

本书简介

本书是一本难得一见的工程性书籍，该书作者在书写过程中尽量避免过多涉及复杂的理论推导，而着重与实际的工程设计。因此阅读该书并不需要很高的技术背景。该书对滤波器设计进行了全面的讲解，阅读该书对滤波器设计和解决实际的 EMC 问题（尤其是传导问题）有很大的帮助。

译者介绍

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北京爱科创业电子技术有限公司

爱科公司是一家专业生产电源滤波器的高新技术企业，并通过了 ISO9001 质量体系认证。目前爱科公司的电源滤波器产品涉及单相、三相、民用和军用等各种用途的滤波器 25 个系列。公司有一只以三位电磁兼容专业博士为技术带头人的研发队伍。可以为客户提供通用或按需定制的滤波器产品。通用产品交货期为一周，定制产品交货期为二周。公司的网址：www.emcare.com.cn，联系电话：010-82922019，联系人：蒋小姐。

译者序

在做了一些年 EMC 技术工作后,常常希望能有一些实用的 EMC 方面的技术书籍。但研发工作是如此的繁重,根本就没有精力去对书籍进行广泛的收集。

在一次偶然的机会看到了这本书的 pdf 文件,由于工作的关系阅读得比较深入。这本书虽只是解决了 EMC 领域中一个小的问题,但是毕竟是实实在在的给出了这个问题的基本答案。所以我就决心将这本书翻译出来,献给所有关心滤波问题的技术人员们。由于原文那个扫描的 pdf 文件十分模糊,因此在翻译稿中将英文一并录入。如果中文翻译有什么问题的话,请参考英文部分。

真心希望我们能常常一起讨论和解决 EMC 的问题。如果您有什么 EMC 的问题,我们可以一起讨论解决。我的联系方式是:电话 01082923717, E-mail: yuxp@emcare.com.cn。

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1

Why Call EMI Filter Black Magic?

为什么把 EMI 滤波器称作 Black Magic?

Most engineers, both designers of electromagnetic interference (EMI) devices and others, call EMI black magic. There are three main reasons for this. First, there has not been a well-defined design method. Second, the input and output impedances are not constant over the band of interest. These impedance are usually just good guesses, at best, because the measurements are costly to make and are rarely repeatable. Third, the filter insertion loss test method specifications often confuse or influence the design method. In the 220A specification, requiring 50- Ω source and load impedance, does the filter engineer design the filter for 50 Ω or for real-world impedance? Even if the design passes the 50- Ω tests, what will happen later in the real world? Is it the designer's responsibility to meet the specification or to stop the real-world emissions?

许多 EMI 设备和非 EMI 设备的设计工程师都把 EMI 称作 Black magic，这主要有以下三个原因：第一、目前还没有一种完善定义的设计方法；第二、在我们关系的频段内，输入和输出阻抗并不是常数。另外阻抗测量十分昂贵且测试结果也不易重复，所以通常只好估算这些阻抗；第三、滤波器插入损耗测试规范常常影响甚至扰乱滤波器的设计。例如，在 220A 规范中，要求源和负载的阻抗均为 50 Ω ，那么设计工程师到底是按 50 Ω 阻抗还是按真实的阻抗来设计滤波器呢？设计者的责任是达到测试规范的要求还是消除真实的发射呢？

The design methods outlined in various publications 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.

许多出版物列出的设计方法十分复杂。其中有很多测试既费时又难以定义。另外，在很多时候，还需要增加一些昂贵的辅助设备来获取必要的参数。

§1.1 A Standard Filter Company Contrasted with an EMI Company

§1.1 通用滤波器制造商与 EMI 滤波器制造商的对比

Most energy in the stopband is reflected to its source, a fact often overlooked in both standard filter and EMI technology. The remaining energy is expended in the inductors through the dc resistance of the coil, through the core losses (eddy currents and hysteresis), and as equivalent series resistance in the capacitors. All designers learn this at some point but often forget it somewhere

along the way. The dissipative filter (discussed in Section 5.5) dissipates the energy in the stopband. The designer has these programs in the computer and knows the input and output impedances needed together with the other information required to run the computer program. When the computer design is complete, the designer should be very close to the result required. I am not saying that this is easy to or that the filters never have to be redesigned. The filter may have to be altered by adding stages, or through the use of a different technique, to achieve the goals desired.

在标准滤波器和 EMI 滤波器设计中常常忽视一种实际情况：位于阻带内的能量多数会被反射回源，余下的能量被电感绕线的直流电阻、磁芯损耗（涡流和磁滞损耗）和电容的等效串联电阻所消耗。几乎所有的滤波器设计者都知道这一现象，但是在实际的设计过程中却常常会忘记。而损耗型滤波器（将在 5.5 节讨论）在阻带内是消耗能量的。实际设计中，设计者有这些滤波器的设计程序，只要知道输入、输出阻抗和其它必须的信息就可以运行这些计算程序。在计算机设计完成时，滤波器设计者就会得到很明确的结果。但我的意思并不是说通过计算机设计滤波器很简单，也不是说计算机设计的滤波器不需要再设计。相反，还需要通过增加滤波器的级数或其它的设计方法来达到目标要求。

Most EMI filter manufacturers 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 manufacturer, especially if the EMI designer uses the techniques mentioned in the preface of this book. These techniques would upset most of the standard filter designers. The filter component values are very flexible, so the engineer can use standard values. These filters are adjusted only to meet the insertion loss specifications.

多数 EMI 滤波器制造商只设计和生产具有 EMI 衰减功能的低通滤波器，而很少制造带通或其它类型的常规滤波器。实际上常规滤波器的设计技术与 EMI 滤波器的设计的确有很大的不同，尤其是本书绪言所提到的那些设计技术。相对与常规滤波器设计 EMI 滤波器设计显得更宽松和不精确。所以，这些 EMI 滤波器设计技术会让标准滤波器设计者感到很混乱和不适应。EMI 滤波器的元件参数是十分灵活的，所以设计者通常都可以使用标准（既然可以灵活选择参数，那么使用标准数值的元件会降低生产成本——译者注）的元件参数。另外，这些 EMI 滤波器设计调整的目标仅仅是为达到插入损耗规范的要求。

Often, the languages spoken by the two groups are also different. True filter houses speak of poles, zeros, group delay, predistortion, 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. Such a harmonic content is especially true for power supplied locally such as by shipboard generators. 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.

这两类技术使用的技术术语也常常各不相同。常规滤波器厂使用诸如极点、零点、群时延、预校正、衰减、滤波器阶数等术语，而 EMI 滤波器设计者考虑的是衰减、插入损耗、压降、达到插入损耗所需要的滤波器节数等方面。尽管电源含有谐波成分，但实际上通过滤波器为设备供电的电源是被严格限制在基频的（当然对于供电部分谐波的确存在，例如船发电机会产生谐波）。由于这些电源线谐波只施加到负载上很少的功率，所以 EMI 滤波器设计者并不关心它们。由于没有涉及使用带通滤波器所要求的平坦的频率响应、低相位失真、低峰—峰纹波等性能，结果就使得在 EMI 滤波器设计中很少提及诸如群时延、纹波和相失真等术语。

§1.2 Power Density Spectrum or Envelope

§1.2 功率密度谱或包络

The EMI designer is not blessed with a density spectrum as described above. The designer knows the power-line frequency and its harmonics, which should be passed on to the load. Most are aware that 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.

上面提到的功率密度谱其实增加了 EMI 滤波器设计者的困难。通常 EMI 滤波器设计者是了解电源线基波和谐波的频率的。虽然谐波为负载提供了一小部分功率，但却不应使用 EMI 滤波器来衰减谐波。因为如需衰减谐波则会出现更大的滤波器电容电流以及更大的磁心涡流和磁滞损耗，这都将增大滤波器内产生的热量。因此滤波器必须具有能够通过大功率的基波和一些奇次谐波的性能，这就需要更大的滤波器元件以便于能将能量导通到负载，另外还需要大的线径来导通电流。这就使得我们不得不增加 EMI 滤波器的成本、尺寸和重量。

§1.3 Power Transfer

§1.3 功率传递

The power to be attenuated is minimal compared to the power transmitted if the subject is not on the much larger electromagnetic pulses (EMPs; discussed in Chapter 8). These conditions must also be treated or attenuated. This typical noise could be any, or all, of the following: (1) the switcher frequency plus the odd harmonics, (2) the initial power supply and/or switcher diode noise, and (3) 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.

在没有处理很大的电磁脉冲（EMPs；将在第 8 章讨论）情况下，与传输功率相比，衰减的功率是非常小的。典型的噪声可能是以下提到的部分或全部：(1) 开关频率和奇次谐波；(2) 初级供电电源和/或开关管噪声；(3) 开关电压的电感或变压器的寄生振荡。这种噪声施加

在负载上的能量量级远小于电源供电的能量。但为了消除此噪声,会需要更大的滤波器元件。

The transfer of power creates (1) larger filters with more weight and cost, and (2) much lower self-resonant frequency (SRF) values. The weight, size, and cost of the inductors and capacitors is directly proportional to the power. The SRF value of these filter components is indirectly proportional to the power demanded by the load. This limits the filter's usefulness at high frequencies. The condition will normally dictate that other components be added to filter to compensate for this lack at higher frequencies. This increases the weight, size, and cost of the filter. The assumption here is that the SRF value of the inductors and capacitors is well below the specified upper frequency limit that must be filtered. These additions to the filter could be handled by placing small ceramic or extended foil Mylar capacitors in parallel with the existing capacitors. The second capacitor will raise the SRF value. Capcon and ferrite beads could also be used (see Section 4.4 for a discussion).

功率传递需要 (1) 更重和成本更高的大滤波器; (2) 更低的自谐振频率 (SRF)。电感和电容的重量、尺寸和成本是直接与功率成正比的。虽然滤波器元件的 SRF 值并不与负载需求的功率成正比。但 SRF 本身限制了滤波器的高频性能, 那么为了弥补这个高频缺陷, 需要为滤波器增加其它的元件, 这也就增加了滤波器的重量、尺寸和成本。此处假设: 实际的电感和电容的 SRF 值会明显低于需要滤波的频率上限。通常通过并联小陶瓷电容或延展膜迈拉电容到已有电容上, 可以提高 SRF 值, 当然也可以使用 Capcon 和磁珠 (参见 Section 4.4 的讨论)。

§1.4 Specifications: Real World or Imagined

§1.4 规范: 实际或理想化的情况

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 50Ω. The filter will, in reality, feed a power supply that is rarely close to 50Ω and work into a source of rarely 50Ω. What does the EMI designer do: match the 50Ω or meet the real-world specification? What is meant here by "real-world specification" is similar to the qual(quality) test that may follow. This test measures the noise out of the system through the filter. The line impedance simulation network ration (LISN) is often used as the source impedance for these tests and is closer to real-world requirements.

造成 EMI 被看作 "black magic" 的另一个原因是规范。这些规范与实际相矛盾, 从而误导了 EMI 滤波器的设计。220A 规范要求的是源和负载阻抗为 50Ω 时的插入损耗, 但实际上, 滤波器的源和负载阻抗很少接近 50Ω。那么 EMI 滤波器设计者要怎么做呢: 符合规范 50Ω 的要求还是按“实际规范”设计? 此处的“实际规范”是指以下的测试 (与质量测试相类似)。在测试系统发出的通过滤波器的噪声时, 常用 LISN 作为源的阻抗, 这也与实际情况很接近。

Some specifications call for an LISN value that gives an output impedance of 50Ω starting well above 100kHz. If the only concern of the designer is to match

this 50 Ω , the filter will be matched to this source impedance of 50 Ω . There are two concerns here. The first is What is the lower frequency of loss required? Is it below 100kHz? The LISN output impedance drops rapidly from 50 Ω 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 Ω .

一些规范要求 LISN 的输出阻抗在高于 100kHz 时为 50 Ω 。如果滤波器的设计者只关心此 50 Ω ，那么就意味着滤波器面对的源的阻抗为 50 Ω 。这样在此处就要注意两点：第一、对于低频损耗要求中的低频是多少？是低于 100kHz 么？如果低于 100kHz，LISN 的输出阻抗会从 50 Ω 迅速下降，使得按 50 Ω 设计的滤波器变得不适合。第二、当源和负载阻抗都不接近 50 Ω 时，会怎样。

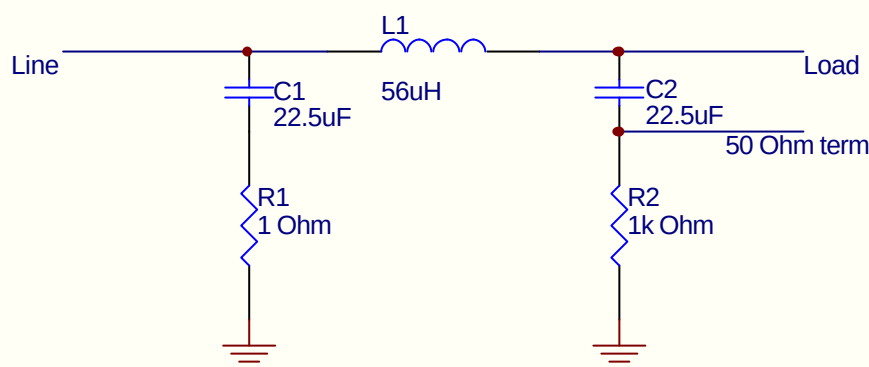


Figure 1.1 Typical LISN.

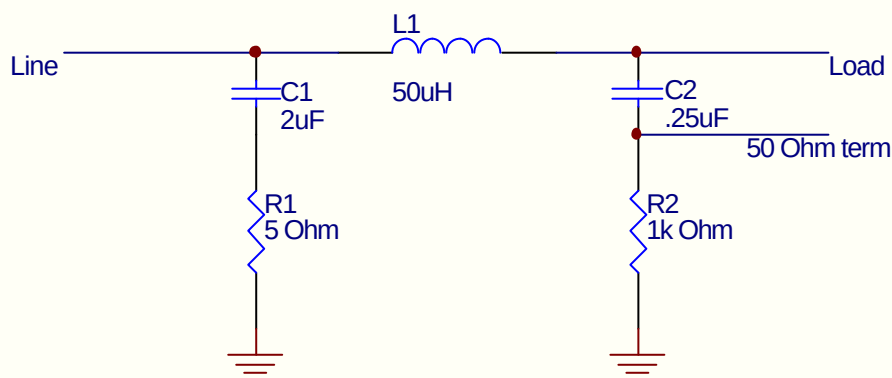


Figure 1.2 Proposed 461 LISN.

Shown in Figure 1.1 is a typical LISN where the line impedance, whatever it is, is shunted by the input network at frequencies above the point where the 22.5- μ F capacitor's impedance is equal to 1 Ω . This is at 7000Hz, so at frequencies above 10,000Hz the line impedance is reduced to the 1- Ω value, while following the same logic, the load side of the LISN looks like 50 Ω , also above 10,000Hz. The impedance of the inductor is low at 10,000Hz (3.5 Ω). At 10 times these frequencies, the inductive reactance is only 35.2 Ω and reaches 50 Ω at 142,000Hz. The entire network at 142kHz generates a source impedance for the filter of 35 Ω at 44°. It obvious that the LISN will not look like

50 Ω until well above 140kHz.

图 1.1 表示的是一个典型 LISN 电路，在 22.5 μ F 电容器阻抗等于 1 Ω 的频率点之上，不管电源线阻抗如何都会被输入网络所旁路。这个频率点是 7000Hz，也就是说，在 10,000Hz 以上的频率电源线阻抗值减小到 1 Ω 。同样，对于 LISN 的负载端，在 10,000Hz 频率以上，看上去为 50 Ω 。在 10,000Hz 的频率下，电感的阻抗为低阻抗 (3.5 Ω)。在此频率的 10 倍频频率下，感抗仅为 35.2 Ω ，并在 142,000Hz 达到 50 Ω 。在 142kHz 整个网络为滤波器提供的源阻抗为 35 Ω ，位相角位 44°。很明显，在没有明显高于 140kHz 的频率时，LISN 的阻抗看上去不是 50 Ω 。

The real intent of the filter is to attenuate conducted emission 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 often pass the insertion loss victoriously and fail when tested along with the system. Often, using a filter that passed the full test as a target or bogie gives disappointing end result. If the filter in question appears to pass, the bogie has often been tagged as bad by the system tester. The 461 specification (Figure 1.6) is more realistic than the 220A, and Robert Hassett, the vice president of engineering at RFI Corp., has given several presentations to the IEEE and other groups that show the advantage of moving away from the 220A test method to the current injection probe (CIP) method. Hassett has run many filters through both methods to show the difference insertion loss curves.

滤波器的真正用途是既衰减电源线发出的又衰减设备发出的共模和差模的传导发射。而（插入损耗的）滤波器的插损测试规范却很少能确保滤波器在实际系统或“实际规范”（50 Ω LISN 的传导发生测试规范）的条件下通过测试。常常会出现这样的情况，滤波器明显通过插入损耗的测试，而对实际系统的测试时则出现失败。实际上 461 规范（见图 1.2—原文为图 1.6，译者注。）比 220A 规范更接近现实。RFI 公司的副总工程师 Robert Hassett 在 IEEE 和其它一些组织的会议上，介绍了以 CIP 方法取代 220A 测试方法的好处。Hassett 用两种插入损耗测试方法测试了许多滤波器，并给出了使用不同方法得到的插入损耗曲线之间的差异。

The curves in Figure 1.3 show the difference using an L filter with the capacitor facing the line CIP as compared to the inductor facing the line CIP (Figure 1.4). The same is also true for the pi filter, due to its input capacitor facing the line CIP. Either filter will look good under 220A tests. This is due to the 50- Ω source impedance. The CIP utilizes a 10- μ F capacitor, and this will reduce the filter loss by 6dB, especially at the lower-frequency end.

图 1.3 给出了电容在电源线的 CIP 端的 L 型滤波器与电感在电源线的 CIP 端的 L 型滤波器（见图 1.4）之间的差异。对于 π 型滤波器，因为输入电容面向电源线 CIP，所以情况也是如此。但在 220A 插入损耗的测试中，二者均表现良好，这是因为 50 Ω 源阻抗的缘故。CIP 使用了一个 10 μ F 的电容器，此电容器将滤波器的插入损耗降低了 6dB，尤其是在更低的频段。

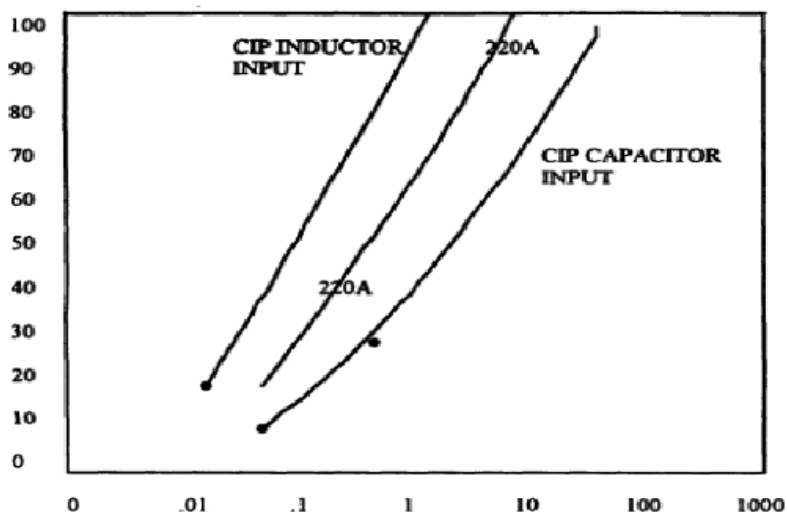


Figure 1.3 L filter with capacitor facing low-impedance source.

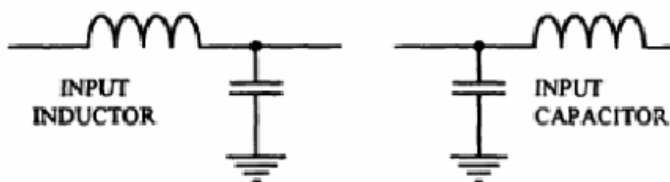


Figure 1.4 L filter with capacitor facing low-impedance source.

In the case of Figure 1.5, the signal generator—now normally from the spectrum analyzer tracking generator—has an output impedance of 50Ω and feeds a coaxial switch (not shown) and the load impedance is the receiver's input impedance (also 50Ω) fed through a second coaxial switch. A calibration path is provided between the two coaxial switches. This test method makes the pi filter function as a true three-pole filter, 18dB per octave, 60dB per decade. The real-world initial losses at the low-frequency end would shunt out the input filter capacitor. This would give initially 12dB per octave of loss until well over 100kHz (depends on the line length). This test method makes this flaw of the pi filter or the L filter with the capacitor facing the low impedance.

在图 1.5 所示的例子中，信号源（现在通常用的是频谱分析仪的振荡发生器）具有 50Ω 的输出阻抗，并馈入同轴切换开关（没有表示出），对于负载阻抗则是接收机的输入阻抗（也是 50Ω）通过另一个同轴切换开关馈入。在两个同轴切换开关之间是一条经过校准的传输路径。这种测试方法使得 π 型滤波器具有三极点滤波器的功能？？每倍频 18dB 的衰减，每 10 倍频 60dB 的衰减。而在按“实际规范”（50Ω LISN 传导发生测试规范）对低频端插入损耗测试时，测试电路将旁路滤波器输入端的电容，这导致在 100kHz（依赖于电源线长）之前，滤波器的插入损耗为 12dB 每倍频。所以用测试方法对 π 型滤波器和电容对应低阻抗输入端口的 L 型滤波器进行插入损耗测试是存在缺陷的。

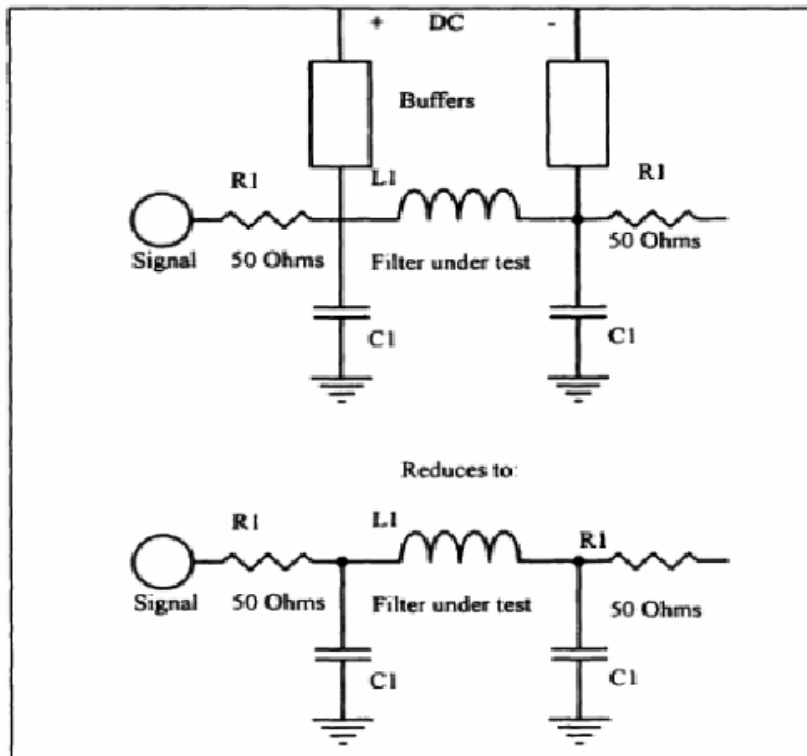


Figure 1.5 220A test setup.

Figure 1.6 shows the method used by Hassett at RFI.

图 1.6 是 RFI 公司的 Hassett 使用的测试方法。

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

在 461 规范中, 使用电流探头馈入测试装置以对比这两个电流。这种测试方法旁路了 π 型滤波器的输入端电容, 在这个 10uF 电容器的 SRF 的频率点之上, 这种测试方法再现了“实际规范”(500 LISN 传导发生测试规范)的情况。这种测试方法测得 π 型滤波器的插入损耗为 12dB/倍频、40dB/十倍频, 而不是原来的 18dB/倍频、60dB/十倍频。

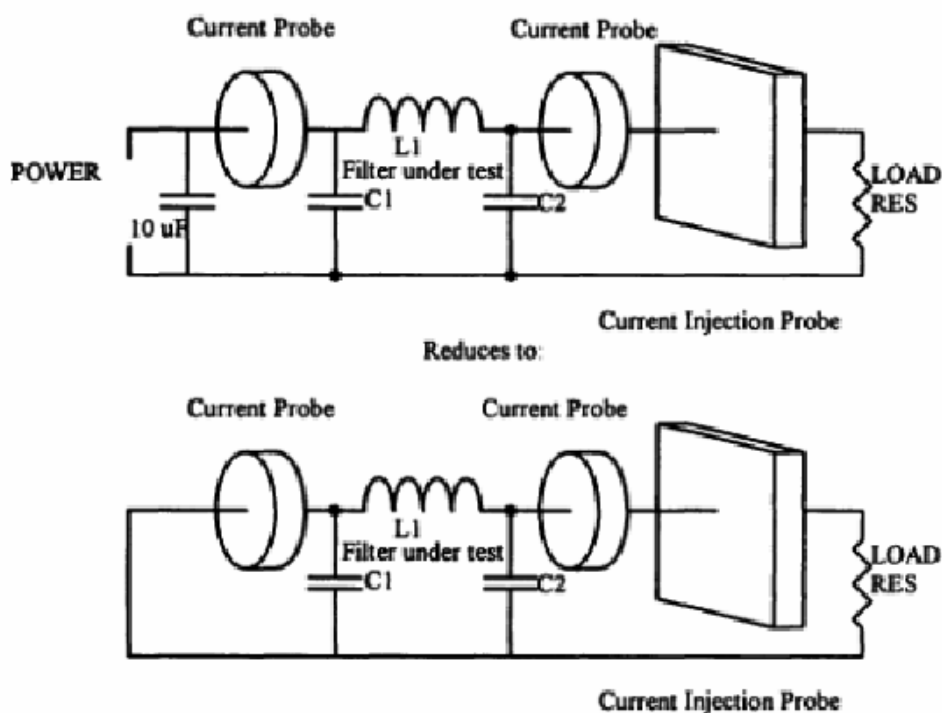


Figure 1.6 CIP test setup.

Other test methods have been suggested. Mitchell Popick, previously with Axel Corp. and a member of the SAE EMI group and dB Bunch, recommends that the load side of the filter face a diode bridge that is properly loaded. This is much better for those that will feed a power supply, which is about 95% of the time. This also shows how the filter handles diode noise. The diode noise is the leading noise competing with switcher noises. These are primarily the odd-order harmonics of the power-line frequency. This works well in three-phase systems as well, but make sure that the total inductive reactance of the filter inductors is much lower than the primary inductance is much lower than the primary inductance of the transformer. This is especially true when multiphase transformer are to be used. These transformers are primarily autotransformers and the primary inductances are much lower than for the isolation transformer type. These two inductances can form a voltage divider reducing the voltage feeding the load. This is another reason to avoid commercial filters, where one filter fits all. These may work fine in some applications and fail in others.

另外其它建议的插入损耗测试方法。如 Axel 公司的 Mitchell Popick (他是 SAE EMI 团体和 dB 组的成员。?? SAE 为美国机动工程师协会, 译者注。) 推荐在滤波器的负载端连接硅桥电路, 这样 95% 的情况下会比负载为另一个电源的效果更好。这种方法同样也表明了滤波器是如何处理二极管噪声的。对于开关电源噪声, 二极管噪声是主要的噪声, 它也是主要的电源基频的奇次谐波源。这种测试方法同样在三相系统中具有良好的效果, 但是需要确保滤波器电感的总感抗远小于变压器的初级线圈电感。在使用多相变压器时同样要确保这一点成立。同样, 那些初级自耦变压器电感和初级线圈电感也要远小于隔离变压器的电感。因为这两种电感 (初级自耦变压器电感和初级线圈电感) 形成的分压器会降低施加于负载的电压。这也是避免使用常规的商用滤波器的另一个原因。通常使用常规的商用滤波器会在一些

设备中工作良好，但在另外一些设备中则完全失效。

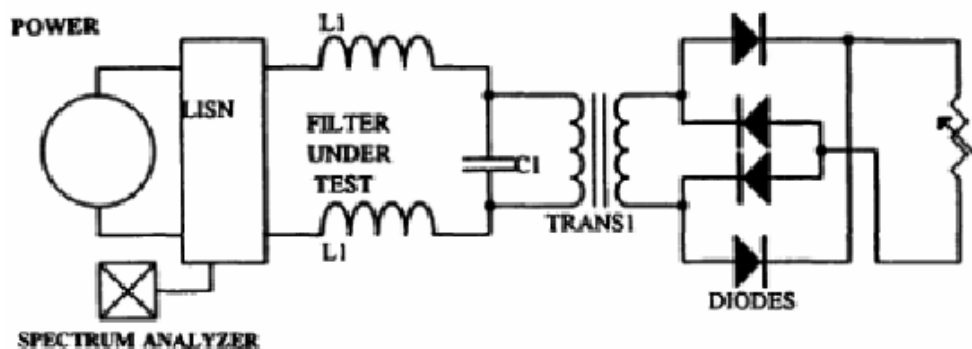


Figure 1.7 Diode test method.

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

图 1.7 中的频谱分析仪可以测试穿过滤波器的二极管、开关和寄生噪声。如果负载是实际的系统而不是图中所画的电阻，那么频谱分析仪测出的是全部信号图像。在一些应用中，常常省略了变压器。负载承载系统承载的全部电流，并在满负载之下变化以便于发现最坏的情况。这比实际规范（500 LISN 传导发生测试规范）更全面，因为在一次测试就需要滤波器处理所有的噪声源。这就是在真实情况下所发生的一切。换句话说，当 CIP 和 220A 方法只是在观测振荡发生器产生的单一频率点时，而在这种测试方法下，却可以完全地检测滤波器。当然，220A 和 CIP 测试是在导通满功率的情况下进行测量的，但上面提到的这种测试方法通样也是如此。

2

Source Impedance of Various Power Lines

不同电源线的源阻抗

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

从事 EMI 领域研究的人常关心的一个主要问题就是源阻抗。目前已经有一些资深科学家和工程师在研究制造商提供的滤波器的转移函数了。他们测量带有负载的电源线阻抗，并试图了解滤波器的转移函数。目前可以通过计算得到转移函数，EMI 滤波器制造商实际上是很少测量它的。如果要求制造商提供转移函数的信息，那么就需要制造商在滤波器和负载之间增加一节测试电路。这样将增加整个系统的共振或其它的不稳定因素。如果保证电源线长、导体间距或导体直径在两次安装（试验和实际安装）时不会有太大的变化，那么滤波器将会工作得很好。当然要获取这个参数是需要宝贵的时间和昂贵的设备租金的。实际上如果把把这个试验过（测定转移参数）的滤波器安装在电缆尺寸与试验相近的这一类船或飞行器上效果就会很好，但是如安装在其它类型的船或其它应用环境中则很有可能会很失败。另外，也需要考虑所有可能出现的负载情况，而不仅仅是峰值负载的情况。如果设备进入待机模式，那么不稳定的情况可能会改变。

The harmonic content of the power-line frequency varies from line to line. In the past, most commercial lines had little harmonic content because of the very low line impedance at these frequencies. This was discussed with a large power

company in 1985, at which time 85% of the power was used for "power and lights," that is, motors and lamps. With the proliferation of computers, copiers, fax machines, and similar equipment now tied to these lines-and very few with power factor correction circuits-the voltage waveform is less sinusoidal. This is due to the voltage drop caused by the high current spikes that these machines demand. The power factor correction capacitors placed across the line by the power company do help to reduce the current gulp, but the voltage is still less sinusoidal. The harmonic content of generators at remote sites and shipboard installations is much greater because of the higher resistance of the generator and lines. The voltage supplied to the end users in these applications is less sinusoidal.

电源线基频的谐波量也会随线的变化而变化。在过去,许多商用线缆因为在这些谐波频率点上具有很低的阻抗,所以谐波量很小。在 1985 年,某个大电力公司讨论过这个问题,在 85% 的情况下电力被用来做“动力和照明”使用,也就是说用在马达和电灯上。但是随着计算机、复印机、传真机和其它类似设备接入电网的数目增加(当然此类设备很少有功率因数校正电路),电源线上的电压波形就很少是正弦波了。这是由于这些设备所需要的电流尖峰导致的电压降所造成的。电力公司在电源线间放置功率因数校正电容器确实会减少电流影响,但是电压仍不是正弦的。这是因为,由于发电机和电源线的高阻抗,处于远端的发动机和船载装置还是有很大的谐波量,因此这些设备传输给终端用户的电压仍不是正弦的。

§2.1 Skin Effect

§2.1 集肤效应

The skin effect can take its toll on the higher-frequency energy on the power lines. This helps dissipate electromotive power (EMP) and other noises traveling on the power line in either direction. These power lines were constructed to handle power at very low line frequencies, not for the higher frequencies creating the loss. Although the characteristic impedance of the line may be 150Ω, the loss of the line per unit length increases with frequency. For copper the equation for the skin effect depth in centimeters is

高频能量在电源线上传输时,集肤效应会很明显。这将增加电动功率(EMP?? 此处的 EMP 不同于 Electromagnetic Pulse 的 EMP 缩写,译者注。)的消耗,并出现其它噪声在电源线任意方向上传播的不良情况。这些电源线的结构设计是为了处理低频功率传输,而不是处理高频功率的。尽管电源线的特性阻抗可能是 150Ω,但是单位长度的电源线的损耗却是随频率增高而增加的。对于铜材,集肤深度用厘米的单位可以表示为下式

$$D = \text{depth (cm)} = \frac{6.61}{\sqrt{F}} \quad (2.1)$$

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

在集肤深度的频率之上,线的导电横截面积(CA)是

$$CA = pR^2 - (R - D)^2 p = D(2R - D^2) p \quad (2.2)$$

(原文公式出现错误—译者注)

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

此处 R 是以厘米为单位的线的半径, D 是前面计算的集肤深度, 单位也是厘米。随着频率增加, D 不断减小, 因此 D 项远小于 $2R$, 那么表达式 (在上限频率点上) 可以简化成

$$CA = 2pRD \quad (2.3)$$

The original cross section, pR^2 , compared to the value of CA , times the original dc resistance, will give an approximate value of the ac resistance at these upper frequencies due to the skin effect.

源于集肤效应, 把电源线原来的截面积 pR^2 除以 CA 并乘以原来的直流阻抗将会得到在这些上限频率上的近似交流阻抗。

$$R_{ac} = \frac{pR^2 R_{dc}}{2pRD} = \frac{RR_{dc}}{2D} \quad (2.4)$$

The lowest frequency would be well above the value of
最低的频率会大于下式的值

$$F = \frac{6.61^2}{D^2} \quad (2.5)$$

Below this frequency, the skin effect is the full radius of the wire. The frequencies we are discussing here should be several times this low frequency. Replacing D with equation (2.5) gives

低于此频率, 集肤深度就是整个线径了。所以我们讨论的频率应该几倍于此最低频率。用 (2.5) 式代替 D 则

$$R_{ac} = \frac{RR_{dc}\sqrt{F}}{13.22} \quad (2.6)$$

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

如果 R_{dc} 是以欧姆为单位的线元阻抗, 那么 R_{ac} 就是这小部分的近似交流阻抗。每米长度的 R_{dc} 、 L 、 C 和 G (线间的电导率) 共同构成了此电源线的一小节, 此时电源线的特性阻抗是

$$Z_0 = \sqrt{\frac{R_{ac} + wL}{G + wC}} \quad (2.7)$$

Substituting (2.6) into (2.7) gives

将 (2.6) 带入 (2.7) 得到

$$Z_o = \sqrt{\frac{0.07564RR_{ac}\sqrt{F} + 2pFL}{G + 2pFC}} \quad (2.8)$$

Simplifying yields

简化后为

$$Z_o = \sqrt{\frac{0.07564RR_{dc}\sqrt{F} + 2pFL}{G + 2pFC}}$$

deleting G because it is much smaller than

因为在本节讨论的频率下 G 远小于 2pFC, 所以忽略 G

$$G \ll 2pFC$$

at the frequencies discussed in this section. The square root of F is removed in the first term in the numerator:

可以化简掉分子式的第一项中的 $\sqrt{F}$, 则

$$Z_o = \sqrt{\frac{0.07564RR_{dc} + 2pL\sqrt{F}}{2pC\sqrt{F}}} \quad (2.9)$$

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

为考察频率对分子、分母的限制, 现在分别对分子、分母求微分

$$\sqrt{\frac{0 + 2pL/2\sqrt{F}}{2pC/2\sqrt{F}}} = \sqrt{\frac{2pL}{2pC}} = \sqrt{\frac{L}{C}}$$

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

分子的第一项为零, 而分子的第二项中的含 F 项与分母抵消

$$Z_o = \sqrt{\frac{L}{C}} \quad (2.10)$$

Therefore, the skin effect term has little or no effect on the characteristic impedance. The dominant terms are the values of L and C. This is the fundamental equation of the characteristic impedance of cables. This all shows that the normal characteristic impedance equation still dominates at the higher

frequencies and that this characteristic impedance does not change with the skin effect, although the loss per unit length does.

因此，集肤效应对特性阻抗影响很小或没有影响，而起主要作用的则是 L 和 C 的值。此式就是线缆特性阻抗的基本公式。通过以上的推导可以看到，即使在高频下通常的特性阻抗仍然是占主导地位的，尽管单位长度的线缆损耗在增加，但是集肤效应还是没有改变特性阻抗。

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

除专为高频设计的同轴电缆外，同轴电缆在高频的表现都类似于上述的情况。实际上有很多不同的电缆都具有相同的特性阻抗。这些线缆的内外径都不相同，一些同轴线具有很小的直径而另外的一些却直径很大，它们之间的每百英尺的 dB 损耗各不相同，并且随频率发生变化。这一点有点像衰减器，阻抗同为 50 Ω 的衰减器，其损耗却是各不相同的。主要的不同点是电源线阻抗是随频率变化的，正是这个阻抗消耗了不需要的能量，而不是传输线特性阻抗消耗的。

§2.2 Applying Transmission-Line Concepts and Impedances

§2.2 传输线概念及其阻抗的应用

Transmission-line concepts (Figure 2.1) may sound like a strange subject to introduce when discussing EMI filters, but the filter designer needs to have a basic understanding of these for several reasons.

在讨论 EMI 滤波器时，提及传输线概念像是一个很奇怪的话题，但是有几个原因使得滤波器设计者应对传输线有基本的了解。

One is to understand why a high impedance is needed at the filter input end for EMP allocations, and a second is to understand that the power-line energy loss is not because of the characteristic impedance of the line but due to the resistance elements along the line. These losses are due to elements such as skin effect, the dc resistance of the line, and the conductance across the line. The third reason is that especially in multiple-element filters, the filter has characteristic similar to those of transmission lines.

其一、可以理解为什么在电动功率分配应用中使用的滤波器其输入端口需要是高阻抗 第二、可以了解导致电源线功率损耗的是线上的阻抗元，而不是线的特性阻抗。功率损耗是电源线的集肤效应、直流电阻和线间电导等因素造成的。第三、尤其是多单元滤波器，其特性与传输线相类似。

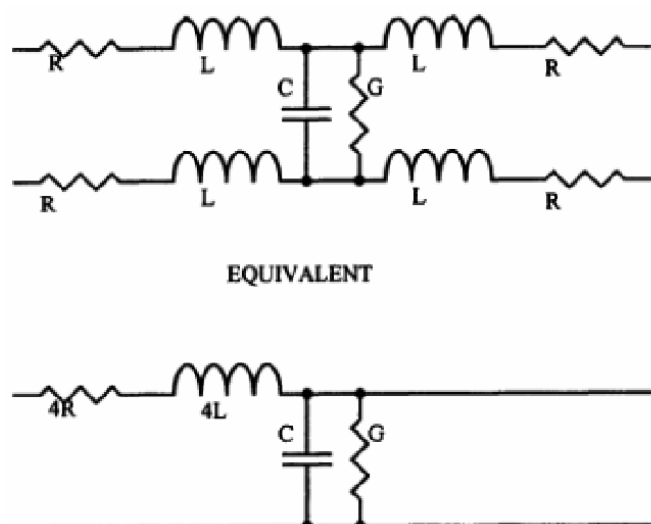


Figure 2.1 Transmission line and equivalent

The shorter power lines start taking on their varying characteristic impedance at much higher frequencies. The characteristic impedance of these power lines varies with frequency and is not as constant as that of coax, twin lead, and twisted pair. The characteristic impedance of the open wire type normally varies typically between 90 and 180Ω, due to the spacing between the conductors and the wire diameter. The paired type, or twisted wires, enclosed in conduit is 50 to 90Ω, for the same reason as above, with the added capacitance between the wires and the conduit. In about 95% of the cases, several different power-line sections will be in tandem. These different power-line impedances will vie back and forth as the different power sections approach resonant lengths. The velocity of propagation is very slow, due to being constructed to carry only the power-line frequencies, not radio frequency. These lines look like they are about eight times their actual lengths. If the power line is struck with an energy pulse, the power line will dissipate some of this energy in the resistive elements. The pulse will travel toward the filter end at the lower velocity of propagation of these cables. The fundamental frequency is around 50kHz and soon assumes the characteristic impedance of the power-line cable rather than the free-space impedance of 377Ω. If the filter input impedance is high or looks like an open impedance to this pulse, the voltage soars (could double), aiding the arrester to function more quickly. If the filter impedance is low or a short to this pulse, the voltage drops to zero and the line current doubles. This slows or keeps the varistor from functioning. This is discussed in Chapter 8.

短电源线在极高的频率下其特性阻抗开始出现变化。实际上电源线的特性阻抗是随频率发生变化的，而不像同轴线、双引线和双绞线等线缆的特性阻抗是常数。随着导线间距和线径的变化，裸线型电源线的典型的特性阻抗在 90~180Ω 之间。基于同样原因，双绞线型、绞线型和带套管型线缆因为线间电容和线套管间电容的存在，导致其特性阻抗在 50~90Ω 之间。在 95% 的情况下，电源线是由不同的电源线类型一前一后构成的。这两个不同的电源线阻抗将前后竞争，形成一个近似的谐振长度。另外，因为电源线的设计结构仅仅是为了传输电源频率而不是射频，所以传播速度是非常缓慢的。因此这些电源线看上去有实际长度的

8 倍长。如果电源线受到脉冲的冲击，那么电源线的阻抗元将消耗其一部分能量。脉冲将以很低的线缆传播速率向滤波器端传播。其基本的频率大约是 50kHz，假设此时电源线特性阻抗大于自由空间阻抗 377 Ω ，那么如果滤波器的输入阻抗对于脉冲是高阻抗或是开路，则电压会剧增（可以加倍），这种情况下需要增加快速动作的过压保险丝。如果滤波器的阻抗对此脉冲是低阻抗或短路，那么电压会下降到零，而线电流则会加倍，这种情况下可以使用慢速动作或持续性能的变阻器。关于这一点将在第 8 章讨论。

I have seen graphs from the IEEE showing some line impedance spikes of almost 500 Ω at one frequency in the megahertz range. Figure 2.2 shows the typical impedance range that the various types of power lines fall between for various twisted-pair cables. The graph is like an average of impedances, and the variations fall primarily between the two curves shown. A very minor percentage of impedances were above and below the curve.

我曾在 IEEE 刊物上见过在兆赫兹级别的频率下一些线缆的阻抗会急升到 500 Ω 。图 2.2 给出了不同类型电源线的阻抗范围的坡度曲线，它处在不同双绞线电缆阻抗的上下限之间。这幅图是阻抗的平均值，不同坡度的曲线处在图中的两曲线间。很少有电源线的阻抗会高于或低于此曲线。

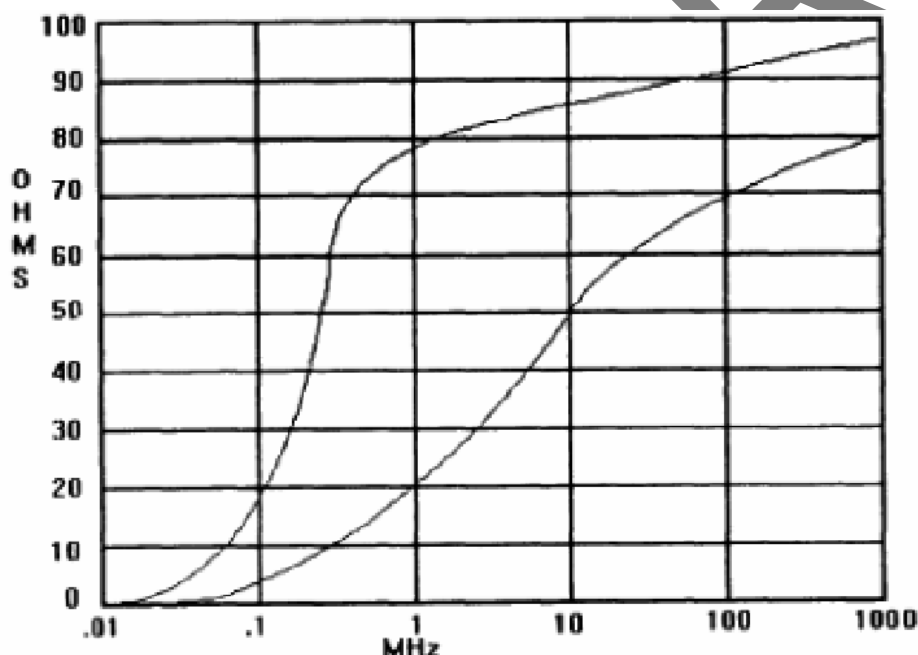


Figure 2.2 Impedance range of a short line: twisted pair

The major part of these losses is due in part to the skin effect and dc resistance, especially at the lower frequencies, for both the long and short lines appear to form link coupling. The point here is that at frequencies below 10kHz, both the longer and shorter lines look resistive and close to the dc resistance value. This value is near 0 Ω for the better commercial power companies, but for most remote power systems, this resistance can be much higher. As the frequency rises, these lines have an impedance near 4 Ω around 10kHz. The impedance then ripples its way to 50 Ω near 100kHz for the longer lines and 250kHz for the shorter lines. It should be obvious that the losses required for MIL STD 461

specification is that the filter must be designed to have the proper losses with the line impedance in the very low or 0-0 region. This should make it clear why the pi filter has trouble meeting the loss for the “real world” and 461 requirements. The same is true using a 10-uF capacitor across the power line. The losses that the filter must meet are at frequencies in the region 10 to 14kHz in the 461 specification and may be as much as 100dB for power-line filters. This is the type of filter used in secure room applications.

尤其是在低频，由于长线和短线呈现链耦合状态，能量损耗主要是由集肤效应和直流阻抗等因素造成的。在低于 10kHz 的频率下，长短线都呈现阻性，其值接近直流阻抗值。对于较好的电力公司，这个线路阻抗值接近 0Ω，但是对于极远端的电力系统，此阻抗值可能会很高。当频率上升时，在 10kHz 线路阻抗会接近 4Ω。对于长线来讲，在接近 100kHz（对于短线则是 250kHz）的频率下，阻抗值会以波动的曲线接近 50Ω。MIL STD 461 规范对损耗的要求是滤波器在电源线阻抗非常低或为 0Ω 时应该具有恰当的损耗，所以 π 型滤波器在满足“实际规范”（50Ω LISN 传导发射测试规范）和 461 的损耗要求上会有困难。对于在电源线间并一个 10μF 电容器的情况也会如此。在 461 规范中，滤波器必须满足规范要求的损耗是在 10~14kHz 的频率范围，或许对于电源线滤波器损耗要求甚至有 100dB。当然这是一种应用在安全室的滤波器。

§2.3 Applying Transmission-Line Impedances to Differential and Common Modes

§2.3 在共模和差模中应用传输线阻抗

The transmission-line losses for the differential mode (Figure 2.3) are all the resistive elements of the line, including the reciprocal of G , the conductance across the line. These losses include the skin effect and some very minor losses due to the equivalent series resistance (ESR) of the capacitor. The dc resistance (DCR) of the inductor adds no loss because the DCR is included in the wire already—it is the wire. Any differential mode pulse, motor, or inductive load switch propagating down the line toward the line filter will be attenuated by the impedances named all along the line in each section.

对于差模来讲（如图 2.3），传输线损耗都是包括线间电导（ G 的倒数）在内的阻性单元所造成的。这些损耗包括集肤效应和电容的等效串联电阻（ESR）等因素所造成的很小一部分损耗。电感器的 DC 电阻（DCR）没有使损耗增加，这是因为 DCR 已经计入整个线缆了（电感线圈也是线缆的一部分）。对于任何沿线向电源线滤波器传播的差模噪声（马达或感性负载切换），将受到沿线（各个小单元）的阻抗的衰减。

The difference in the requirement for common-mode (Figure 2.4) and differential-mode attenuation is that G and ESR value of the capacitor have no effect. A pulse traveling down the line, lightning, or EMP, toward the filter is consumed by the resistance of the line and the line's skin effect. Again, the DCR of the inductor(s) is included with the wire losses (this is the wire). The characteristic impedance also changes because the two lines are in parallel, so the capacitor across the line is out and so is G . The inductances are in parallel, about 1.5μH per meter on each line, or 0.75μH, and the capacitance to ground determines the characteristic impedance. This makes the characteristic

impedance quite high.

共模衰减 (见图 2.4) 与差模衰减的不同点是 G 和电容的 ESR 对衰减不起作用。对于共模, 沿线向滤波器传播的 (照明或电动功率) 脉冲会被线阻抗和线集肤效应所损耗。同样, 电感器的 DCR 也包含在线路损耗内了。由于两线并行, 所以线路特性阻抗也发生了变化。线间电容器和 G 都要忽略, 并联线路的电感是每条线 $1.5\mu\text{H}/\text{m}$, 也就是总计 $0.75\mu\text{H}/\text{m}$ 。此时对地电容决定了线路的特性阻抗, 这使得线路的特性阻抗十分高。

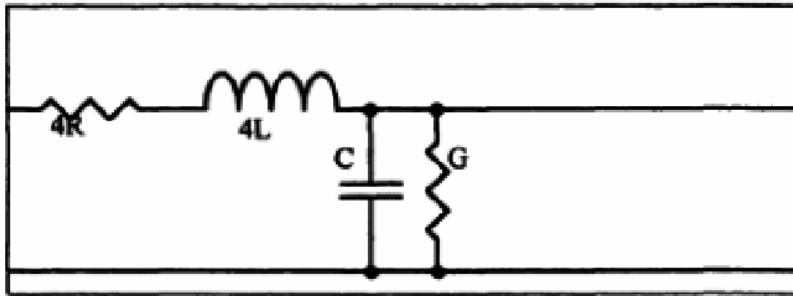


Figure 2.3 Equivalent differential-mode transmission line.

§2.4 Differences among Power-Line Measurements

§2.4 不同电源线测量法的差别

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

许多研究组织都测量了电源线阻抗, 但是通过对比发现他们的结果很少一致, 原因是不同研究组织测量的线的类型差别很大。在所有这些线上波的传播速度都十分缓慢, 这使得这些线看起来比实际的长度要长, 尤其对于一些发动机或变压器等设备端接的线, 其看上去的长度几乎是他们实际长度的 8 倍。这是因为这些设备的线间电容在高频时将传输线路旁路。当然, 这些线的设计是为了传输 50、60 或 400Hz 电功率的, 而不是针对高频传输用的。

§2.5 Simple Methods of Measuring AC and DC Power Lines

§2.5 测试 AC 和 DC 电源线的简单方法

Power-line impedance can be measured, with caution, if the designer wishes to. There should be no need to do this, especially when using the techniques used in this book. This measurement can be accomplished with inexpensive test equipment on both ac and dc lines but would be much faster and more accurate using a network analyzer. Either line, ac or dc, should be resistive loaded, and several runs will be needed. The first reading will be maximum loading, followed by medium loading and then by low loading.

如果设计者试图测量电源线的阻抗, 是可以做到的, 但是这样做却是没有必要的, 尤其是在使用本书的设计技术时。对于 AC 和 DC 电源线都可以使用廉价的测试设备来完成电源线的阻抗测试, 但是如果使用网络分析仪会更快也更精确。无论是 AC 还是 DC 电源线, 都需要

加载阻性负载，还需要依次进行如下几个测试步骤：先满载读数，再半载读数，最后低负载读数。

To measure the ac line, start well above the line frequency by at least a factor of 10 to keep of as much of the ac harmonic power from the readings as possible. This is done to assure accurate readings and to protect the measuring equipment. The frequencies required will be over a wider frequency range, well into the megahertz area. A frequency-selective level meter (FSLM) should be used along with a signal generator and a blocking capacitor. The capacitor should be a high impedance to the line frequency and low to the frequencies to be measured. If the line frequency is 400Hz, the lowest intended frequency to be measured should be well above 4000Hz. If the impedance of the capacitor is 10000 at 400Hz, the impedance would be 1000 at the lowest frequency to be measured. The FSLM should be used in its narrowest input filter selection. The generator frequency should still not be tuned to any multiple, especially the odd harmonics, of the line frequency until 20 times the line frequency is reached. The loss across the capacitor almost cancels except at the lowest frequencies, anyway, so that the 400-Hz loss could be still higher. The only component remaining is a resistor of some value, say 4000, and this should be noninductive to well above the highest-frequency reading. The resistor, capacitor, and signal generator-in that order-are tied across the line with the resistor at the hot end and the signal generator to the neutral. The signal generator is tuned to the desired frequency, the FSLM has its low side tied to the neutral, and two points of measurement are taken. First, the voltage reading is taken between the capacitor and resistor where the FSLM is tuned to peak to the generator's frequency. The second reading is taken at the high line without retuning the FSLM.

为了避免谐波对电源线阻抗读数的影响，在测试时需要至少从电源线频率（基频）的10倍频率开始测量。这样既能保证读数的精度也能保护测试设备。这个测试的频率范围很宽，会达到MHz区域。这时应该使用频率选择型电位计（FSLM）、信号源和一个隔直电容器。这个隔直电容器对电源线频率是高阻抗，而对测试频率是低阻抗。例如：电源线频率为400Hz，那么最低的测试频率应在4000Hz以上。如果隔直电容器在400Hz阻抗为10000，那么在最低的测试频率下其阻抗应是1000。另外，应将FSLM设置在最窄输入滤波选项，并要保障信号源频率不是电源线频率的任何倍频（直至电源频率的20倍频率点），尤其是奇次谐波。除了在最低频率，其它时候隔离电容的损耗几乎是可以忽略的，因为隔直电容在400Hz频率下的损耗会很高。此时的元件只有某一值（如4000）的电阻，这个电阻在高于读数的最高频率下应该是非感性的。这些电阻、电容和信号源按此顺序连接后并在电源线间，其中电阻在火线端、信号发生器在零线端。把信号源调整到期望的频率，并把FSLM的低压端连接在零线上，接下来要做两步测量：第一、当FSLM被调整到信号发生器峰值频率时，读出电容和电阻间的电压读数；第二、不调整FSLM，读线间电压。

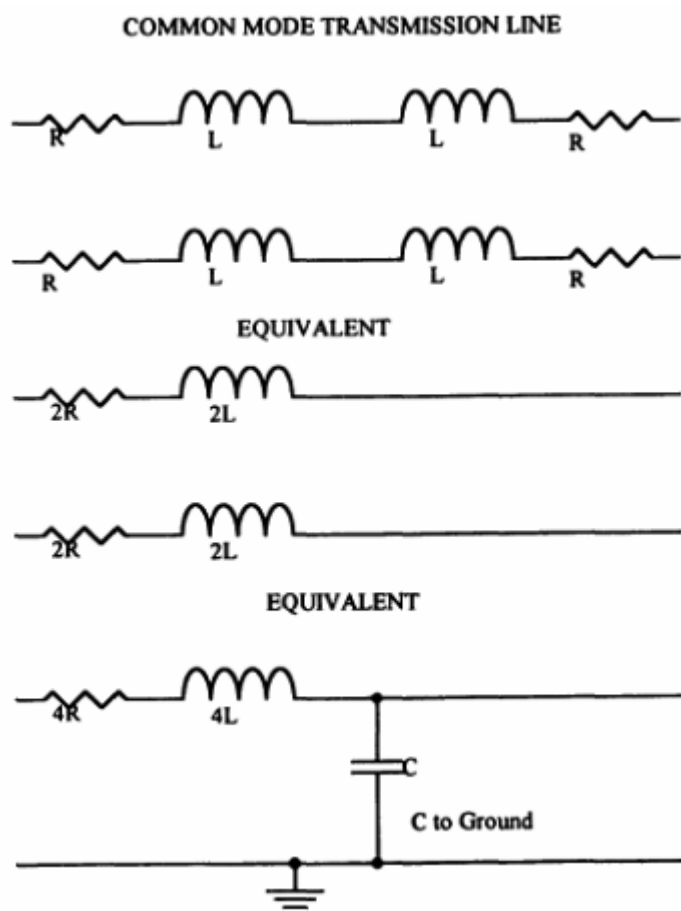


Figure 2.4 Equivalent common-mode transmission line.

Make sure that the FSLM front end can withstand this ac line voltage without blowing the front end. Many frequency-selective voltmeters or level meters have precision input pads that would generate smoke if the hot line is touched. A high-pass filter with a series capacitor input, not a inductive shunt input, could be built in a probe to be installed at the test lead to protect the FSLM's input (Figure 2.5). The cutoff frequency of this filter must be at least half the lowest frequency to be measured.

要确认 FSLM 的前端（对应于 low end）可以承受 AC 线电压（前端没有冒烟现象）。许多频率选择伏特表或电位计都有精确的输入连接器，如果短接火线就会产生烟雾。在 FSLM 探针的引线上装入一个带串联电容（这个电容在输入端口最外端，而不是感性在输入端口最外端）的高通滤波器会保护 FSLM 的输入端（图 2.5）。这个滤波器的截止频率应至少是测试频率最低值的一半。

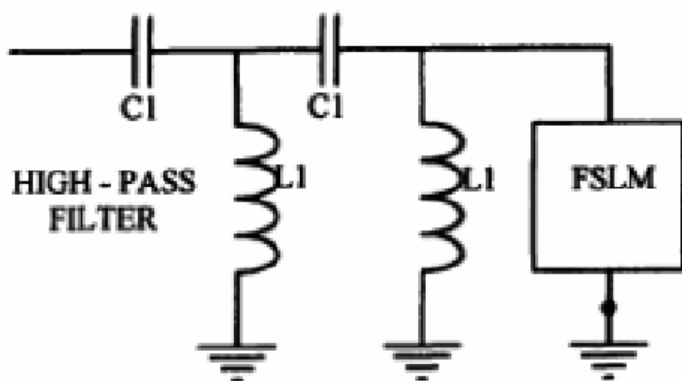


Figure 2.5 High-pass filter

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

这个滤波器在通频段损耗曲线可以有一些波动，并不需要有平坦的 Butterworth（巴特沃思）损耗曲线。如表达式所示，滤波器引起的误差可以忽略，另外在本节后部的示意图中也可以看到这一点。因为在表达式中所忽略的误差都处在同一频率点，所以两次读数的比值是相同的。例如：在 400Hz，最低读数频率为 4000Hz，将基频乘 10 的原因是避开电源线上一些潜在的高次谐波量。此高通滤波器的截止频率应该是 2000Hz，因此此高通滤波器的电容、电感值为

$$L = \frac{Z_{FSLM}}{2\pi F_{cutoff}} = \frac{50}{2\pi \cdot 2000} = 0.004$$

$$C = \frac{1}{2\pi F_{cutoff} Z_{FSLM}} = \frac{1}{2\pi \cdot 2000 \cdot 50} = 1.592 \cdot 10^{-6} \quad (2.11)$$

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

此处假设 FSLM 的输入阻抗是 50Ω。记录下此次读数和下一频率下的读数，则计算式为

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

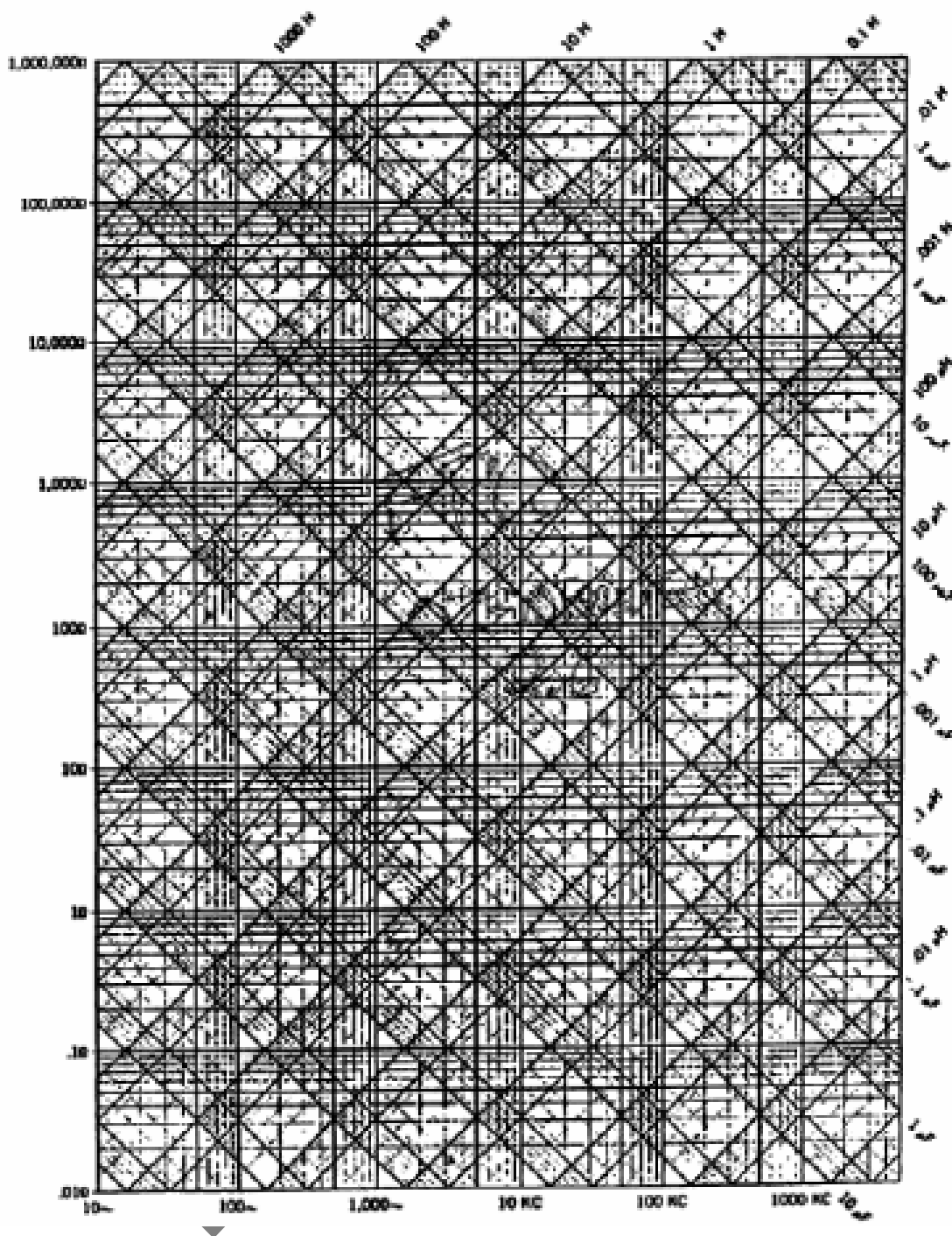


Figure 2.6 Impedance-frequency chart.

The voltage level, V_a , is recorded at F_0 between the capacitor and resistor and V_b is the high line voltage reading, also at F_0 , R is the value of the series resistor between the high side and the capacitor. F is the high-pass filter installed ahead of the FSLM to protect the input pads. The impedances at these various frequencies can be plotted for the various loads and compared (Figure 2.6). It is often better to use a battery-operated FSLM and signal generator to avoid ground loops.

V_a 是在频率 F_0 下，记录下的电容电阻间的电平， V_b 是在 F_0 频率下火线电压读数， R 是在

火线侧和电容间的电阻值。F 是保护 FSLM 输入连接器的高通滤波器。不同负载条件下，阻抗随频率的变化见图 2.6。另外，实际测试中常常使用电池供电的 FSLM 和信号发生器来避免地环路影响，这样效果比较好。

The dc measurements are almost the same, except that lower frequencies can be read and the highest frequency needed will be in the lower kilohertz range (Figure 2.7). Obviously, a high-pass filter is not required. These resistive readings can again be plotted on the graph. The readings will level out (resistive) at a low frequency in the kilohertz range. Here you could replace the FSLM with a ac voltmeter. This setup does not require as much caution as in the ac readings. Make sure, again, that your choice of FSLM or ac voltmeter can handle the dc output or line voltage. If they can not, simply use a capacitor (this replaces the high-pass filter) in series with the meter probe to take both readings. Figure 2.8 shows the changes for the dc line showing either a FSLM or an ac voltmeter and the blocking capacitor, if required. The same equation holds. Again, the various impedances would be plotted on the frequency graphs. These dc plots will normally start with readings in the low-milohm range at the lower frequencies up to around 10Hz and are then followed by a slope of approximately 45° upward (inductive). In some power supplies, the lower frequencies may have negative resistance reading. These plotted points will be on both sides of the best straight line, but the best straight 45° line can be drawn and the inductance read directly off the frequency-impedance graph (Figure 2.9).

对于 DC 的测试几乎与 AC 的测试方法相同，只不过此时可以读取更低频率的读数，而且最高频率也只才达到 kHz 范围(见图 2.7)。显然此测试不再需要使用高通滤波器了。测试的那些读数也可以制成图表，从图中可以看到，在 kHz 范围的读数曲线呈水平(表现为阻性)。在此试验中可以用 AC 伏特表代替 FSLM，实验的搭建也不必像 AC 那么小心。再次强调一下，需要确认选择的 FSLM 或伏特表是否可以处理 DC 输出或线电压。如果不能，那么在表的探头上串联一个电容(用其来取代高通滤波器)就可以读取读数了。图 2.8 表示的是使用 FSLM 或伏特计加隔直电容的 DC 线阻抗测试方法。如果需要，可以导出同样的计算方程，也可以同样画出阻抗—频率的变化曲线图。这些曲线从大约 10Hz 的低频以毫欧姆级读数开始以大约 45° 的曲线斜率上升(呈现感性)。对于许多供电电源，在低频下会有负的阻抗读数，而且这些读数的点可能会落在理想的直线两侧，但是可以通过这些点(忽略误差)画出理想的具有 45° 斜率的直线，而电感的值就可以在此频率—阻抗图上读取(图 2.9)。

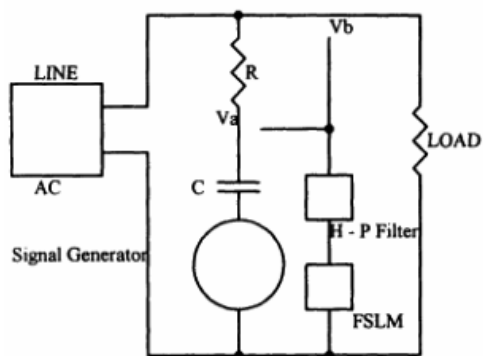


Figure 2.7 ac line impedance measuring set up with equation.

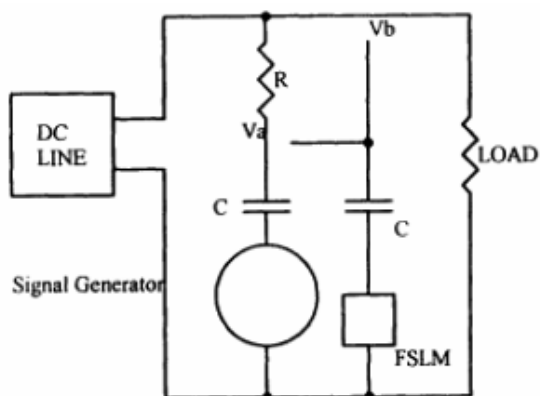


Figure 2.8 dc line impedance measuring setup with equation

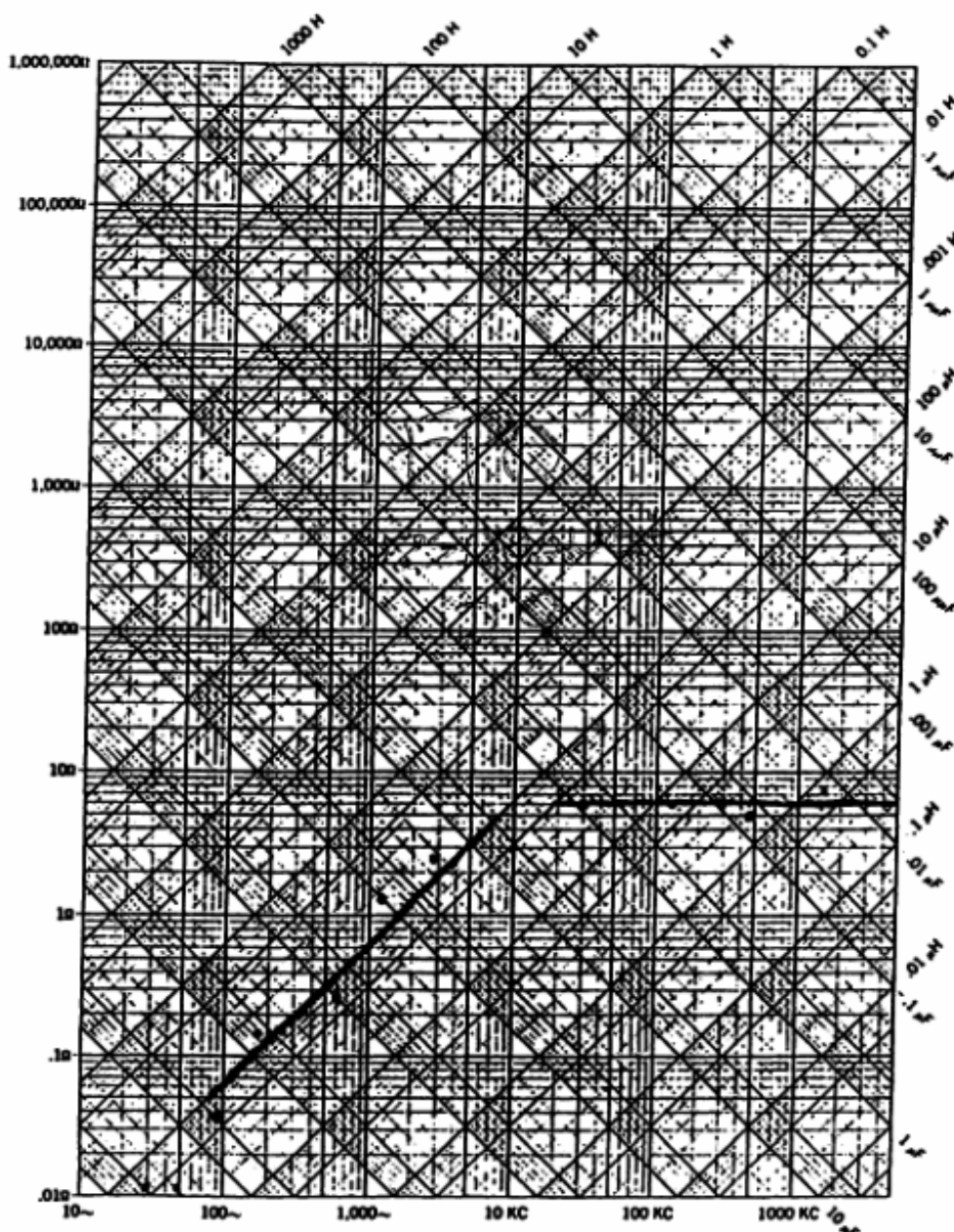


Figure 2.9 Impedance-frequency chart with plots.

The dc power supply readings will be a smoother line, making the reading much easier to read directly off the graph and flatten out at a much lower frequency. These points will then level out flat at the upper frequencies. The upper flat area, in ohms, is then the line output impedance of these upper frequencies. The operating frequencies of the switchers, clocks, or oscillators, which this power supply feeds, will fall well into this upper frequency range. This could starve some circuits being powered by this supply. A line simulation network (Figure 2.10) can be developed using the value of the low ohms in series with the inductor read from the graph. Then these two series elements will be in parallel with the higher upper-frequency ohms.

DC 电源线的读数是一条平滑的曲线，在极低频率下会变得很平坦，这样可以很容易地直接从图表上读数。这些读数点在高频下也呈水平，曲线平坦的上部（欧姆量数量级）就是高频下电源线的输出阻抗，这个电源所供电的元件，诸如开关管、时钟、振荡器等器件的工作频率都处在此曲线上部的所处的频率范围内。因此一些由此电源供电的电路会因此产生“饥饿”现象。使用一个低阻值的电阻串联具有图表上的读数值电感就形成了一个线仿真网络。此仿真网络会与高频电阻并联。

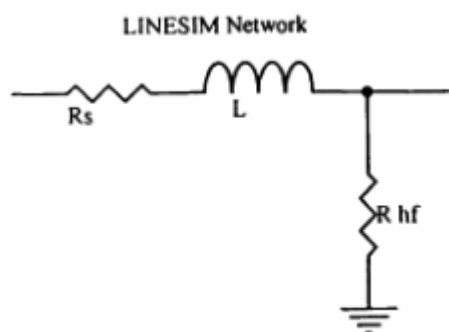


Figure 2.10 Simulation network.

Some power supply designers “cure” this problem with a capacitor tied across the output. This capacitor can oscillate with the inductance, read from the graph, causing ringing at the output of the supply. The best way is to calculate the maximum resistance that the following circuits can tolerate, say 20. Draw the 2-0 line across the same graph used above. Read the frequency where the two lines cross. The first line is the 2-0 line and the second is the inductive line from the graph and say that this cross point is 1800Hz. Divide this by 2.5, here giving 720Hz. Then calculate the capacitor needed to equal 20 at this frequency. The answer for this problem is 110uF in series with the 2-0 resistor tied across the line. If the curve were plotted again using the same setup as above, the curve would start the same way and head up the 45° line inductively to above 720Hz, then head 45° downward capacitively and fatten out on the 2-0 line. A good source for this is the book Power Conversion Design Guide and Catalog, Callex Manufacturing Co. Inc., Pleasant Hill, CA 9452.

一些电源设计者会在输出端接一个线间电容来解决“饥饿”问题，但是此电容会同（从数据图上读出值的）电感发生谐振，这样会在输出端产生振铃现象。最好的解决方法是计算出上述电路所能承受的最大电阻值，例如 20，接着在上面的使用的数据图中画出 2-0 线，然后

读出两条线的交点处的频率，例如交点频率是 1800Hz，将此频率除以 2.5 得到 720Hz，然后就可以计算在此频率下阻抗等于 20 的电容值。此问题的答案是，用 110uF 的电容串联 20 电阻并在线间。如果按上面的方法再次画出改变后（增加了 110uF 电容和 20 电阻）的曲线，那么可以看到，曲线以相同的路径开始，并以 45° 斜率上升到 720Hz 以上，然后以 45° 斜率下降（容性特征），并在 20 线处趋近平坦。详细论述可以参见 Power Conversion Design Guide and Catalog；Calex 国际制造公司, Pleasant Hill 出版社, CA9452。

The output impedance of these supplies is important to EMI design because the wrong simulation impedance network could be used. Without the decoupling capacitor and resistor, the line simulation network looks like a small milliohm resistor in series with a small inductor. These two components are arranged in parallel with resistor value of the upper fattened-out resistance. With the decoupling network used, the output could be replaced by a fixed resistor, as in our case above using 20. The best way to measure the line impedances is still with a network analyzer if you have access to one.

电源的输出阻抗对 EMI 设计很重要，因为这样可以不会用错 LISN。如果没有去耦电容和电阻，线仿真网络看上去是一个小的毫欧电阻与一个小电感串联后，再与具有上述数据图曲线平坦的上限阻抗值的电阻并联。如果使用了带去耦的网络，输出就可以用一个固定的电阻来代替，就像在上面的例子中我们使用 20 一样。总之，如果你已经有了网络分析仪，那么测量电源线阻抗的方法最好是使用网络分析仪。

3

Various AC Load Impedances

不同的 AC 负载阻抗

Load impedance varies with different types of loads. Few loads are resistive-some are capacitive or inductive. Some loads generate high-current pulses at twice the line frequency, others require high-frequency currents. All these statements assume that none of these loads have any input filtering or any other circuitry to offset the conditions mentioned here.

通常负载阻抗随类型不同而发生变化。实际应用中很少有负载是阻性的，多数都呈容性或感性。在这些负载中，有些负载会产生两倍于电源线频率的高电流脉冲，另外一些则需要高频电流。注意，以上的论述都是假设负载端无输入滤波器和消除脉冲的电路的情况。

§3.1 Resistive Load

§3.1 阻性负载

Resistive load is the easiest for the electromagnetic interference (EMI) filter to handle. This assumes that the storage capacitor is missing, as in the circuit in Figure 3.1. The diodes are missing in Figure 3.2, although they can be included with or without the transformer. The power factor for this load is near unity, and inductors can be designed to handle normal RMS current. This means that the method most magnetic engineers use to design inductors give the desired results. These inductors do not saturate at the peak current. If the load uses diodes, these can add severe noise problems to the already severe switcher noises. The entire load noise spectrum must be attenuated by the EMI filter.

电磁干扰滤波器处理阻性负载问题是最容易的。阻性负载情况一般是假设无储能电容的(如图 3.1 所示的电路)或二极管的电路(如图 3.2)，当然此电路可以含有或不含变压器。这样此负载的功率因数接近 1。同时可以设计电感器来处理 RMS 电流。这就意味着，多数电磁工程师用来设计电感器的方法都可以达到预期的结果。这些电感器在电流的峰值不会饱和。但是如果负载使用了二极管，那么这些电感就会增强已有的开关噪声，所以此时必须使用 EMI 滤波器来衰减整个负载的噪声谱。

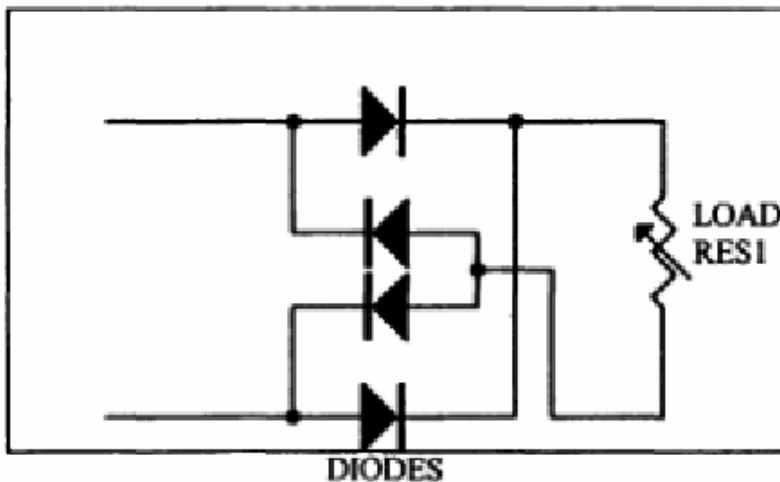


Figure 3.1 Resistive load with diodes.

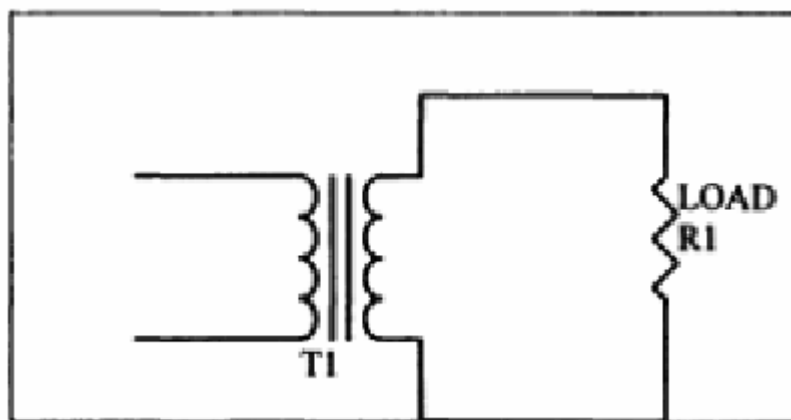


Figure 3.2 Transformer with resistive load.

§3.2 Off-line Regulator with Capacitive Load

§3.2 带容性负载的离线调节器

The most common circuit used today is off-line regulator (Figure 3.3). 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 on or 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 on the capacitor side of the diodes are often called sine wave pulses.

现今大多数共模电路使用离线调节器（图 3.3）。二极管输出端接一个大的储能电容，并为开关电源提供能量。这样负载就在近似短路和近似开路两种状态间变化，当然这主要取决于整流二极管的开和关的状态。如果正弦波电压与二极管电压降（一般硅管为 0.7V，锗管为 0.3V，译者注。）之和大于此时电容的电压，那么二极管导通，此时在电容侧的二极管上的大电流脉冲通常被称作正弦波脉冲。

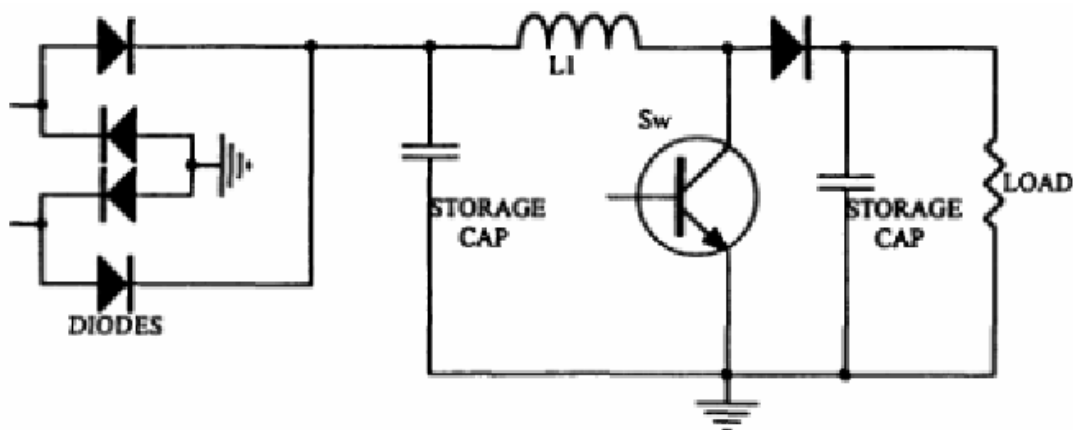


Figure 3.3 Off-line regulator with capacitive load.

The Fourier equation for this is

此波形的傅立叶表达式为

$$\frac{2t}{pT} + \frac{4tT}{p} \sum_{N=2,4,6,8}^{\infty} \frac{1}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \cos \frac{pN}{2} \cos \frac{2pNt}{T} \quad (3.1a)$$

or

$$\frac{2t}{pT} + \frac{4tT}{p} \sum_{N=2,4,6,8}^{\infty} \frac{(-1)^{N/2}}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \cos \frac{2pNt}{T} \quad (3.1b)$$

where $t/2$ 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° out of phase: the fourth, eighth, twelfth, and so on, are in phase.

此处 $t/2$ 是脉冲时间宽度， T 是全波（两个脉冲）时间， N 是谐波次数。另外，此表达式以 1A 为峰值进行了归一化。式中第一项包含所有偶次谐波和直流值；第 2、6、10 项与基波位相差 180° ，而 4、8、12 等项与基波位相相同。

On the line side of the diodes (Figures 3.5), the equation becomes

在二极管的电源线侧（图 3.5），表达式变为

$$\frac{4tT}{p} \sum_{N=1,3,5,7}^{\infty} \frac{1}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \sin \frac{pN}{2} \sin \frac{2pNt}{T} \quad (3.2)$$

and the first sine term can be replaced by

第一个正弦项可以用下式来代替

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



Figure 3.4 Capacitor pulse charging current through the diodes.



Figure 3.5 Capacitor pulse charging current on the line side of the diodes.

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° . The trouble with this is that 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. This would be nice, wouldn't it? The power factor is still the cosine of the angle between the start of the voltage sine wave and the start of the current pulse.

可以看到，(表达式 3.2)缺少直流部分值，并且谐波分量都是奇数次的。1 次、5 次、9 次等为同位相的谐波，3、7、11 等为与基波位相差 180° 的谐波。最后一项表明在 $t=0$ 时，所有的各项都为 0。这给了线电压一个基准的相位，并使功率因数增加到 1。这很不错，不是吗？功率因数仍然还是初始的正弦波电压和初始的电流脉冲之间夹角的余弦值。

These approximate equations [Equations (3.1) and (3.2)] still hold true even if a transformer is inserted ahead of the off-line regulator. This current pulse is not truly a sine wave pulse, as most people describe it. All those I have examined are a somewhat weighted toward the back half of the pulse. These equations were developed for several reasons. Then main reason was so that the switcher noise would show the relationship between the total size and weight of the filter with and without an inductor of critical value (Figures 3.6 and 3.7). This is discussed further in Chapter 11.

如果在离线调节器之前插入变压器，那么这些近似方程[方程 (3.1) 和 (3.2)] 仍然成立。正如很多人认为的那样，此时电流脉冲并不是真实的正弦波。所以我检查过的这些计算式对反向的半波有些加重。之所以推导出这些表达式有几个原因，其中第一个主要的原因是，通过开关噪声我们可以观察到，无电感器（临界电感值）和有电感器（图 3.6 和 3.7）的滤波器在总的尺寸和重量方面之间的对比关系，这将在 11 章进一步论述。

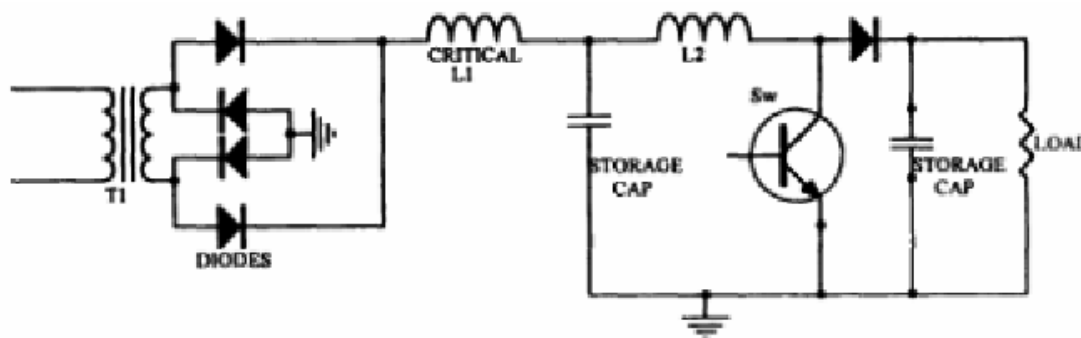


Figure 3.6 Off-line regulator with the added critical inductance.

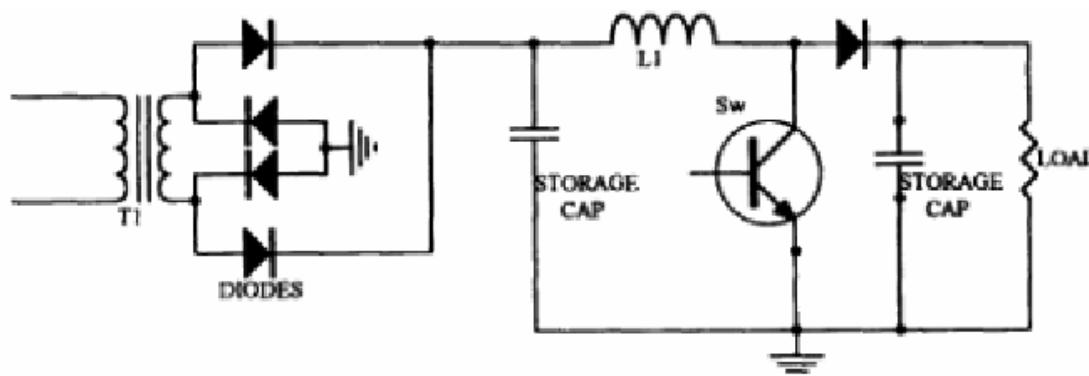


Figure 3.7 Off-line regulator without the critical inductance.

The total size and weight are smaller with the critical inductor for all serious specifications and may not hold for U.S. Federal Communications Commission specifications. The second reason is examined in Chapter 15, which discusses the three-phase filter. Finally, these equations were developed so that the full third, ninth, fifteenth, and so on, harmonic current could be demonstrated for the neutral leg of the three-phase Y, for capacitor design (Chapter 6).

几乎对于所有正规的规范，含电感器（临界电感值）的滤波器总的尺寸和重量会比没有电感器的小，但是 FCC 规范或许并非如此。第二个原因将在 15 章讨论三相滤波器时证明。最后，通过推导这些方程可以计算所有对于三相 Y 型连接的中性点的 3、9、15 等次谐波电流，从而进行电容器的设计（第 6 章）。

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

储存的能量馈送到开关电源，接下来需要考虑一个问题：此 RMS 电流值通常是设计滤波器电感所使用的电流值，但是设计者并不知道设计电感器所必需的峰值电流。目前的问题是峰值电流对滤波器电感有何影响？多数设计者设计电感的方法是，在 RMS 电流条件下取电感的磁通密度约为饱和磁通密度一半的磁通密度。对于设计此 AC EMI 滤波器电感来讲，这

是一种错误的近似，因为这样电感会饱和。电感的计算式为

$$H = \frac{0.4\pi N I_p}{M_{PL}} \quad (3.3)$$

where N is the number of turns, I_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 permeability drops. The permeability μ is a key player in determining the inductance L . The core material is spoken of as "soft" because of the BH curve. These are S shaped, or sigmoid, as in Figure 3.8, and are not made using square loop material. These cores require a strong magnetizing force, H to drive the core into saturation. A "hard" core, or square loop core, is driven quickly into saturation. The soft core is the type of core material needed for EMI filters. Cores that possess squarer loop characteristics are gapped, if they are used at all, reducing the hard magnetic core to a soft core (Figure 3.9). Soft core are also often gapped to make them even softer (less sigmoid) and for dc applications.

此处 N 为匝数， I_p 是电容充电的峰值电流， M_{PL} 是磁路长度， H 是磁化力（应为磁场强度，译者注）。对于铁心型材料，最大的磁通密度 B_m 是一个常数。 B 和 H 间的关联系数是磁导率 μ 。因此，大的电流脉冲使 H 增加时，会导致磁导率下降。而磁导率 μ 是决定电感值的一个关键参数。在 $B-H$ 曲线呈 S 形（如图 3.8），且不是方形环的磁材料时，就是通常所说的“软”磁芯材料。这些“软”磁材料需要有很大的 H 值才可以使之饱和，而对于“硬”磁芯或方形磁心则会很快饱和。软磁芯是 EMI 滤波器设计所需要的磁芯材料。另外，带有气隙的方形材料会使原来硬磁芯变成软磁芯（图 3.9）。通常软磁芯也常开气隙以使其变得更软，这样就可以将其应用在直流环境中。

$$L = \frac{0.4\pi m N^2 A_e \times 10^{-8}}{M_{PL}} \quad (3.4)$$

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

此处 A_e 是磁芯的横截面积，单位是平方厘米。通过上式可以看出， μ 值下降会导致 L 值降低，所以感抗会下降。

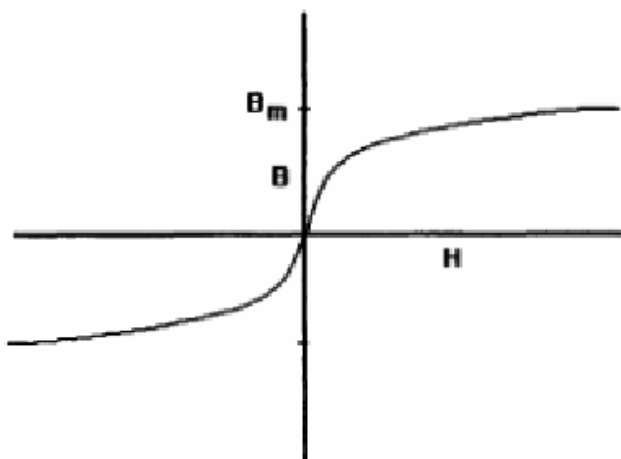


Figure 3.8 BH curve of a 'soft' (sigmoid) core.

What happens if the inductor saturates during this current pulse peak? The answer is that some switcher, diode, and other noise can ride through the saturated filter inductor during this time. These noise spikes can thus show up on the 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 were used, producing a very low impedance to this noise compared with the load and line impedance in parallel. Mclyman, at the Jet Propulsion Laboratory in Pasadena, California gives the following equations in his book, Magnetic Core Selection for Transformers and inductors (Marcel Dekker, 1982). 如果在电流脉冲的峰值处电感饱和，那么会发生什么事情呢？答案是：一些开关电源噪声、二极管噪声和其它噪声会在电感饱和时通过电感，因此噪声尖峰会叠加在正弦波波形上。这其中的原因当然是由于电感在这些未知的峰值电流处饱和所造成的，而这也是电感设计规范所不允许的。此时只能使用高品质的电容来补偿，这些电容会对噪声产生一个并联于负载阻抗和电源线阻抗间的低阻抗。Mclyman（加利福尼亚州 Pasadena 的喷气推进实验室的工程师）在他的书（“变压器和电感器磁芯选择” Marcel Dekker, 1982）中给出了如下的表达式

$$\Delta V_{rms} = \frac{I_0}{2\sqrt{3}FC} \quad \text{or} \quad \Delta V_{rms} = \frac{0.2887I_0}{FC} \quad (3.5)$$

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. Continuing with the discussion in C.W.T. Mclyman's book on page 40 and substituting in Equation (3.5).

此处 I_0 是 DC 负载电流， F 是纹波频率（在单相中其值为 2 倍的电源线频率）， C 是以法拉第为单位的电容值。详情请参见 Mclyman 书中第 40 页的公式，将其带入式(3.5)中，则

$$I_{rms} = 2pFC\Delta V_{rms} = 1.814I_0 \quad (3.6)$$

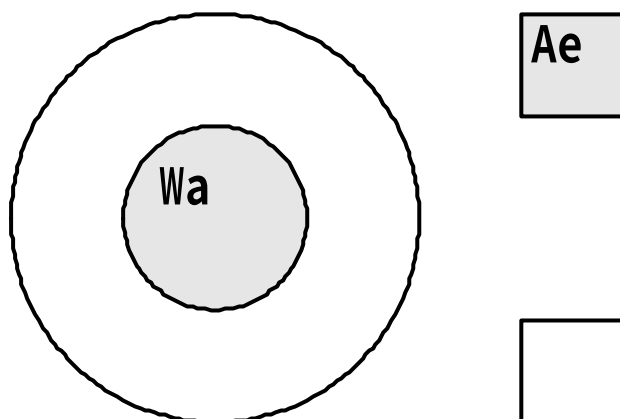


Figure 3.9 Toroid showing the terms A_e and W_a (shaded)

Thus, I_{rms} is a straight function of I_0 , 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.

因此 I_{rms} 是与直流电流 I_0 直接相关的函数，而与储能电容值无关。过去许多人认为，电容尺寸的增加会导致传导宽度降低，这样会增加电流脉冲的幅度，而窄的传导宽度会消减高的电流峰值。这些都会影响电源线的 I_{rms} 值：较高的电流峰值会被较窄的宽度削弱。关键是 EMI 滤波器设计者可以通过计算得到电源线的 RMS 电流值，但是通过 RMS 电流值设计的电感的体积是偏小的，这样电感会在电流脉冲的峰值处饱和。

Another equation relating the capacitance to the peak-to-peak ripple is
纹波电压峰—峰值与电容量的另一个关系式是

$$C = \frac{I_0 \times 10^{-6}}{2F(P-P)} mF \quad (3.7)$$

where P-P is the peak-to-peak voltage and F is the line frequency. This subject is continued in Chapter 11. Another way to avoid this problem is to design an inductor using a gapped core (Figure 3.10). 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.

此处 P-P 是峰—峰电压，而 F 是电源线频率。以后我们将在 11 章继续讨论此问题。另外一个避免电感饱和的方法是使用带气隙的磁芯来设计电感（图 3.10）。气隙使 BH 曲线倾斜，这样只有在很大的磁场强度时电感才会饱和。气隙同样也会使磁导率下降，但是会让磁导率更趋近于常数。开气隙的方法会让电感值在低、中和高电流负载条件下更趋近于常数。

The point is that the EMI filter will be larger, it will weigh more, and it will cost

more when using the off-line regulator because of the high current pulses.

需注意到的是，当使用离线调节器时，因为高电流脉冲的原因常会使 EMI 滤波器变得体积更大、重量更重并且成本更高。

§3.3 Off-line Regulator with Inductor Ahead of the Capacitor

§3.3 在电容器前使用电感的离线调节器

Adding an inductor ahead of the power supply filter capacitor widens the current pulse width and lowers the peak current pulse.(Figure 3.11). If the value of the inductor is equal to or greater than the critical inductance, the current flows all the time and the current through the inductor is the average current. The usual inductance design method will work, again, in this application.

在电源滤波的电容之前加一个电感会增大电流脉冲宽度，并降低电流脉冲峰值（图 3.11）。如果电感值大于或等于临界电感值，那么通过电感的电流会是平均电流。在此类应用中，常见的电感设计方法（依靠平均电流来设计电感器，译者注）却又很有效。

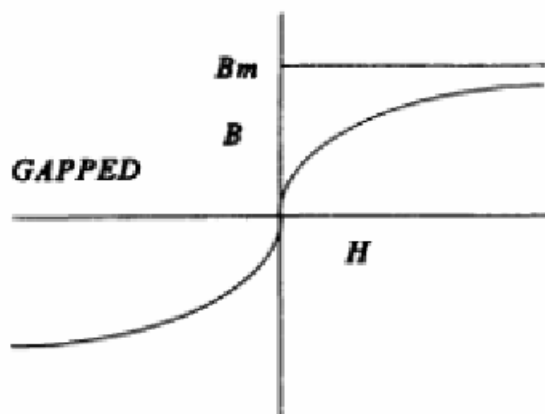


Figure 3.10 Gapped core. The BH curve is more linear and less sigmoid.

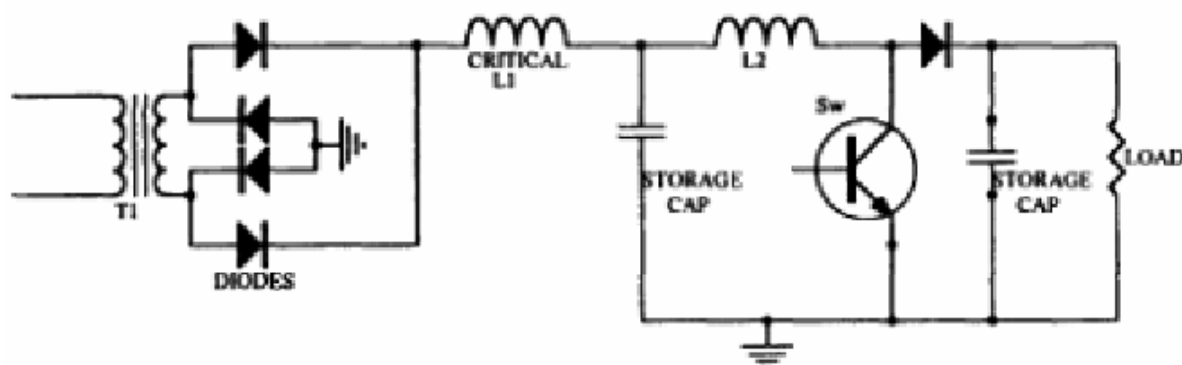


Figure 3.11 Off-line regulator with critical inductance and capacitive load.

The critical inductance is

临界电感值为

$$L_c = \frac{R_0}{6pF} \quad (3.8)$$

where R_0 is the load resistance calculated using the lowest, worst case current and the highest line voltage. F is the line frequency. This 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.

此处 R_0 是在最坏情况下（最低的电流和最高的线电压）计算出的负载阻抗， F 是电源线频率。此方法的缺点是电容上的存储压降从 AC 电压的峰值下降到平均电压，这也增加了电源的重量。但是由于滤波器不需要处理较高的峰值电流脉冲，所以 EMI 滤波器的重量会降低。

A company is involved with power supply designers who are proud of a light and compact power supply they have designed. The power supply designers ask the company to design a filter for this small power supply, but they cannot understand why the filter is larger than their supply and weighs more. To keep their power supply this small and light, they have eliminated all the normal internal EMI protection. Therefore, the output noise is high and the EMI filter is large! This problem is covered more fully in Chapter 11.

曾有这样的例子，有一些电源工程师设计的电源十分轻巧并很引以为豪，但是当这些电源工程师要求滤波器公司为这款电源设计滤波器时，他们却不理解为什么滤波器反而比电源的体积和重量大得多。其实这是因为，为了保证电源比较轻巧，他们去掉了所有的电源内部的 EMI 通用保护元件，结果是输出噪声很高，所以 EMI 滤波器体积就不得不设计得很大！有关这方面的问题会在 11 章更全面地讨论。

§3.4 Power Factor Correction Circuit and Coil

§3.4 功率因数校正电路和线圈

Today in Europe (Canada and the United States are following their lead), power factor correction circuits are a must (Figure 3.12). I have seen power factors as low as 0.43 resulting from the power factor of the power supply and the EMI filter because of larger capacitors. Power supplies that use off-line regulators have power factors as low as 0.7 by themselves, and improper filtering only adds to the problem. Power factor correction circuits work by switching the diode output voltage without initially storing the energy in a larger filter capacitor. Thus, the switcher current is in phase with, or follows, the line voltage. The switcher creates high-frequency pulses that follow the ac voltage sine wave, returning the power factor to near unity. This means that the output impedance of the EMI filter must be very low compared with the conducting load of the power factor correction circuit switcher frequency. The inductor conducts current in the same direction while the switch is open and closed so that the current through the diodes is a sine wave with a small triangular wave superimposed on top of it at the switcher frequency. It is the job of the EMI filter to attenuate this switcher frequency without starving the switcher. This is another reason a filter designed for very similar specifications may work well for one group and fail for the next. The first group might be using an off-line regulator and the last group a power factor correction circuit, and the output impedance of the filter may be inductive (maybe a T). The inductive reactance would be low to the harmonics to the off-line regulator but high to the power

factor correction circuit. This would starve the power factor correction circuit. 现今，在欧洲（加拿大和美国也同样）功率因数校正电路常常是必须使用的（图 3.12）。我曾见过这样一个例子，由于电源本身的功率因数和 EMI 滤波器的大电容的影响，使得功率因数低到 0.43。另外，使用离线调节器的电源，其功率因数也会低到 0.7，所以不正确地滤波只能使问题更加恶化。最初没有滤波器的大电容中储存能量的情况下，功率因数校正电路通过开关二极管输出电源来工作。因此开关电源电流与线电压同相，而开关电源产生的高频脉冲同 AC 正弦波电压一同传输，这就使得功率校正因数非常接近 1。这意味着，EMI 滤波器的输出阻抗与在功率因数校正电路开关电源频率下的传导负载阻抗相比是非常低的。当开关开合时，电感传导电流沿同方向传播，所以通过二极管的电流波形是一个由具有开关频率的三角波叠加在正弦波上所形成的波形。使用 EMI 滤波器的目的就是衰减开关频率下的噪声，同时又不能导致切换开关产生“饥饿”现象。以很相近规范设计出的滤波器常常会在一个系统中工作良好，而在另外一个系统中则出现问题。滤波器工作良好的系统可能使用的是离线调节器，而出问题的系统使用的可能是功率因数校正电路（这个滤波器的输出阻抗可能是感性的，如 T 电路）。这是因为对于谐波，在离线调节器环境中感抗是低阻抗，而在功率因数校正电路环境下感抗是高阻抗。这将使功率因数校正电路出现“饥饿”现象。

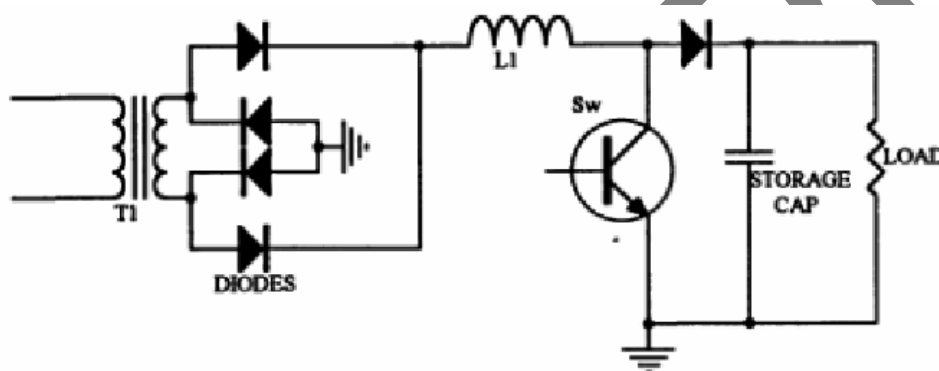


Figure 3.12 Power factor correction circuit.

The disadvantage of power factor correction circuits is that they are not 100% efficient. One of the main reasons to demand power factor correction circuits is to allow more devices to be plugged into wall sockets. These users did not count on the lower efficiencies of these circuits. Thus, little more equipment can be added to the wall outlets. The power required to drive the equipment using power factor correction circuits is now greater yielding little gain for the wiring and circuit breaker.

使用功率因数校正电路的缺点是，其效率不是 100%。但还会常常使用功率因数校正电路的主要原因是：可以让更多的设备插入墙上的插座（减小了设备对电源的影响，因此可以同时在电网上接很多设备—译者注），而这些使用者并不在意功率因数校正电路的低效率。这样，会使多一点的设备使用墙上的插座。由于带功率因数校正电路的设备不断地增加，使得对配线和断路器的使用有了一些增长。

There is also the power factor correction coil (Figure 3.13). In most designs, the EMI filter looks very capacitive at the power line frequency, and some specifications demand a near unity power factor for the filter for two reasons: obviously, for power factor correction, and also for leakage current problems.

This technology is archaic, at best, and is mainly seen on 400Hz power lines.

另外，还有使用功率因数校正线圈的时候（图 3.13）。多数情况，EMI 滤波器在电源线频率下是容性的。一些规范要求滤波器具有整功率因数，这有两个原因：功率因数校正和漏电流问题。使用功率因数校正线圈是一项比较陈旧的技术，多出现在 400Hz 的电源线供电电源中。

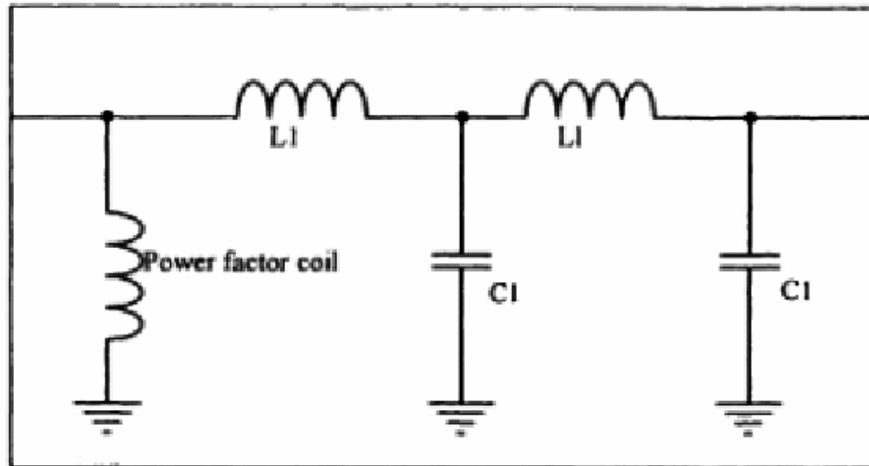


Figure 3.13 Power factor correction coil.

At the line frequency, the impedance of the inductors in the EMI filter is very low so that the capacitors add in parallel to the value, in Figure 3.13, of $2C$. The inductive reactance of the power factor correction coil must be equal to the total capacitive reactance at the line frequency. This returns the power factor to near unity, and the leakage current is reduced.

在电源线频率下，EMI 滤波器电感的阻抗非常低，所以在图 3.13 中的两个电容可以看成是并联的，其总电容值为 $2C$ 。在电源线频率下，如果要求功率因数校正线圈的感抗等于总的容抗，那么功率因数就会接近 1，相应地，漏电流也会降低。

$$L = \frac{1}{4\pi^2 F^2 C} = \frac{1}{4\pi^2 400^2 (6 \times 10^{-6})} = 0.026 \quad (3.9)$$

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

此处 C 是滤波器对地电容的总容量（单位是法拉第）。可以看出，这是一个相当大的电感器，

流过此电感器的电流值等于总电容的漏电流（但是有 180° 位相差）。这样，原来电容的漏电流就变成了这个新形成的储能电路的环流。在决定此储能电路的 Q 值时，需要考虑到两个相互冲突的因素。如果单纯为了尽可能地降低漏电流值，就需要最大的 Q 值；另一方面，考虑长时间使用条件下滤波器老化和工作应力等因素，又要求 Q 值较低。另外，在实际安装使用条件下，例如处于远端的发动机的情况，就常会出现电源线频率在很宽的频率范围内漂移的现象。这种频率漂移现象会使网络偏离中心频率点，并导致并联的储能电路工作在其阻抗曲线的边缘。当然，无论 Q 值是高是低，环流的计算式总是

$$I_{\text{circ}} = QI_{\text{line}} \quad (3.10)$$

where I_{circ} is the circulating current in this tank, the same as the leakage current before adding the power factor correction coil, and I_{line} is the new leakage current. The reason for the concern about leakage current is the danger to anyone touching the filter, or the unit in which the filter is mounted, if the safety ground has been removed or broken off. This can be lethal because of electrical shock when this ground is cut. Equation (3.10) shows that the leakage current has been reduced by a factor of Q .

此处 I_{circ} 是储能电路的环流，其值等于没有安装功率因数校正线圈前的漏电流。 I_{line} 是新的漏电流。此处关注漏电流的原因是，在没有连接安全地时，在接触滤波器或接触安装滤波器的设备时漏电流会对人体产生危害。当安全地与系统断开时，这种电击往往是致命的。式 (3.10) 显示，因子 Q 降低了漏电流。

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

在长岛的 RFI 公司，发货之前对每个滤波器的功率因数校正线圈都做调整。这就要求他们必须将高 Q 值的功率因数校正线圈和滤波器按一一对应的方式安装。如果不能这样做就只能要求把 Q 值限制在 10 左右。RFI 将功率因数校正线圈安装在负载端，而另外一些公司将其安装在电源线侧的前端。这两种安装位置在电性能上没有任何差异。

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

漏电流测量规范更倾向于测量实际的对地电流而不是规定对地的电容量。利用这一诀窍可以设计更大的对地电容以便提高共模性能。使用以上的 Q 值,例如 10,那么 $0.02\mu\text{F}$ (对 400Hz 系统) 可以变为 $0.2\mu\text{F}$ 。这种方法将会减小共模电感器 (此处称作 Z_{or}) 的值和物理尺寸。这会提高共模衰减,所以即使是降低了 Z_{or} 电感的值滤波器也能很轻易地达到插入损耗的要求。此处需注意的是元件值和电源线频率都必须稳定,这样才能确保设备没有不匹配,漏电流也不会升高。同样功率校正因数会接近于 1。在 50 和 60Hz 的应用条件下,功率因数校正线圈的物理尺寸会变得十分巨大,因此在此应用环境中很少使用。

EMCARE

4

The DC Circuit-Load and Source

DC 电路的负载和源

There is a difference between ac and dc circuit designs for filters. It is understood that many off-the-shelf electromagnetic interference (EMI) filter manufacturers sell filters specified to work for both alternating and direct current. This adds to the difficulty because the sources are very different. Filter capacitors must be designed to handle the ac voltage and all the total harmonic current calling for larger margins at 400Hz. This should also increase the margins of the capacitor for the higher ac voltage. The inductor should be gapped to handle the direct current; otherwise the inductor is driven into saturation after a few cycles of operation. This dual requirement adds to the overall cost of the filter. The way the filter manufacturer helps to cut the cost of these filters is to build sizable in-house orders for shelf stock. These filters are usually grouped in families of different current values with similar specifications. The same parts are used for all the different current values, when possible, and all this drives down component costs. This makes the parts less expensive and reduces the cost to the customer. The way for the customer to cut costs is to have more than one application requiring the same filter. This increase in the number of filters required for a period of time also cuts the filter price. This is true for all filter types.

对于 AC 和 DC 电路，滤波器的设计是不同的。许多制造商销售的 EMI 滤波器在交流和直流条件下都可以工作，但因为电源（AC 或 DC）的特点相差很大，因此会增加滤波器的设计困难。例如，滤波器电容需要能处理交流电压，并需要有很大的余量以便于处理所有的基于 400Hz 的谐波电流。对于高 AC 电压的情况同样需要电容有很大的余量。而对于直流电流，则需要电感器开气隙，否则电感器很快就会饱和。另外，以上的双重（AC 和 DC）需求会增加滤波器的制造成本，制造商消减滤波器成本的方法是，按安装托架的要求设计合适尺寸的滤波器，并按不同的工作电流将滤波器按类别分成不同的系列。同一系列的滤波器应用在不同电流环境中，这一系列化的设计方法虽能降低元件的成本，但会出现针对多种应用环境使用同种滤波器的情况。当然同种滤波器产量的增加也会减少生产成本。目前几乎所有的滤波器都采用这样的设计方法。

The loads are also different. Some are resistive, as in heaters that maintain the mountaintop repeaters above a certain temperature. Some equipment may use this supply voltage directly; others may be switchers, for which the inductors should be gapped for pot cores or C cores, but MPPs and powder cores are already gapped as a result of the manufacturing process. Gaps are distributed

all around the core of these powder cores.

滤波器面对的负载情况同样也是各不相同的。如一些加热器，其阻抗是阻性的；一些设备直接使用供电电压，而另外一些则有可能使用开关电源。对于开关电源类负载，其电感器应使用开气隙的罐状或 C 型磁心，但由于制造工艺的原因，往往使得 MPP 类和铁粉磁芯也同样具有气隙。但对于这种磁芯，气隙分布于磁心各处的（铁粉芯的气隙几乎平均分布在磁芯内部的各个地方，这使得铁粉芯不易饱和。其制造工艺故意要产生这些气隙——译者注）。

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

本章推荐的针对电路中元件的设计方法可以减小设备的 EMI，因此可以使整个设备对滤波器插入损耗的要求有所降低，结果降低了滤波器的物理尺寸、重量和制造成本。同时本章讨论了一些其它方面的设计技术。

§4.1 Various Source Impedances

§4.1 不同的源阻抗

The dc power line feed is very short, providing link coupling well into the megahertz range. Often it uses a normal chassis ground, making a balanced circuit impossible. This is where tubular filters play a role, using feedthrough capacitors without line-to-line capacitors. Some have capacitance only; others are Ls and ps and the rest are Ts. This dc power is the type found in aircraft, shipboard, telephone company, and mountain repeater sites. This power is furnished by various arranged systems, such as battery racks, standby generators, solar panels, and wind generators. The battery feed store energy to the various systems and help to regulate the voltage. These large batteries are called deep cycle batteries and can feed the entire system from several hours to many days, depending on the system. These are designed this way when the conditions warrant the higher costs or are referred to as life threatening, the highest cost system requiring more standby batteries, generator fuels, and the like. These types of conditions can be caused by any outside power failures, such as a solar panel on a cloudy day, downed power lines, and fuel supply outages.

因为 DC 电源馈线常常很短，所以可以在 MHz 范围内形成较佳的链耦合状态。通常情况下电源使用连接于设备底盘的大地，这样我们就可以设计平衡的电路结构。在这种应用条件下，我们用的滤波器使用的是穿芯电容而不是线间电容。滤波器的电路结构通常是 C 型、L 型、p 型和 T 型的。DC 电源常常在航空器、船舶、电信公司和位于山上的转接站等设备上使用，这些 DC 电源通常是由诸如电池组、备用发动机、太阳能电池板和风力发动机等设备提供的。电池向不同的系统供电并能调节电压，针对不同的系统有些大电池常常可以提供几个小时到很多天的供电。一些高成本的设备、需要保障生命安全的设备、需要更多备用电池组和发电机燃料的昂贵设备等常会使用电池组的设计方法。当外部电源失效时，（如：太阳能电池在阴天的时候、电源线出现停电、发动机燃料耗尽等情况）就会启用备用电池组。

The plates of the battery rack act as a capacitor and shunt the middle

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

电池架的金属板在中频段作用就如同一个电容器, 可以将传导发射噪声旁路到大地, 但在高频下由于输入电缆和电池架金属板的电感的原因, 对高频噪声没有衰减。电源引线上的射频电流会辐射出 H 场, 通常通过滤波可以避免电源引线的辐射。DC 输出为开关电源供电, 这也会产生很大的射频噪声, 同样开关电源的变压器的寄生振荡和与开关电源相连的二极管也会产生射频辐射。

The other type of dc system is from an ac power supply, and again the feed is very short. This dc normally feeds a switcher(s). The difference is that this output impedance is high at the switcher frequency even though the output impedance is only a few milliohms from dc to 10Hz or so. The output impedance is inductive above this point and rolls off flat or resistive at, say, 5kHz. This power supply looks like a milliohm resistor in series with an inductance, and this is shunted with the higher resistor value. (See Section 2.4.)

另外一种 DC 系统是通过 AC 电源供电的, 同样输入线长也很短。这个 DC 电源同样为开关电源供电, 不同之处在于, 尽管开关电源的输出阻抗在 DC 到 10Hz 频率范围内只有几个毫欧, 但是它的输出阻抗在开关频率下却是高阻的。其输出阻抗在 10Hz 的频率点以上是感性的, 并且在频率达到 5kHz 的时候, 阻抗逐渐趋于平坦并呈阻性。这种电源就如同一个毫欧级电阻串联一个电感的情况, 因此电源会被高阻值 (高频下) 所旁路 (见 2.4 节)。

Without being fixed, this situation starves the switcher because the inductive reactance of the inductor at the switcher frequency, and especially the harmonics, is a high impedance equal to R_{hf} (Figure 4.1). Is this true for all dc power supplies? No: some designers have been made aware of the load conditions the supply must provide by the customer before design. The power supply people can make this output impedance low at the switcher frequencies. Does the filter designer need to know this output impedance for the dc supply feeding the filter before designing the new filter for the remote switcher?

如果没有解决电源的高阻问题, 那么就会因为在开关频率下 (谐波频率更为明显) 电感为阻抗为 R_{hf} (图 4.1) 高阻的, 并使开关电源产生“饥饿”现象。对所有的 DC 电源都是这种情况么? 不, 在设计之前, 一些设计者已经注意到了电源面对客户负载可能会出现的情况。此时可以将电源的输出阻抗在开关频率下设计成低阻抗。那么在为远端的开关电源设计滤波器时, 设计者需要知道 DC 电源输入滤波器端的输出阻抗么?

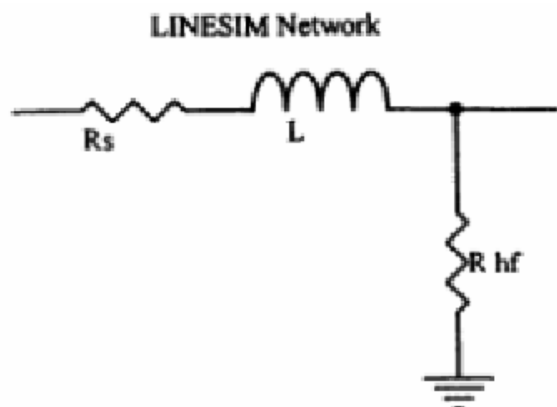


Figure 4.1 Line simulation circuit.

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

如果没有安装滤波器，且 DC 电源的输出阻抗是低阻抗的情况下会怎样？此时，开关电源没有“饥饿”现象，但此时会存在 RF 电流脉冲形成的高辐射，因此会产生很高的 H 场。如果在没有安装滤波器时，DC 电源的输出阻抗是高阻抗的情况下会怎样？再重复一次上面的论述，开关电源会产生“饥饿”现象，但是电流脉冲会很微弱，因此只会产生很微弱的 H 场。上面说的这两种情况在滤波器的应用中都会遇到，所以滤波器设计者只能不关心电源的输出阻抗了。因此滤波器的设计会变得比较容易，在下一节我们同样会在设计滤波器时忽略电源输出阻抗。

§4.2 Switcher Load

§4.2 开关电源负载

Most loads are of the switcher type. The output impedance of the filter at the load or switcher end must be low so as not to starve the load. This assumes that the switcher does not have a capacitor at the switch to lower this impedance. I lump these capacitors as part of the filter. If the capacitor is in the circuit without the filter designer's knowledge, this could detune the output network, lowering the cutoff frequency of this filter. The two capacitor values add because they are in parallel. This assumes that the connecting wire is not a long lead length. This appears as an inductance, tending to split the capacitors.

许多的负载都是开关电源类，为了避免负载的“饥饿”状态，在负载端或开关电源端滤波器的输出阻抗必须是低阻抗。此时假设开关电源在开关处没有降低阻抗的电容。此处我把这些电容概括为滤波器的一部分。如果电路中的电容器设计没有考虑到滤波器的因素，那么这将会使输出网络出现误调从而降低了滤波器的截止频率。这两个电容（滤波器中的电容和电路中的电容—译者注）是并联的，其容值为二者之和。此时是假设连接的电源线是短引线。而

长引线会等效成一个电感，电感的存在会隔离这两个电容。

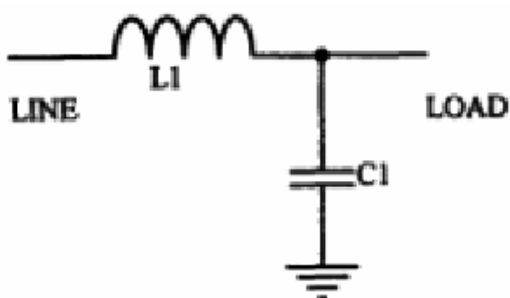


Figure 4.2 Single unbalanced L filter.

This dc application would require a single L, so that the capacitor facing the load is as high as possible (Figure 4.2). This reduces the drop in the dc voltage while the switcher is turned on. If double Ls are used, the value of the final capacitor is smaller than half the original. This may allow the voltage feeding the load to drop more than that of the single L filter. This is still true with the double L even though the ripple voltage on the feed wire is the same (see Section 16.3). If the return is through the chassis, the capacitor facing the load should be a feedthrough type. Otherwise, it should be a line-to-line capacitor in which the system has a return power lead. The input inductor of the filter is high impedance and reduces the H field no matter what the source impedance of the power supply happens to be. These are easy to design, and the method is as follows. The switcher frequency is 60kHz, the dc voltage is 60V, and the peak on current is 1.2A. Divide the voltage by the switcher on current to find the on impedance of the switcher, and divide this on impedance by 10 to minimize the voltage drop. Determine the capacitor value by making the capacitive reactance equal to this impedance at the fundamental switcher frequency, 60kHz. For the inductor, multiply the on impedance by 10, and then solve for the inductor reactance equal to this impedance at , as before, 60kHz.

这种直流应用环境需要的是单 L 型滤波器，这样电容面对负载就可以是高阻抗（图 4.2）。因此当开关电源导通时，可以减少 DC 电压降。如果使用双 L 型滤波器，则后面的电容容量应小于前面电容容量的半值。双 L 型产生的电压降会比单 L 型的大些。对于输入电源线上的纹波电压也同样如此（见 16.3 节）。如果回路是机壳，那么面向负载的电容则需要是穿心电容，如果系统有电源线回路，就需要使用线间电容。滤波器的输入电感是高阻抗的，其作用是无论源阻抗如何都会降低 H 场。简便的设计方法如下，例如，开关频率是 60kHz、DC 电压是 60V 且电流峰值为 1.2A。用电压除以开关电源的电流得到开关电源的阻抗，为使电压降最小实际上要取此阻抗的 1/10，令电容在开关频率 60kHz 频率下阻抗等于刚才计算的阻抗值，进而计算出电容容量。对于电感，将开关电源阻抗乘以 10，并求出 60kHz 频率下感抗等于此阻抗值的电感量。

$$\frac{60\text{Volts}/1.2}{10} = 5\Omega = 0.53mF$$

$$\frac{60V}{1.2} = 500\Omega = 1.326mH$$

This is rather a large inductor, but if the circuit is to be balanced then two inductors of 670uH each are needed. The small capacitor of 0.6uF is wired between the two inductors and faces the load. The 1.4uH would have to only carry about 0.5A, making this possible even unbalanced. This can be made into a simple formula:

计算出的这个电感值相当大，如果使用平衡电路那么每路的电感值就只需要 670uH 了。0.6uF 电容器在面向负载端接在线间。如果 1.4uH 的电感只需要承载 0.5A 的电流，那么使用非平衡电路也是可以的。可以使用简单的公式来计算电路参数：

$$C = \frac{10I_P}{2pF_{SW}V_{dc}} = \frac{10 \times 1.2}{2p \ 60,000 \times 60} = 5.305 \times 10^{-7} = 0.5305mF \quad (4.1)$$

$$L = \frac{10V_{dc}}{2pF_{SW}I_P} = \frac{10 \times 60}{2p \ 60,000 \times 1.2} = 1.326 \times 10^{-3} = 1.326mH \quad (\text{原文为}$$

MH, 译者注)

The equations give inductance in henries and capacitance in farads.

公式中电感的单位是亨利，电容的单位是法拉第。

§4.3 DC Circuit EMI Solutions or Recommendations

§4.3 DC 电路的 EMI 解决方案或建议

This is included here because most problem solutions deal with dc power feeds or are part of the power supply. Some of these may conflict, use what works. 之所以在这里写这方面的内容，是因为多数的问题解决方案是处理 DC 电源馈入或部分电源的。有时这些方法会相互冲突，那就使用实际上有效的。

Most systems grow by: Inception of the idea, choosing a method or concept, the various designs and problem solutions, combining the various circuits or circuit boards to test the system, and system corrections, and then the final, the power supply, oops, the input transformer, and double oops, the EMI filter (in that 2.25 × 1.6 × 0.75 inch or any other small leftover volume into which the designers could not fit anything else). Usually, EMI filter is thought of last. Then people wonder why the filter is so big and bulky and why it cannot fit in that little leftover volume next to the input transformer. The plan for EMI should begin almost at the inception of the idea and should be required on each board or circuit that is applicable. If this is done correctly, the EMI filter might fit the little leftover volume and the cost, along with the weight of the filter, is also greatly reduced.

许多系统是这样开发的：提出一个初始的想法，然后选择设计方法，接下来修改设计并解决问题，进而将不同的电路和板卡安装起来进行系统测试，再进行系统修正。最后是电源，还有输入变压器和 EMI 滤波器（如在尺寸为 2.25 × 1.6 × 0.75 的空间，或是其它不能安装别的设备的剩余空间安装滤波器）。通常都是在最后才考虑到 EMI 滤波器，此时工程师

常会发现滤波器体积很大，以至于不能在与输入变压器相邻的剩余空间中安装。其实，应该在最初的概念阶段就把各个板卡和电路对滤波器的需求进行规划，如果规划得当，滤波器就能够合理地使用很小的剩余空间，同样也会大大降低滤波器的成本和重量。

§4.3.1 Some Ideas for the Initial Power Supply

§4.3.1 设计电源之初的一些想法

Include an inductor of critical value in front of the storage capacitor to remove the high current peaks. The lack of this inductor increases the size of the EMI filter, it weighs more, and it costs more, especially for the military, for which the specifications are harder to meet.

电源设计之初需要考虑在储能电容前加具有临界值的电感，此电感会消除大电流峰值。如果缺少了这个电感，那么 EMI 滤波器的尺寸、重量和成本都会增加。尤其是军用环境，缺少这个电感就会很难达到规范要求。

Isolate the off-line regulator to reduce the common mode and remember that a capacitor is not a capacitor at all frequencies. Most larger power supply storage capacitors are inductive by 20kHz. Parallel them with a good-quality extended foil or ceramic capacitor. Use the proper snubbers in the switchers to remove spikes, and stop reverse current through diodes, for example. This reduces diode noise. Keep the various supplies isolated, and use twisted pairs to power remote units rather than a ground return. Filter the isolated leads, or closely spaced traces on the PC board, at the device end. This way, the primary current in the traces or twisted pair is direct current rather than pulse at the switcher frequency. Lay out the power supply transformer, input, or switcher(s) for minimal magnetic coupling from the transformer to a susceptible receptor either by increasing the distance between these devices or by finding the best orientation of the transformer. Sometimes the magnetic field radiating from around the transformer is stronger in one direction than in the other: the lines of magnetic force may extend farther. Use single-ended converters rather than a flyback transformer, and do not use switches known to be noisy, such as SCRs (this should be obvious). Place sections known to be noisy within a container or shield, which helps to stop the H and E fields. Cold rolled steel must be quite thick to reduce the magnetic field generated by a transformer or any other device. These containers are often silver plated inside and out to enhance surface conduction. The better the conduction, the better the H field is attenuated: the lower the current on the outer skin. This may require a mu metal can or foil than would be thinner and lighter for the same attenuation of the H field but cost much more.

隔离离线调节器会减少共模噪声。但需要注意的是，电容在全频段下并不总是表现为电容特性的。许多大的储能电容器在 20kHz 时就表现为感性，这时可以并联一个高质量的金属薄膜电容或陶瓷电容来解决这个问题。另外，如在开关电源中使用恰当的缓冲器可以消除尖峰，并可以阻断二极管中的反向电流，这样就可以降低二极管的噪声；隔离不同类型的电源，并使用双绞线为远端电路单元供电，而且要避免使用地回路；为单独的引线或空间距离很近的 PC 主板上的印制线滤波。这样，在印制线和双绞线中主要的电流是直流而不是开关频率下

的脉冲。通过增加间距或是将变压器放置在最佳方位等方法对电源变压器、输入或开关电源进行布线，这样会减小变压器和易感接收器间的耦合。通常有些时候变压器在一个方向上的磁场辐射会大于其它方向，而且强磁场方向的磁力线会延伸到很远。最好使用单端转换器而不使用逆程变压器，另外因为噪声的原因也要尽量不使用开关元件如 SCR。将辐射噪声的电路加盖屏蔽 H 和 E 场的屏蔽罩。对于变压器或其它辐射的磁场的设备屏蔽罩就得使用厚度比较大的冷轧板。屏蔽罩表面电阻越小其对磁场的屏蔽就越好，其外表面的噪声电流也越小，所以为了降低屏蔽罩表面电阻，常常在屏蔽罩内外镀银。通常也可以使用 μ 金属罐和薄膜来屏蔽磁场，这样做的优点是可以将屏蔽罩做得很轻也很薄，但缺点是成本高。

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

提高电感或开关电源的变压器的自谐振频率会使寄生振荡的频率增高。这些寄生振荡频率下的噪声通过整个通路最后回到滤波器，并在滤波器处被衰减。可以在电源和开关电源的入口处加一个容值很小的电容来衰减此噪声。

§4.3.2 Other Parts of the System

§4.3.2 系统的其它部分电路

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

时钟和另外一些源产生很大的噪声。有些时候，最初的噪声并不是由时钟产生的，而是由电源或其它由电源供电的电路产生的。应当确保每个供电电源都经过了滤波，滤波器应与被滤波的电路（如时钟或 DAC）尽可能地接近。

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

将地分割成模拟地和数字地，将噪声显著的系统的地也与其它电路分割，这样地线就不会在低噪声电路中感应出电压噪声（如果感应出噪声，这个噪声在某些情况下可能被电路放大）。对噪声源进行屏蔽以减小 H 场辐射。

Keep the noise within the EMI filter inductors and other inductors in nearby areas, even though the cores are toroids (which are quieter and radiate less), by mounting these inductors in quadrature to reduce their coupling (Figure 4.3). This is especially helpful if the cores are not toroids.

虽然圆形磁芯电感的辐射很小，但 EMI 滤波器中的相邻电感的安装方向最好还是相互正交的，这样可以减少它们之间的耦合（图 4.3）。如果磁芯不是圆形的，那么效果就更明显了。

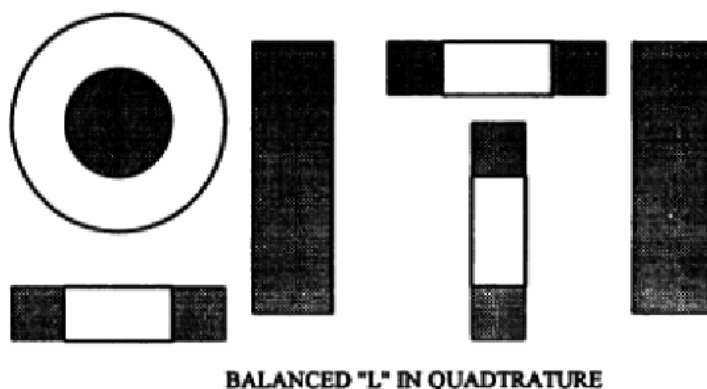


Figure 4.3 Toroids in quadrature.

Figure 4.3 shows a balanced double 'L' in which the first two inductors feed the capacitor using the vee technique (see Chapter 6). The two inductors in quadrature are also using the vee technique. The lower left inductor could radiate to the upper right inductor, but the increased distance and the capacitor between them reduces this.

图 4.3 表示的是平衡电路结构的双 L 型滤波器，此滤波器的头两个电感连接到使用 v 型技术设计的电容上，另外两个按正交方向安装的磁环也如此。这样虽然左侧下面的电感会辐射到右侧上面的电感，通过增加他们彼此的距离可以减弱这种耦合，同时两个磁环中间的电容也会起到减弱耦合的作用。

§4.4 Lossey Components

§4.4 损耗型元件

Another technique is to use lossey components. One such technique is from transmission lines in which high frequencies are absorbed within the dielectric. Coaxial cable developers continue their researches to find dielectrics with lower and lower losses; others have used this phenomenon to enhance losses to make lossey systems. One such company, in New York, is Capcon, Inc. (Figure 4.4). Their material is tubular and can be threaded with a filter hookup wire; it provides substantial losses above 10MHz and well into the gigahertz region. This depends on the total length of the material within the filter. Various inside diameters are available, allowing easy treading for capacitor, inductor, and other hookup leads. This functions better than most ferrite beads because most ferrite beads saturate at around 5A. These beads have a self-resonant frequency of around 50MHz. The lossey components just mentioned do not fail either of these requirements. Only 1 foot of this lossey line material gives as much as 100dB at 10GHz; the ferrite bead gives 20dB loss at the peak of 50MHz, and the loss falls to zero by 100MHz. This material is available in sheets of various thicknesses for enclosures and is also very dissipative. Radiated energy can be reduced by covering the filter enclosure, or a noisy device, with this material.

另一种常用的滤波器设计方法是使用损耗元件。此方法源于传输线理论，传输线上介质会吸收高频能量。虽然同轴电缆的设计是需要低损耗特性的，但是我们却需要增加损耗并利用这一特性来设计损耗系统。如纽约的 Capcon 股份有限公司就是一家损耗材料公司。他们的材

料(图 4.4)是管状的,并可以用滤波器的连接线来配装。这种材料在 10MHz 以上可以提供很大的损耗,直至 GHz 范围其损耗性能都表现良好。这种材料损耗的实际量值直接取决于滤波器内此材料的总长度。这个公司还提供不同内径的材料,这样就很容易对各种电容器、电感器和互连线进行处理。这种材料比多数的磁珠性能要好,因为磁珠多数会在 5A 左右饱和,另外磁珠的自谐振频率在 50MHz 左右,但是上面所提到的这种材料却没有这两种缺点。这种损耗材料在 10GHz 频率下大约有 100dB 的损耗,而磁珠在 50MHz 的峰值点的损耗却只有 20dB,到 100MHz 就下降为零了。此材料也有可以在壳体上使用的不同厚度的平板型。通过在滤波器壳体或其它辐射噪声的设备壳体上覆盖此材料,可以减少辐射能量。

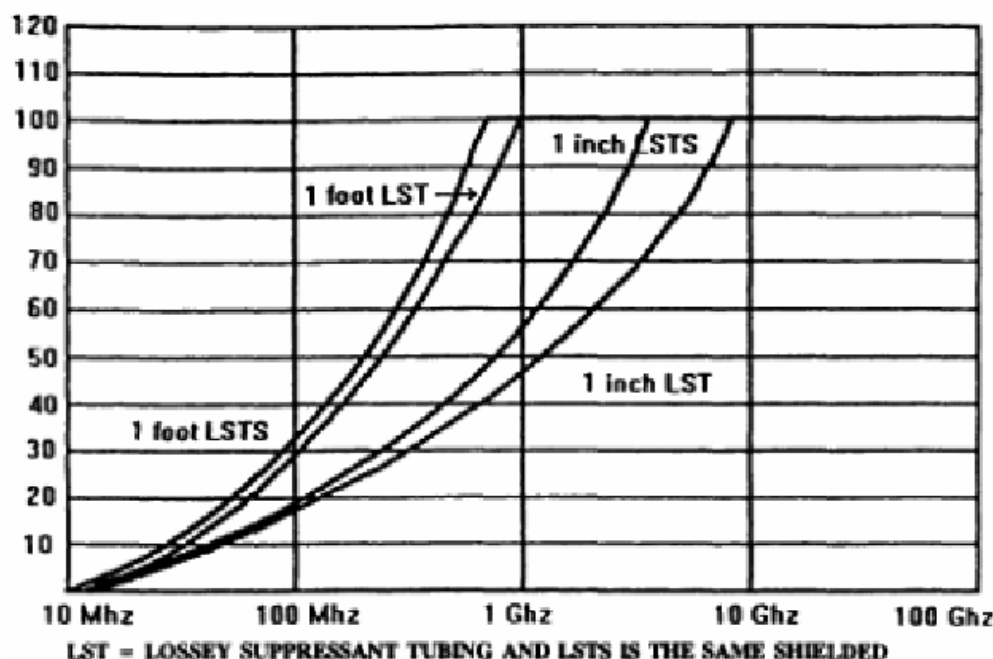


Figure 4.4 Capcon lossey tubing and sheets.

Figure 4.4 shows, the approximate loss of shielded lossey suppressant tubing and the same tubing without shielding measured per Military Standard 220A.

图 4.4 表示的是,按军标 220A 测试的,有和无屏蔽损耗抑制管的近似损耗。

§4.5 Radiation Emissions

§4.5 辐射发射

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

此书的目的是设计抑制传导发射的滤波器的。但我们知道,一般情况下如果传导发射被减弱,那么辐射发射也会降低。因此,在传导发射几乎都被消除的情况下,系统也同时具有很小的辐射发射剩余量。相反,如果传导发射没有消除,那么系统就会有很大辐射。但是如果被测设备的外壳不是导体的话,那么尽管传导发射大幅度衰减,而辐射发射还是会很有问题的。如果外壳是塑料的,那么可以通过导电喷涂的方法将外壳涂一次导电材料,这样会减小辐射。

EMCARE

5

Typical EMI Filters: Pros and Cons

典型的滤波器：问题和结论

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

我们来讨论一下现今常用的典型的 EMI 滤波器。这些滤波器的单元电路主要是 p、T、带考尔电路的 L 型、R-C 傍路，有时也包括损耗型。一般一个滤波器会包含 2 个或 3 个有时甚至是 4 个基本滤波器单元电路。每种 EMI 滤波器都有不同的优缺点，并且每种适用的工作条件也不同。

§5.1 The p Filter

§5.1 p 型滤波器

We start with a discussion of the p filter (Figure 5.1 and 5.2). This filter looks very good under the Military Standard 220A, 500-test specification (5.3). The low frequencies still work in the 500 load and line impedances. If this is the only test requirement specified by the customer, then the p tests well using this test technique. The p filter will be passed with flying colors by the source inspector (if any). This is especially true for the three-phase type, for which the specification states: "measure one phase with the other two phases tied to ground." The input and output capacitors of each p 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.

现在我们开始讨论 p 型滤波器（图 5.1 和 5.2）。在军标 220A/500（图 5.3）的测试方法下，p 型滤波器的测试性能十分好。并且只要在 500 的源和负载条件下即使频率很低的情况下，p 型滤波器仍然能很好工作。当然如果这是客户唯一指定的测试要求，那么 p 型滤波器就会得到比较好的测试结果。同样对于三相滤波器也是如此（标准规定：将其中两相接地，测另外一相，这样 p 电路单元的输入和输出电容器的容量会加倍）。因此在给定的重量和体积条件下，对于 p 型滤波器设计能满足规范要求的插入损耗就比较容易，也就是说，p 型滤波器可以以很小的元件参数、尺寸和重量满足噪声衰减的需求。

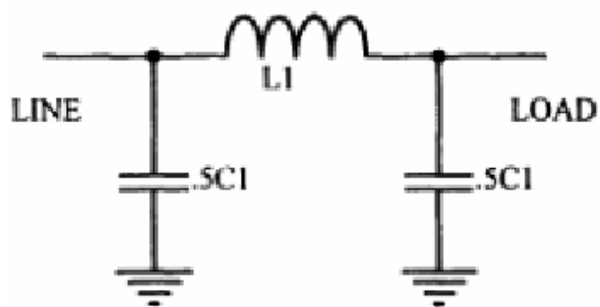


Figure 5.1 The π filter.

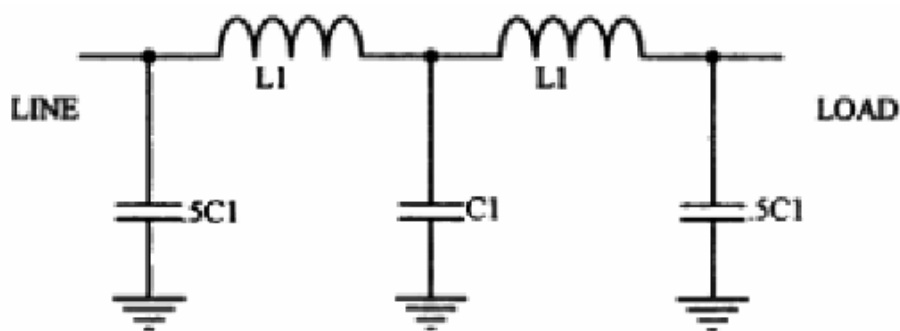


Figure 5.2 The multiple π .

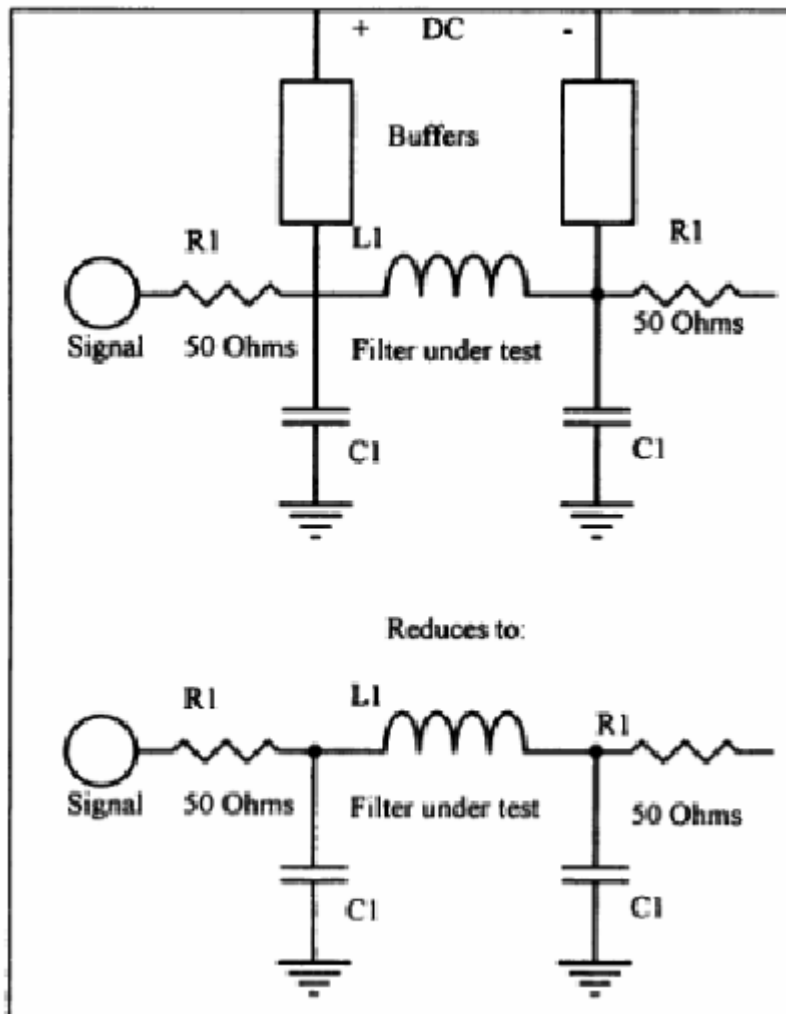


Figure 5.3 Military Standard 220A test setup.

Note that the center capacitor is twice the size of the end capacitors. All inboard capacitors are twice either of the two end capacitors for all multiple p filters.

注意，中间的电容是端口电容容量的二倍。在多级 p 型滤波器中，所有在中间的电容都是两端电容容量的二倍。

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 full peak current. There are two sets of 50 Ohm matching pads, from 6 to 10dB each, that are omitted from these figures.

测试中缓冲器或扩展缓冲器将峰值电流施加在滤波器上。因此在此测试系统中需用值很大的电感来隔离 EMI 测试信号，以免对 DC 电源产生干扰。其中，为滤波器加载峰值电流是为了通过测试确保滤波器电感在满负载的峰值电流下不会饱和。另外该测试系统还有 2 套 50 Ohm 的匹配连接器，每个有 6~10dB 的损耗，但是图中没有画出。

On the other hand, this type of filter looks bad under the Military Standard 461 or 461 naval tests (Figure 5.4). This is where a 10uF capacitor shunted from line

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

另外，这种滤波器在军标 461 或海军 461 测试（图 5.4）中，表现较差。这是因为此测试方法使用了一个 10uF 的线间或对地电容，结果是除非面向 10uF 电容器的滤波器电容量级远大于 10uF，否则它对电路几乎不起作用。但如果此电容容量量级远大于 10uF，那么对于滤波器来讲这个电容值就太高了。并且大电容的自谐振频（SRF）率会很低，由于低的 SRF 会被高品质的小电容所旁路，结果滤波器性能还是比较差。该测试方法是通过接收机或频谱分析仪在两个电流探头间内外切换来测定滤波器损耗。

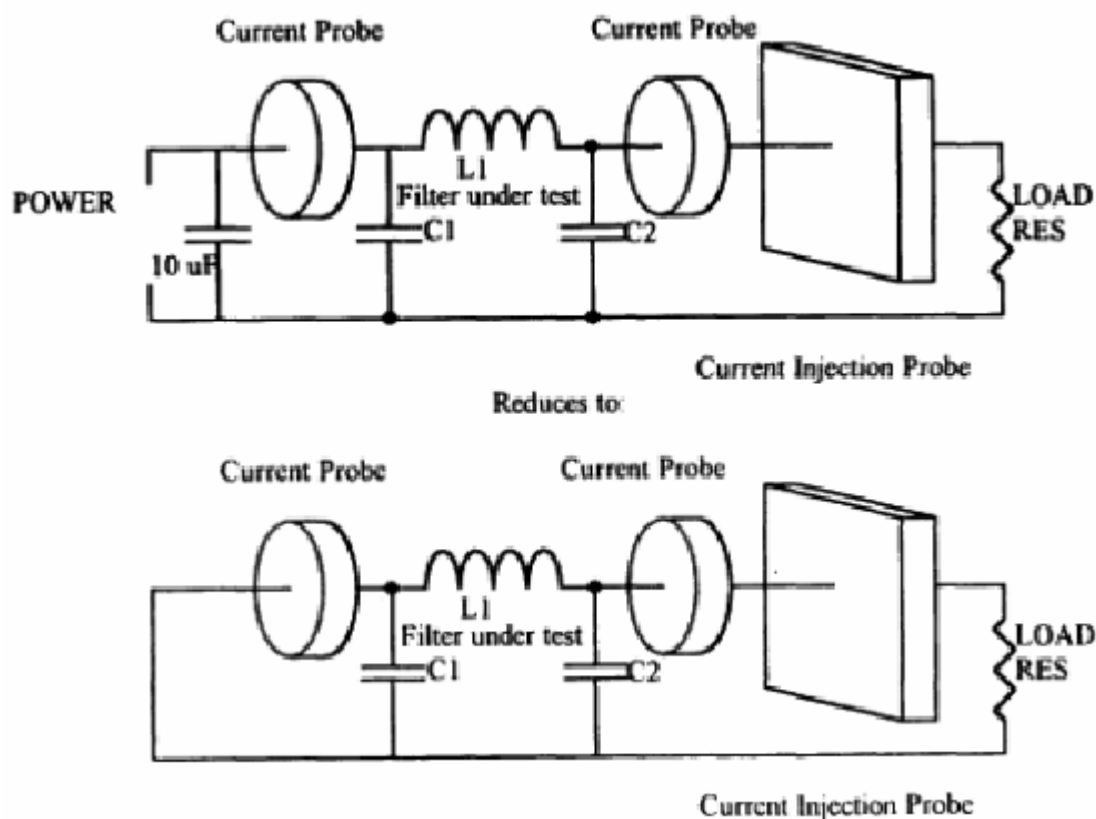


Figure 5.4 The current injection probe test setup.

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

如上所述，在 461 标准和海军测试中，三元件的 p 型滤波器性能在较低频率下看上去就如同两元件的 L 型滤波器。一般情况下每个元件的损耗约为 6dB/倍频，所以在 461 测试方法中 p 型滤波器的 18dB/倍频的损耗就变成了 12dB/倍频。实际上，单 p 型滤波器为 18dB/

倍频，双 p 型滤波器为 30dB/倍频。

The p 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 switchers or cause excessive voltage drop. This assumes that the switcher circuit has not handled this problem. The p 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 5.5). This changes the shape from the p to a square or box shape. It does happen that the system must pass a specification after several prototypes are finished. A qual. test may have to be passed (Figure 5.6). Now the earlier p 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.

p 型滤波器在一些 DC 系统中具有很好的性能，这是因为如果开关频率足够高，那么由于面对负载的是电容器，所以不会导致过分的压降，也不会使开关电源产生“饥饿”现象。当然此处的假设是，开关电源电路还没有处理这个问题（没有加储能电容—译者注）。我们将取值为原 p 型滤波器电感值一半的两个电感分别放置在火线和零线上，这样就实现了平衡线路的电路设计（图 5.5）。此电路更改将 p 型变成了方型或盒型。当电子产品设计者设计了几个原型系统后，就需要进行品质测试了（图 5.6 此处品质测试主要是指传导发射测试）。可能会出现这样的情况，p 型滤波器虽通过了所有的插损测试，但是在品质测试中却效果不佳。这是因为，滤波器前端或电源线侧的电容在低频下失去作用。或许你会问，为什么在 EMI 实验室能通过实验，而实际的传导测试滤波器却出现问题？实际上 220A 和 461 规范并不能真实反映实际规范（500 LISN 传导发射测试规范）的情况。需要提醒的是，220A 和 461 规范相比，461 规范更接近实际情况。

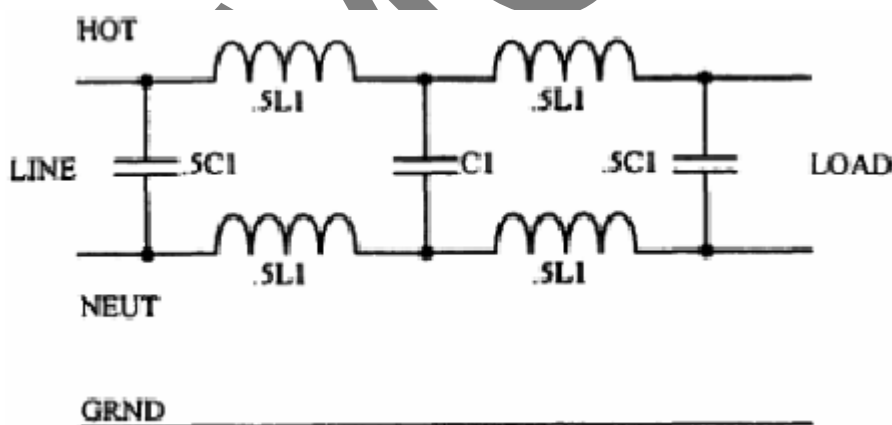


Figure 5.5 The multiple balanced p.

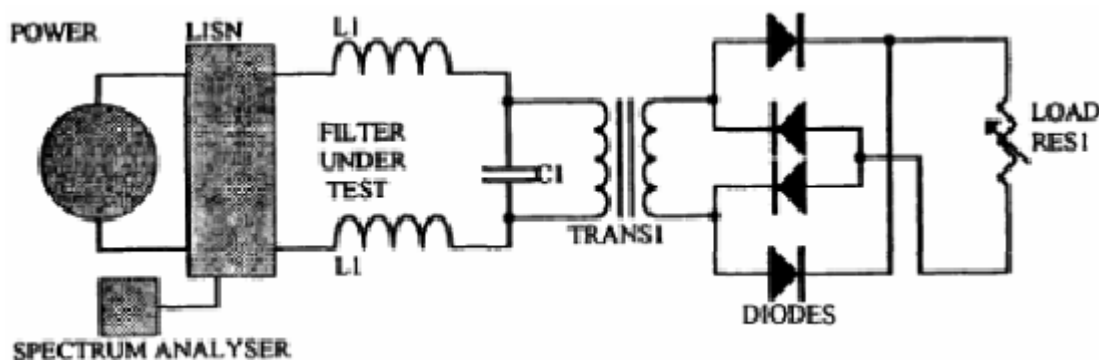


Figure 5.6 The system or diode test method.

§5.2 T Filter

§5.2 T 型滤波器

Ts are discussed next. These give 18dB per octave, and the double T gives 30dB per octave (6dB per element; Figure 5.7). T filters work best in low-impedance lines (high current). The line impedance is very low up to at least 100kHz, but the 461 loss specifications start at either 10 or 14kHz. 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 loads 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. I have seen 115V ac 60Hz lines as high as 132V feeding a light load requirement. This happened because the resonant rise occurred at a very low frequency as a result of higher inductive values.

下面我们来谈一下 T 型滤波器。T 型滤波器可以有 18dB/倍频的衰减，双 T 型会有 30dB/倍频的衰减（每个元件为 6dB，见图 5.7）。T 型滤波器在低阻抗线路中表现性能良好。通常线路阻抗在 100kHz 频率以下的是非常低的。但是需要注意到，461 插入损耗测试规范所规定的测试起始频率是 10 或 14kHz，此时 T 型滤波器的电感阻抗会叠加在低的线路阻抗上，这就使得电容器工作在一个有一定阻抗的环境中。虽然 T 型滤波器的电感值没有设计得很大，T 型滤波器在高电流负载情况下性能还是会十分优良。并且太高的电感值会导致输入到负载的电压升高或下降。我曾经见过 115V AC 60Hz 的电源输入到负载就变得高达 132V，产生这种问题的原因是，高电感值使得在很低的频率点产生了由于谐振引起的电压升高现象。

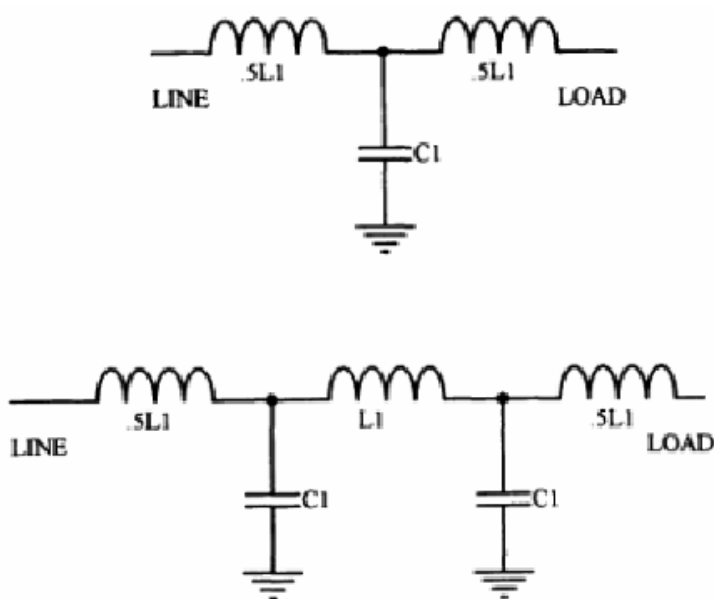


Figure 5.7 The T and multiple T.

The T should never be used in a dc system if the load utilizes mainly switcher because the high impedance of the output inductor facing the load will starve the switchers. The switcher designer may have taken this into account by lowering the impedance with a capacitor at the switch input. This really makes the filter into a maladjusted double L because the capacitor shunts the inductor facing the load. Note in Figure 5.7 that the central inductor is twice the size of either of the two end inductors. All inboard inductors are twice the size of either of the end inductors.

由于 T 型滤波器面对负载的输出端电感是高阻抗，所以会导致开关电源产生“饥饿”现象，因此 T 型滤波器不能应用在以开关电源为主的 DC 系统中。当然，开关电源的设计者可能已经考虑到这种情况，所以常常会在开关电源的输入端加一个低阻抗的电容。但这又会把滤波器变成不适宜的双 L 型滤波器，因为电容会旁路面对负载的电感。注意图 5.7，中间的电感值是两端的电感值的二倍。并且对多极 T 型滤波器来讲，所有中间的电感都是两端电感的二倍。

The 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 5.8).

对 T 型滤波器同样可以进行平衡电路设计，把原电感器值的一半加在零线上与火线对称的位置，就形成了 H 形电路。此电路就是 T 形滤波器的平衡电路（图 5.8）。

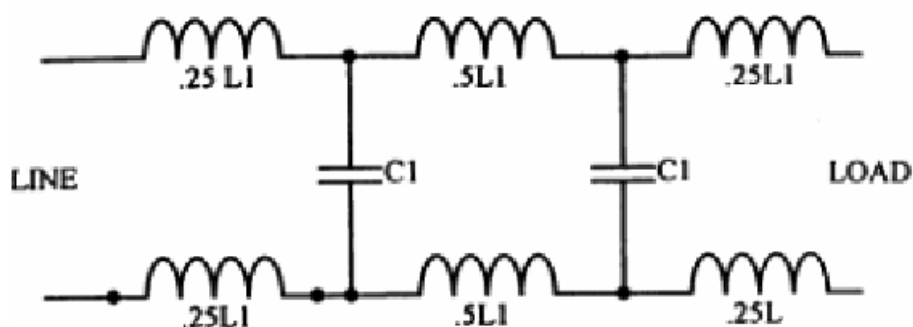


Figure 5.8 The multiple balanced T filter.

§5.3 L Filter

§5.3 L 型滤波器

The L type is the most often used filter. Both the p and the T are three element, or more if multiples are used, and should give 18dB loss per octave. The L, because it is only two elements, provides 12dB loss per octave. All these loss figures refer to the loss starting above the cutoff frequency. A single L often works best in the dc mode if the load has switchers, because a large L would face the dc supply and the large capacitor (of high quality) would provide a low impedance for the switcher frequency (Figure 5.9). 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 (See Section 16.3.). 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 of supply voltage. The L and multiple L work well in higher power applications (Figure 5.10). Again, to balance, split the inductors and place the other half in the neutral leg (Figure 5.11). The double L has 24dB loss per octave.

L 型滤波器（图 5.9）是最常使用的滤波器。p 型和 T 型滤波器都是三个元件的滤波器（如果是多级型则元件还要多些），其损耗大约是 18dB/倍频。对于 L 型滤波器，因为它只有两个元件，所以损耗只有 12dB/倍频。以上所说的损耗曲线是基于截止频率之上的频率的损耗而言的。因为大电感面向 DC 电源而面向负载的大电容（高品质）在开关频率上是一个低阻抗，所以单 L 型滤波器最适合在负载是开关电源的 DC 模式下工作。但为什么不是双 L 型？两个电感（感值为单个的一半）的损耗之和比单个电感的小。同样对电容器也是这样（见 16.3 节）。小于原来电容值 0.5 倍的输出端口的电容，其储存的能量不能满足开关电源的需求，结果就会在开关电源上产生大的峰—峰电压。如果此多级 L 型滤波器连接在 B 类或 Royer 放大器的中间抽头上，那么这个峰—峰电压的频率会是开关频率或是 2 倍的开关频率（开关频率越高，电压降就越低）。在电压降没有超标或是开关频率足够高的条件下，可以使用双 L 型滤波器。同样对纹波的情况也是如此，但在 DC 电源电压的情况下会好些。L 型和多级 L 型滤波器（如图 5.10）比较适合使用在大功率的设备上。同样，为了设计平衡的电路，可以把电感均分在火、零线上（如图 5.11）。另外需提醒的是，双 L 型滤波器具有 24dB/倍频的损耗。

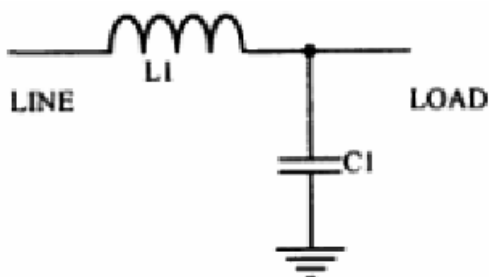


Figure 5.9 Single L filter.

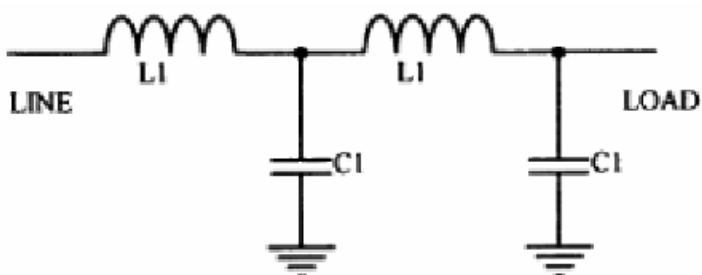


Figure 5.10 Multiple L filter.

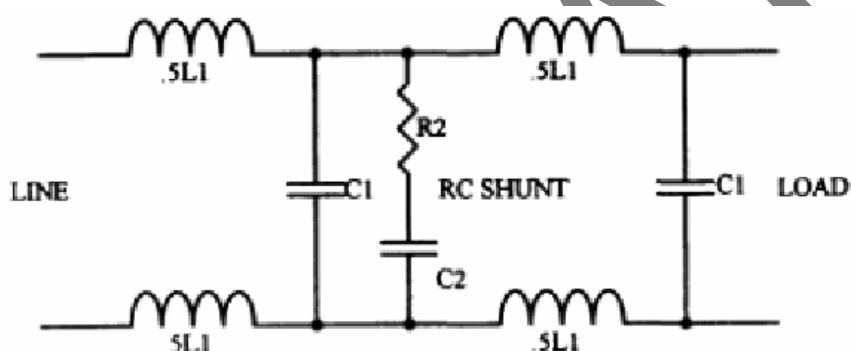


Figure 5.11 Balanced multiple L with R-C shunt filter.

§5.4 Typical Commercial Filter

§5.4 典型的商用滤波器

This is the type used in test equipment, computers, and other commercial electronic equipment. The manufacturer must pass tests for VDE, CSA, and the U.S. Federal Communications Commission (FCC), and the tests are conducted by EMI test houses that help the manufacturer with all the documentation. These filters are the balanced π type often purchased from outside suppliers and are often built offshore. These filter are mainly common mode in appearance, with a capacitor across the input and output from hot to neutral and two other capacitors to ground that must meet leakage current specifications (Figures 5.12 and 5.13). The leakage inductance is often made high by adding washers to the center of the pot cores separating the two windings. This is also accomplished by winding the two windings as far apart as possible on a toroid core. The feedthrough capacitors (CF) are case grounded (Figure 5.12). Some

of these techniques add a differential mode to this common mode filter. This then makes a balanced p type of differential mode. This all works because the losses specified by the FCC start at 450kHz. The inductors and capacitors can be quite small to accomplish these tasks. Another point to remember is that the current through the leakage inductance can saturate this inductor. This is because the common mode inductor is wound on ferrite core with high A_L values. Some of these filters do not use the feedthrough type of capacitor, that the circuit changes to that show in Figure 5.13. These capacitors are less expensive, but the self resonant frequency is lowered by the added lead length.

这是一类常常使用在测试设备、计算机和其它商用电子设备上的滤波器。制造商必须通过 VDE、CSA 和美国联邦通信委员会 (FCC) 的测试认证, 还得具有 EMI 测试实验室出具的所有应有的测试实验数据文件。这些商用滤波器通常是平衡 p 型电路, 并且基本都在海外生产。此滤波器在输入和输出端都有线间电容, 另外还有两个对地电容, 而这两个对地电容是需要满足漏电流规范的 (见图 5.12 和 5.13)。通常在罐型磁芯上加隔离绕组的垫片会大大增加漏感, 同样对于环型磁芯, 绕组彼此尽量远离也会增加漏感。图 5.12 中的穿心电容 (CF) 是接在机壳地上的。通过这些设计, 会使共模滤波器增加了差模功能。通过 Zorro 电感的漏感形成了平衡的 p 型差模电路。因为 FCC 插入损耗测试规范是从 450kHz 的频率点开始的, 所以上面提到的方法都会比较有效。其中, 电感和电容仅以很小容量就能达到性能要求。另外, 须注意因为绕在铁氧体磁芯上的共模电感具有很高的 A_L , 所以通过漏感的电流会使电感饱和。还有一些这种类型的滤波器并不使用穿心电容, 此时电路就会变成图 5.13 所示的电路。这些电容较廉价, 但是引脚长度会降低自谐振频率。

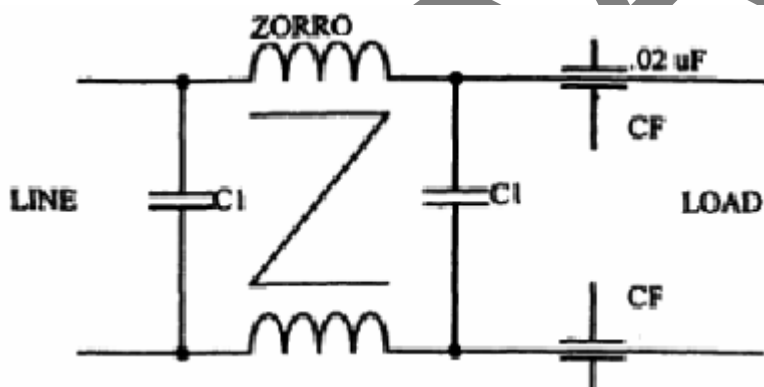


Figure 5.12 Commercial filter with Z and feedthrough.

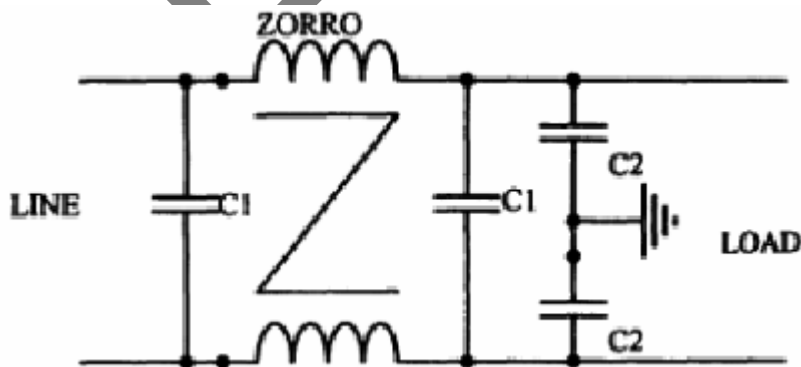


Figure 5.13 Commercial filter with capacitors to ground.

§5.5 Dissipative Filter

§5.5 损耗型滤波器

This filter is rarely seen in the EMI area today. It consists of one inductor and one capacitor along with two resistors (Figure 5.14). The two resistor 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 filters are similar to the line impedance stability net works (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 points of the four resistors (Figure 5.15).

现今在 EMI 领域已经很少能看到损耗型滤波器了。这种滤波器一般由一个电感、一个电容和两个电阻组成（如图 5.14）。这两个电阻分别串联在电感的两端，而电容则连接在两个电阻连接处和大地之间（或其它的线，如相线或零线—译者注）。这类滤波器与 LISN 的阻抗很接近。同样，通过等分电感和电阻到两条电源线上可以实现平衡电路设计（如图 5.15）。

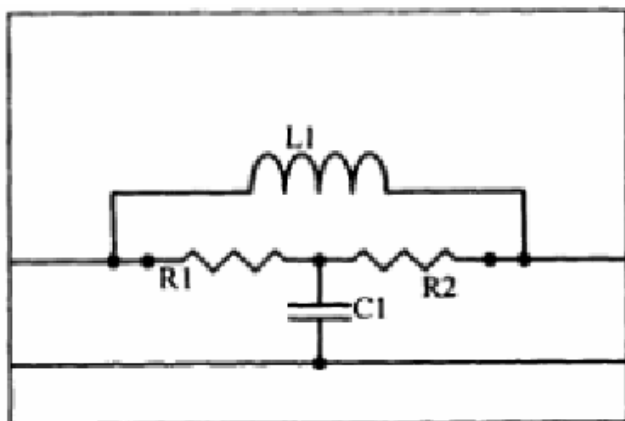


Figure 5.14 Dissipative filter.

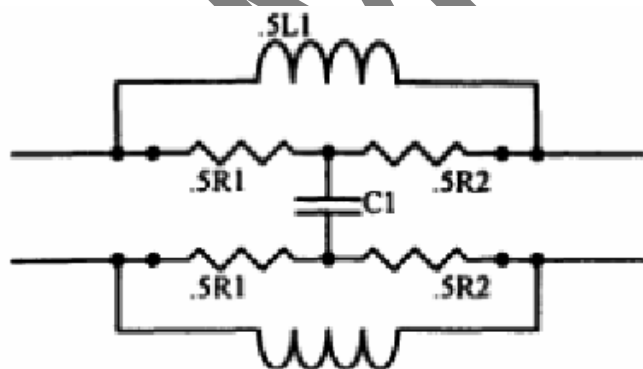


Figure 5.15 Balanced dissipative filter.

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

此类滤波器主要的缺点是虽在理论上能达到 12dB/倍频的损耗，而实际只有 6dB/倍频。另

一个缺点是，仅仅因为这节电路的 6dB 损耗就要增加整个滤波器的体积和成本。其中电阻应是无感的，为了能适合消除电磁脉冲，此电阻至少要有 2W 以上的功率限值。

The advantages are as follows:

损耗型滤波器的优点如下：

1. This filter can be used to match impedance by using two of these at each end (Figure 5.16). Any filter can be used between these two end units and designed to match the impedance of the inbound resistor values. This can work well to meet the 220A specification using four 50Ω resistors and designing the middle unit for 12dB less loss.

1. 通过在每个端口（输入和输出）使用损耗型滤波器（如图 5.16）可以达到阻抗匹配的目的。在这两个滤波器之间可以使用任何与损耗滤波器内部电阻值阻抗匹配的滤波器，通过使用 4 个 50Ω 的电阻，并且位于此二个滤波器之间的元件可以设计成较小的 12dB 损耗，这样还可以很好地满足 220A 规范的要求。

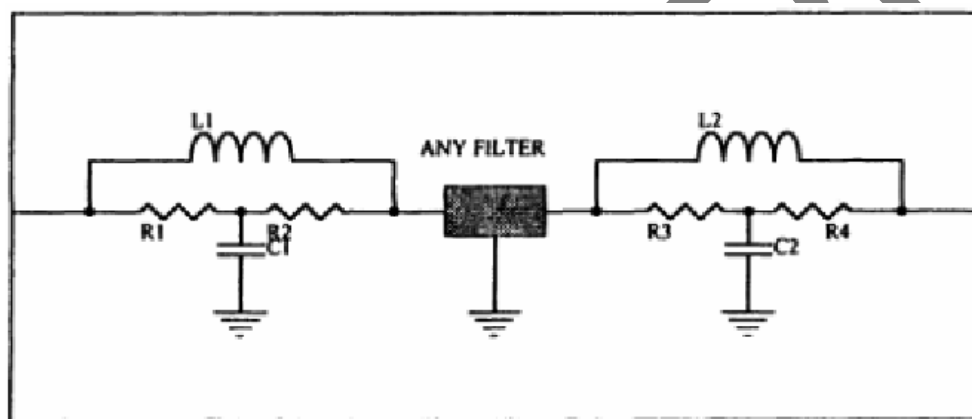


Figure 5.16 DfN and DOUT dissipative filters for both ends.

2. 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.

2. 此滤波器能消耗阻带能量。如果高频噪声通过电源线，那么大多数的能量会被滤波器电源线侧的电阻消耗，负载侧的电阻则用来损耗负载的开关噪声或其它噪声。如果位于各个损耗型滤波器中间的那个滤波器反射噪声，那么与之相对的损耗滤波器的电阻也同样会损耗这部分被反射的能量。

3. 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 5.16.

3. 损耗型滤波器非常类似于廉价的 LISN，如图 5.16 所表示的电路中就有 2 个这样的滤波器。

§5.6 Cauer Filter

§5.6 考尔滤波器

The Cauer, or elliptical, filter is best used in very low impedance circuits (Figure 5.17), usually with multiple L s and T s, in any case, normally a capacitor is shunted across the center inductor. This is used to fix a problem frequency, such as 14kHz. 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 100 and is often the value of the design impedance of the filter. The design impedance is the lowest RMS line voltage divided by the highest RMS line current. If this filter is to be balanced, use half the calculated value of inductance on the two lines, and parallel each inductor with two capacitor resistor networks. The value of the two capacitor networks uses twice the capacitance and half the resistor values, and these would, again, be tied across the two inductors.

在阻抗非常低的电路中最适合使用考尔滤波器（如图 5.17）或椭圆滤波器。通常使用考尔滤波器的情况是，在多级 L 型和 T 型滤波器中，在电路中间的电感上并联一个电容。这种处理方法常常是用来改善某个产生问题的频率点，如 14kHz。但滤波器整个网络特性会在此频率点上有些变化。虽然使用考尔滤波器把位于问题频率点的问题解决了，但是在远高于问题频率时，这个中心的电路小节（考尔滤波器）就等同于不存在了，此时高频会无阻碍通过此节电路。在考尔滤波器中通常会有一个电阻与电容串联，此电阻可以控制旁路的量，其值大约为 100，这通常也是此滤波器的设计阻抗。用最低的线电压 RMS 值除以最高的线电流 RMS 值就是设计阻抗。如果需要是平衡电路结构，那么就分别在两条电源线上串联原电感值一半值的电感，并在每个电感上各自并联电阻与电容串联的电路，其中电容取原值二倍，电阻取原值一半。

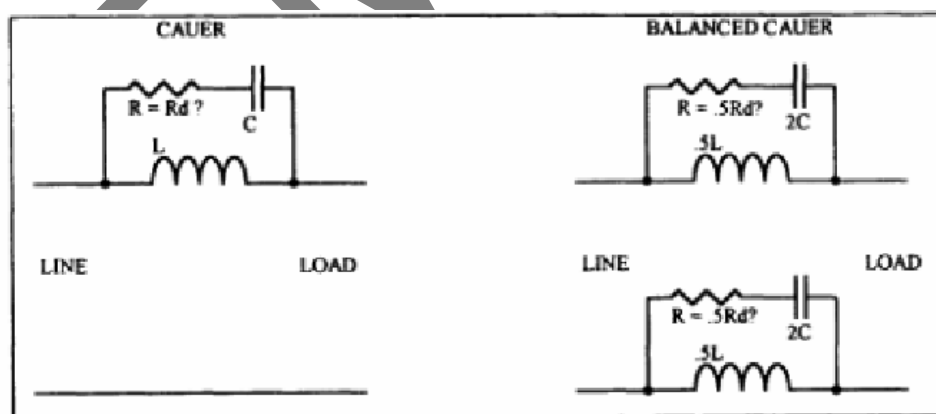


Figure 5.17 Cauer and balanced Cauer.

§5.7 The R-C Shunt

§5.7 R-C 傍路

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 balanced across the line to start with, if required (Figure 5.18). 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, π , or T. Usually the number of resonant rises is 1 less than the multiple number meaning that the single L, π , 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. Find the frequency of the lowest resonant rise, and choose the capacitor value 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 10kHz); choose the capacitor to equal the design impedance at the trouble frequency.

另一种滤波器电路比考尔滤波器更常用，此电路多应用在高阻抗、低电流的电路中。这种滤波器被称作 R-C 旁路（如图 5.18），可以看到 R-C 旁路只使用很少的元件并能自动平衡线路。R-C 旁路由电容串电阻组成的。通常设计滤波器时需要其谐振增益频率低于问题频率，如果整个滤波器是两级或三级 L、 π 或 T 型的多级滤波器，那么就更需要满足此条件了。一般情况下，谐振增益频点数为滤波器级数减 1，这意味着单 L、 π 或 T 型滤波器没有谐振增益，但如果是四级的滤波器，那么谐振增益频点就是 $4-1$ ，即 3。同样如果电路的 Q 值很低也会没有谐振增益，只有高 Q 值的电路网络才会有谐振增益。在设计滤波器时，需要找到最低的谐振增益频率点，然后选择在此频率下阻抗等于滤波器设计阻抗的 R-C 旁路的电容，这个 R-C 旁路会衰减所有的谐振增益和问题频率点上的噪声。如果没有涉及到谐振增益频率（远大于电源 15 次谐波频率或远小于 10kHz 的情况），那就选择在问题频率点电容阻抗等于滤波器设计阻抗的电容容量。

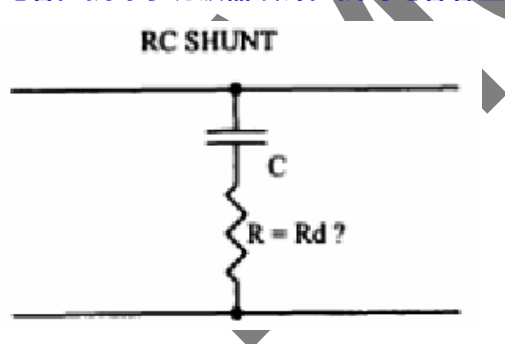


Figure 5.18 Balanced and unbalanced R-C shunt.

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, 100V is the lowest line voltage at 10A. This is the highest current at this lowest line voltage. The resonant rise frequency is 4kHz. Then,

具体示例如下，设计阻抗是将最低线电压除以负载所需要的最高电流，在本例中设电流为 10A 最低的线电压为 100V，这个电流是在此最低的线电压下最高的电流值，滤波器的谐振增益频率为 4kHz，那么，

$$R_d = \frac{100}{10} = 10\Omega \quad C = \frac{1}{2\pi \cdot 4000 \times 10} = 3.979\text{mF} \quad (5.1)$$

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

计算出的电容容量为 3.979uF 或近似为 4uF，将它与 10O 电阻串联接在线间，则此 R-C 傍路将消除 4kHz 的增益并衰减 14kHz 的问题频率点的噪声。如果谐振增益频点处在与设计无关的区域，那么就将计算电容的频率换成问题频率 14kHz，结果在此频率下需要的电容容量为：

$$C = \frac{1}{2\pi \times 14,000 \times 10} = 1.137\text{mF} \quad (5.2)$$

The 10O resistor is in series with a 1.2uF capacitor tied across the filter. This lends itself to multiple L filters 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 impedance.

那么 10O 电阻与 1.2uF 电容串联后接在线间，这将使 R-C 傍路电路参与到具有并联电容的多级 L 型滤波器的电路网络中。R-C 傍路离负载越近，则其越倾向于减少负载的阻抗波动。这是因为，负载为高阻抗而 R-C 傍路在高阻抗条件下工作性能是比较好的。

§5.8 Conventional Filters

§5.8 常规滤波器

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 just discussed (Figure 5.19) because the noninductive resistors would give the proper match needed by these filters, especially if the cutoff frequency of the dissipative filter is about half that of the regular filter.

滤波器设计室很少使用常规滤波器。虽然有时会使用这些滤波器，但是这种滤波器要求输入输出的阻抗是常数才能正常工作，尤其是军用领域需要的低频损耗性能，不幸的是电源和负载阻抗却不能在低频下保持稳定的阻抗，这就限制了常规滤波器的使用。如果这种滤波器安装在两个损耗型滤波器（在 5.5 节刚刚述过，见图 5.19）之间，那么它的性能会很好，因为无感电阻会为常规滤波器提供它所需要的阻抗匹配条件，尤其是当损耗型滤波器的截止频率是常规滤波器的一半时效果会更好。

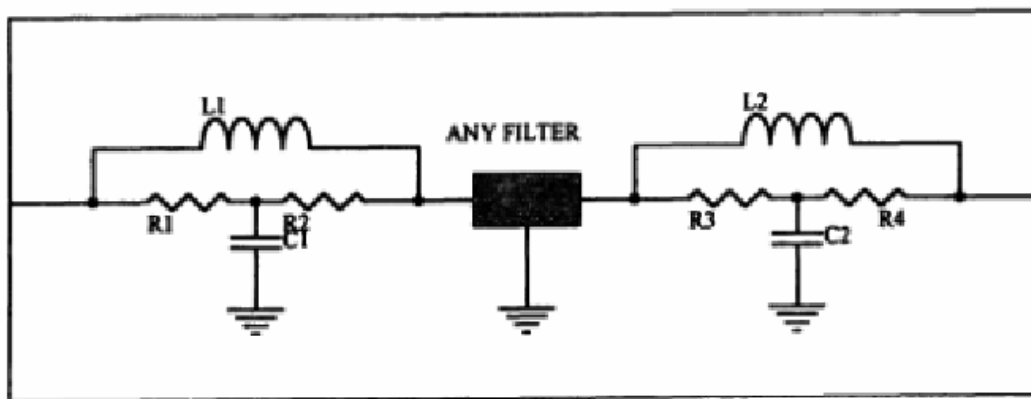


Figure 5.19 Conventional filter used with DIN and DOUT.

§5.9 Filter Matrix for Line and Load Conditions

§5.9 不同源和负载条件下的滤波器列表矩阵

This section covers the filter matrix to use for both the line and load conditions. This is mainly empirical, that these should be regarded as general guidelines. 本节包括对于不同源和负载条件下所使用的滤波器列表矩阵。这些内容主要都是经验性的，我们也可以把这部分内容当作一般的应用指南。

The first is the line side covering the dc to dc, single phase, and three phase (Figure 5.20). The other dimension is FCC, VDE, 220A, 461, and the full system specification. The next part of the matrix is for the load side (Figure 5.21). This is for 60 and 400Hz. The other row is the off-line regulator, with inductor, the power factor correction circuit, and the power line filter. These are general recommendations only.

第一行是源端条件，包括 dc to dc、单相、三相（如图表 5.20）。第一列是规范条件：FCC、VDE、220A、461 和整个系统的规范，下一图表（图表 5.21）的矩阵是针对不同负载情况的列表。第二列分别对应 60 和 400Hz 的源条件，第一列是负载情况：带离线调节器、带电感情况、带功率因数校正电路和带电源线滤波器。这些图表都仅仅是一般的建议用法。

	DC TO DC	SINGLE PHASE	THREE PHASE
FCC	SINGLE L PI ? L	PI, T, L	PI, T, L
VDE	SINGLE L PI ? L	PI, T, L	PI, T, L
220 A	SINGLE L PI ? L	PI, T, L	PI, T, L
461	SINGLE L L	L, T	L, T
SYSTEM	SINGLE L	L, T	L, T

Figure 5.20 Line-side matrix.

OFF LINE REG	INDUCTOR MUST BE DESIGNED TO HANDLE I_{peak} CORE GAPPED ?	SAME AS LEFT. DONT USE POWDERED IRON. $\mu < 125$
OFF LINE WITH INDUCTOR	INDUCTOR TO HANDE I_{rms} . CORE GAPPED ?	SAME AS ABOVE AND LEFT. CAP MARGINS GREATER. WATCH AMPS
PFCC INPUT	Z_o OF FILTER $\ll I_{pk}$ ON LOAD IMPEDANCE. MUST BE CAP OUTPUT.	SAME AS ABOVE SAME AS LEFT
POWER LINE FILTER	SHOULD BE DESIGNED FOR I_{pk} NOT RMS	SAME AS ABOVE SAME AS LEFT

Figure 5.21 Load-side matrix.

6

Differential Mode Components

差模元件

The differential mode components must be of the high-Q type. The individual component Q must be high, and the circuit Q must be lowered to the point at which the filter does not oscillate.

差模元件很可能是高 Q 值的类型。单个元件的 Q 值也很可能会很高，但实际上电路的 Q 值却要求低到滤波器不能振荡。

§6.1 Capacitor Construction and Self-resonant Frequency

§6.1 电容器构造和自谐振频率

Figures 6.1 and 6.2 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 equations to calculate the number of wraps to obtain the capacitance needed for the extended foil type. The design is covered in Section. (Figures 6.1 and 6.2 provided by Robert Hassett of RFI Corp., Bay Shore, New York 11706).

图 6.1 和 6.2 显示的是聚酯（迈拉）电容和陶瓷电容的自谐振频率。两者都是穿心电容，所以自谐振频率都比较高。接下来的内容将介绍延伸金属箔型电容的构成和电容所需缠绕圈数与电容量之间关系的简单计算公式。详细的设计内容也涵盖在本节中了。（图 6.1 和 6.2 由 RFI 公司的 Robert Hassett 提供，地址：海岸，纽约 11706）。

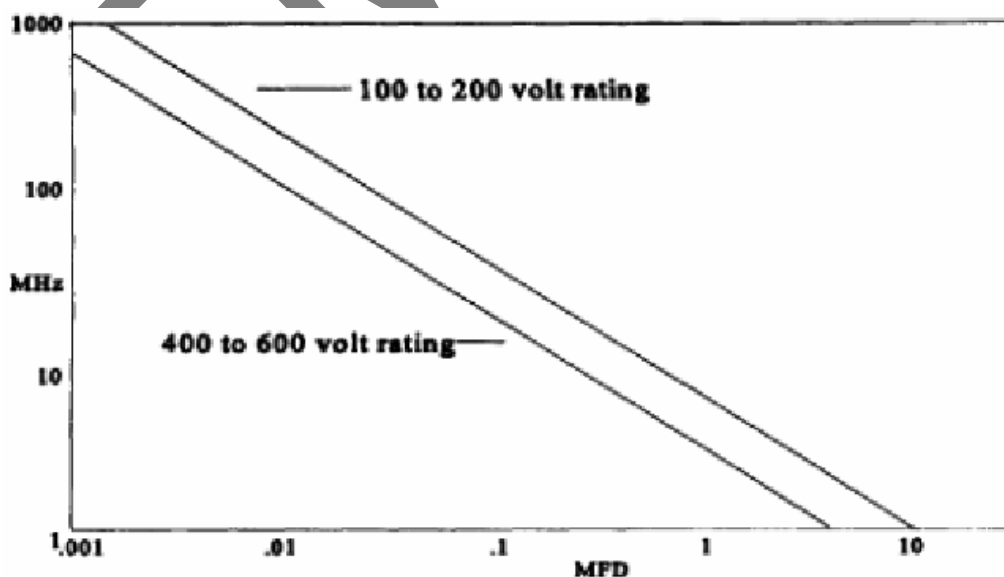


Figure 6.1 Mylar self-resonant frequency chart.

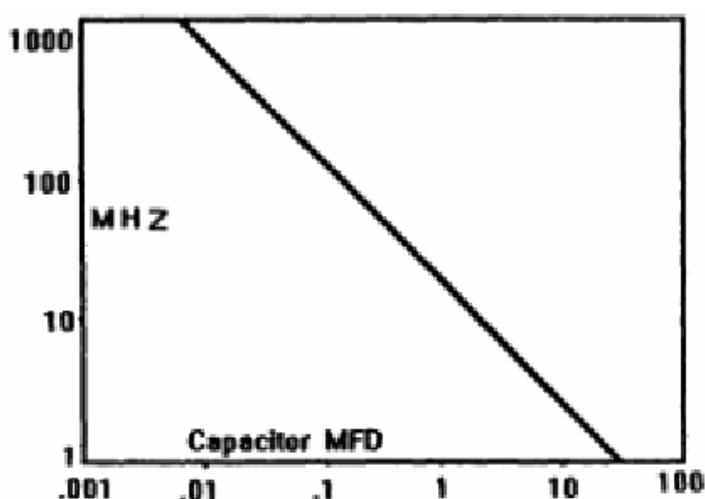


Figure 6.2 Ceramic self-resonant frequency chart.

The advantage of the feedthrough 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. Figures 6.1 and 6.2 show SRFs to 1GHz 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 increase the cost. As the line frequency increase, the line frequency harmonic current increase. 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 skimpy. Many filter builders use the large can type of nonpolarized filter capacitors. Some of these are oil impregnated.

穿心型电容的优点是引脚短，所以具有较低的 ESR 和 ESL（等效串联电阻和电感），这就意味着穿心电容的自谐振频率较高。图 6.1 和 6.2 给出了小陶瓷电容 1GHz 的 SRF。使用穿心电容的问题是，电源线 RMS 电压愈高，为消除爬电和电晕现象所需要留出的余量就越大，这样就导致穿心电容体积增加并同时增加了成本。而当电源线频率升高时，电源线上谐波电流也随之增加，这样由于 dV/dt 的原因也增加了电容的损耗。如果实际取的余量接近较低的限值，那么当电源线频率增加时，原来的余量就需要相应地增加。换句话说，就是除非最初低频条件下余量不是很充裕，否则余量就不是频率的函数。实际设计中，时常有一些滤波器制造者选用大容量桶型封装的无极性滤波器电容，而这些电容多是油浸的。

These oil-filled and large-can type capacitors may be very good for power supplies and other applications requiring nonpolarized capacitors, but watch out for their self-resonant frequencies. Most of these types have very low SRFs of the order of 20kHz. This is when old power supply trick may be handy— "Yeolde paralleling capacitor trick"— which calls for paralleling the capacitor with another capacitor 0.05 the value of the original. This smaller capacitor should have a

much higher SRF and lower ESR and ESL. A recent article in the Institute of Electrical and Electronics Engineers magnetic manuals showed that the gain of only 6dB 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 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. I believe that these large capacitors should not be used. Most of the design techniques in this book should not call for this large a capacitor value. Another trick to lower the ESL and raise the SRF of the capacitor is as follows. In Figure 6.3, 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 time, the new phrase for this technique is "veeing the capacitor" – the old "keep the leads short as possible" trick – especially for the capacitor. The leads on the inductor add only a small amount of inductance to the inductor.

这些油浸的大桶型封装的电容非常适合电源供电系统和其它需要无极性电容的应用,但是在使用中需要注意这些电容的自谐振频率。多数这种电容具有很低的自谐振频率,其量级大约为 20kHz 左右。此时就需要掌握一个解决此问题的窍门——“Yeolde 并联电容诀窍”——即在此电容上并联一个是这个电容容量 0.05 倍的电容。这个小的电容需要具有高 SRF、低 ESR 和 ESL 的特性。最近 IEEE magnetic manuals 分册上的一篇文章研究了这个方法,研究显示假设电容引脚长度相同(引脚相同意味着新并联的电容的 ESL 和 ESR 几乎与原电容相同)的情况下,这两个电容并联的增益只有 6dB。如果新增的并联电容是穿心型的,那么由于穿心型电容的低 ESR、ESL 和高 SRF 特性会使得这种组合十分有效。但是,我认为不应该使用这种大电容,本书的多数设计并不需要这种大电容值。另外还有一个降低电容的 ESL 提高 SRF 的诀窍。在图 6.3 中,有四个电感是多余的,其中电容引脚产生的两个电感增加了电容的 ESL、ESR,并降低了 SRF;另外两个是由电感的引线形成的,它们使原电感的电感值增加,但是这两个电感与原电感相比数量级很小。虽然“使引线尽可能短”是一项比较成熟的技术,但是如今在电容设计上它有了一个新的称谓“V 型电容”。

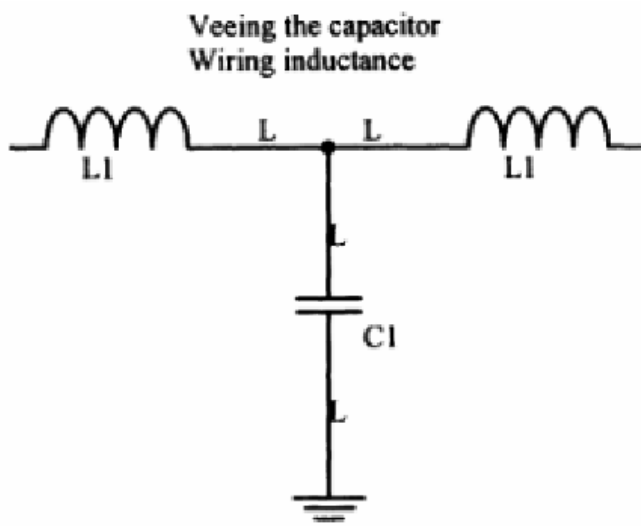


Figure 6.3 The T filter used to demonstrate veeing.

Now, in Figure 6.3, all the inductance in either leg of the vee, adds to the inductors on both sides (Figure 6.4). The capacitor ESL and ESR decrease, increasing the SRF. The same can be done to the bottom half if the circuit is balanced (Figure 6.5).

现在，在图 6.3 中电感的引脚一边弯成 V 型，并把两个这样的电感装在两侧（如图 6.4），这样电容器 ESL 和 ESR 都降低了，而 SRF 也提高了。如果电路是平衡电路，那么图 6.4 中电路的下一半也可以同样处理。

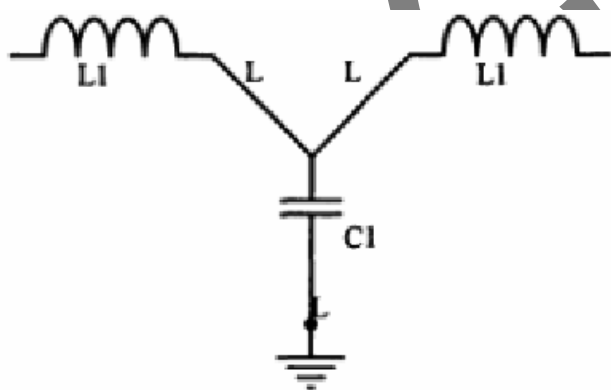


Figure 6.4 Veeing the capacitor.

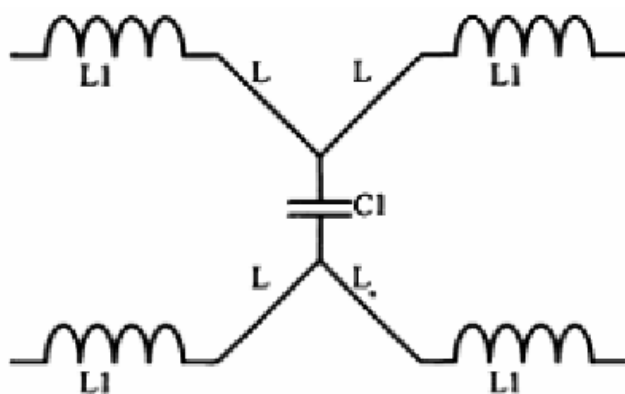


Figure 6.5 The balanced vee.

§6.2 Capacitor Design

§6.2 电容设计

§6.2.1 Wrap and Fill Type

§6.2.1 缠绕和填充的类型

There are three methods used to build these capacitors and two subgroups. 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 types for low-frequency problems. If the filter also has high-frequency problem that the R-C shunt must also aid, the quality of the capacitor dictates the higher quality of the extended foil type. The two subgroups are metallized of 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-filters in which there are high harmonic currents from off-line regulators. Most filters 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.

电容的制作方法有三种和两个子类。最初的制作方法是，在封装、捻、缠电容之前将引线焊接在金属箔片上。因为电流会流过整个金属箔片，使得全部长度的电容箔片都进行充放电，所以我们将这种电容称作感性电容。感性电容的 ESL 和 ESR 都很高，所以 SFR 是非常低

的。除了处理低频问题的 R-C 旁路电路和考尔型滤波器外，感性电容是不能在 EMI 滤波器中使用的。即使在使用 R-C 旁路电路的情况下，如果处理的是高频噪声，那么 R-C 旁路中的电容也不能使用感性电容，而是要使用延伸金属箔型。两个子类是按金属箔片的金属化处理方法来区分的。介质薄膜通过喷涂铝形成作为电容板极的金属薄膜；金属箔片类似于铝薄膜方法，但是金属箔片更厚，所以在处理离线调节器产生的高谐波电流值的应用情况下，在处理脉冲的滤波器上用金属箔片型电容性能会更好。多数滤波器电容都采用这种结构。金属化薄膜有两个优点：第一，在相同容量的电容体积很小，并具有自愈能力。所有的介质材料上都会有很小的孔，所以当给喷涂金属的介质材料加电压时，金属薄膜常常会短路并使薄膜融化。但电容的铝重组就使得电容具有自愈能力。第二个优点是，金属铝可以喷涂在介质的两侧，这样电容的体积就会很小且增强了自愈能力，并且如果总的电容电流足够小时处理脉冲的性能也会比较好。

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 this capacitor type are both very low because the current flow travels only the average 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.

EMI 滤波器使用最多的是延伸金属箔片型电容。金属箔片延伸过绕线柄轴，这样一个极板（一片金属箔片）延伸到左边而对应的极板则延伸到右边。由于左边的箔片和右边的箔片这两个平面存在无重叠区域，因而称之为延伸金属箔片型。延伸出的（无重合）部分分别用来连接电容的两个管脚。因为电流通过电容箔片的距离为电容的平均长度（**因此箔片的边缘都接在电容管脚上**—译者注），所以阻性损耗和电感都很低，因而金属箔片型电容的 ESR 和 ESL 都十分低。当然，每一层箔片都与其它层平行也会降低 ESL 和 ESR。由于 SRF 非常高，所以滤波器设计者常常使用这种电容。

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 turns, and so many degrees are added to each tab so that the tabs are uniformly spaced around the sides of the capacitor. These tabs, extending from each side of the capacitor, are folded over and soldered together to form the contact.

最后一种制作电容的方法叫做片型。这种方法有些类似于延伸金属箔片型，但不同的是它可以使用在具有大的电容电流场合。片型电容的叠片的大小和数目取决于电流大小。叠片是由薄铜片通过浸焊形成，由于电容是缠绕的，所以可以在每层插入叠片，最终形成围绕电容的对称叠片围层。延伸到电容每侧的叠片会焊在一起以形成连接。

Most new people in the EMI design arena are knowledgeable about power supply design in which the main concerns about capacitors are their working voltage

and capacitor value. Others are familiar with the derating of the capacitors, and this is quite helpful. The point is that ac capacitors must be designed to handle ac currents 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 π and multiple filters, at least the last capacitor, the capacitor closest to the load side, should be of the foil type (see the next section).

许多 EMI 设计领域的新人都了解电源设计中所需要关心的电容的主要参数：工作电压和电容量。而只有一些人会比较熟悉电容降级的问题，这一点是十分重要的。因为 AC 电容器必须能够处理电源线频率的 AC 电流和奇次谐波。如果滤波器是单 L 或 T 型，那么电容就应选择金属箔片型，而不是金属化薄膜型。在 π 型和多级滤波器中，至少是最后的电容器（即最接近负载的电容）应选择金属箔片型电容（详见下节）。

Some capacitors are not designed properly to handle the full ac current flow or were designed for dc applications. This is especially true if the 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. Many factories take apart the defunct unit and replace the blown capacitors with the same capacitor, stating, "We've always used this capacitor in these units." They are not aware that the capacitor is the wrong rating for this application. Commonly, the capacitor was designed for dc operation, not for ac, by the capacitor manufacturer and was selected in error by the original filter designer. The filter capacitor must be selected to handle the harmonic currents of the off-line regulator and any other pulse type of harmonic current. (See Figures 6.6, 6.7, and 6.8.)

一些电容器常常不适合处理满负载 AC 电流，而仅仅是使用在 DC 条件下的。金属化薄膜类型的电容器就是此类电容。另外也有一些电容不能处理总谐波电流，处理总谐波电流的电容需要能处理无论是电源线端还是负载端的谐波电流。处理离线调节器产生的谐波电流时就不能使用金属薄膜制作的电容，因为金属薄膜太薄了，以至于不能处理此谐波电流。另外，薄金属薄膜会增加 ESR 损耗，并导致电容温升，会使电容在使用的头一个月就会损坏。所以不能用此类电容处理离线调节器产生的谐波电流。许多厂家拆下失效的元件，并使用相同类型的电容来替换损坏的电容。并说，“我们常常在这些电路中使用此类电容。”但他们却没有注意到在此处选错了电容类型。通常，此类电容是为 DC 应用而不是 AC 应用设计的，所以最初的滤波器设计者就选错了电容。而滤波器电容必须选能处理离线调节器产生的谐波电流和其它谐波电流脉冲（见图 6.6、6.7 和 6.8）。

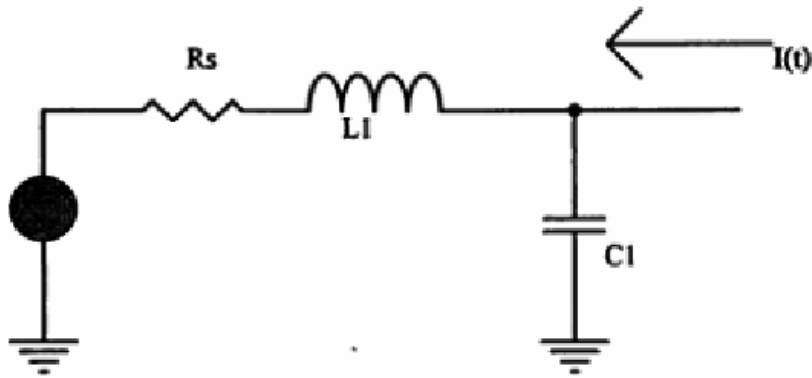


Figure 6.6 Line voltage source E_s , source impedance R_s , a single filter, and $I(t)$

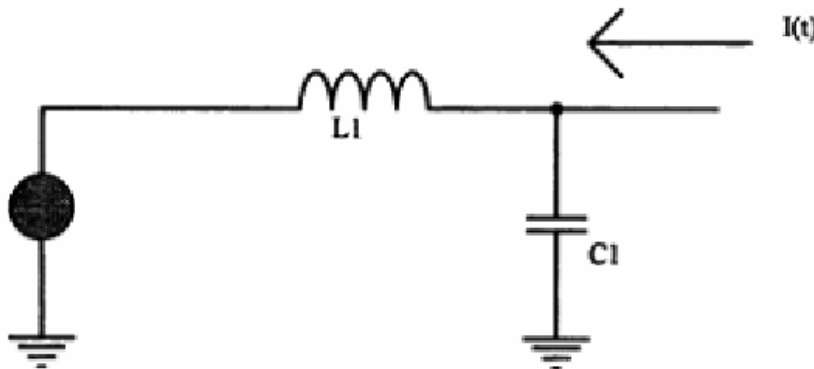


Figure 6.7 Normalized harmonic current without source resistance R_s .

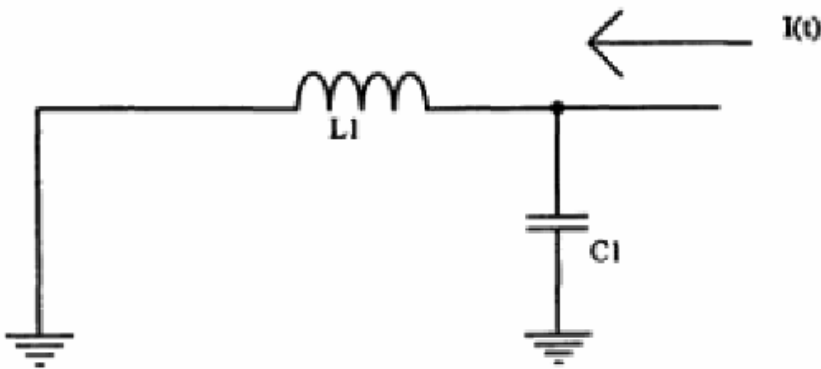


Figure 6.8 Normalized harmonic current without the source voltage.

If the line harmonic current is neglected, the line voltage is approximately sinusoidal and the normalized load current equation is , per Equation (3.2) from Fourier,

如果忽略电源线上的谐波电流，那么线电压近似为正弦波，通过等式（3.2）的以得到归一化的负载电流方程，

$$I_{(t)} = \frac{4tT}{p} \sum_{N=1,3,5,7}^{\infty} \frac{1}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \sin \frac{pN}{2} \sin \frac{2pNt}{T}$$

or

$$I_{(t)} = \frac{4tT}{p} \sum_{N=1,3,5,7}^{\infty} \frac{(-1)^{N/2}}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \sin \frac{2pNt}{T} \quad (6.1)$$

The source resistance is low over the range of harmonics of interest and can be eliminated. The inductors and capacitors making up the filter can be lumped or added because the frequency of interest is 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 $I_{(t)}$.

在我们感兴趣的谐波频率范围内，源阻抗很低并可以消去。并且在我们感兴趣的这些频率下（低于滤波器截止频率），滤波器的电感和电容可以看成是集总元件并可以将容量相加，如果是双 L 型滤波器，那么两个电感就可以相加（2L），电容同样如此（2C，如果电容量相同的话，当然它们的容量是应该相同的）。所以对于 $I_{(t)}$ 来讲，电容和电感各自都是并联的。

The voltage across the reactive pair, E_c is
元件的端电压是 E_c 是

$$E_c = \frac{-jX_C jX_L I_{(t)}}{j(X_L - X_C)} \quad (6.2)$$

The current through the capacitor follows.
通过电容的电流如下

$$I_{(c)} = \frac{E_c}{-jX_C} = \frac{X_L I_{(t)}}{X_L - X_C} = \frac{4p^2 N^2 LC I_{(t)}}{4p^2 N^2 LC - T^2}$$

Separate the sections that belong inside the summation and outside:
将分子的部分等式写在整个等式的外部，则

$$I_{(c)} = 4p^2 LC \frac{N^2}{4p^2 N^2 LC - T^2} I_{(t)} \quad (6.3)$$

and substitute in the $I_{(t)}$ equation (6.1) for the normalized harmonic current,
将 $I_{(t)}$ 带入电流为归一化电流的 (6.1)，有

$$16ptLCT \sum_{N=1,3,5,7}^{\infty} \frac{N^2}{4p^2 N^2 LC - T^2} \frac{(-1)^{(N-1)/2}}{T^2 - N^2 t^2} \cos \frac{pNt}{2T} \sin \frac{2pNt}{T} \quad (6.4)$$

where $t/2$ 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 1A. The multiplier M_p is found in Section 11.1 and Equation (11.22) is

此处 $\tau/2$ 是以秒为单位的脉冲宽度； T 是全波（两个脉冲）时间，单位也是秒； C 是电容

量，单位为法拉第；L 是电感量，单位是亨利；N 是谐波级数。这些等式都以峰值电流为 1A 进行了归一化。在 11.1 节表达式 (11.22) 中的乘数是

$$M_{lp} = \frac{p}{\cos^{-1}[(2V_p FC - I_0)/2V_p FC] + \sin^{-1}[I_0/p(2V_p FC - I_0)]} \quad (6.5)$$

where V_p is the power line voltage peak, C is the value of the storage capacitor, farads, I_0 is the dc load current, and F is the line frequency. The denominator is the value of τ needed in Equation (6.4).

此处 V_p 是电源线电压峰值；C 是储能电容容量，单位是法拉第； I_0 是 DC 负载电流；F 是电源线频率。分母是等式 (6.4) 所需要的 τ 值。

A table can be generated for the normalized odd harmonics. If τ is 0.00028 [from Equation (6.2.5)], T is 0.0025, L is 640uH, and C is 2uF, and treating the last sine as the RMS value (multiply by 0.7071; see Table 6.1). The -1 term can also be eliminated. This was developed using the computer program Mathcad. 可以按归一化的奇次谐波建立一个数据表格。如果 τ 是 0.00028 [从等式 (6.2.5) 计算出]、T 为 0.0025、L 为 640uH、C 为 2uF，并将正弦值转换成 RMS 值 (乘 0.7071，见表 6.1)。而 -1 项同样可以消除。这些数据都是使用 Mathcad 软件 (常用的数学工具软件。一译者注) 来计算的。

Table 6.1 was calculated for the approximate current of all the odd harmonics from $I_{(t)}$. If these are squared and added, and the square root then taken, the total normalized odd harmonic current is about 2.713A through the 19th harmonic. Column 1 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 of 1A. It is high because F_0 is 4448.5Hz, which is close to the 11th harmonic of the line frequency. The 400Hz current through the 2uF is 0.58A. 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.

表 6.1 是由 $I_{(t)}$ 计算出的所有奇次谐波的近似电流值。如果这些值取平方并相加，再取此和的平方根，那么直到 19 次谐波的总归一化奇次谐波的电流约为 2.713A。表 6.1 的第一列是谐波级数；第二列是谐波电流；第三列是所有此谐波之前的谐波 (含此谐波) 电流的均方根值。这些都是基于 1A 的归一化电流。因为频率是 4448.5Hz，而且这个频率比较接近于 11 次谐波，所以谐波电流值很高。在电源线频率是 400Hz 时，通过 2uF 电容的电流是 0.58A。将总谐波电流 2.713 除以 0.58 得到 4.67，可见在此例中，通过电容的谐波电流是基频电源电流的 4.67 倍。这也表明，设计滤波器必须考虑谐波电流，并且滤波器的截止频率不应接近任何电源奇次谐波的值。这同样表明，电容需要选择金属箔型而不是金属化薄膜的类型。

Table 6.1 Harmonic Current Content

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 (See Tables 6.2, 6.3, 6.4 and 6.5). For 1/32 inch and above, use 1/8 if possible. For paper above 4 inches wide, and 1/16 inch to all readings.*

如果通过电容器总的电流过多（见表 6.2、6.3、6.4 和 6.5）那么就应该用片型来替代延伸金属箔型电容。对于 1/32 以上的情况，如可能使用 1/8。对于纸大于 4 英寸宽，那么对于所有的读数都使用 1/16 英寸。*

Note that there are two foils and two dielectrics per thickness in the three columns in Table 6.2. Also, some extra allowance for bends, for example, is included in Table 6.5.

注意在表 6.2 的第三列中，每种厚的有两种金属箔片和两种介质。另外，对于弯曲的材料还有额外的要求，例如表 6.5 中的要求。

As the line frequency rises, the margins must be increased if the margins are one 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 Tables 6.2 through 6.5. This must be done to handle creepage. In the design, the first step is to determine the dielectric.

当电源线频率升高时，如果电容的余量取的较小或由于 dv/dt 导致电容损耗增加，就需要提高电容的余量。如果表 6.2 到 6.5 中所建议的电压升高时，也需要增加电容的余量来防止爬电现象。电容设计的第一步是确定电容的电介质。

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

Table 6.2 Voltage Ratings

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

Table 6.3 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 6.4 RFI Corp. Oil-Impregnated Capacitors: Dielectric Paper to 125°C

Ply thickness X	DC volts	Test		AC volts	Test	
		flash	1 min.		flash	1 min.
$2X$ 0.0002	65	160	130	35	90	70
$2X$ 0.00025	100	250	200	55	140	110
$2X$ 0.0003	160	400	320	85	215	170
$3X$ 0.0002	200	500	400	95	240	190
$2X$ 0.00035	200	500	400	120	300	240
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	2500	2000	400	1000	800
$4X$ 0.0003	1200	3000	2400	440	1100	880
$3X$ 0.0005	1300	3250	2600	500	1250	1000
$2X$ (0.00035 + 0.0004)	1500	3750	3000	550	1375	1100
$4X$ 0.0004	1600	4000	3200	600	1500	1300
$4X$ 0.00045	1800	4500	3600	660	1650	1320
$4X$ 0.0005	2000	5000	4000	720	1800	1440
$5X$ 0.0004	2200	5500	4400	880	2000	1600
$5X$ 0.0005	2500	6250	5000	1000	3500	2000

Source: From Robert Hassett.

Polycarbonate is good without derating from -55 to 125°C and has a good dissipation factor or a higher Q . This type of dielectric is easily contaminated by impurities either in storage or during winding.

聚碳酸酯是一种很好的电介质，它在 $-55\sim 125^{\circ}\text{C}$ 间无温度导致的降级现象，并且它还有很好的损耗因数（介质损耗角）和高 Q 值特性。但此电介质在储存和绕制时比较容易被杂质污染。

Polyester (Mylar) is good in the same range if the voltage is derated by 50%. Polyester is also rated from -55 to $+85^{\circ}\text{C}$ without derating. This has a much worse dissipation factor or lowest Q .

聚酯（迈拉）当电压降低 50%时，额定值变化很小。聚酯介质在 $-55\sim 85^{\circ}\text{C}$ 的温度范围内额定值基本恒定。但它的损耗因数非常差， Q 值在所有电容介质中最低。

Polypropylene is also rated from -55 to 85 and 105°C with a 25% voltage derating. The Q is higher than that for Mylar (polyester).

聚丙烯在 105°C 电压变化到 25%时额定值基本恒定。在温度范围在 $-55\sim 85^{\circ}\text{C}$ 区间时额定值也保持恒定。其 Q 值比迈拉（聚酯）电介质要高。

Paper in oil-filled caps is also good to 125°C , and the quality of the capacitor is similar to that of mylar.

油浸纸介质电容到 125°C 时额定值的稳定性也比较好，其品质方面的特性类似于迈拉介质。

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. Each has a different value of K , the dielectric constant, needed to calculate the active width of the capacitor A_w . Next, the full thickness of the dielectric plus the film must be determined based on the voltage derating and the preceding tables.

电介质厚度的要求是与电容的工作电压有关的。相关表中推荐的值往往是基本厚度的两倍、三倍或四倍。每种厚的都具有不同的介电常数 K ，在实际计算电容器宽度 A_w 时往往要用到 K 值。接下来，确定介质加薄膜的总厚的需要考虑到介质导致的电压额定值下降或按上面提到的数值表来计算。

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

如果应用条件是在 105°C 200V，那么考虑到聚丙烯和金属箔片导致的电压额定值下降，介质厚度的选择就需要选择为 266V 电压条件的厚度比较适合。

Table 6.5 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
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

§6.2.2 Round Capacitor: Extended Foil

§6.2.2 圆形电容：延伸金属箔型

Solve for the number of turns, N (see Figures 6.9 and 6.10):

缠绕圈数 N (见图 6.9 和 6.10) 的计算为

$$N = \frac{D_s - D_a}{2t} \quad (6.6)$$

where t is the total thickness of the two dielectrics and the two foils, as in Figure 6.10. The value of t is in inches. T is the value of the thickness, mils, for one dielectric only.

此处 t 是图 6.10 中的两层介质和两层金属箔片的总厚， t 的单位是英寸。 T 是一层介质的厚的，单位是 mil。

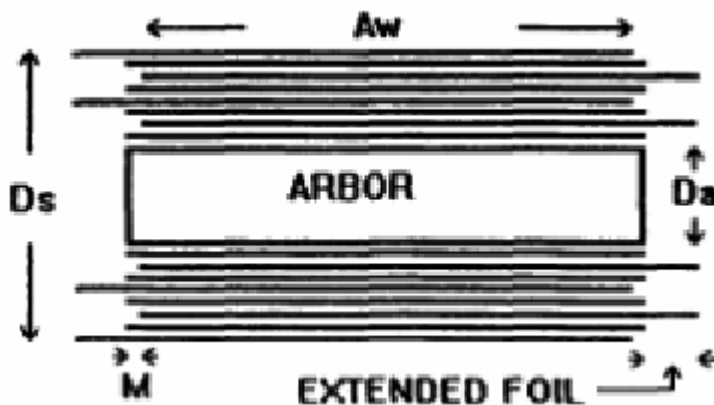


Figure 6.9 Extended foil capacitors.

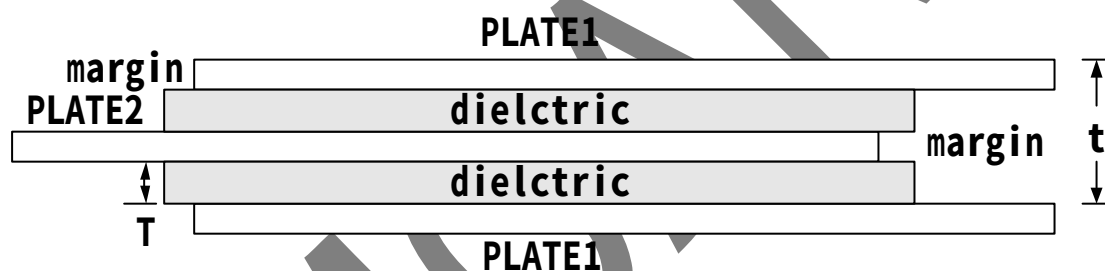


Figure 6.10 Close-up view of extended foil plates with dielectric. T is the dielectric thickness in mils while t is the thickness of two dielectrics and two foil plate thicknesses in inches.

图 6.10 显示的是带介质的延伸箔片金属板的特写图。 T 是以 mil 为单位的介质厚度，而 t 是以英寸为单位的两层介质和两层金属板的总厚度。

Solve for the mean diameter D_m :

平均直径 D_m 是：

$$D_m = \frac{D_s - D_a}{2} \quad (6.7)$$

Solve for the mean length of turn L_m :

每圈的平均长度 L_m 是：

$$L_m = \pi D_m \quad (6.8)$$

Solve for the active length L_r :

有效长度 L_r 为：

$$L_r = N L_m \quad (6.9)$$

Substituting Equations (6.6), (6.7), and (6.8) into (6.9),
将等式 (6.6)、(6.7) 和 (6.8) 带入到 (6.9) 中, 则

$$L_r = p \frac{D_s^2 - D_a^2}{4t} \quad (6.10)$$

To continue, we need to know the K value of the chosen material, find the dielectric thickness in mils (also listed in the charts), and move the decimal over three places. The value 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.

接下来需要知道选定材料的 K 值和以 mil 为单位的介质厚度；并保留小数点后三位。K 值是介质形状的函数，且介质的选取是基于损耗因数（介质损耗角）和尺寸的。其中迈拉的这些参数在正常的温度范围内是最小的。

Typical K value are polyester (Mylar), 900; polycarbonate, 840; and paper (resin or PBT), 580 (wet).

典型的 K 值为：聚酯（迈拉）为 900；聚碳酸酯为 840；纸材料（涂树脂或 PBT）为 580（湿）。

Solve for the active width (A_w):

有效宽度 (A_w) 的解为：

$$A_w = \frac{CKT}{L_r} \quad (6.11)$$

where T is the thickness of the dielectric only and the value of the material is in mils. C is the capacitance value, uF, and K is the dielectric constant.

此处 T 只是以 mil 为单位的介质厚度。C 是电容值，单位为 uF，K 是介电常数。

Substituting Equation (6.10) into Equation (6.11),

将 (6.10) 带入 (6.11) 则，

$$A_w = \frac{4CKTt}{p(D_s^2 - D_a^2)} \quad (6.12)$$

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 $A_w + 2M$, where M is the margin for the corona.

如果为此电容器分配的空间确定了，那么通过此等式可以计算电容器的宽度。最终的宽度值会略微大于 $A_w + 2M$ 。此处 M 是为避免电晕而留的余量。

§6.2.3 Pressed Capacitor

§6.2.3 压制型的电容

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:

绕薄片用的芯轴在绕完就拆掉了，然后将电容器压成平板。如果高度（ D_y ）和压制后的平板宽度（ D_x ）是已知的，那么绕制电容器所需的芯轴直径 D_a 就可以确定下来了：

$$D_a = \frac{2(D_y - D_x)}{p} \quad (6.13)$$

Solve for the number of turns:

绕制的匝数为：

$$N = \frac{D_a}{2t} \quad (6.14)$$

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

此处 t 和 D_x 、 D_y 的单位都还是英寸，

The mean diameter is

平均直径为

$$D_m = \frac{2(D_y - 0.215D_x)}{p} \quad (6.15)$$

The mean length is

平均长度为

$$L_m = pD_m = 2(D_y - 0.215D_x) \quad (6.16)$$

The active length is

有效长度为

$$L_r = L_m N \quad (6.17)$$

Substituting Equations (6.13) through (6.16) into (6.17),

将 (6.13) ~ (6.16) 式带入 (6.17) 得

$$L_r = \frac{D_x(D_y - 0.215D_x)}{t} \quad (6.18)$$

$$A_w = \frac{CKTt}{D_x(D_y - 0.215D_x)} \quad (6.19)$$

The K values are slightly different from those listed earlier: polyester (Mylar), 800; polycarbonate, 750; and paper (resin or PBT), 515 (wet).

此时 K 值与当初列出的有些差异：聚酯（迈拉）为 800；聚碳酸酯为 750；纸（树脂涂覆或 PBT）为 515（湿）。

Again, with the active width and knowing the needed margins, the full length can be determined for the space needed to fit the capacitor. Remember, 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.

此外，知道了有效宽度和需要的余量后，就可以确定出电容的整体长度。记住， T 值是以

mil 为单位的一层介质的厚度，而 t 值是以英寸为单位的金属箔片和介质的总厚度。

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

与平板或圆型电容器缠绕相关的其它等式如下

Dielectric width = $A_w + 2M$

Foil width = $A_w + M + P$

Gage = $A_w + 2M + 2P$

介质宽度 = $A_w + 2M$

箔片宽度 = $A_w + M + P$

线 规 = $A_w + 2M + 2P$ (Gage = Gauge 为线规之意，译者注)

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 feedthrough type.

此处 P 为金属箔片的伸出量，设置箔片伸出量是为了在每个终端通过阴极真空喷镀制作引线接点或为穿芯型电容制作终端接触面。

For the feedthrough type (Figure 6.11), the arbor tube is left in place and the maximum filter current, not the capacitor current, must be known. A conductive bolt, probably not copper because copper lacks the strength necessary to handle the torque required by the nuts, is silver plated by a military specification plating company (assuming a military contract). This bolt is designed to carry the full filter current and must fit through the center of the arbor. The diameter needed for this material to handle the current with low heat rise is often the factor determining the inside arbor diameter. This bolt is often plated to help handle the current and normally forms the outside stud for electrical connections for the output.

对应穿芯型电容（如图 6.11），芯轴管要从左侧安装，同时需要知道滤波器最大电流而不是流过电容器的电流。导电的螺钉可能不是铜的，这是因为铜的强度可能达不到螺母所需的扭矩。在军用合同中，导电螺钉常常根据军用规范电镀上银。另外，此螺钉要能够承载滤波器满负荷电流，同时尺寸也比较适当，以便于能穿过芯轴管。决定芯轴内径的常用条件是，在满负荷电流下能具有较低温升的螺钉的直径。此螺钉常常通过电镀来增加处理电流的能力，而且会在输出端制作成适合电连接的外部钉头。

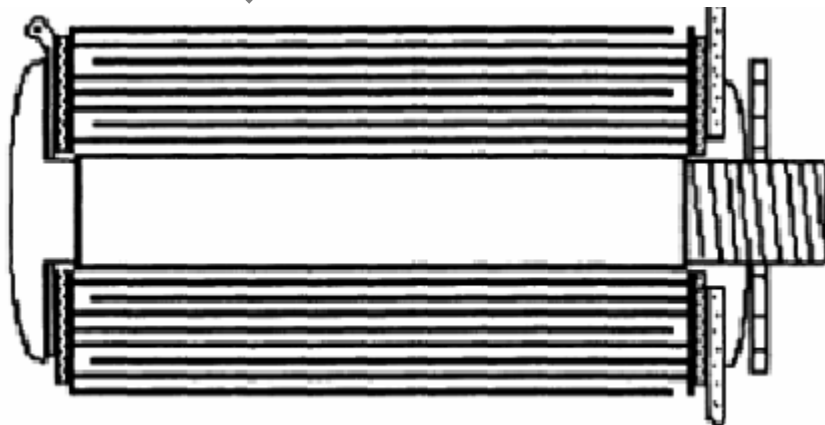


Figure 6.11 Feedthrough capacitor construction showing the case.

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

其结构如下，电镀的导电螺钉穿过连接用的垫圈或接线终端，这个电镀的导电垫圈和电容穿过芯轴孔。垫圈给芯轴孔内部电容的（阴极真空喷镀）终端施加压力以便于良好接触。导电垫圈的直径要等于电容的直径，这样此银电镀垫圈就可以覆盖整个延伸箔片的侧端表面。所有以上所说的元件都位于另一个导电垫圈之后，此导电垫圈在更大的容器孔内安装。使用一个具有一定扭矩的螺母来给最后的垫圈和电容对地终端施加压力，使之成为一个统一的连线终端。在此螺母后，通常用陶瓷制作的一个绝缘垫圈安装在螺钉终端并粘接在外筒上。火线侧是由位于外筒外面的螺钉和外筒里面的连接点共同组成的。有时会使用一个刻划过的金属板（称作地帽），并将其压入延伸箔片的侧面以替换两个镀银的垫圈。这些众多的锯齿刻划沟槽会增加电接触表面。

§6.3 Inductor Construction and SRF

§6.3 电感的构造和自谐振频率

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 transformers 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 6.12).

EMI 滤波器需要使用术语称之为软磁芯的材料。软磁芯的含义是磁芯饱和的很慢，而不像脉冲变压器和磁放大器使用磁芯具有的快速饱和特性。通过开气隙可以使硬磁芯变成软磁芯。此技术可以使 BH 曲线趋向饱和的趋势变得平缓（见图 6.12）。

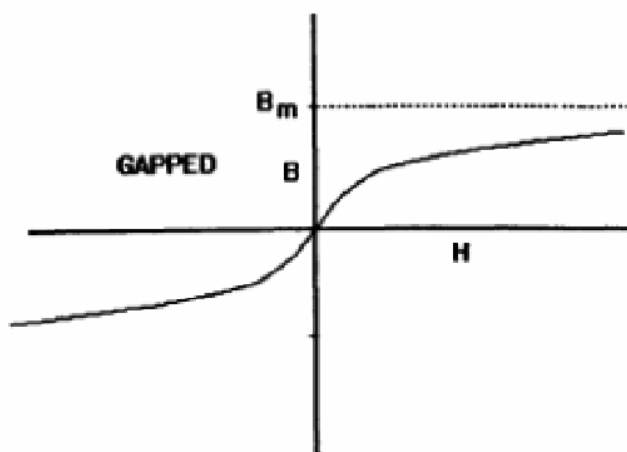


Figure 6.12 BH curve for "soft" core material

The core types used for EMI filters are often Molypermalloy toroids and sometimes the high flux type. Powdered iron toroids are also sometimes used but often cause problems at 400Hz because they often overheat. For the same reason, the permeability of MPP or HF types should be limited to 125 and below at 400Hz line frequency. This is developed as follows using MPP cores from Magnetics, Inc. through Colonel W. T. McLyman, Jet Propulsion Laboratory, NASA, in Pasadena, California.

EMI 滤波器的磁芯常常是 Molypermalloy 环, 有时也有一些高磁通型。有些时候也使用铁粉芯, 但是铁粉芯在 400Hz 使用条件下会出现过热问题。基于同样原因, 在 400Hz 的电源线频率下, MPP 或 HF 型的磁导率应不高于 125。在后面我们将看到喷气推进力实验室的 Colonel W. T. McLyman 按此原则使用 Magnetics 公司 MPP 环的例子。

The core manufacturer 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 (as Magnetics, Inc. does), the designer can choose as many circular mils per ampere as needed. The maximum number of turns that will fit in the core window can be calculated. As the current required increases, the turns drop, and as the needed current falls, the number of turns increases.

磁环制造商常常提供一些磁环的重要信息, 如外径、内径、窗口面积 (W_a)、截面积 (A_c)、重量和磁路长度 (M_{pl})。如果磁环窗口面积以圆密耳 (1 circular mils = 0.0005067062573155716 square millimeters, 译者注) 的单位给出, 设计者可以按设计需要的 ‘圆密耳数/安培’ 来选择磁环。磁环窗口能容纳的最大匝数也可以通过此参数来计算。当需要导通的电流升高时, 匝数会减少; 而电流下降时, 匝数就增加。

For any circular mils per ampere that is chosen, NI is a constant, making H a constant for the core. From H , B can be found utilizing the information provided by the manufacturer and using the winding or fill factor of 0.4 for the window of the toroid. 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.

对于任何选定的圆密耳/安培数值， NI 都是一个常数，这使得磁环的 H 保持常数。对于 H 、 B ，可以通过制造商提供的信息和圆环绕线常数（此常数为 0.4）来确定。但这并不是电感的设计方法，这样做只是为了保证一个系列中的每个磁环都以相同的绕线常数来绕制。

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

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.

此处 W_a 是以圆密耳为单位的磁环窗口面积， C_m 是适当的线径和低 DCR 所需要的圆密耳/安培数， M_{pl} 是磁路长度，0.4 是圆环的典型的窗口绕线常数。

The Magnetics, Inc. core for 125 permeability is 55438, with window area, 842,700 circular mils; magnetic path length (M_{pl}), 10.74cm; and weight, 0.4 pounds.

Magnetic 公司磁导率为 125 的环磁通密度为 55438，窗口面积为 842700 圆密耳，磁路长度为 10.74cm，重量为 0.4 磅。

Using the circular mils per ampere as 600, H follows.

如果选取圆密耳/安培为 600 时， H 为如下值

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

Using the BH curves furnished by the manufacturer yields the approximate values of B for each permeability: (Table 6.6). The given, or initial, permeability is listed as a reference, and B was estimated from the BH curve. With this information, the approximate value of the actual permeability of the core was calculated. The watts/pound follows from these formulas, also provided by the manufacturer for MPP cores (Table 6.7). From these equations W /pound was determined, (the last column in Table 6.7) followed by multiplying the W /pound by the pounds furnished by the manufacturer, here 0.4.

使用制造商提供的 BH 曲线可以计算出每个磁导率对应的 B 值：(见表 6.6)，给定的或最初的磁导率在表中列出作为参考，而 B 可以通过 BH 曲线估算出来。通过这一信息，磁环磁导率的实际值也可以计算出来。制造商同样也给出了通过这些公式计算出的 MPP 磁环的 瓦特/磅’值（见表 6.7）。通过这些公式确定了‘瓦特/磅’的数值（表 6.7 最后一列），接下来需要将重量与‘瓦特/磅’值相乘，此处重量取 0.4。

Table 6.6 Permeability and Flux of a Core

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

Table 6.7 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	$0.890 \times 10^{-11} \times F^{1.26} \times B^{2.11}$

The watts/pound and the total loss columns in Table 6.7 indicate a hefty jump from 60 to 125 permeability and continue hefty increase upward. Using a permeability of 125 is questionable. The same thing happens with powdered iron at even lower permeabilities. Reducing the frequency back to 60Hz make 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. This is fair warning of things that have happened at several EMI filter companies. Refer again to Table 6.6.

表 6.7 中的瓦特/磅参数和总损耗显示了磁导率从 50 到 125 的急剧上涨情况，并在后来仍保持急剧向上增加的趋势。因而使用磁导率为 125 来标定磁环是值得怀疑的。同样对于更低的磁导率的铁粉环也是如此。如果将频率降回到 60Hz 会产生很大的差异，因此所有的磁导率值都可能被使用到。如果违反了这一原则，那么磁环会产生过热现象。但是如果磁环的设计很保守（绕线常数取很小值，绕线匝数远低于最大值），也就是说当磁环的窗口面积较少被绕线填充，那么可以忽略这一因素。这是在几个 EMI 滤波器公司都出现了一件特别需要警告的事情。重新参考表 6.6。

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

注意到，无论如何磁导率在 125 以上时磁通密度还是太高了。

Another way of showing the same idea is to require the same inductance at the same current. The following equations of A_l values for Table 6.8 were developed using the same core as before at 5 A RMS at 380 uH. See the next section for a discussion of A_l values.

同样可以用另外一种要求在相同电流时有相同电感量的方法来显示同样的道理。表 6.8 所对应的 A_l 值的表达式使用的是与前面相同的磁芯，是在 5A RMS 380uH 电感的条件下推导出来的。详细情况参见下节关于 A_l 值的讨论。

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

$$B = mH \quad H = \frac{0.4\pi NI_p}{M_{pl}}$$

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

使用 600 圆密耳/安培是线径是一个常数，此时 No.16 (AWG)线正适合。电流峰值 I_p 为 7.07A，并可以通过它确定 H 值。

The B term was determined using the initial permeability, and the error increases with increased permeability. These could have been read from the BH curve, and the values would be less than those listed in Table 6.8. Regardless of how the value of B is determined, these should not be used much above 3000G so that a permeability of 125 might be okay, but 147 and above should be avoided.

虽然可以用起始磁导率来确定 B 项，但是误差会随磁导率的增加而增加。当然这些数值也可以从 BH 曲线上读取，但实际上读取的值会比表 6.8 列出的要小些。不管 B 值是如何确定的，在使用时也不能高于 3000G，所以磁导率为 125 是可以的，但要避免使用高于 147 的磁导率。

Table 6.8 Constant Current at 5A for a MPP Core Showing A_l Values

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

C cores are used especially at the higher current lines, mainly the power line filter used for shelters, EMI test houses, screen rooms, and secure communications applications. 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 current 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 costs of the core and the filter are increased by either method. 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 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. I have known many engineers who wanted to use Litz wire and high-frequency cores, which would increase the cost by an order of magnitude.

C 型磁芯多使用在大电流情况下，主要在屏蔽应用领域的电源滤波器、EMI 测试室、屏蔽室和可靠的通信设施上使用。C 型磁芯上磁芯缠带的厚度应按以下原则选取：在处理电源 15 次谐波和离线调节器所能产生的谐波电流时，缠带厚度取不会使磁环变热的厚度。电源线上的谐波和离线调节器产生的谐波是导致磁环温度升高的主要原因。如果滤波器在馈入电源是阻性且谐波量很小的条件下使用时，可以不遵守上面提到的这个原则。但实际上高电流尖峰使电压变得平坦，并产生了很高的谐波电流，另外还有负载会产生一些谐波电流，面对这种应用条件前面提到的原则就不能忽略了。另外一些设计选项是，使用非常低的磁通来设计电感器、允许谐波使磁环温度升高。这两种设计方法都会导致磁环和滤波器的成本增加。第一种方法会用一些更小的磁环，但是价格也更贵；后一种方法可以使用比较廉价的磁环类型，但是又常常会由于更大的物理尺寸和重量使得成本重新上升。另一个参数也不能忽略，这就是使用第一种方法所需的最小缠带厚度，薄的缠带厚度具有低的绕组占空因数，这样会提高磁环截面积 A_c 值。高的开关频率和负载端的寄生噪声无论缠带的厚度如何，都会在磁环材料上通过涡流和磁滞损耗等机制来损耗能量。而线的集肤效应会增强这些高频噪声的损耗。我了解到有许多工程师想使用绞合线和高频磁芯来设计滤波器，但这样做会使成本上升一个数量级。

The C core is easy to gap using shims of half the gap value in each leg, showing that this would be good for dc operations and would make the core "softer".

They are easy and quick to wind on a bobbin: toroids much more slowly. 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.

在每一边使用厚度为半个缝宽的薄垫片就可以很容易地完成 C 型磁芯的开气隙工作，开气隙后对 DC 应用效果良好，并能使磁芯变软。而且还能很快完成绕制工作，而圆形磁芯的绕制工作就比较慢。如果需要很多这种电感器，可以将多个（数目取决于线径）C 型环同时绕制线圈。但圆环却一次只能绕一个。因此，C 型环生产速度更快。

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 6.13). Continue until all the turns are on the core and the number of turns can be placed around the core without making much more than 320° of the circumference of the core window. Place a barrier or space between the point of start to the point of finish. This barrier is used so that the two ends cannot slide together, and can be of any material, even tape. This technique decreases the turn-to-turn capacitance in series across the inductor. This method works because the capacitance varies with the voltage difference between the turns and the voltage difference is small for nearby turns using this technique. The disadvantage is that it looks like a 5-year-old child wound it. People normally cover it with tape to help hide this, because if uninformed customers see it, they may question the ability of the company.

电感的 SRF 值是磁环尺寸和绕线匝数的函数。对于圆环，可以有几个方法来提高 SRF。第一种方法是使用渐进的绕线技术，如 6 圈向前绕，接下来 5 圈在前面 6 圈绕线层上向回绕，然后再 6 圈向前绕在前面的绕线上面，接下来是另外 5 圈向回的绕线（见图 6.13）。继续绕线直到所有的线匝都绕完，同时要保证绕线不能超过磁芯窗口圆周的 320° 。在绕线的起点和终点之间放置一个屏障，它能保障绕线的两个端口不会滑到一起，这个屏障可以用任何材料来制作，甚至可以使用缠带。这种方法降低了与电感并联的匝间电容。实际上这种方法很有效，这是因为这个电容是随匝间电位差的变化而变化的，而使用这种绕线方法使匝间电位差变得很小。这种绕线方法的缺点是绕出的线圈像 5 岁小孩绕的线圈，人们通常使用缠带来掩藏绕线，如果不专业的客户看见了这种绕线，他们会对生产商的能力提出质疑。

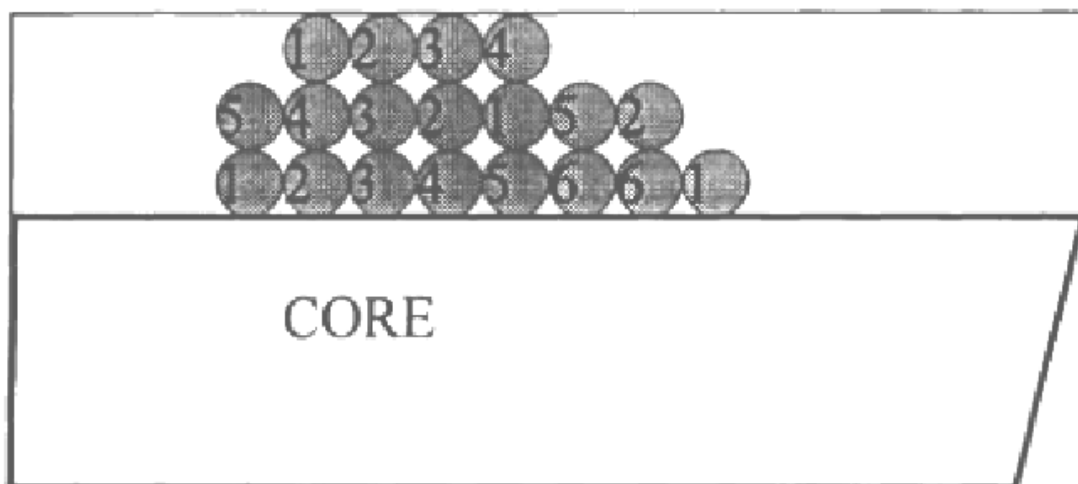


Figure 6.13 Progressive wind: a better SRF.

Side-by-side windings in which there is less than a full 360° 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. Tape the core with several thicknesses to keep the turn-to-core capacity as low as possible (Figure 6.14). Toroides can be purchased that are already coated. Often a layer of tape is also needed to raise the wire farther from this core. 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 6.14 shows tape with a single layer of turns. If many layers are needed, use the progressive technique, just discussed, along with this technique. If a single layer is adequate, try to separate the turns slightly, which also lowers the turn-to-turn capacitance. Do not allow the two ends to come too close together. Try to keep them at least 20° apart.

另一种方法是并排绕线，且使绕线在磁环圆周上少于 360° ，这种方法也很充分。此方法基于同样的原因会使匝间电容变得很小，匝间电压也很低。但是，当需要很多层绕线时，很显然 2 层与 1 和 3 层相比，会有更多的具有高匝数差的绕线，这会产生更高的电位梯度，因而匝间电容就更高。结果会导致较低的 SRF，最终影响滤波器的使用功能。第二种方法是让绕线不直接绕在磁环上，而将磁环用缠带绕几层，确保匝—环间电容尽可能小（图 6.14）。当然也可以购买带涂层的磁环，但通常还需要在这种磁环上再绕一层缠带以增加线圈和磁环的间距。这个匝—环间电容更趋向于并联，所以会很显著、很直接地增加与电感并联的电容容量。图 6.14 表示的是缠带上有一层绕线的情况。如果需要多层绕线，就要使用先前提到的渐进绕线方法，并结合此方法来绕线。如果单层绕线就满足要求，试着稍微疏散一下各匝，这也能降低匝间电容。并且不要让绕线两端距离太近，将它们分开至少 20° 的间隔。

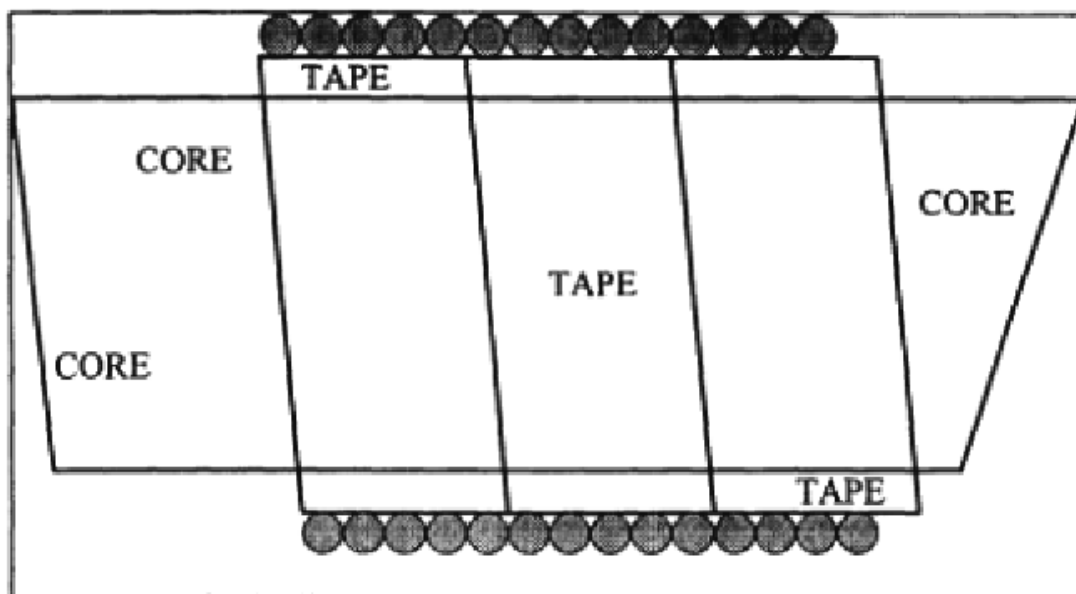


Figure 6.14 Taped core for better SRF.

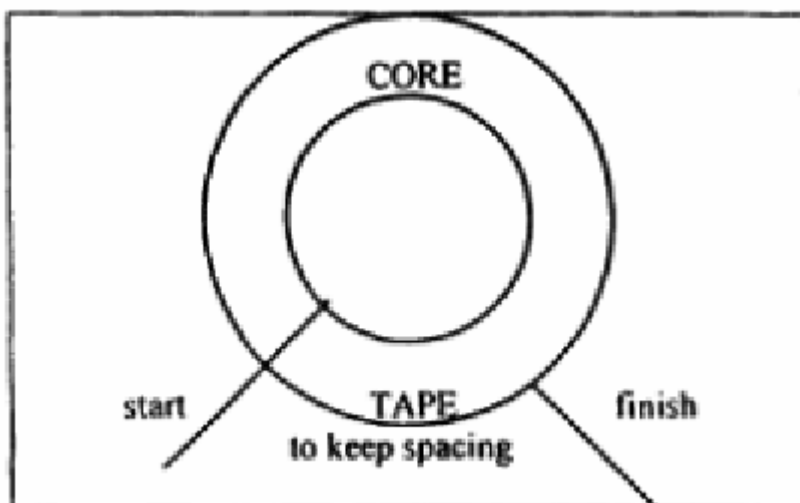


Figure 6.15 Core showing the start to finish spacing.

Another way to raise the SFR is to use heavier insulation. This forces a higher spacing from turn to turn and helps raise the copper farther from the core and decrease the turn-to-turn capacity (Figure 6.15).

另外一种提高 SFR 的方法是使用强绝缘的方法。这种方法会增加匝间铜芯的间距和铜芯与磁芯的间距，因而会降低匝间电容（见图 6.15）。

§6.2.3 Inductor Design

§6.2.3 电感的设计

The design of inductor was covered for the most part in Chapter 3. 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. Nanohenries per turn is rarely used for the type of filters discussed in this book.

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 was developed from the following:

第三章的许多内容已经叙述了电感的设计。此处介绍的另一种常规方法是使用 A_l 值设计。 A_l 通常用毫亨/1000 匝和微亨/100 匝的单位来标定。而纳亨/匝在电源滤波器中很少使用。只要生产商列出了 A_l 值或通过其它方法得到 A_l 值, 就可以据此来设计所有类型的电感, 包括共模电感。

$$L = \frac{0.4\pi\mu N^2 A_c \times 10^{-8}}{M_{pl}} \quad (6.20)$$

$$A_l = \frac{0.4\pi\mu A_c \times 10^{-8}}{M_{pl}} \quad (6.21)$$

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 A_l value is normally provided by the manufacturer.

所有这些 A_l 项不是常数就是磁环选用所使用的通用值。此处相关的变量为: A_c 是铁心的截面积, μ 是磁导率, N 为匝数。通常 A_l 值由制造商提供。

$$L = A_l N^2 \quad (6.22)$$

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

其中, 电感值是已知的, 这样就可以计算出匝数。由于 A_l 值已给出, 所以

$$N_2 = N_1 \sqrt{\frac{L_1}{A_l}} \quad (6.23)$$

where N_2 is the needed number of turns, N_1 is the known number of turns (1000), 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 (1000).

此处 N_2 是所需的匝数, N_1 是已知的匝数 (1000), L_1 是所需的电感值, 其单位与 A_l (毫亨) 相同。 A_l 是给定匝数 (1000) 条件下的毫亨值。

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

例如, Magnetics 有限公司 55351 MPP 磁环列出的 A_l 值为 51mH/1000 匝。我们需要 0.8mH (800uH) 的电感, 即 L_1 等于 0.8, A_l 等于 51。那么 0.8 除以 51 等于 0.01569, 其平方根值为 0.1252。此值的 1000 倍为 125 匝。

$$N_2 = 1000 \sqrt{\frac{8}{51}} = 125.245$$

This is followed by the current multiplied by circular mils to determine the window area to determine whether the turns fit. This is a small core, but we need 1A. Using 600 circular mils per ampere would lead to 23 AWG at 650 circular mils; 650 multiplied by 125, the number of turns just calculated, equals 81,250 circular mils, but the winding factor for this toroid core is only 0.4. The 81,250 divided by 0.4 yields 203,125 circular mils. Magnetics, Inc. lists the window area as 298,800 circular mils, so the windings will fit. The wire size could be changed to 22 AWG, giving 253,125, and still fit. This would reduce the copper losses and lower the temperature rise.

接下来将电流乘以圆密耳数计算出磁环窗口的尺寸,通过这个尺寸来确定此磁环是否能满足需绕制的线匝数。本例使用的是一个小磁芯,但我们只需要 1A 的电流。通过 600 圆密耳/安培可以确定 23 AWG 线 (650 圆密耳); 650 乘 125, 得到全部匝数的线圈需要 81,250 圆密耳,但是这种圆磁芯的绕线因数仅为 0.4。81,250 除以 0.4 得到 203,125 圆密耳,这就是实际需要的磁芯窗口面积。Magnetics 有限公司列出的此磁芯的窗口面积为 298,800 圆密耳。所以此磁环可以容纳绕线。线径也可以换成更大一些的 22 AWG (253,125 圆密耳), 同样也适合本例的需求。这会更有助于减少铜损降低温升。

I recommend McLyman's programs for selecting the proper inductor core and turns. The heat rise is a common filter problem, and the filter design should be rather conservative: McLyman's programs meet these requirements.

我推荐使用 McLyman 的程序来选择合适的磁芯和绕线匝数。温升是滤波器的一个常见问题,所以滤波器设计尤其需要谨慎, McLyman 的程序能满足这样的需求。

§6.5 Convert from Balanced to Unbalanced or the Reverse

§6.5 从平衡电路到非平衡电路的变换或反变换

Conversion from balanced to unbalanced 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. This happened to a customer who specified a complete filter design. We tried to tell them that the bottom half of the filter was not working, but it was never changed because it went through equal testing that way. The change would have cost them almost \$100,000.

实际应用中常常需要从平衡电路到非平衡电路的变换或反变换。这是因为,计算公式都是基于非平衡电路的。如果电源或设备还没有接地,那么平衡电路的滤波器就比较好。还有一种情况常常出现在定型了滤波器设计的客户身上。虽然我们试图告诉他们滤波器下一半电路并不起作用,但是客户并没有改变电路,这是因为客户使用这种滤波器并通过了测试实验。而改变电路会花费他们 100,000 美元的费用 (需要重新做测试,译者注)。

To balance the filter, as in the 800uH discussed in the last section, use a double L with some value of capacitor, say 1uF with a limit of 0.1uF to ground. The

picture would be 800uH series, 1uF Y capacitor shunt, followed by two just like the first two, and then finish with the two feedthroughs on both lines. Simply divide both 800uH inductors in half so that the circuit ends us with 400uH on both the hot and neutral, with the X capacitors of the same values across the lines. The inductors are smaller, and the SRF should be higher. Mount the four inductors in quadrature (the two inductors that would be 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 6.16 illustrates before and after.

在上节讨论的 800uH 的例子中,使用两个 L 和某一容值的电容如 1uF,另外对地还有 0.1uF 的限制。电路图为 800uH 串联电感, 1uF 旁路的 Y 电容,后面的两个元件与前面的相同,最后在电源线上接两个穿心电容。接下来平衡该电路:简单地取两个 800uH 电感器的半值,并分别串于火线和零线上,另取容值相同的 X 电容并于线间。结果是电感容量变小了,而 SRF 相应地提高了。以正交位置安装此四个电感器(每二个电感需要远离安装,以期望额外的距离会减小耦合),并令电容器引脚尽可能地短。图 6.16 表示的是开始和完成的电路图。

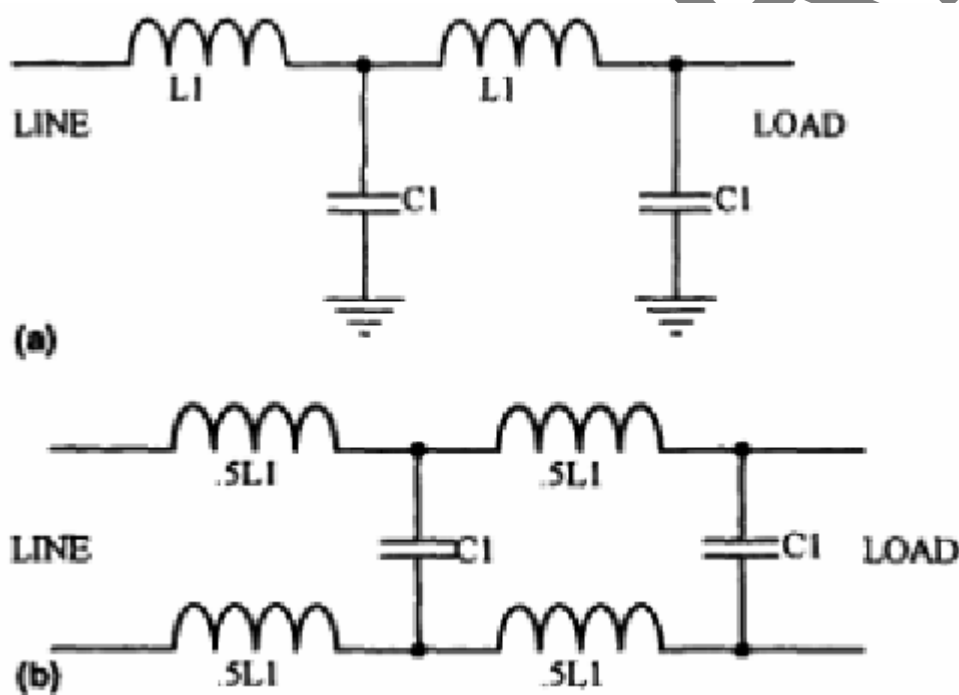


Figure 6.16 The multiple L network converting from unbalanced (a) to balanced (b).

7

Common Mode Components

共模元件

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. The problem is another specification limiting the leakage current. This is the capacitive reactance current flow through the capacitor between the line and ground, another misnomer.

通常产生电磁脉冲或尖峰的原因有：闪电、切断大电感设备和核放射产生的磁脉冲。闪电和核放射在线—地间生成共模脉冲。切断大电感设备则在线间生成差模脉冲。在负载侧，开关电源是主要噪声的源。在离线调节器的负载侧，任何对地的电流脉冲都表现为线上的共模噪声。加隔离变压器会减小这些共模传导发射脉冲（此时，唯一一个损害共模性能的因素是在初级和次级线圈间的电容），此时就可以不使用向地泄放（或从地吸取）电流的滤波器对地电容了。为了消除共模噪声，EMI 滤波器常使用共模扼流环和对地电容这些元件。但问题是另外一个规范会限制漏电流。用一个不专业的解释是，漏电流是在线—地间通过电容容抗的电流。

§7.1 Capacitor to Ground

§7.1 对地电容

For 400Hz, the limit of the capacitor to ground is 0.92uF 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 100uA. This means that most power supply people want the filter restricted to 20-40uA. It is difficult to have common mode losses that meet common mode loss specifications with capacitors to ground this small. There are two schools of thought on this problem because the specifications governing it, as usual, are not clearly written. The first group thinks that this is the total capacitor to ground. In three-phase four-wire circuits, the capacitor limit value for 400Hz is 0.02uF. This capacitor value is then shared, or

在 400Hz 应用条件下, 461 军标中对地电容的限值为 0.92uF。而对于医疗的漏电流规范就更难满足了。如果是接触病人的医疗设备, 那么系统总漏电流会限制在 100uA 以下, 这意味着大多数电源的滤波器漏电流应限制在 20—40uA。此限制决定了对地电容是如此之小, 以至于很难满足规范所规定的共模损耗。另外, 因为限制漏电流的规范并没有明确写明要求, 所以这里有两种不同的观点需要提出。第一是对地电容指全部对地电容总和。如在三相四线电路中, 400Hz 应用条件下的电容限值为 0.02uF。此电容值是要共享分配的, 即

$$\frac{0.02}{4} = 0.005mF$$

that is , 0.005uF for each of the four legs to ground. The second group thinks that this is the maximum per line whatever the number of lines. If the system is well balanced, the currents nearly cancel through these capacitors. I hold with the latter group because it makes my job as a filter designer easier. However, ground fault equipment does not allow capacitance to ground that produces a current above the current threshold of the ground fault device, or it would have to be installed in the circuit after the filter.

也就是说四路对地电容每路为 0.005uF。第二种观点是, 无论线数为多少, 这个值是每线的最大值。并且如果系统三相平衡, 那么通过这些电容器的电流相互抵消接近于 0。我倾向于后者的想法, 因为这会让滤波器设计工作变得相对容易些。但要注意, 地产生故障的设备不允许对地电容产生高于故障设备的门限电流, 或者可以考虑将此电容接于滤波器后的电路中。

§7.2 Z for Zorro

§7.2 Zorro 电感

Ferrites are often used for common mode inductors because they have the very high A_l 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. The line-to-line capacitors are of no value for the common mode, but the capacitors from line to ground are. In the three-phase four-wire case, the total capacitor to ground (my method) would be

共模电感常常使用铁氧体材料, 这是因为铁氧体具有共模滤波器所需的高 A_l 值。在共模条件下, 所有电源线都等效为连接在一起, 所有平衡电路中的差模电感都等效为相互并联 (上一半与下一半并联后串联在电路中, 译者注), 线间电容没有容量, 电容中只有对地电容工作。在三相四线情况下, 总的对地电容 (采用我的方法, 见上节两类想法的论述, 译者注) 为

$$0.02 \times 4 = 0.08mF$$

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 7.1). Split the value of the feedthrough capacitor limit to ground by 2. Two of these smaller feedthrough capacitors are then installed in this shield. For best results using this method, place the Zorro inductor 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.

为获得所需的共模损耗，设计上会常常使用共模扼流环。通常所有电感和所有对地电容还不能满足损耗的需求，所以常常在滤波器中心位置放置良好接地的隔断或屏蔽层来接决这个问题（见图 7.1）。具体方法是，将对地的穿心电容限值一分为二，并将这两个半值的穿心电容安装在屏蔽层上。为得到较好的性能，需将 Zorro 电感安装在低阻抗端或电源线侧，将穿心电容安装在高阻抗端或负载侧。此处，元件值都经过了半值拆分。第一个 Zorro 放置在电源线侧而第一个穿心电容安装在电感后的屏蔽层上。滤波器的第二节先是安装了 Zorro，接下来安装穿心电容。在此例中，应尽量选取偶数个差模滤波器，这样它们就可以均匀地分布在两个腔体中了。如需要两个 L 型滤波器时，那么在第一个腔体的始端安装 Zorro，接下来安装一个 L 滤波器，再往下是穿心电容。对于第二个腔体其配置方式与第一个腔体相同。

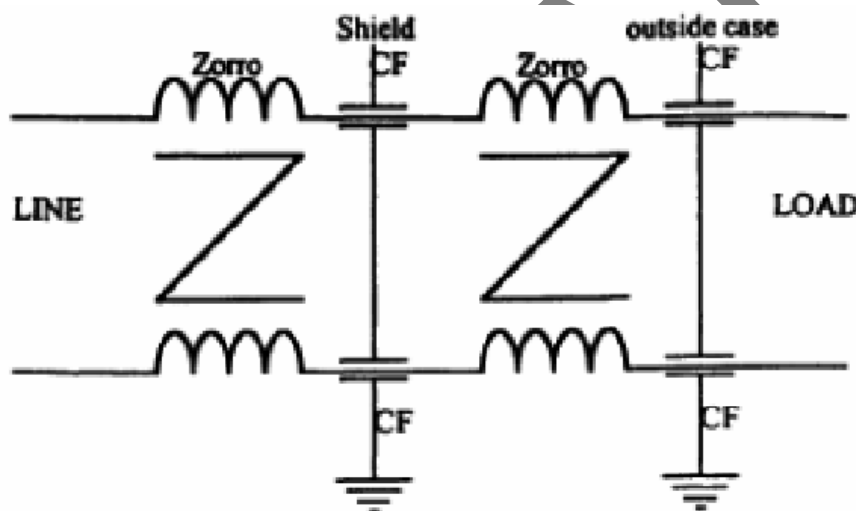


Figure 7.1 Double-Zorro common mode filter

Place a second common mode inductor 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. Remember that the inductor's magnetic fields buck or cancel for the differential mode and have a high magnetic gain for the common mode. Figure 7.2 shows a three-phase delta with spacing between the windings.

将两个共模电感置于滤波器中，一个位于前一半电路（空间）中，另一个位于后一半电路（空间）中，这样对于共模就形成了双 L 型滤波器，因此会大大降低共模电感的尺寸，而且多数情况下都可以满足规范的损耗要求。要记住，电感内部的磁场对差模是相互抵消的，而对于共模是相互增强的。图 7.2 显示了三相△型共模电感的绕线间空间的分配。

Notes

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

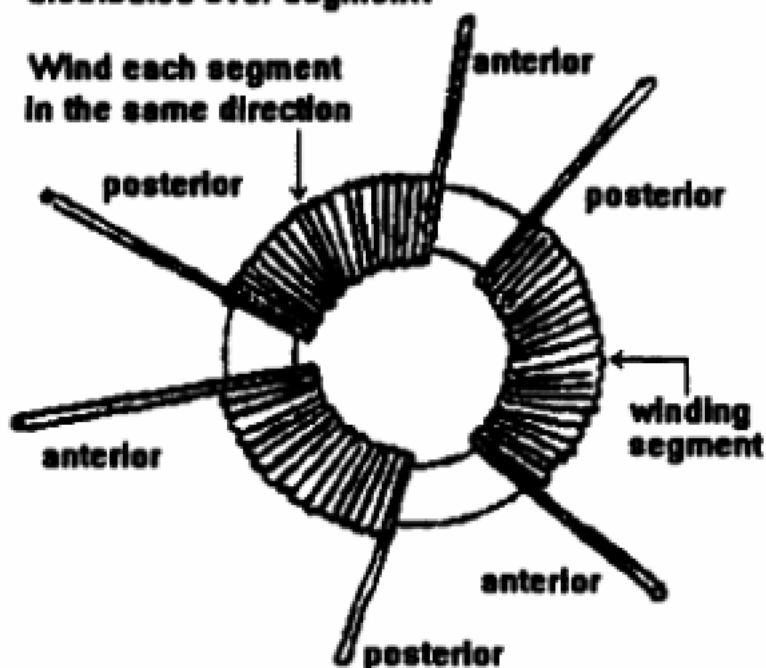


Figure 7.2 RFI Corp. three-phase three-wire common mode inductor.

For the Y, another set of windings is required, demanding a larger core for the same form. These are excellent for creepage, but the leakage inductance is greater, giving a differential inductance. These separate windings create 0.5-1% leakage or differential inductance. Figure 7.3 shows them wound together, reducing the leakage inductance, but this arrangement may be subject to creepage and corona problems.

对于 Y 型连接，需要另外一种绕线方法，这样就需要更大的磁芯。Y 型电感处理爬电性能较好（对于分离绕线方式），但形成差模电感的漏感会比较大。这些隔开的绕线会产生 0.5~1% 的漏感或差模电感。图 7.3 显示各相绕线绕在了一起以减小漏感的情况，但是这样会产生爬电和电晕的问题。

Notes

- 1 - Quadfilar winding on one core as shown.
- 2 - Wire to be tightly wound and evenly distributed.
- 3 - Lead length to be 4" minimum.

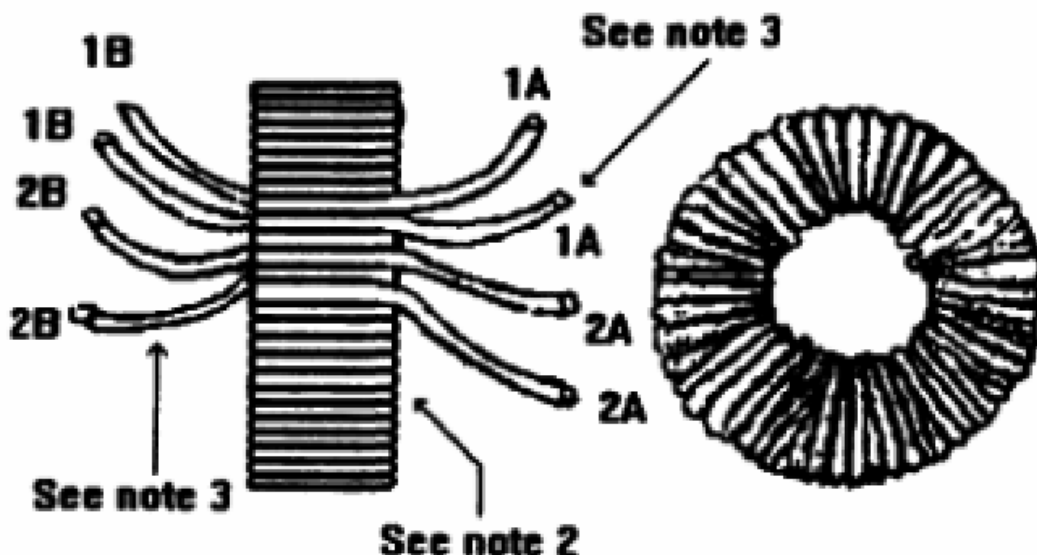


Figure 7.3 RFI Corp. tightly coupled-low leakage-common mode inductor.

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

一些设计者会将分离绕线方法换成 4 线型绕线以消除振铃。振铃是由电感和其它导线中的漏感和寄生电容形成的。

§7.3 Converting Common Mode to a Differential Mode Filter

§7.3 将共模滤波器转换差模滤波器

The common-mode inductor has one value assigned to it. The inductance value is written above the Z for Zorro, 10mH, for example. Either winding should read the indicated inductance if the measurement employed a good inductance bridge. The reason is from the inductance formula

共模电感有一个值来标定它。对于 Zorro 电感值写在 Z 上面，如 10mH。如果采用一个电感桥来测量，对共模电感的每边绕线都应能读出标定的电感值。由此，共模电感的公式为

$$L = \frac{0.4 \mu m N^2 A_c \times 10^{-8}}{M_{pl}} \quad (7.1)$$

If an inductance bridge is used to read each winding in this example, the reading of both windings is 10 mH. If the inductance bridge is used to measure the two windings, what is the aiding inductance? From Equation (7.1), the turns are squared, so that twice the number of turns gives four times the inductance, or

40mH. However, the windings are split, half on each line, so each half is 20mH. These two windings are now measured in parallel, so we're back to 10mH! In other words, either winding would measure X and both measured in parallel-aiding-would measure X. If this were measured with the windings opposing, this would read as the leakage inductance.

在本例中，如果使用电感桥来读取每边线圈的电感值，那么两个读数都是 10mH。如果用电感桥来同时测量两边，那么由互感增加的电感是多少？按式 (7.1) 来看，式中含匝数的平方项，所以两倍的匝数其电感值为四倍 40mH。实际上线圈呈一边一半的状态，所以每边应为 20mH。现在将这两个线圈并联测量，所以读数就回到 10mH 上了！换句话讲，每边单独测量读数为 X，那么两边并联读数也为 X。如果反相串联两边，那么读数为漏感。

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 common mode is that the winding or window fill factor is no longer 0.4 but now is reduced to 0.2 to fit the two windings.

接下来可以使用磁芯的 A_l 值来设计铁氧体共模电感。因为双线型要求非常低的漏感，只能使用这种材料。设计共模和差模电感的唯一不同点是，占空因数不再是 0.4 而是能适合两组线圈的 0.2。

The single-layer ferrite toroid winding can also be found when the core I_d 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

当使用电流下磁芯的 I_d 和线径（美线规 AWG）都确定下来时，铁氧体的单层绕线量也就确定下来了。如果已知线径 W_d （与磁芯的 I_d 单位相同），那么总绕线匝数 N_t 为：

$$N_t = \frac{p(I_d - W_d)}{W_d} \quad (7.2)$$

If the wire diameter is much smaller than the core diameter (and it always is except in those types discussed in Chapter 16), this approaches

如果线径远小于磁芯直径（这种情形在 16 章讨论的那些类型中总属于例外情况），可以近似为

$$N_t = p \frac{I_d}{W_d} \quad (7.3)$$

Divide the turns by 2, and use 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 .

当单层和小于半圆的绕线弧度就能满足匝数的需求时，将这个匝数除以 2，并取整数匝数来设计电感。如果小于需要的电感值，就再选另外一个更大的 I_d 值的环。

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

number of turns required using the A_l equation. Find the number of turns necessary, and use the higher integer.

如果我们找到了一个磁芯，并且用它制作的电感可以比需要的值大，此时就可以通过 A_l 公式来计算所需的匝数。通过计算找到所需的匝数并将结果取整（向上取整）。

For example, 8mH is needed; current = 2A peak; AWG = 18 picked for the current. The diameter is 0.0429 inches. Choose the core: 42915-TC from Magnetics, Inc. The l_d in inches (same units)=1.142:

例如需要 8mH 电感，峰值电流为 2A，对于此电流可以选取 AWG 18 号线，AWG 18 号线的线径为 0.0429 英寸。选取环：Magnetics 有限公司的 42915-TC 环，其 l_d 为 1.142 英寸，则

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

divide this by 2 = 41.81, and use the integer. The integer is 41 turns. The A_l value is 3868 mH per 1000 turns using F material. Here, the backward formula for A_l is used for 41 turns:

将上式结果除以 2 等于 41.81，并取整为 41 匝。使用 F 材质的环的 A_l 值为 3868mH/1000 匝。将 41 匝带入到前面提到的 A_l 相关的表达式中：

$$L = \frac{A_l (41)^2}{(1000)^2} = 6.502mH$$

This is a little low, so a larger core must be selected.

这个电感值还有些低，所以还要选取更大一些的磁环。

The diameter in inches is 0.0429. Choose the core: 43615-TC from Magnetics, Inc. The l_d in inches (same units) = 1.417.

考虑到线的直径为 0.0429 英寸。选取 Magnetics 有限公司的另一款磁环：43615-TC，其 l_d 值为 1.417 英寸。

$$N_t = \frac{p 1.417}{0.0429} = 103.768$$

divide this by 2 = 51.88.

将重新计算的匝数除以 2 得 51.88。

The integer is 51 turns. The A_l value is 4040 using F material as before.

取整为 51 匝。与前面一样此环亦使用 F 材质，其 A_l 值为 4040。

$$L = \frac{4040 \times (51)^2}{(1000)^2} = 10.508mH$$

This is 10.5 mH (原文为 2.5mH，译者注) more than our goal of 8 mH. Solve for the turns needed with the normal A_l formula:

结果为 10.5mH，比我们需要得电感量 8mH 大。现在可以通过标准的 A_L 公式，得到所需的匝数。

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

Use the upper integer, or 45 turns.

向上取整得 45 匝。

Use the 45 turns for each half-winding. Keep the end gaps between the two halves as far apart as possible. This creates a visible winding gap on the core: make this gap as large as possible. The difference between 51.88 and 45 gives this gap spacing between both ends of the two windings: 6.88 multiplied by the wire diameter of 0.0429 inches approximates this spacing. This is on an arc: the entire turn will not touch, eating some of this circumference, so that the value will be less than the 0.295 inches calculated. Again, I would tape, or fill, these spaces so that both gaps between the two windings are secure (Figure 7.4).

按计算结果，每边绕线 45 匝，并使两边的线圈的终端尽可能远离。这样在磁环上就会有一个明显的绕线缝隙，当然实际操作上要让此缝隙尽可能大。51.88 匝是最大绕线匝数，而实际绕线为 45 匝，这样就使两组绕线终端之间可以出现空间间隙：用最大匝数和实际匝数之间的差值 6.88 乘以线径 0.0429 得到的结果就是这个空间间隙的近似值。这个值是弧形间距的弧长。但是由于绕线的线间总有间隙，会消耗掉一部分周长，所以两组绕线的空间间隙会小于计算值 0.295 英寸。另外，我们需要使用胶带或其它材料填满这些绕线终端间的空间间隙，这样就会可靠保障绕线终端间的间隙。

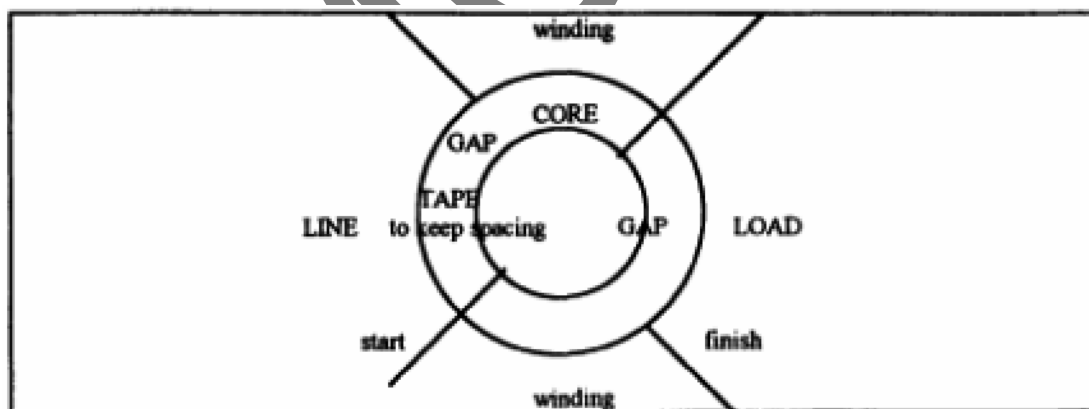


Figure 7.4 Single-layer common mode inductor showing the spacing between windings.

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.

现在的问题是，虽然已经有了选取合适的磁环尺寸的方法，但是如何确定 Zorro 的电感值为 10mH 呢？这里主要涉及两点。第一如何将共模转换成差模、平衡电路（共模电路）转换成差模，差模通常被称作常模。如果在共模电感后面的是平衡的常模网络，那么就可以将之转换回非平衡的差模电路以便于计算。

In figure 7.5, 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 7.6. 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 400uH, 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 400Hz is 0.02uF but now totals 0.04uF.

在图 7.5 中，Z 电感下面依次是两个 L 型滤波器，再接下来是穿心电容。前面已经讨论过共模电感，并将其转换成单一的电感容量为 10mH。对于共模，接下来的两个电感每个 0.5L，并联后的有效值为 0.25L。而 C1 对于共模则等效于 0。只剩下最后的元件 C2（应为 CF，译者注），其值等效为 2 倍容值的单个电容。等效的结果见图 7.6。两个差模线圈的总感值为 0.5L，但是它们常常只作为滤波器共模部分的裕量而不计算在内。这是因为，此二线圈的值只有 400uH 比 Zorro 的电感值要低一个数量级。CF 的值也常常被规范限制，如在美军标 461 中，对于 400Hz 使用条件下最大不超过 0.02uF，两线相加总共也只有 0.04uF。

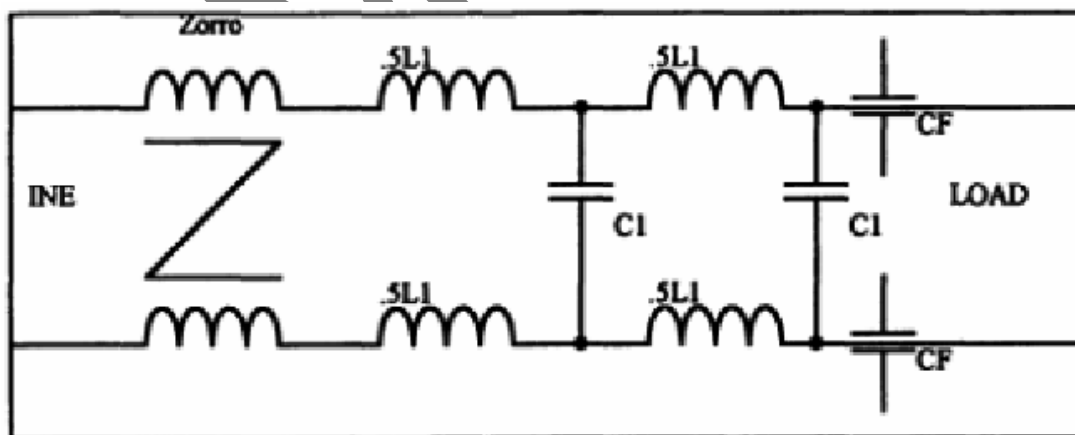


Figure 7.5 The common mode and differential mode filter.

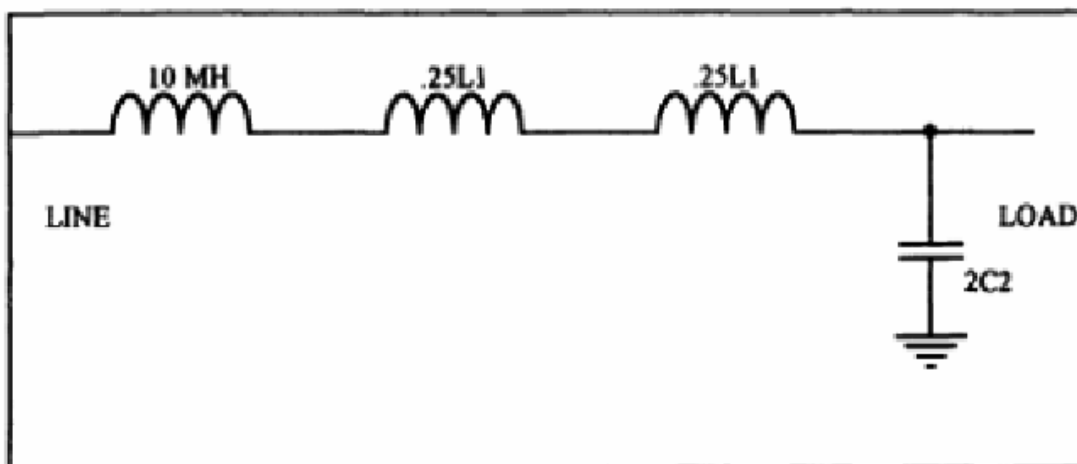


Figure 7.6 The common mode converted to the unbalanced filter of Figure 7.5.

The problem is now in a form in which the actual value of the common mode inductor can be solved using the techniques in Chapters 12 and 13 (Figure 7.7).
现在电路变成了图 7.7 的形式，这样就可以使用 12 章和 13 章的方法来计算出共模电感的值了。

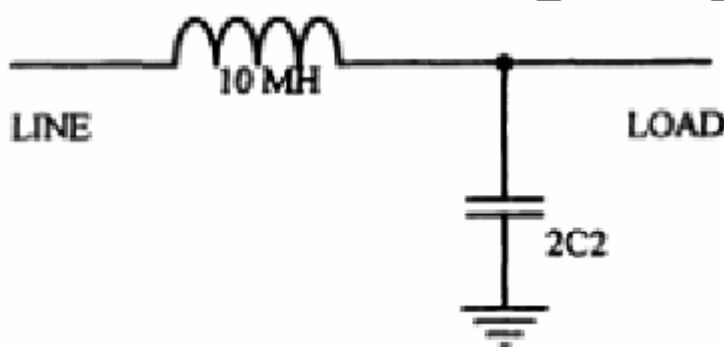


Figure 7.7 The final conversion of the common mode to the unbalanced filter of Figure 7.5

The second point is the test setup normally used to test the common mode inductor (Figure 7.8). The two inputs and outputs are shorted together. The input is fed from a generator with an output impedance of 50Ω , the two outputs, also shorted together, feeding the load. The load impedance is also 50Ω . This was another reason to convert from the balanced to the unbalanced filter as before.

第二点是，常使用图 7.8 的实验设置形式来测试共模电感。输入两端口和输出两端口分别短接在一起。电源输入是输出阻抗为 50Ω 的发电机，负载阻抗也是 50Ω 。这也是将滤波器平衡电路结构转换成非平衡电路的另一个原因。

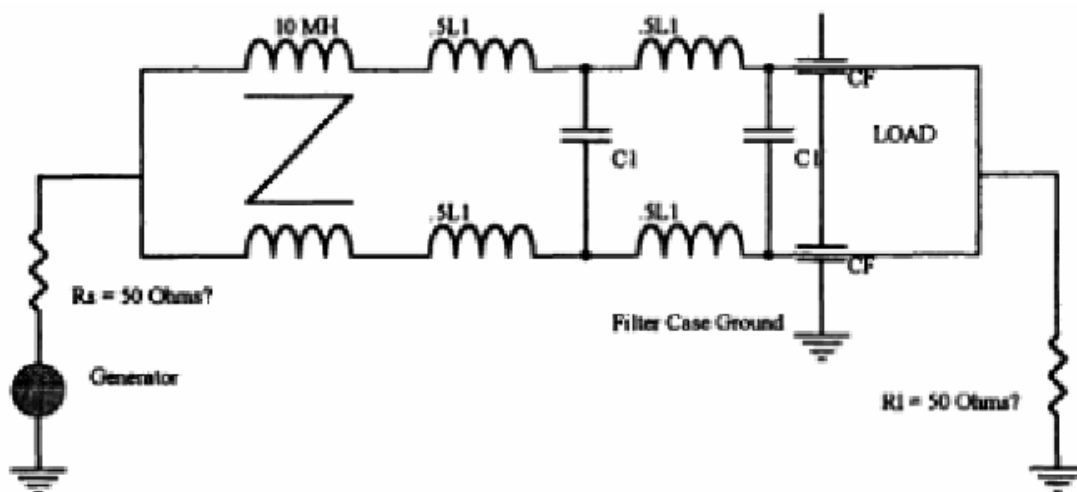


Figure 7.8 The 220A test method for differential mode section of the filter.

This all works out to the circuit shown in Figure 7.9, and the equations are in the next section.

上图可以等效成图 7.9 所示的电路，其表达式见下节。

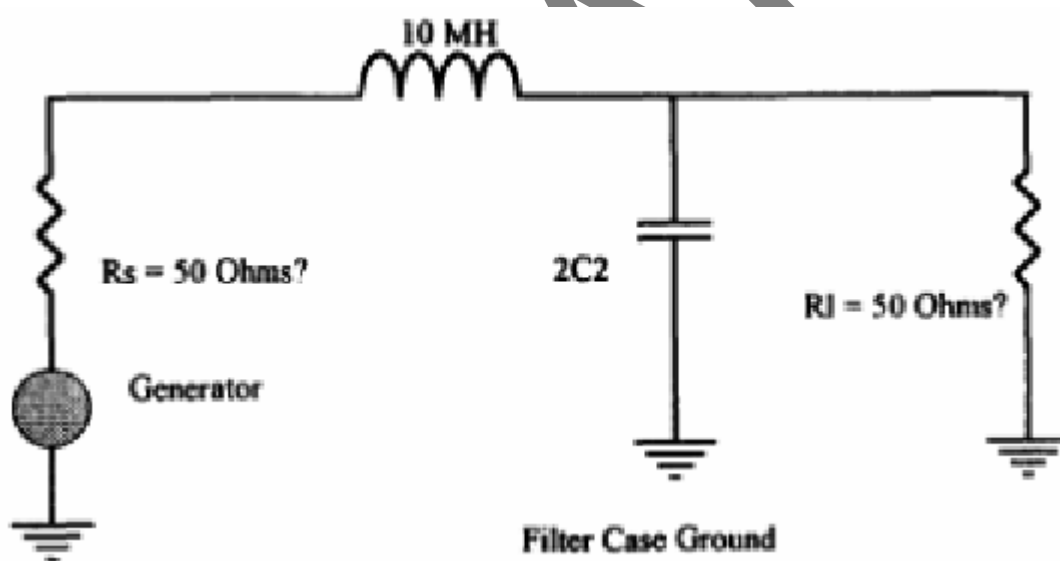


Figure 7.9 The converted test method to calculate common mode values.

§7.4 Equation for the Common Mode via the Differential Mode of Section 7.3

§7.4 通过 7.3 节的差模等效形式得到的共模的计算式

The method used to arrive at Figure 7.9 can be converted to the circuit here, where C is equal to two times $C2$, L is the common mode inductor needed for the proper insertion loss, R_s is the source impedance, and R_l is the load impedance. Because the equation includes R , L , and C , equations based on Q (charge) rather than I (current) of the two networks are the easiest to work with, generating the matrix in Figure 7.10. This is from impedance matrix equations.

将图 7.9 转换为新电路,其中 C 为两倍的 C2 值,L 为能达到需要的差入损耗的共模电感值, R_s 为源阻抗, R_l 为负载阻抗。由于方程包含 R、L 和 C,所以由图 7.10 导出的方程采用基于电荷 Q 的表达形式比使用电流 I 做变量更容易得到答案。

$$\begin{pmatrix} V_{in} \\ 0 \end{pmatrix} = \begin{bmatrix} LS^2 + R_s S + \frac{1}{C} & -\frac{1}{C} \\ -\frac{1}{C} & R_l S + \frac{1}{C} \end{bmatrix} \begin{pmatrix} Q_1 \\ Q_2 \end{pmatrix} \quad (7.4)$$

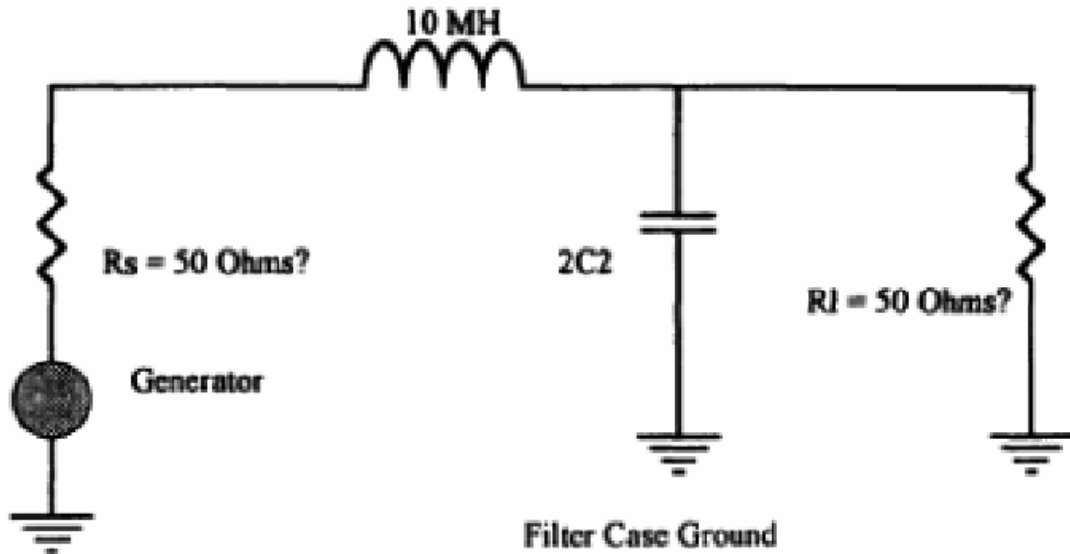


Figure 7.10 The converted test method to calculate the common mode values.

The Δ determinant of the matrix is next:

矩阵的 Δ 值如下:

$$\Delta = \frac{LCR_l S}{C} \left[S^2 + \frac{(R_s R_l C + L)S}{LCR_l} + \frac{R_s + R_l}{LCR_l} \right] \quad (7.5)$$

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

将最初的 V_{in} 和 0 值带入上式中,解 Q_2 得:

$$\begin{bmatrix} \left(LS^2 + R_s S + \frac{1}{C} \right) V_{ins} \\ -\frac{1}{C} 0 \end{bmatrix} \quad (7.6)$$

and

$$Q_{2s} = \frac{V_{ins}}{C\Delta} = \frac{V_{ins}}{LCR_l S \left[S^2 + (R_s R_l C + L)S / LCR_l + (R_s + R_l) / LCR_l \right]} \quad (7.7)$$

but Q_2 is not the goal, V_o is . In reality, the goal is the ration V_{os}/V_{ins} , where

$$Q_{2s}S = I_{2s} \text{ and } I_{2s}R_l = V_{os} :$$

但是我们需要解并不是 Q_2 ，而是 V_{os} 。实际上我们要解的是 V_{os}/V_{ins} 的比值。其中 $Q_{2s}S = I_{2s}$
 $I_{2s}R_l = V_{os} :$

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

This equation has been published in many articles, but most often the authors do not include R_s , the source impedance, and often do not allow for the two feedthrough capacitors, in parallel, which are now doubled in value.

许多技术文献都给出了此等式，但多数没有考虑源阻抗 R_s ，并常常没有考虑那两个并联的穿心电容（此处等效后的单个电容值为原值的二倍）。

Most solve this by completing the square. This means, Equation (7.8) 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 R_s , making the value of a reduced to $1/(2RC)$, the dampening factor. This also reduces the last term of the denominator to $1/(LC) - \omega^2$.

通常在取平方后求解。这意味着，等式（7.8）的分母需要加上头两项的平方值并从最后一项减去。此项通常会远大于共模应用条件下分母中的最后一项，它就是新的最后一项，值为 ω^2 ，符号为负。按推荐的测试设置方法，这会使得解为双曲函数且损耗非常大。多数情况下会忽略 R_s ，使得阻尼因数值为 $1/(2RC)$ 。同样分母的最后一项也减小为 $1/(LC) - \omega^2$ 。

$$\left(\frac{R_s R_l C + L}{2LCR_l} \right)^2 \quad (7.9)$$

$$a = \frac{R_s R_l C + L}{2LCR_l}$$

$$w = -\sqrt{\left(\frac{R_s R_l C + L}{2LCR_l} \right)^2 - \frac{R_s + R_l}{LCR_l}} \quad (7.10)$$

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

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

通过上式给出如下的解，此处 a 为衰减因子：

$$\frac{1}{LC} (e^{-at} \sinh(wt)) \quad (7.11)$$

This may now look easy, but there is another way to know that the common mode will always be lossy: forget the sin or sinh solution, and set the main denominator Equation (7.8) to

这看上去很简单。另外还有一种方法可以证明共模常常是损耗型的：忘记 sin 或 sinh 形式的解，现在把分母的等式（7.8）设成

$$(S + a)(S + b) \quad (7.12)$$

So that a+b is equal to the middle term and ab is equal to the end term of the main denominator Equation (7.8), repeated here:

因此 a+b 等于式（7.8）分母的中间项而 ab 等于最后项，此处再次列出：

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

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

这是一个简化的解，且 a 和 b 都包含在同一二次方程中。不管 a 还是 b 都可以指定为符号为正的平方根，但是取 b 比 a 大会比较好。

$$a = \frac{(R_s R_l C + L) - \sqrt{(R_s R_l C + L)^2 - 4LCR_l(R_s + R_l)}}{2LCR_l} \quad (7.13)$$

$$b = \frac{(R_s R_l C + L) + \sqrt{(R_s R_l C + L)^2 - 4LCR_l(R_s + R_l)}}{2LCR_l}$$

and takes the form of
并取如下形式

$$\frac{e^{-at} - e^{-bt}}{b - a} \quad (7.14)$$

The term b-a reduces to
b-a 项简化为

$$\frac{\sqrt{(R_s R_l C + L)^2 - 4LCR_l(R_s + R_l)}}{LCR_l} \quad (7.15)$$

This cancels LC, so the final answer is
这样就消去了 LC，所以最终的解为

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

In the normal test arrangement, R_s and R_l are both 50Ω , and C , because of leakage current, is whatever the specification requires. Remember that the value of the capacitor is doubled here because the two feedthroughs are in parallel and, therefore, add. Now that the common mode is reduced to a simple single L filter, the required value of the common mode inductor can be easily solved. See Chapters 15 and 16.

在通常的测试设置中, R_s 和 R_l 都为 50Ω , 且 C 因为漏电流的缘故无论如何都需要满足规范的限制。记住, 因为两个穿心电容并联的缘故, 此处电容的值是加倍的。现在共模电路简化成单 L 型滤波器, 其所需的共模电感就可以很容易地计算出来了。详细的计算见 15 和 16 章。

§7.5 Common Mode Inductor Used for Differential Mode

§7.5 共模电感用于差模条件

Some filter manufacturers have designed common mode inductors that also function partially as differential mode inductors. This is done as earlier with a very wide winding spacing that generates the leakage inductance needed to provide this differential inductance. Some flux of one coil fails to cut the other, creating this leakage inductance. Another way this is accomplished is with pot cores (Figure 7.11). These use a split bobbin so that some flux in one half of the bobbin fails to cut the other half. This has been expanded to finding two separate bobbins that fit in the core with additional room to place a washer between the two bobbins. This washer is cut or split to avoid the washer acting as a shorted turn. The material of the washer has little to do with the differential inductance that is created. The separation of the two windings causes the leakage inductance.

一些滤波器厂商设计的共模电感也部分地具有差模电感的功能。因为线圈间的间隙会产生形成差模电感的漏感。其中一个线圈的部分磁通不能与另外一个线圈抵消, 这样就产生了漏感。另外一种产生漏感的方法是使用罐状磁芯 (图 7.11), 罐状磁芯使用了一种特殊的分离式绕线轴, 所以其中一半绕线轴的部分磁通不能与另外一半相抵消。当然这还需要寻找安装到磁芯上的分离绕线轴, 此线轴在两部分绕线轴间有多余的空间来安装垫圈。垫圈常采用切割和分离的结构以防止其成为磁短路通路。而垫圈的材料需要选对形成差模电感不起作用的材料。最后需强调的是, 是两组绕线的分离导致了漏感。

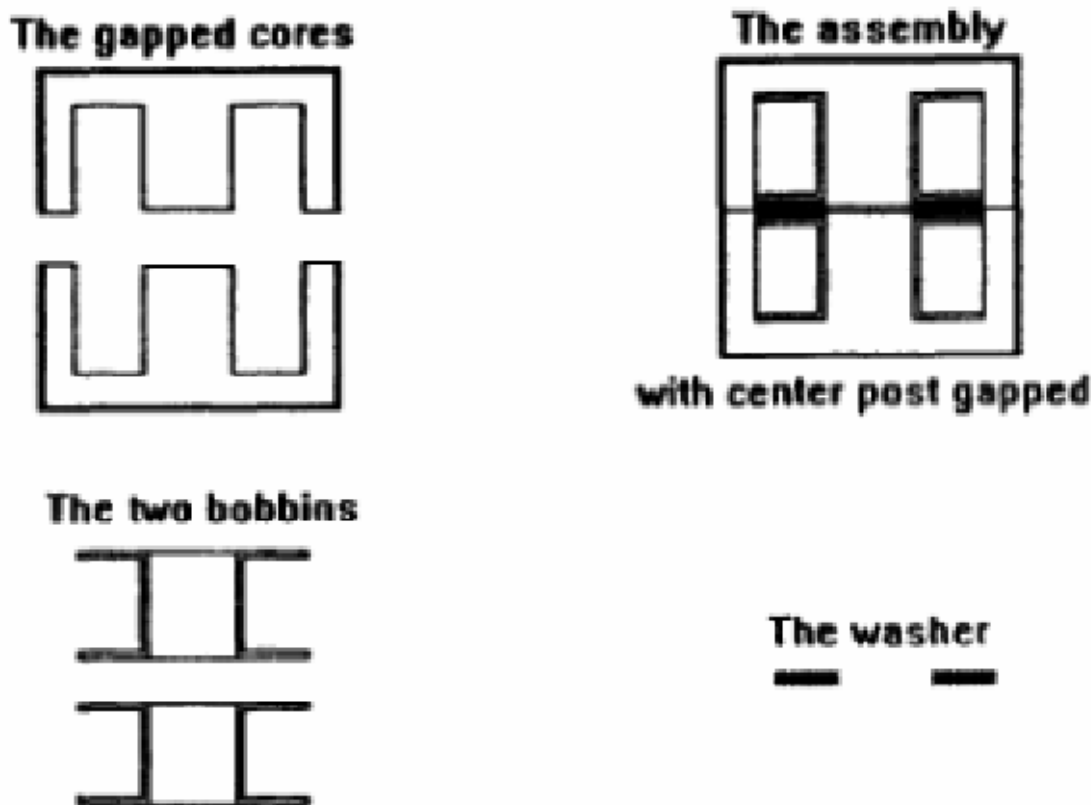


Figure 7.11 The gapped pot core and assembly.

The leakage inductance is easy to measure with an inductance bridge. Shunt one winding of the common mode inductor, and read the inductance of the other winding. If all the flux lines cross or cut the other turns, the reading is zero. This is truly impossible to accomplish because all the turns cannot be this tightly coupled. The difference is the leakage inductance measured by the inductance bridge. Another way is to measure both legs together, opposing, and the bridge reads the leakage inductance.

用电感桥很容易测量漏感。具体方法是，旁路共模电感的一组绕线，并读取另一组绕线的读数。如果所有磁力线穿过另一组绕线，那么读数就是 0。当然这是不可能达到的，因为所有的线圈不能紧耦合，此时的读数就是电感桥读出的漏感。另一种方法是将两组绕线反接，读出的读数也是漏感。

§7.6 Other Wave Shapes

§7.6 其它波形

Some engineers to whom I have spoken think that common mode inductors require a well-balanced sine wave for the common mode to work properly. In the single-phase common mode inductor, the sum of the magnetic fields caused by the currents will still cancel no matter what the wave shape. Because the current is equal and opposite, these two flux fields cancel. If the currents are not equal, for example a current difference created by the capacitors to ground, the two fields almost cancel and give some differential mode inductance. This is

the problem in speaking to purists: most of them claim that feedthrough capacitors unbalance the common mode because they are not an exact match. All this is true, but to what degree? It is also true of any arresters from line to ground because they have different values of capacitance across them and leakage currents through them.

一些工程师认为共模电感需要极平稳的正弦波以使共模电感良好工作。但实际上，在单相共模电感中，不管波形如何电流产生的磁场都将相互抵消。这是因为电流是等幅和反向的，所以磁通会相互抵消。如果电流不相等，如对地电容产生电流差值的情况，这样两个场会近似抵消，而未抵消部分则产生了差模电感。绝对论者会讲：穿心电容之间不精确相等会导致共模电路不平衡。这是对的，但又会在多大程度上如此呢？一些线对地的避雷器也会导致这种情况，因为它们的容值不同穿过的漏电流也不同。

In three phases this is also true: if the phases are 120° apart, the wave shape still need not be sinusoidal. Any current of any shape gives a magnetic field that cancels in the single-phase, three-phase, or dc system. If the reverse is true, the common mode would not work for three phases. Because of the imbalance of the phases and the harmonic content, the voltage is not very sinusoidal: the common mode will not work. The three-phase Zorro requires three windings for the delta and four wires-the added neutral-for the Y. If the three legs are not balanced, the difference current is carried by the neutral and magnetic flux will still cancel because of the fourth neutral winding.

同样在三相条件下也是如此：如果相位差为 120° ，波形同样不需要正弦波。在单相、三相或交流系统中，任何波形、任何电流产生的磁场都会相互抵消。如果位相条件不满足，那么共模电感就不能工作在三相条件下。相位和谐波量不平衡导致电压不是正弦，共模电感就不能起作用。△接法的三相 Zorro 需要三组线圈，对于 Y 型则需要增加零线，在三相不平衡时，差值电流会流过零线，因为零线圈的作用磁通仍然会相互抵消。

8

Electromagnetic Pulse and Voltage Transients

电磁脉冲和电压瞬态

The circuit discussed here consists of a battery, a transmission line, and a switch. The end piece differs: either a short or an open. The battery has an output impedance of Z and equals the impedance of the transmission line. When the switch is closed, a step function travels down the line at the velocity of propagation. The voltage divides between the characteristic impedance of the line and the source impedance. The main interest is what is at the far end of the line. If the line sees an open at the far end, the voltage doubles. The wave travels back, at the same velocity, to elevate each segment of the transmission line to full voltage. At the time the wave reaches the source, or battery, end, the current drops to zero and the full battery voltage is impressed across the transmission line (Figure 8.1).

此处讨论的电路包括一个电池、一条传输线和一个开关。这个电路具有不同的终端条件：短路或开路。电池的输出阻抗 Z 等于传输线的输出阻抗。当开关闭合时，就会有一个阶跃函数以传播速度在线上传播。电压在线特性阻抗和源阻抗间分压。我们主要关心的是线远端的情况。如果线远端开路，电压会加倍。波会以相同的速度返回会将传输线的每一部分的电压都提高到满幅值。在波到达源、终端的时间点，电流下降为 0，并且将满幅值电压施加于传输线上（图 8.1）。

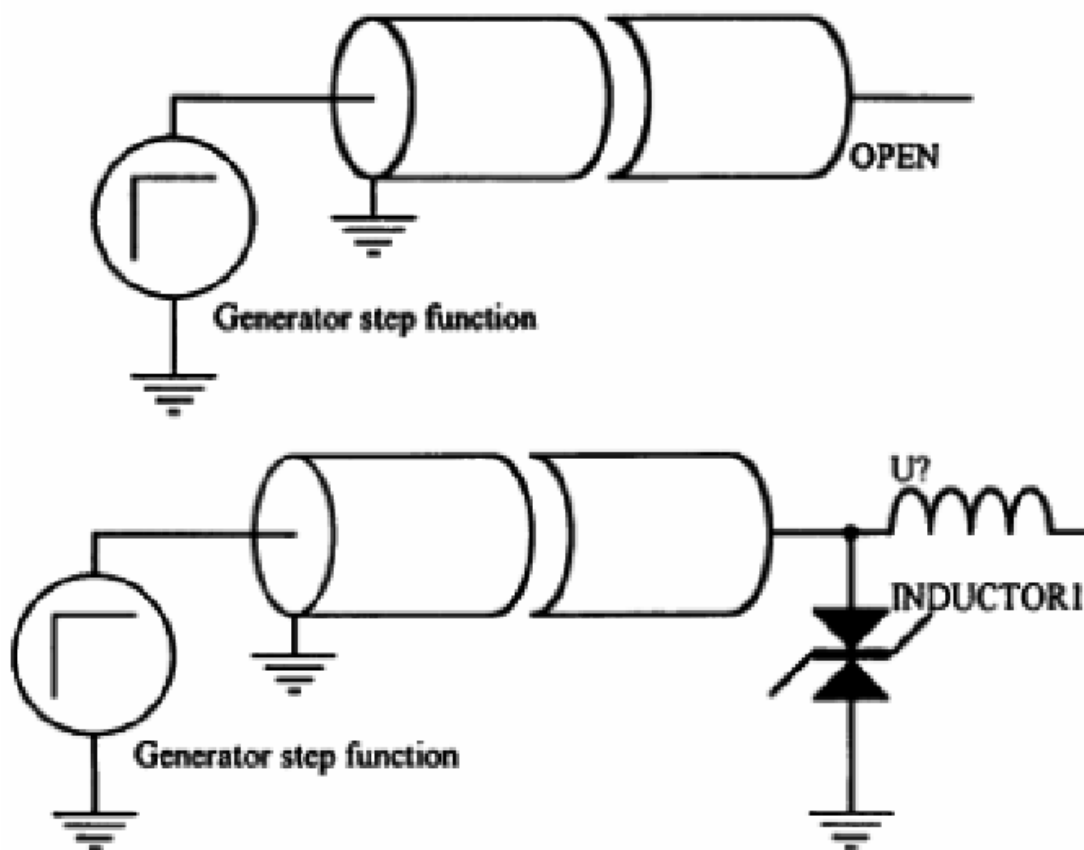


Figure 8.1 The open-ended transmission line applied to the transorb.

If the far end is shorted instead, the current doubles when the wave reaches this far end. One-half of this current depletes the initial stored voltage, segment by segment; the other half continues to flow through the line to the battery Z . This maintains the proper voltage ratios. When the pulse reaches the battery end, the line is fully discharged and the double current flows into the battery, or source. The full voltage is dropped across the internal source impedance Z . If the line and battery Z_s are not equal, the pulses iterate back and forth until equality is reached. The main point here is to see the difference between the two far end conditions, one near open and one near short. In the link, or short line, the source is thought to be very low, and again, the question is what is at the filter end of the line, a shorted condition or a nearly open impedance condition. If the initial condition is a high impedance for the load, the full strike voltage is impressed across the suppressor, aiding turn-on or firing. The device then drops to a low impedance, and a high current passes through the suppressor. The suppressor and the line impedance dissipate the energy—we hope. If the energy is too great for the suppressor, the suppressor normally first shorts and then blows open.

如果远端短路，那么当波到达远端时电流加倍。一半电流将耗尽传输线每个单元原来的储能电压，另一半持续从线到电池 Z 流动。这将保持特有的电压比。当脉冲到达电池端，线完全放电，并且有两倍的电流流入电池，且满幅值电压施加在源内阻 Z 上。如果线和电池

内阻 Z_s 不相等, 那么波会多次反复传播直到达到电压平衡。此处的重点是远端条件的差别, 一个是近似短路, 一个是近似开路。问题是, 在线的终端是滤波器时, 在短路或断路的阻抗条件下, 情况会如何。如果初始条件是高负载阻抗, 那么冲击电压就会施加于干扰抑制器上并产生火花。之后负载降为低阻抗, 就会有很大电流流过干扰抑制器, 线和干扰抑制器会消耗这部分能量, 但对于干扰抑制器来说如果此能量太大, 干扰抑制器会先短路后熔断。

The power in the pulse is dissipated in the MOV, line, and source impedance. If the condition is near short, the voltage divides, according to the ratio, and delays the firing of the protector (Figure 8.2).

通常脉冲的能量被 MOV、线和源阻抗消耗。如果条件接近短路, 依照电压比率电压将被分压, 并延迟保护器燃烧 (见图 8.2)。

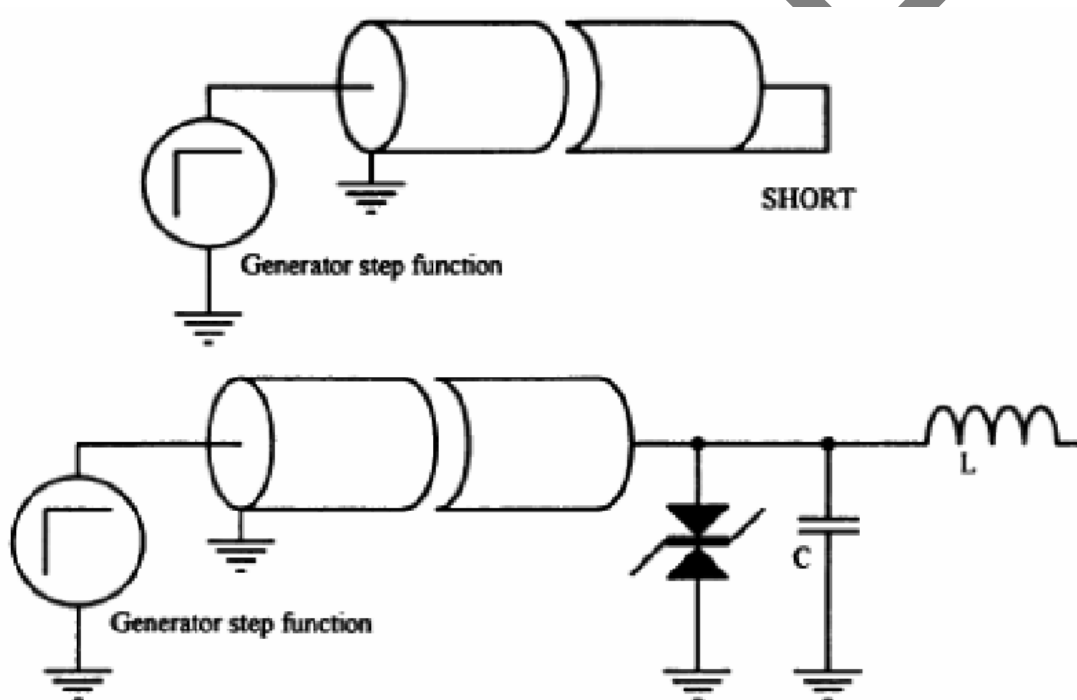


Figure 8.2 The shorted transmission line applied to the transorb.

Some engineers place capacitors across the transorbs, or MOVs, making the condition similar to the shorted condition just discussed. Transorbs and MOVs are much faster on turn on, these days, but the capacitor must first charge to well past the turn-on voltage before the arrester can act. Another point is that the current through the capacitor is of the order of twice the initial line current, or several 100A. All this action slows, or delays, the turn-on of the transorb. This can also blow the capacitor, especially if the capacitor is the metallized film type, if it was not rated for this amount of current, and again if the capacitor voltage rating was exceeded. All this can stress not only the capacitor but also all the remaining filter components. Some engineers argue that all they are trying to do is to protect the arrester, but when a high-energy pulse occurs, both

the capacitor and arrester can be blown. If the filter is designed to withstand this stress, it may well hold up, but this also adds cost to the filter. If the far end (filter input end) is open-transorb is directly across the line followed by an input series inductor-the voltage will rise. The voltage will not necessarily double; the impedance ratio is not known. The voltage will help the arrester to fire, speeding turn-on. The transorb may blow, but the filter and the following equipment will be protected. Most of these arresters conduct in less than a microsecond under these conditions, so the peak current through the inductor and the voltage stored across the following filter capacitor are reduced.

一些工程师会在 transorb 管或 MOV 管上并电容,使得终端条件类似于刚刚讨论过的短路状态。现今的 transorb 管和 MOV 管的开关速度非常快,这首先就要求电容要在避雷器动作前能导通开关电压。另外,通过电容的电流会是原来线电流的二倍的量级,或者能够达到几百安培。另外,所有这些电容的作用都会减慢或延迟 transorb 管的开关速度。如果电容不能承载这个量值的电流或超过了电容的额定电压,那么电容器就会发生爆炸,金属化薄膜类型的电容尤为如此。所有电流和电压超过额定值的情况都会对电容器或其它滤波器元件有的影响,一些工程师可能会争辩说这样设计是为了保护避雷器,但是当高能量脉冲出现时,电容器和避雷器都会烧断。如果设计可以承受此苛刻应用条件的滤波器,虽然滤波器可以很好工作,但是同样也会增加滤波器的成本。如果远端(滤波器输入端)是开路状态——transorb 管直接并在线间,其后是输入串联电感器——那么电压会上升。但因为阻抗比是未知的,所以这个电压值未必会加倍。此电压将导致避雷器燃烧,也会加速开关速度。当然这个 transorb 管可能会烧断,但是滤波器和后面的设备都会得到保护。在这种使用条件下,多数避雷器动作时间会少于一个毫秒,所以通过电感器的峰值电流和储存在滤波器电容的电压都会减少。

§8.1 The Three Theories

§8.1 三个理论

The first theory is from the purists. They say that the arrester is not necessary, and it is the job of the filter to handle these pulses of energy. This might be true if the components were designed to handle them. In this case, the inductor must be designed to handle the full pulse voltage and pulse duration without arcing. The following capacitor must withstand twice the pulse voltage without blowing. The inductor must be wound with spacings between the turns or with very heavy insulation. To calculate this, the total number of turns and the maximum spike voltage must be known:

第一个理论源自于纯理论者。他们宣称不需要避雷器,处理这些脉冲能量是滤波器的工作。如果把元件设计成能处理这些脉冲的话,这样说或许是正确的。在此例中,电感的设计必须保证能处理满幅值脉冲电压和持续脉冲而不能出现电弧现象。接下来的电容必须能承载两倍于脉冲电压的电压值而不发生烧断。电感的绕制要保证匝间距或进行强绝缘。为了计算绝缘强度,需要总匝数和最大尖峰电压等参数:

$$\text{Maximum spike} = 10,000\text{V} = V_{\text{max}} \text{ turns} = 50 = N \quad (8.1)$$

The volts per turn is

每匝电压为

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

Most wire insulation will withstand 500V and should not require special insulation; the turns, here, could touch each other. This is easy to measure. Take the length required for the inductor, and strip one end. Bury this length of wire, with both ends exposed, in a basket of shot pellets (the container is conductive). Apply the test potential to the wire end that was stripped and the other lead to the shot container. What happens if there are only 10 turns?

多数的导线绝缘层都能经受 500V 的电压所有并不需要特殊的绝缘处理；此处每匝线圈可以相互接触。这很易于调节。取电感需要的线长，剥开一端。将这个线长的线圈埋入装小球的篮子中（里面装的是导电物），并将两端露在外边。将测试电压设备的一极施加在剥开的一端，另一极加在小球上。如果线圈只有 10 匝时会发生什么？

The volts per turn is

每匝电压是

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

This is marginal, so special insulation should be used, and as an extra margin of safety, the turns should not touch. Granted, the wire should withstand 1000V from wire to wire, or turn to turn, but this is too risky. This spike inductor should also have only one layer or be wound once around the core and raised from the core to protect the windings from arcing to the core. Many inductors are wound on forms without cores and are wound in or on a solenoid form. In the 10-turn case, these techniques must be employed, but for extra safety, the 50-turn case, some extra measures should be employed: use better wire insulation, keep the turns apart so they cannot touch, and keep the turns as far as possible away from the core, if any (Figure 8.3).

这个数值超过了限制，所以需要特殊的绝缘措施。作为安全余量，同时也需要匝间不能有接触。合乎要求的线需要能承受的线间或匝间电压为 1000V，但是这样还是太危险。另外，这个处理尖峰的电感还应在磁芯上只绕一层，并增加线圈和磁芯的间距以防止线圈和磁芯间产生电弧。基于这个原因许多电感采用无磁芯的绕线形式或以螺线管的形式绕线。在这个 10 匝的例子中，这些方法都需要采用。但对于 50 匝的例子，为了保持一定的安全余量，还是需要这些额外的方法的：使用良好绝缘的线，使匝间远离以保证彼此不会接触，如果有磁芯时就需要让绕线尽量与磁芯远离（见图 8.3）。

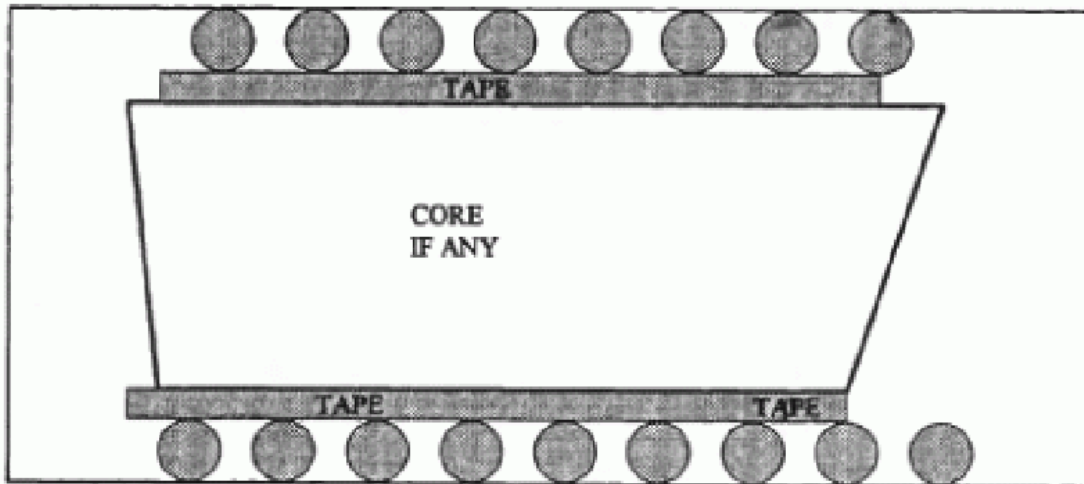


Figure 8.3 The solenoid, or core with tape, showing insulated wire with spacing

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

第二个理论是为了保护设备，而不是滤波器。所以避雷器置于滤波器的接设备的端口。这些工程师还争辩道，他们并不希望电源线上的每个脉冲都启动避雷器动作。问题是，“如果脉冲使滤波器烧断，那么设备仍然可以正常工作么？”这些工程师会用另一个问题来回答：“如果 transorb 管烧断，设备还可以正常工作么？”对于他们的问题，答案或许是可以。原因有二。

1. The arresters often blow shorted; then with the high current through them, they then blow open. If the arrester is then open, the equipment can still function.

1. 避雷器常常对很快熔断，如果大电流通过，那么避雷器会熔断。如果避雷器开路，设备就可以正常工作。

2. I like placing the arresters on the outside of the filter at the input. With proper access to the filter, these blown units, shorted or open, can be cut out and the equipment can operate until the new arresters can be installed.

2. 我喜欢将避雷器放置在滤波器外边的输入端上。并适当接近滤波器，这些熔断器件短路或短路时可以切断电源，所以等安装上新避雷器时设备还可以正常工作。

The third group locates the arresters at the front end of the filter to protect the entire equipment. The opening comments also pertain to this solution. I like the MOV at the front end with an inductive input to follow as part of the filter (Figure 8.4).

第三个理论是，将避雷器放置在滤波器的前端以保护整个设备。开路属性的避雷器适合此

解决方案。我更倾向于在 MOV 的前端后面加一个输入电感,这个电感是滤波器的一部分(见图 8.4)。

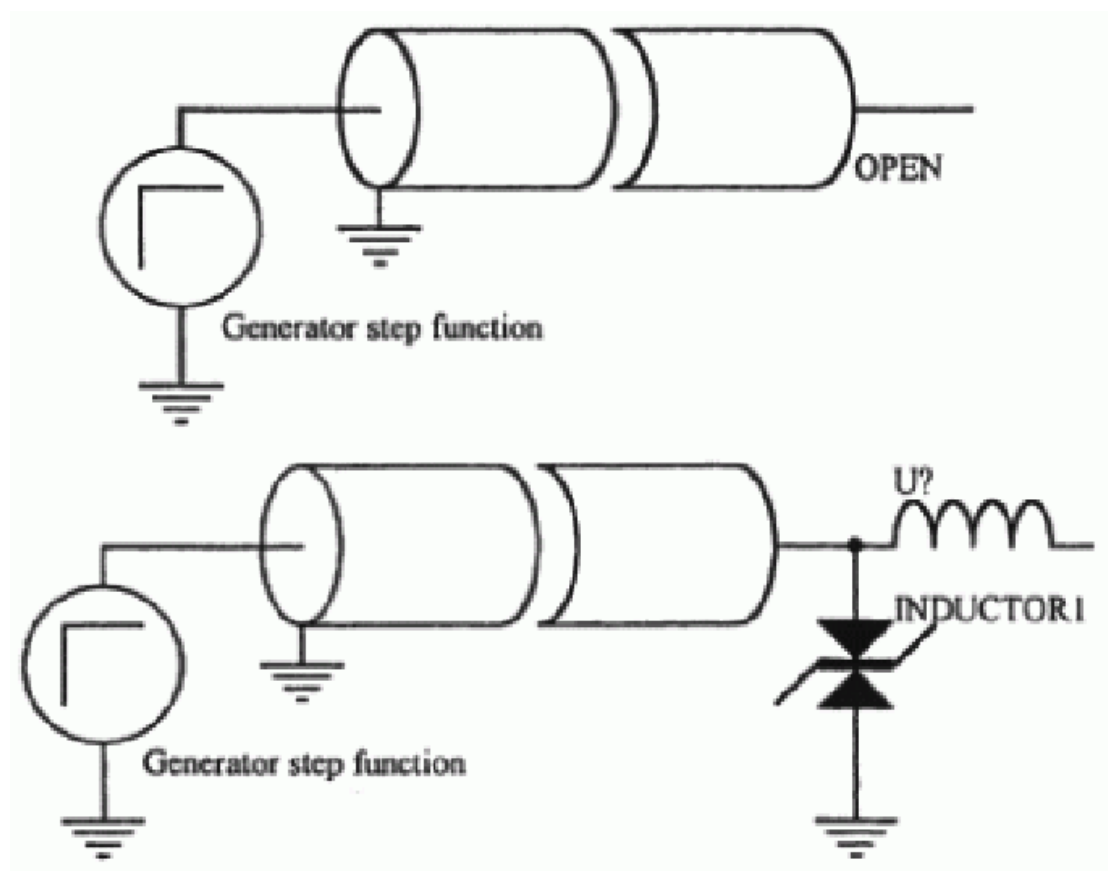


Figure 8.4 The open-ended transmission line applied to the transorb.

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

当脉冲到达避雷器和电感时,脉冲看到的电感阻抗为高阻抗,且电压上升,并启动 MOV。原来的线电流持续通过避雷器,且脉冲被 MOV 和线阻抗所消耗。滤波器的电压为 MOV 的钳位电压。同样,后面的电容应能承受至少两倍于 MOV 或 transorb 管的钳位电压。

There is a fourth group, a combination of groups 1 and 3. They want the input inductor split, with the arrester tied to the junction of the inductors (Figure 8.5). The first half helps to limit the arrester current but must be able to withstand the pulse. This first inductor is often wound on a bobbin without a core to eliminate the potential of any arcing to the core.

将理论 1 和 3 结合就形成了第四个理论。他们希望将输入电感拆分,而避雷器就连接在拆分后的电感间的连接点上(图 8.5)。这样第一半可以限制避雷器电流,但需要能承受脉冲。这个电感常常绕制在线轴上,而没有使用磁芯。这样就可以消除线圈和磁芯间可能产生的电弧。

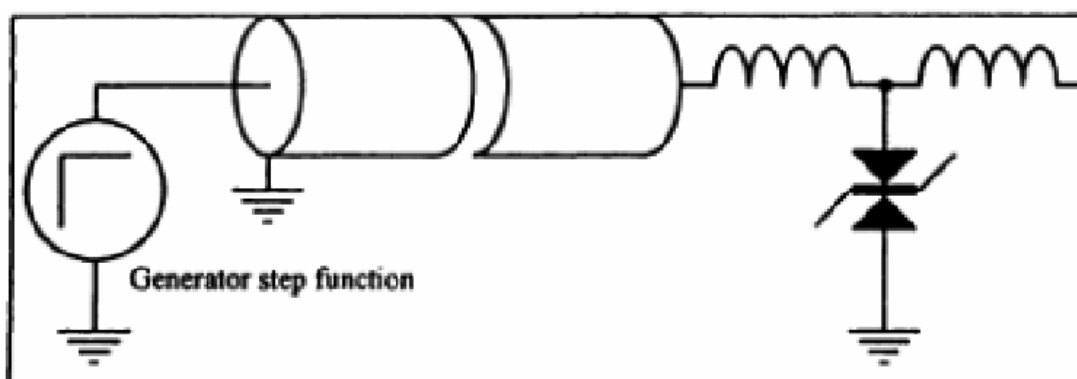


Figure 8.5 The split input inductor to limit the transorb current.

§8.2 Location of the Arrester

§8.2 避雷器的位置

From the preceding section, it should be clear that I prefer the MOV at the head end of the filter, the filter having a series inductor at the filter input (Figure 8.6). I also prefer MOVs to be mounted outside the filter, where there should be access to change them if the need ever arises.

从前述的小节中可知，我倾向于 MOV 置于滤波器的前端，且在滤波器的输入端滤波器有一个串联电感（见图 8.6）。我同样倾向于 MOV 安装在滤波器外部，这样就可以在需要时更换它们。

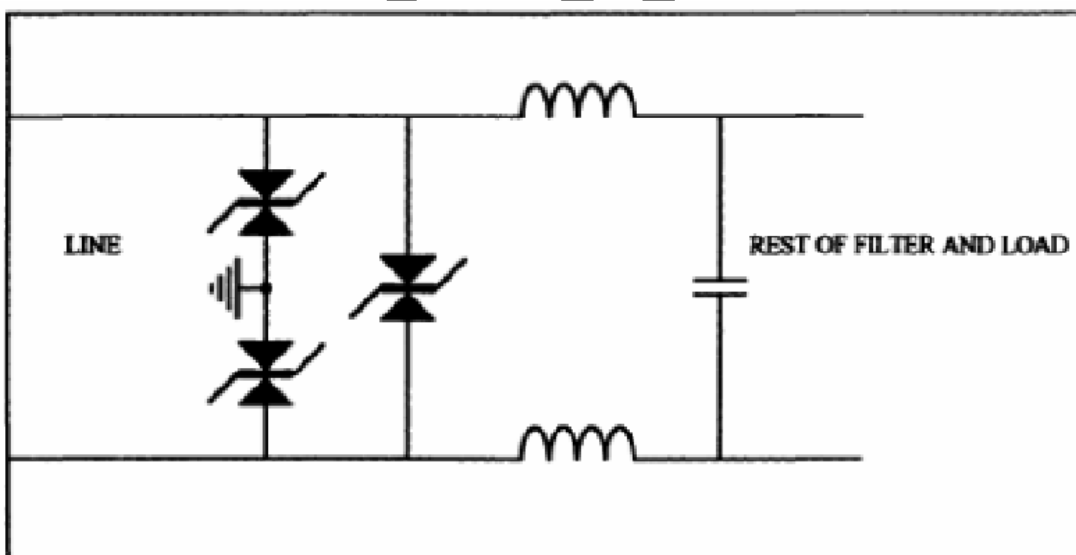


Figure 8.6 The Single phase with arresters.

In the single-phase balanced circuit, three transorbs are required: one from hot to ground, one from neutral to ground, and one from line to line. The first two protect the equipment from common mode pulses-pulses from both lines to ground. The last is any pulse between the lines. In the three-phase filter, in which all three lines are treated within the one-filter enclosure, six transorbs are

required. Three are wired from the three lines to ground, and three are wired from line to line. If the units are internal, the entire filter must be replaced after an electromagnetic pulse or other pulse that blows the transorbs. This can create an inventory problem requiring expensive filters in stock. The other method, in which the arresters are located outside, the blown arresters can be cut out. The equipment can be operated without protection in an emergency until the transorbs can be replaced.

在单相平衡电路中，需要三个 transorb 管：一个火线对地，一个零线对地，一个线对线。头两个保护设备不受线对地的共模脉冲的干扰。最后一个用来消除线间脉冲。在三条线都在一个外壳里的三相滤波器上，需要 6 个 transorb 管。3 个分别对于三线对地，另 3 个安装在线间。如果这些 transorb 管安装在滤波器内部，那么在电磁脉冲或其它脉冲熔断了 transorb 管后就需要更换整个滤波器。这就有需要（昂贵的）滤波器库存的问题。而对于避雷器置于滤波器外部的情况，熔断的避雷器则是可以拆除的，而设备还可以在紧急的情况下在没有保护的条件下工作，直到更换新的 transorb 管为止。

§8.3 How to Calculate the Arrester

§8.3 如何计算避雷器

The specification of the shape and size of the strike is usually furnished by the end user of the device. The job of the filter designer is to calculate the joules needed. The style of transorb is selected, and the clamp voltage for the peak current is found from the transorb specification sheet. From this and the various shapes, the joules are calculated. Once the joules are known, check that the transorb initially selected can handle this amount of joules. The method follows. The peak current is 50A, and the circuit is three phase, with 208V line to line and 120V line to ground. Six transorbs are needed, with three from line to line and three from lines A, B, and C to ground. The latter three can be changed to fit the 120V line, say 150V, but I recommend keeping all the same because production people do not know the difference. The radial lead is selected because of ease of installation. The voltage is specified for the root-mean-square voltage, so 230 is the lowest we can use; 250 is the next lowest. With the 230V transorb, the clamp voltage per the catalog is 700V at 50A. Next is the shape of the wave, also given here. Next, we need the shape factor K (Figure 8.7).

通常是由设备的终端客户给出冲击的波形和幅度的说明。此时滤波器设计者的工作就是要计算出需要的焦耳数。如果选定了 transorb 管的类型，就可以在规格书上查到峰值电流下的钳位电压。通过这个参数和不同的波形可以计算出焦耳数。一旦知道了所需的焦耳数，就要检查原来选择的管子是否可以承载此焦耳数。方法如下，峰值电流是 50A，电路是三相的，并且线间电压为 208V，线对地为 120V。此时需要 6 个 transorb 管，3 个线间另 3 个对地。后 3 个也可以换成适合在 120V 线电压下使用的管子，如 150V，但我推荐使用同样的管子，因为生产的人很难区分这两种管子的差别。另外为了便于安装，需要选用轴向引线的管子。电压是用均方根值标定的，所以 230 是我们能选用的最低值，下一个可以选用的低值为 250。230V 的 transorb 在 50A 条件下的钳位电压是 700V。接下来需要波形，此处亦已经给出。再接下来我们还需要形状参数 K（图 8.7）。

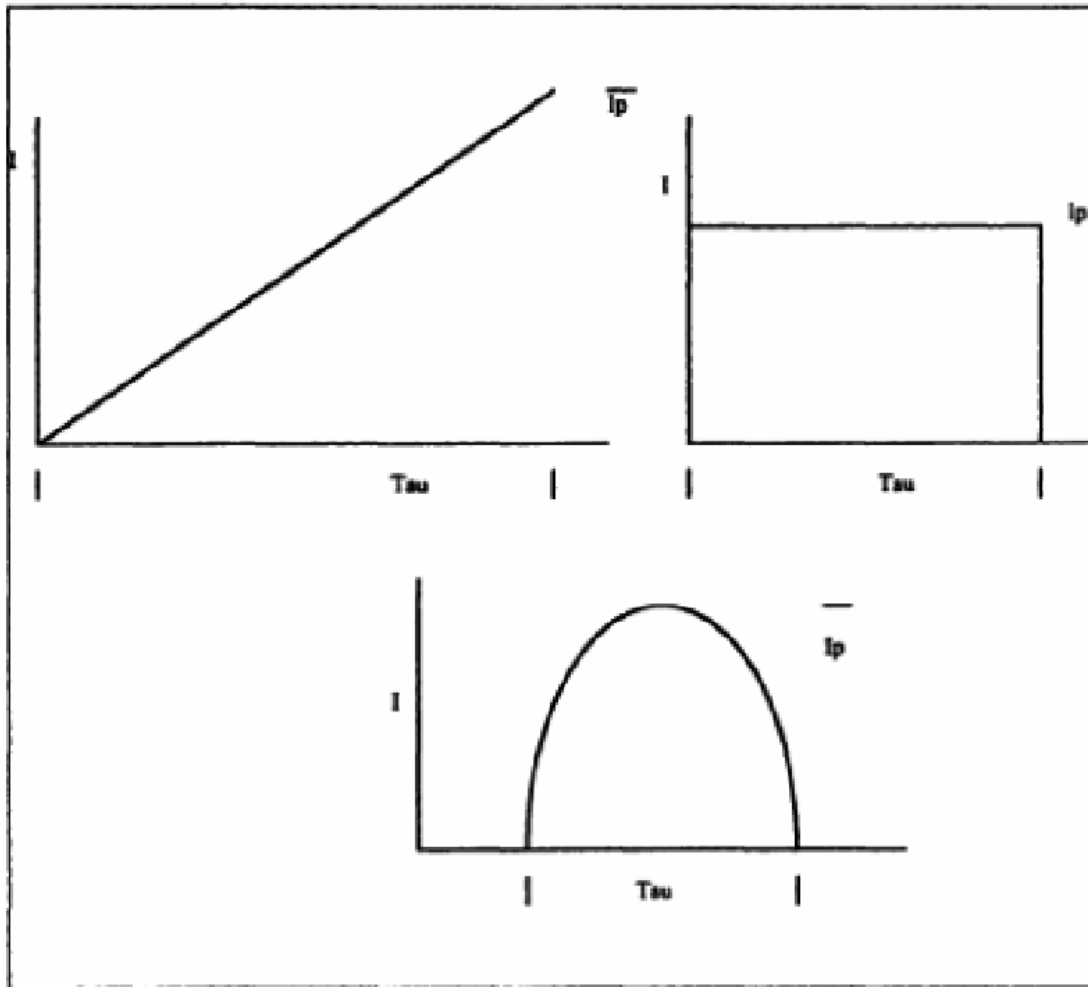


Figure 8.7 The k factor calculated using the full time of the pulse.

The K factors for the pulses in Figure 8.8 are as follows: ramp = 0.5, a constant height = 1.00, and sine pulse = 0.637. The K factors based on the pulses in Figure 8.8 are as follows: a dampened sine wave = 0.86, and a dampened exponential = 1.4. If the drawing or specification has a ramp followed by the dampened exponential with the time of 5 μ s for the full ramp time and a total of 50 μ s to the 50% current point of the exponential (Figure 8.9), the calculation is as follows:

图 8.8 中脉冲的 K 参数如下：斜线=0.5，幅值为常数=1.00，正弦脉冲=0.637。图 8.8 中基于脉冲的 K 参数如下：阻尼正弦波=0.86，阻尼指数=1.4。如果图中或规范中的斜线具有以下阻尼指数：满幅斜线时间 5 μ S，到 50%电流点总时间为 50 μ S（图 8.9），计算结果如下：

Energy of the damp =

$$kV_c I_p t = 0.5 \times 700 \times 50 \times 5 \times 10^{-6} = 0.087$$

Energy of the dampened wave=

$$kV_c I_p t = 1.4 \times 700 \times 50 \times (50 - 5) \times 10^{-6} = 2.205$$

where the time is the difference. This gives a total of 2.292. Multiply this by 1.2 for a safety factor:

此处时间是不同的。上两式给出的总能量为 2.292。将这个结果乘以 1.2 的安全参数：

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

and round this to 2.75 joules. The joules are listed as 20 W · s or J, giving good headroom. This was calculated using a Harris V230LA4.

并四舍五入为 2.75 焦耳。列出的焦耳值为 20 W · s 或 J，这给出了很好的余量。这个结果是用 Harris V230LA4 来计算的。

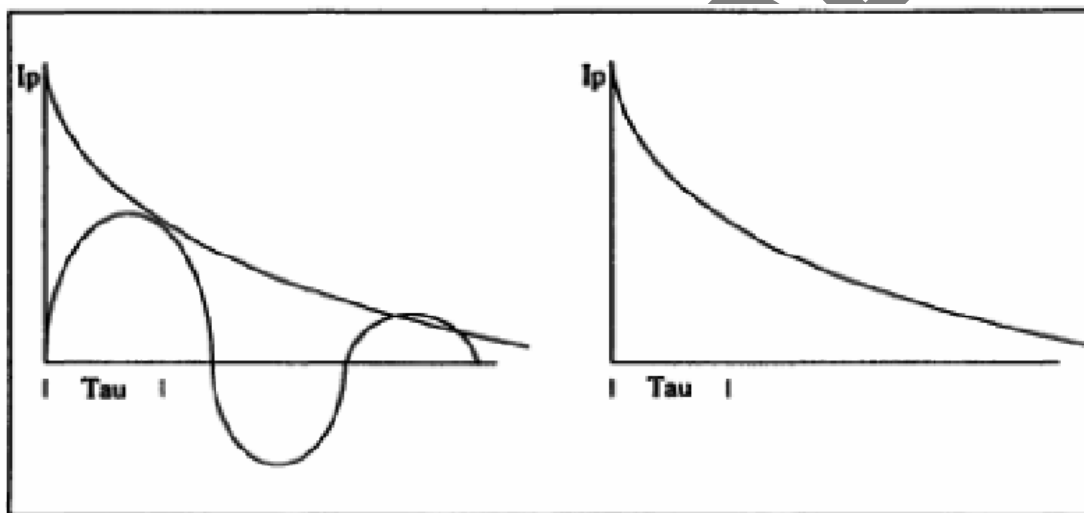


Figure 8.8 The k factor based on the time to 50% of peak current.

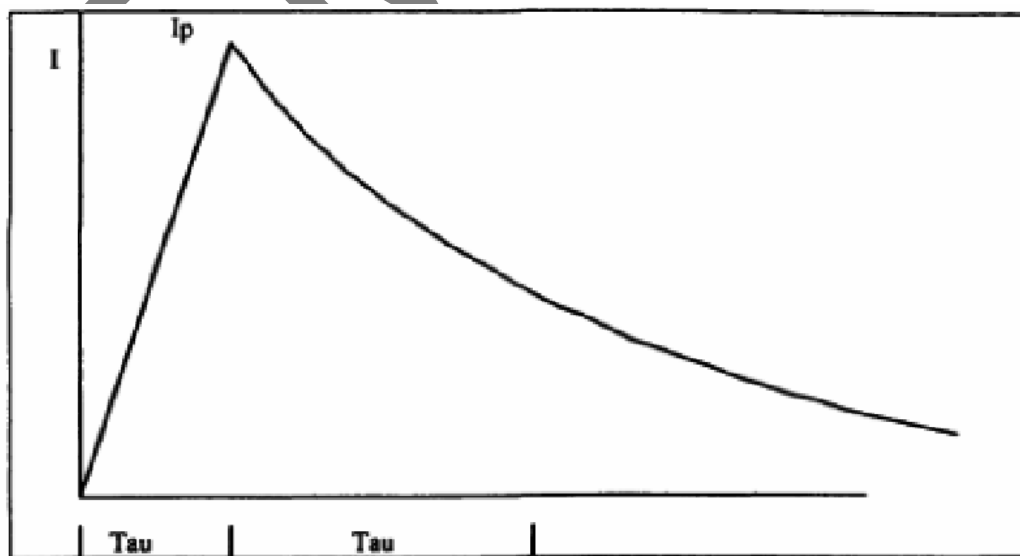


Figure 8.9 A typical pulse calculation.

9

What Compromises the Filter?

是什么损害了滤波器的性能

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. As mentioned in an earlier chapter, some filters are compromised by ground faults. The case we discussed was a balanced filter with the bottom half shunted around. This reduced the loss by 50% and was caused by the obvious lack of communication between two different engineering groups, doubling the weight, size, and cost while halving performance. More candidates for failure are discussed here.

滤波器在用规范的方法测试时性能良好，但由于太多的原因实际仍工作不良，此处会讨论最常出现的原因。如早先章节讨论的，一些滤波器由于接地错误会损害性能。我们讨论过的例子是带下一半电路的平衡电路滤波器。这会将损耗减小 50% 并导致在两个不同的工程系统间的出现明显的通信缺陷，并在性能减半的情况下其重量、尺寸和成本却增加了一倍。以下将讨论更多的滤波器失败原因。

§9.1 Two or More Filters in Cascade

§9.1 两个或多个滤波器串联

This happens when more than one filter type follows another (Figure 9.1). For example, in a secure room or screen room, the entire room is powered by a large three-phase filter. Each power leg and neutral has an insert rated to carry 100A or more, and these inserts in turn are enclosed in a larger cabinet. This can be fed to cabinets in the room that powers a full 19 inch rack of equipment through a filter in each rack, and the individual equipment mounted in the rack is also filtered. This situation places three filters in tandem-in cascade. If the power line filter insert feeding the rack requires 100dB at 10kHz, this may force the designer to cut well below the normal recommended cutoff frequency to achieve the proper attenuation or insertion loss for any reasonable cost, size, and weight. The filter in the bottom of the rack may be a single π filter and may be chased by another π filter. 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.

当多个不同类型的滤波器连接时会出现滤波器串联的情况（图 9.1）。例如，在安全室或屏蔽室内，整个房间由一个大的三相滤波器供电。每相引线和零线插入的滤波电路可以承载 100A 或更多的电流，这些在回路中插入的电路都装在一个大的机壳内。它为在房间中的机柜提供馈电，而这个房间中的机柜通过每个机架的滤波器为 19 英寸机柜的设备供电，并且

每个机架内安装的独立设备也进行了滤波。这种情况下前后层叠安装了三个滤波器。如果给机架供电的电源线滤波器的插入损耗要求在 10kHz 为 100dB，这就要求设计者在低于正常推荐的截止频率就要有很好地衰减，以便于在合理的成本、尺寸和重量条件下获得合适的衰减或插入损耗。在机架底部的滤波器可能为一个单 π 型滤波器，之后又连接了另一个 π 型滤波器。这些滤波器会彼此误调，尤其是在部分滤波器或全部滤波器具有 2 以上的高 Q 值电路时。高电路 Q 值增加了这些滤波器振荡的可能性，而这种振荡会将截止频率移入到通常的带通区域。这种振荡导致的问题常常会降低线电压，导致机架设备工作失效。

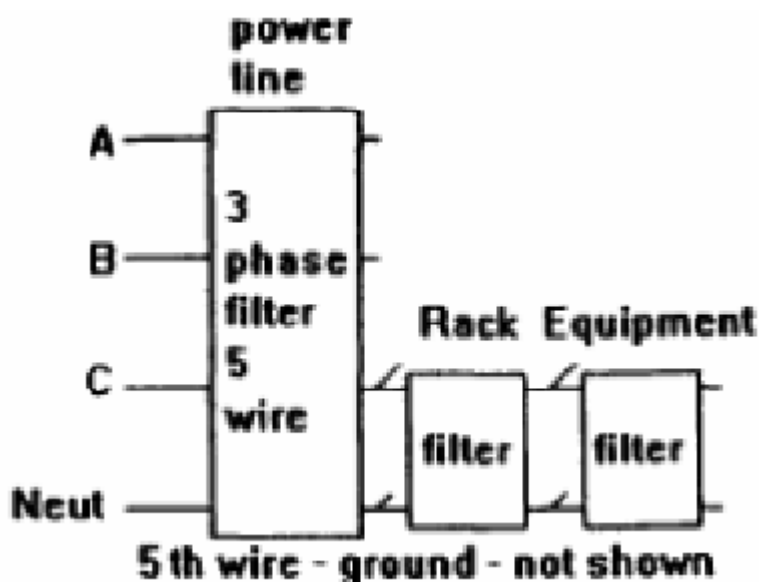


Figure 9.1 Three filters in cascade (in tandem).

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. The problems, as mentioned earlier, this situation can create should now be obvious.

如果这其中的一些滤波器原来就设计错误，这样就会加重问题，如果在同一电源线上前后层叠连接了多个滤波器，问题就更严重了。前后层叠的总电容会导致大的线电流和谐波电流，这样会增加滤波器产生的热量。同样总电容也会在滤波器链中增加一些附加的谐振增益或谐振损耗。如早先我们提出的，在这种滤波器层叠情形下产生的问题，问题的情况就更明显了。

§9.2 Poor Filter Grounding

§9.2 滤波器接地不良

A properly designed filter may check out fine in the electromagnetic interference (EMI) test laboratory or at the EMI filter design house. You may have been in these laboratories and have seen all the grounding techniques necessary to test the filter (Figure 9.2). The test bench is covered with a sheet

of copper that is well grounded. The equipment, of 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 cutout holes in the chassis. If this ground is not provided, the filter fails to live up to its decibel rating. Some people scream that the filter is no good even though they were at the laboratory when the filter was checked. Without a good ground, the filter's feedthrough capacitors and other components to ground cannot work (Figure 9.3). The transorbs and MOVs tied to ground cannot function.

设计的滤波器会比较好地通过 EMI 测试实验室或 EMI 滤波器设计室的测试。或许你在这些实验室工作并已经看到这些实验室测试滤波器所需要的所有接地方法（图 9.2）。测试台覆盖了一层良好接地的铜板。安装滤波器的被测设备常与此铜板 C 型紧密连接。多数滤波器被设计成通过穿心螺柱或连接器直接安装在大地上。滤波器通过底板的孔并两面使用 EMI 衬垫进行安装。这样滤波器就通过螺母和垫圈紧密地安装在大地上。通过此方法使用的衬垫能提供数以千计的接地点，也能通过底板上镂空孔接入底板地平面。如果没有这个大地，滤波器就不能达到其设计性能。一些人宣称，甚至在实验室滤波器测试时就没有良好接地。没有良好接地时，滤波器的穿心电容和其它对地器件就不能正常工作（图 9.3）。另外，接于大地的 tansorb 管和 MOV 也不能正常实现功能。

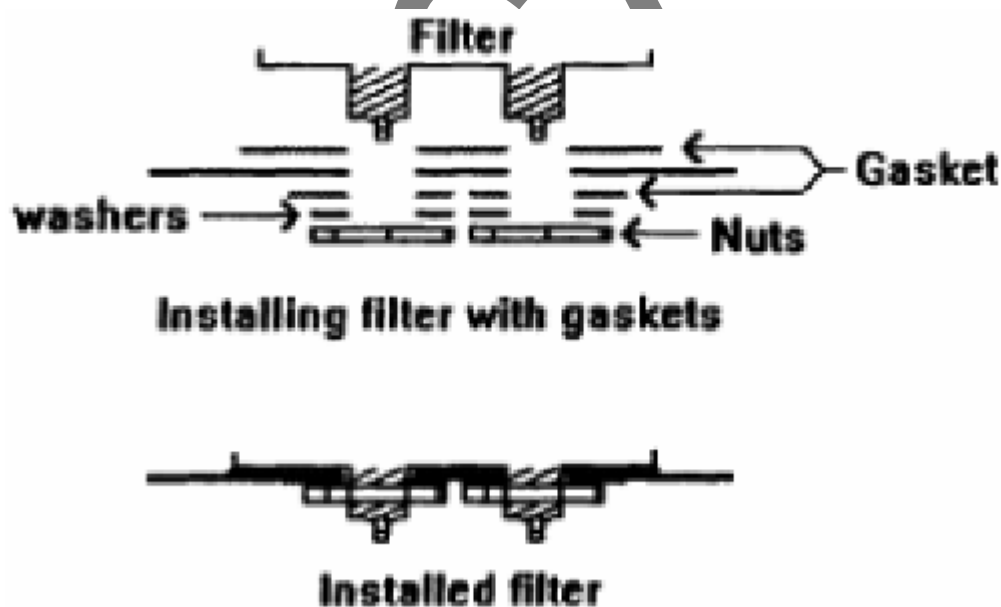


Figure 9.2 Filter with feedthroughs showing the grounding through the mounting.

It is easy to see how the filter in Figure 9.3 does not function properly if the filter case is not grounded or is poorly grounded. The two load-side feedthroughs would be out of the circuit, along with the two transorbs to ground. The two similar capacitors on the input are noncapacitive input terminals or connectors.

如果滤波器外壳没有接地或接地不良，通过图 9.3 可以很容易地看出滤波器是如何不正常工作的。如果接地不良，那么就等效于负载端的两个穿心电容和两个 transorb 管从电路中去掉。另外，两个位于输入端的电容也会被去掉，结果使得输入端口呈非容性。

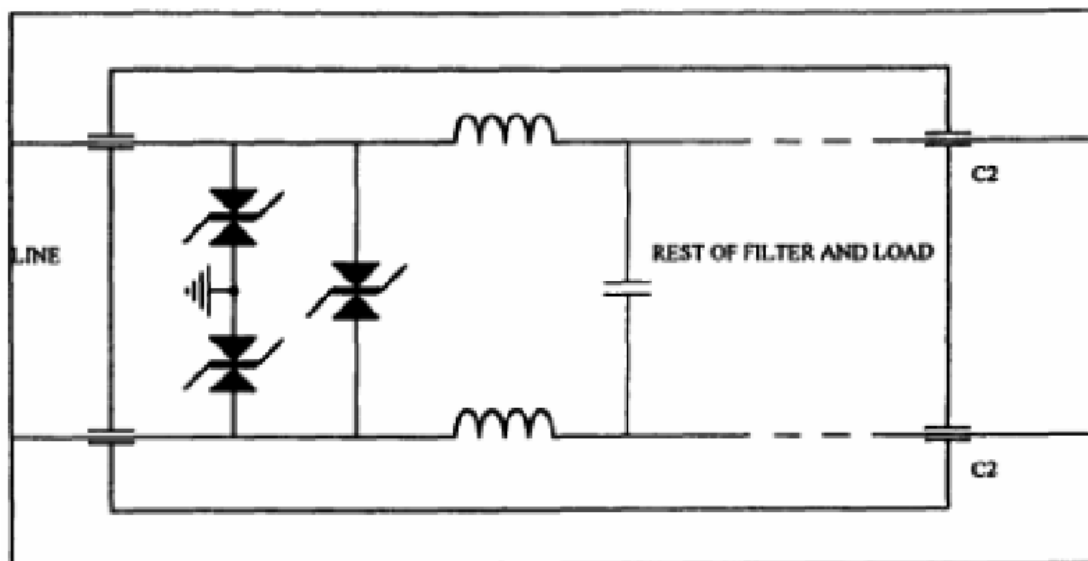


Figure 9.3 The π filter showing the lack of a good ground.

§9.3 The “Floating” Filter

§9.3 “浮地”滤波器

This follows from the preceding topic. The filter was designed according to the method just described. This filter was to be mounted to ground as described, and the users complained that the filter did not function. Sure enough, the filter was mounted or hung in air using a plastic hanger. A 6 inch green wire (normal hookup wire, not Litz) ran from one lug on the filter through a cable harness to a ground (Litz wire could have carried most of the upper frequencies). Common hookup wire acts as a good point to radiate an H field to the surrounding wires in the same cable form. I was never able to induce these users to make the change, although I was able to show some of them the folly of their ways. They were may behind schedule, with the customer pressing them to ship, and could not find time to make these changes. I think they must have gotten a release from the EMI specification from the customer.

本节的内容是前面话题的继续。滤波器是按刚叙述过的方法设计的。这个滤波器按前述的方法接地，但是用户还是抱怨滤波器不工作。毫无疑问滤波器是被安装在塑料架上或悬挂在空中的。一条 6 英寸长的绿线（通常是连接线而不是绞合线）通过电缆束从滤波器的接线片接到大地上（绞合线常能承载更高上限频率）。通常连接线是一个能向同一个电缆束其它线辐射磁场的辐射点。尽管我向他们展示了这种接地方法的缺点，但我还是不能劝说这些用户去改变这种情况。他们按计划时间表工作，因为客户迫使他们按时发货，所以没有时间来做修改。我想他们的客户一定对他们的产品的 EMI 规范做出了宽限。

The green wire just mentioned has the following properties (Figure 9.4):
上面刚刚提到的绿线有如下特性（图 9.4）：

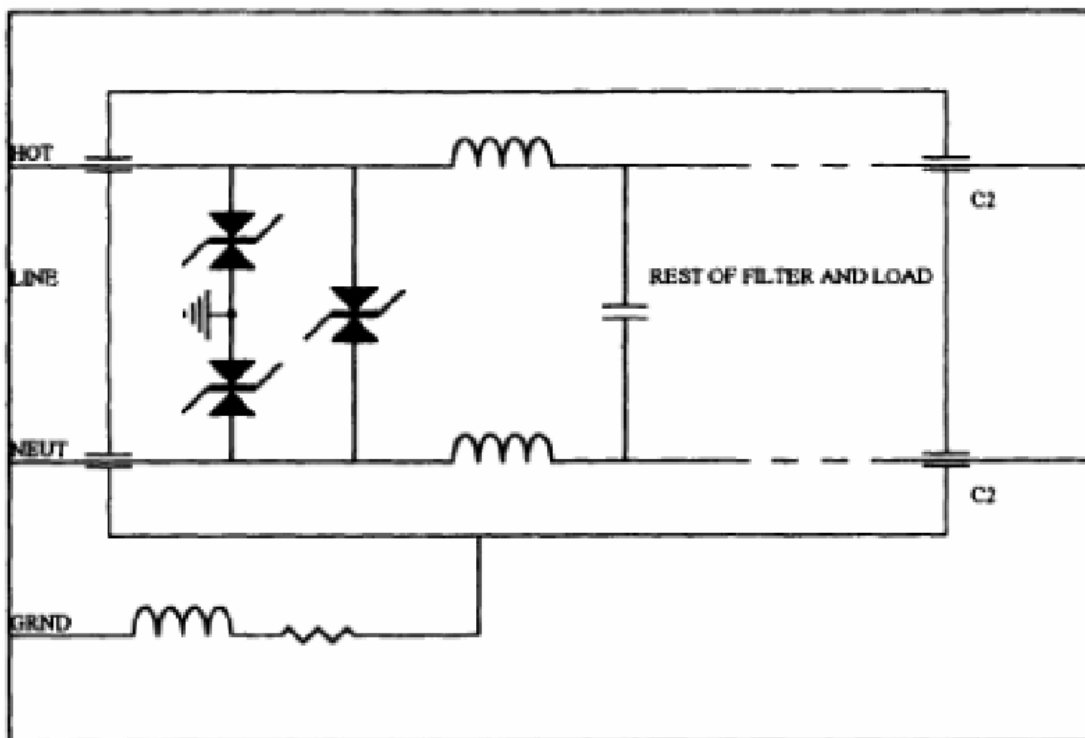


Figure 9.4 The little green wire.

1. A skin effect adds to the ac, or RF, resistance, making the ground more resistive.
1. 趋肤效应增加了 AC 或 RF 阻抗, 使得大地呈阻性。
2. A slow velocity of propagation, making the apparent length about eight times longer, adds inductance to the lead and further impedes the RF current.
2. 较慢的传播速度, 使得实际的长度有原来的 8 倍长, 这样就增加了导线电感并进一步阻碍了流向地的 RF 电流。
3. From 1 and 2, a series RL circuit is formed.
3. 通过 1 和 2, 形成了一个串联 RL 电路。
4. Whatever the RF current in the lead, this would radiate to the surrounding areas and wires.
4. 无论导线中的 RF 电流如何, 该导线都将向周围区域和导线辐射。

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

因为较慢的传播速度, 这条绿线的电感会接近 1.5uH。这导致 6 英寸的线长增加到 8 倍的长度, 接近 48 英寸。此时整个长度就至少等于 1 米, 则此线的典型电感为 1.5uH/m。

§9.4 Unknown Capacitor in the Following Equipment

§9.4 在滤波器后面的设备中的未知电容

This is similar to Section 9.1 discussing filters in tandem and applies only to dc filters. This is also another reason a filter may be praised by one group and cursed by another. Granted, this double capacitor may not compromise every situation in dc filtering, but the additional component adds cost, volume, and weight (Figure 9.5). The added capacitor detunes the filter.

这种情况与在 9.1 节讨论的前后层叠的滤波器相类似，当然这种情况只出现在 DC 滤波器中。这也是一个滤波器会被某一用户认可而却有可能被另一个用户否定的另一个原因。当然，这种双电容的情况或许不会损害 DC 滤波器的性能，但多余的元件却增加了成本、体积和重量（图 9.5）。另外，增加的电容器还会误调滤波器。

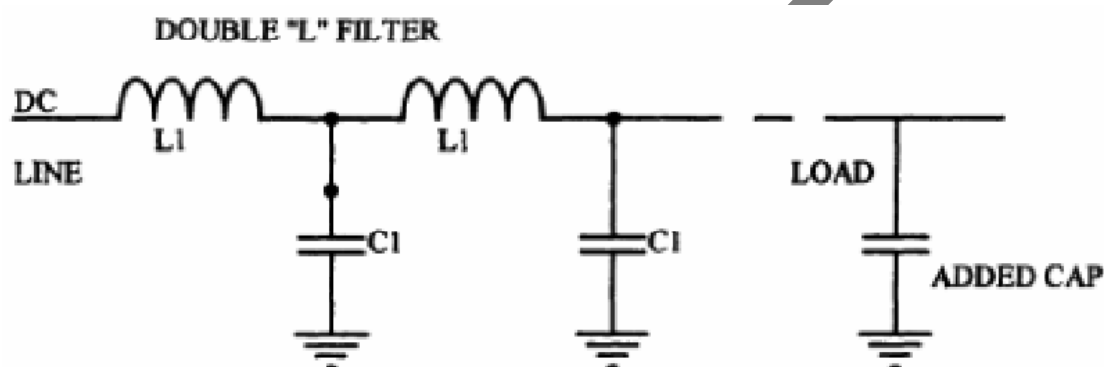


Figure 9.5 The added unknown or hidden capacitor.

§9.5 The Input and Output Too Close Together

§9.5 输入和输出间隔太小

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

我曾看到过将滤波器输入和输出螺柱都置于前面板的滤波器设计。实际应用中应尽量避免这样做：这样输入和输出之间的间隔太小了。这样的面板还有另一个问题就是它大约只有 2 × 3 英寸大，这就意味着滤波器输入和输出的元件空间距离十分近。结果使得设计的 40dB 滤波器因为 RF 从输入辐射到输出（或滤波器整体相互耦合）的原因而只有 26dB。我曾见过滤波器输入和输出都设计在一个连接器上，并且输入和输出线都扎在同一线缆束中的例子。

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 section toward the opposite end. Figure 9.6 shows an input connector

with threads so that EMI gaskets can be used on both sides of the enclosure material. The internal gasket should be the full width and height, and the outside gasket should be the same diameter as the mounting washer. This is followed by the mounting nut. The output terminals and feedthrough capacitors are on the far end.

最好的滤波器物理尺寸设计是采用长体结构，且每个端口都位于相反方向的小平面上。这种结构就使得当信号从滤波器一端通过滤波器向相对端口传播时，不需要的能量会很好地得到清除。图 9.6 显示的是一个带螺纹输入连接器，这样就可以在壳体的两面都使用 EMI 衬垫。在内的衬垫应是和滤波器此端面等宽等高的，在外的衬垫应与安装垫圈的直径相同，接下来安装螺母。输出端口和穿心电容都处在滤波器的远端。

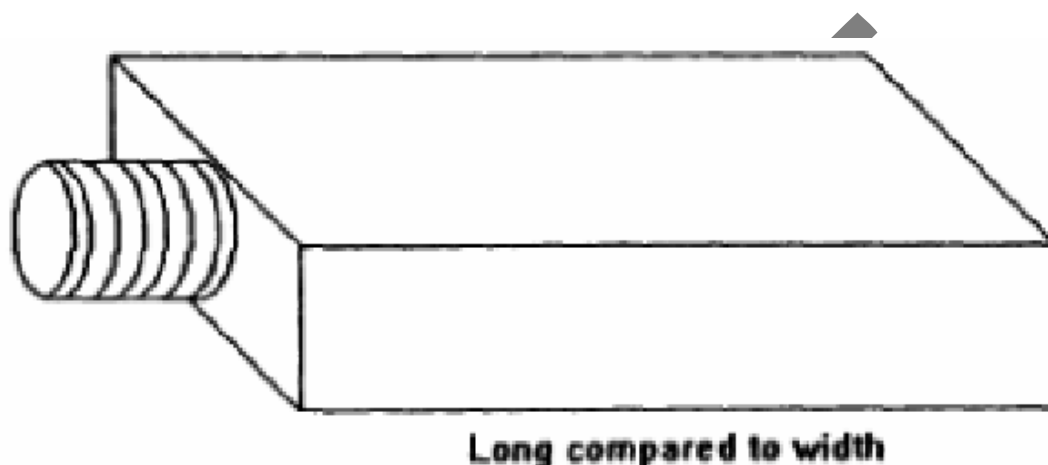


Figure 9.6 Filter using a connector.

If you cannot escape this condition, say in the 2×3 inch section just mentioned, placed a shield between the two halves. Do not allow the harnesses feeding the filter to carry both feeds. There are two ways to divide the 2×3 area, and the choice is determined by the component size. First, divide the 3 inch width by 2, and install a shield that runs well to the other end of the filter. This makes the shield 2 inches high (Figure 9.7). The input or output section is now 2×1.5 inches. The input studs and components should be installed on one side of the filter, and the additional components should be installed toward the rear, then, turn the corner and head toward the front again. If some components are larger than the compartments formed by the shield, the second method of using the shield to split the 2 inch height can be used. Continue with the same technique.

如果你避不开前面所说的 2×3 英寸的滤波器截面的使用条件，那么都要在两半电路间放置屏蔽层。并且不允许馈给滤波器的电缆同时承载输入和输出线缆。有两种方法分离此 2×3 的区域，但最终选择哪种方法是由元件尺寸决定的。第一种，将 3 英寸宽度一分为二，并安装一个与滤波器另一端有效屏蔽的屏蔽板。这样屏蔽层有 2 英寸高（图 9.7）。此时输入和输出部分的截面为 2×1.5 英寸。输入螺柱和元件安装在滤波器的一侧，接下来的元件向后排布，并在滤波器尾部转角后再向前排布。如果一些元件大于由屏蔽层形成的隔间，那么就要用第二种方法了：将 2 英寸高一分为二。接下来同样要按前面的方法安装。

If the components still do not fit, it is best to change the layout so that the filter

can run from end to end without doubling back as in Figure 9.6.

如果元件还装不下，那么最好更改布线，以便于滤波器结构变成一端输入另一端输出，而不是像图 9.6 那样的往返排布形式。

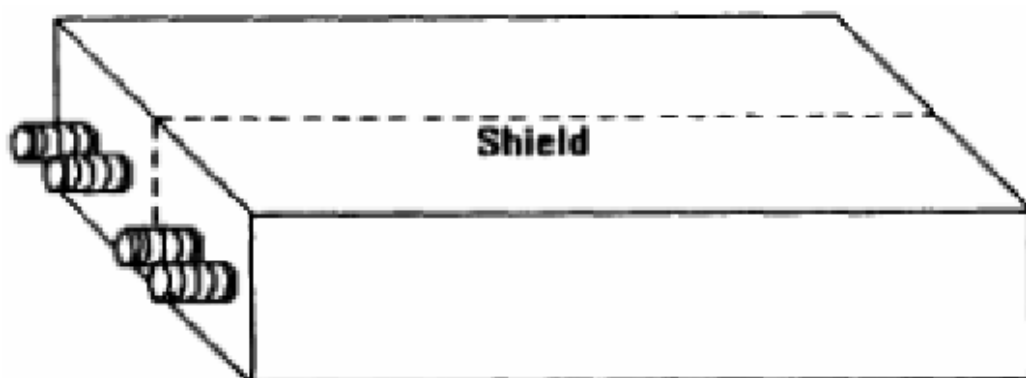


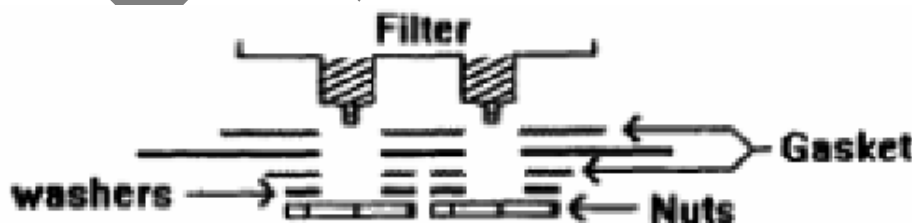
Figure 9.7 Input and output on same filter face.

§9.6 Gaskets

§9.6 衬垫

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 (Figure 9.8). 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. Some times, the filter is removed for various reasons or replaced, and the same gaskets are reused. This again reduces the effectiveness of the ground, and the components to case ground of the filter are less effective. On the other hand, it is better to reuse the gasket than to do without a gasket.

EMI 滤波器常通过壳体上的小孔安装，此时就需要能使滤波器外壳良好接地的衬垫（图 9.8）。通常会遗忘衬垫，这样滤波器的接地回路会较差，因此滤波器内任何穿心电容和对地（或对滤波器壳）MOV 都会受到影响。有时，滤波器由于更换等原因会重复使用同一个衬垫，这同样也会降低接地效果，也同样会降低滤波器对地元件的效能。另外，即使重复使用导电衬垫也比没有衬垫强。



Installing filter with gaskets



Installed filter

Figure 9.8 EMI gasket used to install a feedthrough input filter.

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).

另一个使用衬垫的原因是，可以完善滤波器安装孔导致的地缺陷。滤波器较适宜的安装方法需要两个衬垫：一个位于滤波器和机壳间，另一个位于壳外和垫圈间。

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.

衬垫厂商会提供 EMI 衬垫所需的合适的扭矩要求，这样就能既保障了良好接地又不会导致衬垫压坏。在这个扭矩作用下，滤波器和地间的 DC 和 AC 阻抗最小。

When two objects are tightened as earlier without gaskets or with old, reused gaskets, few points touch, creating a resistive path. When new gaskets are used, thousands of points touch, creating a low-impedance path. Glitsch Special Products in Dallas, Texas is a good source for EMI gaskets of all kinds, including the type used to keep out humidity and salt spray.

当滤波器和壳体间虽结合紧密，但却没有衬垫时（即使连旧的重复利用的衬垫也没有），那么结合面的接触点会很少，因而生成了阻性路径。当使用了新衬垫时，会产生数以千计的接触点，此时会生成一个低阻抗路径。德克萨斯州达拉斯的 Special Products 公司是一个非常好的 EMI 衬垫供应商，它提供所有类型衬垫包括阻燃和防盐雾衬垫。

10

Waves as Noise Source

作为噪声源的波

The waves we discuss 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.

此处我们讨论的波会在电子系统中常常遇到。这些波不会像本章画出的那么简单，因为一些变压器的寄生振荡会叠加在这些波形上。这些波形同样也要考虑加上上升和下降时间，我们此处画出的一些阶跃函数实际是不可能达到的。每个电压或电流的波形都是通过 Fourier 公式得来的，并在本章列出了其区别点。

§10.1 The Spike Wave

§10.1 尖峰波

Spikes are a common waveform, often seen when double-ended, or Royer, circuits are used (Figure 10.1), 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 500-2000uF range. If the spike frequency is 60kHz, twice the 30kHz 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 tied at the junction of the 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.

尖峰波是一种常见的波形，常见于使用双头或 Royer 电路时（图 10.1），即当其它的开关打开时另一个开关关断。这两个电流是相加的，最终导致电流近似加倍。其它元件如二极管等也与此电流相加，结果总的电流尖峰常常会是平均开关电流的许多倍。在每个开关频率周期内有两个尖峰，所以尖峰的频率是开关频率的两倍。在没有校正电路的 DC 系统的开关电路中，这种高尖峰电流能使 DC 电压下降。这会影响开关的正常操作，并且多数尖峰能量会被送返到 EMI 滤波器上。对于带离线调节器的电路，储能电容常常会很大，大约在 500~2000 μ F 范围。如果尖峰频率为 60kHz（2 倍于开关频率 30kHz），这个频率就会远大于其 SRF，结果使储能电容呈开路或感性状态。如果需要此节电路在此频率点不被傍路，那么就应并一个高质量的穿心电容。如果此节被地隔离，那么就需要在储能电容间并一个好的线间电容（X 电容）。另一个放置电容的较佳位置是在电源馈入和开关电源的变压器连接处。如果开关电源不被地隔离时，就应使用高质量穿心电容，隔离时使用高质量的线间电容。

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

简单计算穿心电容或线间电容的方法如下：估算出较低 DC 存储电压，最差的工作阻抗等于此电压除以正常开关电流的二倍（大约是尖峰电流的峰值）。需增加的线间电容的阻抗应是在尖峰频率（此处为 60kHz）下最差工作阻抗的 0.1 倍。线间电容的引线应尽可能短，并保持此电容的 SRF 远大于 10 倍尖峰频率（600kHz）。将计算值近似到高的电容标准值。

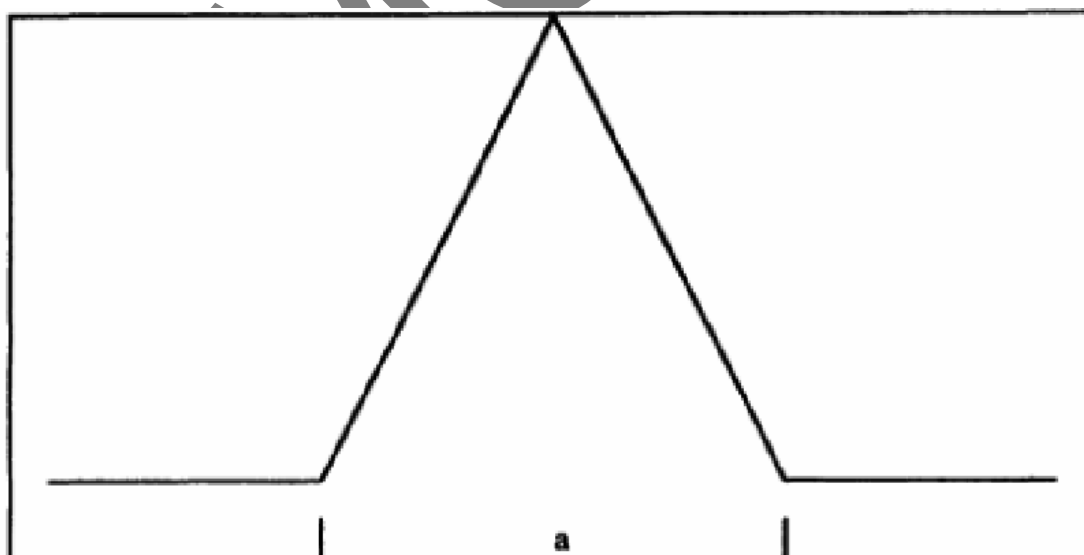


Figure 10.1 The spike wave.

This spike is 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 here. The leads must be short because recent articles published by the Institute of Electrical and Electronics Engineers state that a capacitor with the same lead length will not give the higher SRF required. 这个尖峰是脉冲波，因此它需要能处理高脉冲电流的元件，因此需要将金属箔电容降级使用，而不能使用金属化薄膜电容。此处陶瓷电容会工作良好，但需要短的电容引线。这是因为最近 IEEE 上的一篇文章表明，长电容引线无法得到所需的高 SRF 值。

The equation for the spike is as follows:

此尖峰信号的傅立叶表达式如下：

$$\frac{Ea}{T} + \frac{Ea}{T} \sum_{N=2,4,6,8}^{\infty} \left(\frac{\sin X}{X} \right)^2 \cos \frac{2pNt}{T} \quad (10.1)$$

where a is the pulse width and

此处 a 是脉冲宽度，且

$$X = \frac{2pNa}{T}$$

and the differential is

微分表达式为

$$- \frac{E}{2pa} \sum_{N=1,3,5,7}^{\infty} \frac{1}{N} \left(\frac{\sin X}{X} \right)^2 \sin \frac{2pNt}{T} \quad (10.2)$$

§10.2 The Pulse Wave

§10.2 脉冲波

The pulse is similar to the quasi-square waveform but pulses in the same direction, like the spike (Figure 10.2). 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.

脉冲波与准方波较类似，但脉冲波的脉冲如同尖峰波一样都在同一个方向上（图 10.2）。这与 Royer 类似，其每半波保持时间少于半个时间周期。因此，脉冲波在每个周期生成了两个电流脉冲，每半电路产生一个脉冲，同样脉冲周期是开关周期的二倍。对脉冲波的设计和注意事项与尖峰波相同。

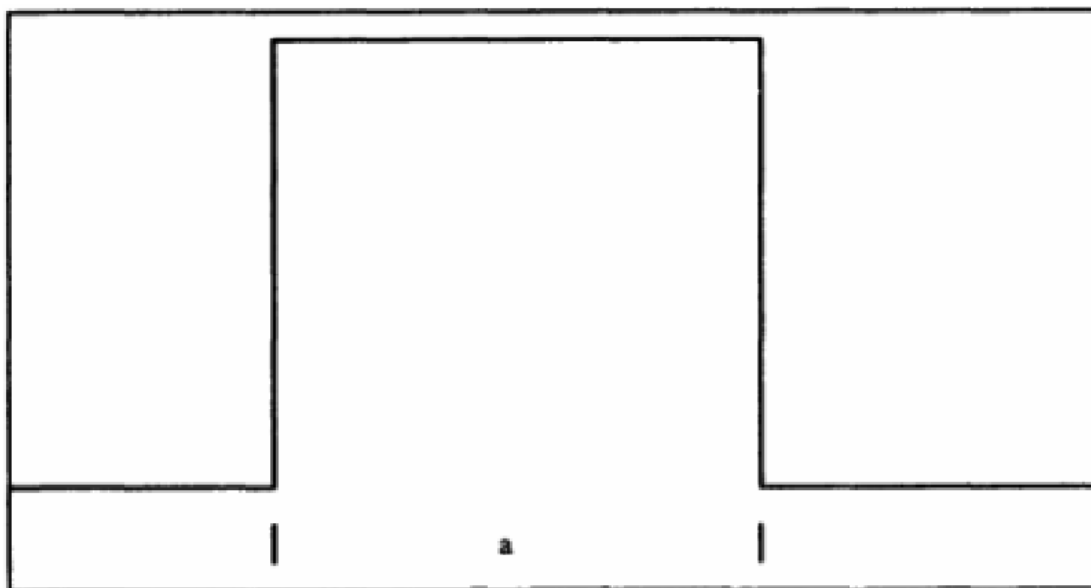


Figure 10.2 The pulse wave.

The equation for the pulse is

脉冲波的傅立叶方程为

$$\frac{Ea}{T} + \frac{2E}{p} \sum_{N=2,4,6,8}^{\infty} \frac{1}{N} \cos pNt \sin \frac{pNa}{T} \quad (10.3)$$

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

此处 a 为脉冲宽度。微分表达式如下：

$$- \frac{4E}{T} \sum_{N=2,4,6,8}^{\infty} \sin pNt \sin \frac{pNa}{T} \quad (10.4)$$

§10.3 The Trapezoid Wave

§10.3 梯形波

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 10.3). 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.

因为脉冲波和准方波是一个阶跃函数，因此上升和下降沿时间为零，所以就这一点来讲梯形波更接近实际情况。梯形波具有像实际波形那样的上升和下降沿时间（图 10.3）。同样，梯形波的 EMI 谱能量也较少。梯形波的缺点是会降低电源效率，因为在上升和下降时间中有太多的能量被消耗。

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.

可以按尖峰波所用的方法选择一个电容器与主储能电容并联。同样： T_0 是脉冲宽度， T_r 是上升和下降沿时间。

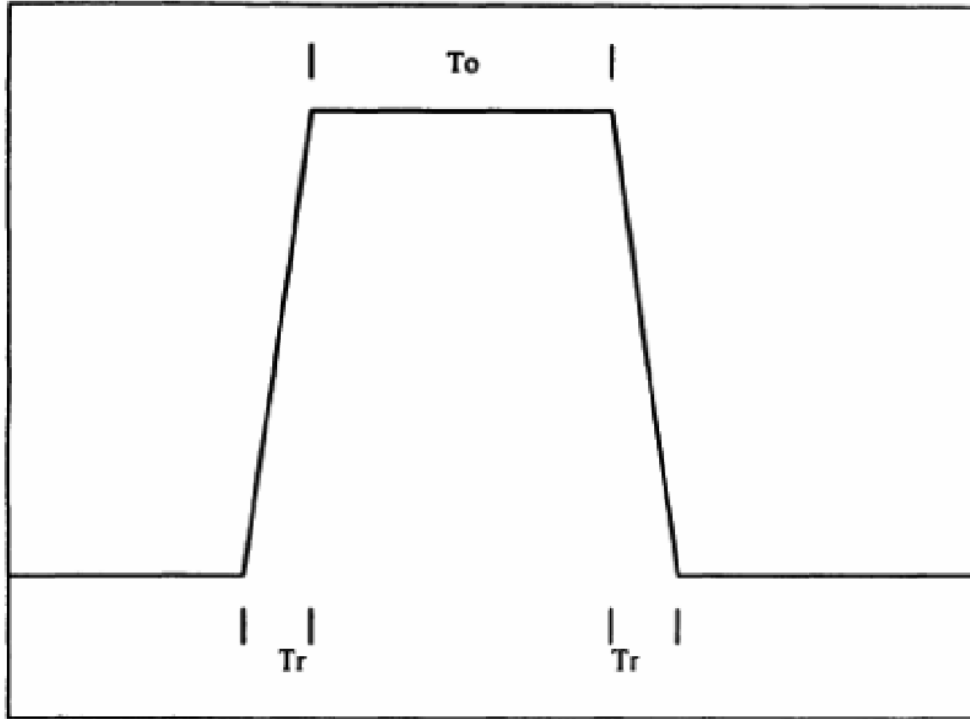


Figure 10.3 The trapezoid wave form.

$$\frac{2E(T_0 T_r)}{T} \sum_{N=1,2,3,4}^{\infty} \frac{\sin Np(T_0 + T_r)/T}{pN(T_0 + T_r)/T} \frac{\sin pNT_r/T}{pNT_r/T} \cos \frac{2pNt}{T} \quad (10.5)$$

The differential is as follows:

其微分形式如下：

$$-4E \sum_{N=1,2,3,4}^{\infty} \sin \frac{Np(T_0 + T_r)}{T} \frac{\sin pNT_r/T}{pNT_r} \sin \frac{2pNt}{T} \quad (10.6)$$

§10.4 The Quasi-square Wave

§10.4 准方波

The quasi-square is the wave applied to the gate, or control, of the switch, and the reciprocal is applied to the opposite control (Figure 10.4). 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. The pulse is filtered by the same technique described in Section 10.1.

准方波是开关的门控制信号，其反向部分是施加在反相控制电路上的（图 10.4）。其脉冲宽度为 a ，驻留角是脉冲之间角度的一半，每个周期有四个驻留角，且每个驻留角在脉冲前和后产生。每半周期内这个方波信号都会开启正/反向工作的两个设备中的一个。并且每次开启的时间会小于半个周期。其输出是电流脉冲，此脉冲为准方波频率的两倍。这种脉冲可以用 10.1 节介绍的技术来滤除。

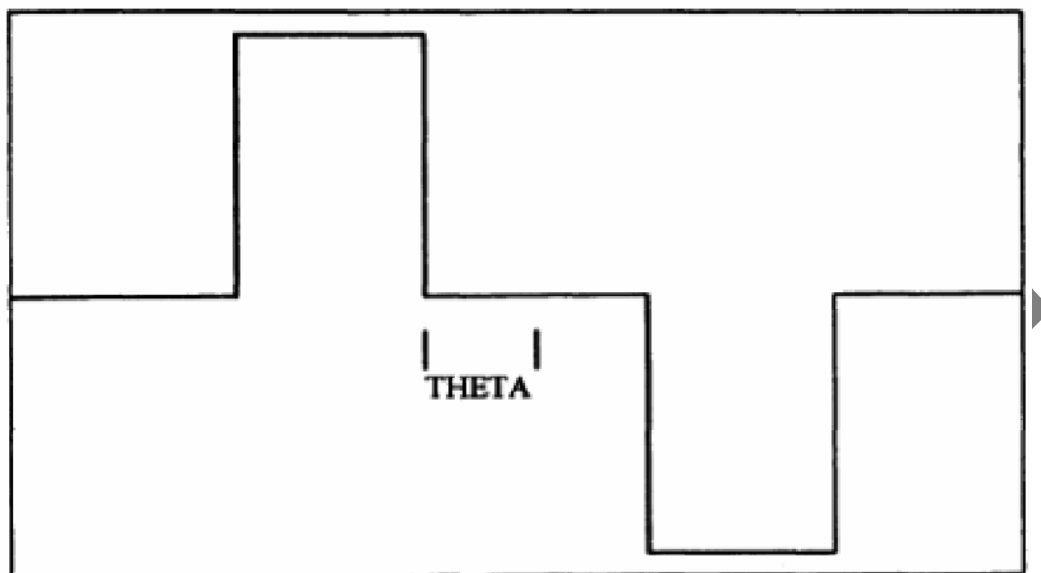


Figure 10.4 The quasi-square.

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 charge, 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 10.5.

这种准方波与其输出脉冲波都不是真实存在的，它们缺少实际波形所具有的上升/下降沿时间，这些上升/下降沿时间是由回路中的小电容和电感产生的。对于所有的设备和电路板，导线寄生电容和开关输入电容的充放电都会使波形具有斜面，并使得方波、准方波变成具有上升/下降沿时间的梯形波。此时准方波更像图 10.5 所表示的波形。

The equation for the quasi-square wave is

准方波的傅立叶变换等式为

$$\frac{4E}{p} \sum_{N=1,3,5,7}^{\infty} \frac{1}{N} \sin Nwt \cos Nf \quad (10.7)$$

The differential is as follows:

其微分形式如下：

$$\frac{8E}{T} \sum_{N=1,3,5,7}^{\infty} \cos Nwt \cos Nf \quad (10.8)$$

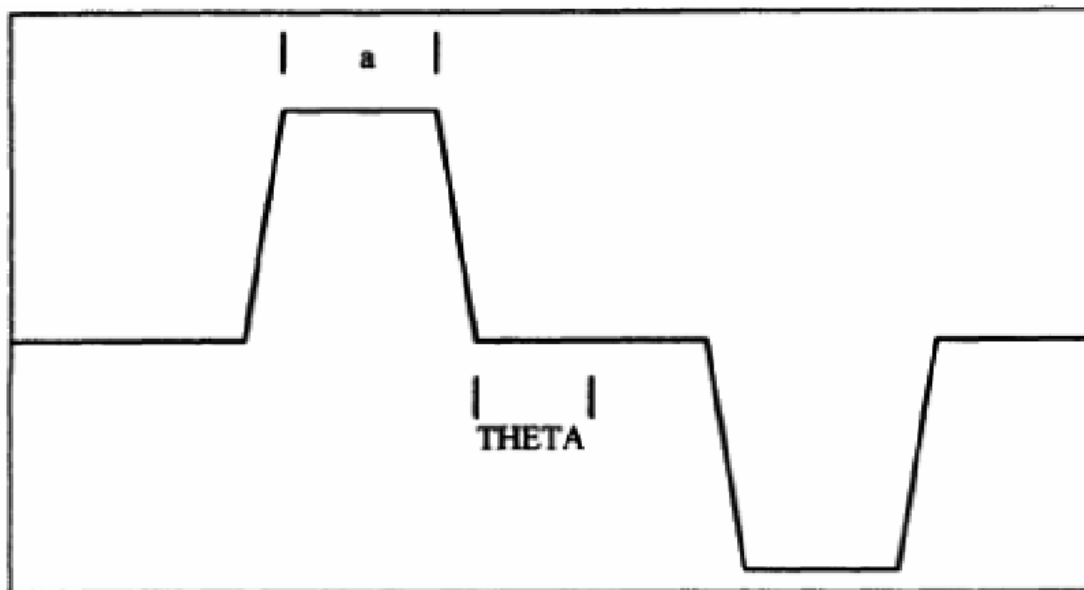


Figure 10.5 The quasi-square with rise and fall times.

§10.5 Why Differentiate?

§10.5 微分的意义

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 wave is only a small change in the dc current state, the radiated H field is small. A steady-state dc current generates only a steady-state H field. This H field cuts the surrounding surfaces only during turn-on and turn-off. These waves are often pulses with a peak amplitude of many amperes. They generate strong varying H fields, leading to radiated energy (Figure 10.6). The formula is as follows:

如果每种波代表电流脉冲，那么可以通过波形计算出其辐射的磁场。如果过每种波在直流状态下变化很小，那么辐射的磁场是很小的。如稳态 DC 电流只产生稳态磁场。这种磁场只在波形处在打开和关闭状态时才会切割周围的平面。这些波常常是具有很大幅度的脉冲，它们会产生强变化磁场，并向外辐射能量（图 10.6）。磁场的计算公式如下：

$$H = \frac{0.4\pi NI_p}{M_{pl}} \quad (10.9)$$

where N is the number of turns, I_p is the peak current, and M_{pl} is the magnetic path length. If this is a wire carrying this I_p current, the turns are then equal to 1 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:

此处 N 是匝数, I_p 是峰值电流, M_{pl} 是磁路长度。如果直导线的峰值电流为 I_p , 那么匝数就是 1 而 M_{pl} 是围绕直导线的圆周。在此例中, 半径为参考平面到导线的距离 (通常是到最近的金属盖或其它导线)。如果导线到金属盖的长度是 R_r , 那么等式变为:

$$H = \frac{0.4\pi N I_p}{2\pi R_r} = \frac{0.2 N I_p}{R_r} = \frac{0.2 I_p}{R_r} \quad (10.10)$$

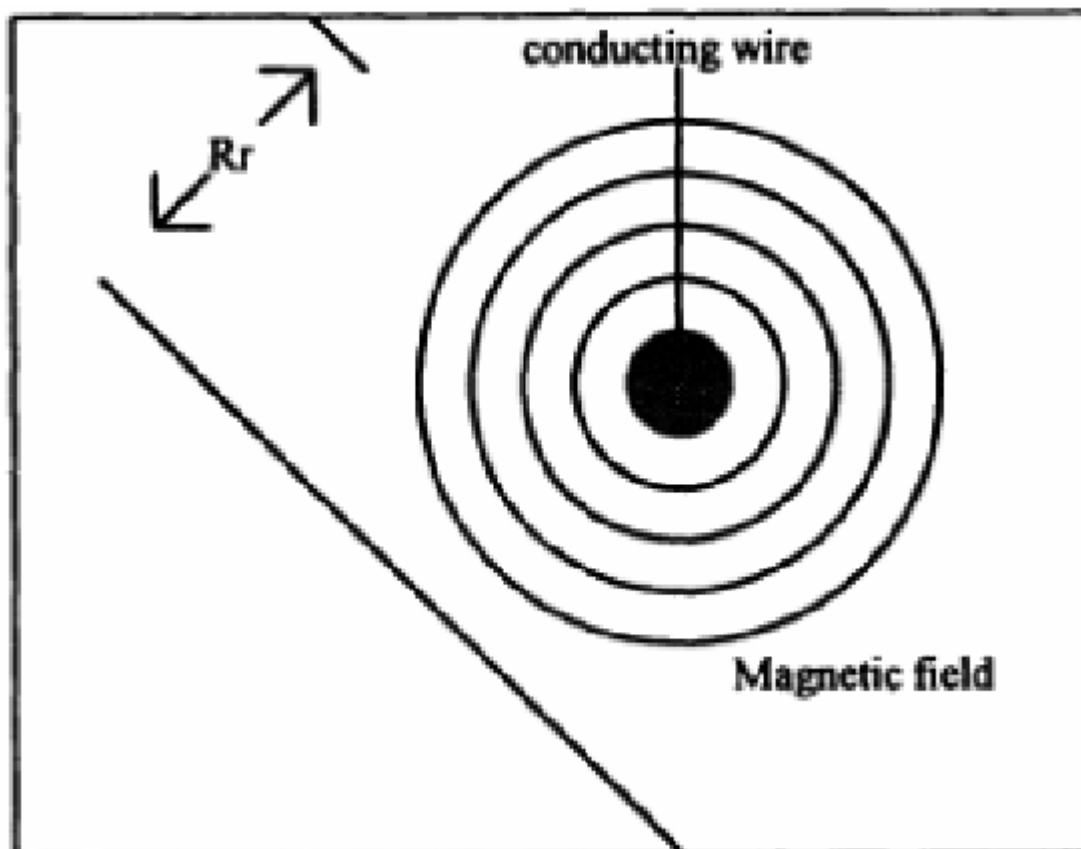


Figure 10.6 The radiated field.

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 .

我们所需要的是 H 对 t 的微分, 且在右侧的等式中依赖于 t 的变量只有 I_p 。

$$\frac{dH}{dt} = \frac{0.2}{R_r} \frac{dI_p}{dt} \quad (10.11)$$

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 is in R_r meters.

所以, 在距离 R_r 米时, 此问题的近似的磁场密度为 0.2 除以 R_r 后乘以波的微分式 (此时 $N=1$)。

§10.6 The Power Spectrum

§10.6 功率谱

How much power is in the various waves discussed in the earlier sections? These form envelopes that provide the peak power, which varies with the amplitude of the current I . If the current pulse width is a and T is the period, the equation is as follows (Figure 10.7):

上节讨论的各类波的功率是多少呢？这常常用峰值功率来表示，它是随电流 I 的幅度而变化的。如果电流脉冲宽度为 a 且周期为 T ，那么脉冲功率表达式如下（图 10.7）：

$$\text{Pulse power} = 20 \log \frac{2Ea \times 10^6}{T} \text{ dB / MHz} \quad (10.12)$$

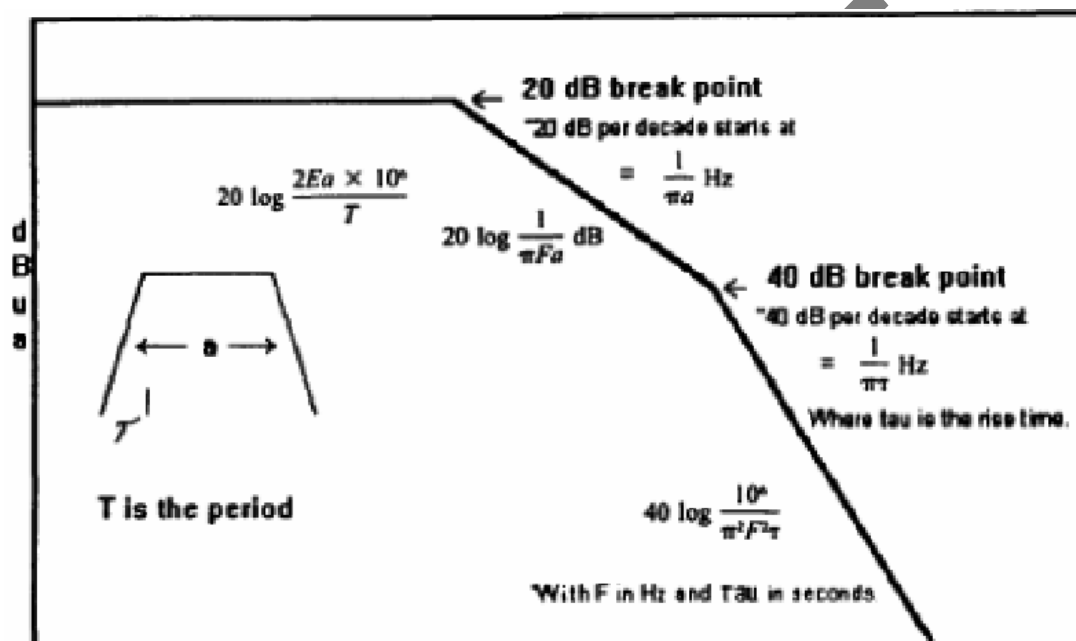


Figure 10.7 The dBuA per megahertz.

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

此处脉冲宽度为 a ， E 为幅度。在 DC 到转折点之间是一条平直的曲线，之后是 20dB/十倍频或 6dB/倍频的斜线。此转折频率的始点值为：

$$\frac{1}{\pi a} \text{ Hz} \quad (10.13)$$

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

此处 a 是以秒为单位的脉冲宽度，在此点后幅度的 dB 值为

$$20 \log \frac{1}{\pi F a} \text{ dB} \quad (10.14)$$

There is also a 40dB per decade breakpoint, 12dB 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.

接下来是另一个转折点，此转折点的频率值是由上升时间决定的，上升沿时间越长，转折点出现的就越早。之后就是 40dB/十倍频或 12dB/倍频的斜线。这是一种较好的折衷，可以减小滤波器的尺寸、重量和成本。基本上，在 40dB/十倍频转折点以上的频率，如果上升沿时间是原来的 10%，那么就需要在已有的滤波器上增加一个 L 型滤波器。

$$\frac{1}{p\tau} \text{ Hz} \quad (10.15)$$

The added loss after the 40 dB per decade breakpoint is
在 40dB/十倍频转折点后的频率下，损耗为：

$$40 \log \frac{10^6}{p^2 F^2 \tau} \quad (10.13)$$

where F is in hertz and τ is in seconds.

此处 F 的单位是 Hz， τ 的单位是秒。

11

Study of the Off-Line Regulator

离线调节器的研究

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

通过利用面积乘积参数 A_p , Mclyman 计算出了体积和重量的近似关系。滤波器的体积和重量与磁芯类型和绕线方法是紧密相关的。此处研究这个问题的目的是，如果在线性电源中使用具有临界值的电感，那么滤波器总的体积和重量会减少。这个电感位于二极管和储能电容之间，它主要用在低频插损要求较高的场合，如军标 461 规范要求的场合。一年前，工程师将这种电源称作感性输入电源。另外，这个电感也常用在功率校正电路上，其作用是消除电流尖峰。通常这个电流尖峰会在每个周期内为储能电容充电两次。当然不论使用何种技术，因为临界值电感的原因 EMI 滤波器的尺寸和重量都会减小，尤其是军标 461 等类似规范要求的应用场合。

The following section uses the same equations to show how the added harmonic content on the power lines increases. This happens because most electronic equipment making up the load today does not employ power factor correction circuits. Another reason for the higher harmonic content is the lack of the inductive input technology discussed in Section 11.1. This improves as the loads meet the newer standards requiring power factor correction circuits, but the equipment making up the load will take years to replace.

后面的章节同样使用这些表达式来显示电源线上的谐波量是如何增加的。目前谐波问题是很普遍的，这是因为多数电子设备都没有使用功率校正电路来补偿负载。另一个原因是，电子设备缺乏 11.1 节讨论的感性输入技术。当负载满足新标准（要求带有功率校正电路）时，这种情况会得到改善，但是这需要很多年才能完全把电子设备都替换成带有负载补偿功能的设备。

§11.1 With or Without Critical Value of Inductance: Size and Weight Differences of Filters

§11.1 有或无临界值电感时：滤波器的尺寸和重量

In referring to this section, many have said that they know that the critical inductor will make the EMI filter smaller, yet rarely is this application seen in practice. In other words, the weight and volume of the EMI filter without the inductor are much greater than those of the filter and inductor together using the critical inductor (Figures 11.1 and 11.2). This is especially true if the EMI filter must have some attenuation at very low frequencies, such as that required by Military Standard 461 (Figure 11.3). This may not be a good trade-off if the only goal is to meet U.S. Federal Communications Commission and similar specifications. If the decibel losses of the filter must be tens of decibel at 10 kHz, this should be looked into.

按前面提到的，这个临界值电感可以令 EMI 滤波器体积更小，但是至今在实际应用中却很少见到使用。无邻近值电感的EMI滤波器在体积和重量上远大于有此电感的EMI滤波器(图 11.1 和图 11.2)，如果在很低的频率下对EMI滤波器的损耗有要求，如军标 461(图 11.3)，这种情况尤为严重。如果 EMI 滤波器的目标仅仅是为满足美国联邦通信委员会（FCC）或类似标准的要求，那么这种方法并不是很好。但是如果在 10kHz 要求滤波器的损耗为几十 dB 的话，那么就需要考虑使用临界值电感了。

The formula for the critical inductance, L_c , is as follows:

临界值电感 L_c 的表达式如下：

$$L_c = \frac{R_0}{6pF} \quad (11.1)$$

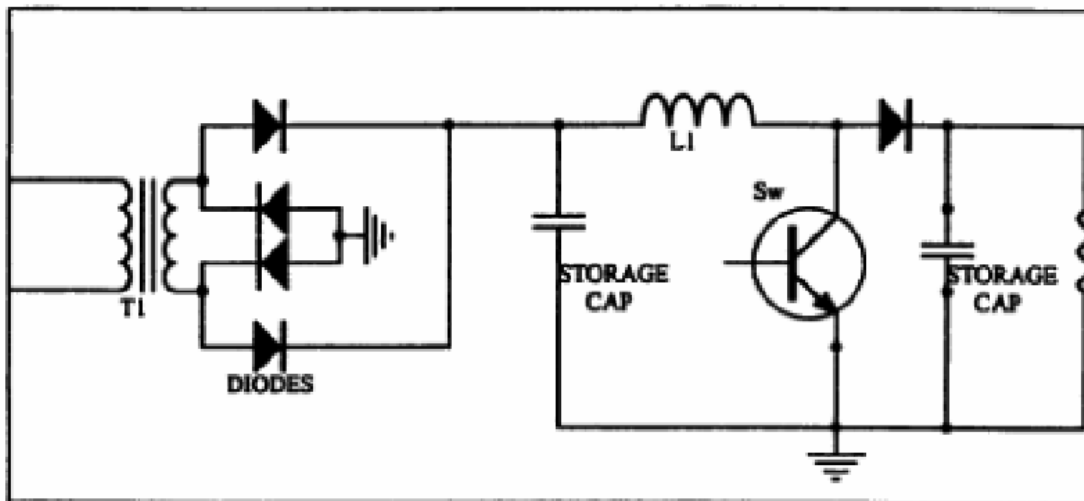


Figure 11.1 The off-line regulator with capacitive load and without critical inductance.