

# CHAPTER 2

## PASSIVE COMPONENTS

The selection of passive component elements in a circuit is often overlooked as these components are usually chosen to bias or complement their more exciting active component counterparts. The passive component does have a significant effect on the overall electromagnetic compatibility (EMC) of a circuit, as these components cannot only be the cause of circuit problems in themselves, but can also upset the stability of the active circuit they are connected to.

Passive component electromagnetic interference (EMI) problems could result in the expensive mistake of replacing a perfectly good active device, such as an op-amp, because a bias resistor is acting inductively at a certain frequency for example. The change could not only be more expensive, the addition of extra filtering to reduce some noise source, but may simply be masking the real problem and not truly solving it.

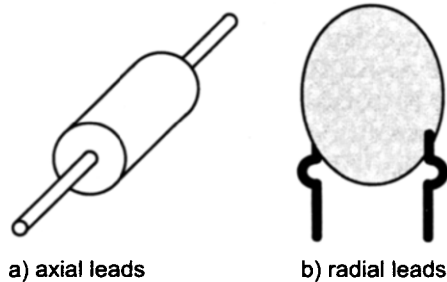
Care therefore needs exercising in the choice of passive component for certain circuit applications. There will be instances where absolute value is not as important as construction or component material for the EMC performance of the circuit. Few manufacturers like to admit that their component is not suitable for any particular application, it is therefore left up to the designer or production engineer to decide if component choice, say using a carbon instead of metal film resistor, is going to change the EMC performance of the circuit. Without the right background knowledge this type of decision cannot be made correctly.

### 2.1 Passive Component Packaging

There are essentially only two types of package for all electronic components, these are leaded or leadless. The two package types use different technology for final assembly of the circuit.

Leaded components are mounted on the opposite side to the tracking and the leads pass through the printed circuit board (PCB) to make electrical contact with the circuit. Consequently, leaded components are sometimes referred to as through hole components (due to the common practice of plating through holes on a PCB this type of component is also referred to by the abbreviation PTH component). Leaded technology is the elder method of component attachment, virtually all component

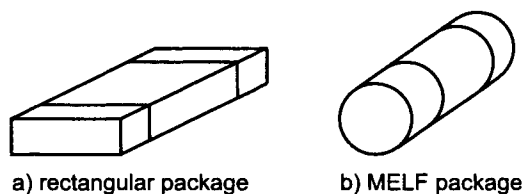
types are available in a leaded package. Types of leaded package are relatively numerous, however passive components tend to stick to two form factors; axial or radial. Axial leaded components feature a cylindrical component structure with leads concentric with the component body at either end. Radial leaded components have leads which extend from the base of the component structure.



**Figure 2.1**

Through hole components

Leadless components are mounted on the same side as the tracking connecting the component to the circuit, hence these are often referred to, probably more correctly, as surface mount (SM) components. Certainly the term leadless is a slight misnomer as there are of course some form of lead or termination between the component structure and the circuit contacts. Although SM has been around for some time, it is not as universally used as through hole technology on low volume circuits. Primarily due to the cost of the component handling equipment, SM is more commonly found on high volume products and more recent circuit designs requiring a high packing density. Not every type of passive components is available in a SM form, hence there is not always as much choice as with leaded types. Leadless passive components also have a limited variety of package styles, the most dominant is the rectangular body with rectangular end terminations. There are circular bodied types (MELF), but these tend not to be as popular due to handling and placement issues.



**Figure 2.2**

Surface mount packages

The package parasitic of leaded components are dominated by the lead length. At high frequencies the lead forms a low value inductor, a typical value for lead inductance is 1 nH/mm per lead (i.e. a component with 10 mm leads will have a parasitic 20 nH inductance in series with it, 10 nH per lead). The end terminations can also produce a small capacitive effect, in the region of 4 pF (based on axial shape with metal end caps on a 10 mm body), but it is usually the lead inductance that is the dominant parasitic component. Consequently, the lead length should be reduced as much as possible. Drilled through holes in the PCB should be spaced just longer than the body length for axial components and the component should be mounted close to the PCB surface. Radial leaded components naturally allow for shorter leads as the body can be maintained flush with the PCB surface and no bend is required for the body diameter (as with axial parts). A radial leaded part can have a lead length almost equal to the PCB thickness. Some lead forming may be required for the manufacturing process (i.e. to stop the parts falling off the PCB during wave soldering), this forming should be made with the shortest possible lead lengths to hold the component body flush to the PCB surface.



**Figure 2.3**

Loop area of through hole components

SM technology has the shortest lead lengths possible by design, hence there is little the user can do to reduce further the parasitic effects of this package type. There is still a parasitic inductive effect at high frequencies and the main benefit of SM over leaded components is that this is much better controlled and stable, the variations in lead length that can be observed in similar type of leaded parts are not manifest in an SM package. Typically 0.5 nH of parasitic inductance is present in most SM packs with a small end termination capacitance of about 0.3 pF. As with leaded parts it is the parasitic inductance that dominates at high frequency. The different SM sizes (usually quoted by reference numbers, e.g. 1206, 0805) tend to produce reasonably similar parasitic values of inductance and capacitance. Certainly the variations between types are small because as body sizes increase, end terminations also get larger and the net effect is a similar package parasitic.

The different mounting options therefore produce different additional parasitic component elements to be considered. The preferred choices from an EMC viewpoint would be first SM, then radial leaded and lastly axial leaded. At the end of the day it is most likely that component package choice will be down to assembly technology and part type availability, but being aware of some of the EMC issues at the package level can help minimise potential EMC problems.

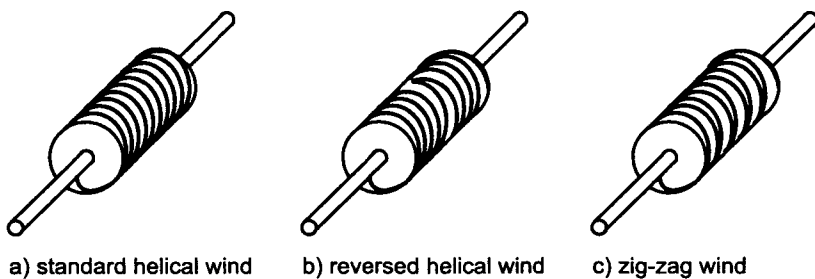
## 2.2 Resistors

The simplest element in almost all circuits, surely the good old resistor cannot affect the EMC performance of a circuit? Unfortunately, the wrong choice of resistor type can actually affect several aspects of circuit performance, not only EMC but higher frequency functional performance. The choices are mainly connected with the physical construction of the resistor and their parasitic effect.

### 2.2.1 Resistor Construction

Leaded resistors are almost exclusively formed in an axial package style. There is generally a choice of three common materials: carbon, metal film or wire wound. Other materials do exist but these are the most common forms, and much of the discussion can be applied to other materials knowing the material properties and the resistor construction.

The construction of film type (metal and carbon film) and wire wound types of leaded resistor are identical, essentially a helical track of the resistive element is placed on to a thermally conductive body (ceramic being the preferred body material). The resistance is governed by the thickness and length of the track. The shape can lead to the potential for a large inductive element, which can be virtually eliminated with tracked film resistors, as the tracks can be made to zigzag rather than perform complete circles. With wire wound resistors this can be performed to some extent by reversing the direction of the helix midway through the winding, but this is not easy to produce and not a common technique.



**Figure 2.4**

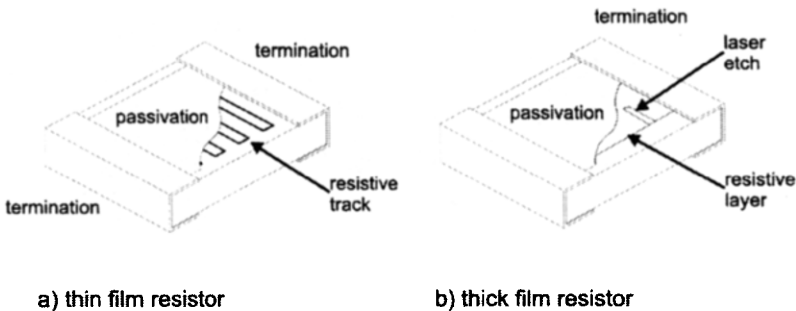
Film and wire wound resistor construction

The other type of carbon resistor is produced from a solid body of a carbon compound, therefore no tracking is present. This resistor's resistance is affected by the mix of the carbon compound forming the body. Carbon-bodied resistors tend to be lower power rated than metal film or wire wound types as the heat generated by the resistive element in a film or wire type is easily conducted away from the

element. With solid-bodied resistors the heat can affect the resistance if allowed to rise to a high enough value.

There is an obvious hierarchy or preference in leaded resistors with regards to EMC performance, with preference for the carbon body types. Wire wound resistors are highly inductive and should be avoided in any frequency sensitive circuit. The shape of a spiral wound wire can be considered as a solenoid, hence the component will not only behave inductively in circuit, but is quite a susceptible design and could introduce signals into the circuit. Metal film tend to have lower inductance but this still dominates as a parasitic element at relatively low frequencies (MHz region). Carbon film is not a particularly good conductor, hence its use as a resistive element, this also reduces its ability to form an inductor. Carbon bodied-leaded resistors are dominated by the parasitic end terminations and leads rather than the resistor construction.

SM leadless resistors come in two basic construction types: thick film and thin film. Thick film types have a resistive layer on a carrier (again usually ceramic) and the film is 'trimmed' along one edge to produce the desired resistance value. Thin film resistors have a 'snake' of resistive film on a carrier forming the resistor. Neither type has a significant inductive parasitic and the end terminations tend to be the dominating parasitic component at high frequencies. Typically in small sizes (0603, 0805) this end termination gives 0.3 pF of capacitance, in larger body sizes (2010, 2512) this can become as low as 0.05 pF. There is some termination inductance, in the region of 0.5 nH, but this is insignificant except at extremely high frequencies. There is no hierarchical preference for either construction of SM resistor, both are preferable to leaded types for their EMC performance.



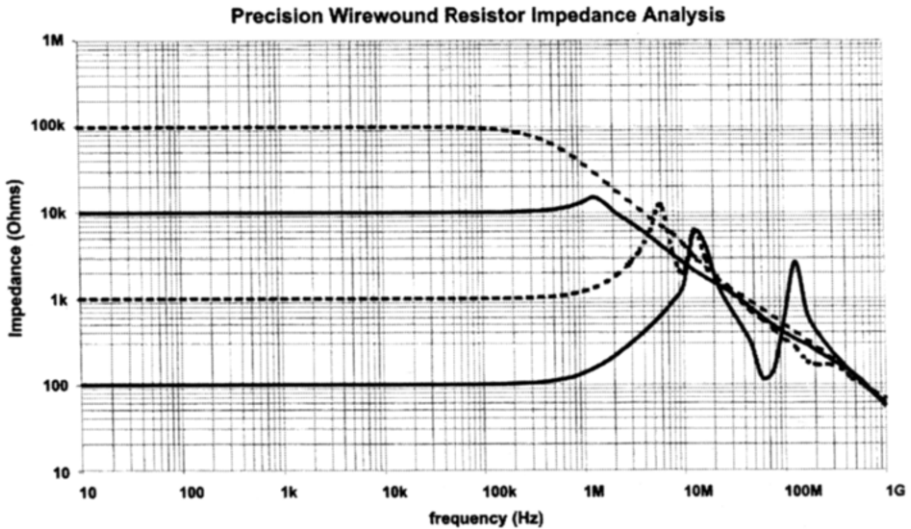
**Figure 2.5**

Surface mount resistors

### 2.2.2 Resistance Value

The absolute resistance value can have an effect on the EMC performance. In wire wound types for instance the parasitic inductance will dominate at a lower frequency the more wire that is on the part (i.e. at higher resistance values). Precision wire

wound resistors can be constructed by using an equal number of turns and changing the wire gauge to produce the required resistance value, hence equal parasitic inductance and capacitance is present regardless of resistance value.



**Figure 2.6**

Wire wound resistor impedance

With leaded carbon and metal film types the lower the resistance the more prominent the lead inductance, hence at high resistance values lead inductance can be swamped and the end cap terminations become the dominating parasitic element, this only affects the component performance at very high frequencies (>100 MHz).

In SM packages the end termination parasitics dominate regardless of construction. The parasitic capacitance is in the sub-pF range so only becomes noticeable with higher resistance values and at very high frequencies.

Resistors also have a noise voltage associated with their construction. In wire wound types this is extremely small, metal film types have a noise voltage of about  $0.1 \mu\text{V/V}$  (i.e.  $0.1 \mu\text{V}$  of noise is generated for every 1 V of applied bias). Carbon types have a noise voltage dependent on the resistance, given by the equation:

$$V_n = 2 + \log\left[\frac{R}{1000}\right] (\mu\text{V/V}) \quad 2.1$$

The value of noise voltage is small even for large resistor values, for example a 100 k resistor only generates  $4 \mu\text{V/V}$ . In the majority of EMC critical applications it will be the effect of the reactive parasitic elements which will limit performance not the noise voltage. Noise voltages of this level are of more concern for circuits which

require a low noise threshold for their intrinsic function (e.g. high precision converters, precision voltage references, audio amplifiers). The frequency of resistor noise voltage also tends to fall as a  $1/f$  law, hence this noise voltage has little effect on the EMC performance of a resistor.

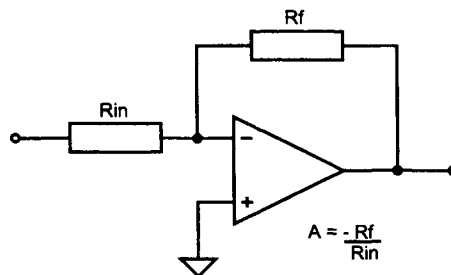
### 2.2.3 EMC Critical Resistor Applications

The most important consideration for the majority of circuit designers is what is the best type for my application? Although the application of resistors is so varied it is next to impossible to answer all applications within any single book, it is possible to generalise on a few circuits by their normal operating frequency.

If possible, SM types are always preferable, these offer the lowest parasitic elements and are relatively stable over temperature and life due to their film on ceramic construction. With through hole parts the carbon body or carbon film types are the best option, with metal film very good where a higher power density or high accuracy is required. Wire wound resistors should only be used where the high power handling of this type is essential. Some of the very high power wire wound types are available in a metal heat sink body cladding, this can be connected to a ground or even one end of the resistor to reduce the inductive effect of the construction and provide some shielding from any internally generated fields.

In gain setting on amplifier circuits the effect of increased impedance due to inductive reactance can cause large differences in the circuit gain at higher frequencies. In high frequency amplifiers the resistor gain setting network needs to be located very close to the amplifier, especially at the input terminal (whether a discrete or integrated amplifier). In audio and lower frequency circuits placement tends not to be quite as critical, but minimising distances between connected circuit components is generally a good practice.

Biasing resistors in general do not have to be of a high tolerance, hence low cost carbon-bodied types are perfectly adequate. If biasing the outputs of a fast switching



**Figure 2.7**

Locate gain setting resistors close to amplifier

transistor or integrated circuit (IC) then placement needs to be close to the active component terminals and power (pull-up) or ground (pull-down) connection at the local supply point of the active device. Parasitic inductance on fast-edged switches can create some ringing, and hence requires minimising.

On DC bias points resistor placement and the parasitic lead effects tend not to be as critical. The exception for DC bias circuits are regulator and reference devices which are minimally decoupled to improve the transient response times, therefore will require close placement of the resistors to the active device.

In RC filter networks the wire wound resistor may give a much higher roll-off curve than other types due to its highly inductive nature. Some knowledge of the inductive effect will need to be known if a controlled roll-off is required, otherwise roll-off may occur earlier than expected due to the parasitic inductance. This is one situation in which the inductive parasitic of a wire wound resistor can be used to benefit the EMC performance of the circuit. Care needs to be taken for the possibility of local oscillations if a resonance mode is produced between the parasitic inductance and any capacitor in the circuit, effectively forming a tank circuit, or even within the resistor construction itself (see 100Ω wire wound resistor impedance in Figure 2.6).

Overall, with the exception perhaps of wire wound types, resistors are relatively stable elements. Care should be taken if designing very high frequency circuits that gain roll off is not occurring due to parasitic inductive or capacitive impedance effects rather than the resistance itself.

It is not possible here to cover every resistor type or application, next time your supplier calls ask for further information on the parasitic elements present in the types in use in your circuit. The suppliers are the best source of information and may even be able to change their component construction to suit your application if you require enough resistors of a specified type.

## 2.3 Capacitors

The capacitor is potentially the easiest and cheapest way to solve many EMC problems. However, the correct choice of capacitor needs careful consideration as not all capacitors behave the same over a wide frequency range. Capacitor selection can be a headache due to the number of types available, their individual behaviour and the possibility of parallel resonance when different values or types are used in the same circuit. Choice should be made on the basis of application and with a good idea of the frequency range over which the capacitor is intended to work.

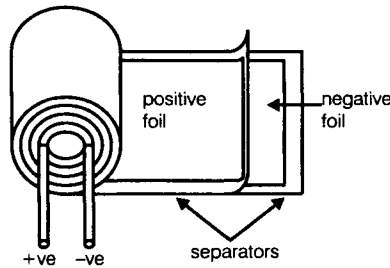
### 2.3.1 Construction

The construction of a capacitor is to a certain extent similar for all types, essentially there are two plates of metal with a sandwich of dielectric between them. It is the dielectric material used for charge storage which determines the main capacitor



characteristics. There is quite a diversity of constructional methods and so for simplicity only a few of the most common types will be described here. One comforting factor throughout the whole range of dielectric and constructional types is that the same parallel plate capacitor equation is used with all types to calculate the designed capacitor value:

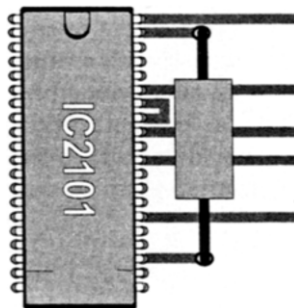
$$C = \epsilon_0 \epsilon_r \frac{A}{d} \tag{2.2}$$



**Figure 2.8**  
Aluminium electrolytic capacitor construction

Where  $\epsilon_0$  is the dielectric constant of free space (8.854 pF/m),  $\epsilon_r$  is the relative dielectric constant of the dielectric material,  $A$  is the overlapping plate area and  $d$  is the distance between plates.

Leaded capacitors are available in both axial and radial forms, with radial being the preferred form for most dielectric types. The advantage of radial leads are the ease of producing very short lead lengths on to the PCB, leads can be cropped almost flush with the PCB. Axial-leaded parts should be mounted with the body flush to the



**Figure 2.9**  
Axial decoupling capacitor bridging PCB tracks

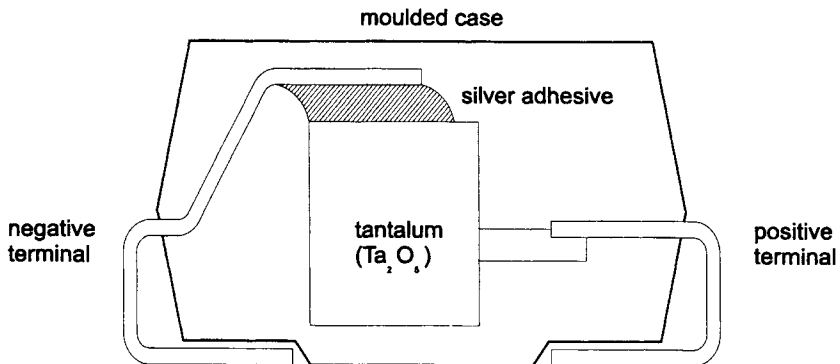
PCB and with minimum excess lead lengths. There are situations where an axial leaded part is advantageous, if for example there are tracks which need bridging to decouple an IC, the capacitor body can be used to form the bridge. At high frequencies the inductance of the leads reduces the effectiveness of the capacitor and in the majority of cases radial-leaded components would be the preferred choice of leaded part.

Aluminium electrolytic capacitors are usually constructed by winding metal foils spirally between a thin layer of the dielectric. This constructional method gives high capacitance per unit volume but increases the internal inductance of the part. Tantalum types are manufactured from a block of the dielectric with direct plate and pin connections, and hence have a lower internal inductance than aluminium electrolytic types.

Ceramic capacitors have a construction consisting of multiple paralleled metal plates within a ceramic dielectric (SM types are often referred to as multilayer ceramic chip capacitors, MLCC). The dominant parasitic is the inductance of the plate structure and this usually dominates the impedance for most types in the lower MHz region. The type of ceramic dielectric appears to have little effect on the parasitic inductance (standard ceramic materials have very close natural resonant frequencies: COG; 40 MHz, X7R; 44 MHz, Y5V; 46 MHz). This makes the estimation of useful frequency range relatively easy as a parasitic package inductance of 0.4 nH can be used and the self-resonant frequency (the frequency at which the capacitor has minimum impedance) can be estimated from the nominal capacitance value:

$$f = \frac{1}{2\pi \sqrt{L_p C}} \quad 2.3$$

The above equation can be rearranged to calculate the maximum value of capacitance above which parasitic lead inductance limits frequency response. For ceramic capacitors this value is approximately 40 nF, that is below 40 nF ceramic

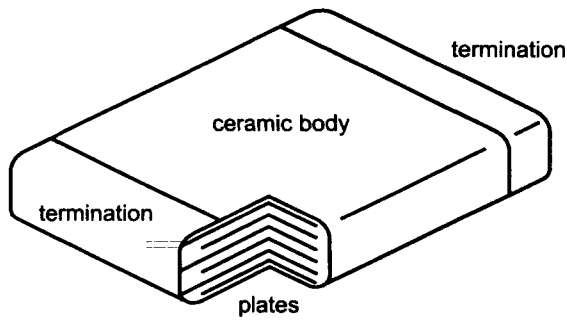


**Figure 2.10**

Surface mount tantalum capacitor construction

capacitors of Y5V, X7R and COG dielectric are limited by the natural frequency response of the material not by package parasitics.

As with resistors, package type is important and for the same reason, the internal structure and end terminations behave inductively at high frequency. Preference would again be for SM types; however, the choice and range of capacitors available in SM is limited compared with through hole types. More types are becoming available in SM packages and as well as the popular ceramic types, aluminium and tantalum dielectric types are now available in a SM package from many suppliers.



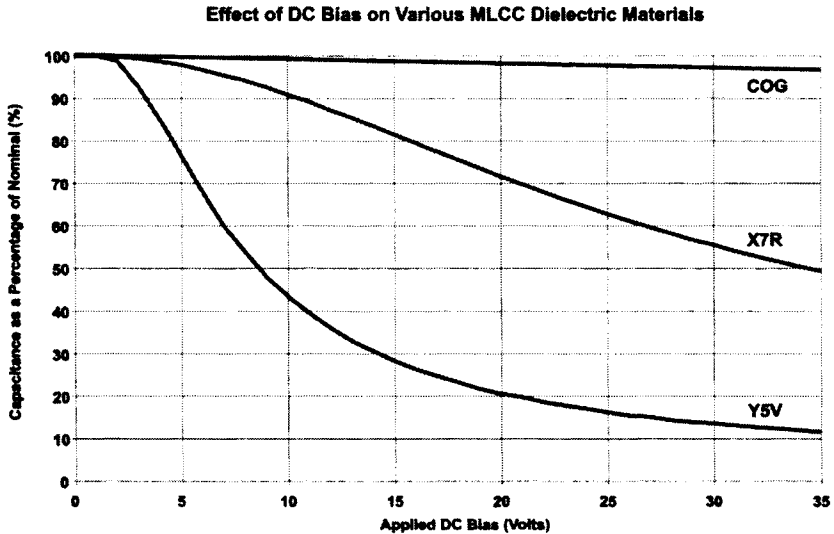
**Figure 2.11**

Ceramic capacitor

### 2.3.2 Capacitor Dielectric

The dielectric material used in the construction of a capacitor to some extent defines its application area, each dielectric has a limited frequency range over which it is usable. Aluminium and tantalum electrolytic types dominate at the low frequency end, mainly in reservoir and low frequency filtering applications. In the mid frequency range (kHz to MHz) the ceramic capacitor (multilayer structured types) is the dominant dielectric for decoupling and higher frequency filters. Special low loss ceramic and mica capacitors are available, at a cost, for very high frequency applications and microwave circuits.

The use of ceramic capacitors needs careful consideration, as some of the lower cost ceramic dielectrics (Y5V or Z5U and X7R) have a voltage bias dependence and temperature variation. The effect is a significant fall of capacitance with DC bias voltage, hence the effective capacity available in a circuit may be much less than is required or has been designed for. This may be a problem in decoupling applications, particularly at high bias voltages (12 V and higher), so care should be taken in selection even of ceramic types. Even at 5 V the lowest cost ceramic dielectric (Y5V and Z5U) exhibits a loss of 20% of its nominal capacitance value, this is ignoring any additional tolerance variation and thermal effects.



**Figure 2.12**

Ceramic dielectric bias dependence

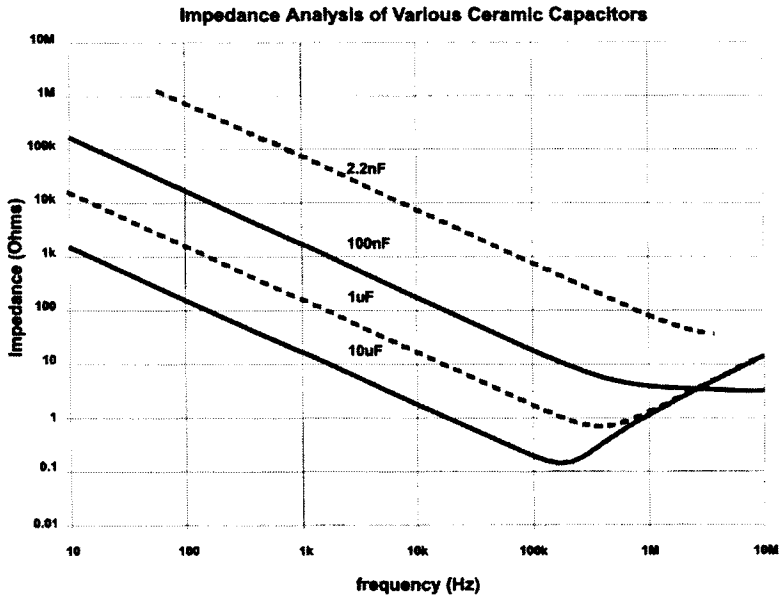
There is of course a cost and sometimes a size penalty with better ceramic dielectric. The COG and NPO types are very stable but do not have the storage density of the lower cost materials, and, hence, often a larger body size may be required for the same value of capacitor. For example, using Y5V dielectric may enable a 1  $\mu\text{F}$  capacitor to be produced in an 0805 size body, in X7R the same capacitance value may only be available in a 1206 size and in 1210 for COG. At the values used for decoupling in digital circuit applications (in the nanofarad range) this should not be a great problem as all values can be obtained in small (0805 or less) body sizes. Also the absolute value of capacitance is not quite as critical for high frequency decoupling applications, even a 50% loss of nominal capacitance should still maintain most of the decoupling effectiveness.

### 2.3.3 Capacitor Impedance Plot

One of the best ways to assess if a capacitor is suitable for an application, with respect to frequency, is its impedance plot. This is a graph of the impedance between the terminals as a function of the applied frequency. The ideal capacitor equation would suggest that the impedance ( $Z_c$ ) would continue falling indefinitely and is dependent solely on the capacitance ( $C$ ) for any applied frequency ( $f$ ):

$$Z_c = \frac{-1}{j2\pi fC} \quad 2.4$$

The actual plot shows the effect of parasitic inductance ( $L_p$ ) in that the impedance eventually starts rising again. Hence the capacitor exhibits a self-resonant frequency.



**Figure 2.13**

Impedance plot of various ceramic capacitors

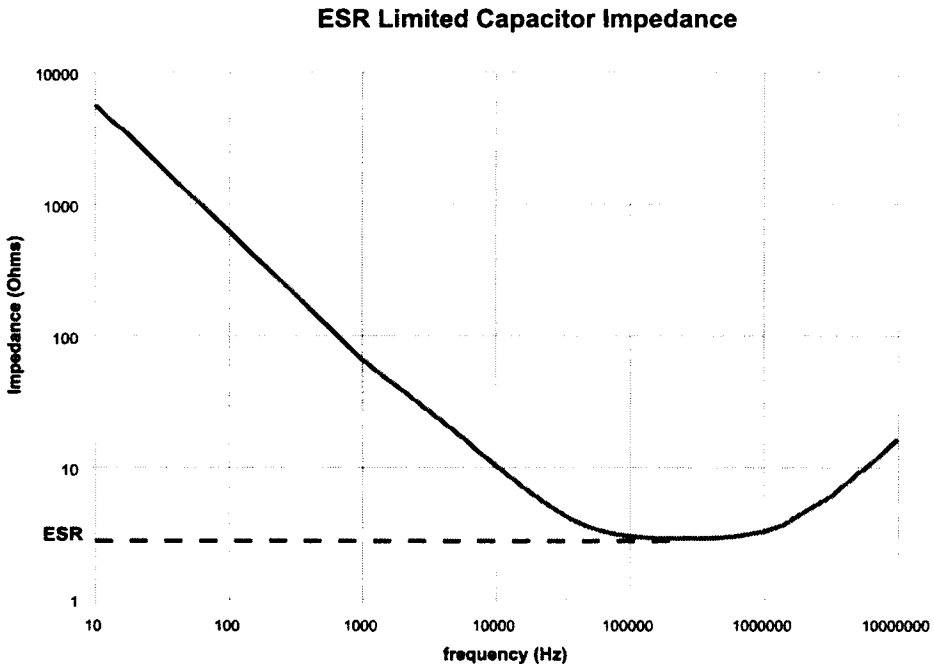
This is the frequency at which the parasitic inductance and design capacitance have an equal reactance. Self-resonance frequency ( $f_o$ ) can be calculated, if not quoted in the data sheet, from:

$$f_o = \frac{1}{2\pi \sqrt{L_p C}} \quad 2.5$$

On some capacitors the lowest value of the impedance plot is not limited by this self-resonance effect, but by the equivalent series resistance (ESR). The ESR is a figure for the effective short circuit ability of the dielectric. The ESR value for ceramic dielectrics is typically under  $1\Omega$ , often close to  $0.1\Omega$  or lower. Tantalum materials exhibit ESR values in the  $1\text{--}5\Omega$  range, with aluminium dielectrics having higher ESR values up to  $10\Omega$ . The equivalent series resistance value limits the lowest value at which the impedance can fall to, and often disguises the actual self-resonance frequency. For best EMC performance it is important to have a low ESR value as this provides a higher attenuation to signals, especially frequencies close to the self-resonant frequency of the capacitor in use.

The equation for the impedance ( $Z$ ) can be rewritten to include the parasitic inductive impedance ( $Z_L$ ) and ESR value if a more detailed analysis is required:

$$Z = Z_L + Z_c + ESR = j2\pi f L_p - \frac{1}{j2\pi f C} + ESR \quad 2.6$$



**Figure 2.14**

ESR limited capacitor

It is generally the self-resonant frequency that limits the useful frequency range of a capacitor, hence the need to limit parasitic inductance. It can also be observed that a higher operating frequency range is obtained from lower value capacitors in the same dielectric and package type until limited by the dielectric material frequency limit.

### 2.3.4 Bypass Capacitor

One primary use of capacitors in many circuit designs is to act as a high frequency bypass source for switching demands. The bypass capacitors also tend to be used as supply voltage hold-up capacitors and act as a ripple filter to reduce the transient circuit demand on the power supply unit (PSU) directly. The capacitor reduces transient current demands being routed around other circuits attached to the PSU by providing the short demands from the local reservoir bypass capacitor. A slower current demand on the PSU is made by the capacitor for recharging. The current demand is considered as 'bypassing' the PSU and being supplied by the capacitor.

Bypass capacitors tend to be in the range of 10 to 470  $\mu\text{F}$  per PCB or circuit function. The value required can be calculated if the transient current demand ( $\Delta I$ ), the allowable voltage droop ( $\Delta V$ ) and the PSU to PCB lead inductance ( $L_{PSU}$ ) are known.

The maximum allowable supply impedance ( $Z_{PSU}$ ) is given by:

$$Z_{PSU} = \frac{\Delta V}{\Delta I} \quad 2.7$$

The maximum switching frequency the PSU can tolerate ( $f_{PSU}$ ) without a bypass capacitor is estimated by the ratio of lead impedance to supply impedance:

$$f_{PSU} = \frac{Z_{PSU}}{2\pi L_{PSU}} \quad 2.8$$

This value usually calculates to be in the 10 kHz to 1 MHz region; therefore, unless the target circuit operates at low speeds (e.g. audio circuits) or on DC only, a bypass capacitor will be required. The bypass capacitor value can then be calculated from the switching frequency available from the supply to match the impedance of the target circuit demand:

$$C_{bypass} = \frac{1}{2\pi f_{PSU} Z_{PSU}} \quad 2.9$$

For example, consider a PCB containing 50 HC CMOS digital ICs (10 ns rise/fall time), each with four gate drivers into loads of 5 pF per gate operating with a 3.3 V supply rail. The current demand for a simultaneous switch into these capacitive loads is given by:

$$\Delta I = nC \frac{V_{cc}}{\tau_r} = (200) \times (5 \text{ pF}) \times \left(\frac{3.3}{10 \text{ ns}}\right) = 0.33 \text{ A} \quad 2.10$$

The lowest voltage droop allowable is 0.1 V, this is the design safety margin given in the specification for the circuit. The supply impedance to this board is therefore  $0.303\Omega$ . The PCB has 10 cm leads to the PSU (approximately 200 nH lead inductance), hence the supply can provide adequate bypass up to a frequency of 241 kHz. If the digital clock exceeds this frequency (most digital circuits use a higher clock frequency than this) then a bypass capacitor of more than 2.2  $\mu\text{F}$  is required.

Note that these calculations give the lowest recommended bypass capacitor value; for the above example a 4.7  $\mu\text{F}$  or 10  $\mu\text{F}$  would provide a better bypass. Often the actual current demand and voltage droop are difficult to gauge and the calculations cannot be made. A general rule of thumb is to use between 10 and 470  $\mu\text{F}$  per PCB, using higher value bypass capacitors for longer leads between PCB and PSU, larger PCBs with higher numbers of ICs and faster rise times circuits. The high value of the bypass capacitor usually excludes ceramic as the dielectric in most bypass applications. Consequently, aluminium and tantalum dielectric capacitors are often used in bypass applications.

### 2.3.5 Decoupling Capacitor

The values calculated for bypass capacitance can be observed on an impedance plot to not be effective at the frequencies that are typically used for switching most digital ICs (most 10  $\mu\text{F}$  capacitors for instance have a self-resonant frequency about

100 kHz). This dichotomy occurs due to the fact that the bypass is primarily aimed at preventing the local supply from drooping by more than the allowable noise margin and not directly decoupling the switching frequency noise itself. The switching frequency noise needs decoupling to ground by a lower value capacitor, specifically employed to ground out this higher frequency noise and located as close as possible to the source of the noise (i.e. close to each IC or discrete circuit).

The bypass capacitor can also have a decoupling capacitor sited with it at the PCB supply inlet. This helps filter high frequency noise and a value of approximately 1/100–1/1000 of the bypass capacitor is commonly used (e.g. for a 10  $\mu\text{F}$  bypass capacitor a 100 nF or 10 nF decoupling capacitor is placed in parallel at the inlet connection).

Even with a decoupling capacitor in parallel with the bypass there is often insufficient noise rejection from switching at the individual ICs. There is still a relatively large track impedance between the supply inlet decoupling capacitor and each IC or discrete circuit function. There needs to be a decoupling capacitor at each IC and close to any discrete circuit performing a high speed function (e.g. transistor mixer circuits). Consequently, for improved EMC performance there are going to be a number of discrete low value capacitors distributed around the circuit and located close to each IC or discrete switching function.

The traditional method is to use the standard capacitance-current and voltage-time derived equation as used to calculate the current demand in the bypass capacitor example above. The equation for current to/from the capacitor is:

$$i = C \frac{dV}{dt} \quad 2.11$$

Rearranging for the capacitance ( $C$ ) for a given current demand ( $i$ ), assuming the rise time ( $\tau_r$ ) and allowable noise voltage ( $\Delta V$ ) are known;

$$C = i \frac{\tau_r}{\Delta V} \quad 2.12$$

Take an example of an oscillator circuit with a rise time of 2 ns (regardless of actual oscillator frequency), driving a 50 mA line and as this is in the same circuit as our previous bypass capacitor example, a maximum voltage noise of 100 mV. The calculated decoupling capacitor is 1 nF, again a slightly higher value adds in some margin for error, say 2.2 nF or 4.7 nF. Too large a value may not decouple the harmonics adequately and again reference to the capacitor impedance plot may be necessary. These equations only really give boundary conditions for capacitor selection and do not necessarily compute the optimum value.

A less rigorous and less laborious method is to use a value selected from the impedance analysis curves (doing the bypass calculation for a few PCBs is not too problematic, but for hundreds of ICs calculating the decoupling capacitance gets a bit tedious). Capacitors which are still capacitive and have a low ESR at the maximum switching frequency will provide suitable decoupling. Even capacitors which are beginning to behave inductively may be suitable providing their



impedance is adequately low, as a general guide an impedance below  $1\Omega$  is adequate decoupling for most digital ICs.

Using the absolute switching frequency may not be adequate if there are very fast rise ( $\tau_r$ ) and fall ( $\tau_f$ ) times involved as these tend to produce harmonics that can dominate the frequency spectra. An estimate of the frequency content of a fast pulse edge ( $f_{edge}$ ) can be obtained from the equation:

$$f_{edge} = \frac{1}{\pi\tau_r} \text{ or } f_{edge} = \frac{1}{\pi\tau_f} \quad 2.13$$

Note: Usually the rise time is the dominant edge in EMC design and fall times are generally ignored. This is primarily due to the predominance of rising edge triggered logic.

For an ALS logic gate for example, even when driven by a 4 MHz clock, the rise time is faster than 4 ns, which gives a harmonic content of almost 80 MHz. Some may argue that significant harmonic content can still be present at up to 10 times this figure (see section on integrated circuits for suggested decoupling capacitor values for use with specific logic families).

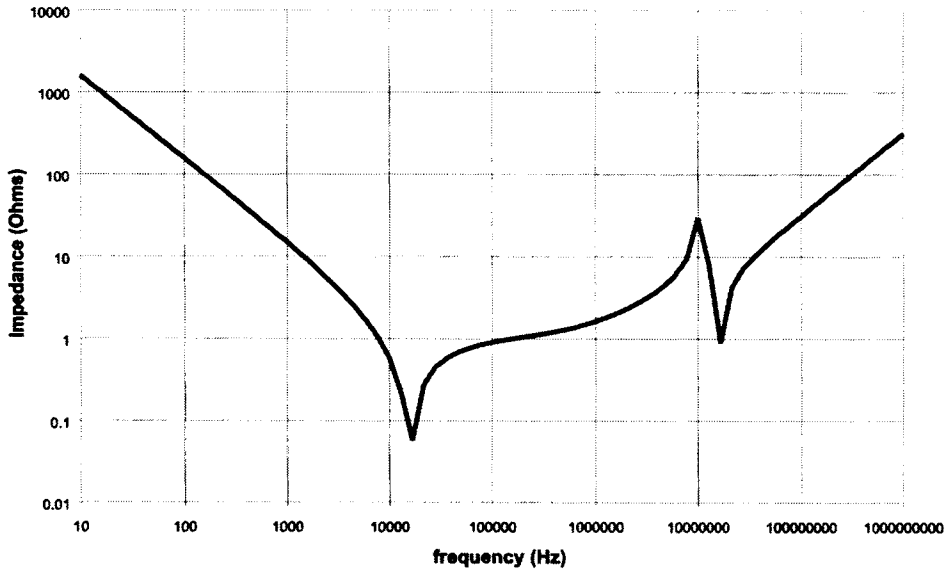
The exact choice of capacitor for decoupling is difficult to predict accurately as many different effects may come in to play. If the rise and fall times are 10 ns or less, a value chosen on the clock frequency alone is usually adequate. With rise and fall times under 10 ns use the above equation to determine the upper operating frequency range of the decoupling capacitor and select a suitable value that meets this frequency specification.

Some recent experimental studies even suggest that values between 4.7 nF and 100 nF will be suitable for virtually all IC applications with clock frequencies above 33 MHz. The study also suggested that a 10 nF decoupling capacitor is an optimum value for digital circuits up to 100 MHz. Certainly this saves a lot of time in calculating suitable values and saves cost if all decoupling capacitors are of an equal value.

One thing is important to note, lead and track length connecting decoupling capacitors is even more critical than with bypass capacitors. The decoupling capacitor must have a minimal parasitic inductance if it is to operate effectively. Often the values of capacitance are already low due to the frequency range required, hence to have any decoupling effect at all lead and track length must be minimised (this is covered in the PCB section).

### 2.3.6 Parallel Resonance

When different values of capacitor, or different dielectric types, are connected in parallel (as with a bypass and supply decoupling capacitor) then a resonant mode can be introduced. Connecting capacitors in parallel is not necessarily a bad idea as it extends the operating frequency range of the capacitive effect (bypass or decoupling) as well as maintaining a low impedance to a large range of frequencies. However,

**Figure 2.15**

Capacitor parallel resonance

due to phase interaction and multiple parasitic inductances between each capacitor self-resonant frequency, there is a third resonance mode introduced.

These resonant modes will be present regardless of capacitor choice, the designer has little they can do other than to be aware of their presence. Because of this effect it is usually recommended that only one decoupling capacitor and one bypass capacitor be connected directly in parallel. If more filtering than can be achieved by two capacitors in parallel is required it is recommended that a series inductor or ferrite bead is used to separate sections of capacitors to increase filtering and reduce the interaction of these resonance peaks.

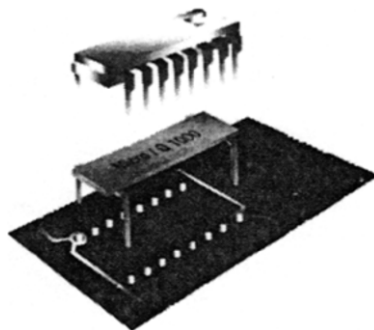
### 2.3.7 EMC Specific Capacitors

There is a special construction of capacitor called a feedthrough capacitor which exhibits very flat impedance-frequency response once the ESR value is reached. This type is used for filtering of signals into systems from external sources and can have a metallised body contact for direct mounting on to grounded panels or equipment bulk heads. They have been designed specifically with EMC in mind for signal and power line filtering applications, consequently they are expensive and offer only a limited range of values. There are ranges of SM feedthrough capacitors emerging in the market for low level signal lines, but limited values are available and their

effectiveness relies on low ground impedance, hence proper PCB layout considerations for signal filter grounding is required.

In bypass applications there is a range of higher frequency aluminium electrolytic types which include an organic semiconductor in their dielectric. One type, sold under the trade name OS-CON™ (a trade mark of Sanyo Electric Company Ltd), offers extremely low ESR and high self-resonant frequency at high capacitance values (for example, a 47  $\mu\text{F}$ , 16 V type has ESR  $<0.1$  and self-resonant frequency  $>1$  MHz). Certainly for circuits which contain high frequency switching power devices these types offer a better bypass solution compared with standard aluminium and tantalum capacitors of a similar value. As might be expected they are more expensive and are not available from as many sources as standard electrolytic capacitors. Similar aluminium electrolytic types with special polymer dielectrics are also being developed by other capacitor manufacturers.

For decoupling applications in combinational logic and memory circuits there are a range of flat plate capacitors which sell under the trade name MICRO/Q™ (a trade mark of Circuit Components Incorporated). These sit beneath various sizes of dual-in-line (DIL) and pin grid array (PGA) through hole ICs and some plastic leadless chip carrier (PLCC) SM packages. The parts are constructed from ceramic dielectric and fit into the same PCB holes as the IC and can be retrofitted to existing designs. These devices exhibit very low parasitic elements and occupy no additional PCB space, hence are extremely neat solutions to the decoupling of ICs, especially on PCBs which do not have a ground plane. The down side is that they are considerably more expensive than a single discrete capacitor, the supply and ground pins have to be in a specific configuration (i.e. 7 and 14 on a 14-pin DIL IC) and the range of values is limited. Despite the limitations the devices provide an extremely convenient solution to existing designs which were not correctly decoupled and new designs where there is no space for a discrete decoupling capacitor.



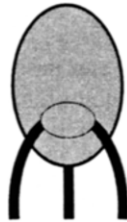
**Figure 2.16**

MICRO/Q decoupling capacitor for standard DIL logic gates, courtesy of Circuit Components Inc

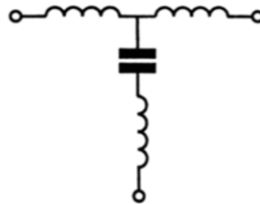
The simplest way to avoid problems of self-resonance in standard capacitor types is to use components specifically designed for high frequencies (UHF, VHF and microwave applications, often on porcelain ceramic). This type of capacitor, commonly referred to as high-K, are more expensive than standard ceramic capacitors and may only be available in SM packages. The range of capacitance values are limited (up to 1000 pF) but handling and availability are comparable with standard SM ceramic types (MLCCs).

Three terminal capacitors are an attempt to accept that a lead has inductance and use a third lead to compensate by forming a T-filter. The component gives significantly improved high frequency performance compared with an equivalent valued 2 terminal type, but lead lengths should still be minimised, especially the ground lead connection. The frequency performance can be made to exceed SM ceramic capacitors if the ground lead is sufficiently short and connected directly to a low impedance ground (i.e. a ground plane).

3 Terminal Capacitor



Circuit Representation

**Figure 2.17**

3 terminal capacitor

The EMC regulations have probably given one of the biggest marketing boosts in capacitor technology development over the last 20 years. This is likely to continue for some time with other dielectric materials, such as relaxor dielectrics, making a niche for themselves for certain EMC-specific problems.

### 2.3.8 EMC Critical Capacitor Applications

The most critical EMC application for capacitors are the bypass and decoupling applications already covered. There are further uses in filters and suppression circuits which also rely on capacitor performance to achieve the required circuit function.

In filtering applications it should be remembered that the capacitor has a limited frequency range over which it behaves capacitively. Low impedance is not a suitable

parameter to select filter capacitors on as after self-resonance the phase angle is rotated  $180^\circ$  from  $-90^\circ$  to  $+90^\circ$ . If producing a filter choose a capacitor that is still acting as would be expected (i.e. capacitively). The best indications are a high self-resonant frequency and a low equivalent series resistance value. Also use the dielectric frequency table to ensure you have an appropriate choice of material. Much of a capacitor's ability to operate is lost when a long or thin PCB trace is used to connect it to its target circuit (this places an inductance in series with the capacitor terminal), the capacitor must be located close to its target circuit and connected with short traces.

In digital circuit decoupling applications it is a low ESR value that is more important than the capacitor's self-resonant frequency. Capacitors that are beginning to behave inductively, but still offer a low impedance path to ground can still provide adequate decoupling for most digital gates. This argument cannot be applied to analogue circuits as phase interactions can usually affect the circuit function.

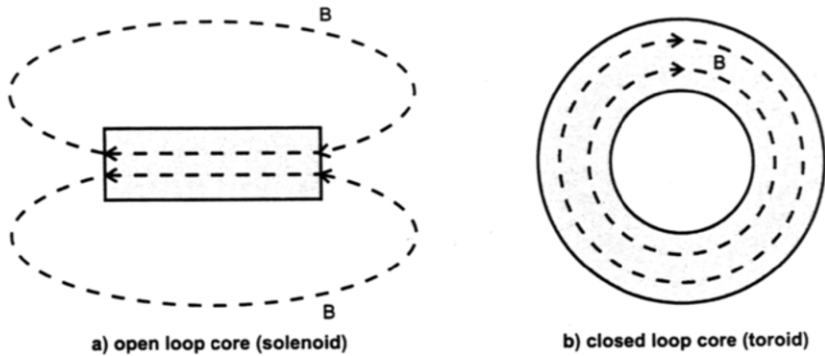
## 2.4 Inductors

Inductors are a circuit element that a lot of designers prefer to avoid, probably due to a lack of understanding of their application and uses. Like capacitors, inductors are a potential cure for many EMC-related circuit problems and it is expected that their use will increase. Consequently, it is well worth spending some time re-evaluating them as circuit elements.

One of the reasons designers may have avoided inductive components is their potential to cause EMI as well as suppress it. The circuit forms a link between magnetic and electric fields, hence are potentially more susceptible than other components as they have an inherent ability to interact with magnetic fields. This conception may be based more on supposition than measured fact, even so it is worth considering the form or shape of the inductor construction prior to use as certain shapes are more problematic than others.

### 2.4.1 Construction

There are numerous shapes for the core materials used in inductor construction, but these can be reduced to just two types: closed loop and open loop. Open loop types are those in which the magnetic field deliberately passes through air to complete the magnetic circuit. The easiest open loop shape to imagine is the rod core (solenoid), in which there is a large magnetic field from the ends of the rod passes through the air around the rod to complete the magnetic circuit loop. Closed loop shapes can be considered as those in which the magnetic field is contained in the core itself (or is intended to be contained). The easiest closed loop to visualise is the toroid or ring core, in which the magnetic field is contained solely within the form factor of the core material.

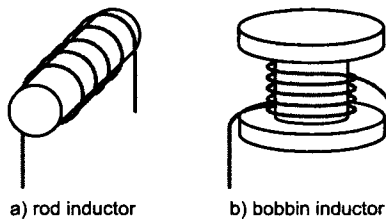


**Figure 2.18**

Magnetic field in inductor core

Leads do not tend to be a problem on inductors as the parasitic inductance value is usually very small compared with the design value. Consequently, unlike most other components, the use of SM or leaded makes little difference to the EMC performance; this is dominated primarily by the core shape and winding styles. It is worth noting that SM inductors are only capable of supporting relatively low current values (up to a few hundred mA in most cases). Surface mount types are primarily aimed at signal lines where low current handling is not a problem. In power line applications the physical size and weight of the inductor core makes through hole the preferred mounting option.

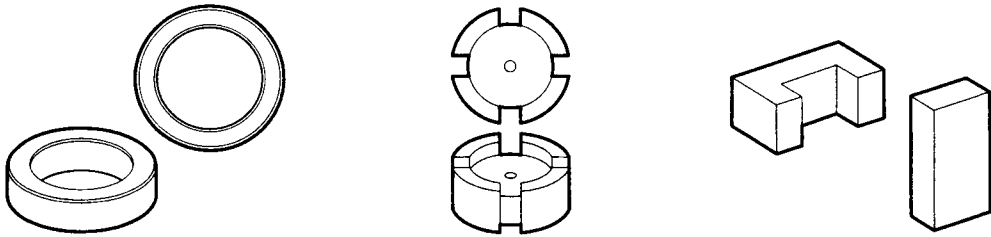
The open loop inductor shapes are obviously the biggest potential EMC problem. The fact that the magnetic field deliberately exceeds the component body means that the potential for radiating a magnetic field is higher, and susceptibility to radiated fields is also increased. The coupling to other components positioned close to these open loop inductors is potentially high and they can be difficult to place in a circuit where high frequencies may be present or in which component density is high. The rod inductor shape is definitely the worse for EMC performance, the magnetic field fringes a large area around the rod core. The bobbin shape, which is also an open loop form and probably more popular than the rod inductor, has a relatively localised magnetic field, fringing to the open section of the core shape in quite a small air loop.



**Figure 2.19**

Open loop conductors

Closed loop cores are often more expensive than open loop shapes and more expensive to wind, hence only a few closed loop core shapes are used for inductor components. More closed loop shapes are used in transformer design and will be discussed later. The most popular closed loop inductor is the toroid, which is also the least problematic as far as EMC is concerned. The toroid not only contains the generated magnetic field within the core itself, but any incident field radiating on to the shape creates an equal and opposite field in the winding, hence has a self-cancelling effect. The one other popular closed loop inductor shape is the pot core. This type has a closed loop surface, but requires exit slots for the wires. Consequently, some fringing around these slots can occur, but generally this is very small and highly localised. Pot cores are two part symmetrical cores and offer high inductance values and high operating frequencies, hence tend to be used in more specialised applications.



**Figure 2.20**

Some closed loop cores

To reduce cost of materials, inductors are often wound directly on to the core itself. The exception to this is the pot core where a bobbin former is used to hold the winding and enclosed in the core material at a later assembly stage. The winding sometimes dominates the cost of the finished part, hence the open core being less expensive as these shapes are simple to machine wind.

Unlike capacitors which have a large range of materials, inductors have basically only two choices of core material: iron or ferrite. This is a bit of a narrow classification as iron cored materials can include iron powder cores and laminated silicon-iron types. Likewise, there are a number of ferrite material compounds, the main two which account for in the region of 95% of ferrites are manganese-zinc (MgZn) and nickel-zinc (NiZn). The iron-cored devices are used for lower frequency applications (up to a few tens of kHz) as the iron compounds tend to have a limited magnetic frequency response. The ferrite compounds can be used into the high MHz frequency range, but are limited below a few kHz. Consequently, the use of inductors in EMC applications tends to be based on the ferrite cored devices.

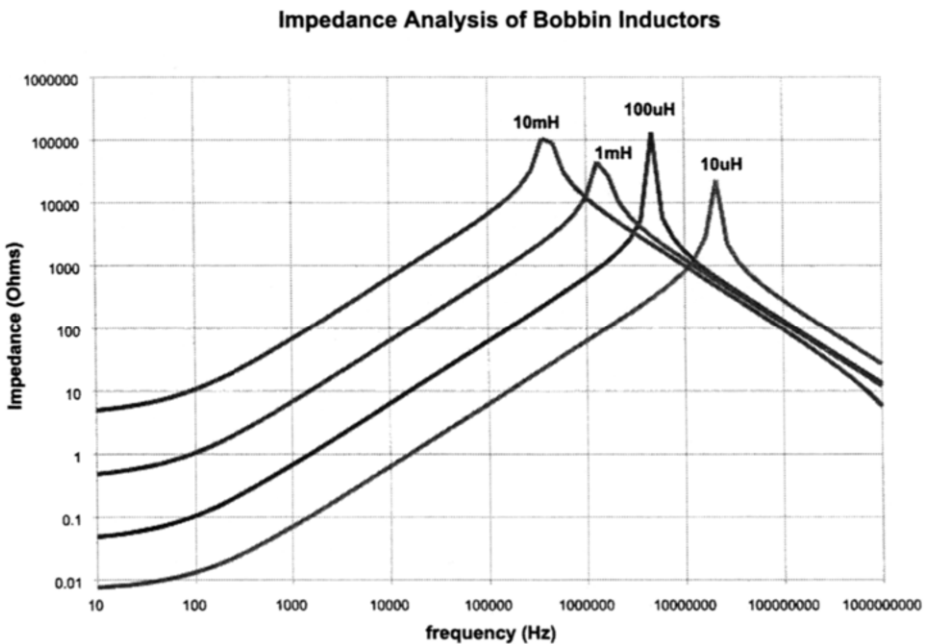
There are also a range of inductors with no magnetic core: air-cored inductors. These are usually low cost, low inductance devices that operate over a wide frequency range consisting of a stiff piece of wire wound into a core shape (most often a rod). In general, they are not recommended for EMC applications as they exhibit

extremely large flux leakage, offer very small inductance values (up to 1  $\mu\text{H}$  typically) and act as receiving antennae to stray fields.

## 2.4.2 Inductor Impedance Analysis

Inductors, like capacitors, have a self-resonant frequency ( $f_o$ ). If an inductor consisted of a single turn the self-resonant frequency would represent the ferromagnetic frequency limit (i.e. the limit that the material can 'flip' magnetic domains). In most inductors, with more than 1 turn, this self-resonant frequency is limited by the self-capacitance of the winding wire ( $C_w$ ):

$$f_o = \frac{1}{2\pi\sqrt{LC_w}} \quad 2.14$$



**Figure 2.21**

Impedance of various inductors

The impedance curve of an inductor is therefore similar to that of a capacitor inverted, that is the impedance rises from DC to a high impedance peak, then falls as the parasitic wire capacitance dominates the impedance curve. There is a lower limit to the impedance determined by the DC resistance of the wire ( $R_{dc}$ ) and an upper limit to the peak determined by magnetic loss in the core, analogous to the ESR of a



capacitor. The magnetic loss can be considered for analysis as a resistive element ( $R_p$ ) in parallel with the inductor in the electrical circuit.

Fortunately (unlike most capacitor data sheets), the value of self-resonant frequency is usually quoted in the inductor specification. This allows not only the appropriate inductor to be chosen for the circuit application, but gives some idea of the likely resonant modes that may appear with a wide band signal applied (i.e. likely attenuation of EMI).

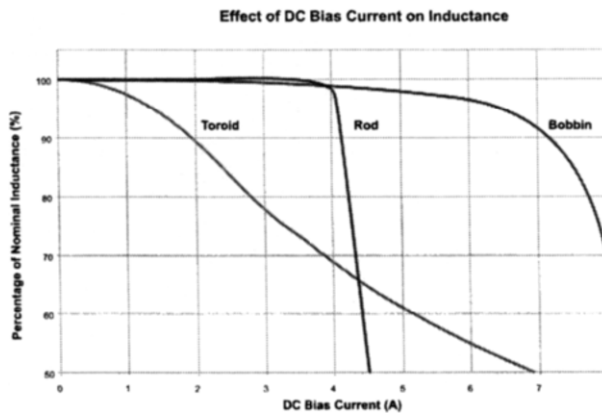
The core loss factor is not quoted directly but implied by a parameter called the quality factor ( $Q$ ). The quality factor is a measure of the inductive reactance to magnetic loss in the ideal operating band of the inductor frequency range (i.e. at a frequency where the impedance is dominated by the inductive reactance):

$$Q = \frac{R_p}{2\pi f_o L} \quad 2.15$$

This equation can also be utilised in a circuit where the resonant peak is too sharp and causes reflection or ringing problems. Rearranging the equation for the resistance ( $R_p$ ) can give a suitable resistor value to place in parallel with the inductor to produce a lower  $Q$  circuit. (Note that the quality factor of a single inductor is analogous to the quality factor of an LC tank oscillator circuit.)

### 2.4.3 Bias Dependence

Inductors also have a bias dependency similar to a ceramic capacitor voltage bias, but current rather than voltage dependent. The inductance value falls under applied DC bias current as part of the core is magnetised and consequently unavailable to attenuate AC signals.



**Figure 2.22**

Inductor DC bias current

Not all inductors actually suffer this as much as others, high saturation cores (e.g. rod and gapped pot cores) tend to exhibit a thermal runaway effect before DC current saturation occurs. Toroid inductors tend to offer the lowest DC saturation for unit magnetic volume due to the enclosed core shape. Bobbin-shaped inductors suffer some saturation; however, the inductance value tends not to 'collapse' as rapidly as in toroid cores. Closed loop cores can be designed with the same values of current handling as open loop shapes; however, they are larger as they require a greater volume of core to achieve the same saturation characteristics.

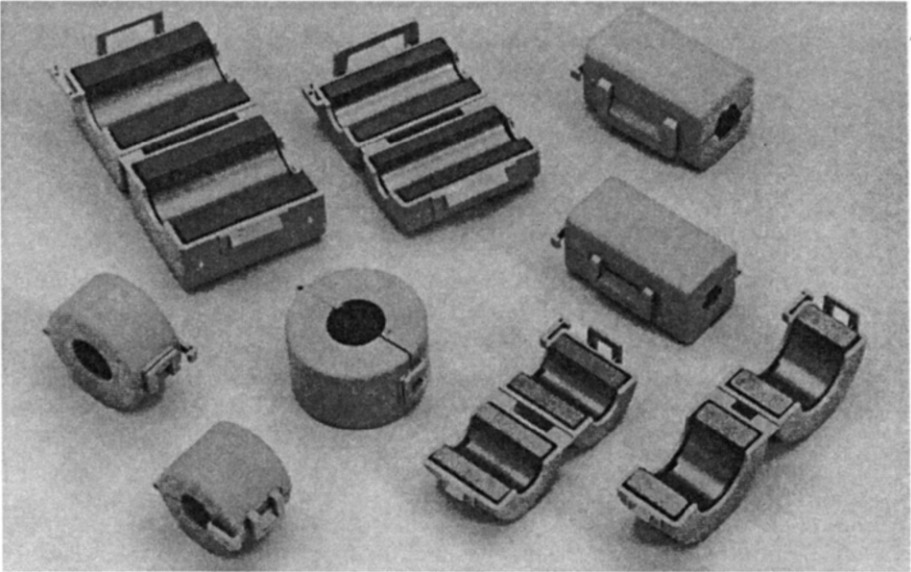
There is therefore a hierarchy of selection on DC bias current which runs counter to the case for low emissions, as the lower emission cores (closed loop) suffer the most due to current saturation. The best way to avoid saturation when using closed loop cores is to ensure that the DC bias current is below the maximum value specified for the inductor. This current specification does not include any AC value, if there is a known significant AC current (often the AC value is not known in a noise suppression choke) the r.m.s. value plus any additional DC bias should not exceed the maximum DC current specification ( $I_{dc}$ ).

The effect of this current bias-related droop of inductance is a reduction in the attenuation by the inductor on a noise or signal source. In an EMC-related application, such as a filter, this results in a reduction in the rejection and change in the effective frequency range. Usually, the changes are relatively small (less than 10% of nominal value) and a filter should always be designed to encompass the possibility of change in the passive elements under different load conditions. The effect is quite easy to simulate in a circuit simulation program (e.g. SPICE) or even in a spreadsheet.

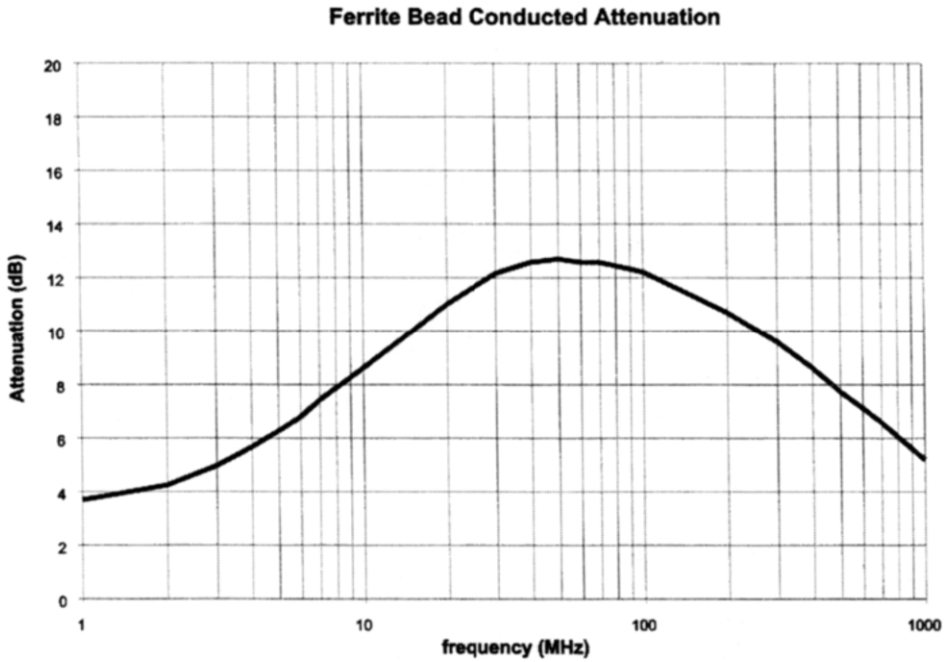
### 2.4.5 EMC Specific Inductors

Ferrite beads are actually single turn inductors. The bead is usually slipped over a wire or has a single lead through the ferrite material to form the one turn. The advantage of single turns inductors is very high frequency attenuation range (the ferrite material is specifically formulated for high frequency operation) with very low losses at DC and low frequencies (up to several hundred kHz). Ferrite beads can also offer advantages in that they are easily retrofitted to a wire after the system has been designed. Hence if a high frequency problem is occurring that has only come to light at the end of a product design cycle these parts are easily added. The down side is that they have very low attenuation (typically 10 dB) and have to have a wire or placement available in the circuit.

Ferrite clamps act in a similar manner to ferrite beads, except they are usually applied over a cable bunch or ribbon cable. They offer a degree of both common mode (CM) and differential mode attenuation. Attenuation again occurs in the high MHz region and the clamp can be retrofitted to an existing design or cable installation. As with ferrite beads only a low level of attenuation is obtained (10–20 dB). Retrofitting is simple, most cable forms can have a clamp placed over them which has been specifically manufactured for the cable form. The effect this

**Figure 2.23**

Ferrite clamps, courtesy of Richco International

**Figure 2.24**

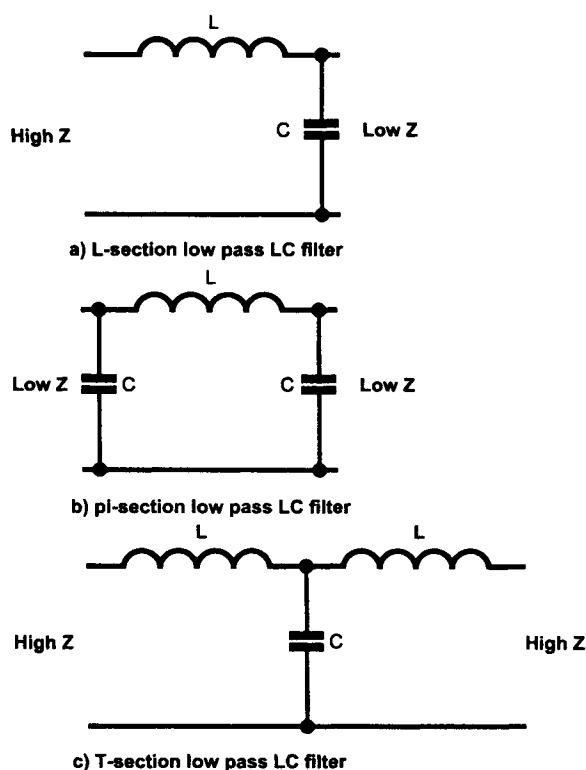
Typical attenuation of ferrite bead

could have on signal skew should be borne in mind if there are high frequency signals in the cable, especially single-ended signals.

Both beads and clamps are especially useful in dissipating transient energy from electrostatic discharges induced into a wire or cable loom. The ferrite dissipates much of the fast rising edge energy and dampens the transient seen at the exit of the wire or cable. More attenuation can be achieved by using larger ferrite beads and clamps or by adding more to either the same wire/cable or to additional wires and cables. These components do not necessarily have to be designed in at the start, but several SM bead components which include the single wire are coming on to the market, suggesting that these products are not used solely as retrofit components.

### 2.4.6 EMC Critical Inductor Applications

Inductors are commonly used for energy storage in DC-DC converters where the inductor form can have a significant effect on the EMC performance of the circuit. In power circuits the preference would be for toroid cores due to near zero emissions;



**Figure 2.25**

LC filter sections

however, if the toroid saturates high current conducted noise could be introduced. Often the bobbin shape is used as a good compromise between low emissions and high saturation current.

The best use for inductors is in the filtering of power supply lines where the low DC resistance produces very low losses. In particular, LC filters can be easily arranged to form an impedance matching bridge between different impedance circuits, such as a low impedance supply and a high impedance digital circuit. High impedances should be connected via a series inductor whereas low impedance circuits can have a bypass capacitor in parallel. The simple passive filter is easy to design and implement and high order frequency rejection can be easily obtained by cascading filter sections.

## 2.5 Transformers

The application area for a transformer usually decides the selection criteria and the EMC benefits are often considered 'in-built' due to the offer of galvanic isolation. There are areas which can help improve the EMC performance but often these can be at the expense of functional performance.

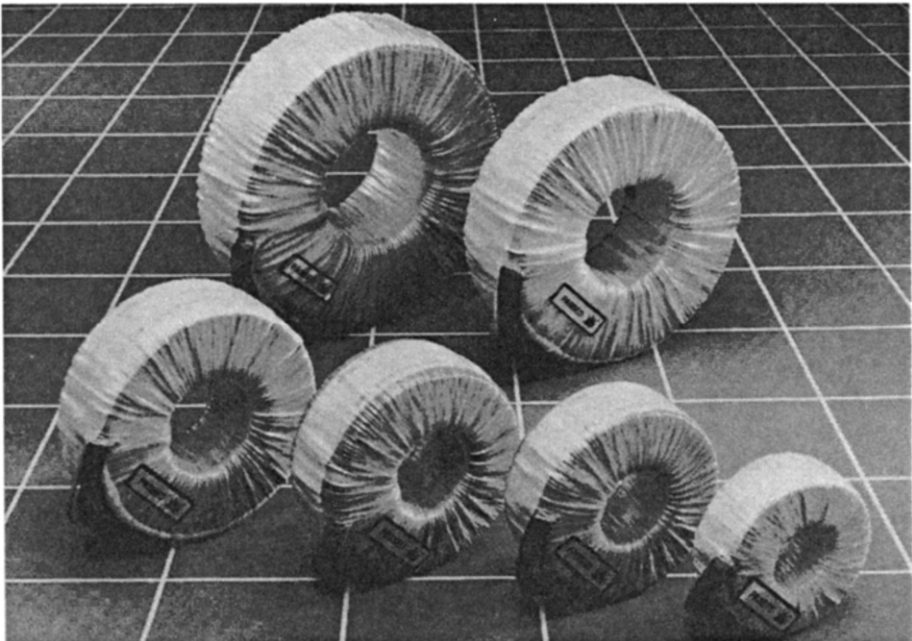
The transformer as a passive component is often overlooked in the same manner as inductors. In mains isolation applications often the lowest cost and smallest power units are used, in switched mode power supplies size of component dominates and usually means the higher frequency the better, and in signal isolation pulse shape determines choice. In all these application areas there are other considerations which can be used to ensure EMI is minimised.

### 2.5.1 Construction

The construction of mains isolation transformers with laminated iron cores for linear supplies is limited to virtually two choices: E cored or toroidal. The E core is the most popular due to lower cost and small size. Toroidal cored mains transformers cost more in both materials and winding cost and are usually significantly heavier as well as larger than an E cored transformer of the same power rating. As with inductors, however, the toroid offers advantages for reduced emissions and susceptibility even at low frequencies.

Switched mode power supplies (SMPS) usually utilise a ferrite core material and the most popular core form factors are pot cores, RM cores and E cores (including planar E core for very high frequencies). The toroidal core is unfortunately not as popular in SMPS due to its saturation capability (see section on inductors); hence from an EMC viewpoint the pot core and RM form offer the best shapes for reduced radiated emissions and susceptibility.

Signalling transformer construction will depend on the frequency of signal to be transmitted. Audio and telecommunication interface transformers are usually similar to mains linear transformers, that is an E core of laminated iron construction. Similar arguments can be applied as for the mains transformer. Low kHz signalling



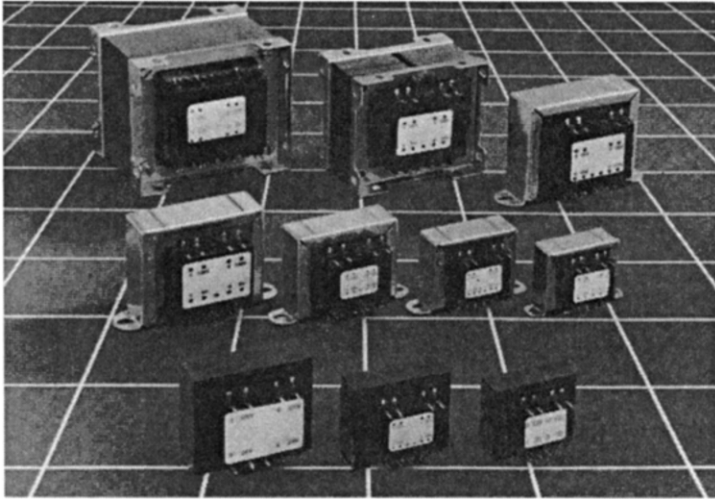
**Figure 2.26**

Toroidal mains transformers, courtesy of Newport Components Ltd

transformers and pulse transformers require shapes offering high pulse retention but in small sizes, hence CI cores, pot cores and EP shapes dominate this frequency spectrum. Higher frequency signals (100 kHz to MHz) are commonly wound on toroidal ferrites, offering the best EMC performance of all signalling types.

### 2.5.2 Power Supply Transformers

Mains isolation transformers are generally designed to operate at low frequencies (50 or 60 Hz) and the use of laminated iron cores means they are not particularly good at conducting high frequencies (the 3 dB point of most mains transformers is typically about 1 kHz). However, at higher frequencies it is capacitive coupling not magnetic coupling that links circuits on either side of the isolation barrier. The parameter which gives some indication of the amount of conducted noise that may be transmitted is the isolation or interwinding capacitance. This should be low for linear mains transformers as the physical separation requirement for safety regulations can result in only a few hundreds or a few tens of picofarads of interwinding capacitance, resulting in very good high frequency noise rejection up to several MHz. With in-line filters and decoupling capacitors usually attached to rectification circuits the linear power supply can offer a very good conducted noise barrier for high frequencies, often sufficient in themselves to require no further noise filtering if power demand is low (under 50 W). Higher power linear supplies can usually be improved by the addition of simple line filtering using appropriately rated inductors and capacitors.



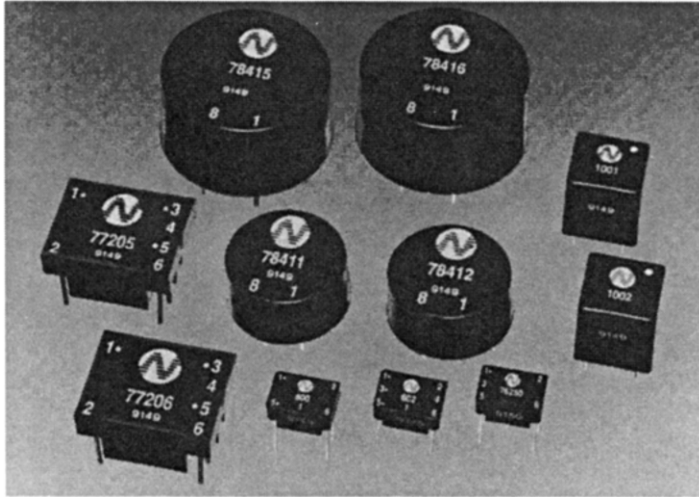
**Figure 2.27**

Laminated EE cored mains transformers, courtesy of Newport Components Ltd

Owing to the nature of SMPS operating at high frequencies (typically 25 kHz and higher) the circuit produces much more conducted noise than linear supplies. The SMPS has gained popularity due to high efficiency and small size (about 10 times reduction in size compared with similar power linear designs) and not its EMC friendliness. Additional filtering will be required if using an SMPS circuit as a direct off-line supply. Even for in-board DC–DC conversion SMPS circuits produce a high conducted noise and require filter considerations (see section on DC–DC converters). The filtering required to reduce the noise from the SMPS circuit itself is usually a good filter for other incident noise generated by the target circuit or injected into the input line. However, as the filter is usually not specifically constructed for frequencies much above the SMPS oscillator frequency, additional line filtering and protection may be required. Another potential drawback with SMPS circuits is that the interwinding capacitance of the transformer is often quite high to reduce stresses on the power components. This can result in SMPS transformers providing a lower level of immunity to conducted noise than a linear transformer.

### 2.5.3 Pulse and Signalling Transformers

At low frequencies up to the audio band (<20 kHz) signal transformers tend to be quite bulky as signal hold-up requires significant energy storage. For telecom and some audio circuits the silicon–iron EE core construction dominates. These offer a good high frequency noise rejection until the coupling capacitance begin to be the dominating impedance and high frequency signals can capacitively couple over the isolation barrier. Unless the attached circuit already has some high frequency rejection (i.e. slow amplifier circuits) then some additional simple filtering of the



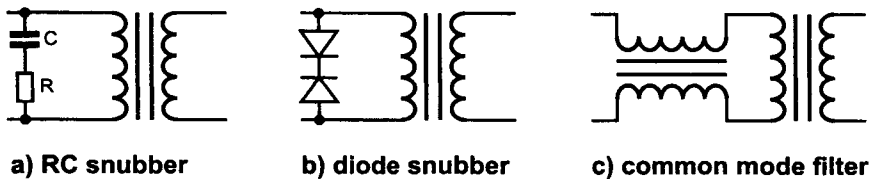
**Figure 2.28**

Pulse transformers, courtesy of Newport Components Ltd

signal, such as an RC or diode snubber, may be required. Another simple method is to use a similar transformer with a 1:1 ratio as a CM choke.

The best low kHz frequency construction for radiated emissions are those with almost closed outer surface, such as EP cores, as there is little or no fringing along these surfaces. Other shapes that can work well are those which approach the configuration of a toroid (e.g. C-I cores) and attempt to maintain the field within a predefined loop structure.

At higher signalling rates (20 kHz up to several hundred kHz) toroidal transformers are common and radiated EMI is close to zero. Conducted noise is usually controlled by matching between signalling transformers and careful selection of cabling and stub locations (e.g. Ethernet). Separated windings offering low interwinding capacitance can be easily implemented in high frequency signalling parts, improving the noise immunity further.



**Figure 2.29**

Transformer filter circuits



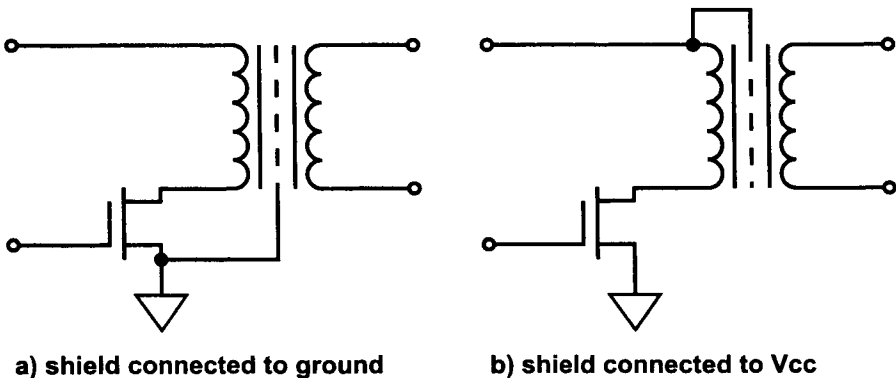
Layered windings help maintain reduced coupling capacitance and internal physical separation also helps. Some damping to minimise overshoot can also be applied either resistively or with a diode snubber type circuit external to the transformer.

Additional care to reduce the effect of reflections and potential ringing within signalling systems can be achieved by careful selection of the impedance profile of the signalling transformers. The transformers are optimised to have a peak impedance across their network or signal side at the optimum transmission rate. Consequently, the impedance at the signalling rate is higher than the out of band impedance. The peak impedance can, however, be too high and present a sharp impedance boundary which can cause overshoot in the signal and hence create a high frequency harmonic. The in-band impedance should be between 10 and 100 times that of the just out-of-band impedance, but less than 1000 times. This more gentle roll off of impedance produces a small benefit to the overall EMC of a single system, but as signalling can affect several systems is often worth the additional consideration (see section on transformer impedance analysis).

#### 2.5.4 Transformer Shielding

Interwinding capacitance (sometimes referred to as isolation or coupling capacitance) of transformers can be significantly reduced by the use of shielding between the windings. An interwinding shield, termed a Faraday shield or electrostatic screen, consists of a thin conductive sheet layered over the windings. This can be connected to a common point on either primary or secondary winding, usually the local ground or the supply is adequate. If the ground is not a reference point at the transformer (e.g. a switched ground circuit) the shield should be connected to the steady-state terminal and not the switched terminal.

Additional shielding can be added for primary and secondary local grounds with an intermediate safety or mains ground between them. This level of isolation shielding



**Figure 2.30**

Transformer shield connection

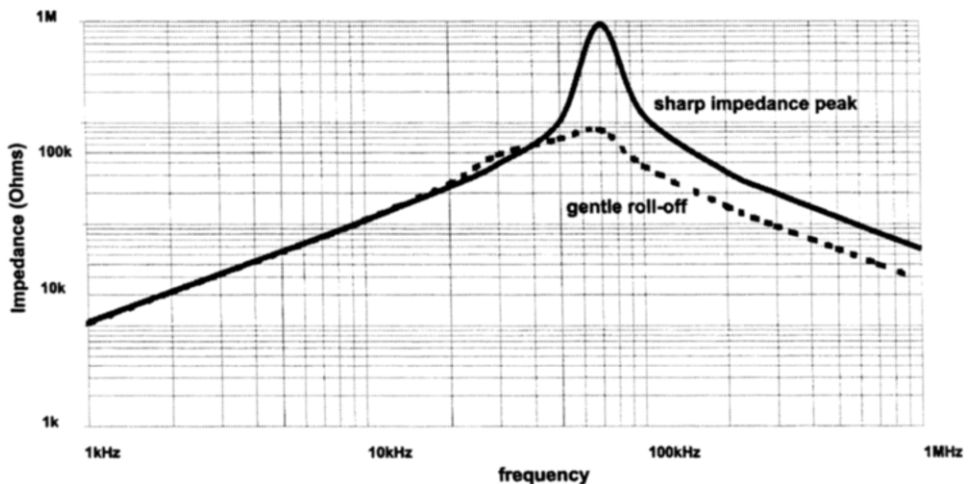
can produce interwinding capacitance in the sub-pF range but adds to the expense of the transformer.

In low frequency and power applications the wave shape is usually unaffected by shielding but in signal applications a reduction in rise time could result if shielding is applied to high frequency signals (the shield provides capacitive coupling to ground, hence acts as a low pass filter). Also in high power SMPS applications this reduction in interwinding capacitance is often associated with a rise in leakage inductance which can stress the power components.

Completely enclosed shielded transformers are available, particularly for mains applications. The actual benefit of the shield is mainly in preventing low frequency mains hum from interfering with components located close to the transformer. The shield may give the winding wire some additional immunity to RF interference but this is not the original intention of the shield and the benefits are small.

### 2.5.5 Transformer Impedance Analysis

As with inductors, transformers have a self-resonant frequency. This is not often quoted for mains and low frequency transformers, but for signalling transformers this is usually close to the peak signalling frequency. The peak impedance shape can be used cautiously as another indicator of EMC performance. A very sharp peak impedance gives very high rejection to 'out of band' signals but can produce an easily resonant circuit with cabling capacitance, particularly with signal edge harmonics. A smoother peak curve can offer good rejection while reducing the possibility of edge harmonics resonating and creating a false signal.



**Figure 2.31**

Network transformer impedance analysis

The impedance analysis of a transformer can be a useful plot to have for selecting impedance matching transformers where signal isolation over cables is required. Additional series resistors may be required, in series with the cable, if insufficient characteristic impedance at the transmission frequency is not available at the terminating end. Incorrect termination impedance causes reflections as well as signal attenuation. The greatest potential threat from a reflection are false data, emission problems would be stronger from the original signal source.

### 2.5.5 Common Mode Chokes

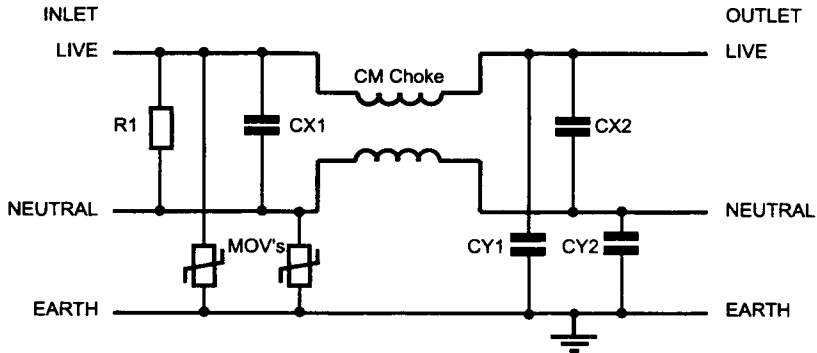
CM chokes are usually treated as inductors as their primary objective is filtering, but their construction is identical to a 1:1 transformer. CM chokes are most commonly wound on a toroidal core to minimise emission and susceptibility problems as they may be placed in close proximity to other filter components and at signal input/output ports.

This type of device is designed specifically for noise reduction and filtering, the important parameters are primary and leakage inductance. The primary inductance provides differential mode filtering, similar to a single inductor on the line. The leakage inductance produces a filter for CM noise (i.e. common to both lines) between each line and the system ground if capacitively coupled, or simply magnetic absorption of CM noise between the two lines when used without reference capacitors. The level of filtering will depend on the additional capacitors added or the parasitic capacitances within the system.

The easiest way to consider the action of a CM choke is in a filter application, such as a mains filter. Examination of a mains filter provides a good background to any passive filter design. The additional considerations in mains filter design is the earth leakage current requirement. This is a factor that should be borne in mind at each stage of design when using capacitors to provide an RF sink to a safety ground. If the safety ground or chassis is common with the mains earth then leakages must be summed to ensure the safety leakage current requirements are still met.

A standard mains line filter usually consists of five components; an X class capacitor at the inlet between live and neutral (CX1), a CM choke (sometimes referred to as a current compensated choke) between inlet and outlet (L1), an X class capacitor across the outlet sockets live and neutral (CX2) and finally two Y class capacitors between the outlet sockets and earth (CY1, CY2). Note that the mains supply is connected at the inlet side and the equipment at the outlet side. In most systems it is also advisable to add a high value discharge resistor (R1) to ensure that the capacitor charge is dissipated once power is removed, this prevents possible shock risk at the exposed mains pins. In off-the-shelf mains filter units the discharge resistor is often omitted as the user can fit this, or if feeding directly to a linear transformer this resistor is not necessary. Also shown are metal oxide varistors for protection against line transients; again these are optional and rarely included in standard filter units.

The mains filter operates in two filtering modes, differential (between the AC supply lines) and CM (between either AC supply and earth). The differential mode of operation filters out the noise between the supply lines reaching the mains and



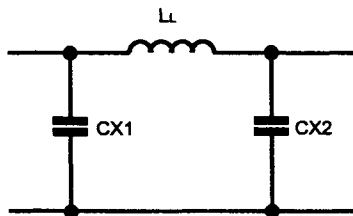
**Figure 2.32**

Mains filter circuit

likewise noise on the mains reaching the system. The CM filtering reduces noise from ground loops and earth offsets between systems.

The equivalent circuit for the differential mode signal is shown as a standard pi-filter. The X class capacitors dominate this circuit and often, where these have significantly greater capacitance than the Y class capacitors, the Y terms can be ignored. The effective inductance in each line is the leakage inductance ( $L_L$ ) of the CM choke, and consequently this is much lower than the value usually quoted for the single inductance of one winding (for example in a typical 1.6 mH mains choke, the leakage inductance is only 30  $\mu$ H). The leakage inductance is defined more by the constructional method for mains CM chokes, due to the safety requirement for separating live and neutral there is often little choice between types and a low value of leakage inductance is to be expected. In signal line filters, where the safety requirement is not critical, bifilar wound CM chokes can be used; however, leakage inductance will still be low compared with the primary inductance value.

Choice of capacitor value can be limited, typically X class capacitors are available from 1 nF to 2.2  $\mu$ F, in mains applications values between 0.1 and 0.47  $\mu$ F are common place. The 2 X class capacitors can be of different values, in fact this can



**Figure 2.33**

Differential mode equivalent circuit

be beneficial as it is unlikely that the impedance of the input mains line and system are equal. This does make frequency analysis of the filter more difficult without recourse to simulation, but allows impedance matching at both inlet and outlet terminals. For high frequency signal filtering lower value capacitors would be used.

As far as the standard EMC conducted tests are concerned we know that the differential mode impedance between live and neutral is 100Ω (CISPR 16 50Ω/50 μH LISN). If the filter is supplying a switched mode power supply, it is likely to have an impedance in the 0.1Ω region, hence matched X class capacitors are not desirable and removing CX2 completely may be beneficial. Most off-the-shelf filters are designed for 50Ω circuit impedances, and hence may not be the ideal for many situations.

The cut-off frequency is not as easy to estimate due to the inductors in both branches; however, with equal value X class capacitors and negligible Y class (CX > 50 CY) then an approximation can be derived from:

$$f_c = \frac{1}{2\pi\sqrt{L_L C_X}} \tag{2.16}$$

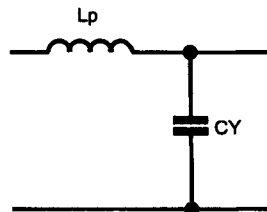
The 2 X class capacitors (and any Y class capacitance) produce the pi-filter arrangement, giving a theoretical fall off of 60 dB/decade after the cut-off frequency (in practice 50 dB/decade is more typical).

The CM equivalent circuit is a simple LC low pass filter, present in both the live and neutral lines. The inductance available is the full inductance of each single arm of the CM choke, and the combination of Y class capacitor completes the circuit. The cut-off frequency for CM signals is:

$$f_c = \frac{1}{2\pi\sqrt{L_p 2C_Y}} \tag{2.17}$$

after which the attenuation is close to 40 dB/decade based on 50Ω terminating impedances.

CM chokes for signal lines can be analysed and designed into a circuit in the same way as the above mains filter. DC current saturation is usually much higher than with single inductors as the DC magnetic field is self-cancelling, hence a high saturation current can be obtained on a toroid core.



**Figure 2.34**  
Common mode equivalent circuit