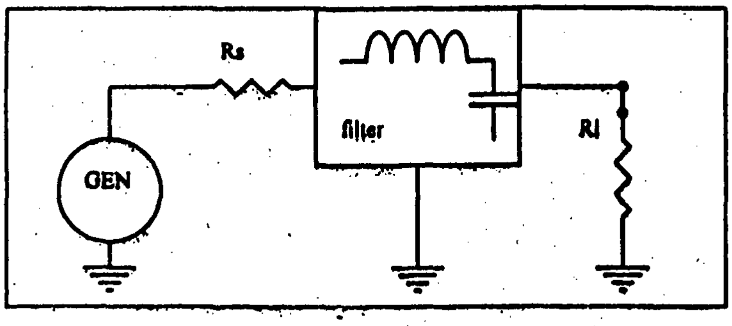


Figure 9.1 The R_s matrix.

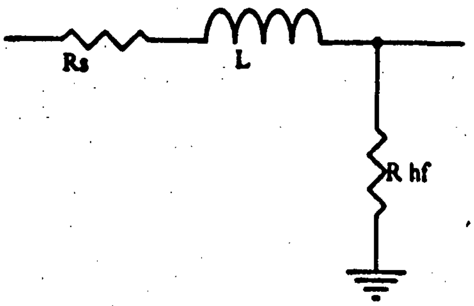


Figure 9.2 The LINESIM network.

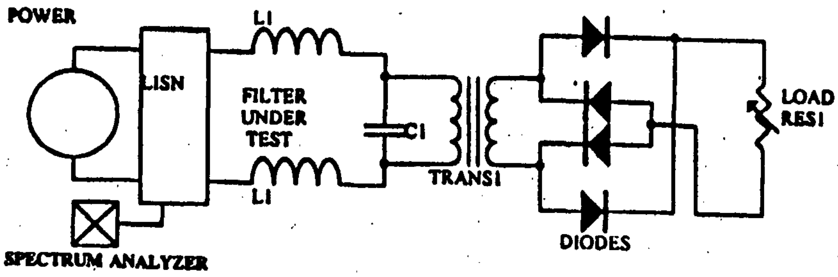


Figure 9.3 The LISN used for testing.

the matrix in the equation):

$$\begin{bmatrix} 1 & 4 \\ 0 & 1 \end{bmatrix}$$

No further information is needed for the unit matrix, which describes the line, not the filter.

9.3 THE LINESIM MATRIX

The LINESIM matrix is used to match a known line or the output impedance of a dc power supply (Figure 9.2). The small series resistor is the value of R_s from the "Common Terms", but the value of the shunt resistor R_{hf} and the series inductor L_1 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_1 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:

$$F = KF_0$$

This makes X_l equal to

$$X_l = 2\pi FL = 2\pi KF_0L$$

R_s, R_{hf}, and L_1 are given so that values can be inserted directly into the matrix, 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.

The LINESIM matrix is

$$\begin{bmatrix} V_i \\ I_i \end{bmatrix} = \begin{bmatrix} \dfrac{R_{nf} + R_s}{R_{nf}} & + \quad \dfrac{JX_l}{R_{nf}} & R_s + JX_l \\ \dfrac{1}{R_{nf}} & + \quad JO & 1 + JO \end{bmatrix} \begin{bmatrix} V_0 \\ I_0 \end{bmatrix}$$

For example, F_0 is 4000, R_s is 4, R_{nf} is 50, L is 90 μH, and floats, so it can be plotted.

X_l = 2.262K, so the matrix becomes

$$\begin{bmatrix} 1.08 + 0.0452JK & 4 + 2.262JK \\ 0.02 + OJ & 1 + 0J \end{bmatrix}$$

When this matrix is multiplied by the others 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.

9.4 THE LISN MATRIX

The input section formed by R_{in} and C_{in} is in series shunt across the line. The mudsection is made up of one inductor in series, L_1. The output section, with R_0 and C_0 in series values are from a standard LISN but can be changed. The default values in any of these sections can always be used. LISN units are used by electromagnetic interference (EMI) test laboratories and are often required for the variation test specification (Figure 9.3). If the test specification calls for an 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.

The LISN matrix is the most difficult to format and is formed by the following multiplication of the three matrices:

$$\begin{bmatrix} 1 & 0 \\ \dfrac{R_1 + JX_{c1}}{R^2_1 + X^2_{c1}} & 1 \end{bmatrix} \begin{bmatrix} 1JX_{l1} \\ 0 \quad 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{R_2 + JX_{c2}}{R^2_2 + X^2_{c2}} & 1 \end{bmatrix}$$

Solving the first two,

$$\begin{bmatrix} 1 & 1JX_{l1} \\ \dfrac{R_1 + JX_{c1}}{R^2_2 + X^2_{c1}} & 1 + \dfrac{JX_{l1}\,(R_1 + JX_{c1})}{R^2_1 + X^2_{c1}} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{R_2 + JX_{c2}}{R^2_2 + X^2_{c2}} & 1 \end{bmatrix}$$

The full solution is

$$\begin{bmatrix} 1 + \dfrac{JX\ R_2 + JX_{c2}}{R^2_2 + X^2_{c2}} & JX_{l1} \\ \dfrac{R_1 + JX_{c1}}{R^2_1 + X^2_{c1}} + \dfrac{R_2 + JX_{c2}}{R^2_2 + X^2_2} + \dfrac{Jx_{l1}\,(R_1 + JX_{c1})\,(R_2 + JX_{c2})}{(R^2_1 + X^2_{c1})\,(R^2_2 + X^2_{c2})} & 1 + \dfrac{JX_{l1}\,(R_1 + JX_{c1})}{R^2_1 + X^2_{c2}} \end{bmatrix}$$

Typical LISN values with F_0 at 4000 are as follows :

Line section			Midsection			Load section		
R_1	=	1Ω	L_1	=	56 μH	R_2	=	50Ω
C_1	=	22.5 μF				C_2	=	22.5 μF

$$X_{c1} = \frac{1}{2\pi K F_0 C_1} = \frac{1.768}{K}$$

$$X_{l1} = 2\pi K F_0 C_1 = K$$

$$X_{c2} = \frac{1}{2\pi K F_0 C_2} = \frac{1.768}{K}$$

Using the standard outline of the 2 x 2 matrix with eight terms, the equations for the individual terms follow.

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

The terms for the martix with the preceding values simplified are as follows :

$$A = 1 - \frac{X_{L1}\ X_{C0}}{R^2_0 + X^2_{C0}} = 1 - \frac{1.407 \times 1.768}{50^2 + 1.768^2/K^2} = 1 - \frac{1.407 \times 1.768K^2}{50^2K^2 + 1.768^2}$$

$$B = J\frac{R_0\ X_{L1}}{R^2_0 + X^2_{c0}} = J\frac{50 \times 1.407K}{50^2 + 1.768^2/K^2} = J\frac{50 \times 1.407K^3}{50^2\ K^2 + (1.768)^2}$$

$$C = 0$$

$$D = JX_{L1} = J1.407K$$

$$E = \frac{R_i\ R_0\ (R_i + R_0) + R_i\ X^2_{C0} + R_0\ X^2_{C0} - X_{li}\ (X_{C0}\ R_0 + R_i\ X_{c0})}{(R^2_i + C^2_{ci})\ (R^2_0 + X^2_{c0})}$$

X_{C0} is often the same as X_{C1}. This reduces the equation to:

$$\frac{(R_1 + R_2)\,|R_1 R_2 + X^2_{c2} - X_{l1} X_{c2}|}{(R^2_1 + X^2_{c2})(R^2_2 + X^2_{c0})}$$

where R_1 and R_{in} are the same, which also follows for C_1 and C_{in}. R_0 (R_{out}) and C_o are equal to R_2 and C_2. These values are interchangeable, as in the LISN shown in Figure 9.4.

$$E = \frac{51K^2\,(47.51K^2 + 1.768^2)}{(K^2 + 1.768^2)\,(50^2 K^2 + 1.768^2)}$$

$$F = J\,\frac{X_{c1}(R^2_0 + X^2_{c0} - X_{Li} X_{c0}) + X_{c0}(R^2 i + X^2_{ci}) + R_i R_0 X_{Li}}{(R^2_i + X^2_{ci})(R^2_0 + X^2_{c0})}$$

X_{c0} is often the same as X_{cin}. This reduces the equation to:

$$F = J\,\frac{X_{c0}\,[R^2_0 + R^2_i + X_{c0}(2X_{c0} - X_{Li})] + R_i R_o X_{Li}}{(R^2_i + X^2_{c0})(R^2_0 + X^2_{c0})}$$

$$F = J\,\frac{[70.35K^5 + 4417.34K^3 + 11.053K]}{(K^2 + 3.126)\,(2500K^2 + 3.126)}$$

$$G = \frac{R^2_i - X_{Li} X_{ci} + X^2_{ci}}{R^2_i + X^2_{ci}}$$

$$G = \frac{3.126 - 1.488K^2}{K^2 + 3.126}$$

$$H = J \frac{R_i X_{Li}}{R^2_i + X^2_{ci}} = J \frac{1.407K^3}{K^2 + 3.126}$$

In a custom made computer program, all these may be already provided but the component values must be entered if they are different. Finally, the component values (A-H) are entered based on the simpler equations:

$$P = 2\pi F_0 K$$

$$A = \frac{1 + P^2 R^2_2 C^2_2 \quad - P^2 L_1 C_2}{1 + P^2 R^2_2 C^2_2}$$

$$J \quad \begin{matrix} P^3 C^2_0 L_1 R_0 \\ 1 + P^2 C^2 R^2 \end{matrix}$$

$$C = 0$$

$$D = JPL$$

$$E = P \frac{\{R_i C^2_i [1 + P^2 C_0 (R^2_0 - L)] + R_0 C^2_0 [1 + P^2 C_i (R^2_i - L_i)]\}}{(1 + P^2 C^2_i R^2_0)(1 + P^2 C^2_0 R^2_i)}$$

$$F = JP \frac{C (1 + P^2C^2_0R^2_0) + C_0 (1_0 + P^2C^2_iR_i) + P^2C_iC_0L_1 (P^2R_1R_0C_iC_0 - 1)}{(1 + P^2C^2_iR^2_0)(1 + P^2C^2_iR^2_i)}$$

$$G = \frac{1 + P^2 C_{in} (C_{in} R^2_{in} - L)}{1 + P^2 C^2_1 R_1}$$

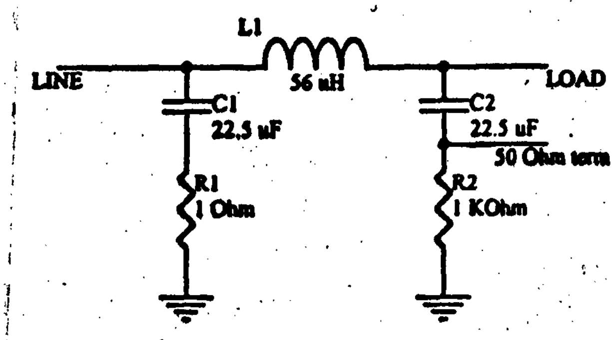


Figure 9.4 The common LISN with equal capacitors.

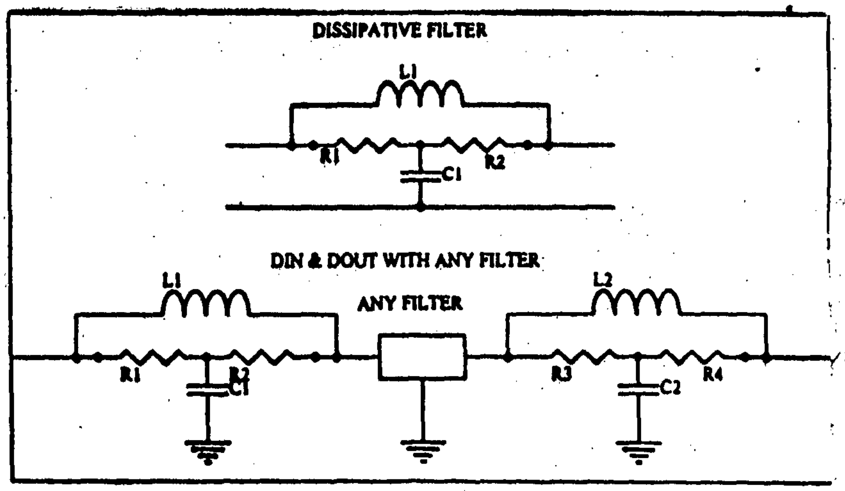


Figure 9.5 The DIN and DOUT.

$$H = J \frac{P^3 C^2_1 R_1 L_1}{(1 + P^2 C^2_1 R_1)}$$

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 normalized frequency, which has the value of 1 at the cutoff frequency.

9.5 THE DIN AND DOUT MATRICES

A dissipative filter gives a loss of 6dB per octave. The DIN faces the line and the DOUT faces the load if both filters are used. 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 Figure 9.5 should be the desired filter input impedance, such as 50Ω for the DIN filter. R_1, the output impedance needed, should match the load for the DOUT, which can also be 50Ω. This would be a nice feature for the 220A spcification. Here, the terms is R_{out} (R_2 in Figure 9.4, bottom) for DIN and R_{in}(R_3 in Figure 9.5) for DOUT, can be used to match any filter wanted - Butterworth, π or other - and these can also be, here, designed for 50Ω.

The Δ is not always equal to 1 because of conversion back and forth between Y and A matrices. This leads to minor errors, or rounding 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_i, and R_0 should be listed with this matrix. Both DIN and DOUT are the same, but R_i and R_0 should switch values. As 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 K_{ratio} is 0.5, or making its cutoff = 8000 Hz and still have the full filter give a good impedance match at 10,000 or 14,000 Hz, say 50Ω, the DIN and DOUT equations are:

$$K_d = KK_{ratio} \qquad R_d = (R_{in}\ R_{out})$$

$$L = \frac{R_d\ K_{ratio}}{2\pi F_0} , \qquad C = \frac{K_{ratio}}{2\pi F_0\ R_d}$$

where K_d is the equivalent of the value of K in the other filters, R_d is the design impedance for the DIN and DOUT only, and F_0 is the cutoff frequency for all the filters, including the DIN and DOUT. The format of the DIN or DOUT matrix is the same as the others :

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

If the design calls for all the R to be 50Ω, K_d is equal to 1, and $K_{ratio} = 1$, the various elements A-H, equal the end values given after each element :

$$A = \frac{4\ K^2_d\ R^2_d\ -\ K^2_d\ R_i\ (R_i + R_0) + (R_1 + R_0)^2}{4\ K^2_d\ R^2_d + (R_i + R_0)^2}$$

$$A = 0.75$$

$$B = \frac{2\ JK^3_d\ R_d\ R_d}{4\ K^2_d\ R^2_d + (R_i + R_0)^2}$$

$$B = J0.25$$

$$C = \frac{K\ R\ (R + R)}{4\ K\ R\ + (R + R)}$$

$$C = 12.5$$

$$D = \frac{J\,(\,2\,K^3_d\,R^3_d + K_d\,R_d\,(\,R_i + R_o\,)^2\,)}{4\,K^2_d\,R^2_d + (\,R_i + R_0\,)^2}$$

$$D = J37.5$$

$$E = \frac{K^2_d\,(\,R_i + R_0\,)}{4\,K^2_d\,R^2_d + (\,R_i + R_0\,)^2}$$

$$E = 0.005$$

$$F = \frac{J(2\,K^3_d\,R^2_d + K_d\,(\,R_i + R_0\,)^2\,)}{R\,(\,4\,K^2_d\,R^2_d + (\,R_i + R_0\,)^2\,)}$$

$$F = J0.015$$

$$G = \frac{4\,K^2_d\,R^2_d\ - K^2_d\,R_0\,(\,R_i + R_0\,) + (\,R_i + R_0\,)}{4\,K^2_d\,R^2_d + (\,R_i + R_o\,)^2}$$

$$G = 0.75$$

$$H = \frac{2\,JK^3_d\,R_d\,R_0}{4\,K^2_d\,R^2_d + (\,R_i + R_0\,)^2}$$

$$H = J0.25$$

Typical values are $F_0 = 4000$, $K_{ratio} = 1$, $R_i = 50$, and $R_0 = 50$. Then, $K_d = K$,

$R_i + R_0 = 100$, and

$$\begin{bmatrix} 0.75 + J0.25 & 12.5 + J37.7 \\ 0.005 + J0.015 & 0.75 + J0.25 \end{bmatrix}$$

as shown by the preceding solutions of A-H. Note that the Δ is equal to 1. These are the values at 4000 Hz.

DIN and DOUT are the same, but R_i and R_o, if the two values are different, exchange values, so that the R_i of DIN faces the line and the Ro of DOUT faces the load. The K_{ratio} is usually 1 but must be the same if both DIN and DOUT are used. This means that two matrices are required, one for DIN and one for DOUT.

9.6 THE RCSHU MATRIX

The RCSHU matrix is used to lower the Q of the filter and correction for resonant rises and/or a problem frequency, such as insufficient loss at a given 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. This filter section can often cure several problems at once. If, for example, a three-stage L is needed to achieve the required insertion loss at 10 kHz, two resonant rise appear in the graph. As the load impedance deviates from the design impedance, the peak value of the resonant rise changes. Also, the loss near 10 kHz may be at, or even somewhat over, the insertion loss limit. The lower resonant rise frequency is determined, and the resultant network is tied 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 should be eliminated. The loss around 10 kHz, assuming that the first

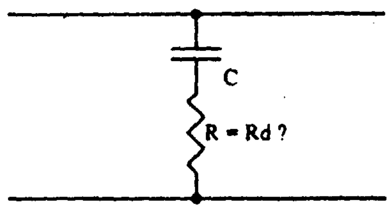


Figure 9.6 The RC shunt.

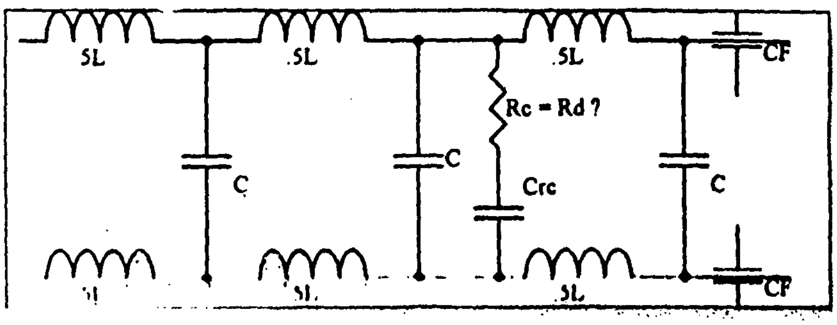


Figure 9.7 The multiple L with RC shunt.

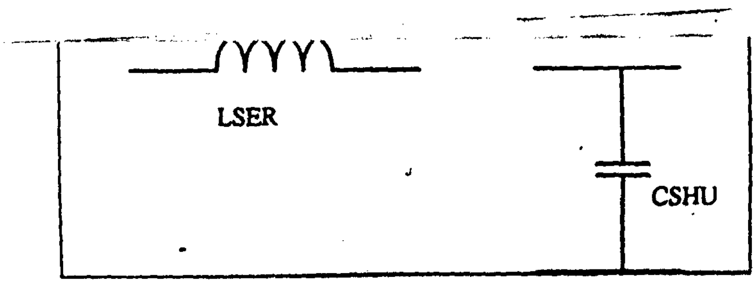


Figure 9.8 The LSER and CSHU.

resonant rise is well below 10 kHz, is improved. Circuit Q is reduced to impede any oscillations by the addition of an RC shunt. (Figure 9.6 and 9.7). This is easier to design than the Cauer and is automatically balanced.

The matrix is simple to from Ri 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. 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 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 the frequency and can be plotted.

$$X_c = \frac{1}{2\pi F_0 KC}$$

$$E = \frac{R_c}{R^2_c + X^2_c} \qquad F = \frac{JX_c}{R^2_c + X^2_c}$$

$$\begin{bmatrix} 1 + J0 & 0 + J0 \\ E + F & 1 + J0 \end{bmatrix}$$

9.7 THE SERIES INDUCTOR LSER AND THE SHUNT CAPACITOR CSHU

There are many uses for each of these matrices, 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 feedthrough capacitor. The LSER, the series inductor, can be a common mode inductor. LSER and CSHU can work together to form part of the common mode filter (Figure 9.8).

The LSER is

$$X_l = 2\pi F_0 KL$$

It is obvious that the value of X_l is a function of K and is placed in the matrix at D

$$\begin{bmatrix} 1 + J0 & 0 + JX_l \\ 0 + J0 & 1 + J0 \end{bmatrix}$$

The CSHU is

$$X_c = \frac{1}{2\pi F_0 KC}$$

This is the same as RCSHU except that Rc is zero, making the E value zero and F the reciprocal of X_c :

$$Xc = \frac{1}{2\pi F_0 KC}$$

9.8 THE L MATRIX

Here we show the origin of the values that make up the L filter, where K is the normalized frequency and Rd is the design impedance (Figure 9.9).

$$F = KF_0$$

$$L = \frac{R_d}{2\pi F_0}$$

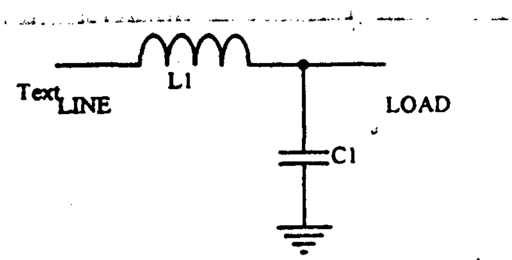


Figure 9·9 The single L filter.

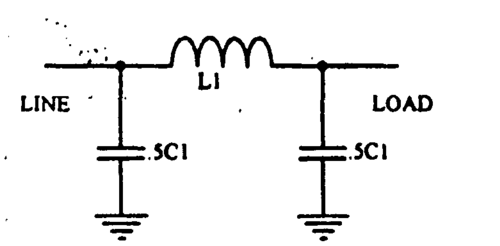


Figure 9·10 The π filter.

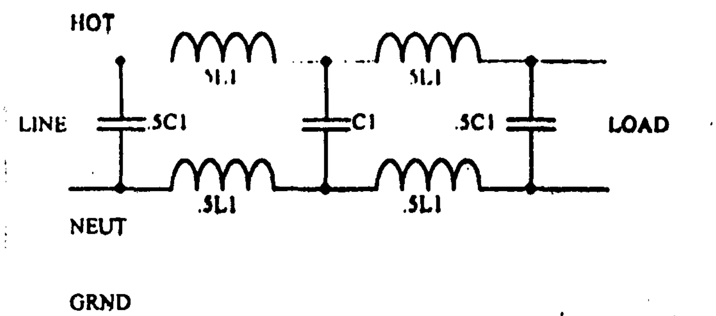


Figure 9·11 The multiple balanced π filter.

$$X_1 = \frac{2\pi F R_d}{2\pi F_0} = K R_d$$

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

$$X_c = \frac{2\pi F_0 R_d}{2\pi F} = \frac{R_d}{K}$$

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 :

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} = \begin{bmatrix} 1 + J0 & 0 + JKR_d \\ 0 + J0 & 1 + J0 \end{bmatrix} \begin{bmatrix} 1 + J0 & 0 + J0 \\ 0 + \dfrac{JK}{R_d} & 1 + J0 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

Eliminating the unnecessary zeros here:

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} \begin{bmatrix} 1 & JKR_d \\ J0 & 1 \end{bmatrix} \begin{bmatrix} 1 & J0 \\ \dfrac{JK}{R} 1 & I + J0 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

The new matrix becomes

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} = \begin{bmatrix} 1 - K^2 & JKR_d \\ \dfrac{JK}{R_d} & 1 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

Then, inserting the other zeros so that the L has the same form as the rest,

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} \begin{bmatrix} 1 - K \quad + J0 & 0 + JKR_d \\ 0 + \dfrac{JK}{R} & 1 + J0 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

9.9 THE π MATRIX

In the π filter (Figure 9.10 and 9.11), 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. The capacitive reactance of both halves in then doubled. Yet, the reciprocal of this impedance placed in term JF or half the value is then used to make up the L matrix in term JF.

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{JK}{2R} & 1 \end{bmatrix} \begin{bmatrix} 1 & JKR_d \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{JK}{2R} & 1 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

The π matrix is

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} = \begin{bmatrix} 1 - \dfrac{K^2}{2} + J0 & 0 + JKR_d \\ 0 + \dfrac{JK\,(4 - K^2\,)}{4R_d} & 1 - \dfrac{K^2}{2} \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

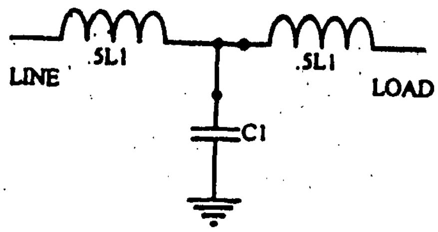


Figure 9·12 The T filter.

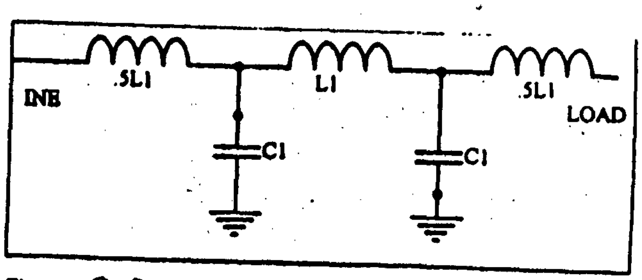


Figure 9·13 The multiple T.

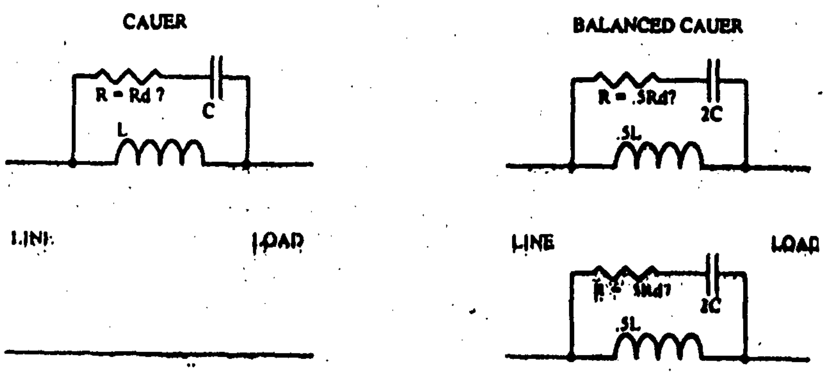


Figure 9·14 The Cauer matrix.

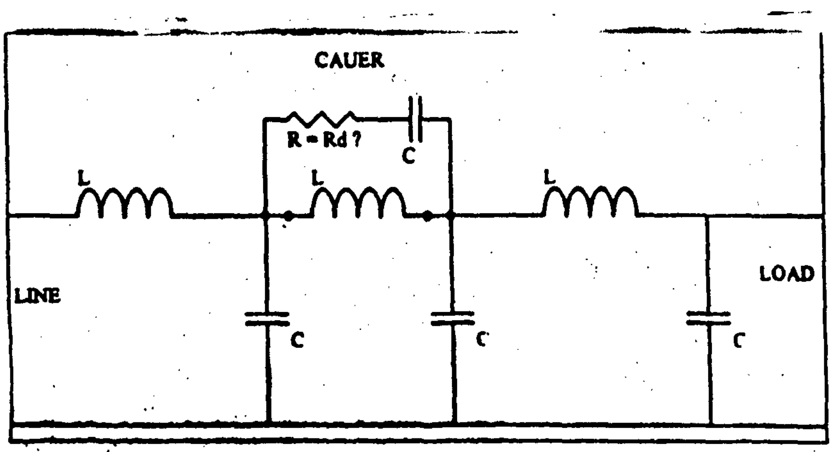


Figure 9·15 The Cauer or elliptical filter.

the same form as the others.

9.10 THE T MATRIX

In the T filter (Figure 9.12 and 9.13), the inductor is split, with half of the inductor in series with line and the other half in series with the load. This makes the JD term half value of the π or L. The full T matrix is:

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} \begin{bmatrix} 1 & \frac{JKR_d}{2} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{JK}{R_d} & 1 \end{bmatrix} \begin{bmatrix} 1 & \frac{JKR_d}{2} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

which becomes

$$\begin{bmatrix} V_{in} \\ I_{in} \end{bmatrix} \begin{bmatrix} 1 - \frac{K^2}{2} + JO & 0 + \frac{JKR_d}{2}\ (2 - \frac{K^2}{2}) \\ 0 + \frac{JK}{R_d} & 1 - \frac{K^2}{2} + JO \end{bmatrix} \begin{bmatrix} V_{out} \\ I_{out} \end{bmatrix}$$

9.11 THE CAUER MATRIX OR ELLIPTICAL FILTER

The Cauer filter (Figure 9.14) is used to eliminate a problem frequency. In the 461 specifications in which 100 dB or so is needed at 10-14 kHz, a Cauer filter is sometimes used. F_m is the problem frequency, and R_c is the resistance in series with the cap. 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 (Figure 9.15). The inductor value is the same as before :

$$L = \frac{R_d}{2\pi F_0}$$

and the Cauer capacitor must resonate with L at the problem frequency F_m.

$$C = \frac{F_0}{2\pi F_m \; R_d}$$

The impedances are

$$X_1 = KR_d \qquad X_c = \frac{2\pi F^2_m \; R_d}{2\pi F_0 \; F} = \frac{M^2 \; R_d}{K}$$

where M is the multiplier to convert from F_0 to F_m :

$$F_m = MF_0$$

$$C = \frac{K^4 \; R_d \; R^2_d}{K^2 \; R^2_c + R^2_d \; (K^2 - M^2)}$$

and

$$D = J \frac{(K^3 \; R^2_c \; R_d - K^3 \; M^2 \; R^3_d + KM^4 \; R^3_d)}{K^2 \; R^2_c + R^2_d \; (K^2 - M^2)}$$

R_c and R_d should but here are listed separately is for any reason they should be different for a special problem an engineer may encounter. If the two terms are equal, C and D reduce to

$$C = \frac{K^4 R_d}{K^2 + (K^2 - M^2)}$$

which approaches R_d at the higher k values, and

$$D = J \frac{(K^3 R_d - K^3 M^2 R_d + KM^4 R_d)}{K^2 R^2_c + R^2_d (K^2 - M^2)}$$

This approaches

$$D = - J \frac{R_d M^2}{K}$$

which is the capacitive reactance, and this approaches zero as the frequency increases.

The matrix for the Cauer filter is

$$\begin{bmatrix} 1 + J0 & C + JD \\ 0 + J0 & 1 + J0 \end{bmatrix}$$

and without the zero is

$$\begin{bmatrix} 1 & C + JD \\ 0 & 1 \end{bmatrix}$$

CHAPTER - X

MATRIX APPLICATIONS

The first step is to find the filter type needed for an application. Either through documentation from the electromagnetic interference (EMI) test house or mathematically the required insertion loss must be known. This is often 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 line frequency must be known to obtain the design impedance R_d and find the minimum cut off frequency. Calculate the value of R_d by dividing the voltae 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. 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 this can be plotted.

10.1 THE SINGLE - PHASE AC FILTER

For example, an L is chosen by the filter designer. The design impedance is found to be 10Ω, 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 4n at 20 KHz. The second matrix is the L matrix chosen by the engineer. If these are automatically multiplied as a system in the computer, all the other matrices in the chain contain unit matrices. 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 40dB loss by varying the value of F_0 or by using a program like GOAL-SEEKER (a shareware program). In the newest

Lotus program, this feature is inbiult. The engineer should add 6 dB for headroom to the loss. This gives the filter designer some latitude on component values without fear of harming the performance of the filter. This also gives some additional attenuation to an unknown frequency spike that could push the equipment when using the filter over the limit. The value of R_s here is 1200 Hz to obtain the desired loss of 46dB (40 + 6 headroom), and this is sitting on the third harmonic, way too low. Another L filter is added. One matrix is skipped by, and it is left as a unit matrix. This allows changing it later to a RCSHU, or whatever if needed. To review:

Matrix 1	:	R
Matrix 2	:	L
Matrix 3	:	unit
Matrix 4	:	change to L
All remaining matrices	:	unit

This is followed by the column matrix with elements V_0 and I_0. 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. The next L filter is added. This changes the cutoff frequency to 400 Hz, which is well above the 400 Hz recommended limit. The initial inductance was 1.3 MH, and now the requirement is two filter at 345 µH. The same is true for the capacitor. This can be balanced as in Figure 10.1. This required an RCSHU to be added to the third matrix; add the common mode components to Figure 10.2. The new matrix layout is as follows :

Matrix 1	:	R
Matrix 2	:	L
Matrix 3	:	now an RCSHU
Matrix 4	:	L
All remaining matrix	:	units

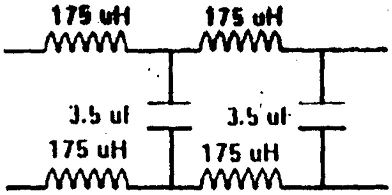


Figure 10·1 The balanced L.

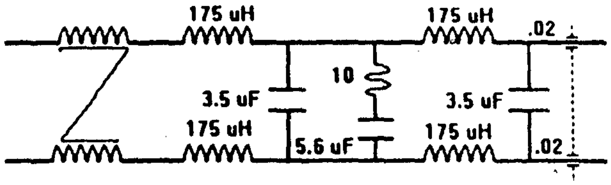


Figure 10·2 The fully balanced double L with Zorro, RCSHU, and feedthroughs.

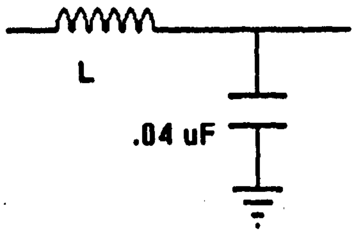


Figure 10·3 The common mode part of the filter in Figure 15.2.

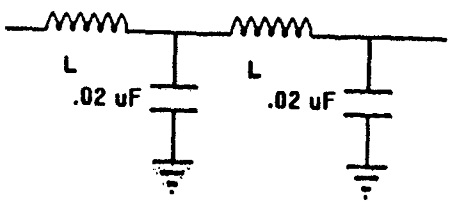


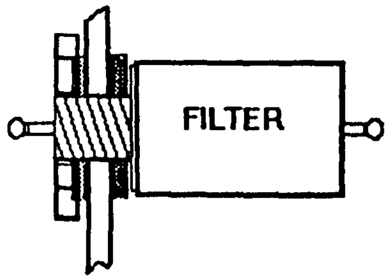
Figure 10·4 The calculation of the double Zorro.

This matrix group is still followed by the column matrix with elements V_0 and V_0/R_d. All dB losses at the other frequencies listed in the specification are checked to meet their requirements for this filter. Most expect all the losses at the higher frequencies pass if the lowest passes. This is usually true but there are exceptions, so that all losses must be checked. In general 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 (Figure 10.3). 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, a total of only 175 μH for the four, adding little to the common mode insertion loss. Yield all this to headroom. This is reduced to a single L.

Removing or replacing these elements with the unit matrix, except the R_s and matrix 2, which is an L. 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 1125Ω and a cutoff frequency of 3550 Hz. The Zorro is 50.4 MH. Zorro should be fixed, some would substitute any reasonable lower value.

All that remains is to design the two different inductors and capacitors or choose a supplier. This filter may be out slightly when tested but should require only minor changes at best. High-frequency problems? Add Capcon or ferrite beads. Differential problems at the switcher? Improve the quality of the RCSHU capacitor, or increase its value and/or raise the inductors to 180 μH, or 200 if needed. Common mode? Increase the Zorro somewhat. Too high a value for the common mode inductor, as earlier? Change the feed through values to 0.01, and add a second section. This is now a double L, as in Figure 10.4. The new value of each Zorro is 7.9 MH. Now, the case could still be presented again by examining the contribution of the differential mode to the common mode filter. One set of parallel differential inductors aids each



The installed feedthrough filter.

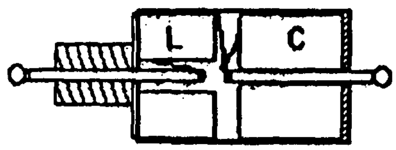


Figure 10.5 The Construction of the single-phase feedthrough.

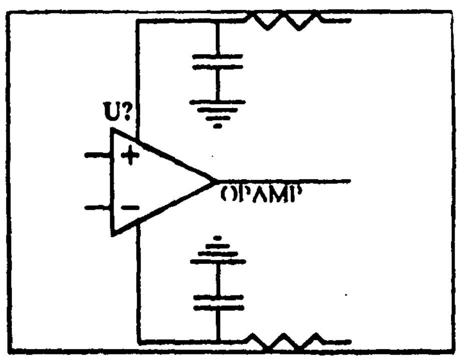


Figure 10.6 The active filter.

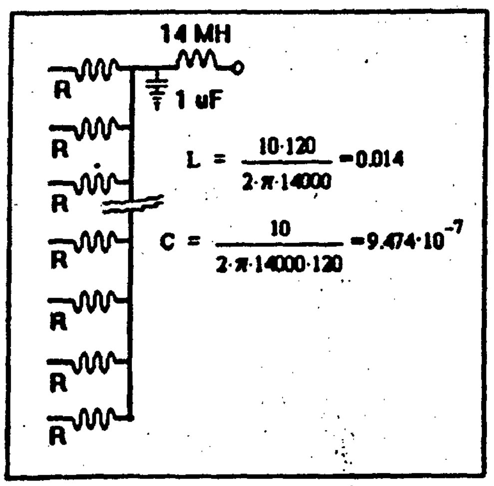


Figure 10.7 The filter required to filter the active filters.

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.

10.2 THE DC-TO-DC FILTER

The dc-to-dc type is often the tubular type and is unbalanced. The filter is often a single feed through capacitor, giving only 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 dB that the L or T would give.

The construction is as follows (Figure 10.5). 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 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.

10.3 LOW-CURRENT FILTERS

The low-current filter suffers from the opposite conditions that affect the high-current filter. Here, the inductances increase while the capacitor size decreases. One method 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 filter (Figure 10.6). They are small and light and can be designed with many poles, but sometime the higher frequencies suffer because of the open-loop gain of the operational amperes. What is often overlooked here is that the direct current feeding the operational amperes must be very clean. Therefore, the filter(s) needs a filter (Figure 10.7).

There have been cases in which a number of high-impedance lines must be filtered. The plus and minus voltage for each operational ampere 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.

$$R \ll \frac{B^+}{10\ I_{max}}$$

If the voltage is 12 V and the maximum current is 10 mA, the maximum value of R is 120 Ω. Making R 100Ω, and since the capacitive reactance value is also 100Ω at half the needed cutoff frequency. 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 operational ampere cutoff frequency, then a 7 kHz, C is equal to 0.22 μF; round this to a standard value. The same is true for the other B, supply. If there are a group of these, say, 10, then the total B' current is 100 mA. This gives an impedance of 120Ω.

The impedance of the inductor should be 10 times the 120Ω just calculated, and the impedance of the capacitor should be one-tenth this impedance of 120Ω. A good quality 1 μF capacitor should remove the noise so that one operational ampere does not add noise to the next. The inductor removes any of these signals from the main supply.

The filter can almost be designed without a computer program. In other words, filter can almost be designed without a calculator if the design is simple. This means that the design should be without RCSHU or Cauer circuits. The goal in the single phase design 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 a sharp stick to write in the wet send.

Frequeny (kHz)	dB
20	46
10	34
5	22
2.5	10
1.25	- 2

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, 22 dB level that a single L does 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. The engineer has many miles of wet sand and starts over a few feet away in clean sand with a double L at 24 dB per octave :

Frequeny (kHz)	dB
20	46
10	22
5	- 2

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 10Ω design impedance must also be known, so the value of the inductors and capacitors can be calculated. For the inductor:

$$\frac{10}{2\pi 5000} = 318\ \mu H$$

and obtains 318 μH. For a balanced filter, this is divided by 2, giving 160 μH not that far away from the 175. Actually, the value of 3 for π would be close enough for rough work. This would have given the value of 333 μH. The capacitor design follows:

$$\frac{1}{2\pi 500{,}010} = 3.18\ \mu F$$

The designer knows that the capacitor value can be solved by dividing the inductance value by the square of the impedance. This is 10Ω, and squared is 100: again, close to the 3.5 used earlier. The designer would round this to 3.2 μF, anyway. The Zorro remains to be calculated, and the designer knows that the feedthrough limit is 0.02 μF, or parallel, giving 0.04 μF to ground :

Frequeny (kHz)	dB
50,000	46
25,000	34
12,500	22
6,250	10
3,125	- 2

This is close to the 3550 value earlier. The zorro inducutor value follows :

$$\frac{10}{2\pi 3125} = 509.3 \ \mu H$$

And the capacitor follows but should be held to two times the maximum to ground, or 2 x 0.02, or 0.04. Again, the value of π, 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 large. So the designer multiplies the inducctor by 127.32 here, and obtains 64.8 MH. This value is somewhat close to the 50.4 MH from the earlier design. This value of inductance is too large. also a conclusion in the previous design. Adding another L filter, makes the common mode a double L. The double L has 24 dB of loss octave :

Frequeny (kHz)	dB
50,000	46
25,000	22
12,500	- 2

The calculation of the inductor follows :

$$\frac{10}{2\pi 12.500} = 127.3 \ \pi H$$

and the capacitor,

$$\frac{1}{2\pi 1,25,010} = 1.273 \ \mu F$$

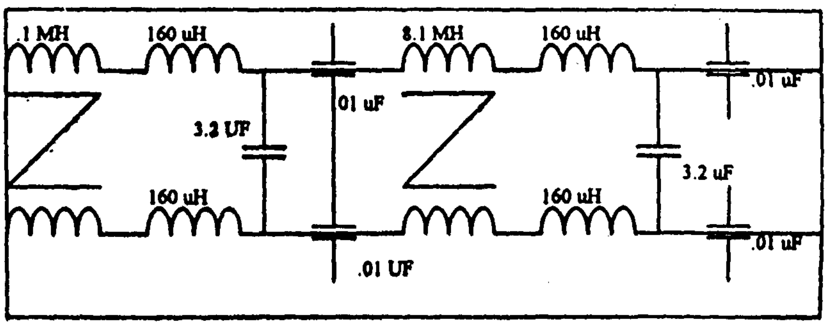


Figure 10-8 The fully designed filter using F_0 the easy way.

which is 63.65 times too large compared with the total of 0.02 μF (four feedthroughs, 0.001 each), so the inductor is multiplied by 63.65, giving 8.1 MH, which is close to the earlier 7.0. Note that the impedance of the common mode filter section is no longer 10Ω.

$$\sqrt{\frac{8100}{0.02}} = 636.396 \quad \text{or} \quad 10 \times 63.65$$

and the cutoff frequency is still 12,500 Hz :

$$\frac{1}{2\pi\sqrt{0.0081 \times 0.02 \times 10}} = 12{,}500 \text{ Hz}$$

The final design is shown in Figure 10.8.

This technique of trading inductance for capacitance can be used in common mode filter design but not is 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 that with the single Zorro. 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 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 that of differential mode impedance. This fights against the common mode Q increase and reduces the opportunities to oscillate. The same technique works on Ts and πs, except that 18 dB is used-replacing 12 dB for the L-for each level as the frequency is divided by 2. Using 30 dB per octave for two or multiple networks, adding 12 dB for each additional network. Three networks and gives 42 dB per octave.

CHAPTER - XI

REVIEW OF FILTER DESIGN

In the last section it is pertinant to elucidate certain details of design and finally give out an overview of design technique. As some ideas that were mentioned earlier but were not clarified.

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. If this is a single-phase 50 Hz line, a minimum of 100 V RMS should do and the SRF line current could be raised by 10%. Attempt to determine the type of load this filter must feed, though : off-line regulator, power factor correction circuit, inductive input filter, or resistive. For a dc system, the load should be a switcher. If this filter is being designed for multiple use or for filter sales, the worst cases should be accounted for within the design.

(a) Find the best filter type by knowing the type or needs of the load. Also, the specification will help here. Knowledge of the specification and losses required will help in assigning a line impedance:

(b) The design impedance from the lowest voltage divided by the highest current is calculated.

(c) Determine the filter cutoff frequency, both differential and common mode. The second method will get the filter designer on the proper ballpark property and the first will get the designer onto the playing field. This can be a computer program written based on earlier discussion. (It also can be purchased through vendors.) For the first method, the actual equations for the cutoff frequency are

$$L = \text{loss for } \pi \text{ or } T$$

$$= (2N + 1)^6.$$

$$L = \text{loss for an } L = 12N$$

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 the equation that follow. N is the number of filters in tandem.

$$\text{Loss required} = \text{dB}$$

$$\text{Loss per octave for filter} = L$$

$$\text{No of octaves} = \frac{X}{L} = \text{dB}$$

$$\text{Frequency at needed loss} = F$$

$$\text{Cutoff frequency} = F_0 = \frac{F}{2^x}$$

$$= \frac{F}{2^{dB/l}}$$

Another way to select the value for the loss L is as follows. Choose the loss based on the filter type. Table 11.1 lists the loss per octave for the filter type listed on the left, the value of loss L in the proceeding equations. Once 10 times the power line frequency, or more, is reached, the filter type and the cutoff frequency are known. For example, for three frequencies provided by an electromagnetic interference test house, the approximate dB losses required to meet the specification and the associated peak frequencies are listed in Table 10.

Table 11.1 F the Easy Way : Loss per Octave Based on

Filter	Number of filter in tandem						
	1	2	3	4	5	6	7
L	12	24	36	48	60	72	84
π	18	30	42	54	66	78	90
T	18	30	42	54	66	78	90

If the power line frequency is 60 Hz, the single L filter will work and the cutoff frequency is 1.4 kHz, but if the power line frequency is 400 Hz, the single L must be replaced by the double L filter at 12.98 kHz. Round either of these to a convenient lower frequency. The filter cutoff frequency is often determined by the lowest problem frequency. All listed problem frequencies must be checked. Do not assume that it is only the lowest frequency listed that will give the proper cutoff frequency.

(d) 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, π, T, and L. The difference is that the value of the inductor 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 last inductors are half values, but the values of the central inductors are the full value. Further the value of the capacitor above, is also divided for the π. Again, the central values for the capacitors for a multiple π will be the whole value and the first and last capacitors will be split.

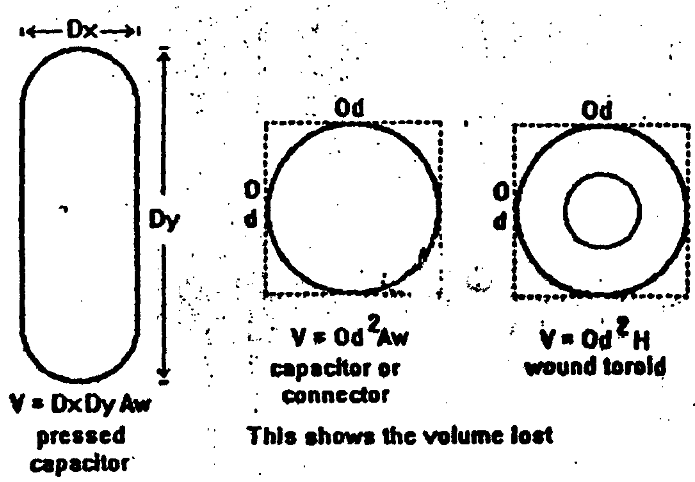


Figure 11-1 Volume of components to calculate total filter volume.

The T and the π filter are exactly opposite each other. Choose the filter layout based on Chapter - IV. Balance the filter if possible.

(e) Find the size of the components, design the components (Figure 11.1). For connector, round capacitors, or any round components, the O_d is the squared times the length. For a pressed capacitor, the volume is D_x D_y $A_{(i)}$. 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 x 3 x 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 x 3 inch direction or the height must change. This example may work, but quadrature is violated and cross talk may be a problem.

Table 11.2 F_0 to Evaluate the Required Filter (Single L and Double L)

Test house trouble frequencies	F1	F2	F3
Frequencies	36 kHz	80 kHz	120 kHz
Measured needed loss	30 dB	52 dB	71 dB
Add 6 dB for head room	36 dB	58 dB	77 dB
F_0 for 50 Hz-okay single L	4.5 kHz	2.8 kHz	1.4 kHz
F_0 - double L required for 400 Hz	17.2 kHz	14.98 kHz	12.98 kHz

(f) Design and build the case, or container, based on the components and volume just determined, and make sure that there are no solder or weld voids, which can allow radiation or destroy the potential of passing environmental tests, such as for humidity and salt spray. Have the case silver plated for better surface conduction. Also, aluminum is often used for better conduction, giving lower radiation. This is plated for even better conduction for lower radiation and for ease of soldering. This container will be silver plated by a military specification plater if the unit is for military uses.

(g) Install the components, and test the filter in the open container. Tack the lid down only for easy alteration or adjustment.

(h) Adjust the filter for the desired loss, if needed, by the following steps : Add lossey components. Add small line-to-line capacitors (X capacitors) in parallel to the existing capacitors. Keep the lead length as short as possible. Add ferrite beads if the current is low enough (typically 5 A 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). Adjustments move the designer from either the ballpark property or the playing field to home plate with lossey components or line-to-line capacitors.

(j) Make sure that the end product is buildable and repeatable for production.

11.1 FILTERS IN TANDEM

Now consider a filter expanding from one section to two sections to three sections, and so on. As mentioned earlier 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 is solved first for the L then the T and followed by the π filter, only to prove that the same holds true for all three types (Tables 11.3, 11.4, and 11.5).

Table 11.3 L Filter Change as Section Are Added

Sl No	F_0	L	C	Extension L	Extension C
1.	405	7.860	19.6	7.860	19.6
2.	2820	1.130	2.82	2.260	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	5.7

Table 11.4 T Filter Change as Section Are Added

Sl No	F_0	L	— C	Extension L	Extension C
1.	940	1.700	8.46	3.40	8.46
2.	3130	0.508	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.	8035	0.198	0.99	2.37	5.94

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 is μH. The first L and C are the individual component values, but the final value for L and C are the total component values for

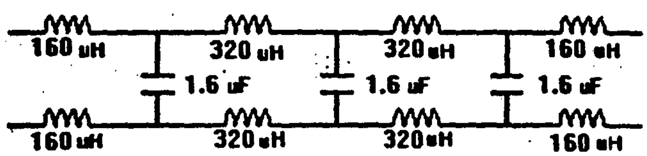


Figure ft·2. The balanced multiple T.

the filter. Note how the total L and C drop in value from one to three sections in Table 11.3, 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 11.4 is the same as the L in Table 11.3. It is ridiculous to go beyond three sections. The 314 μH inductance 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. These are rounded to even values. The balanced circuit is shown in Figure 11.2.

The π filter values of the cutoff frequency F_0 is much lower for the various sections in Table 11.5 compared with the T. Both the T and the π are said to have 18 dB loss per octave, but the T has this loss but it is missing in the π. This is because of the low line impedance at 20 kHz, estimated at 4Ω. If this had been equated with 220A specifications, the cutoff frequencies would have been similar to those for the T. Under these conditions, three sections may be too low, requiring a fourth section. This is caused by the value of F_0.

Table 11.5 π Filter Changes as Section Are Added

Sl No	F_0	L	C	Extension L	Extension C
1.	555	5.735	7.16	5.735	14.32
2.	2520	1.263	1.578	2.526	6.312
3.	4730	0.672	0.841	2.016	5.046
4.	6360	0.500	0.626	2.000	5.008
5.	7450	0.427	0.534	2.135	5.340
6.	8155	0.390	0.488	2.340	5.850

F_0 is cutting into the band pass, which is decided by the tenth 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 Figure 11.1. 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, I would 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 will almost double the inductor SRF, and this should help. Here the concern is for the SRF mainly, adding a section may help. This is because, unless the discussion is very high current filters, the typical value of inductor required compared with the capacitor has an order of magnitude of SRF lower than the capacitor. 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 filtering when the line impedance rises to a reasonable value. An easy way to calculate this improvement in lower total inductance and capacitance is through the equations listed earlier. These are repeated here: they cannot be differentiated because they are not continuous functions.

Loss required = dB

Loss per octave for filter = L

No of octaves = $X = \dfrac{dB}{L}$

Frequency at needed loss $=$ F

Cutoff frequency $= F_0 = \frac{F}{2^x}$

$$F_0 = \frac{F}{2^{dB/l}}$$

As the value of L doubles and triples because of a higher number of filters in tandem, the denominator is reduced and the cutoff frequency rises. The loss L can be calculated from:

L = loss for π or T = $(2N + 1)6$

L = loss for an L = $12N$

The lower value of (loss) L is for the π and T filter; the upper value of (loss) L is for the L filter (Table 11.6).

Table 11.6 Loss per Octave as Tandem Number Increases:

Filter	Number of filter in tandem						
	1	2	3	4	5	6	7
L	12	24	36	48	60	72	84
π	18	30	42	54	66	78	90
T	18	30	42	54	66	78	90

11.2 Q

If the Q is low, enough, the number of resonant rises is 1 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 the Q if 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, than the inductors. For a single L circuit, the equation for Q is

$$L = \frac{R_d}{2\pi F_0} \quad X = \frac{2\pi F R_d}{2\pi F_0} = \frac{F R_d}{F_0} = K R_d$$

$$Q = \frac{X_l}{R_d} = K$$

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, or 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 T and π filters, the equation for the value of L in the loss equation 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.

$$F = \frac{F}{dB/(2N + 1)6}$$

$$L = \frac{R}{2\pi F_0} = \frac{R_d^{2\ dB/(2N\ +\ 1)6}}{2\pi F_t}$$

$$X = \frac{2\pi FR_d\ dB/\ 2^{(2N+1)6}}{2\pi F_t} = \frac{FR_d\ \ 2^{dB/2(2N+1)6}}{F_t}$$

$$Q = \frac{NX_1}{R_d} = \frac{FN2^{dB/(2N+1)6}}{F_t}$$

The terms are restated here. The needed loss at the problem frequency (dB); the tandem number N, the problem frequency F_t, and the Q needed are known. Substitute 1 for Q, and solve for F. This is the approximate frequency, where Q is 1 (because of losses not accounted for, the actual value of F will be higher). Does this frequency fall 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 Xc will equal Rd at this frequency F. If another RC shunt is required for other reasons, only one is needed. Opt for the lower frequency RC shunt. If there are higher frequency Q problems, opt for a higher quality capacitor that will have a higher Q and a higher SRF. Also, in this case, make sure that the resistor use in the RC shunt is noninductive in this case.

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