Soft Ferrites and Accessories

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DATA SHEET STATUS DEFINITIONS

DATA SHEET STATUS	PRODUCT STATUS	DEFINITIONS
Preliminary specification	Development	This data sheet contains preliminary data. Philips Components reserves the right to make changes at any time without notice in order to improve design and supply the best possible product.
Product specification	Production	This data sheet contains final specifications. Philips Components reserves the right to make changes at any time without notice in order to improve design and supply the best possible product.

DISCLAIMER

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PRODUCT STATUS DEFINITIONS

STATUS	INDICATION	DEFINITION
Prototype	prot	These are products that have been made as development samples for the purposes of technical evaluation only. The data for these types is provisional and is subject to change.
Design-in	des	These products are recommended for new designs.
Preferred		These products are recommended for use in current designs and are available via our sales channels.
Support	sup	These products are **not** recommended for new designs and may not be available through all of our sales channels. Customers are advised to check for availability.

THE NATURE OF SOFT FERRITES

Composition

Ferrites are dark grey or black ceramic materials. They are very hard, brittle and chemically inert. Most modern magnetically soft ferrites have a cubic (spinel) structure.

The general composition of such ferrites is $MeFe_2O_4$ where Me represents one or several of the divalent transition metals such as manganese (Mn), zinc (Zn), nickel (Ni), cobalt (Co), copper (Cu), iron (Fe) or magnesium (Mg).

The most popular combinations are manganese and zinc (MnZn) or nickel and zinc (NiZn). These compounds exhibit good magnetic properties below a certain temperature, called the Curie Temperature (T_C). They can easily be magnetized and have a rather high intrinsic resistivity. These materials can be used up to very high frequencies without laminating, as is the normal requirement for magnetic metals.

NiZn ferrites have a very high resistivity and are most suitable for frequencies over 1 MHz, however, MnZn ferrites exhibit higher permeability (μ_i) and saturation induction levels (B_s) and are suitable up to 3 MHz.

For certain special applications, single crystal ferrites can be produced, but the majority of ferrites are manufactured as polycrystalline ceramics.

Manufacturing process

The following description of the production process is typical for the manufacture of our range of soft ferrites, which is marketed under the trade name ‘Ferroxcube’.

RAW MATERIALS

The raw materials used are oxides or carbonates of the constituent metals. The final material grade determines the necessary purity of the raw materials used, which, as a result is reflected in the overall cost.

PROPORTIONS OF THE COMPOSITION

The base materials are weighed into the correct proportions required for the final composition.

MIXING

The powders are mixed to obtain a uniform distribution of the components.

PRE-SINTERING

The mixed oxides are calcined at approximately 1000 °C. A solid state reaction takes place between the constituents and, at this stage, a ferrite is already formed.

Pre-sintering is not essential but provides a number of advantages during the remainder of the production process.

MILLING AND GRANULATION

The pre-sintered material is milled to a specific particle size, usually in a slurry with water. A small proportion of organic binder is added, and then the slurry is spray-dried to form granules suitable for the forming process.

FORMING

Most ferrite parts are formed by pressing. The granules are poured into a suitable die and then compressed. The organic binder acts in a similar way to an adhesive and a so-called ‘green’ product is formed. It is still very fragile and requires sintering to obtain the final ferrite properties.

For some products, for example, long rods or tubes, the material is mixed into a dough and extruded through a suitable orifice. The final products are cut to the required length before or after sintering.

SINTERING

The ‘green’ cores are loaded on refractory plates and sintered at a temperature between 1150 °C and 1300 °C depending on the ferrite grade. A linear shrinkage of up to 20% (50% in volume) takes place. The sintering may take place in tunnel kilns having a fixed temperature and atmosphere distribution or in box kilns where temperature and atmosphere are computer controlled as a function of time. The latter type is more suitable for high grade ferrites which require a very stringent control in conditions.

FINISHING

After sintering, the ferrite core has the required magnetic properties. It can easily be magnetized by an external field (see Fig.2), exhibiting the well-known hysteresis effect (see Fig.1). Dimensions are typically within 2% of nominal due to 10- 20% shrinkage. If this tolerance is too large or if some surfaces require a smooth finish (e.g. mating faces between core halves) a grinding operation is necessary. Usually diamond-coated wheels are used. For high permeability materials, very smooth, lapped, mating surfaces are required. If an air-gap is required in the application, it may be provided by centre pole grinding.

Magnetism in ferrites

A sintered ferrite consists of small crystals, typically 10 to 20 μm in dimension. Domains exist within these crystals (Weiss domains) in which the molecular magnets are already aligned (ferrimagnetism). When a driving magnetic field (H) is applied to the material the domains progressively align with it, as shown in Fig.2.

During this magnetization process energy barriers have to be overcome. Therefore the magnetization will always lag behind the field. A so-called hysteresis loop (see Fig.1) is the result.

If the resistance against magnetization is small, a large induced flux will result at a given magnetic field. The value of the permeability is high. The shape of the hysteresis loop also has a marked influence on other properties, for example power losses.

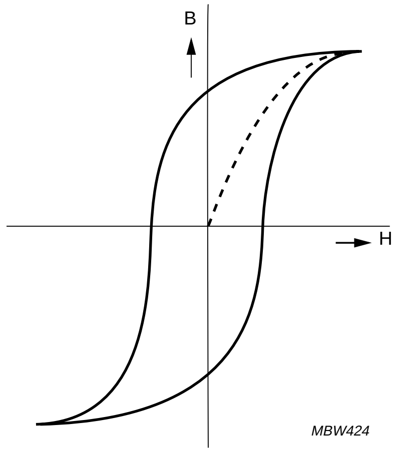


Fig.1 Hysteresis loop.

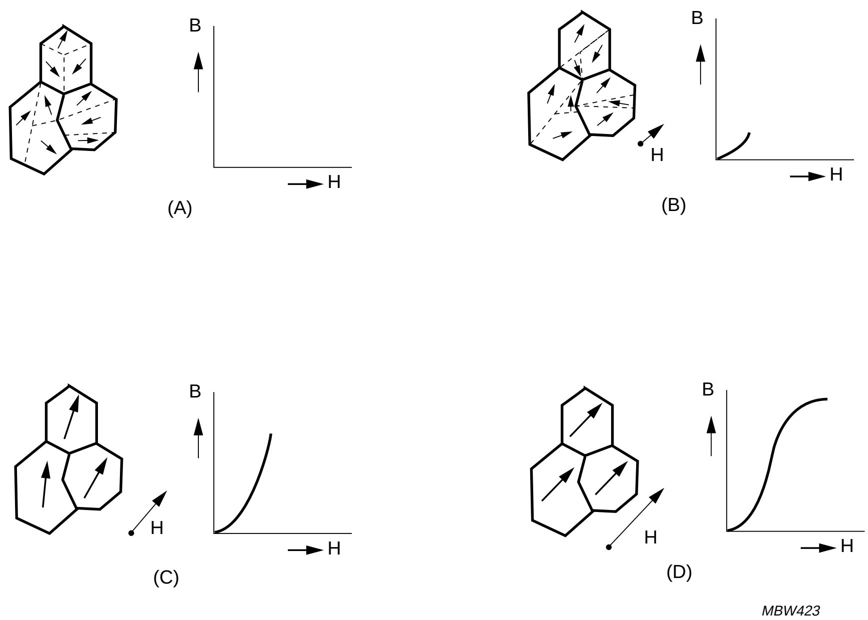


Fig.2 Alignment of domains.

EXPLANATION OF TERMS AND FORMULAE

Symbols and units

SYMBOL	DESCRIPTION	UNIT
A_e	effective cross-sectional area of a core	mm^2
A_{min}	minimum cross-sectional area of a core	mm^2
A_L	inductance factor	nH
B	magnetic flux density	T
B_r	remanence	T
B_s	saturation flux density	T
$\hat{B}$	peak flux density	T
C	capacitance	F
D_F	disaccomodation factor	–
f	frequency	Hz
G	gap length	μm
H	magnetic field strength	A/m
H_c	coercivity	A/m
$\hat{H}$	peak magnetic field strength	A/m
I	current	A
l_e	effective magnetic path length	mm
L	inductance	H
N	number of turns	–
P_v	specific power loss of core material	kW/m^3
Q	quality factor	–
T_c	Curie temperature	°C
V_e	effective volume of core	mm^3
α_F	temperature factor of permeability	K^{-1}
$\frac{\tan\delta}{\mu_i}$	loss factor	–
η_B	hysteresis material constant	T^{-1}
μ	absolute permeability	–
μ_o	magnetic constant ($4\pi \times 10^{-7}$)	Hm^{-1}
μ_s'	real component of complex series permeability	–
μ_s''	imaginary component of complex series permeability	–
μ_a	amplitude permeability	–
μ_e	effective permeability	–
μ_i	initial permeability	–
μ_r	relative permeability	–
μ_Δ	incremental permeability	–
ρ	resistivity	Ωm
$\Sigma(l/A)$	core factor (C1)	mm^{-1}

Definition of terms

PERMEABILITY

When a magnetic field is applied to a soft magnetic material, the resulting flux density is composed of that of free space plus the contribution of the aligned domains.

$$B = \mu_0 H + J \quad \text{or} \quad B = \mu_0 (H + M) \qquad (1)$$

where $\mu_0 = 4\pi.10^{-7}$ H/m, J is the magnetic polarization and M is the magnetization.

The ratio of flux density and applied field is called absolute permeability.

$$\frac{B}{H} = \mu_0\left(1 + \frac{M}{H}\right) = \mu_{absolute} \qquad (2)$$

It is usual to express this absolute permeability as the product of the magnetic constant of free space and the relative permeability (μ_r).

$$\frac{B}{H} = \mu_0 \mu_r \qquad (3)$$

Since there are several versions of μ_r depending on conditions the index ‘r’ is generally removed and replaced by the applicable symbol e.g. μ_i, μ_a, μ_Δ etc.

INITIAL PERMEABILITY

The initial permeability is measured in a closed magnetic circuit (ring core) using a very low field strength.

$$\mu_i = \frac{1}{\mu_0} \times \frac{\Delta B}{\Delta H}_{(\Delta H \to 0)} \qquad (4)$$

Initial permeability is dependent on temperature and frequency.

EFFECTIVE PERMEABILITY

If the air-gap is introduced in a closed magnetic circuit, magnetic polarization becomes more difficult. As a result, the flux density for a given magnetic field strength is lower.

Effective permeability is dependent on the initial permeability of the soft magnetic material and the dimensions of air-gap and circuit.

$$\mu_e = \frac{\mu_i}{1 + \frac{G \times \mu_i}{l_e}} \qquad (5)$$

where G is the gap length and l_e is the effective length of magnetic circuit. This simple formula is a good approximation only for small air-gaps. For longer air-gaps some flux will cross the gap outside its normal area (stray flux) causing an increase of the effective permeability.

AMPLITUDE PERMEABILITY

The relationship between higher field strength and flux densities without the presence of a bias field, is given by the amplitude permeability.

$$\mu_a = \frac{1}{\mu_0} \times \frac{\hat{B}}{\hat{H}} \qquad (6)$$

Since the BH loop is far from linear, values depend on the applied field peak strength.

INCREMENTAL PERMEABILITY

The permeability observed when an alternating magnetic field is superimposed on a static bias field, is called the incremental permeability.

$$\mu_\Delta = \frac{1}{\mu_0}\left[\frac{\Delta B}{\Delta H}\right]_{H_{DC}} \qquad (7)$$

If the amplitude of the alternating field is negligibly small, the permeability is then called the reversible permeability (μ_{rev}).

COMPLEX PERMEABILITY

A coil consisting of windings on a soft magnetic core will never be an ideal inductance with a phase angle of 90°. There will always be losses of some kind, causing a phase shift, which can be represented by a series or parallel resistance as shown in Figs 3 and 4.

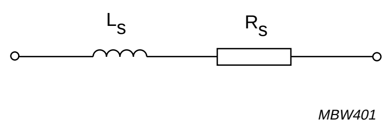


Fig.3 Series representation.

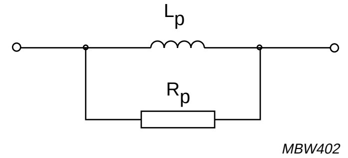


Fig.4 Parallel representation.

For series representation

$$\bar{Z} = j\omega L_s + R_s \quad (8)$$

and for parallel representation,

$$\bar{Z} = \frac{1}{1/(j\omega L_p) + 1/R_p} \quad (9)$$

the magnetic losses are accounted for if a resistive term is added to the permeability.

$$\mu = \mu_s' - j\mu_s'' \quad \text{or} \quad \frac{1}{\bar{\mu}} = \frac{1}{\mu'_p} - \frac{1}{\mu''_p} \quad (10)$$

The phase shift caused by magnetic losses is given by:

$$\tan\delta_m = \frac{R_s}{\omega L_s} = \frac{\mu''_s}{\mu'_s} \quad \text{or} \quad \frac{\omega L_p}{R_p} = \frac{\mu'_p}{\mu''_p} \quad (11)$$

For calculations on inductors and also to characterize ferrites, the series representations is generally used (μ'_s and μ''_s). In some applications e.g. signal transformers, the use of the parallel representation (μ'_p and μ''_p) is more convenient.

The relationship between the representations is given by:

$$\mu'_p = \mu'_s(1 + \tan\delta^2) \quad \text{and} \quad \mu''_p = \mu''_s\left(1 + \frac{1}{\tan\delta^2}\right) \quad (12)$$

LOSS FACTOR

The magnetic losses which cause the phase shift can be split up into three components:

1. Hysteresis losses
2. Eddy current losses
3. Residual losses.

This gives the formula:

$$\tan\delta_m = \tan\delta_h + \tan\delta_f + \tan\delta_r \quad (13)$$

Figure 5 shows the magnetic losses as a function of frequency.

Hysteresis losses vanish at very low field strengths. Eddy current losses increase with frequency and are negligible at very low frequency. The remaining part is called residual loss. It can be proven that for a gapped magnetic circuit, the following relationship is valid:

$$\frac{(\tan\delta_m)_{gapped}}{\mu_e - 1} = \frac{\tan\delta_m}{\mu_i - 1} \quad (14)$$

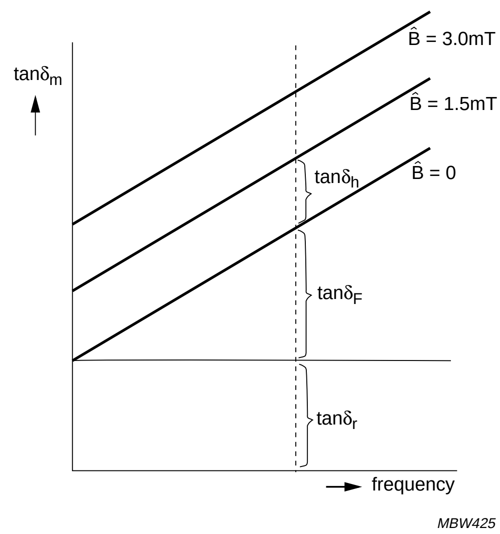


Fig.5 Magnetic losses $(\tan\delta_m)$ as a function of frequency.

Since μ_i and μ_e are usually much greater than 1, a good approximation is:

$$\frac{(\tan\delta_m)_{gapped}}{\mu_e} = \frac{\tan\delta_m}{\mu_i} \quad (15)$$

From this formula, the magnetic losses in a gapped circuit can be derived from:

$$(\tan\delta_m)_{gapped} = \frac{\tan\delta_m}{\mu_i} \times \mu_e \quad (16)$$

Normally, the index ‘m’ is dropped when material properties are discussed:

$$(\tan\delta)_{gapped} = \frac{\tan\delta}{\mu_i} \times \mu_e \quad (17)$$

In material specifications, the loss factor ($\tan\delta/\mu_i$) is used to describe the magnetic losses. These include residual and eddy current losses, but not hysteresis losses.

For inductors used in filter applications, the quality factor (Q) is often used as a measure of performance. It is defined as:

$$Q = \frac{1}{\tan\delta} = \frac{\omega L}{R_{tot}} = \frac{\text{reactance}}{\text{total resistance}} \quad (18)$$

The total resistance includes the effective resistance of the winding at the design frequency.

HYSTERESIS MATERIAL CONSTANT

When the flux density of a core is increased, hysteresis losses are more noticeable. Their contribution to the total losses can be obtained by means of two measurements, usually at the induction levels of 1.5 mT and 3 mT. The hysteresis constant is found from:

$$\eta_B = \frac{\Delta \tan\delta_m}{\mu_e \times \Delta\hat{B}} \qquad (19)$$

The hysteresis loss factor for a certain flux density can be calculated using:

$$\frac{\tan\delta_h}{\mu_e} = \eta_B \times \hat{B} \qquad (20)$$

This formula is also the IEC definition for the hysteresis constant.

EFFECTIVE CORE DIMENSIONS

To facilitate calculations on a non-uniform soft magnetic cores, a set of effective dimensions is given on each data sheet. These dimensions, effective area (A_e), effective length (l_e) and effective volume (V_e) define a hypothetical ring core which would have the same magnetic properties as the non-uniform core.

The reluctance of the ideal ring core would be:

$$\frac{l_e}{\mu \times A_e} \qquad (21)$$

For the non-uniform core shapes, this is usually written as:

$$\frac{1}{\mu_e} \times \Sigma\frac{l}{A} \qquad (22)$$

the core factor divided by the permeability. The inductance of the core can now be calculated using this core factor:

$$L = \frac{\mu_0 \times N^2}{\frac{1}{\mu_e} \times \Sigma\frac{l}{A}} = \frac{1.257 \times 10^{-9} \times N^2}{\frac{1}{\mu_e} \times \Sigma\frac{l}{A}} \text{(in H)} \qquad (23)$$

The effective area is used to calculate the flux density in a core,

for sine wave:

$$\hat{B} = \frac{U\sqrt{2} \times 10^9}{\omega A_e N} = \frac{2.25U \times 10^8}{fNA_e} \text{(in mT)} \qquad (24)$$

for square wave:

$$\hat{B} = \frac{0.25\hat{U} \times 10^9}{fNA_e} \text{(in mT)} \qquad (25)$$

where:

A_e is the effective area in mm^2.

U is the voltage in V

f is the frequency in Hz

N is the number of turns.

The magnetic field strength (H) is calculated using the effective length (l_e):

$$\hat{H} = \frac{IN\sqrt{2}}{l_e} \text{(A/m)} \qquad (26)$$

If the cross-sectional area of a core is non-uniform, there will always be a point where the real cross-section is minimal. This value is known as A_{min} and is used to calculate the maximum flux density in a core. A well designed ferrite core avoids a large difference between A_e and A_{min}. Narrow parts of the core could saturate or cause much higher hysteresis losses.

INDUCTANCE FACTOR (A_L)

To make the calculation of the inductance of a coil easier, the inductance factor, known as the A_L value, is given in each data sheet (in nano Henry). The inductance of the core is defined as:

$$L = N^2 \times A_L \text{ (nH)} \qquad (27)$$

The value is calculated using the core factor and the effective permeability:

$$A_L = \frac{\mu_0 \mu_e \times 10^6}{\Sigma(l/A)} = \frac{1.257\mu_e}{\Sigma(l/A)} \text{ (nH)} \qquad (28)$$

MAGNETIZATION CURVES (H_C, B_R, B_S)

If an alternating field is applied to a soft magnetic material, a hysteresis loop is obtained. For very high field strengths, the maximum attainable flux density is reached. This is known as the saturation flux density (B_s).

If the field is removed, the material returns to a state where, depending on the material grade, a certain flux density remains. This the remanent flux density (B_r).

This remanent flux returns to zero for a certain negative field strength which is referred to a coercivity (H_c).

These points are clearly shown in Fig.6.

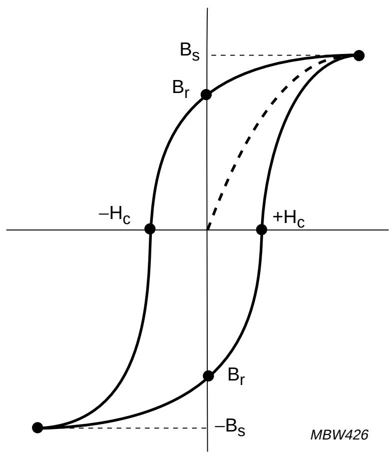


Fig.6 Typical BH curve showing points B_S, B_r and H_c.

TEMPERATURE DEPENDENCE OF THE PERMEABILITY

The permeability of a ferrite is a function of temperature. It generally increases with temperature to a maximum value and then drops sharply to a value of 1. The temperature at which this happens is called the Curie temperature (T_C). Typical curves of our grades are given in the material data section.

For filter applications, the temperature dependence of the permeability is a very important parameter. A filter coil should be designed in such a way that the combination it forms with a high quality capacitor results in an LC filter with excellent temperature stability.

The temperature coefficient (TC) of the permeability is given by:

$$TC = \frac{(\mu_i)_{T2} - (\mu_i)_{T1}}{(\mu_i)_{T1}} \times \frac{1}{T_2 - T_1} \quad (29)$$

For a gapped magnetic circuit, the influence of the permeability temperature dependence is reduced by the factor μ_e/μ_i. Hence:

$$TC_{gap} = \frac{\mu_e}{(\mu_i)_{T1}} \times \frac{(\mu_i)_{T2} - (\mu_i)_{T1}}{(\mu_i)^2_{T1}} \times \frac{1}{T_2 - T_1} = \mu_e \times \alpha_F \quad (30)$$

So α_F is defined as:

$$\alpha_F = \frac{(\mu_i)_{T2} - (\mu_i)_{T1}}{(\mu_i)^2_{T1}} \times \frac{1}{T_2 - T_1} \quad (31)$$

Or, to be more precise, if the change in permeability over the specified area is rather large:

$$\alpha_F = \frac{(\mu_i)_{T2} - (\mu_i)_{T1}}{(\mu_i)_{T1} \times (\mu_i)_{T2}} \times \frac{1}{T_2 - T_1} \quad (32)$$

The temperature factors for several temperature trajectories of the grades intended for filter applications are given in the material specifications. They offer a simple means to calculate the temperature coefficient of any coil made with these ferrites.

TIME STABILITY

When a soft magnetic material is given a magnetic or thermal disturbance, the permeability rises suddenly and then decreases slowly with time. For a defined time interval, this ‘disaccommodation’ can be expressed as:

$$D = \frac{\mu_1 - \mu_2}{\mu_1} \quad (33)$$

The decrease of permeability appears to be almost proportional to the logarithm of time. For this reason, IEC has defined a disaccommodation coefficient:

$$d = \frac{\mu_1 - \mu_2}{\mu_1 \times \log(t_2/t_1)} \quad (34)$$

As with temperature dependence, the influence of disaccommodation on the inductance drift of a coil will be reduced by μ_e/μ_i.

Therefore, a disaccommodation factor D_F is defined:

$$D_F = \frac{d}{\mu_i} = \frac{\mu_1 - \mu_2}{\mu_1^2 \times \log(t_2/t_1)} \quad (35)$$

The variability with time of a coil can now be predicted by:

$$\frac{L_1 - L_2}{L_1} = \mu_e \times D_F \quad (36)$$

RESISTIVITY

Ferrite is a semiconductor with a DC resistivity in the crystallites of the order of 10^{-3} Ωm for a MnZn type ferrite, and about 30 Ωm for a NiZn ferrite.

Since there is an isolating layer between the crystals, the bulk resistivity is much higher: 0.1 to 10 Ωm for MnZn ferrites and 10^4 to 10^6 Ωm for NiZn and MgZn ferrites.

This resistivity depends on temperature and measuring frequency, which is clearly demonstrated in Tables 1 and 2 which show resistivity as a function of temperature for different materials.

Table 1 Resistivity as a function of temperature of a MnZn-ferrite (3C80)

TEMPERATURE (°C)	RESISTIVITY (Ωm)
–20	≈10
0	≈7
20	≈4
50	≈2
100	≈1

Table 2 Resistivity as a function of temperature of a NiZn-ferrite (4C6)

TEMPERATURE (°C)	RESISTIVITY (Ωm)
0	≈5.10^7
20	≈10^7
60	≈10^6
100	≈10^5

At higher frequencies the crystal boundaries are more or less short-circuited by heir capacitance and the measured resistivity decreases, as shown in Tables 3 and 4.

Table 3 Resistivity as function of frequency for MnZn ferrites

FREQUENCY (MHz)	RESISTIVITY (Ωm)
0.1	≈2
1	≈0.5
10	≈0.1
100	≈0.01

Table 4 Resistivity as function of frequency for NiZn ferrites

FREQUENCY (MHz)	RESISTIVITY (Ωm)
0.1	≈10^5
1	≈5.10^4
10	≈10^4
100	≈10^3

PERMITTIVITY

The basic permittivity of all ferrites is of the order of 10. This is valid for MnZn and NiZn materials. The isolating material on the grain boundaries also has a permittivity of approximately 10. However, if the bulk permittivity of a ferrite is measured, very different values of apparent permittivity result. This is caused by the conductivity inside the crystallites. The complicated network of more or less leaky capacitors also shows a strong frequency dependence.

Tables 5 and 6 show the relationship between permittivity and frequency for both MnZn and NiZn ferrites.

Table 5 Permittivity as a function of frequency for MnZn ferrites

FREQUENCY (MHz)	PERMITTIVITY (ε_r)
0.1	≈2.10^5
1	≈10^5
10	≈5.10^4
100	≈10^4

Table 6 Permittivity as a function of frequency for NiZn ferrites

FREQUENCY (MHz)	PERMITTIVITY (ε_r)
0.001	≈100
0.01	≈50
1	≈25
10	≈15
100	≈12

QUALITY

Quality standards

Our ferrite cores are produced to meet constantly high quality standards. High quality components in mass production require advanced production techniques as well as background knowledge of the product itself. The quality standard is achieved in our ferrite production centres by implementation of a Quality Assurance System based on ISO9001 and our process control is based on SPC techniques.

To implement SPC, the production is divided in stages which correspond to production steps or groups of steps. The output of each stage is statistically checked in accordance with MIL STD 414 and 105D.

The obtained results are measured against built-in control, warning and rejects levels. If an unfavourable trend is observed in the results from a production stage, corrective and preventive actions are immediately taken. Quality is no longer “inspected-in” but “built-in” by continuous improvement.

The system is applicable for the total manufacturing process including,

- Raw material
- Production of process
- Finished products.

All our production centres are complying with the ISO 9000 quality system.

Aspects of quality

When describing the quality of a product, three aspects must be taken into account:

- Delivery quality
- Fitness for use
- Reliability.

DELIVERY QUALITY

After production, the ferrite components are tested once again for their main characteristics. Tests are conducted in accordance with the guidelines specified by IEC 60367. A sampling system, in accordance with IEC 60410 is used, and the Acceptable Quality levels (AQL's) are set for different classes of defects, major defects having lower AQL's than minor defects.

Customers may follow the same system to carry out incoming inspections. If the percentage of defects does not exceed the specified level, the probability that the batch will be accepted is high (>90%), but rejection is still possible.

If the reject level is much lower than specified, quality complaints will disappear. We aim at very low reject levels to eventually allow any customers to dispose with incoming inspection.

FITNESS FOR USE

This is a measure of component quality up to the point where the component has been assembled into the equipment and is quoted in parts per million (PPM). After assembly, the component should function fully. The PPM concept covers the possibility of failures that occur during assembly. It includes line rejects that may occur for any reason.

For ferrite cores, co-operation between the component supplier and the customer is a very important aspect. The core is generally a building block for a wound component and many things can go wrong during the assembly process, but the core is not always the problem. A mutual quality control programme can be established to minimize line rejects for a specific application. For some product lines, levels of 30 PPM have already been realized.

RELIABILITY

Ferrite cores are known for their reliability. Once the assembly process has been successfully concluded, no real threats for the life of the ferrite are known.

Reliability is mainly governed by the quality of the total assembly of the wound component. Extreme thermal shocks should be avoided. Some data are available for RM cores assembled with the recommended Philips bobbins and clips.

Vibration test, IEC 60068-2-6 (test Fc)

- No failures
- Less than 0.1% drift of inductance value.

Bump test, IEC 60068-2-29 (test Eb)

- No failures
- Less than 0.03% drift of inductance value.

Classification defects

If a component does not comply with the specification published in this handbook, it is considered to be defective. Defects are divided into two classes:

- Major defects
  These defects lead to malfunction of the finished wound components.
- Minor defects
  These defects do not have a severe influence on the function of the wound component. Often, they have a negative effect on the visual appearance of the end product, or they slightly disturb the assembly process.

Classification of defects per product line

CORE TYPE	CLASSIFICATION OF FAILURES	
	MAJOR	MINOR
RM; P; X; EP; H; PH; RM/I; P/I; PQ; PT; PTS	A_L; critical dimensions	power loss; secondary dimensions
E; planar E; EFD; ETD/ER; EC; U; I	A_L; critical dimensions	power loss; secondary dimensions
ring cores rods tubes beads wideband chokes bobbin cores cup and mushroom cores	A_L; critical dimensions; Z_{min}	A_{Lmax}; power loss; dielectrical strength of coating; secondary dimensions

Classification of defects per product line

Tighter AQL levels can be agreed upon for customized products. Also ppm agreements with customers are encouraged.

CORE TYPE	APPLICATION AREA	CLASSIFICATION OF FAULT				
		FAULT TYPE	MAJOR		MINOR	
			AQL	LEVEL	AQL	LEVEL
P; RM; X	filters	electrical	1%	(I)	2.5%	(S3)
		mechanical	0.65%	(I)	4%	(S3)
P; RM; X; EP; H	general purpose transformers	electrical	1.5%	(S4)	4%	(S3)
		mechanical	0.65%	(I)	4%	(S3)
E; EFD; ETD/ER; EC; U;I	power transformers	electrical	1%	(I)	4%	(S3)
		mechanical	1%	(I)	4%	(S3)
ring cores rods, tubes beads chokes	EMI-suppression	electrical	0.25%	(II)	2.5%	(II)
		mechanical	0.25%	(II)	2.5%	(II)

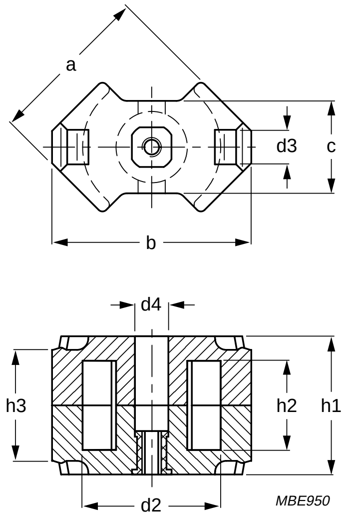


Fig.1 RM core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
RM	1	$h_2$min	a
		h_3	b
		$d_2$min	c
		$d_3$max	h_1
		d4	$h_2$max
			$d_2$max
			$d_3$min

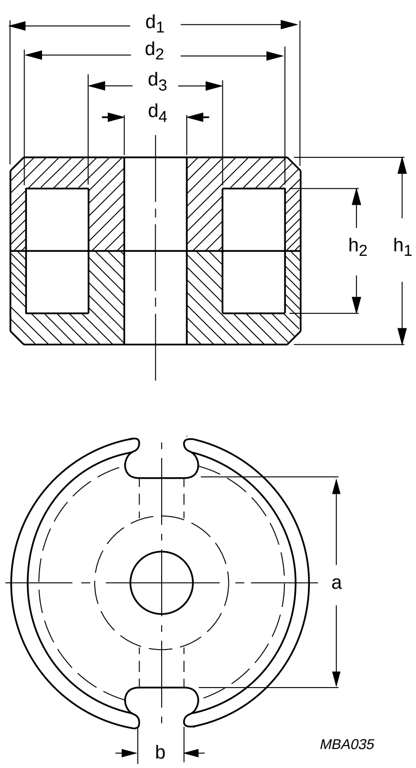


Fig.2 P core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
P; P/I; PT; PTS;	2	$h_2$min	a
		$d_2$min	b
		$d_3$max	h_1
		$d_1$max	$h_2$max
		d_4	$d_2$max
			$d_3$min
			$d_1$min

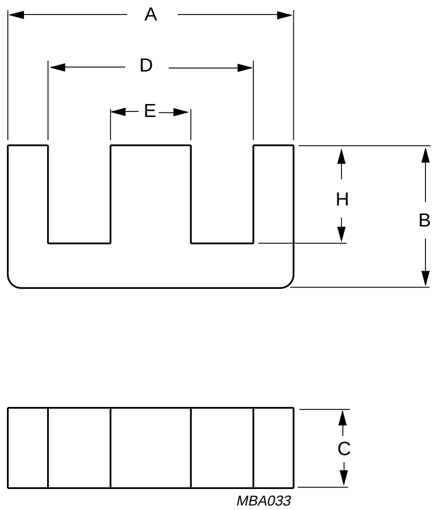


Fig.3 E core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
E; Planar E	3	Amax	Amin
		Bmax	Bmin
		Cmax	Cmin
		Dmin	Dmax
		Emax	Emin
		Hmin	Hmax

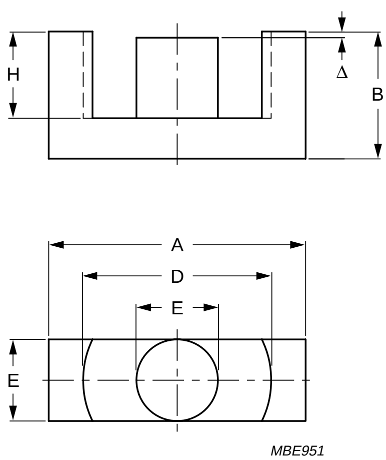


Fig.4 ETD core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
ETD/ER EFD	4	Amax	Amin
		Bmax	Bmin
		Cmax	Cmin
		Dmin	Dmax
		Emax	Emin
		Hmin	Hmax

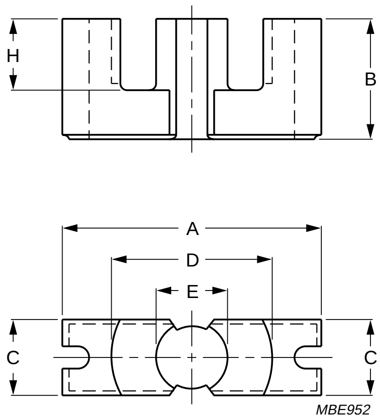


Fig.5 EC core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
EC	5	Amax	Amin
		Bmax	Bmin
		Cmax	Cmin
		Dmin	Dmax
		Emax	Emin
		Hmin	Hmax

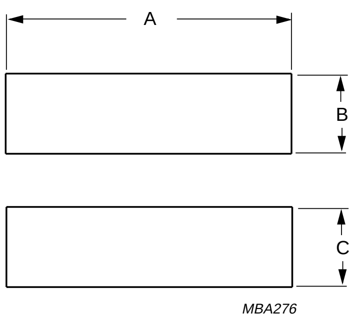


Fig.6 I core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
I	6		A
		Bmax	Bmin
		Cmax	Cmin

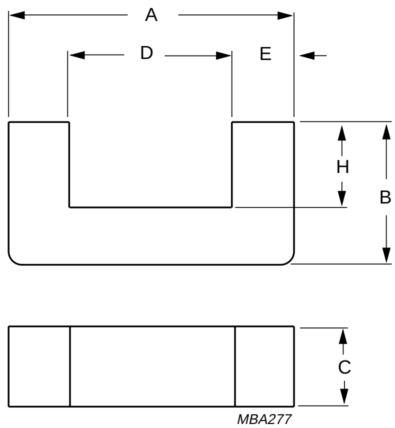


Fig.7 U core.

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
U	7		A
			B
		Cmax	Cmin
		Dmin	
		Emax	Emin
		Hmin	

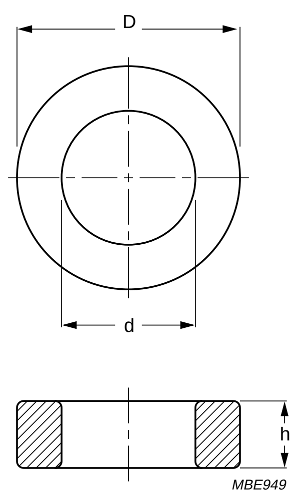


Fig.8 Ring cores (toroids).

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
ring cores (toroids)	8	hmax	hmin
		Dmax	Dmin
		dmin	dmax

Classification of mechanical defects

CORE TYPE	FIGURE	FAULT CLASSIFICATION	
		MAJOR	MINOR
rods; tubes; beads; multi-hole tubes	9	Dmax	Dmin
		dmin	dmax
			L
			H

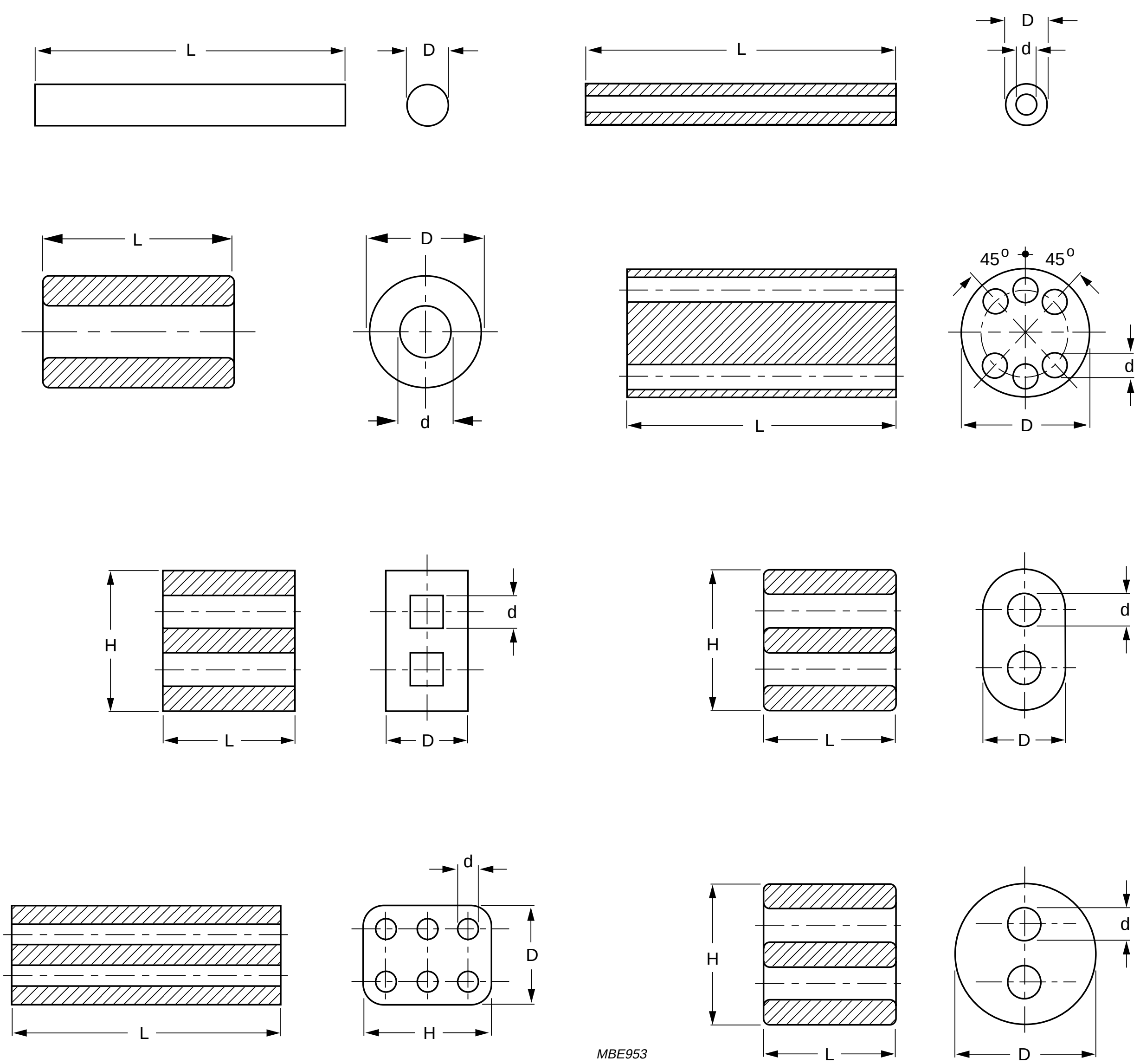


Fig.9 Rods, tubes, beads, multi-hole tubes.

ENVIRONMENTAL ASPECTS OF SOFT FERRITES

Our range of soft ferrites has the general composition $MeFe_2O_4$ where Me represents one or several of the divalent transition metals such as manganese (Mn), zinc (Zn), nickel (Ni), or magnesium (Mg).

To be more specific, all materials starting with digit 3 are manganese zinc ferrites based on the MnZn composition.

Their general chemical formula is:

$Mn_{\delta}\ Zn_{(1-\delta)}\ Fe_2\ O_4$

Materials starting with digit 4 are nickel zinc ferrites based on the NiZn composition. Their general chemical formula is:

$Ni_{\delta}\ Zn_{(1-\delta)}\ Fe_2\ O_4$

Materials starting with digit 2 are magnesium zinc ferrites based on the MgZn composition. Their general chemical formula is:

$Mg_{\delta}\ Zn_{(1-\delta)}\ Fe_2\ O_4$

General warning rules

- With strong acids, the metals iron, manganese, nickel and zinc may be partially extracted.
- In the event of fire, dust particles with metal oxides will be formed.
- Disposal as industrial waste, depending on local rules and circumstances.

ORDERING INFORMATION

The products in this handbook are identified by type numbers. All physical and technical properties of the product are expressed by these numbers. They are therefore recommended for both ordering and use on technical drawings and equipment parts lists.

The 11-digit code, used in former editions of this data handbook, also appears on packaging material.

Smallest Packaging Quantities (SPQ) are packs which are ready for shipment to our customers. The information on the barcoded label consists of:

- Technical information:
  - type number
  - 11-digit code number
  - delivery and/or production batch numbers
- Logistic information:
  - 12-digit code number
  - quantity
  - country of origin
  - production week
  - production centre.

The Philips 12-digit code used on the packaging labels, provides full logistic information as well.

During all stages of the production process, data are collected and documented with reference to a unique batch number, which is printed on the packaging label. With this batch number it is always possible to trace the results of process steps afterwards and in the event of customer complaints, this number should always be quoted.

Products are available troughout their lifecycle. A short definition of product status is given in the table “Product status definitions”.

Product status definitions

STATUS	INDICATION	DEFINITION
Prototype	prot	These are products that have been made as development samples for the purposes of technical evaluation only. The data for these types is provisional and is subject to change.
Design-in	des	These products are recommended for new designs.
Preferred		These products are recommended for use in current designs and are available via our sales channels.
Support	sup	These products are **not** recommended for new designs and may not be available through all of our sales channels. Customers are advised to check for availability.

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.

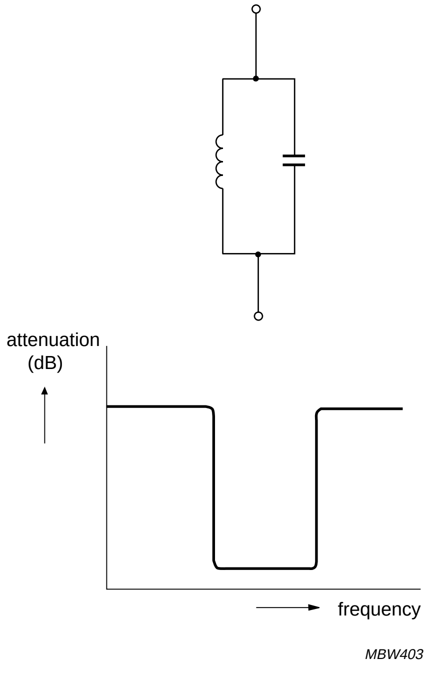


Fig.1 Filter application.

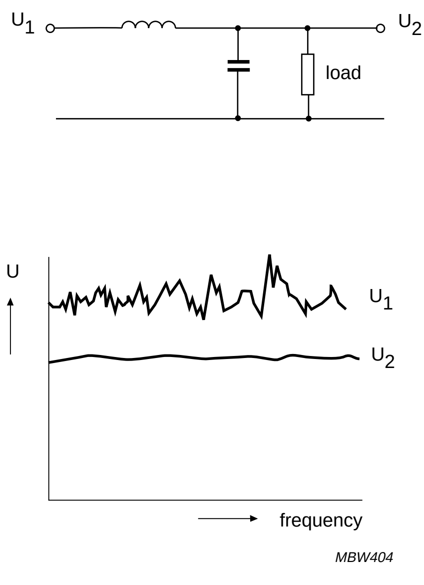


Fig.2 Suppression application.

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.

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: 3C15, 3C30, 3C34, 3C90, 3C94, 3C96 2P-iron powder.

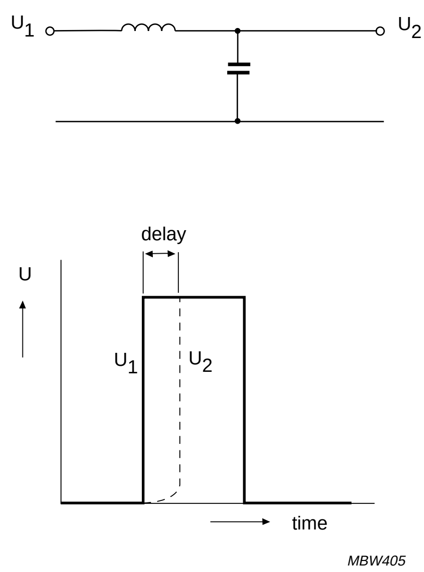


Fig.3 Pulse delay application.

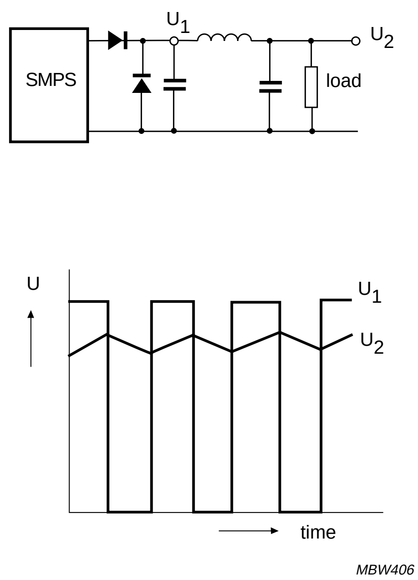


Fig.4 Smoothing/storage application.

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, 3C94, 3C96, 3F3, 3F35, 3F4, 4F1.

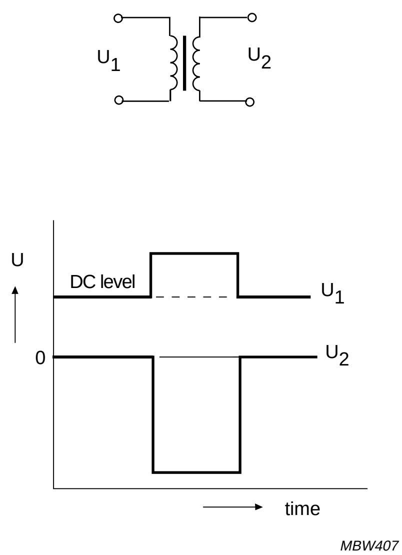


Fig.5 Pulse and general purpose transformer.

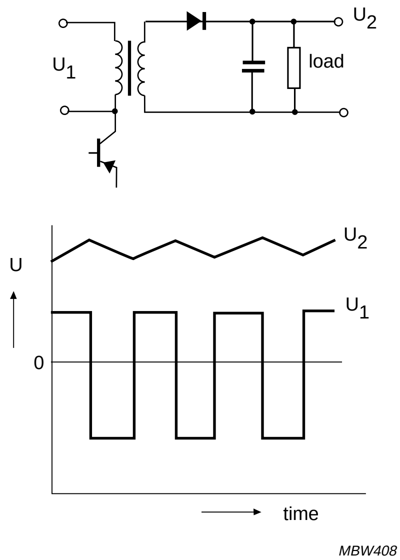


Fig.6 Power transformer application.

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, 4D2, 4E1.

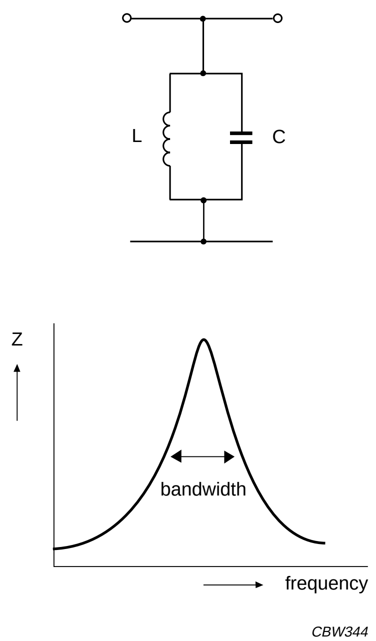


Fig.7 Tuning application.

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.

PULSE 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.

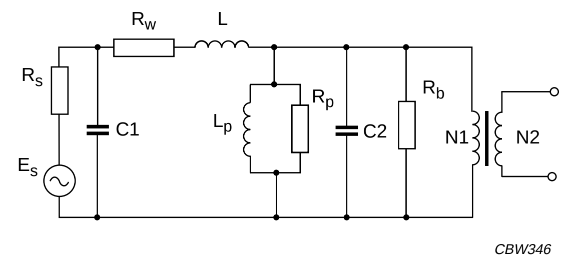


Fig.8 Simplified equivalent circuit of a transformer.

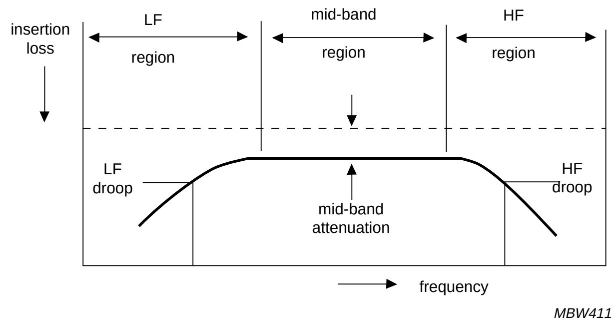


Fig.9 Transmission characteristic of a wideband transformer.

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

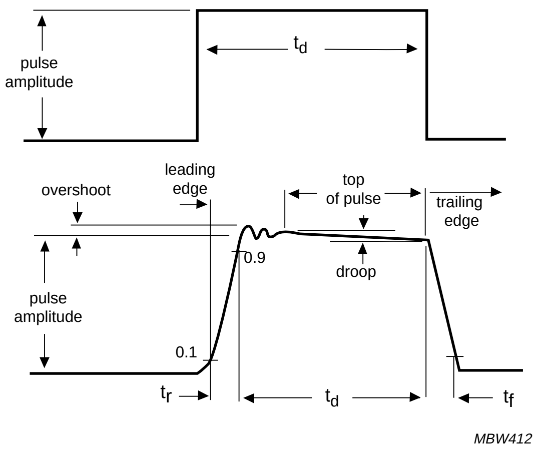


Fig.10 An ideal rectangular pulse and the main pulse distortions that may be introduced by a transformer.

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 a major application area for modern ferrites. Originally designed for use as line output transformers in television receivers, power cores are now being used in a wide range of applications. The introduction of Switched Mode Power Supplies (SMPS) has stimulated the development of a number of new ferrite grades and core shapes to be used in the manufacture of power transformers, output chokes and input filters.

Power transformers and inductors generally operate under loss or saturation limited conditions which require special power ferrites with high saturation levels and low losses at elevated temperatures.

Output chokes must tolerate high DC currents; this means a gapped magnetic circuit or a special material with very high saturation level such as iron powder.

Input chokes prevent mains pollution generated by the SMPS. Therefore grades are used which provide maximum blocking impedances at the switching frequencies.

SWITCHED MODE POWER SUPPLY CIRCUITS

The basic arrangement of a Switched Mode Power Supply (SMPS) is shown in Fig.11.

In this configuration, the power input is rectified and the resulting DC voltage is chopped by a switch at a high frequency. The chopped waveform is applied to the primary of a transformer and the secondary output is rectified and filtered to give the required DC output. The output voltage is sensed by a control circuit which supplies a correction signal to the drive circuit to vary the ON/OFF time of the switched waveform and compensate for any change at the output.

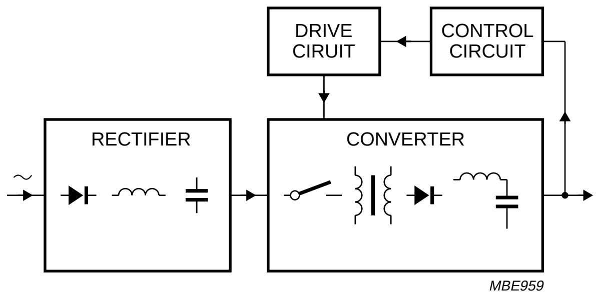


Fig.11 Block diagram of a Switched Mode Power Supply.

Numerous circuit designs can be used to convert DC input voltage to the required DC output voltage. The requirements for the transformer or inductor depend largely on the choice of this circuit technology.

If the circuits are analyzed in this way, three basic converter designs can be distinguished, based upon the magnetic converting device.

These are:

- Flyback converters
- Forward converters, and
- Push-pull converters.

FLYBACK CONVERTER

Figure 12 shows the basic circuit of a flyback converter and its associated waveforms.

When the switch is closed (transistor conducts), the supply voltage is connected across the inductor and the output diode is non-conducting. The current rises linearly, storing energy, until the switch is opened. When this happens, the voltage across the inductor reverses and the stored energy is transferred into the output capacitor and load. By varying the conduction time of the transistor at a given frequency the amount of energy stored in the inductor during each ON cycle can be controlled. This allows the output of the SMPS to be controlled and changed.

This basic circuit can be developed into a practical circuit using an inductor with two windings (see Fig.13).

In a flyback converter, all the energy to be transferred to the output capacitor and load is, at first, stored in the inductor. It is therefore possible to obtain line isolation by adding a secondary winding to the inductor (although an inductor with more than one winding appears in schematic diagrams as a transformer, it is referred to as an inductor in accordance with its function).

Another advantage of the flyback converter is that no smoothing choke is required in the output circuit. This is important in high-voltage supplies and in power supplies with a number of output circuits (see Fig.14).

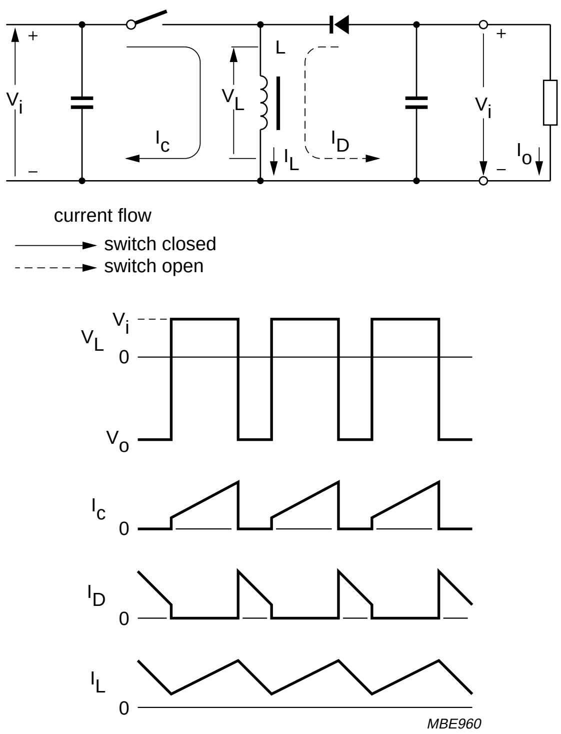


Fig.12 Basic circuit of a DC to DC flyback converter with associated waveforms.

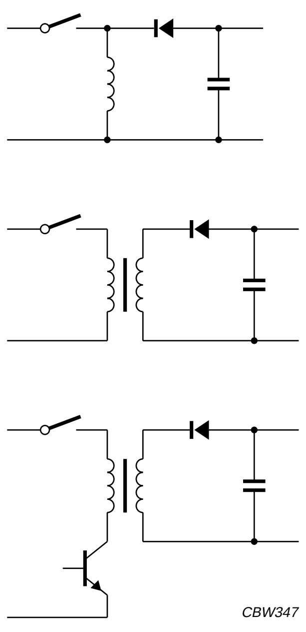


Fig.13 Practical flyback converter circuits.

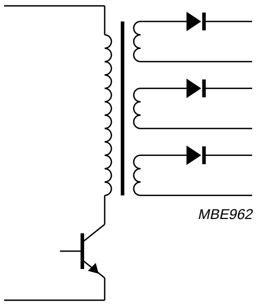


Fig.14 Multiple output flyback converter circuit.

A disadvantage of this type of converter is that the output capacitor is charged only during the transistor's OFF cycle. Hence the output capacitor ripple current is high when compared with the other types of converters.

Another disadvantage of the flyback converter concerns the energy stored in the inductor. The inductor is driven in one direction only; this requires a larger core in a flyback design than for an equivalent design using a forward or push-pull converter.

FORWARD CONVERTER

The basic circuit of the forward converter, together with its associated voltage and current waveforms is shown in Fig.15.

When the switch is closed (transistor conducts), the current rises linearly and flows through the inductor into the capacitor and the load. During the ON cycle, energy is transferred to the output and stored in the inductor 'L'. When the switch is opened, the energy stored in the inductor causes the current to continue to flow to the output via the diode.

As with the flyback converter, the amount of energy stored in the inductor can be varied by controlling the ON/OFF cycles. This provides control of the output of the forward converter.

A more practical forward converter circuit with a line-isolation transformer is shown in Fig.16.

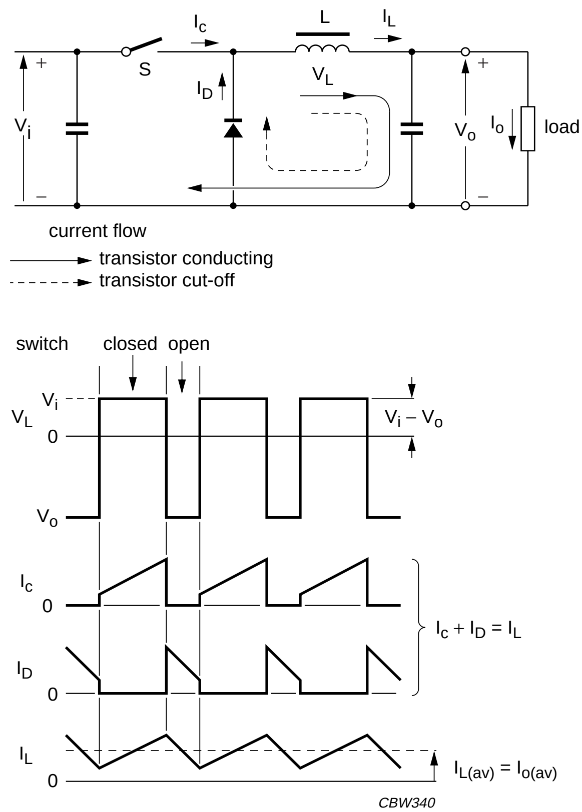


Fig.15 Basic circuit of a DC to DC forward converter with associated waveforms.

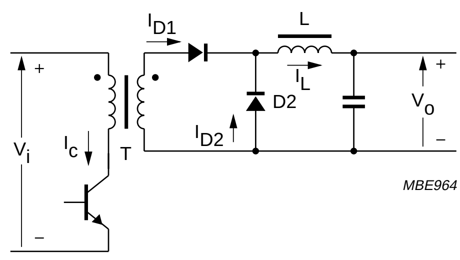


Fig.16 Forward converter with line isolation.

PUSH-PULL CONVERTER

The basic circuit of the push-pull converter, with voltage and current waveforms is shown in Fig.17.

The push-pull converter is an arrangement of two forward converters operating in antiphase (push-pull action). With switch S1 closed (Fig.17a) diode D2 conducts and energy is simultaneously stored in the inductor and supplied to the load. With S1 and S2 open (Fig.17b), the energy stored in the inductor continues to support the load current via the parallel diodes D1 and D2, which are now acting as flywheel diodes. When switch S2 closes (Fig.17c), diode D1 continues to conduct, diode D2 stops conducting and the process repeats itself.

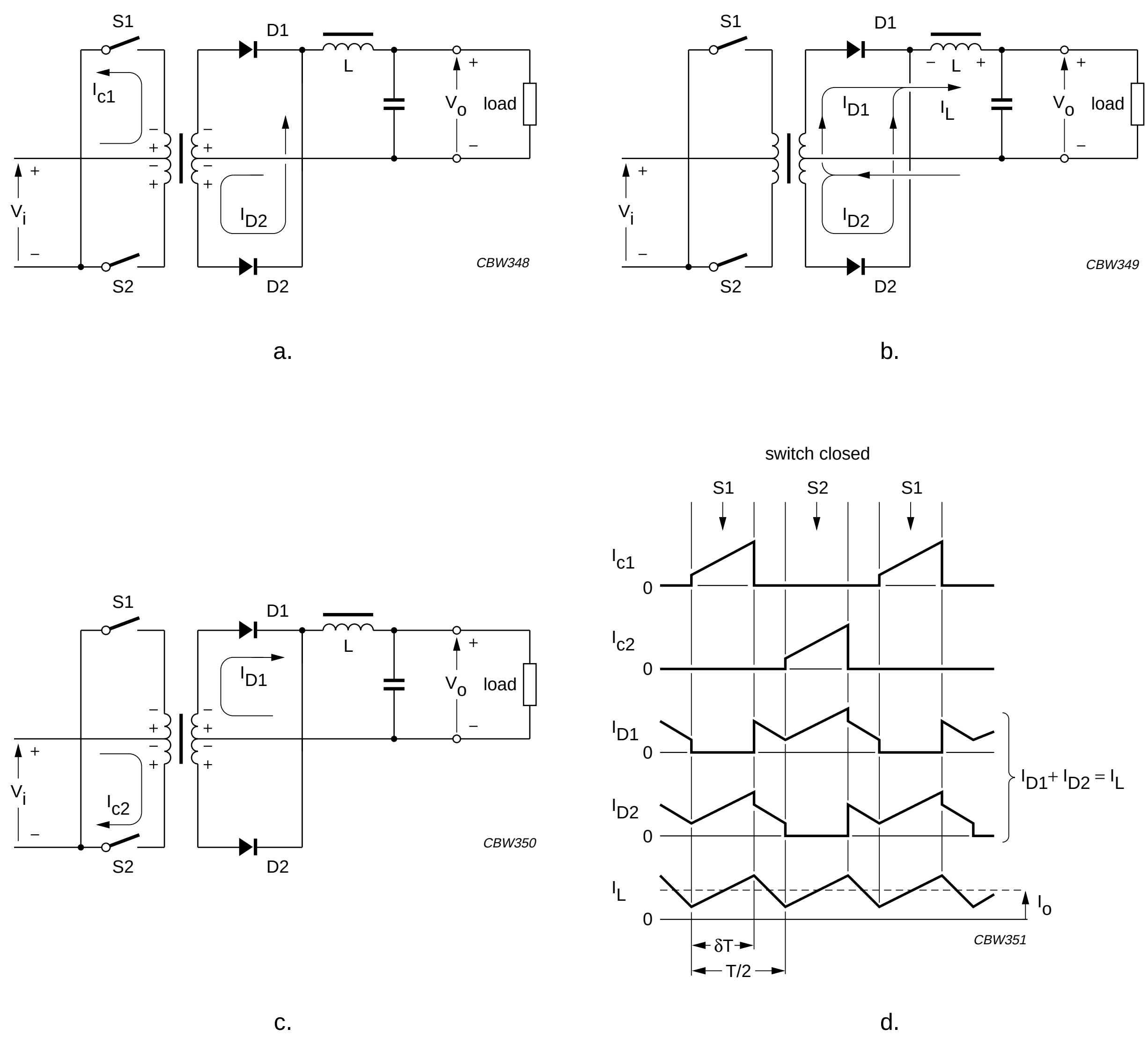


Fig.17 Basic circuit of a DC to DC push-pull converter with associated waveforms.

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.

Multiple outputs can be constructed by using several secondary windings, each with its own output diodes, inductor and smoothing capacitor.

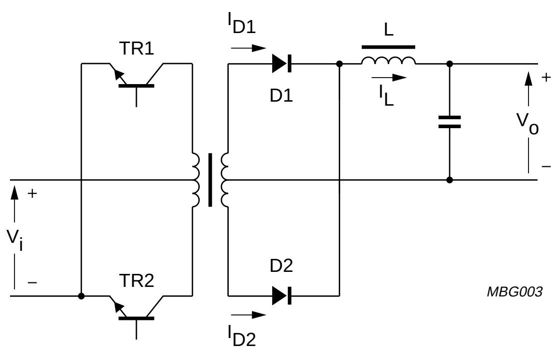


Fig.18 Conventional push-pull converter circuit.

CONVERTER SELECTION

In each of the three basic converter designs there are several different circuit possibilities. In the flyback and forward converters, single and two-transistor designs can be used. If two transistors are used, they will switch simultaneously. This type of circuit preference is determined by the allowable collector-emitter voltage and collector current of the transistor. In push-pull converter designs, the primary of the transformer can be connected in several ways (see Fig.19).

Depending upon how the transformer primary is driven, it is possible to differentiate between single-ended (see Fig.19a), push-pull (see Fig.19b) and full-bridge circuits (see Fig.19c). Decisions on circuit details are determined by the transistor capabilities.

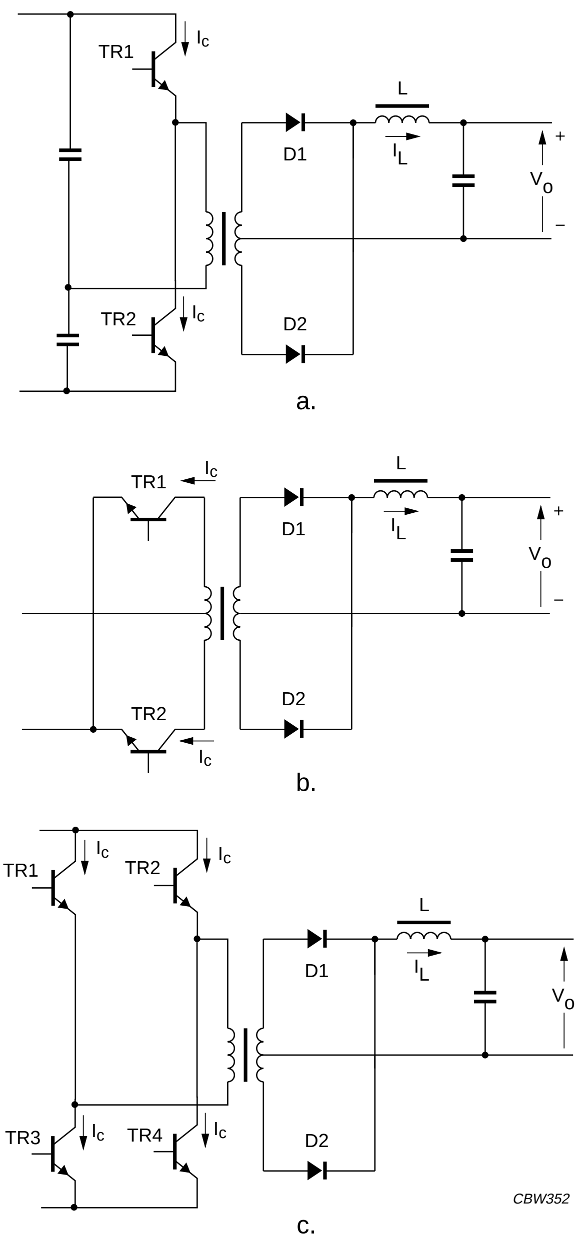


Fig.19 Several push-pull converter circuits.

For a practical converter design, the first selection that should be considered is the type of converter circuit to use. To aid in this initial converter circuit selection, Fig.20 offers a rough guide to the type of converter, its output voltage and power capability. This selection has to be considered along with other requirements, including line isolation, ripple content, overall efficiency, multiple outputs, etc.

Table 1 summarizes the most significant properties of a converter design. It shows the relative strengths and weaknesses of the three types of converters with regard to these characteristics.

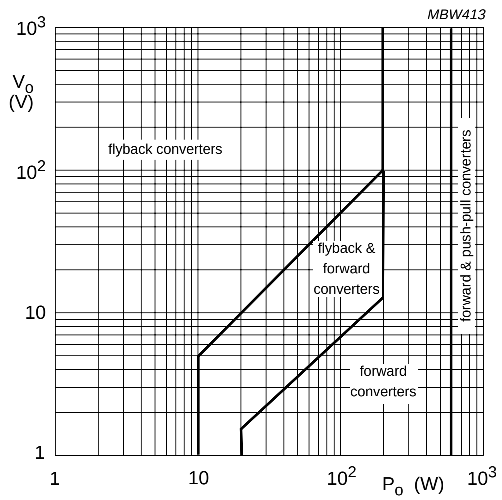


Fig.20 Converter circuit selection as a function of output voltage and throughput power.

For a high performance, high power, single output supply, where ripple is well below 1%, the push-pull design is the obvious choice. For smaller power versions of this type of supply, the forward, or double-forward converter provides a useful alternative to push-pull converter.

In high-voltage supplies, the flyback converter is the most suitable circuit and should be considered as a preference. In multiple-output supplies, the flyback converter is again normally the first choice because it avoids the necessity of providing a number of output windings on the inductor, together with a single diode and capacitor for each.

Table 1 Converter design selection chart (I)

FUNCTION	TYPE OF CONVERTER CIRCUIT(1)		
	FLYBACK	**FORWARD**	**PUSH-PULL**
Circuit simplicity	+	0	–
Number of components	+	0	–
Drive circuitry	+	0	–
Output ripple	–	0	+
Choke volume	not required	0	+
Transformer volume	–	0	+
Mains isolation	+	–	+
High power	–	0	+
High voltage	+	0	0
Multiple outputs	+	0	0

Note

1. ‘+’ = favourable; ‘0’ = average; ‘–’ = unfavourable.

CORE SELECTION

Table 2 shows which core type could be considered suitable for the different types of converter design.

The power-handling capability of a given core is determined by frequency and material grade, its geometry and available winding area, and by other factors which depend on the specific application.

Table 2 Converter design selection chart (II)

FUNCTION	TYPE OF CONVERTER CIRCUIT[1]		
	FLYBACK	FORWARD	PUSH-PULL
E cores	+	+	0
Planar E cores	–	+	0
EFD cores	–	+	+
ETD cores	0	+	+
ER cores	0	+	+
U cores	+	0	0
RM cores	0	+	0
EP cores	–	+	0
P cores	–	+	0
Ring cores	–	+	+

Note

1. ‘+’ = favourable; ‘0’ = average; ‘–’ = unfavourable.

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.

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 3 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 3 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; T56; U25; U30; E42; EFD30
200 to 500	ETD44; ETD49; E55; EC52; E42; P42/29; U37
<500	E65; EC70; U93; U100

Choice of ferrite for power transformers

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

3C15

Low frequency (<100 kHz) material with improved saturation level. Suitable for flyback converters e.g. Line Output Transformers.

3C30

Medium 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 50 °C.

3C90

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

3C91

Medium frequency (<200 kHz) material with loss minimum around 50 °C.

3C94

Medium frequency material (<400 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 3 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.

There are two remedies against this effect:

- The use of gapped ferrite cores
- The use of a material with low permeability and high saturation.

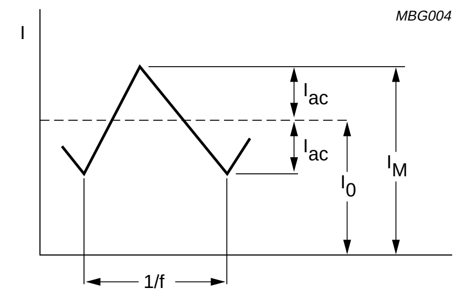


Fig.21 Choke waveform.

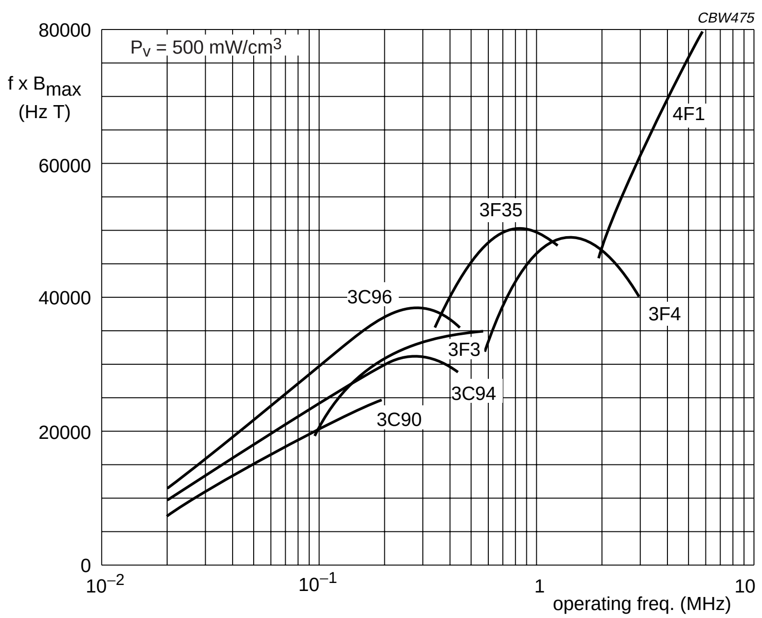


Fig.22 Performance factor ($f \times B_{max}$) at PV = 500 mW/cm^3 as a function of frequency for power ferrite materials.

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.

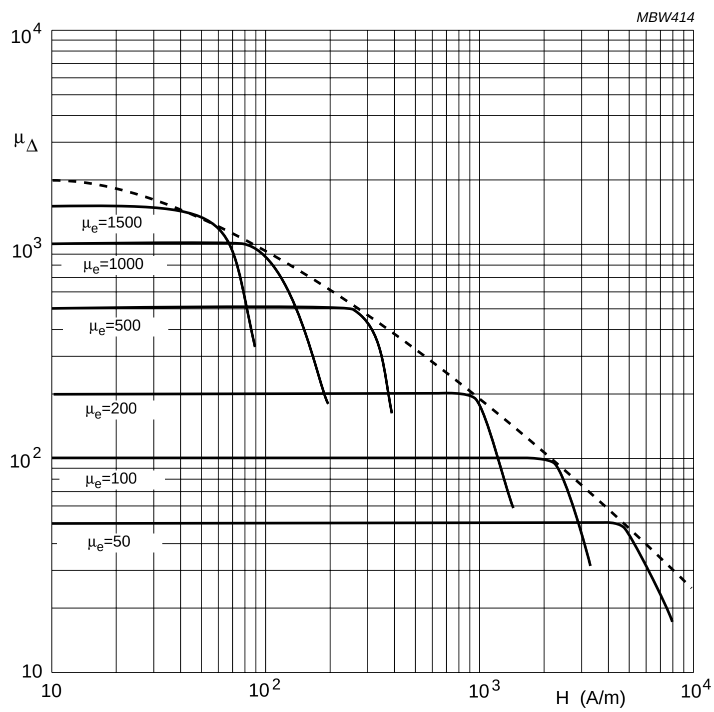


Fig.23 Effect of increased gap length.

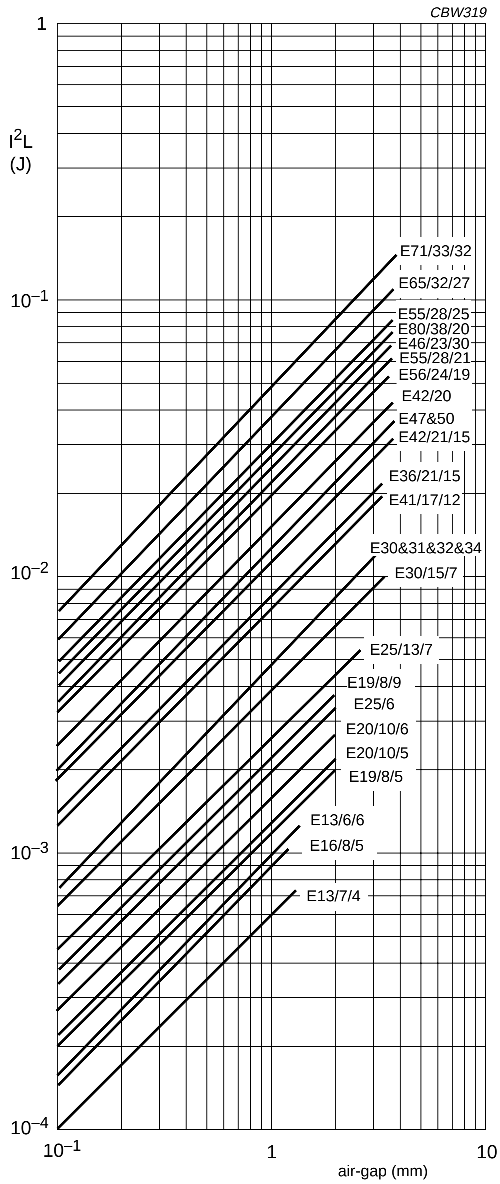


Fig.24 I^2L graph for E cores.

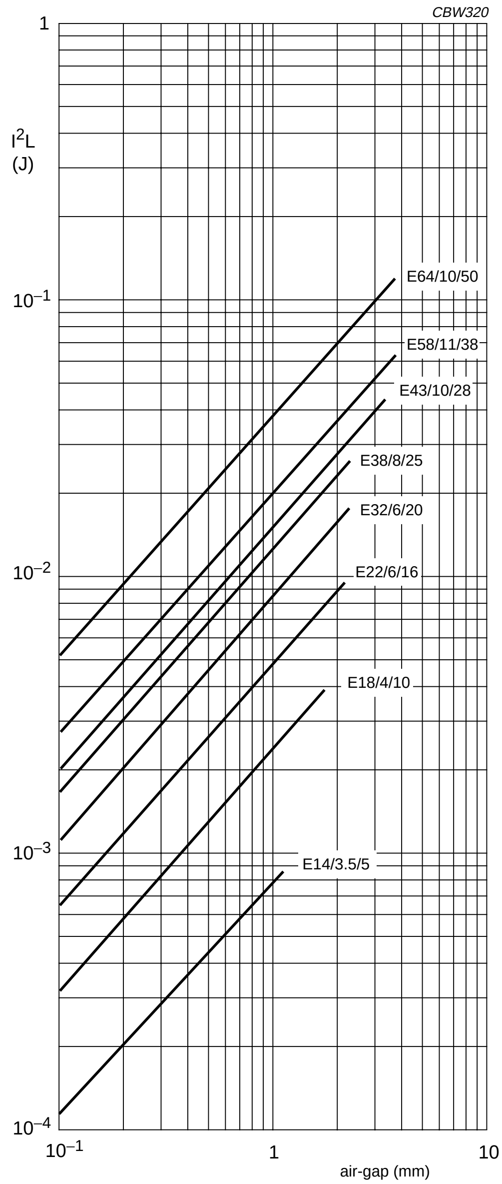


Fig.25 I^2L graph for planar E cores (valid for E + E and E + PLT combinations).

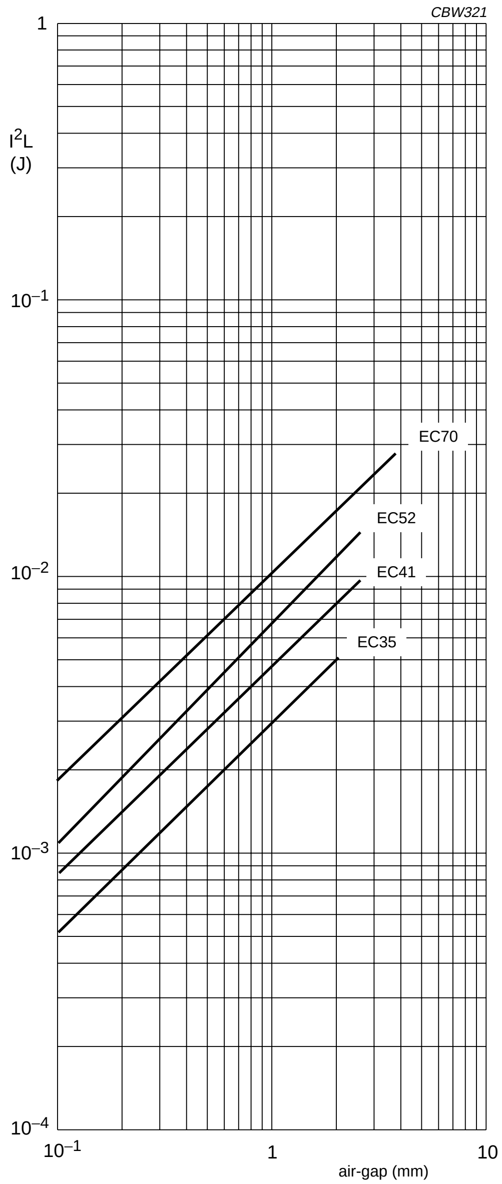


Fig.26 I^2L graph for EC cores.

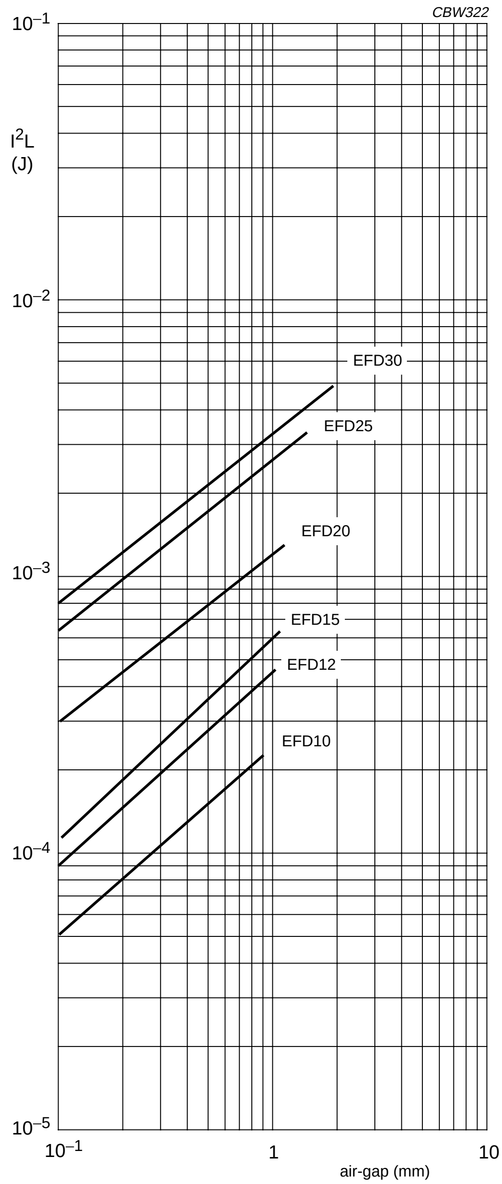


Fig.27 I^2L graph for EFD cores.

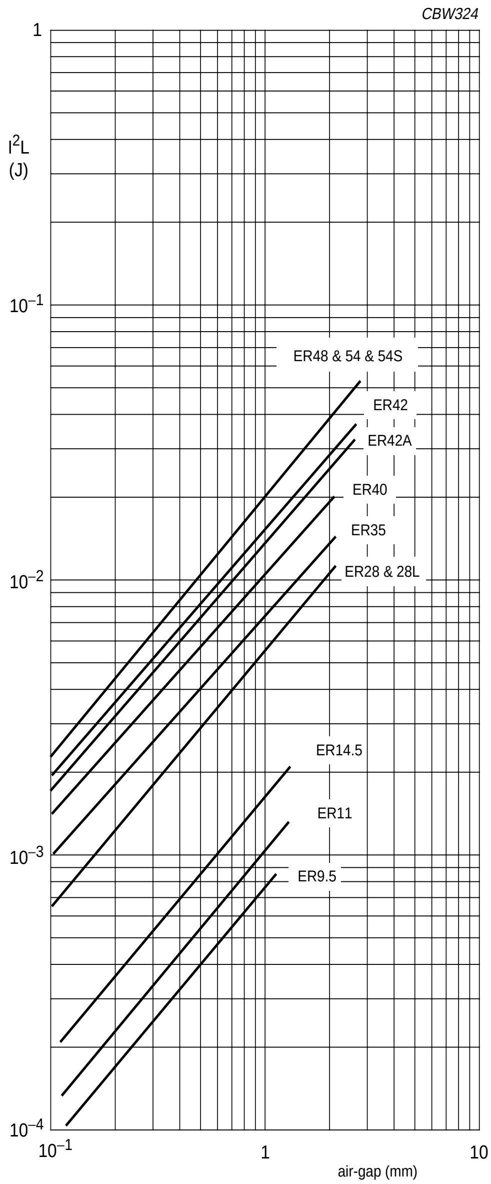


Fig.28 I^2L graph for ER cores.

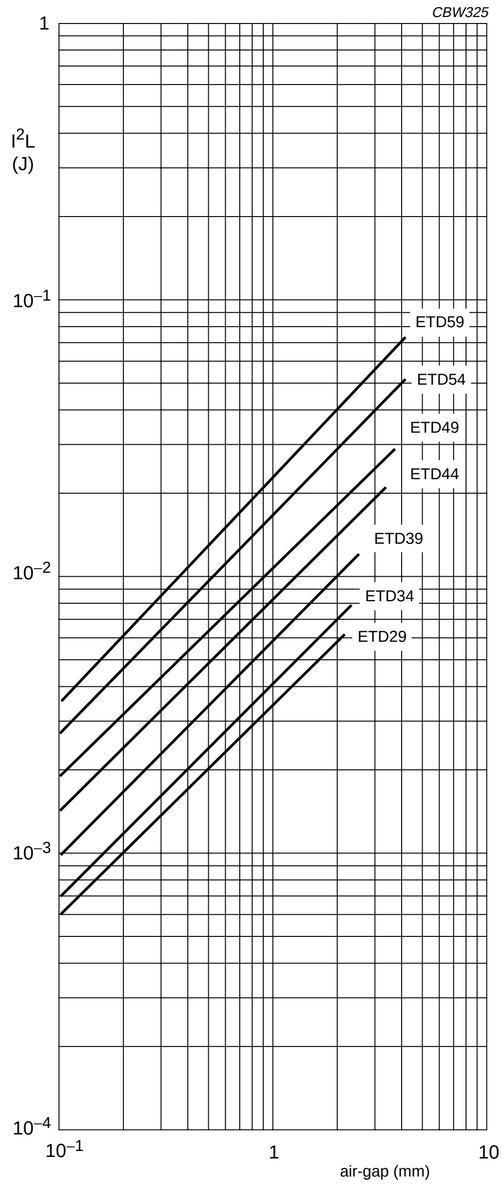


Fig.29 I^2L graph for ETD cores.

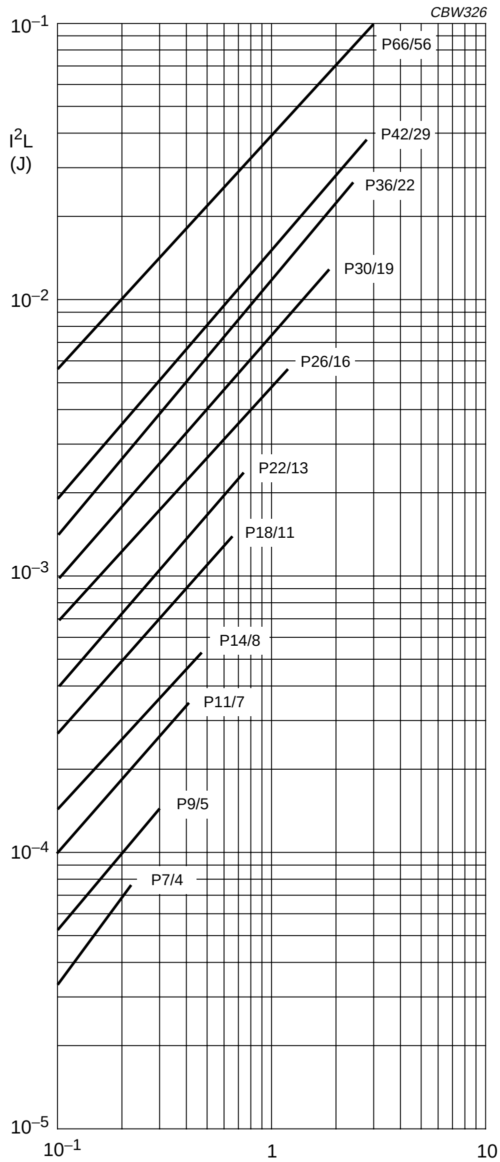


Fig.30 I^2L graph for P cores.

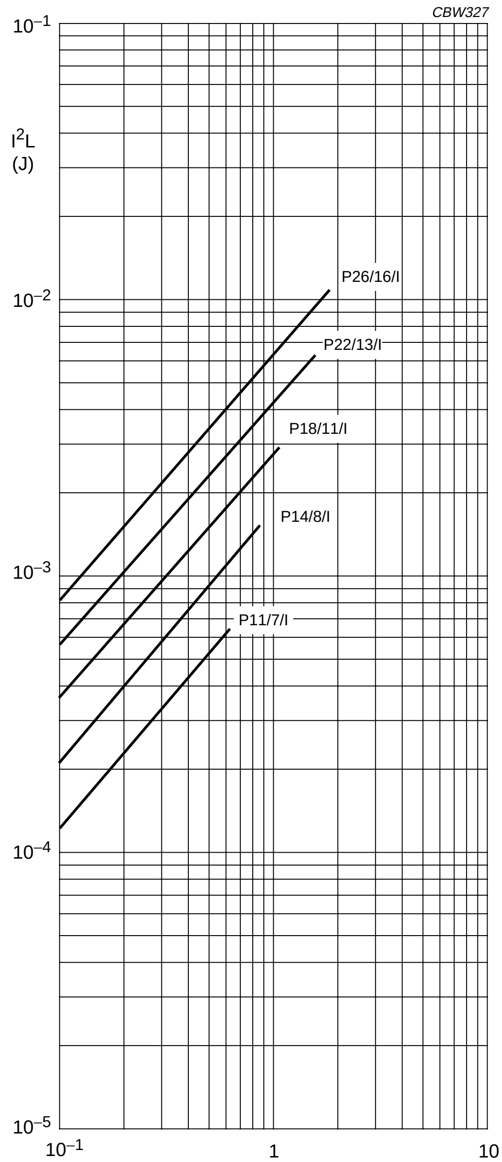


Fig.31 I^2L graph for P/I cores.

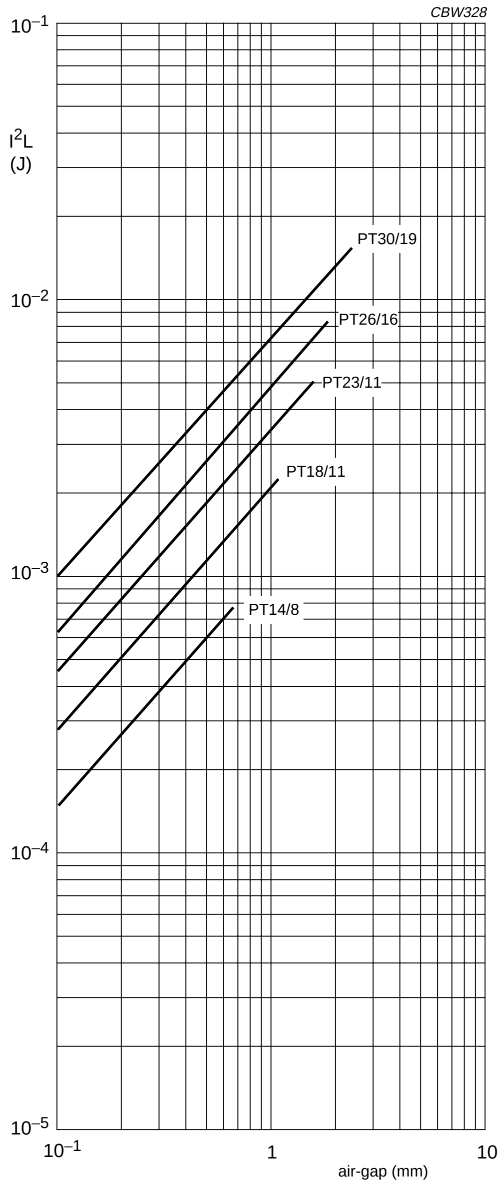


Fig.32 I^2L graph for PT cores.

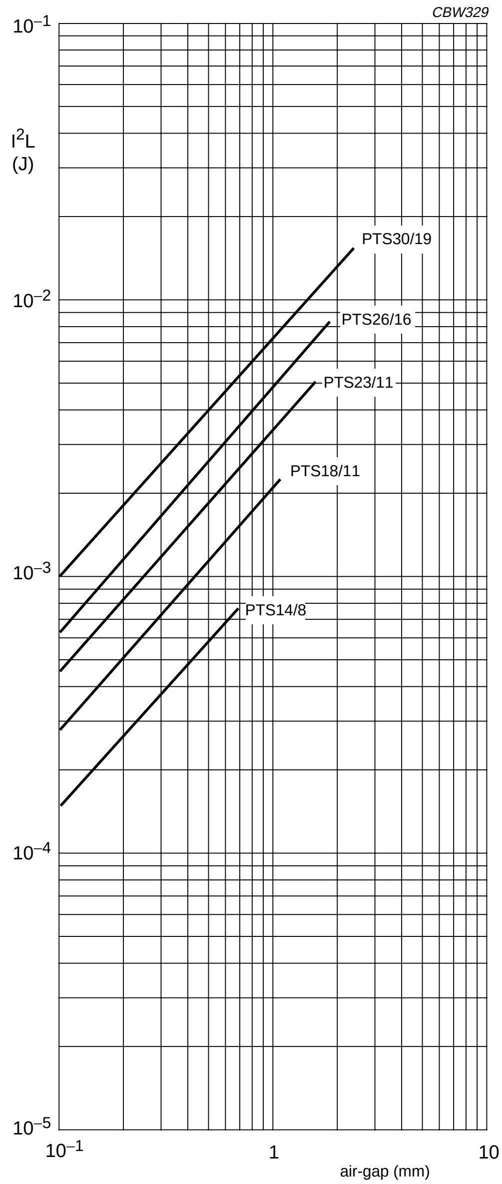


Fig.33 I^2L graph for PTS cores.

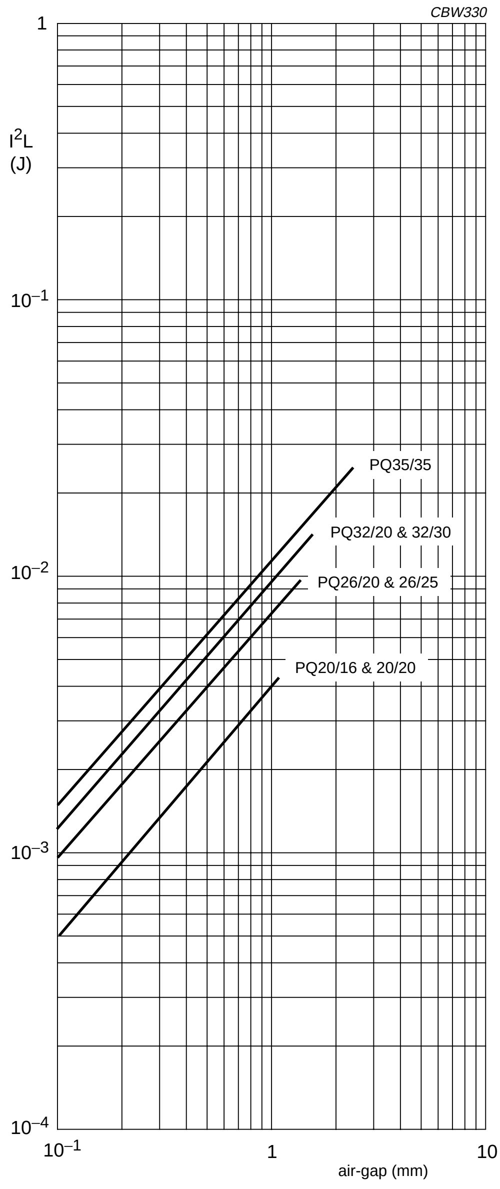


Fig.34 I^2L graph for PQ cores.

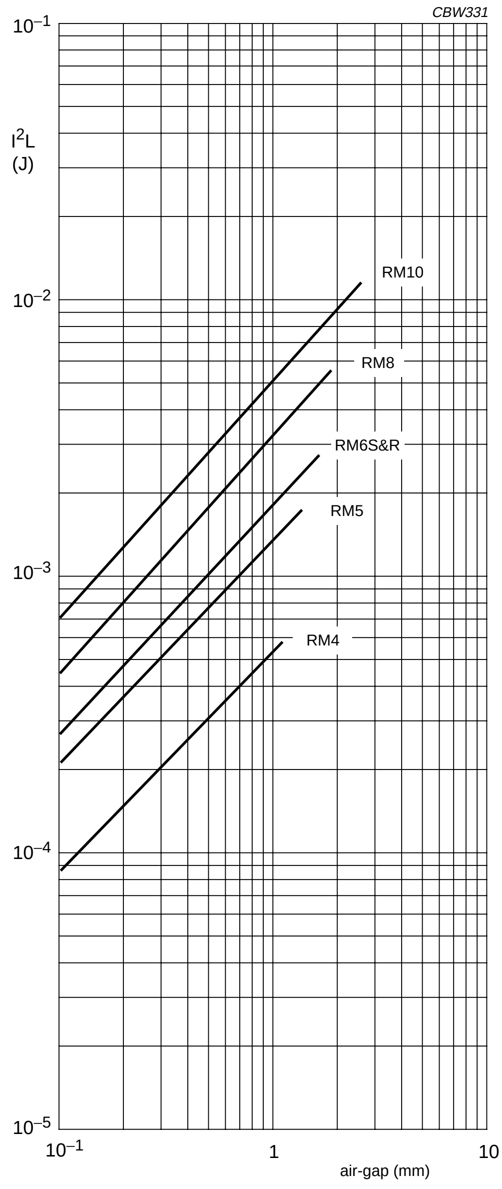


Fig.35 I^2L graph for RM cores.

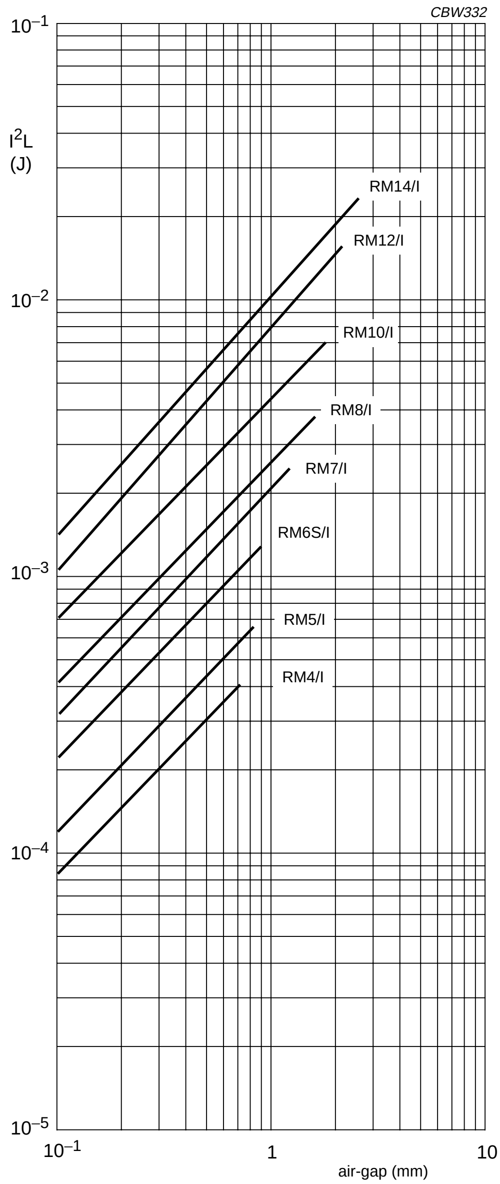


Fig.36 I^2L graph for RM/I cores.

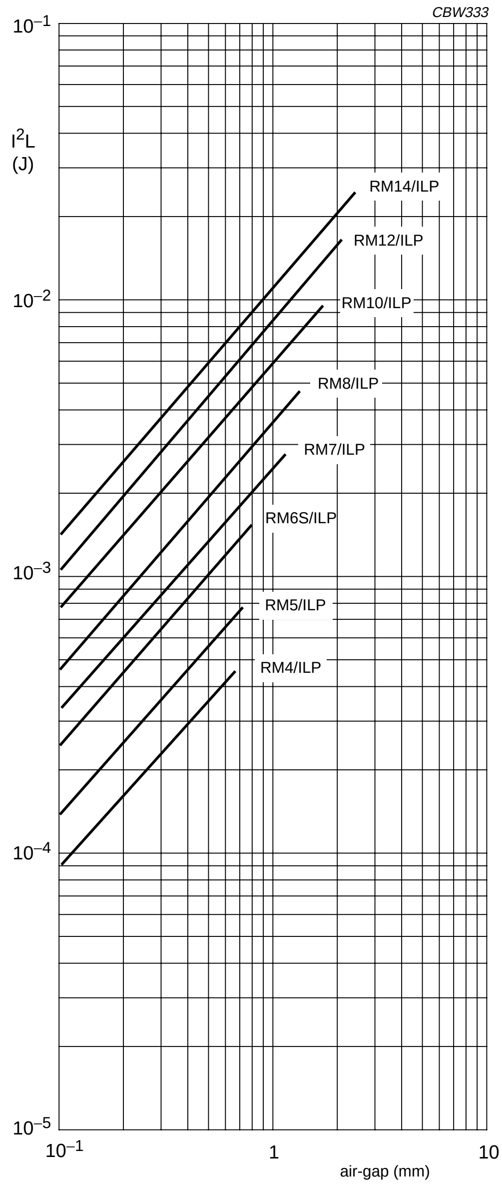


Fig.37 I^2L graph for RM/ILP cores.

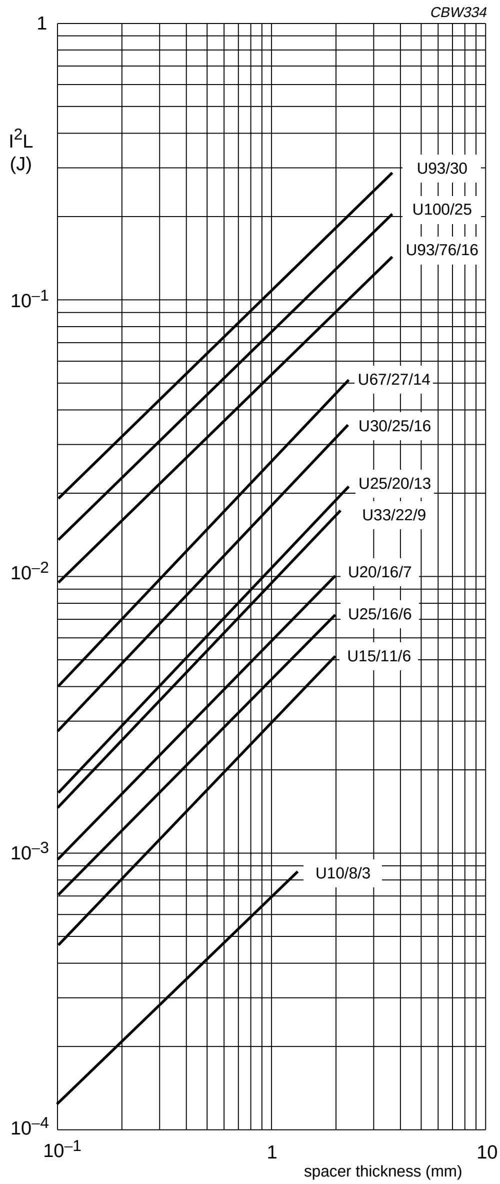


Fig.38 I^2L graph for U cores.

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)

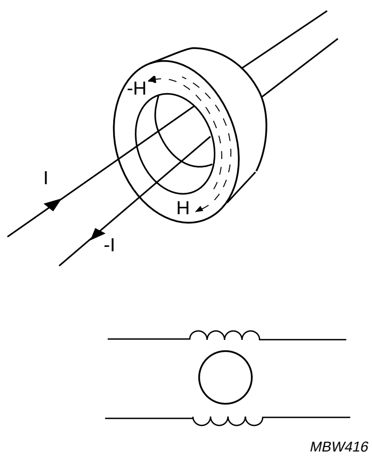


Fig.39 Common mode choke.

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.40.

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. Figure 41 provides a quick method of choosing the right ferrite for the job.

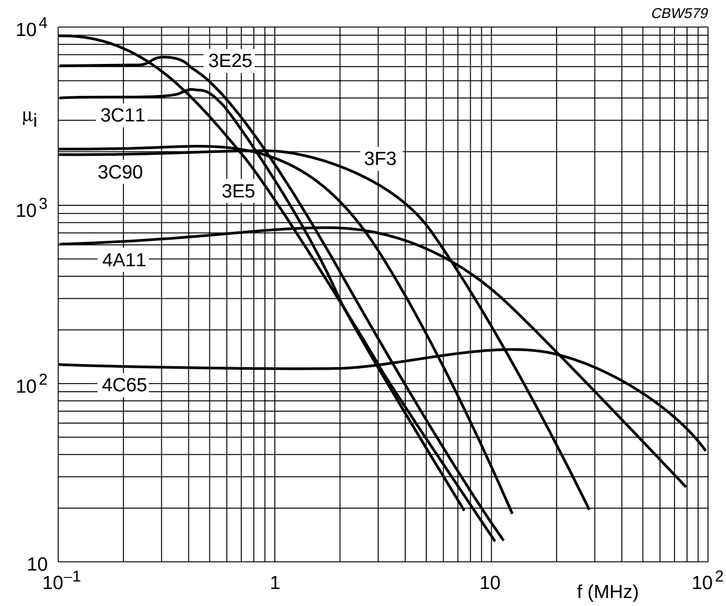


Fig.40 Permeability as a function of frequency of different materials.

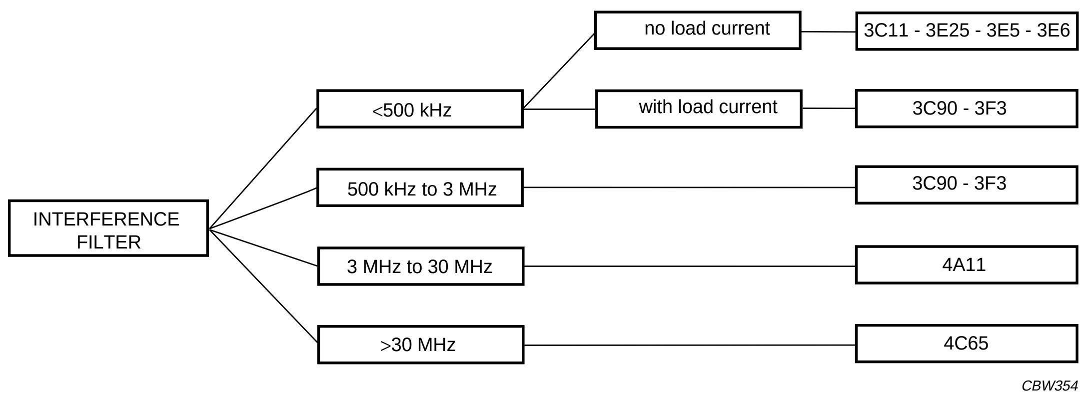


Fig.41 Selection chart for materials used in input filters.

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.42.

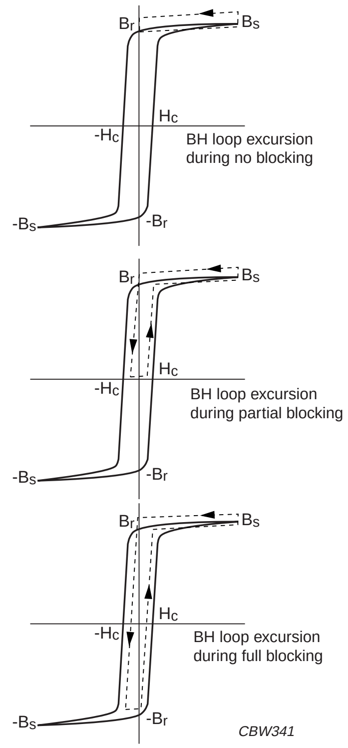


Fig.42 Behaviour of the ferrite material in a saturable inductor.

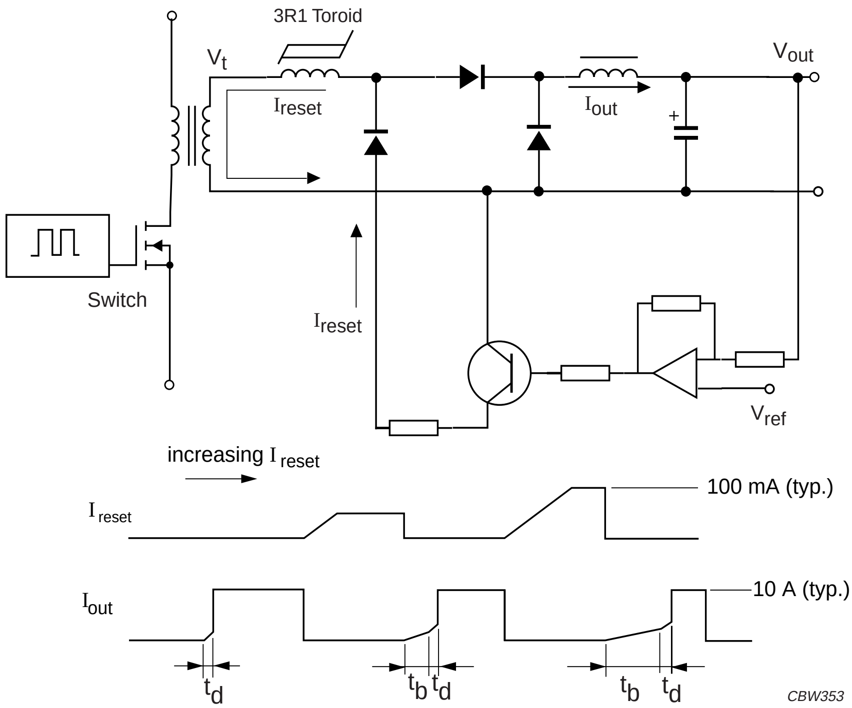


Fig.43 Schematic of a saturable inductor and associated waveforms (with regulation).

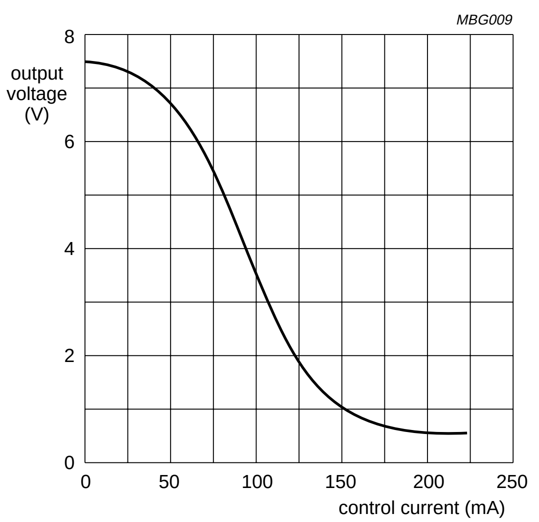


Fig.44 Typical control curve for a 3R1 ring core (size 14 × 9 × 5 mm, with 15 turns).

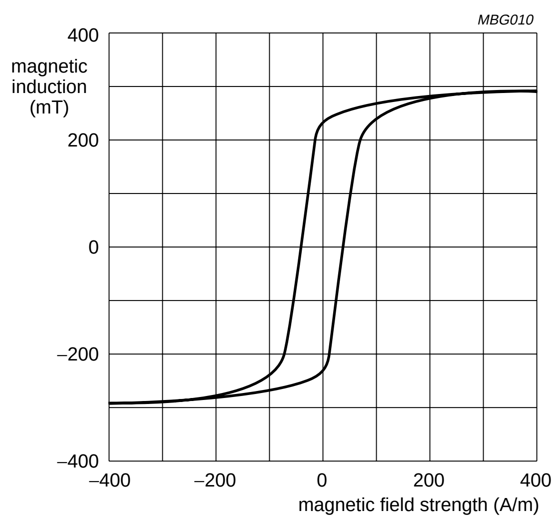


Fig.45 Properties of 3R1 ferrite material; f = 100 kHz, T = 25 °C.

Ferrites for Interference Suppression and Electromagnetic Compatibility (EMC)

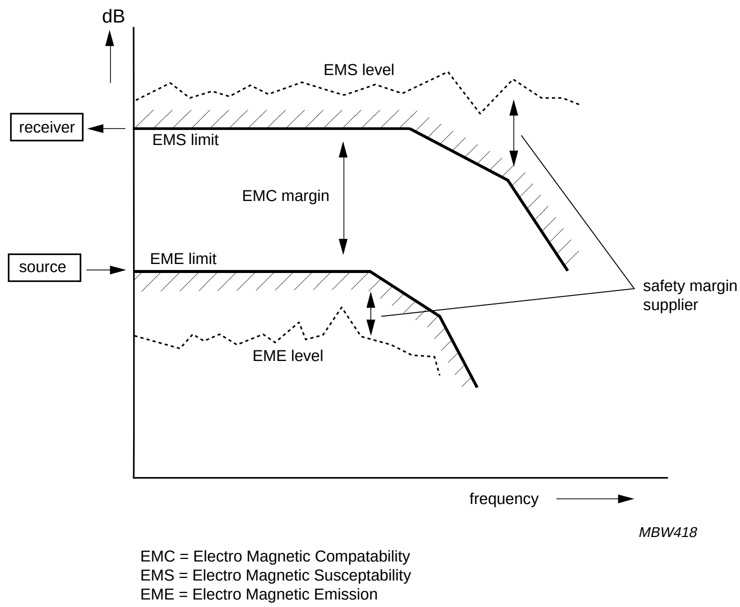


Fig.46 Principles of Electromagnetic Compatibility (EMC).

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 CISPR 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.47.

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.

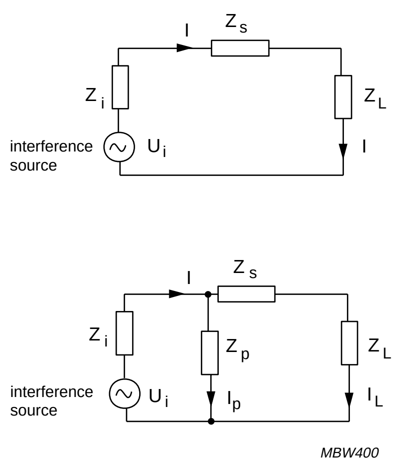


Fig.47 Basic suppression circuits.

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.

Philips have a full range of ring cores, beads, multilayer beads, 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.

New in the program are the Integrated Inductive Components (IIC).

SAMPLE BOXES

As the design process in these areas is often based on trial and error, we have assembled 6 different **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.

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 4S2 can be a better choice (see Figs 48, 49 and 50).

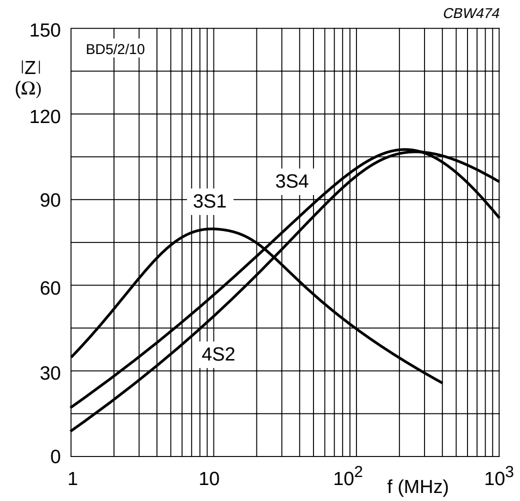


Fig.48 Impedance as a function of frequency for material grades 3S1, 3S4 and 4S2; bead size 5 × 2 × 10 mm.

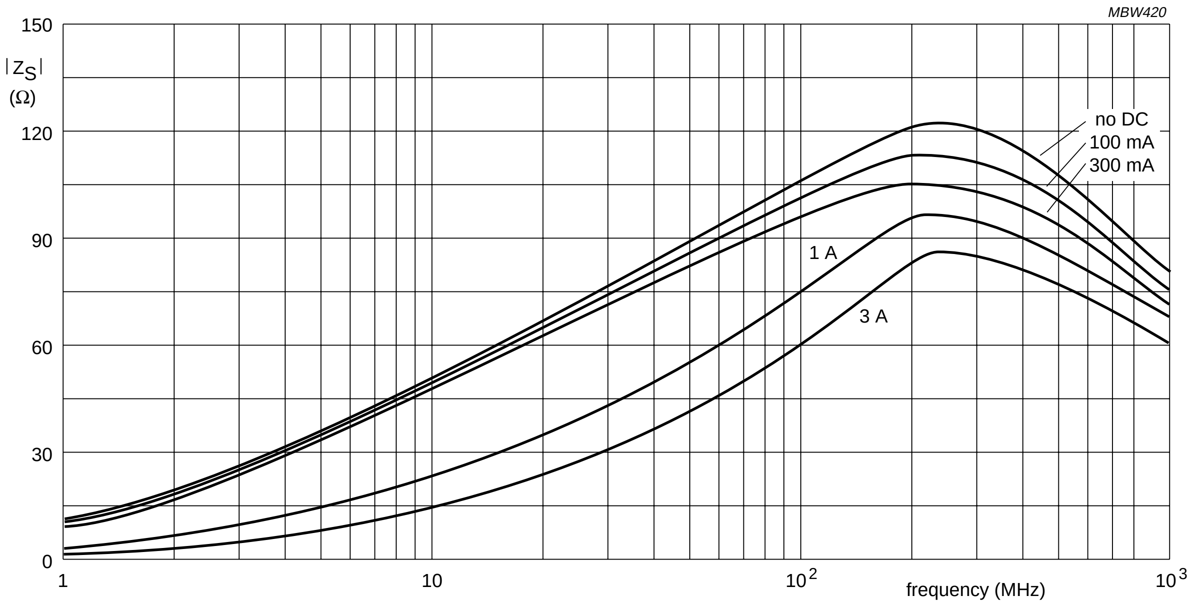


Fig.49 Impedance as a function of frequency at different DC levels for material grade 4S2.

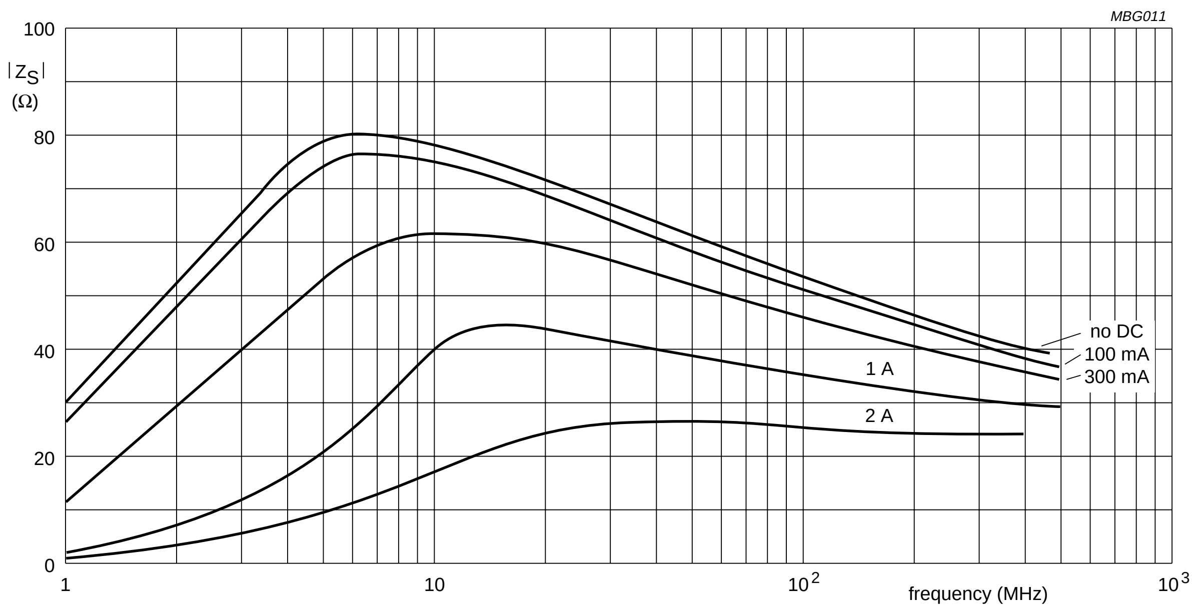


Fig.50 Impedance as a function of frequency at different DC levels for material grade 3S1.

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.

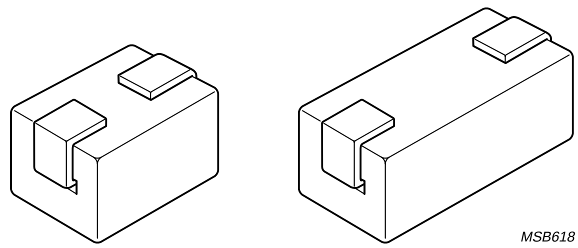


Fig.51 Outline of SMD beads.

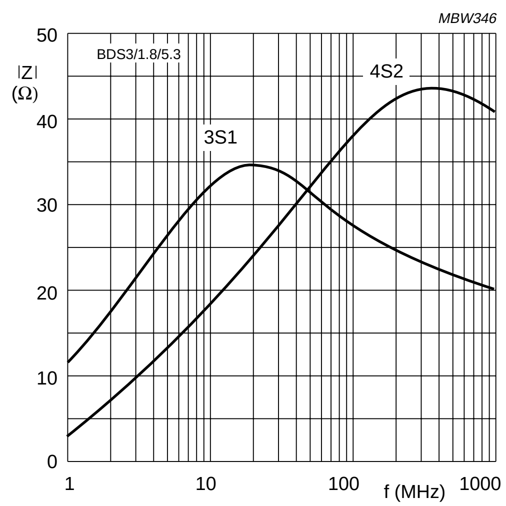


Fig.52 Impedance as a function of frequency for SMD beads.

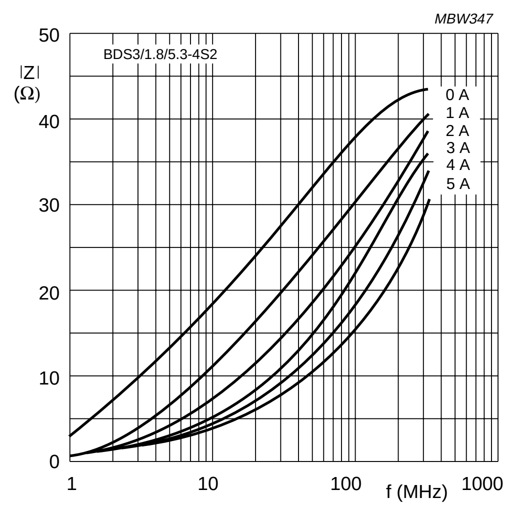


Fig.53 Impedance as a function of frequency for an SMD bead with bias current as a parameter.

SMD FERRITE BEADS FOR COMMON-MODE INTERFERENCE SUPPRESSION

Philips Components has introduced a new 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 Philips' new 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.

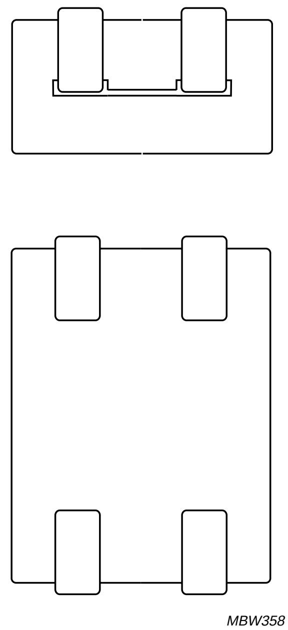


Fig.54 Outline of an SMD common-mode choke.

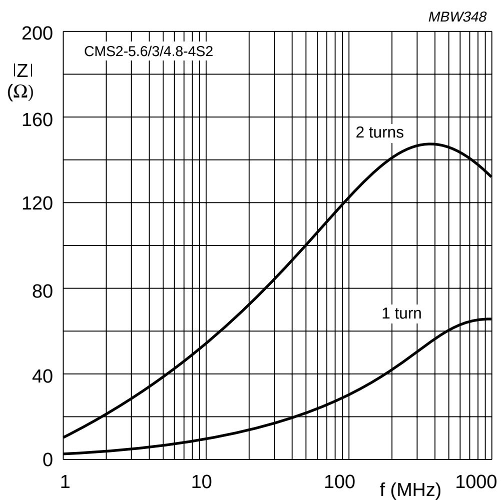


Fig.55 Impedance as a function of frequency of an SMD common mode bead with two conductors.

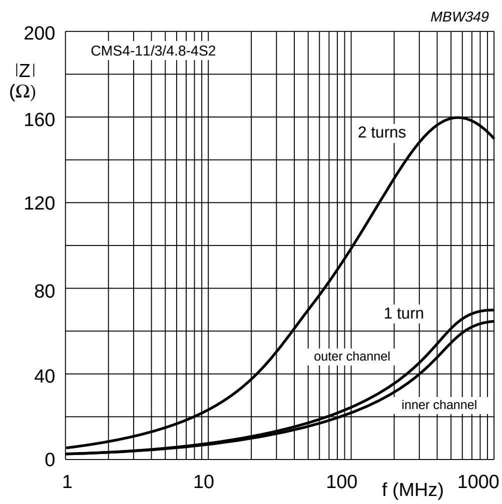


Fig.56 Impedance as a function of frequency of an SMD common mode bead with four conductors.

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.

Recently the range was extended with several new types, e.g. with isolation and taped on reel.

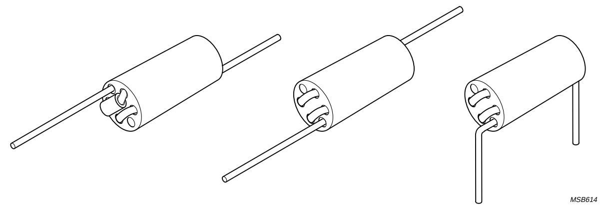


Fig.57 Outline of wideband chokes.

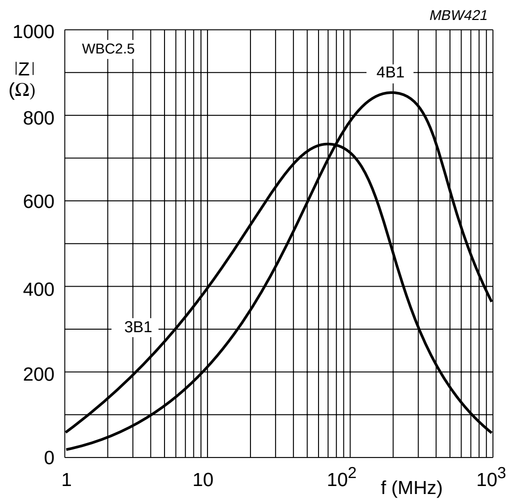


Fig.58 Impedance as a function of frequency for a wideband choke.

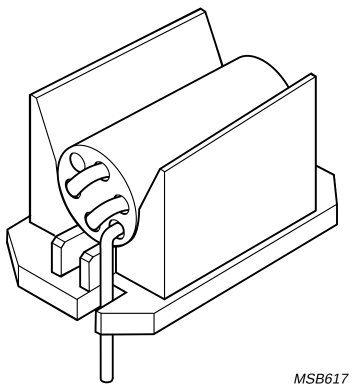


Fig.59 Outline of a wideband choke with support plate.

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.

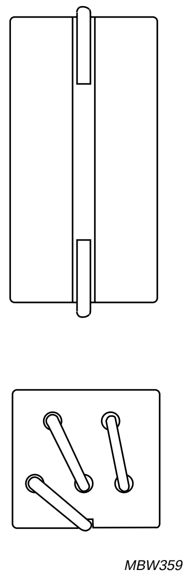


Fig.60 Outline of an SMD wideband choke.

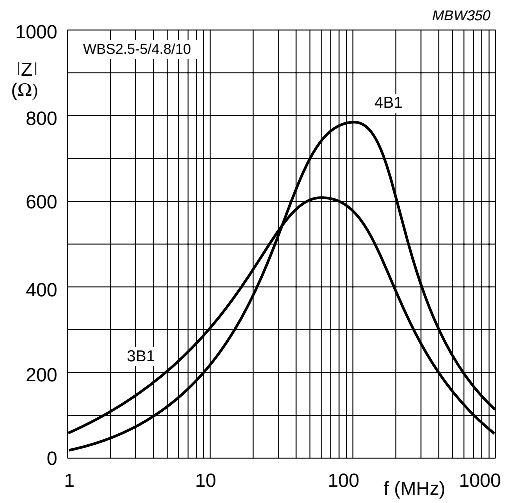


Fig.61 Impedance as a function of frequency for SMD wideband chokes.

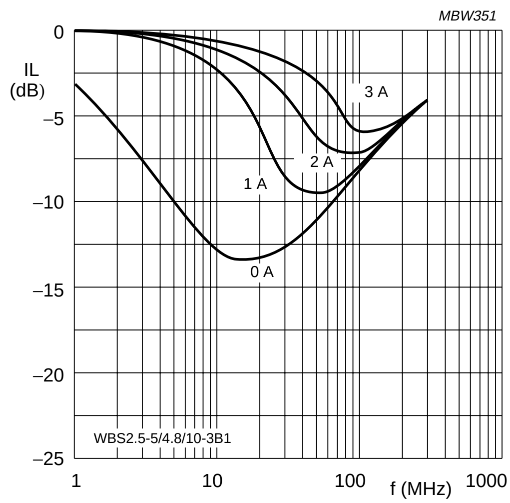


Fig.62 Insertion loss of a 3B1 SMD wideband choke as a function of frequency (50 Ω circuit).

Cable shields

New 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 new 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.

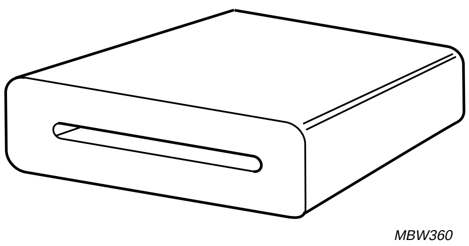


Fig.63 Outline of a cable shield.

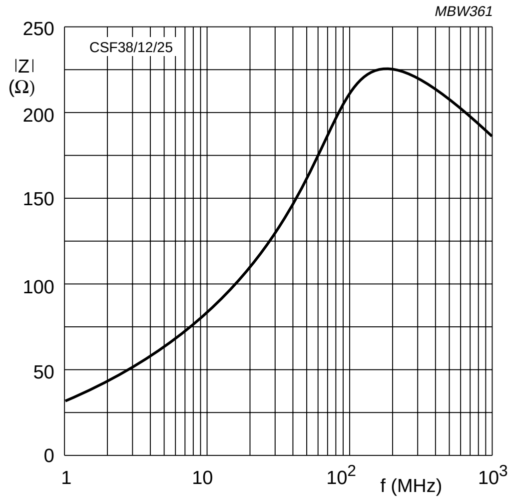


Fig.64 Impedance of a cable shield as a function of frequency.

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.65) 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.

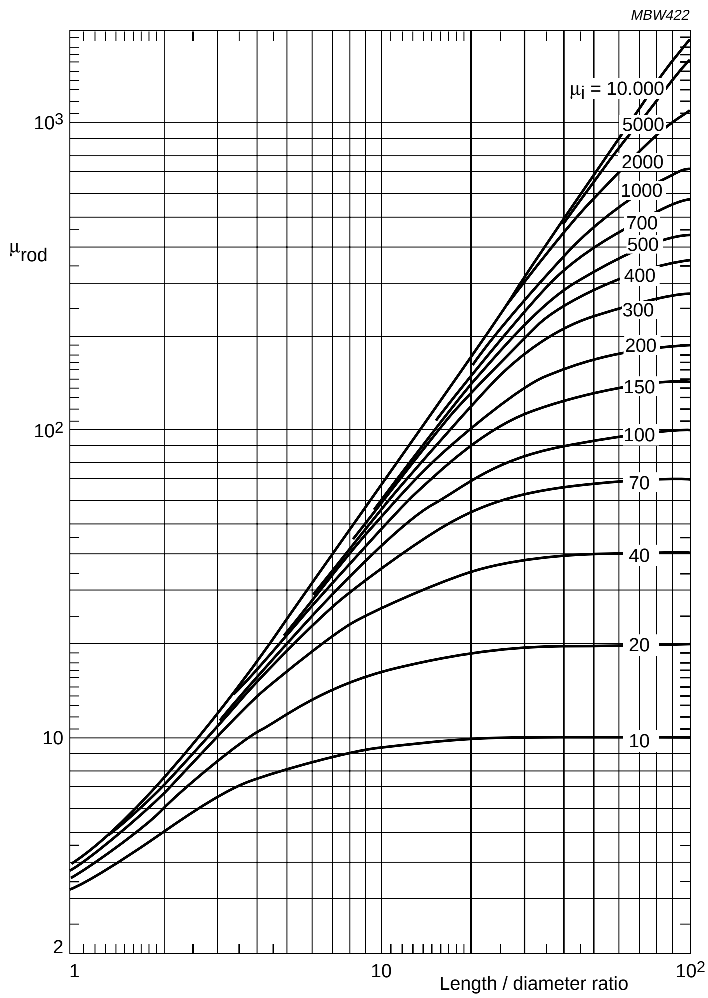


Fig.65 Rod permeability (μ_{rod}) as a function of length to diameter ratio with material permeability as a parameter.

Literature and reference materials

PHILIPS COMPONENTS APPLICATION LITERATURE

For the latest application literature, refer to the website at: **www.acm.components.philips.com**

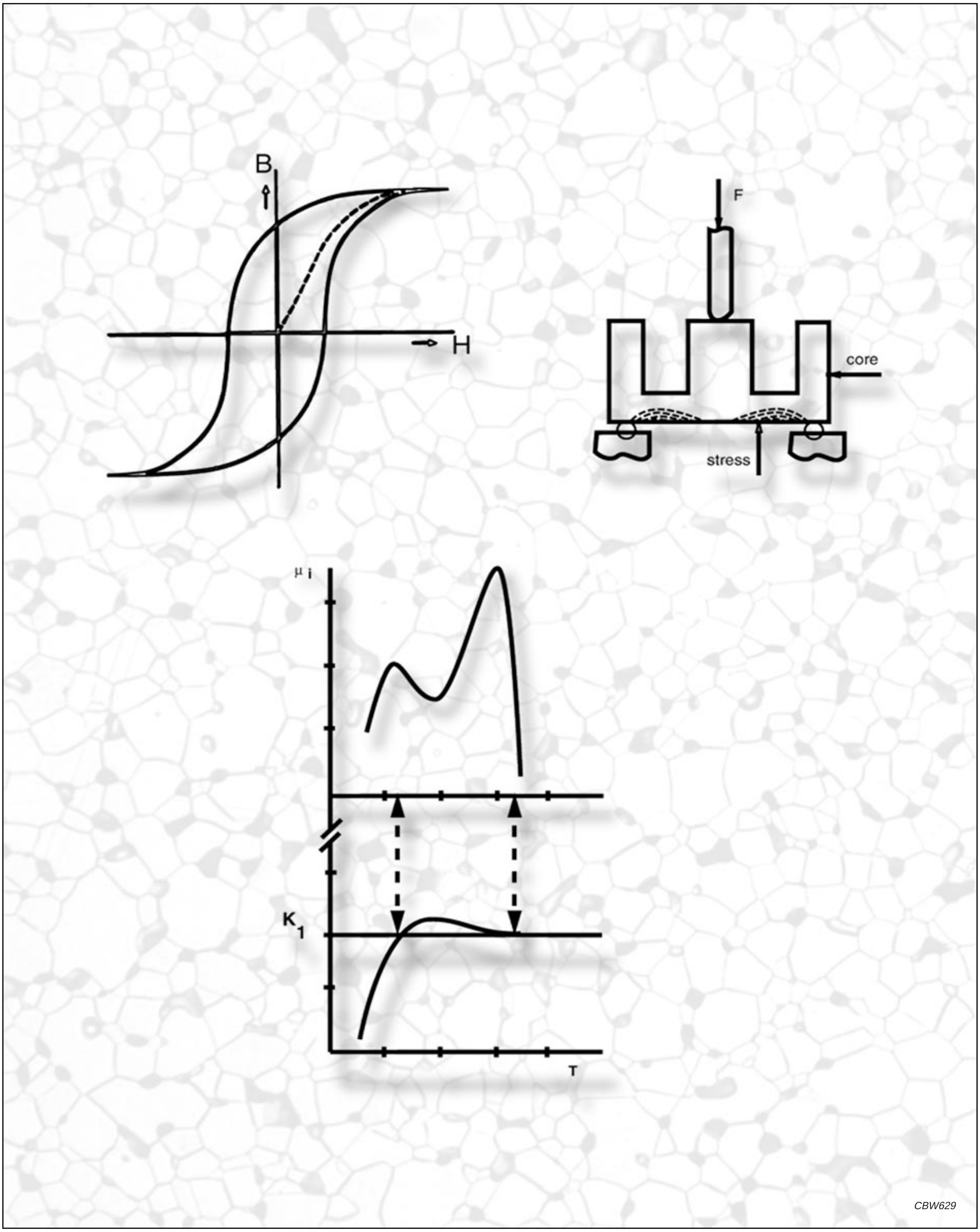
IEC STANDARDS ON SOFT FERRITES

133 (1985)	Dimensions for pot cores made of magnetic oxides and associated parts
205 (1966)	Calculation of the effective parameters of magnetic piece parts
205A (1968)	First supplement
205B (1974)	Second supplement
226 (1967)	Dimensions of cross cores (X cores) made of ferromagnetic oxides and associated parts
367	Cores for inductors and transformers for telecommunications
367-1 (1982)	Part 1: Measuring methods
367-2 (1974)	Part 2: Guides for the drafting of performance specifications
367-2A (1976)	First supplement
424 (1973)	Guide to the specification of limits for physical imperfections of parts made from magnetic oxides
431 (1983)	Dimensions of square cores (RM cores) made of magnetic oxides and associated parts
525 (1976)	Dimensions of toroids made of magnetic oxides or iron powder
647 (1979)	Dimensions for magnetic oxide cores intended for use in power supplies (EC cores)
1185 (1992)	Magnetic oxide cores (ETD cores) intended for use in power supply applications - Dimensions
1246 (1994)	Magnetic oxide cores (E cores) of rectangular cross-section and associated parts - Dimensions

Literature and reference materials

REFERENCE BOOKS ON MAGNETIC COMPONENT DESIGN

1.	Soft Ferrites, Properties and Applications	2nd Edition, E.C. Snelling, Butterworths Publishing, 80 Montvale Ave., Stoneham, MA 02180 Tel: (617) 928-2500
2.	Ferrites for Inductors and Transformers	C. Snelling & A. Giles, Research Studies Press, distributed by J. Wiley & Sons, 605 Third Ave., New York, NY 10016
3.	Transformer and Inductor Design Handbook	C. McLyman, Marcel Deckker, 207 Madison Ave., New York, NY10016
4.	Magnetic Core Selection for Transformers and Inductors	C. McLyman, Marcel Deckker, 207 Madison Ave., New York, Ny10016
5.	Handbook of Transformer Applications	W. Flanigan, McGraw Hill Publishing Co., 1221 Ave. of Americas, New York, NY 10020
6.	Transformers for Electronic Circuits	N. Grossner, McGraw Hill Publishing Co., 1221 Ave. of Americas, New York NY 10020
7.	Magnetic Components-Design and Applications	S. Smith Van Nostrand Reinhold Co., 135 West 50th St., New York, NY 10020
8.	Design Shortcuts and Procedures for Electronic Power Transformers and Inductors	Ordean Kiltie, O. Kiltie & Co. 2445 Fairfield, Ft. Wayne, IN 46807
9.	Switching and Linear Power Supply, Power Converter Design	A. Pressman, Hayden Book Co. Inc., 50 Essex St., Rochelle Park., NY 07662
10.	High Frequency Switching Power Supplies	G. Chrysiss, McGraw Hill Publishing Co, 1221 Ave. of Americas, NY
11.	Design of Solid State Power Supplies	3rd Edition, E. Hnatek, Van Nostrand Reinhold Co., New York, NW 10020
12.	Power Devices and Their Applications	Edited by: Dr. F. Lee & Dr. D. Chen, VPEC, Vol. III, 1990. Tel: (703) 231-4536
13.	Application of Magnetism	J.K. Watson, John Wiley & Sons, Inc. 605 Third Ave., New York, NY 10016
14.	Applied Electromagnetics	M.A. Plonus, McGraw Hill Publishing Co., 1221 Ave. of Americas, New York, NY 10020
15.	Transmission Line Transformers	J. Sevick, American Radio Relay League, 225 Main Street, Newington, CT 06111


B
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Soft Ferrites

Properties specified in this section are related to room temperature (25 °C) unless otherwise stated. They have been measured on sintered, non ground ring cores of dimensions ∅25 × ∅15 × 10 mm which are not subjected to external stresses.

Products generally comply with the material specification. However, deviations may occur due to shape size and grinding operations etc.

Specified product properties are given in the data sheets or product drawings.

MATERIAL GRADE SURVEY

Ferrite material grade survey

FERRITE MATERIAL	μ_i at 25 °C	B_{sat} (mT) at 25 °C (3000 A/m)	T_C (°C)	ρ (Ωm)	FERRITE TYPE	MAIN APPLICATION AREA	AVAILABLE CORE SHAPES
3D3	750	≈400	≥200	≈2	MnZn	telecom filters signal transformers pulse transformers delay lines	RM, P, PT, PTS, EP, E, ER, RM/I, RM/ILP, toroids
3B7	2300	≈450	≥170	≈1	MnZn		
3H3	2000	≈400	≥160	≈2	MnZn		
3E1	3800	≈400	≥125	≈1	MnZn		
3E4	4700	≈400	≥125	≈1	MnZn		
3E5	10000	≈400	≥125	≈0.5	MnZn		
3E55	10000	≈350	≥100	≈0.1	MnZn		
3E6	12000	≈400	≥130	≈0.1	MnZn		
3E7	15000	≈400	≥130	≈0.1	MnZn		
3E8	18000	≈350	≥100	≈0.1	MnZn		
3E25	6000	≈400	≥125	≈0.5	MnZn		
3E27	6000	≈400	≥150	≈0.5	MnZn		
3E28	4000	≈400	≥145	≈1	MnZn		
3C15	1800	≈500	≥190	≈1	MnZn	power conversion general purpose transformers	E, Planar E, EC, EFD, EP, ETD, ER, U, UR, I, RM/I, RM/ILP, P, P/I, PT, PTS, PQ, toroids
3C30	2100	≈500	≥240	≈2	MnZn		
3C34	2100	≈500	≥240	≈5	MnZn		
3C81	2700	≈450	≥210	≈1	MnZn		
3C91	3000	≈450	≥220	≈5	MnZn		
3C90	2300	≈450	≥220	≈5	MnZn		
3C94	2300	≈450	≥220	≈5	MnZn		
3C96	2000	≈500	≥240	≈5	MnZn		
3F3	2000	≈450	≥200	≈2	MnZn		
3F4	900	≈450	≥220	≈10	MnZn		
3F35	1400	≈500	≥240	≈10	MnZn		
4F1	80	≈350	≥260	$\approx 10^5$	NiZn		
3S1	4000	≈400	≥125	≈1	MnZn	EMI-suppression	EMI beads, beads on wire, SMD beads, Multilayer suppressors, common-mode chokes, cable shields, rods, ring cores (toroids), wideband chokes, U cores
3S3	350	≈350	≥225	$\approx 10^4$	MnZn		
3S4	1700	≈350	≥110	$\approx 10^3$	MnZn		
4S2	700	≈350	≥125	$\approx 10^5$	NiZn		
4S4	250	≈300	≥130	$\approx 10^5$	NiZn		
4S7	200	≈300	≥140	$\approx 10^5$	NiZn		
4C65	125	≈400	≥350	$\approx 10^5$	NiZn		
4A11	700	≈350	≥125	$\approx 10^5$	NiZn		
4A15	1200	≈350	≥125	$\approx 10^5$	NiZn		
3C11	4300	≈400	≥125	≈1	MnZn		
3E25	6000	≈400	≥125	≈0.5	MnZn		
3E26	7000	≈450	≥155	≈0.5	MnZn		

FERRITE MATERIAL	μ_i at 25 °C	B_{sat} (mT) at 25 °C (3000 A/m)	T_C (°C)	ρ (Ωm)	FERRITE TYPE	MAIN APPLICATION AREA	AVAILABLE CORE SHAPES
4E1	15	≈200	≥500	≈10^5	NiZn	tuning suppression	rods, tubes, wideband chokes
4D2	60	≈240	≥400	≈10^5	NiZn		
4B1	250	≈350	≥250	≈10^5	NiZn		
3B1	900	≈400	≥150	≈0.2	MnZn		
3R1	800	≈450	≥230	≈10^3	MnZn	magnetic regulators	toroids
4B3	300	≈400	≥250	≈10^5	NiZn	scientific particle accelerators	large toroids
4E2	25	≈350	≥400	≈10^5	NiZn		
4M2	140	≈350	≥200	≈10^5	NiZn		
8C11	900	≈350	≥125	≈10^5	NiZn		
8C12	1200	≈300	≥125	≈10^5	NiZn		

Iron powder material grade survey

IRON POWDER MATERIAL	μ_i at 25 °C	B_{sat} (mT) at 25 °C (3000 A/m)	MAXIMUM OPERATING TEMPERATURE (°C)	MAIN APPLICATION AREA	AVAILABLE CORE SHAPES
2P40	40	950	140	suppression	toroids
2P50	50	1000	140		
2P65	65	1150	140		
2P80	80	1400	140		
2P90	90	1600	140		

Typical mechanical and thermal properties

PROPERTY	MnZn FERRITE	NiZn FERRITE	UNIT
Young's modules	(90 to 150) × 10^3	(80 to 150) × 10^3	N/mm^2
Ultimate compressive strength	200 to 600	200 to 700	N/mm^2
Ultimate tensile strength	20 to 65	30 to 60	N/mm^2
Vickers hardness	600 to 700	800 to 900	N/mm^2
Linear expansion coefficient	(10 to 12) × 10^{-6}	(7 to 8) × 10^{-6}	K^{-1}
Specific heat	700 to 800	750	$Jkg^{-1} \times K^{-1}$
Heat conductivity	(3.5 to 5.0) × 10^{-3}	(3.5 to 5.0) × 10^{-3}	$Jmm^{-1}s^{-1} \times K^{-1}$