APPLICATIONS

Introduction

Soft ferrite cores are used wherever effective coupling between an electric current and a magnetic flux is required. They form an essential part of inductors and transformers used in today's main application areas:

- Telecommunications
- Power conversion
- Interference suppression.

The function that the soft magnetic material performs may be one or more of the following:

FILTERING

Filter network with well defined pass-band.

High Q-values for selectivity and good temperature stability.

Material requirements:

- Low losses
- Defined temperature factor to compensate temperature drift of capacitor
- Very stable with time.

Preferred materials: 3D3, 3H3.

INTERFERENCE SUPPRESSION

Unwanted high frequency signals are blocked, wanted signals can pass. With the increasing use of electronic equipment it is of vital importance to suppress interfering signals.

Material requirements:

- High impedance in covered frequency range.

Preferred materials: 3S1, 4S2, 3S3, 3S4, 4C65, 4A11, 4A15, 3B1, 4B1, 3C11, 3E25, 3E5.

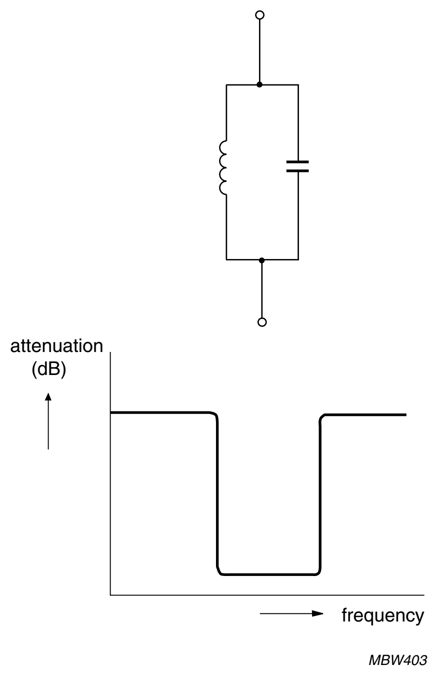

Fig.1 Filter application.

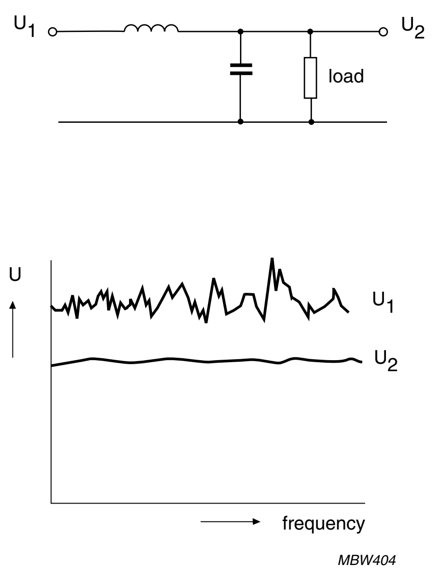

Fig.2 Suppression application.

DELAYING PULSES

The inductor will block current until saturated. Leading edge is delayed depending on design of magnetic circuit.

Material requirements:

- High permeability (μ_i).

Preferred materials: 3E25, 3E5, 3E6, 3E7, 3E8, 3E9.

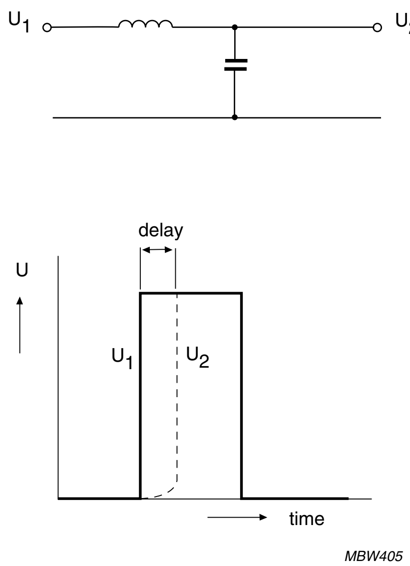

Fig.3 Pulse delay application.

STORAGE OF ENERGY

An inductor stores energy and delivers it to the load during the off-time of a Switched Mode Power Supply (SMPS).

Material requirements:

- High saturation level (B_s).

Preferred materials: 3C30, 3C34, 3C90, 3C92, 3C96 2P-iron powder.

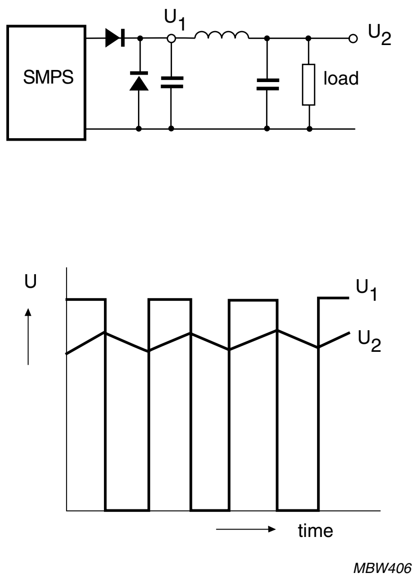

Fig.4 Smoothing/storage application.

PULSE TRANSFORMERS/GENERAL PURPOSE TRANSFORMERS

Pulse or AC signals are transmitted and if required transformed to a higher or lower voltage level. Also galvanic separation to fulfil safety requirements and impedance matching are provided.

Material requirements:

- High permeability
- Low hysteresis factor for low signal distortion
- Low DC sensitivity.

Preferred materials: 3C81, 3H3, 3E1, 3E4, 3E25, 3E27, 3E28, 3E5, 3E6, 3E7, 3E8.

POWER TRANSFORMERS

A power transformer transmits energy, transforms voltage to the required level and provides galvanic separation (safety).

Material requirements:

- Low power losses
- High saturation (B_s).

Preferred materials: 3C15, 3C30, 3C34, 3C81, 3C90, 3C91, 3C93, 3C94, 3C96, 3F3, 3F35, 3F4, 3F45, 3F5, 4F1.

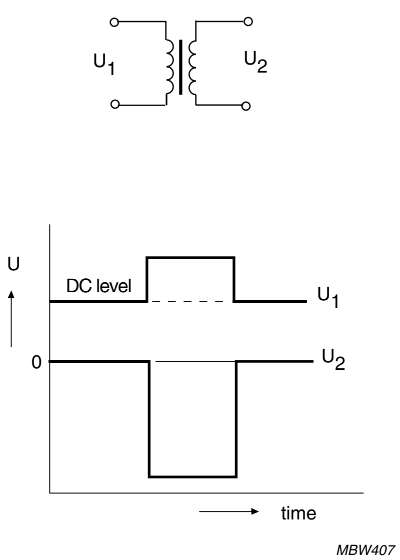

Fig.5 Pulse and general purpose transformer.

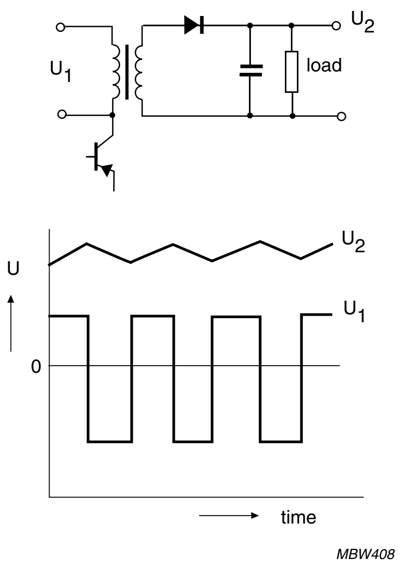

Fig.6 Power transformer application.

TUNING

LC filers are often used to tune circuits in audio, video and measuring equipment. A very narrow bandwidth is often not wanted.

Material requirements:

- Moderate losses up to high frequency
- Reasonable temperature stability.

Preferred materials: 3D3, 4A11, 4B1, 4C65, 4D2, 4E1.

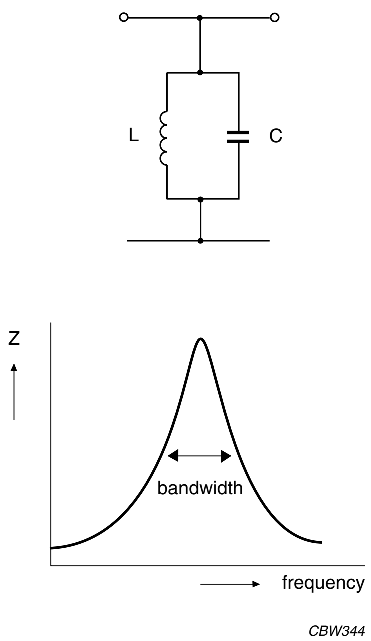

Fig.7 Tuning application.

Ferrites for Telecommunications

Telecommunications is the first important branch of technology where ferrites have been used on a large scale. Today, against many predictions, it still is an important market for ferrite cores.

Most important applications are in:

- Filter inductors
- Pulse and matching transformers.

FILTER COILS

P cores and RM cores have been developed specially for this application.

The P core is the oldest design. It is still rather popular because the closed shape provides excellent magnetic screening.

RM cores are a later design, leading to a more economic usage of the surface area on the PCB.

For filter coils, the following design parameters are important:

- Precise inductance value
- Low losses, high Q value
- High stability over periods of time
- Fixed temperature dependence.

Q VALUE

The quality factor (Q) of a filter coil should generally be as high as possible. For this reason filter materials such as 3H3 and 3D3 have low magnetic losses in their frequency ranges.

Losses in a coil can be divided into:

- Winding losses, due to the DC resistance of the wire eddy-current losses in the wire, electric losses in insulation
- Core losses, due to hysteresis losses in the core material, eddy-current and residual losses in the core material.

Losses appear as series resistances in the coil:

$$\frac{R_{tot}}{L} = \frac{R_0}{L} + \frac{R_{ec}}{L} + \frac{R_d}{L} + \frac{R_h}{L} + \frac{R_{e+r}}{L} \ (\Omega/H)$$

As a general rule, maximum Q is obtained when the sum of the winding losses is made equal to the sum of the core losses.

DC resistive losses

The DC resistive losses in a winding are given by:

$$\frac{R_0}{L} = \frac{1}{\mu_e} \times \frac{1}{f_{Cu}} \times \text{constant} \ (\Omega/H)$$

The space (copper) factor f_{Cu} depends on wire diameter, the amount of insulation and the method of winding.

Eddy-current losses in the winding

Eddy-current losses in a winding are given by:

$$\frac{R_{ec}}{L} = \frac{C_{wCu} V_{Cu} f^2 d^2}{\mu_e} \ (\Omega/H)$$

Where C_{wCu} is the eddy-current loss factor for the winding and depends on the dimensions of the coil former and core, and V_{Cu} is the volume of conductor in mm^3, d is the diameter of a single wire in mm.

Dielectric losses

The capacitances associated with the coil are not loss free. They have a loss factor which also increases the effective coil resistance:

$$\frac{R_d}{L} = \omega^3 LC\left(\frac{2}{Q} + \tan\delta_c\right) \ (\Omega/H)$$

Hysteresis losses

The effective series resistance due to hysteresis losses is calculated from the core hysteresis constant, the peak flux density, the effective permeability and the operating frequency:

$$\frac{R_h}{L} = \omega \eta_B \hat{B} \mu_e \ (\Omega/H)$$

Eddy-current and residual losses

The effective series resistance due to eddy-current and residual losses is calculated from the loss factor:

$$\frac{R_{e+r}}{L} = \omega \mu_e (\tan\delta/\mu_i) \ (\Omega/H)$$

INDUCTOR DESIGN

The specification of an inductor usually includes:

- The inductance
- Minimum Q at the operating frequency
- Applied voltage
- Maximum size
- Maximum and minimum temperature coefficient
- Range of inductance adjustment.

To satisfy these requirements, the designer has the choice of:

- Core size
- Material grade
- A_L value
- Type of conductor (solid or bunched)
- Type of adjuster.

FREQUENCY, CORE TYPE AND MATERIAL GRADE

The operating frequency is a useful guide to the choice of core type and material.

- Frequencies below 20 kHz:
  the highest Q will be obtained with large, high inductance-factor cores of 3H3 material. Winding wire should be solid, with minimum-thickness insulation.

  Note: high inductance factors are associated with high temperature coefficients of inductance.
- Frequencies between 20 kHz and 200 kHz:
  high Q will generally be obtained with a core also in 3H3. Maximum Q will not necessarily be obtained from the large-size core, particularly at higher frequencies, so the choice of inductance factor is less important. Bunched, stranded conductors should be used to reduce eddy-current losses in the copper. Above 50 kHz, the strands should not be thicker than 0.07 mm.
- Frequencies between 200 kHz and 2 MHz:
  use a core of 3D3 material. Bunched conductors of maximum strand diameter 0.04 mm are recommended.

SIGNAL LEVEL

In most applications, the signal voltage is low. It is good practice to keep wherever possible the operating flux density of the core below 1 mT, at which level the effect of hysteresis is usually negligible. At higher flux densities, it may be necessary to allow for some hysteresis loss and inductance change.

The following expression for third harmonic voltage U_3 may be used as a guide to the amount of distortion:

$$\frac{U_3}{U_1} = 0.6 \tan\delta_h$$

For low distortion, materials with small hysteresis loss factors should be used (e.g. 3H3).

DC POLARIZATION

The effect of a steady, superimposed magnetic field due to an external field or a DC component of the winding current is to reduce the inductance value of an inductor. As with other characteristics, the amount of the decrease depends on the value of the effective permeability. The effect can be reduced by using a gapped core or by choosing a lower permeability material.

A_L VALUE

Since the air gap in ferrite cores can be ground to any length, any value of A_L can be provided within the limits set by the core size. In practice, the range of A_L values has been standardized with values chosen to cover the majority of application requirements.

If a core set is provided with an asymmetrical air gap, this air gap is ground in the upper half. This half is marked with the ferrite grade and A_L value.

For very low A_L values (e.g. 16 to 25) the contribution of the stray inductance will be quite high, resulting in a marked influence of the position of the coil in the core and its number of turns.

Most pre-adjusted cores are provided with an injection-moulded nut for the adjuster.

Continuously variable adjusters can be supplied for pre-adjusted cores of most A_L values. These are specially recommended for filter coils. Maximum adjustment range is 10% to 30%, depending on core type and adjuster.

The A_L factor is the inductance per turn squared (in nH) for a given core:

$$L = N^2 \times A_L \ (\text{nH})$$

The measured A_L value of a core will depend slightly on the coil used for this measurement.

FERROXCUBEPULSE AND SIGNAL TRANSFORMERS

Pulse and signal transformers, also known as wideband transformers, are frequently used in communication systems, including modern digital networks such as, for example ISDN and XDSL.

They provide impedance matching and DC isolation or transform signal amplitudes. Signal power levels are usually low. In order to transmit analog signals or digital pulses without much distortion, good wideband characteristics are needed.

The principal function of the transformer core is to provide optimum coupling between the windings.

The general equivalent circuit of a signal transformer is shown in Fig.8.

The elements of the circuit depicted in Fig.8 may be defined as follows:

- E_s = source voltage
- R_s = source resistance
- R_w = total winding resistance = R1 + R2, where R1 is the primary winding resistance and R2 is the secondary winding resistance referred to the primary
- L = total leakage inductance = the primary inductance with the secondary shorted
- L_p = open circuit inductance
- R_p = the shunt loss resistance representing the core loss
- N1, N2 = the primary and referred secondary self or stray capacitance respectively
- R_b = load resistance referred to the primary turns ratio.

A high permeability core with polished pole faces results in a large flux contribution, improving the coupling. Open circuit inductance will be high, leakage inductance is kept low compared to this main inductance.

Ring cores are very suitable since they have no air gap and make full use of the high permeability of the ferrite.

The frequency response of a practical transformer is shown in Fig.9.

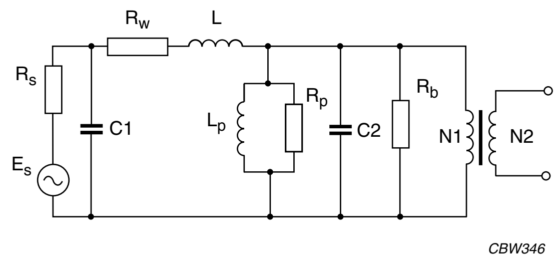

Fig.8 Simplified equivalent circuit of a transformer.

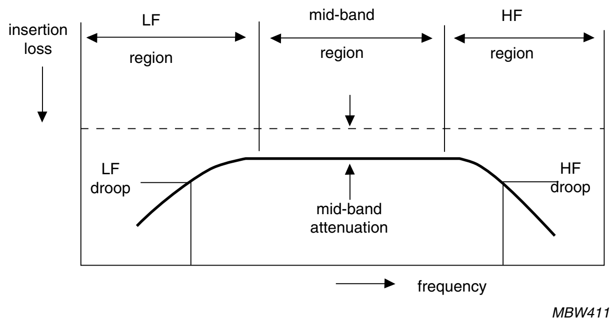

Fig.9 Transmission characteristic of a wideband transformer.

The corresponding distortion of a rectangular pulse by the same circuit is shown in Fig.10.

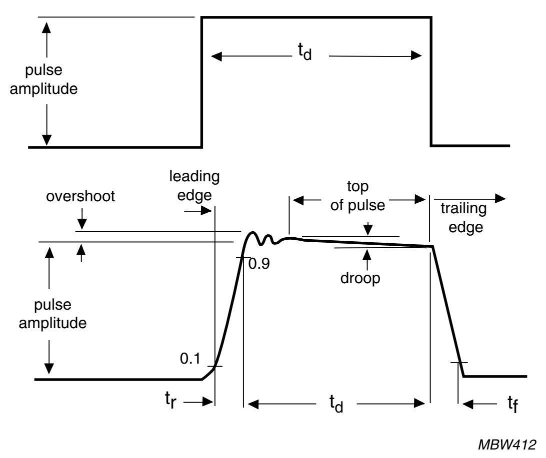

Fig.10 An ideal rectangular pulse and the main pulse distortions that may be introduced by a transformer.

The shunt inductance (L_p) is responsible for the low frequency droop in the analog transformer since its reactance progressively shunts the circuit as the frequency decreases. In the case of the pulse transformer, the shunt inductance causes the top of the pulse to droop, because, during the pulse, the magnetizing current in L_p rises approximately linearly with time causing an increasing voltage drop across the source resistance.

The winding resistance is the main cause of the mid-band attenuation in low frequency analog transformers. In a pulse transformer, it attenuates the output pulse but usually has little effect on the pulse distortion.

The high frequency droop of an analog transformer may be due to either the increasing series reactance of the leakage inductance or the decreasing shunt reactance of the self-capacitances, or a combination of both as the frequency increases. In a pulse transformer, the leakage inductance, self-capacitances and the source or load resistance combine to slow down, or otherwise distort the leading and trailing edge responses.

Suitable core types for this application in the materials 3E1, 3E4, 3E27, 3E28, 3E5, 3E55, 3E6, 3E7 and 3E8 are:

- P cores
- RM cores
- EP cores
- Ring cores
- Small ER cores
- Small E cores.

If the signal is superimposed on a DC current, core saturation my become a problem. In that case the special DC-bias material 3E28 or a lower permeability material such as 3H3, 3C81 or 3C90 is recommended.

Gapping also decreases the effect of bias currents.

Ferrites for Power conversion

Power conversion is one of the major application areas for ferrites and is generally realized by using a switched mode power supply (SMPS). The basic arrangement of a SMPS is shown in Fig.11. In a SMPS, power is efficiently converted into the voltage and current levels, required by the end-application. The wide area in which SMPSs are applied can be divided into four parts: DC-DC, DC-AC, AC-DC and AC-AC. Although every converter type can be found for power conversion, most SMPS applications are based upon the DC-DC (e.g. battery operated equipment) and DC-AC types (e.g. inverters of lamp drivers). Note that many of these converters still have an AC-DC front-end, most of the times nothing more than a rectifier, a smoothing capacitor and a filter for EMC reasons. This front-end does not belong to the SMPS itself and ferrites used in the EMC filter will be treated in the part about interference suppression.

Numerous converter types exist, but most SMPS applications make use of one of the following types :

- Buck or down converter (DC-DC)
- Boost or up converter (DC-DC)
- Flyback converter (DC-DC)
- Forward converter (DC-DC)
- Half and full bridge converter (DC-AC)

Their basic operation principle will first be discussed and after this the focus is put on how to choose the appropriate core material.

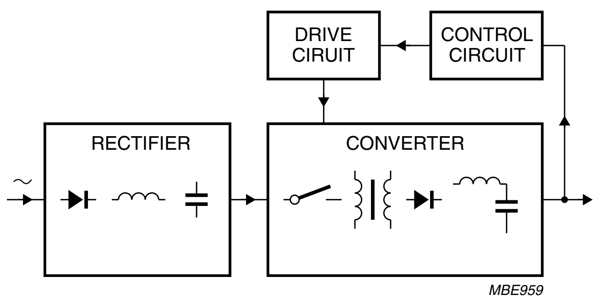

Fig.11 Block diagram of a Switched Mode Power Supply.

BUCK OR DOWN CONVERTER

With a buck or down converter, as shown in Fig.12, it is possible to adapt the input voltage to a lower level. It also means that the average output current is larger than the input current. The basic operation is as follows. During the on-time of the mosfet, a linearly rising current is flowing from the input to the output and energy is stored in the inductor (note that always the largest part of the energy is stored in the air gap and a minor part in the ferrite itself). By the end of the on-time, defined by the ratio of the output voltage and the input voltage, the mosfet is switched off. According to Lenz' law, the inductor voltage reverses and the stored energy results in a decreasing output current via the diode.

Dependent on the operating condition defined by the load, the current through the inductor will be mainly DC (continuous mode) with a triangular ripple on top of it. This means that the ferrite is operating around a DC bias point of B and H. Around this point a minor BH loop can be found.

In all those applications where a lower voltage is needed than the available supply voltage (e.g. automotive), buck converters can be found. Another application can be a so-called voltage regulated module (VRM) behind the standard computer power supply or Silver Box to deliver a stable processor voltage even under high load variations.

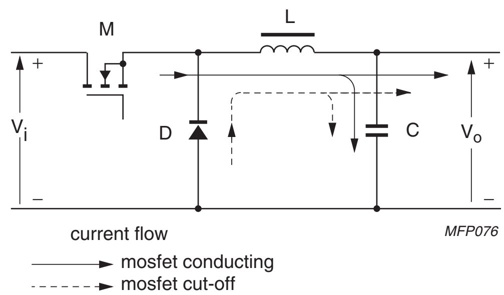

Fig.12 Circuit diagram of a buck or down converter.

BOOST OR UP CONVERTER

Re-arranging the circuit components of the buck converter results in the boost or up converter of Fig.13. It adapts the DC input voltage to a higher output level. When the mosfet is on, the inductor voltage is equal to the input voltage and the linearly rising current stores energy in the inductor. Switching off the mosfet stops the storage of energy and the inductor voltage reverses. The output voltage becomes the sum of the input and inductor voltage and a decreasing inductor current will be forced to flow to the output via the diode.

Typical boost converter applications can be found in battery operated equipment, e.g. a laptop where higher internal voltages are needed than supplied by the battery. In order to meet the stringent requirements on EMC, boost converters can also be used as power factor correction (PFC) circuits in between the mains and a SMPS. A PFC circuit ensures that a sinusoidal voltage and current are drawn from the mains, which is not possible with a SMPS only.

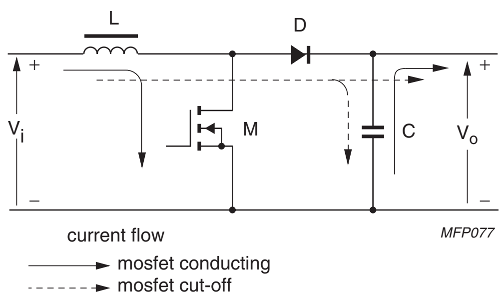

Fig.13 Circuit diagram of a boost or up converter

FLYBACK CONVERTER

One of the major drawbacks of both buck and boost converter is the absence of galvanic isolation between in- and output, which can be required by some applications. Introducing a magnetically coupled coil, like in the flyback converter of Fig.14, solves this point. But the big advantage of the flyback converter is that the output voltage can be higher or lower than the input voltage, depending on the turns ratio N. More secondary windings result in more output voltages. During the on-time of the mosfet, a linearly rising current is flowing through the primary winding and energy will be stored in the coupled coil. By the end of the on-time the primary voltage reverses and the stored energy introduces, via the magnetic coupling, a linearly decreasing current in the secondary winding. The dots close to the primary and secondary windings indicate the winding direction, necessary for good operation.

The galvanic isolation between in- and output and the possibility of multiple outputs make the flyback converter one of the most popular SMPS types. Flyback converters can be found in many applications from small low power stand-by supplies of less than 1 W to big power supplies delivering over a few kWs.

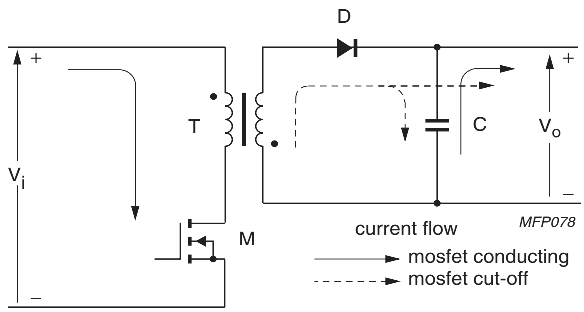

Fig.14 Circuit diagram of a flyback converter

FORWARD CONVERTER

The forward converter of Fig.15 is basically a buck converter with galvanic isolation realized by the transformer. With the turns ratio, the output voltage can be made higher or lower than the input voltage. When the mosfet is on, current is flowing through both the primary and secondary winding of the transformer and it will be magnetized. The secondary current stores energy in the coil. Switching off the mosfet releases the energy and a decreasing current is flowing to the output.
De-magnetizing of the transformer is achieved by a third winding having an equal number of turns but opposite winding direction. With it's higher component count, the forward converter is less attractive than the flyback converter.

A push-pull converter is an arrangement of two forward converters operating in antiphase (push-pull action). A push-pull converter circuit doubles the frequency of the ripple current in the output filter and, therefore, reduces the output ripple voltage. A further advantage of the push-pull operation is that the transformer core is excited alternately in both directions in contrast to both the forward and flyback converters. Therefore, for the same operating conditions and power throughput, a push-pull converter design can use a smaller transformer core.

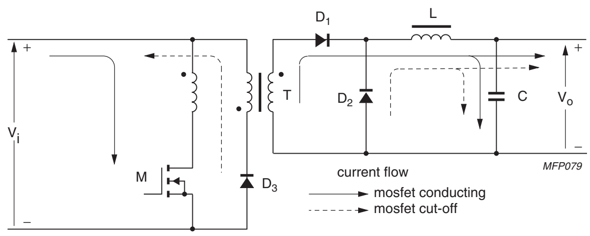

Fig.15 Circuit diagram of a forward converter.

HALF AND FULL BRIDGE CONVERTER

With the half bridge converter of Fig.16, one side of the primary winding is at a voltage potential equal to half the supply voltage. Switching the mosfets puts the other side alternately to the supply voltage and ground and therefore the primary voltage is half the supply voltage. However, with the full bridge converter, see Fig.17, and using the same transformer, the primary voltage is equal to the supply voltage. This makes the full bridge converter more efficient, but the control of two pairs of mosfets is more complicated. Transformer de-magnetizing is in both converters realized by the body diodes of the mosfets. For example, magnetizing of the transformer core is done with M1, while the de-magnetizing is done by the body diode of M2. An advantage of this principle is that M2 can be switched on during the de-magnetizing process and no switch-on losses (the so-called zero voltage switching ZVS principle) occur and less EMI is generated.

The advantage of the bridge converters compared to the previous ones (except the push-pull converter) is that the transformer is excited in two directions and therefore the full BH loop from -Bsat to +Bsat can be used. For equal throughput power, the transformer of a bridge converter can be smaller than e.g. the transformer of a forward converter operating on the same frequency.

With a secondary circuit identical to that of a forward converter, the DC-AC converter is transformed into a DC-DC converter. Still, the operating frequency of the energy storage inductor is twice the control frequency and the ripple current has been halved. Therefore, the core volume of the inductor can also be smaller.

Half and full bridge converters are normally the basis for resonant converters. In these converters, the primary inductance is a part of a resonant tank made with one or more capacitors and/or inductors. Although the resonant tank has a squarewave input voltage, sinusoidal voltages and currents appear in the tank. This means that no harmonics are introduced and in combination with the ZVS of the mosfets, it makes resonant converters very attractive for high frequency designs. Note that resonant converters directly deliver their energy to the load and no energy storage inductor is necessary.

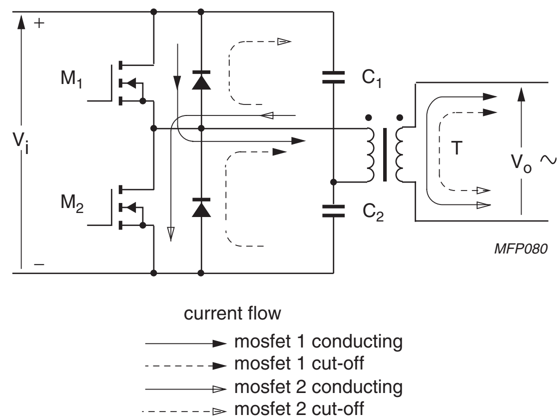

Fig.16 Circuit diagram of a half bridge converter.

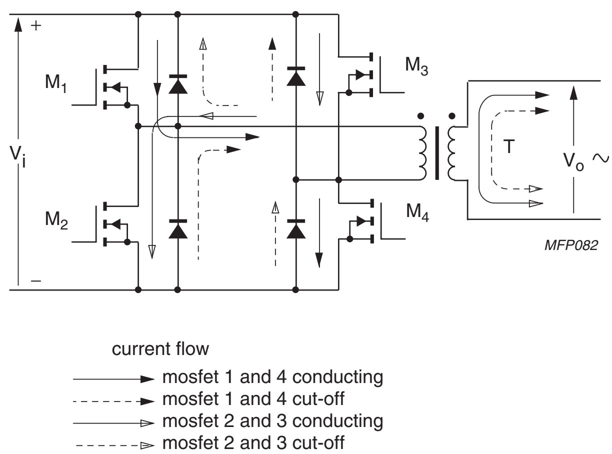

Fig.17 Circuit diagram of a full bridge converter.

FERRITE SELECTION

Dependent on the converter type, the ferrites used in these converters operate under saturation or loss limited conditions, which require special power ferrites with high saturation and low loss levels even at elevated operation temperatures.

The operating frequency is one of the parameters defining which core material can be used in an application. For

those SMPSs which are connected directly to a high input voltage (e.g. the rectified mains can be up to 400 V in some parts of the world), semiconductors with a high breakdown voltage are needed. A high breakdown voltage limits the switching frequency otherwise severe switching losses inside the semiconductor occur (on the other hand, a low voltage device can be used at much higher operating frequencies). For flyback converters things are even worse. When the mosfet is switched off, it's drain-source voltage is the sum of input and rectified secondary voltage. Therefore the operating frequency of a high voltage input SMPS is limited by the capabilities of the used semiconductors and switching frequencies up to 300 kHz can be found nowadays even when the semiconductors are connected to a heatsink. This means that most power ferrites of the 3Cxx series will mainly be used under saturation limited conditions, see also the performance factor graph in Fig.19. On top, for many applications the operating frequency is only a few tens of kHz due to EMC regulations. The reason is that these requirements can relatively easily be met in the frequency area below 150 kHz.

Converters which are not directly connected to a high input voltage and/or soft-switching power supplies, like half and full bridge (resonant) converters can overcome this problem and operating frequencies into the MHz range can be found. Power ferrites for this range are gathered in the 3Fxx series and 4F1.

The energy storage inductor of all converters (except flyback and resonant) normally operates at a bias level, therefore ferrites with a high saturation value at the application temperature, like 3C92, result in the smallest core volumes.

In case of post regulators, the operating frequency can be chosen much higher as the input voltage is much lower and the generated EMI will be sufficiently attenuated by the SMPS in front of the post regulator. Now ferrites from the 3Fxx series and 4F1 are the best choice.

All the inductors, including the coupled inductor of the flyback converter, need an air gap necessary for the energy storage, while the transformers can be made without gap.

CORE SELECTION

OPERATING FREQUENCY

The preferred operating frequency of a Switched Mode Power Supply is greater than 20 kHz to avoid audible noise from the transformer. With modern power ferrites the practical upper limit has shifted to well over 1 MHz.

Ambient temperature

Ambient temperature, together with the maximum core temperature, determines the maximum temperature rise, which in turn fixes the permissible total power dissipation in the transformer. Normally, a maximum ambient temperature of 60 °C has been assumed. This allows a 40 °C temperature rise from the ambient to the centre of the transformer for a maximum core temperature of 100 °C. There is a tendency however towards higher temperatures to increase power throughput densities. Our new material 3C93 meets these increased temperature requirements with a loss minimum around 140 °C

Flux density

To avoid saturation in the cores the flux density in the minimum cross-section must not exceed the saturation flux density of the material at 100 °C. The allowable total flux is the product of this flux density and the minimum core area and must not be exceeded even under transient conditions, that is, when a load is suddenly applied at the power supply output, and maximum duty factor occurs together with maximum supply voltage. Under steady-state conditions, where maximum duty factor occurs with minimum supply voltage, the flux is reduced from its absolute maximum permissible value by the ratio of the minimum to maximum supply voltage (at all higher supply voltages the voltage control loop reduces the duty factor and keeps the steady-state flux constant).

The minimum to maximum supply voltage ratio is normally taken as 1 : 1.72 for most applications.

SELECTING THE CORRECT CORE TYPE

The choice of a core type for a specific design depends on the design considerations and also on the personal preference of the designer. Table 1 gives an overview of core types as a function of power throughput and this may be useful to the designer for an initial selection.

Each of the core types has been developed for a specific application, therefore they all have advantages and drawbacks depending on, for example, converter type and winding technique.

Table 1 Power throughput for different core types at 100 kHz switching frequency

POWER RANGE (W)	CORE TYPE
< 5	RM4; P11/7; T14; EF13; U10
5 to 10	RM5; P14/8
10 to 20	RM6; E20; P18/11; T23; U15; EFD15
20 to 50	RM8; P22/13; U20; RM10; ETD29; E25; T26/10; EFD20
50 to 100	ETD29; ETD34; EC35; EC41; RM12; P30/19; T26/20; EFD25
100 to 200	ETD34; ETD39; ETD44; EC41; EC52; RM14; P36/22; E30; T58; U25; U30; E42; EFD30
200 to 500	ETD44; ETD49; E55; EC52; E42; P42/29; U67
> 500	E65; EC70; U93; U100; P66/56; PM87; PM114; T140

Choice of ferrite for power transformers and inductors

A complete range of power ferrites is available for any application.

3C30

Low frequency (< 200 kHz) material with improved saturation level. Suitable for flyback converters e.g. Line Output Transformers.

3C34

Medium frequency (< 300 kHz) material with improved saturation level. Suitable for flyback converters e.g. Line Output Transformers.

3C81

Low frequency (< 100 kHz) material with loss minimum around 60 °C.

3C90

Low frequency (< 200 kHz) material for industrial use.

3C91

Medium frequency (< 300 kHz) material with loss minimum around 60 °C.

3C92

Low frequency (< 200 kHz) material with a very high saturation level. Specially recommended for inductors and output chokes.

3C93

Medium frequency (< 300 kHz) material with loss minimum around 140 °C.

3C94

Medium frequency material (< 300 kHz).
Low losses, especially at high flux densities.

3C96

Medium frequency (< 400 kHz) material. Very low losses, especially at high flux densities.

3F3

High frequency material (up to 700 kHz).

3F35

High frequency material (up to 1 MHz).
Very low losses, around 500 kHz.

3F4

High frequency material (up to 2 MHz).
Specially recommended for resonant supplies.

3F45

High frequency material (up to 2 MHz).
Specially recommended for resonant supplies.

3F5

High frequency material (up to 4 MHz).
Specially recommended for resonant supplies.

4F1

High frequency material (up to 10 MHz).
Specially recommended for resonant supplies.

Performance factor of power ferrites

The performance factor ($f \times B_{max}$) is a measure of the power throughput that a ferrite core can handle at a certain loss level . From the graph it is clear that for low frequencies there is not much difference between the materials, because the cores are saturation limited. At higher frequencies, the differences increase. There is an optimum operating frequency for each material. It is evident that in order to increase power throughput or power density a high operating frequency and a better ferrite should be chosen.

OUTPUT CHOKES

Output chokes for Switched Mode Power Supplies have to operate with a DC load causing a bias magnetic field H_{DC}.

In a closed ferrite circuit, this can easily lead to saturation. Power ferrites such as 3C90 or 3F3 start saturating at field strengths of about 50 A/m. Permeability drops sharply, as can be seen in the graphs of the material data section. The choke loses its effectiveness. The new material 3C92 is optimized for use in power inductors. It features a very high saturation level as well as a high Tc, making it the best material for power inductors, especially at elevated temperatures.

There are two remedies against the saturation effect:

- The use of gapped ferrite cores
- The use of a material with low permeability and high saturation, like iron powder 2P.

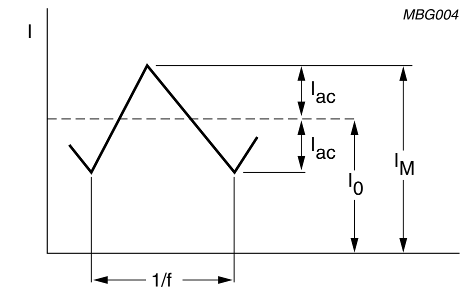

Fig.18 Choke waveform.

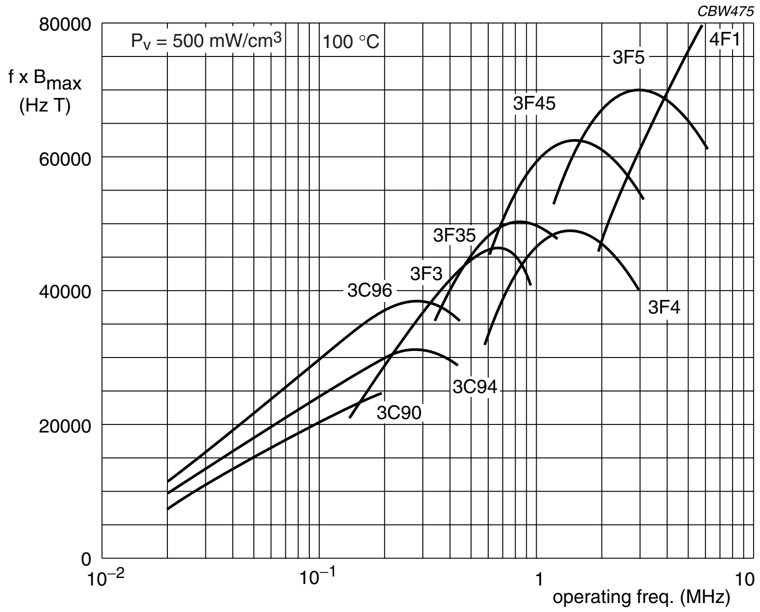

Fig.19 Performance factor ($f \times B_{max}$) at PV = 500 mW/cm^3 as a function of frequency for power ferrite materials.

GAPPED CORE SETS

The effect of an air gap in the circuit is that a much higher field strength is needed to saturate a core.

For each operating condition an optimum air gap length can be found. In a design, the maximum output current (I) and the value of inductance (L) necessary to smooth the ripple to the required level are known.

The product I^2L is a measure of the energy which is stored in the core during one half cycle.

Using this I^2L value and the graphs given on the following pages for most core types, the proper core and air gap can be selected quickly at a glance.

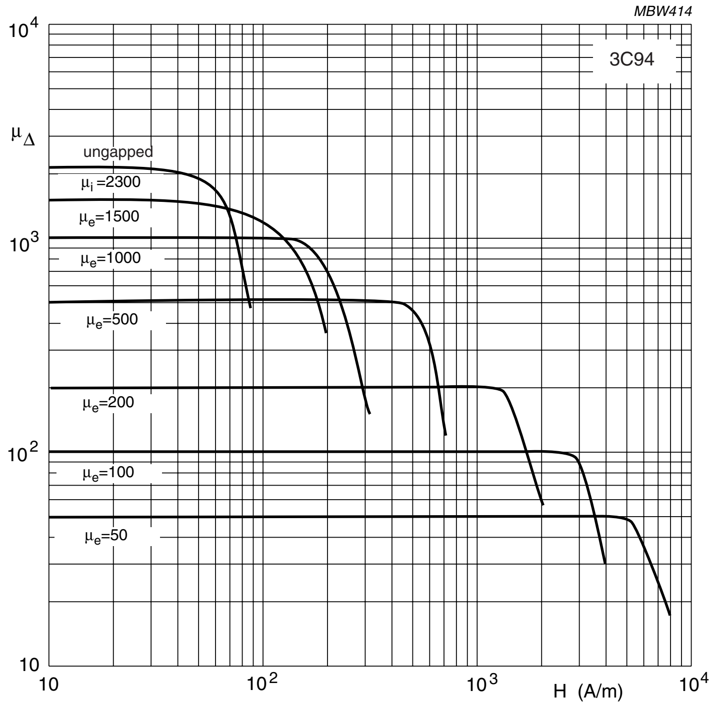

Fig.20 Effect of increased gap length.

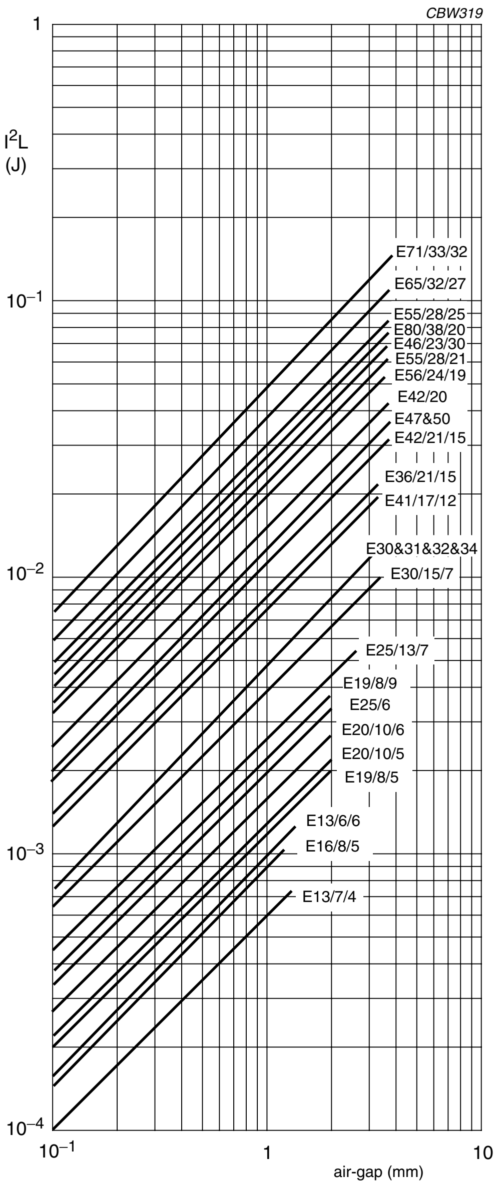

Fig.21 I^2L graph for E cores.

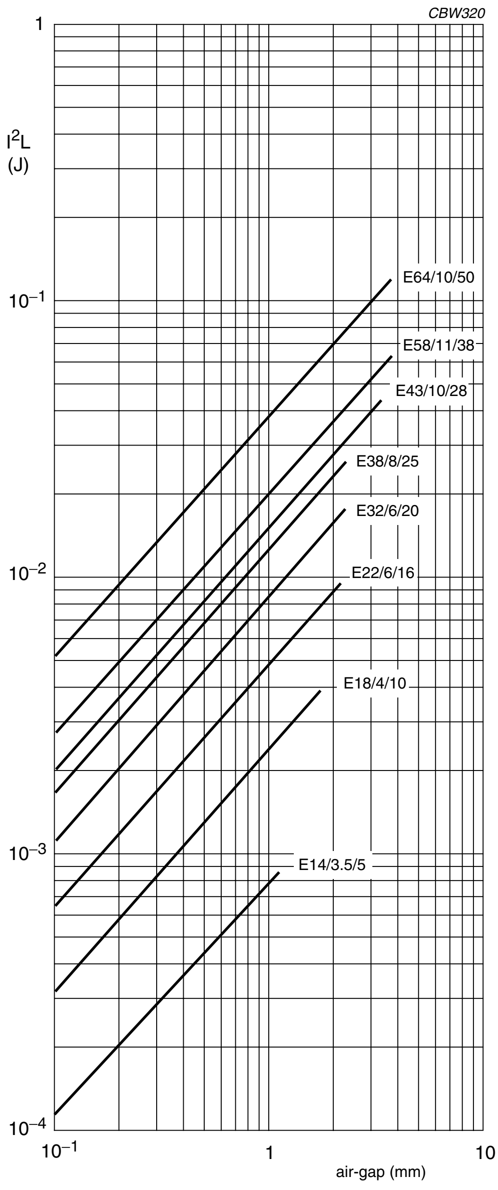

Fig.22 I^2L graph for planar E cores (valid for E + E and E + PLT combinations).

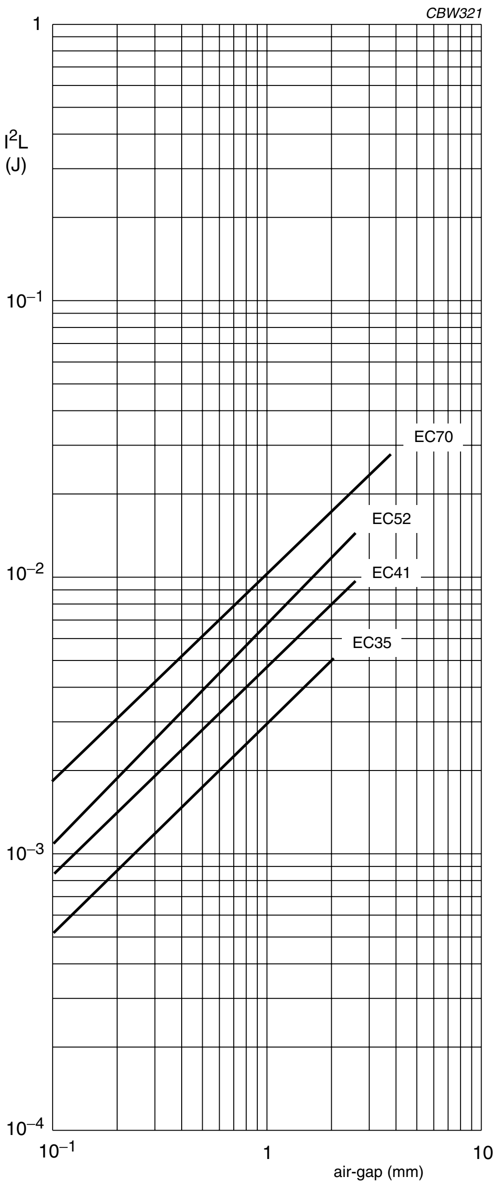

Fig.23 I^2L graph for EC cores.

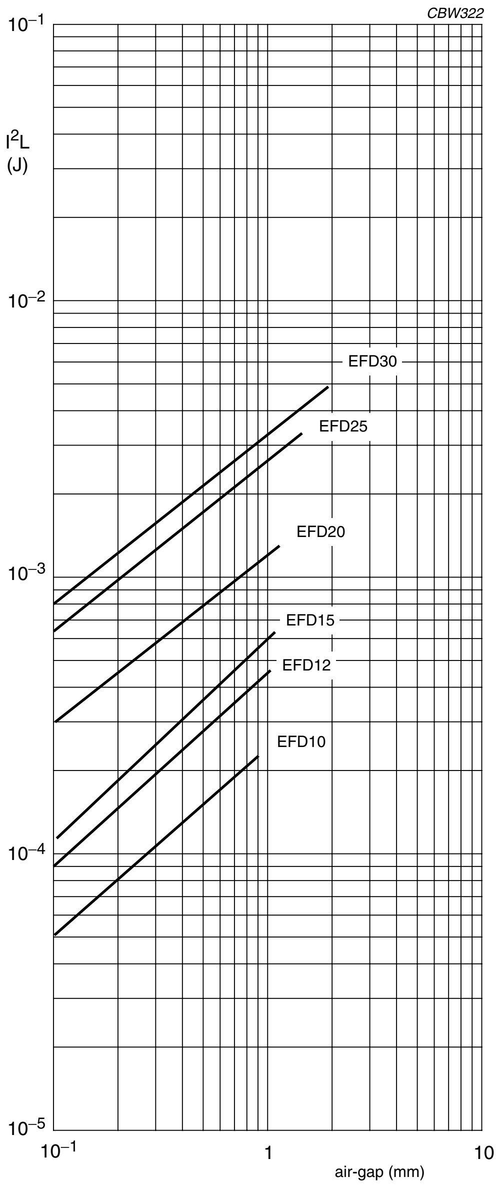

Fig.24 I^2L graph for EFD cores.

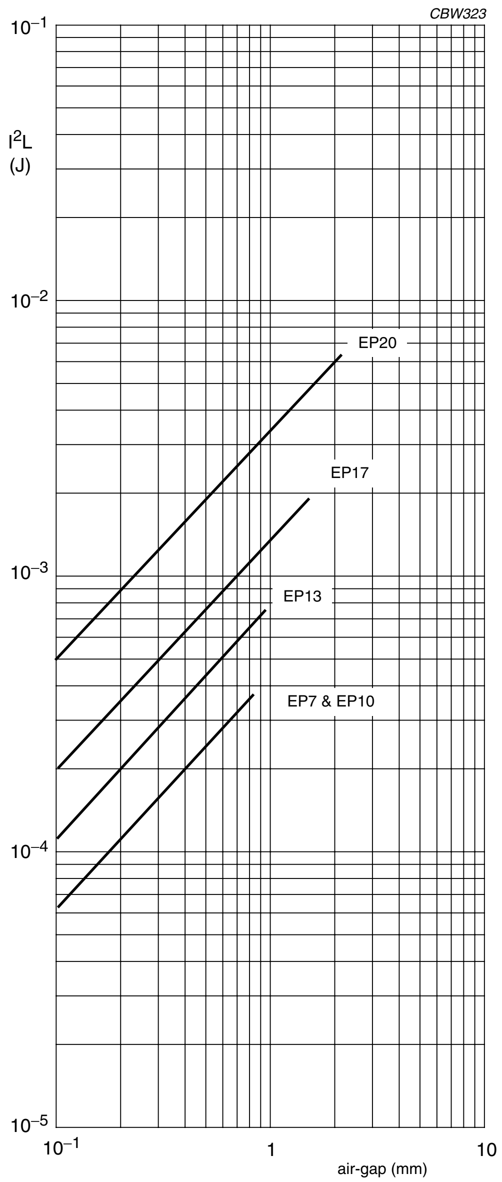

Fig.25 I^2L graph for EP cores.

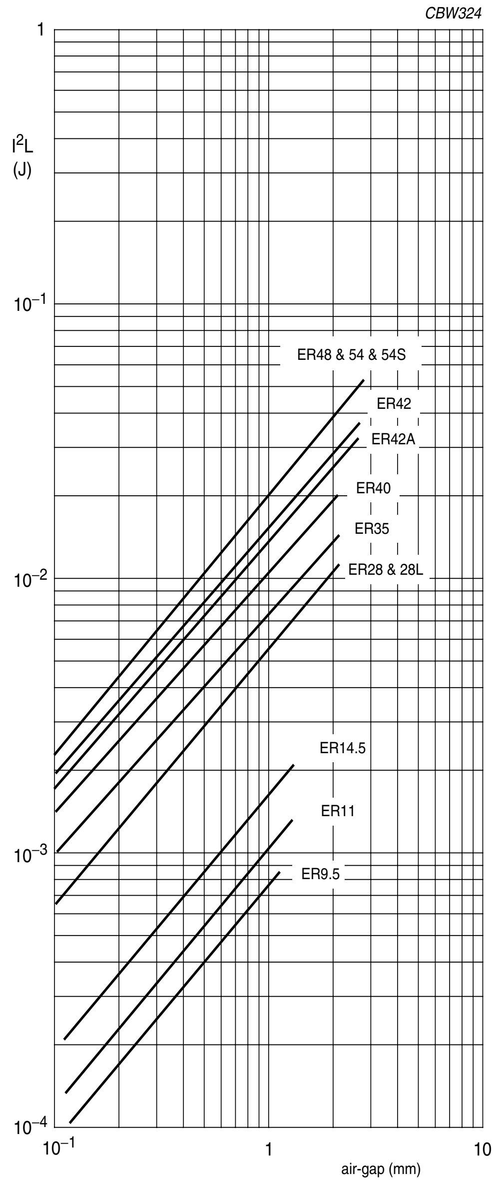

Fig.26 I^2L graph for ER cores.

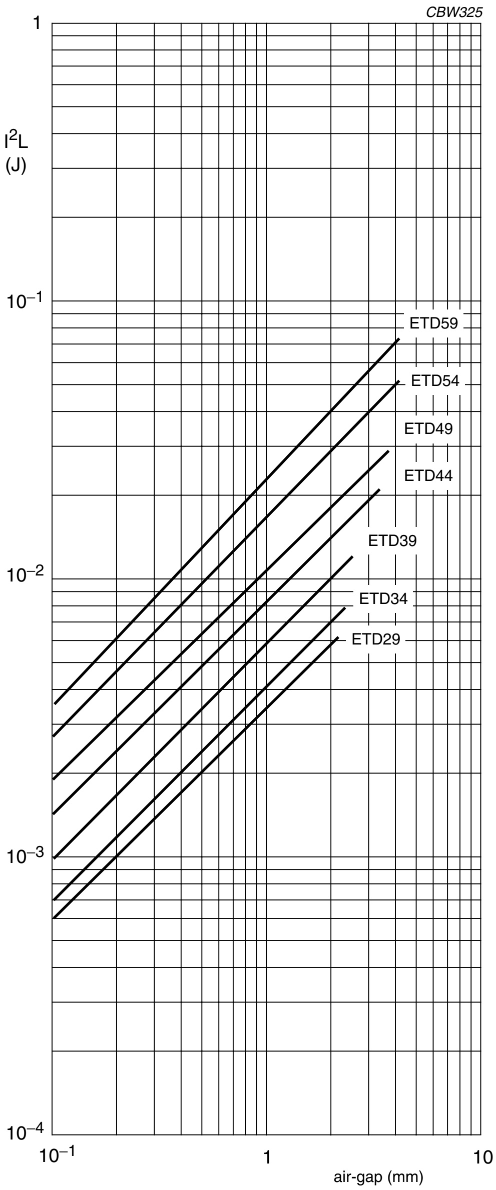

Fig.27 I^2L graph for ETD cores.

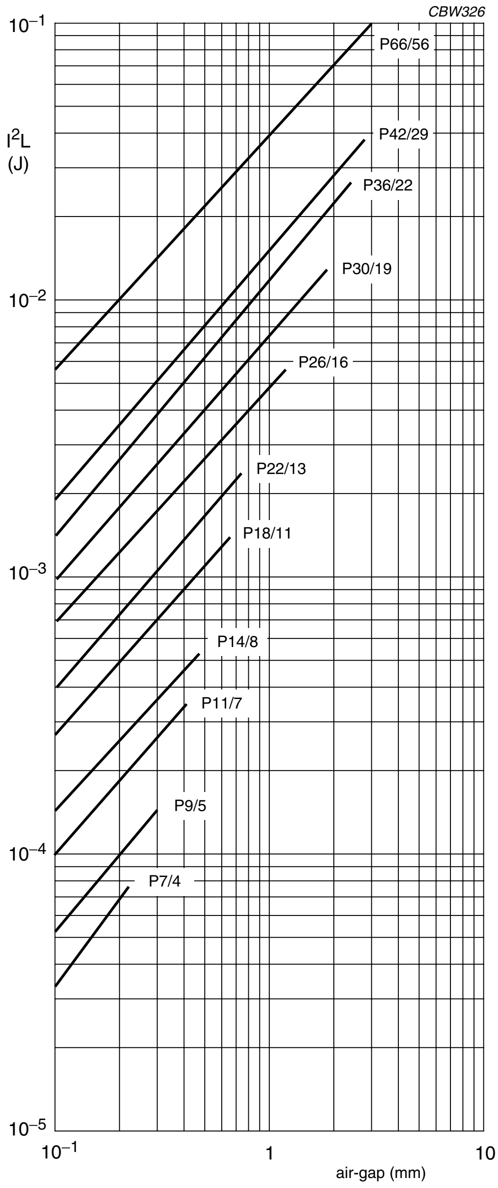

Fig.28 I^2L graph for P cores.

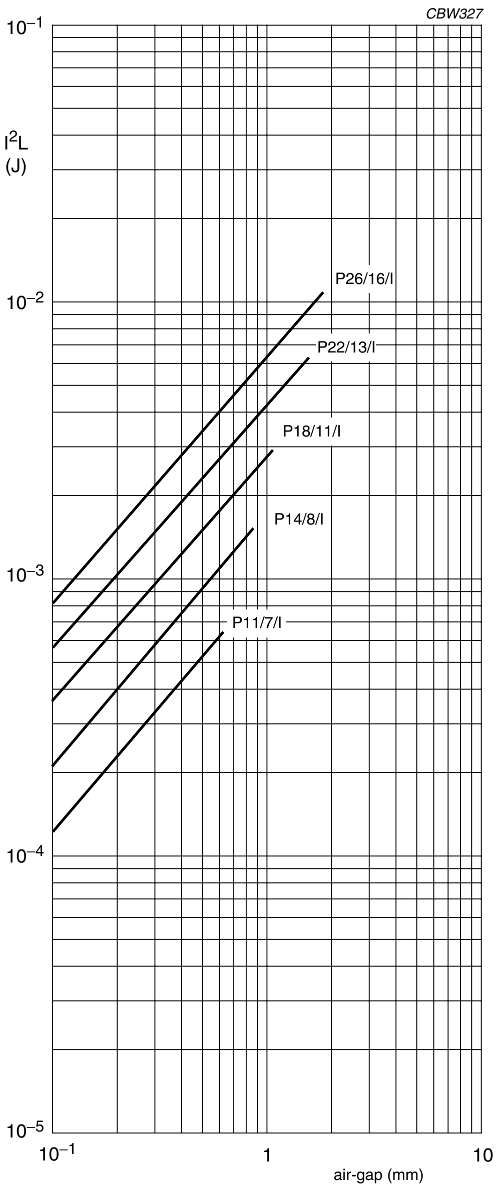

Fig.29 I^2L graph for P/I cores.

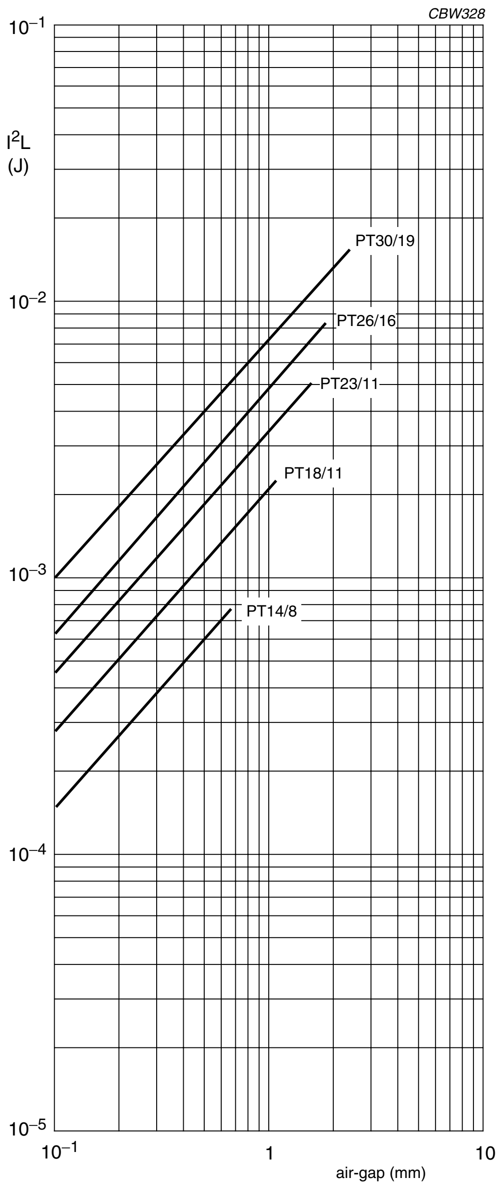

Fig.30 I^2L graph for PT cores.

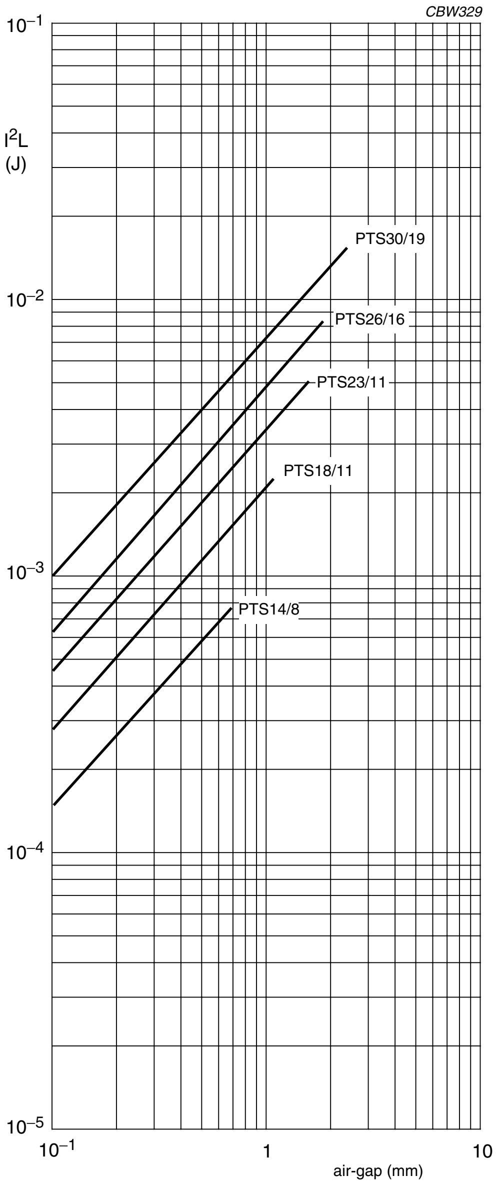

Fig.31 I^2L graph for PTS cores.

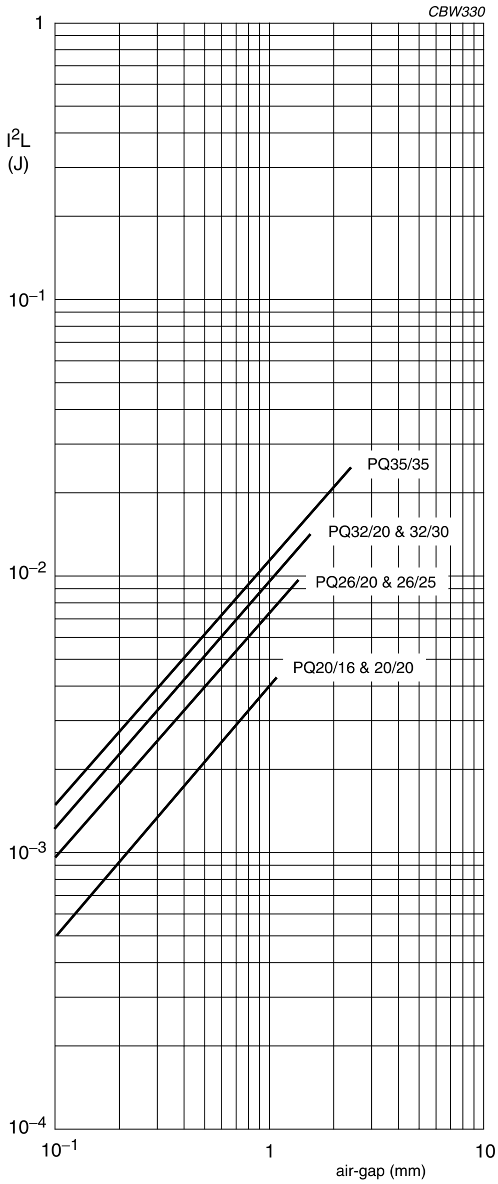

Fig.32 I^2L graph for PQ cores.

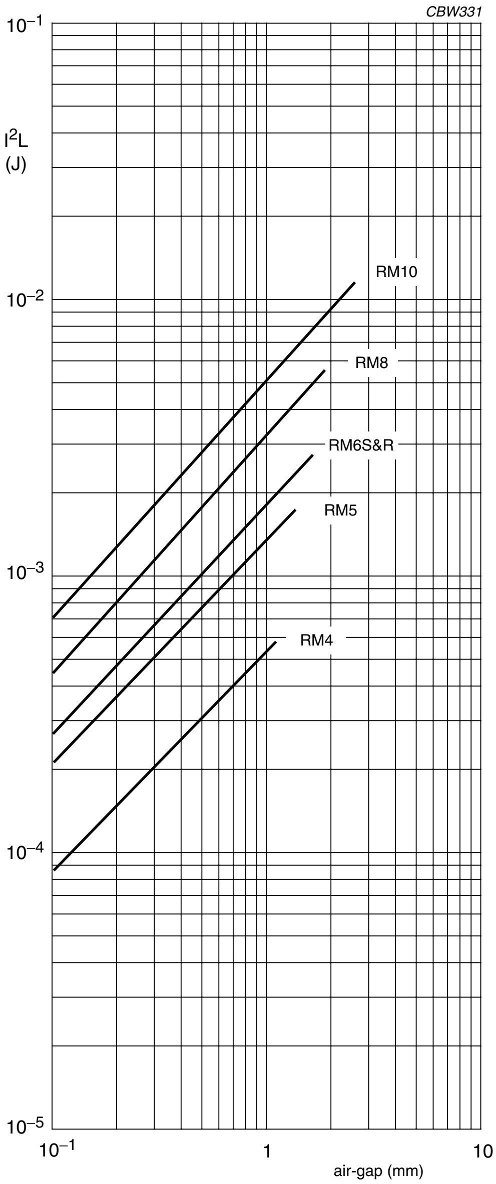

Fig.33 I^2L graph for RM cores.

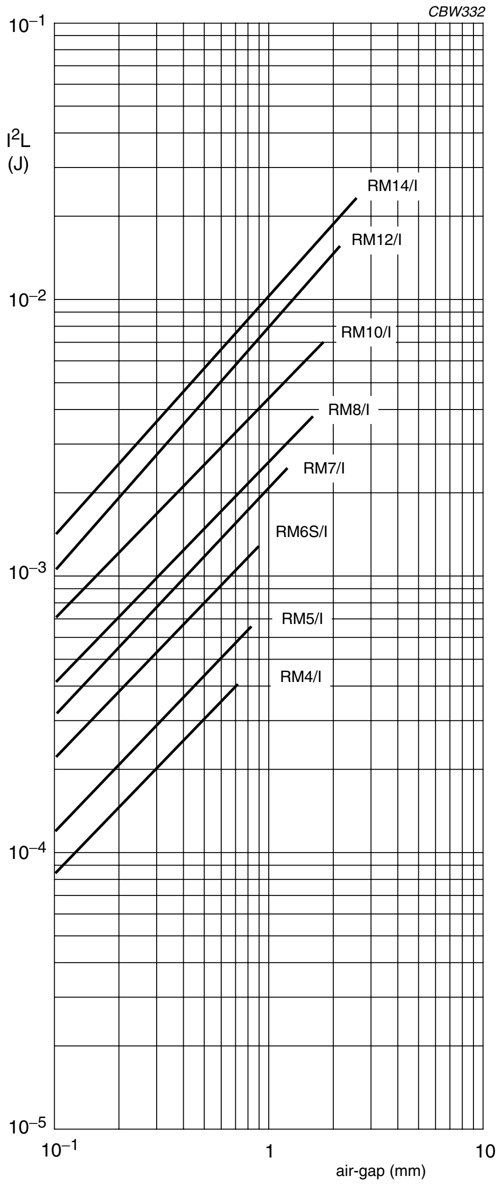

Fig.34 I^2L graph for RM/I cores.

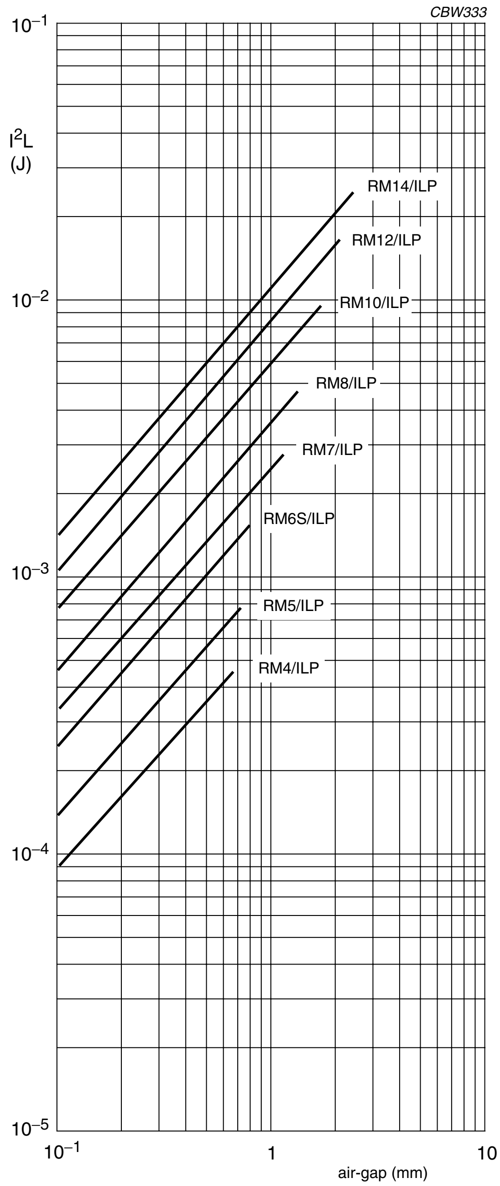

Fig.35 I^2L graph for RM/ILP cores.

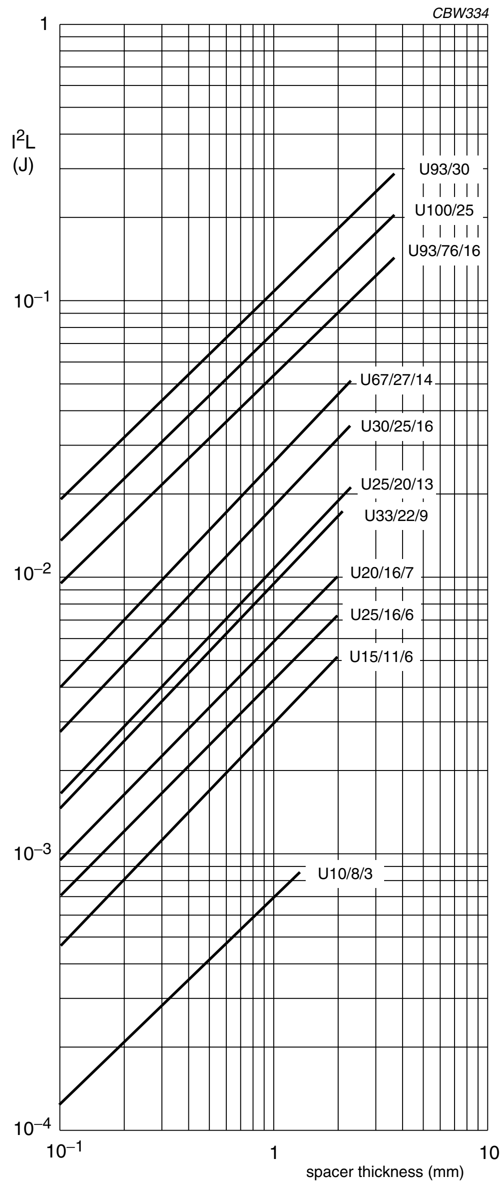

Fig.36 I^2L graph for U cores.

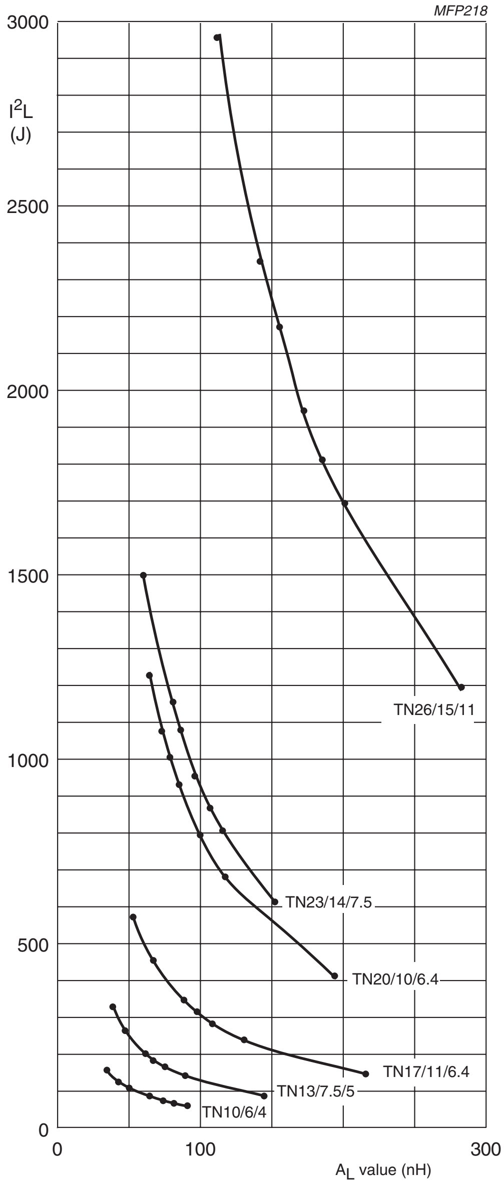

Fig.37 I^2L graph for gapped toroids.

IRON POWDER RING CORES

Ring cores made from compressed iron powder have a rather low permeability (max. 90) combined with a very high saturation level (up to 1500 mT). The permeability is so low because the isolating coating on the iron particles acts as a so called distributed air gap. Therefore, our 2P ring core range can operate under bias fields of up to 2000 A/m.

INPUT FILTERS (COMMON MODE CHOKES)

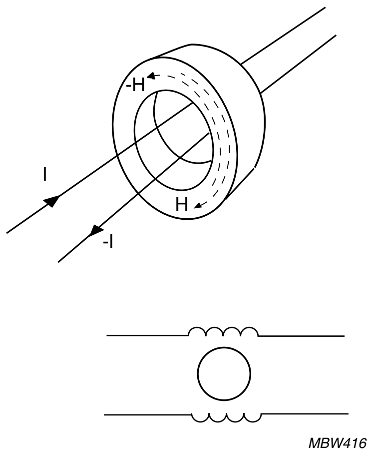

Fig.38 Common mode choke.

To avoid the conduction of switching noise from a SMPS into the mains, an input filter is generally necessary. The magnetic circuit in these filters is usually a pair of U cores or a ring core.

Since the noise signal is mainly common mode, current compensation can be used to avoid saturation.

Two separate windings on the core cause opposing magnetic fields when the load current passes through them (current compensation). The common mode noise signal however, is blocked by the full inductance caused by the high permeability ferrite.

If, for some reason, current compensation is not complete or impossible, high permeability materials will saturate. In that case one of the power materials may be a better compromise. Another important factor in the design process is the frequency range of the interference signal. High permeability ferrites have a limited bandwidth as can be seen from Fig.39.

These materials only perform well as an inductor below the frequency where ferromagnetic resonance occurs. Above this cut-off frequency, a coil will have a highly resistive character and the Q-factor of the LC filter circuit will be limited an thus, also the impedance. A better result could have been obtained with a grade having a lower permeability. Fig.40 provides a quick method of choosing the right ferrite for the job.

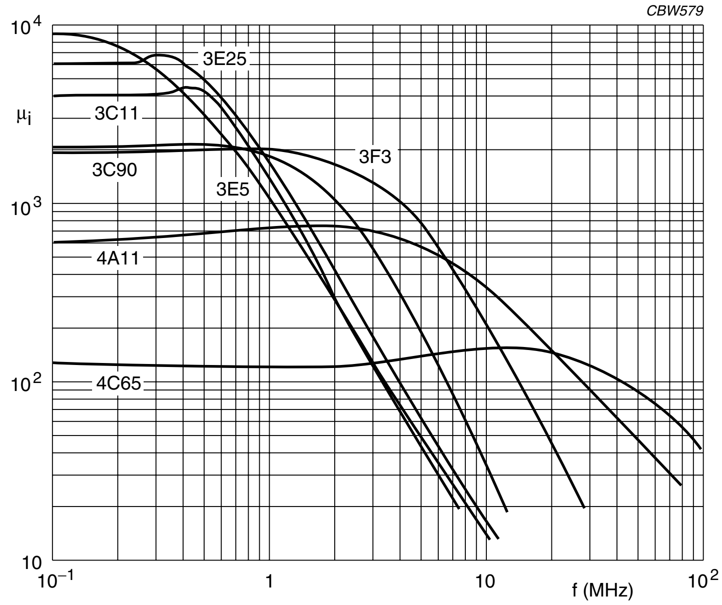

Fig.39 Permeability as a function of frequency of different materials.

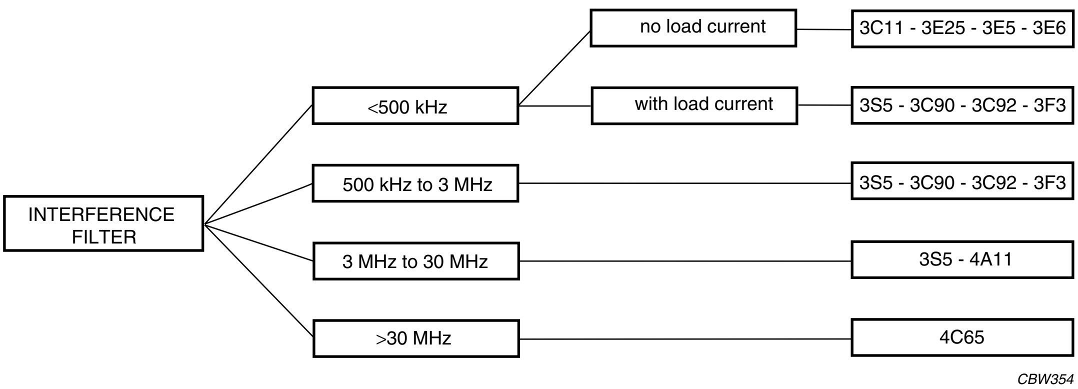

Fig.40 Selection chart for materials used in input filters.

3R1 TOROIDS IN MAGNETIC REGULATORS

Saturable inductors can be used to regulate several independent outputs of an SMPS by blocking varying amounts of energy from the secondary of the transformer. The rectangular BH loop of our 3R1 ferrite toroids makes them ideal for magnetic regulators with reset control. The circuits required are both simple and economic and can be easily integrated.

Operating principles

When the main switch is ON (t_{on}) the output current (I_{out}) flows through the winding of the saturable inductor to the output inductor and from there to the load.
During OFF time this current falls to zero and so does the magnetic field H. Because the saturable inductor has a rectangular B-H loop, the flux remains at the high level B_r even when the driving field H has fallen to zero.

When no reset current is applied, the flux in the toroid remains at the level of B_r until the next ON time starts. There is only a short delay (t_d) because the flux rises from B_r to B_s. After that, the current rises sharply to its maximum value, limited only by the load impedance. The output voltage has its maximum value, given by:

$$V_{out} = V_t \times \frac{t_{on} - t_d}{T}$$

When V_{out} is higher than V_{ref} a reset current flows during OFF time, regulated by the transistor. This current can only flow through the winding of the saturable inductor. Because this current causes a magnetic field in reverse direction it will move the ferrite away from saturation. Resetting to $-H_c$, for instance, causes some extra delay (t_b) because of the larger flux swing. Full reset causes a flux swing of almost $2 \times B_s$, resulting in a maximum delay ($t_d + t_b$) and the blocking of a major part of the energy flowing from the transformer to the load. The output voltage is regulated to the required level and is given by:

$$V_{out} = V_t \times \frac{t_{on} - t_d - t_b}{T}$$

In this way a reset current in the order of 100 mA can regulate load currents in the order of 10 A or more, depending on the layout of the saturable inductor. For this reason the described circuit is called a magnetic regulator or magnetic amplifier.

The performance of the material 3R1 is comparable to that of amorphous metal making it an excellent choice for application in magnetic regulators. However, since the value of H_c is higher for the ferrite than for most amorphous metal compositions, a simple replacement will often fail to deliver the expected results. A dedicated design or a slight redesign of the regulating circuit is then required, for which we will be glad to give you advice.

Behaviour of the ferrite material in a saturable inductor is shown in Fig.41.

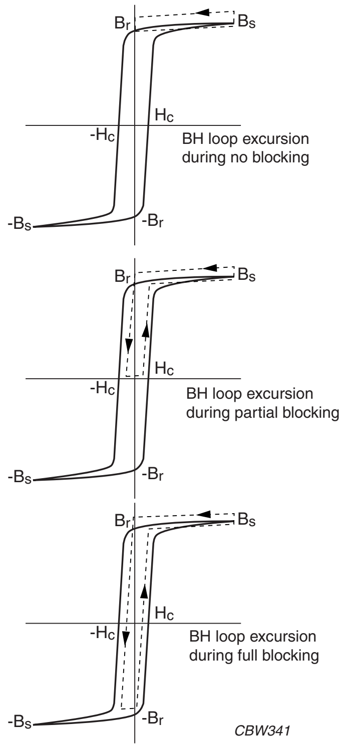

Fig.41 Behaviour of the ferrite material in a saturable inductor.

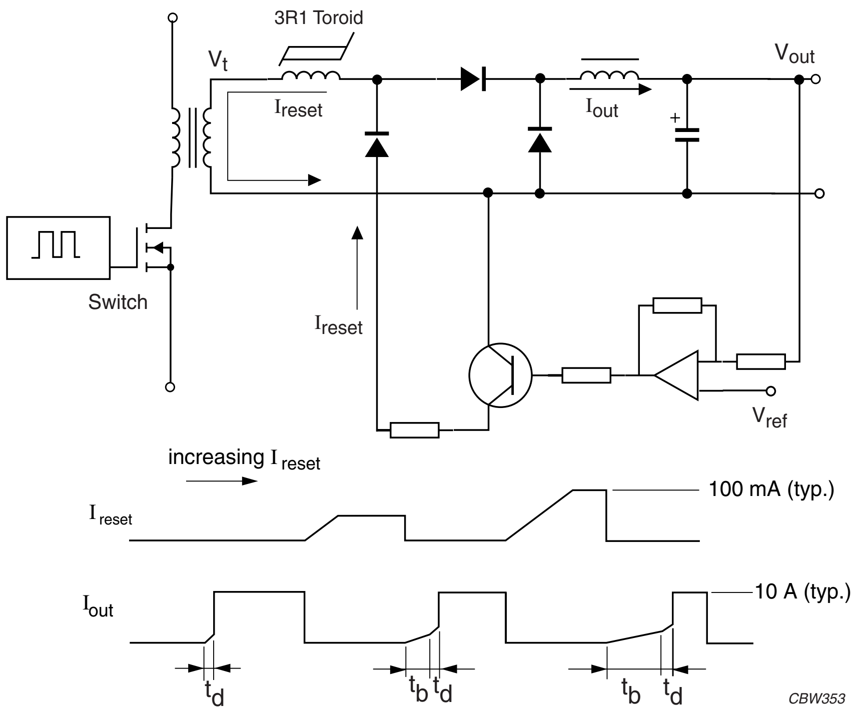

Fig.42 Schematic of a saturable inductor and associated waveforms (with regulation).

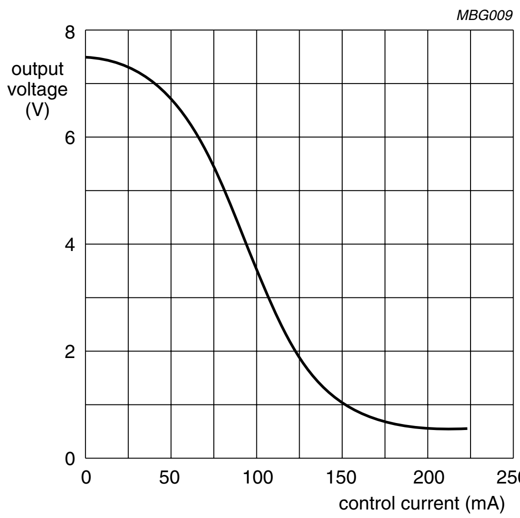

Fig.43 Typical control curve for a 3R1 ring core (size 14 × 9 × 5 mm, with 15 turns).

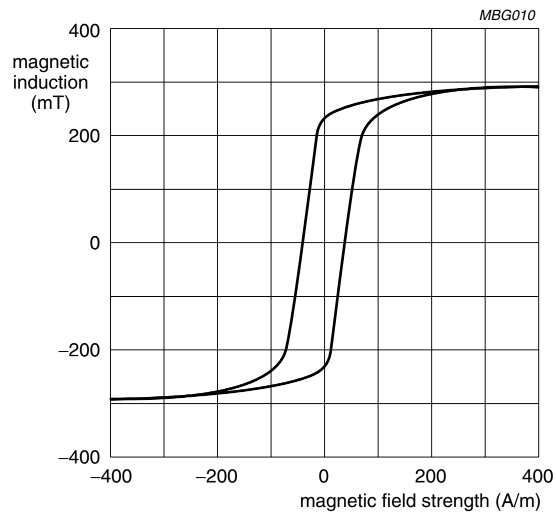

Fig.44 Properties of 3R1 ferrite material; f = 100 kHz, T = 25 °C.

Ferrites for Interference Suppression and Electromagnetic Compatibility (EMC)

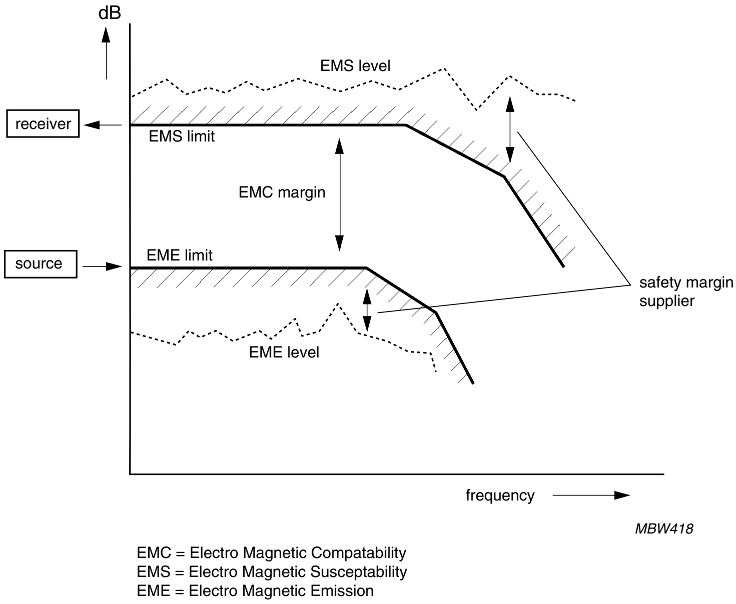

Fig.45 Principles of Electromagnetic Compatibility (EMC).

With the ever increasing intensive use of electronic equipment Electromagnetic Compatibility (EMC) has become an important item. Laws specify limits of the level of interference caused by equipment (EME) and also the sensitivity of equipment to incoming interference (EMS).

Limiting curves are defined by organizations such as EU and FCC. Since the density of equipment increases, laws will become more stringent in the near future.

During the design phase, problems with interference can be avoided to some extent. Often additional suppression components such as capacitors and coils will be necessary to meet the required levels. Inductive components are very effective in blocking interfering signals, especially at high frequencies. The principles of suppression are shown in Fig.46.

Capacitors are used as a shunt impedance for the unwanted signal.

Unfortunately for high frequencies, most capacitors do not have the low impedance one might expect because of parasitic inductance or resistance.

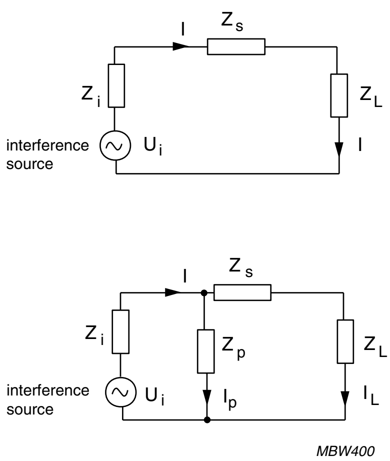

Fig.46 Basic suppression circuits.

Suppressors are used in series with the load impedance. They provide a low impedance for the wanted signal, but a high impedance for the interfering, unwanted, signal.

Ferroxcube have a full range of ring cores, beads, multilayer suppressors and inductors, beads on wire, SMD beads, wideband chokes and cable shields to suit every application. Rods and tubes are also often used for this application after they have been coiled by the user.

SAMPLE BOXES

As the design process in these areas is often based on trial and error, we have assembled several **designers' sample boxes**. Each box is filled with a selection from our standard ranges, which aims at a specific application area. The boxes also contain a booklet with full information about the products and their applications. These sample boxes are:

- Sample box 9: SMD beads and chokes
- Sample box 10: Cable shielding
- Sample box 11: EMI suppression products
- Sample box 12: Multilayer suppressors.
- Sample box 13: Multilayer inductors.

INTERFERENCE SUPPRESSION BEADS

A range of beads is available in two material grades, especially developed for suppression purposes.

They can easily be shifted on existing wires in the equipment:

- 3S1 for frequencies up to 30 MHz
- 3S4 for frequencies from 30 to 1000 MHz
- 4S2 for frequencies from 30 to 1000 MHz.

The materials and beads are fully guaranteed for their main feature, impedance as a function of frequency.

The grade 3S1 has a high permeability and is therefore rather sensitive for DC load. In applications where a high DC current is flowing 3S5 is a better choice, especially at elevated temperatures.

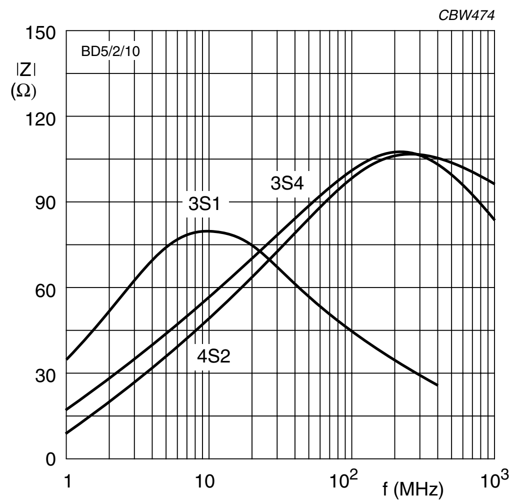

Fig.47 Impedance as a function of frequency for material grades 3S1, 3S4 and 4S2; bead size 5 × 2 × 10 mm.

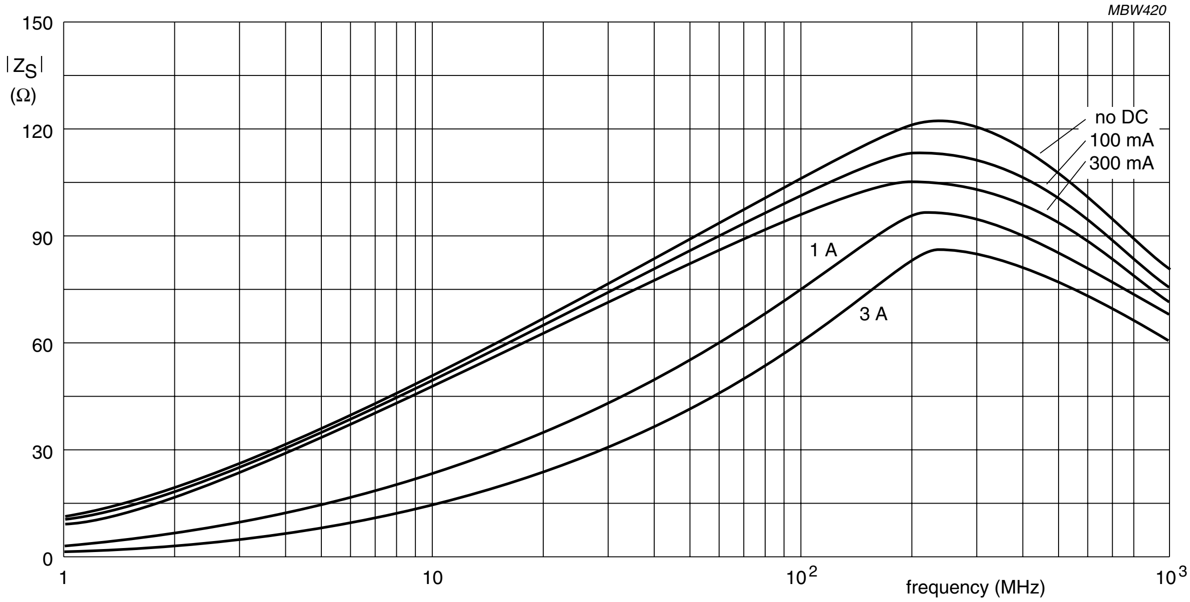

Fig.48 Impedance as a function of frequency at different DC levels for material grade 4S2; bead size 5 × 2 × 10 mm.

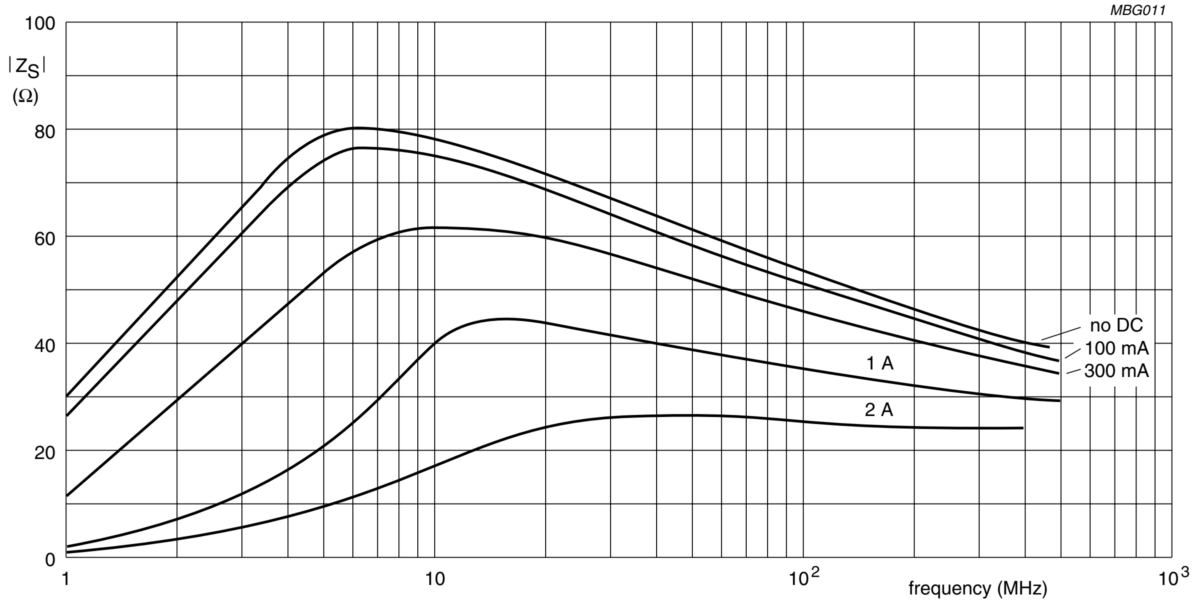

Fig.49 Impedance as a function of frequency at different DC levels for material grade 3S1; bead size 5 × 2 × 10 mm.

BEADS ON WIRE

This product range consists of suppression beads, already mounted on pre-soldered 0.6 mm wire and taped on standard reels. These can be handled by automatic placement machines.

SMD FERRITE BEADS

In response to market demands for smaller, lighter and more integrated electronic devices a series of SMD beads was added to our range. They are available in different sizes and 2 suppression ferrite grades.

Basically these beads consist of a ferrite tube with a rectangular cross-section and a flat tinned copper wire which is bent around the edges and forms the terminals of the component. This design offers many superior mechanical and electrical features.

Some examples of their impedance as a function of frequency and the influence of bias current are given in the graphs.

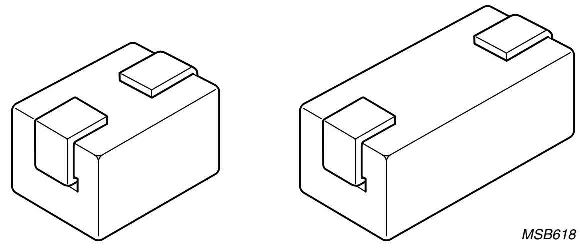

Fig.50 Outline of SMD beads.

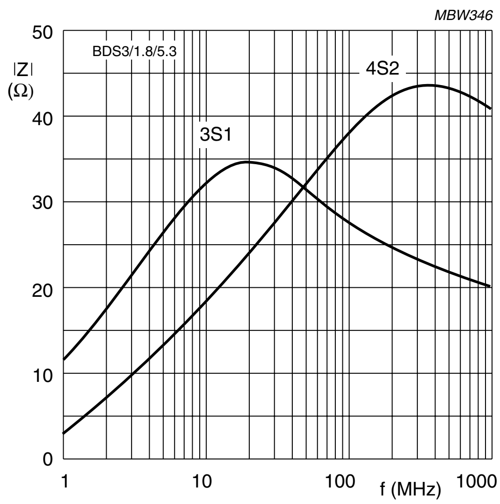

Fig.51 Impedance as a function of frequency for SMD beads.

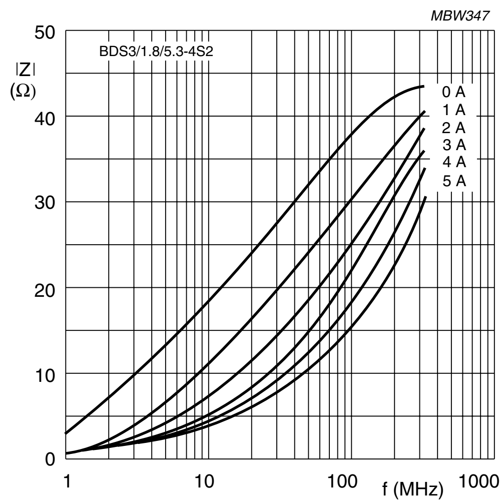

Fig.52 Impedance as a function of frequency for an SMD bead with bias current as a parameter.

SMD FERRITE BEADS FOR COMMON-MODE INTERFERENCE SUPPRESSION

Ferroxcube has a range of soft ferrite SMD beads for common-mode interference suppression.

With standard suppression methods in a signal path, the wanted signal is often suppressed along with the interference, and in many modern applications (EDP for instance) this leads to unacceptable loss of signal.

In Ferroxcube's interference suppression beads, a pair of conductors within a single soft ferrite block are connected along their lengths by an air gap.

Common-mode signals (interference signals passing in the same direction along the input and output channels of a device, an IC for instance) serve to reinforce the magnetic flux around both conductors and are therefore attenuated.

In contrast, the wanted signal passing along the input and output channels serves to cancel the flux around the conductors and therefore passes unattenuated.

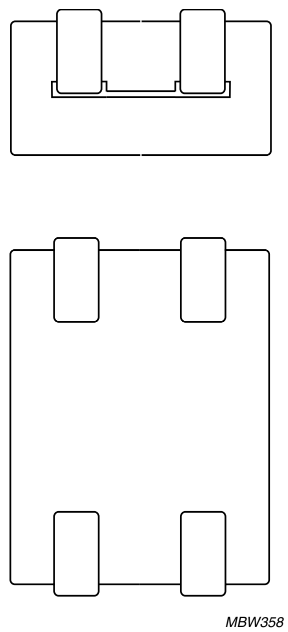

Fig.53 Outline of an SMD common-mode choke.

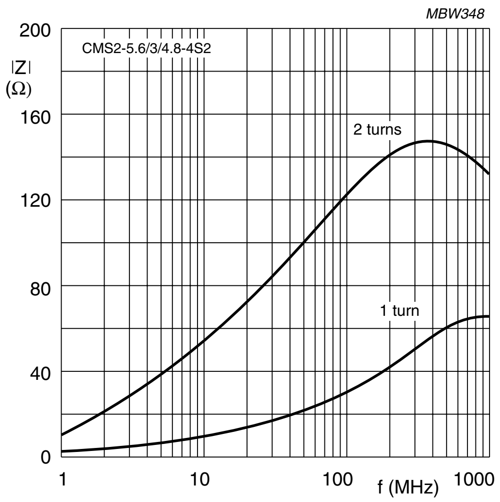

Fig.54 Impedance as a function of frequency of an SMD common mode bead with two conductors.

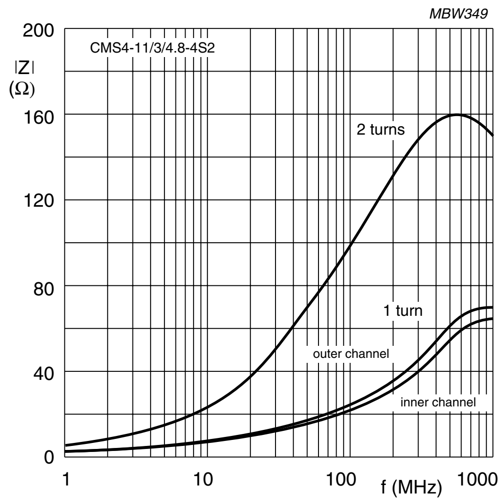

Fig.55 Impedance as a function of frequency of an SMD common mode bead with four conductors.

WIDEBAND CHOKES

Wideband chokes are wired multi-hole beads. Since they have up to 2½ turns of wire their impedance values are rather high over a broad frequency range, hence their name.

The magnetic circuit is closed so there is little stray field. The DC resistance is very low since only a short length of 0.6 mm copper wire is used.

These products already have a long service record and are still popular for various applications.

The basic range has been extended with several types, e.g. with isolation and taped on reel.

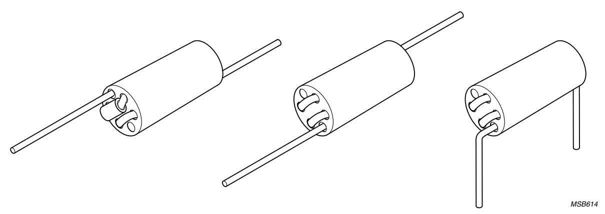

Fig.56 Outline of wideband chokes.

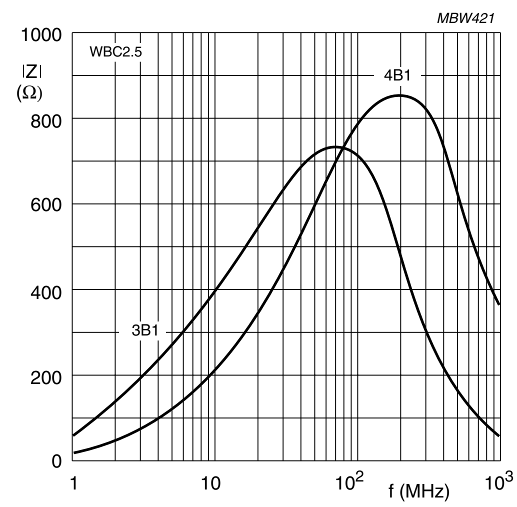

Fig.57 Impedance as a function of frequency for a wideband choke.

SMD WIDEBAND CHOKES

SMD wideband chokes are an alternative to a SMD bead when more impedance or damping is required.

The design of this product is based on our well known range of wideband chokes.

In these products the conductor wire is wound through holes in a multi-hole ferrite core, thus separating them physically and reducing coil capacitance.

The result is a high impedance over a wide frequency range, a welcome feature for many interference problems.

The present SMD design preserves the excellent properties and reliability of the original wideband chokes by keeping the number of electrical interfaces to an absolute minimum.

A plated version is available to increase the soldering surface. The metallization does not extend to the edge of the core to allow for side-to-side mounting.

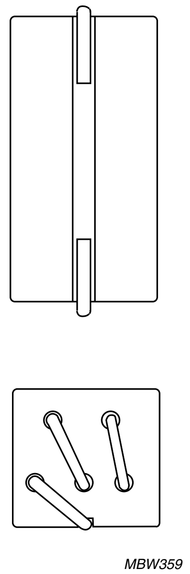

Fig.59 Outline of an SMD wideband choke.

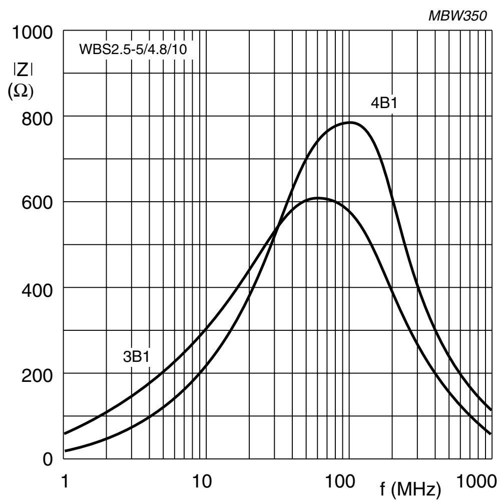

Fig.60 Impedance as a function of frequency for SMD wideband chokes.

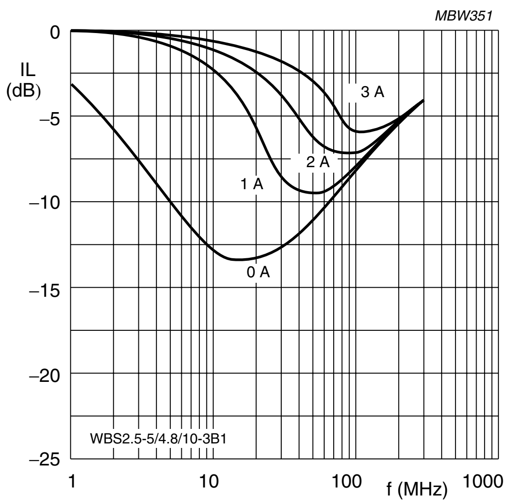

Fig.61 Insertion loss of a 3B1 SMD wideband choke as a function of frequency (50 Ω circuit).

CABLE SHIELDS

Also in our range are so-called cable shields. These products are an effective remedy against common-mode interference on coaxial or flat cables. They come in several shapes: round tubes, rectangular sleeves and split sleeves to mount on existing cable connections.

Our suppression material 3S4 is very suitable for this application. It combines a high permeability (1700) for high impedance in the lower frequency range with an excellent high frequency behaviour for true wideband suppression.

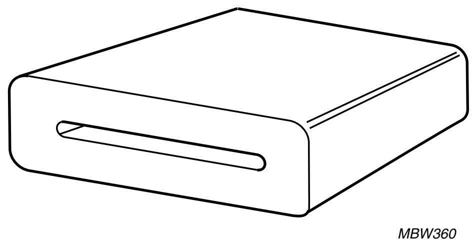

Fig.62 Outline of a cable shield.

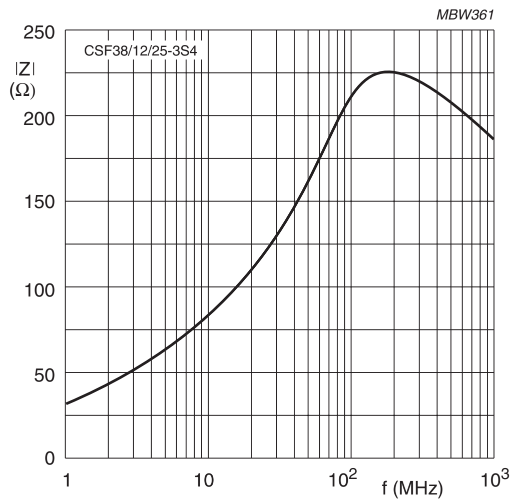

Fig.63 Impedance of a cable shield as a function of frequency.

Rods and tubes

Rods and tubes are generally used to increase the inductance of a coil. The magnetic circuit is very open and therefore the mechanical dimensions have more influence on the inductance than the ferrite's permeability (see Fig.64) unless the rod is very slender.

In order to establish the effect of a rod on the inductance of a coil, the following procedure should be carried out:

- Calculate the length to diameter ratio of the rod (l/d)
- Find this value on the horizontal axis and draw a vertical line.

The intersection of this line with the curve of the material permeability gives the effective rod permeability.

The inductance of the coil, provided the winding covers the whole length of the rod is given by:

$$L = \mu_0 \mu_{rod} \frac{N^2 A}{l} (H)$$

where:

N = number of turns

A = cross sectional area of rod

l = length of coil.

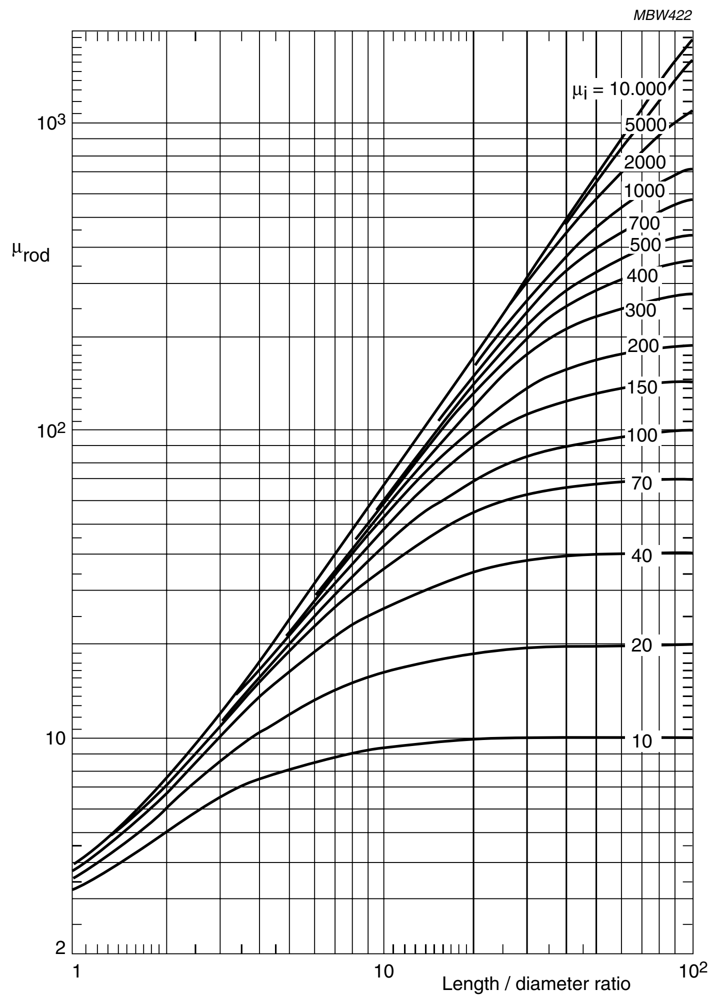

Fig.64 Rod permeability (μ_{rod}) as a function of length to diameter ratio with material permeability as a parameter.