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.

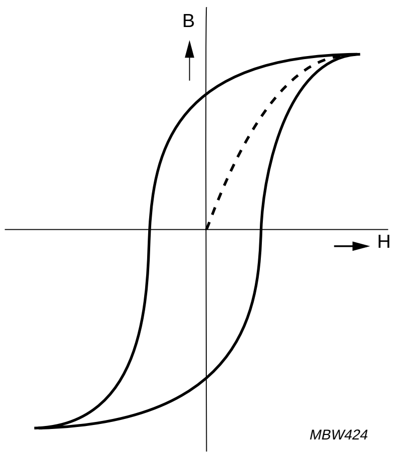

Fig.1 Hysteresis loop.

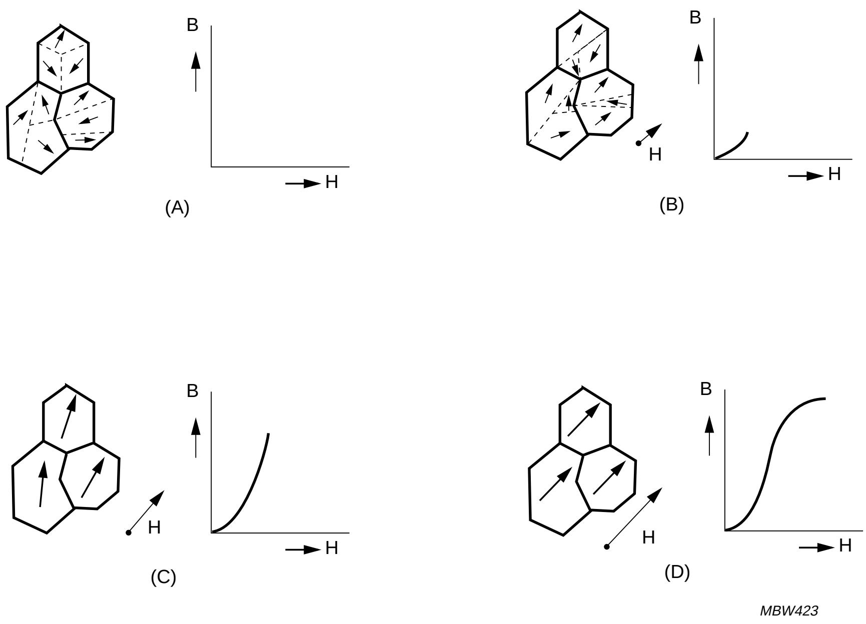

Fig.2 Alignment of domains.

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
U	voltage	V
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
THD/μ_a	Total Harmonic Distortion factor	dB
V_e	effective volume of core	mm^3
α_F	temperature factor of permeability	K^{-1}
$\tan\delta/\mu_i$	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_{rev}	reversible 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.

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

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.

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

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

$$\frac{B}{H} = \mu_0 \mu_r \tag{3}$$

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.

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

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.

$$\mu_e = \frac{\mu_i}{1 + \frac{G \times \mu_i}{l_e}} \tag{5}$$

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.

$$\mu_a = \frac{1}{\mu_0} \times \frac{\hat{B}}{\hat{H}} \tag{6}$$

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.

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

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.

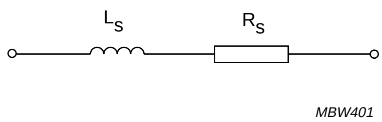

Fig.3 Series representation.

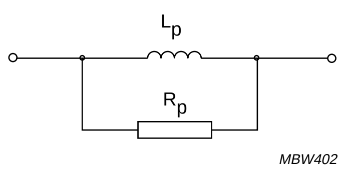

Fig.4 Parallel representation.

For series representation

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

and for parallel representation,

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

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

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

The phase shift caused by magnetic losses is given by:

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

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:

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

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:

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

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:

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

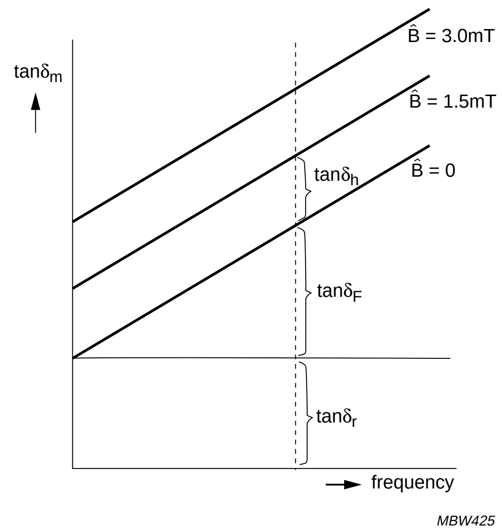

Fig.5 Magnetic losses $(\tan\delta_m)$ as a function of frequency.

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

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

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

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

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

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

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:

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

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:

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

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

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

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:

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

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

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

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

$$L = \frac{\mu_0 \times N^2}{\frac{1}{\mu_e} \times \Sigma \frac{l}{A}} \quad (23)$$

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

for sine wave:

$$\hat{B} = \frac{U\sqrt{2}}{\omega A_e N} = \frac{U}{\pi\sqrt{2} \times fNA_e} \quad (24)$$

for square wave:

$$\hat{B} = \frac{\hat{U}}{4 \times fNA_e} \quad (25)$$

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

$$\hat{H} = \frac{IN\sqrt{2}}{l_e} \quad (26)$$

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:

$$L = N^2 \times A_L \quad (27)$$

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

$$A_L = \frac{\mu_0 \mu_e}{\Sigma(l/A)} \quad (28)$$

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.

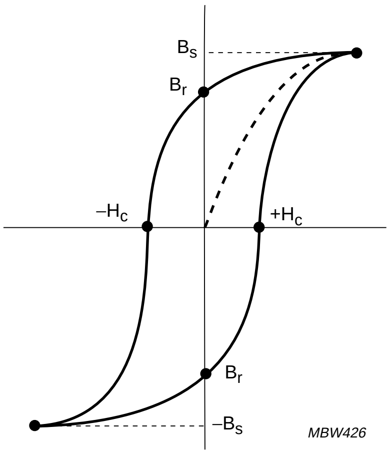

Fig.6 Typical BH curve showing points B_s, B_r and H_c.

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:

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

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

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

$$= \mu_e \times \alpha_F$$

So α_F is defined as:

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

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

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

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.

TOTAL HARMONIC DISTORTION (THD)

Harmonic distortion is generated when a sine wave magnetic field H, which is proportional to the current, induces a non-sinusoidal flux density B. This is due to a non linear relation between B and H in the ferrite core of a transformer. Consequently the induced output voltage, which is proportional to the flux density B, is also not a pure sine wave, but somewhat distorted. The periodic voltage signals can be decomposed by writing them as the sum of sine waves with frequencies equal to multiples of the fundamental frequency.
For signals without bias, the THD is defined as the ratio of: the square root of the sum of the quadratic amplitudes of the (uneven) higher harmonic voltages and, the amplitude of the fundamental frequency (V_1). It is often sufficient to consider only the strongly dominant third harmonic for the THD. In that case the definition of THD can be simplified to:

THD $\approx V_3 / V_1$ or 20 x ^{10}log (V_3 / V_1) [dB]

Introducing an airgap in a core set reduces the THD in the same way as it reduces temperature dependence and magnetic losses, which shows that the THD is not a pure material characteristic. It can be shown by calculation and measurement that THD/μ_{ae} is a real material characteristic. It is a function of flux density (B), frequency (f) and temperature (T), but not of the airgap length in a core set. THD/μ_{ae} is defined as the THD-factor, denoted as THD_F.

The term μ_{ae} stands for effective amplitude permeability of the ferrite material. It is a more general term than the effective permeability μ_e which is only defined for very low flux densities (< 0.25 mT).

Published data of this THD-factor (THD_F) as a function of frequency (f), flux density (B) and temperature (T) can directly be used to predict the THD in gapped core sets (THD_C) at the applicable operating conditions of f, B and T.

$$THD_C = THD_F + 20 \times {}^{10}\log(\mu_{ae}) \text{ [dB]} \quad (33)$$

THD MEASUREMENTS

Measured THD values as well as accuracies depend on the impedances in the measuring circuit used.

Fig.7 shows an equivalent THD test or measuring circuit. In Fig.8 a simplified equivalent circuit is shown with the generated (V_{F3}) and measured third harmonic voltage (V_{M3}).

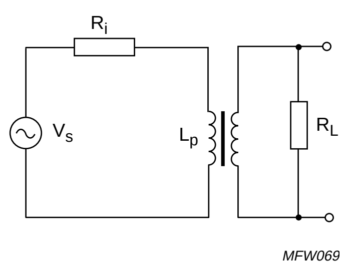

Fig.7 Equivalent test circuit for THD measurement.

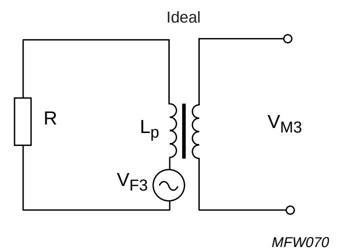

Fig.8 Equivalent test circuit for THD measurement.

The test circuit consists basically of a voltage source and a measuring device capable of measuring the third harmonic voltage or directly the THD. Both devices are often combined in one instrument like e.g. an audio analyzer which is represented by V_s in Fig.7.

R_i represents the total equivalent resistance in the primary circuit, which consists of the internal resistance of the voltage source, possibly in combination with other resistors in this part of the circuit. L_p is the inductance of the transformer under test connected to the load resistance R_L.

The generated third harmonic voltage V_{F3} will cause a current flow through the impedances R_i and R_L, resulting in a voltage drop. These impedances are combined to one equivalent resistance R as shown in Fig.8. This equivalent resistance can be calculated with:

$$R = \frac{R_i \times R_{L_P}}{R_i + R_{L_P}} \quad (34)$$

in which R_{Lp} is R_L referred to the primary side:

$$R_{L_P} = \left(\frac{N_p}{N_s}\right)^2 \times R_L \quad (35)$$

Hardly any voltage drop will occur when R is very high compared to the impedance $3\omega L_p$. In that case the measured third harmonic voltage V_{M3} would be equal to the real generated third harmonic V_{F3} multiplied by the transformation ratio N_s/N_p.

The measuring situation would be fully current driven. However in practical situations the resistance R will play a role and V_{F3} can be calculated with equation:

$$V_{F3} = V_{M3} \times \frac{N_p}{N_s} \times \sqrt{1 + (3\omega L_p / R)^2} \quad (36)$$

MEASUREMENT PRECAUTIONS

In general it is advised to check measuring conditions and the test circuit with impedances R and ωL_p in order to keep the circuit correction factor as low as possible. This avoids measuring in non-discriminating ranges (< –80 dB), which may lead to inaccurate or useless results. It is recommended to use low measuring frequencies, preferably < 25 kHz, for several reasons. At high frequencies it will often be difficult to reach the required flux level in the core of the transformer or inductor because of output voltage limitations. The real generated THD by the ferrite core ($THD_C \approx V_{F3}/V_{F1}$) can be related to the THD which is measured in the circuitry ($THD_M \approx V_{M3}/V_{M1}$) by knowing that $V_{F1} = V_{M1} \times (N_p/N_s)$. By using equation [36] this relation is as follows :

$$THD_M = \frac{1}{\sqrt{1 + (3\omega L_p / R)^2}} \times THD_C = CCF \times THD_C \quad (37)$$

The inverse square root term in equation [37] is the circuit correction factor (CCF). To get the measured THD in terms of the factor THD_F, equation [37] must be combined with [33] which gives in units of dB :

$$THD_M = THD_F + 20 \times {}^{10}\log(\mu_{ae} \times CCF) \quad (38)$$

To make use of equation [38] in practice, the following route can be followed :

The first step is to determine the voltage which will appear across the transformer. This is the voltage V_{Lp} across the inductance L_p in figure 8. If this value is not known from the (test) specification, it can be derived from the source voltage V_s. The relation between the source voltage V_s, the primary voltage V_{Lp} and the secondary voltage V_{RL} is given in equations [39] and [40] :

$$V_{Lp} = V_s \times \left| \frac{j\omega L_p \parallel R_{Lp}}{R_i + (j\omega L_p \parallel R_{Lp})} \right| \tag{39}$$

or

$$V_{Lp} = V_s \times \frac{1}{\sqrt{(1 + R_i/R_{Lp})^2 + (R_i/\omega L_p)^2}} \tag{40}$$

and $V_{RL} = (N_s/N_p) \times V_{Lp}$.

The second step is to use Faraday's law for induction to find the flux density B in the transformer. In case the voltage V_{LP} is a sinusoidal rms voltage, the relation to the peak flux density B_{peak} can be written as :

$$V_{Lp} = ½\sqrt{2} \, . \, \omega \, . \, N_1 \, . \, A_e \, . \, B_{peak} \tag{41}$$

The third step is to use the published curves on THD_F (as e.g. in fig. 4, 5 and 6 for 3E55) to determine the THD_F under the application conditions of f, B and T.

The last step is to use equation [38] to calculate the THD which will be measured and, to check whether this value is in line with the requirement (specification).

Time stability

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

$$D = \frac{\mu_1 - \mu_2}{\mu_1} \tag{42}$$

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

$$d = \frac{\mu_1 - \mu_2}{\mu_1 \times \log(t_2/t_1)} \tag{43}$$

Where t_1 and t_2 are time intervals after the disturbance. 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:

$$D_F = \frac{d}{\mu_i} = \frac{\mu_1 - \mu_2}{\mu_1^2 \times \log(t_2/t_1)} \tag{44}$$

Usually ferrite cores are magnetically conditioned by means of a saturating alternating field which is gradually reduced to zero. Measurements for our data sheets are taken 10 and 100 minutes after this disturbance.
The variability with time of a coil can now easily be predicted by:

$$\frac{L_1 - L_2}{L_1} = \mu_e \times D_F \tag{45}$$

L_1 and L_2 are values at 2 time intervals after a strong disturbance.

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 (3C94)

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 (4C65)

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