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Figure 1 shows the current radiation limits as defined by FCC Rules Part 15, for class A (industrial) and class B (mass-market) equipment.

Contrary to a decade ago when these regulations were first enforced and designing for EMI protection was often an afterthought rather than a forethought, a major portion of today's circuitry is incorporating EMI safeguards in the intial design. Many approaches can be used to comply with design or specification limits for EMI. Attention to basic circuit design, component layout, shielded enclosures and other use of shielding materials may be considered. For reducing or eliminating interference on printed circuit boards in wiring and cables, ferrite components have been used very successfully for decades. Figures 2 and 3 are photographs of a representative sampling of Fair-Rite suppressor cores.

There are basically three different ways to use ferrites as suppressors of unwanted signals, conducted or radiated. The first, and least common, is as actual shields where ferrite is used to isolate a conductor, component or circuit, from an environment of radiated stray electromagnetic fields. In both the second and the third application, ferrites are used as components to protect against conducted EMI. The second application is in conjunction with a capacitive element to create a low pass filter that is basically inductance-capacitance (LC) at low frequencies and dissipative at higher frequencies. The third and most common use will be addressed in this section of the catalog. In this case the cores are used alone on component leads or in board level circuitry either to prevent any parasitic oscillations or to attenuate unwanted signal pickup or transmissions which might travel along component leads or interconnecting wires, traces, or cables.

The Magnetics

Although permeability and quality factor play a role in the performance of a ferrite EMI suppressor, in virtually all instances, the frequency of usage puts these parameters beyond the point of meaningful definition, and as will be discussed later, the cores impedance is specified instead.

The permeability of a material is a complex parameter consisting of a real and an imaginary part. The real component represents the reactive portion and the imaginary component represents the losses. These may be expressed as series components (μ_s', μ_s'') or parallel components (μ_p', μ_p'').

The curves in Figure 4, are typical graphs of the permeability (μ_s') vs. frequency for various ferrite materials. The curves of Figure 5 are plots of loss factor ($\tan\delta/\mu_i$) in parts per million (ppm) vs. frequency for the same materials. The total loss tangent ($\tan\delta$) is the reciprocal of the Q factor and is a measure of the energy lost or incurred as the magnetization alternates. The real part of the permeability (μ_s') of these materials range from 40 to 10,000. In almost all cases μ_s' of the material first remains constant with frequency, then rises to a maximum value after which it falls off sharply. The loss component (μ_s'') rises to a peak as μ_s' falls. This is principally due to the ferrimagnetic resonance or spin precession resonance. It should be noted that the higher the permeability the lower the frequency at which this occurs. This was first observed by J. L. Snoek and given the relationship:

$$f_{res} = \gamma M_{sat}/3\pi(\mu_i - 1)\ \text{Hz} \qquad \text{eq.[1]}$$

where: f_{res} = frequency at which μ_s'' is maximum
γ = gyromagnetic ratio ~0.22 x 10^6 $A^{-1}m$
μ_i = initial permeability
M_{sat} ~250-350 x10^3 Am^{-1}

this same relationship can also be approximated by:

$$f_{max} \sim B_{sat}/\mu_i\ \text{MHz} \qquad \text{eq.[2]}$$

where: B_{sat} ~ 3500 -4500 gauss

Here f_{max} is the frequency limit of useful core inductance, not to be confused with maximum usable frequency for attenuation applications.

Figure 6 represents curves of impedance vs. frequency for the same ferrite materials in figures 4 and 5. Comparison of the three graphs shows that as the permeability decreases, the losses increase, and the impedance rises to a maximum, then either levels off or declines.

The impedance of a wound ferrite core is considered to be a series combination of the inductive reactance,($j\omega L_s$) and the equivalent loss resistance,(R_s), both of which are frequency dependent. At

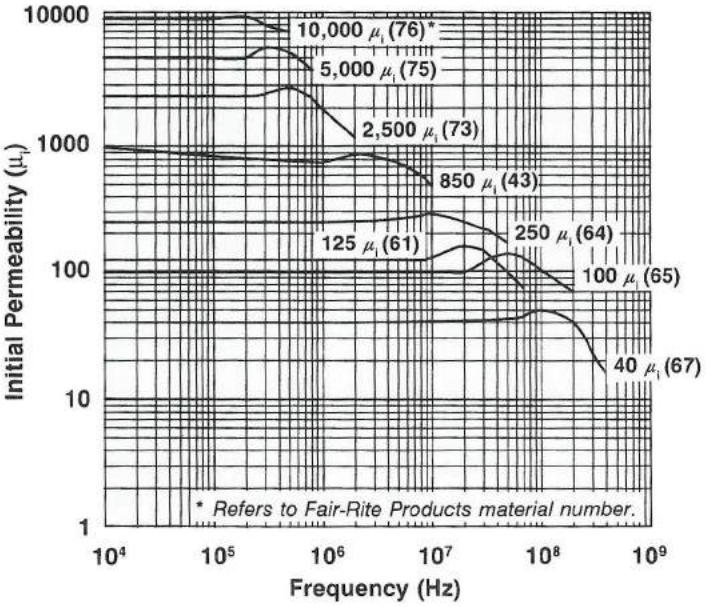


Figure 4 Initial Permeability vs. Frequency for various ferrite materials. Measured on Fair-Rite PN 26--000101 using HP 4275A and HP 4191A.

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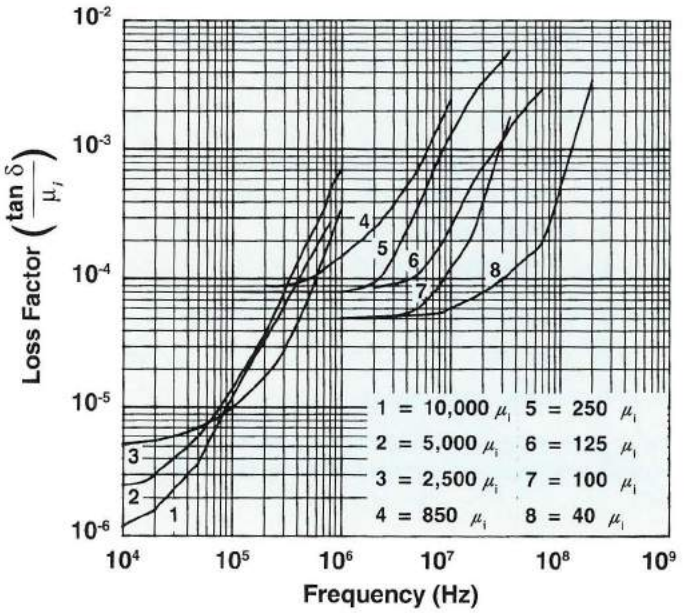


Figure 5 Loss Factor vs. Frequency for same the ferrite materials as in Figure 4. Measured on Fair-Rite PN 26--000101 using HP 4275A and HP 4191A.

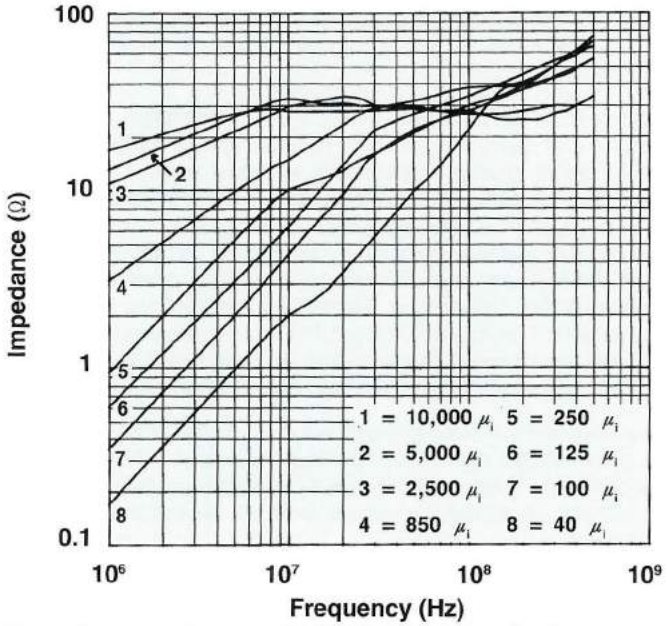


Figure 6 Impedance vs. Frequency for same ferrite materials as in Figures 4 and 5. Measured on Fair-Rite PN 26--000101 using HP 4191A.

low frequencies the impedance of the suppressor core is primarily the inductive reactance, which is a function of the material's permeability, and unwanted signals are mostly reflected. As the frequency increases, the inductive reactance decreases. Even so, the total impedance increases due to the increasing losses, and unwanted signals are absorbed. Figure 7 is the simplest representation of the equivalent circuit of a ferrite core, an inductive reactance in series with a resistor.

Figure 8 is the vector representation of the relationship between the components of the complex permeability and the components of the complex impedance of a ferrite core.

The following equations relate the series impedance and the complex permeability:

$$Z = j\omega L_s + R_s = j\omega L_o(\mu_s' - j\mu_s'') \text{ ohm} \qquad \text{eq.[3]}$$

so that

$$\omega L_s = \omega L_o \mu_s' \text{ ohm} \qquad \text{eq.[4]}$$

$$R_s = \omega L_o \mu_s'' \text{ ohm} \qquad \text{eq.[5]}$$

where: L_o = Air Core Inductance

$$\text{and } \tan\delta = \frac{R_s}{\omega L_s} = \frac{\mu_s''}{\mu_s'} = \frac{1}{Q} \qquad \text{eq.[6]}$$

$$\tan\Phi = \frac{\omega L_s}{R_s} = Q \qquad \text{eq.[7]}$$

$$Z = [(\omega L_s)^2 + R_s^2]^{1/2} \text{ ohm} \qquad \text{eq.[8]}$$

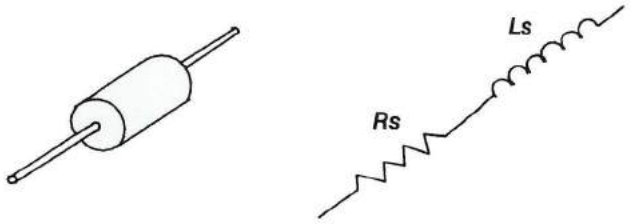


Figure 7 Figure 8 is the simplest representation of the equivalent circuit of a ferrite core, an inductor in series with a resistor.

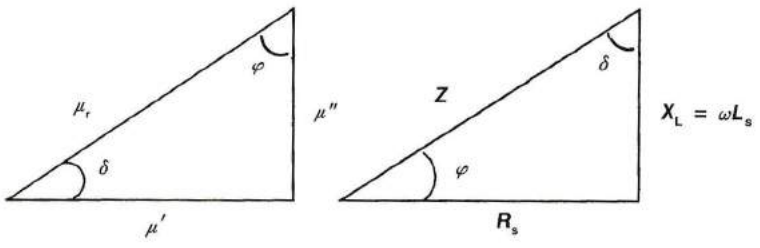


Figure 8 Vector relationship between real and imaginary components of impedance, complex permeability, series resistance and inductive reactance.

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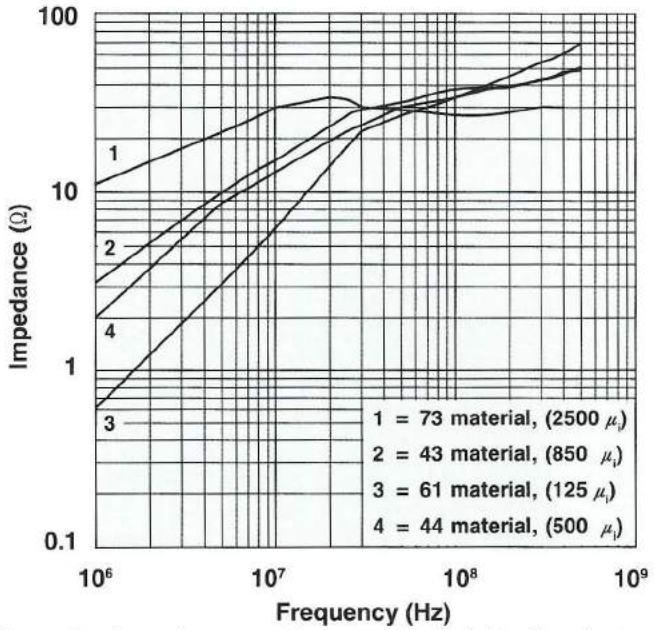


Figure 9 Impedance vs. Frequency for Fair-Rite Products four major suppressor materials. Measured on Fair-Rite PN 26---000101 bead using HP 4191A.

ohm-cm), manganese zinc material, which is the material of choice for suppressing signals of 30 MHz or less. A "wideband" suppressor material, 43, is a medium permeability, 850μ_i, higher resistivity (10^5 ohm-cm) nickel zinc material, designed for suppression of signals between 25 and 200 MHz. The third material, 61, is a low permeability, 125μ_i, high resistivity (10^8 ohm-cm) nickel zinc material that is recommended for use at frequencies above 200 MHz. The fourth material, 44, is similar in many respects to 43 material. It provides much higher volume resistivity (10^9 ohm-cm), but somewhat lower impedance over the same frequency range.

After comparing Figure 9 with Figure 6 it becomes obvious why manufacturers have limited production to several choice materials. Simply stated, why offer eight materials, when three or four will cover the same frequency range effectively?

Covering the broadest frequency range with several materials is economical for both the manufacturer and the consumer, but if the application frequency is specific, rather than over this range, the most effective material may not be one of those offered. As an example, if optimum attenuation must be achieved at between 1 and 10 MHz, a 5000 or 10,000 permeability material could be much more effective than the standard 2500 permeabilty. Figure 10 is the impedance vs. frequency curve for the same bead in three different materials. Comparison of the impedance curves for these three materials reveals that the impedance of the 10,000 permeability material is superior up to 6 MHz, at which point the 5000 permeability material becomes more effective and at approxi-

How to Choose the Right One

The designer or the retrofitter choosing the ferrite for optimum performance in combatting EMI, should know the following:

1. The frequency of the unwanted signals.
2. The source of the EMI.
3. Environmental conditions, temperature, and field strengths, ac and dc.
4. If high resistivity is required because of multiple turns, conductor pins in a connector filter plate or position in the circuit.
5. The circuit source and load impedances.
6. How much attenuation is required.
7. The allowable space on the board.

The Material

The first task is choosing the best ferrite material. This is based on the frequency or frequency range of the interference that must be suppressed.

Although most manufacturers of ferrite cores produce more than ten materials, only a select few have been offered for use in the suppression of EMI. The impedance vs. frequency curves for Fair-Rite's four major suppressor materials are shown in Figure 9. 73 material is a high permeability, 2500μ_i, low volume resistivity (10^2

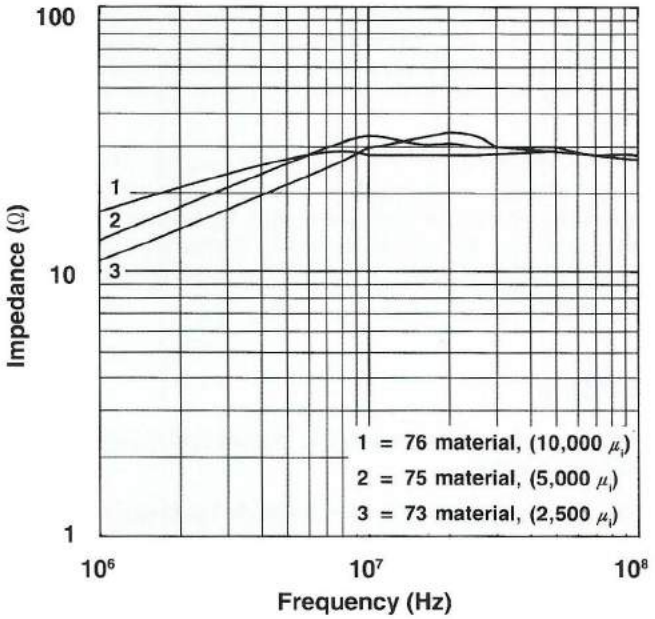


Figure 10 Impedance vs. Frequency for Fair-Rite Products three high permeability materials demonstrating the low frequency impedance. Measured on Fair-Rite PN 26--000101 bead using HP 4191A.

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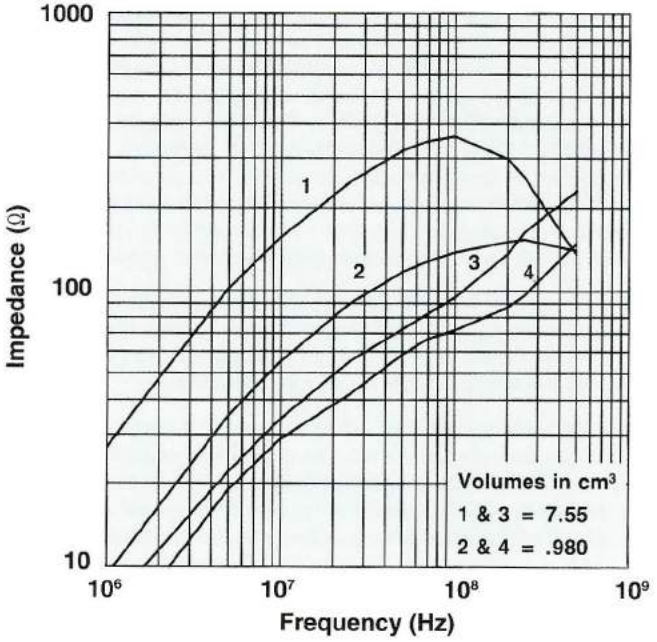


Figure 11 Impedance vs. Frequency for Fair-Rite shield beads, in the same material, two each with the same volume using HP 4191A.

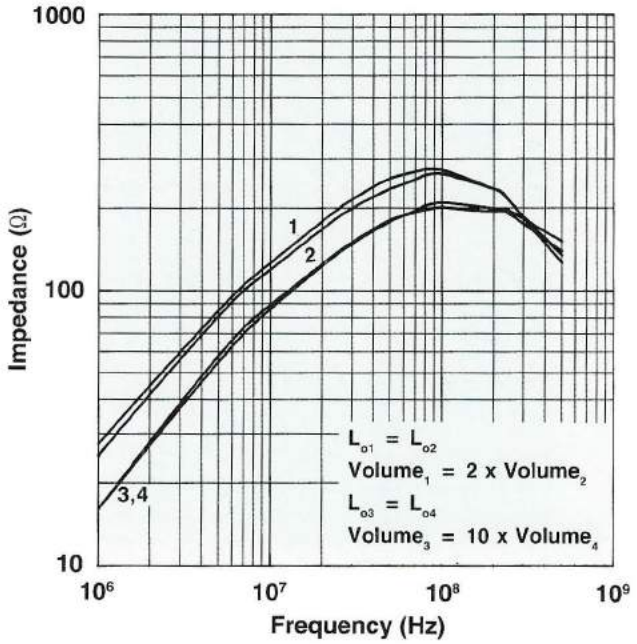


Figure 12 Impedance vs. Frequency for Fair-Rite shield beads, in the same material, all with different volumes. Two each with the same L_o using HP 4191A.

mately 14 MHz, the standard 2500 permeability material takes over. Therefore, if the requirement is for maximum attenuation at below 15 MHz and the ferrite manufacturer doesn't offer the higher permeability material in its standard suppressor line, try to get samples.

Generally speaking, the higher the permeability, the lower the optimum attenuation frequency, and conversely, the lower the permeability, the higher the attenuation frequency. This is frequency limited at both ends, since low frequency attenuation is reflective and high frequency attenuation is limited by the core and circuit resonance.

The Core

Once the material has been chosen, the size and shape of the core should be selected.

Figure 11 is a set of curves of impedance vs. frequency for four different beads, in the same material, two each with the same volume. Figure 12 is a set of curves of impedance vs. frequency for the same material, with different volumes. These curves show that increasing the cores volume does not guarantee an increase in impedance.

The same relationship that exists between the low frequency inductance factor (A_L) of a core and the core's initial permeability, can be used to approximate a core's impedance. This relationship is the core's air core inductance or Lo. It can be used as a proportion between two cores of the same material at the same frequency to approximate the impedance of one, knowing the impedance of the other. It is valid only below the resonant frequency, which is a function of the material and geometry of the core and/or winding/circuit conditions. The relationship is:

$$Z = K \times L_o$$

$$\text{Toroidal } L_o = .046 \times N^2 \times \text{Log}_{10}(OD/ID) \times Ht \times 10^{-8} \text{ H}$$

$$Z = .046\ K \times N^2 \times \text{Log}_{10}(OD/ID) \times Ht \times 10^{-8} \text{ ohm} \quad \text{eq.[9]}$$

All dimensions are in mm.

where: $K = Z/L_o$ (ohm/H) 10^8

N = Number of turns = 1

This is a very useful tool for maximizing the impedance of a core within the allowable space, after the best material has been chosen. It reconfirms the fact that in most cases greater impedance will be obtained by increasing the height of the core rather than the diameter for the same increase in volume.

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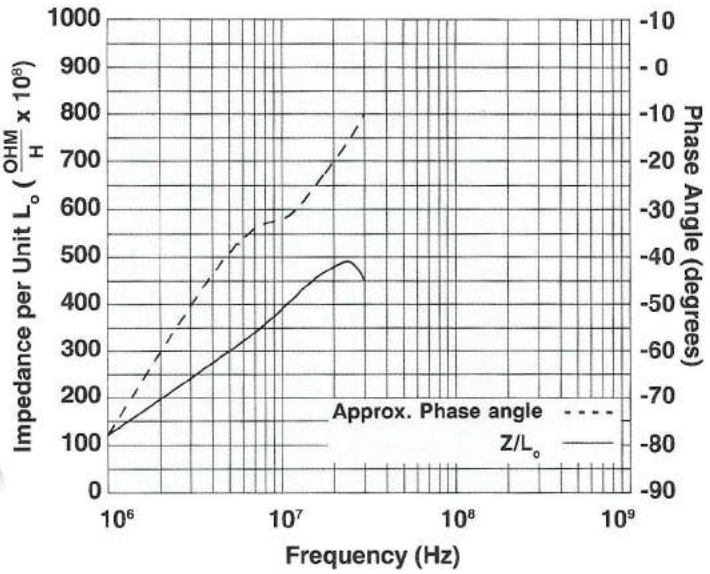


Figure 13 Nominal Impedance Per Unit L_o and Average Phase Angle vs. Frequency for Fair-Rite 73 material using HP 4191A.

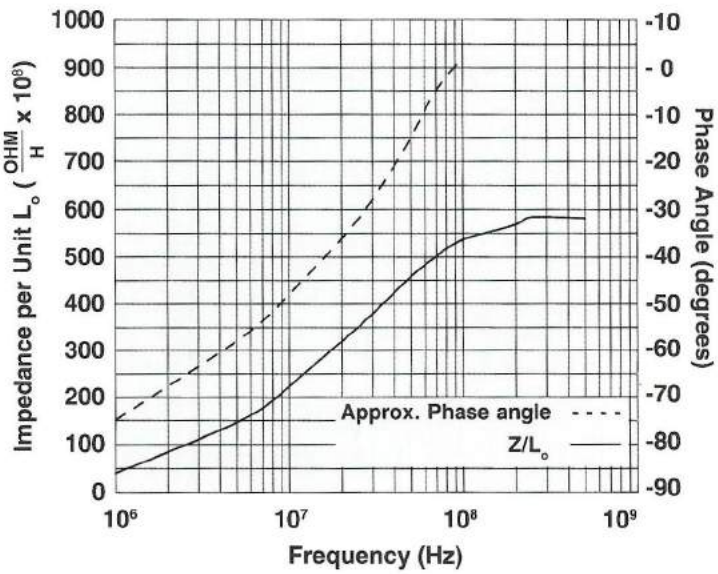


Figure 14 Nominal Impedance Per Unit L_o and Average Phase Angle vs. Frequency for Fair-Rite 43 material using HP 4191A.

Figures 13, 14, and 15 are graphs of the average impedance per unit Lo, and phase angle (Φ) for 73, 43, and 61 material respectively. To determine the impedance of a core using these graphs, the designer needs to know the physical dimensions of the core, the frequency of concern, and the material.

As an example:

A designer utilizing a ferrite core wants to guarantee maximum impedance for the frequency range of 25 MHz to 150 MHz. 43 material is chosen. The core must accept a #22 awg wire, and fit in a space of 10 mm by 5 mm. The inside diameter will be specified as .8 mm.

Using the graph of Figure 14, Z/Lo at 25 MHz for 43 material is 350 x 10^8 ohm/H. Solving for the estimated impedance by first using 10 mm for the outside diameter of the bead and 5 mm for the length...

$$Z = 350\ 10^8 x.046 \times \log_{10}(10/.8) \times 5 \times 10^{-8} = 88.3 \text{ ohm}$$

then using 5 mm for the outside diameter and 10 mm for the length

$$Z = 350\ 10^8 x.046 \times \log_{10}(5/.8) \times 10 \times 10^{-8} = 128.1 \text{ ohm}$$

In this instance, as in most, maximum impedance is achieved by using the smaller OD with the longer length. If the ID was larger, for instance 4 mm, the reverse would have been true. The same approach may be used for different materials, dimensions, and frequencies.

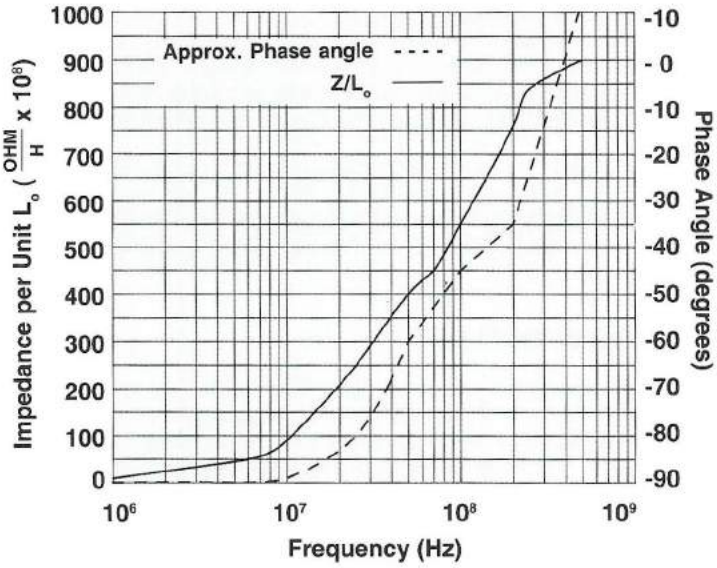


Figure 15 Nominal Impedance Per Unit L_o and Average Phase Angle vs. Frequency for Fair-Rite 61 material using HP 4191A.

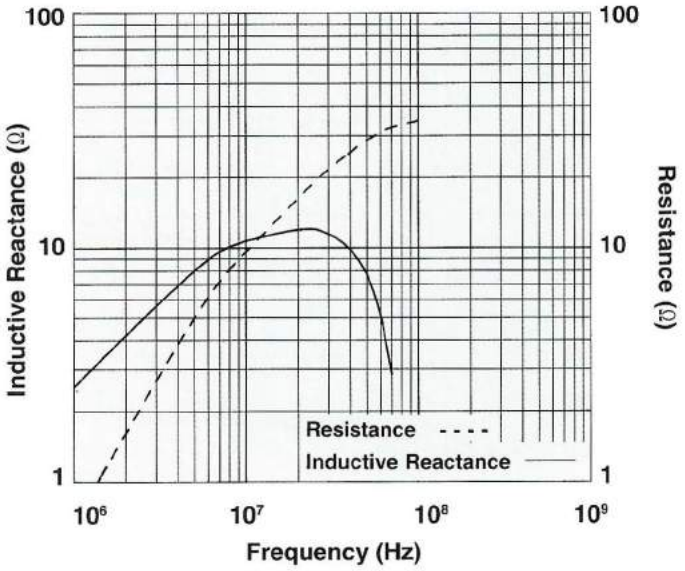


Figure 16 Graph of Inductive Reactance,(ωL_s), and Series Resistance, (R_s) for Fair-Rite PN 2643000101.

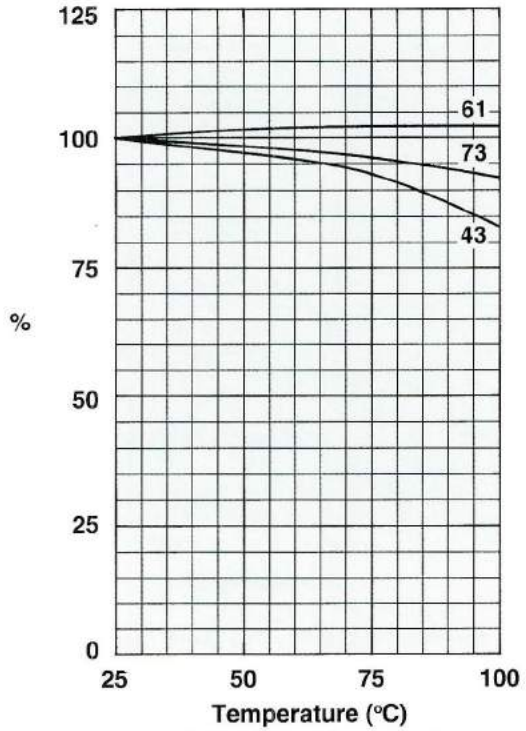


Figure 17 Percent of Original Impedance vs. Temperature at a frequency of 25 MHz for Fair-Rite's three major suppressor materials. Measured on 26--000801 using HP 4193A.

The previous discussion does assume that the core of choice is cylindrical. If the ferrite core being used is for flat ribbon, or bundled cable, or a multi-hole plate, the calculation for the L_o becomes more difficult, and fairly accurate figures for the cores path length and effective area must be obtained in order to calculate the air core inductance. In all cases though, the increase in impedance will be directly proportional to the height/length as long as the increase in height/length does not cause the core to be in resonance.

The phase angle plotted for each of the three materials is an average. The inductive reactance, and the series resistance can be calculated using these graphs and equations 7 and 8. Figure 16 is just such a graph for a Fair-Rite bead in 43 material.

The Environment

Ferrite's magnetic parameters can be affected by temperature, and field strength. Figures 17 and 18 are graphs of impedance vs. temperature for a Fair-Rite core in 73, 43, and 61 material at 25 and 100 MHz. The most temperature stable of these materials is 61, with a decrease in impedance of 8% at 100°C and 100 MHz as compared with over 25% for 43 at the same frequency and temperature. These curves may be used to adjust the specified room temperature impedance if desired attenuation is to be at elevated temperatures.

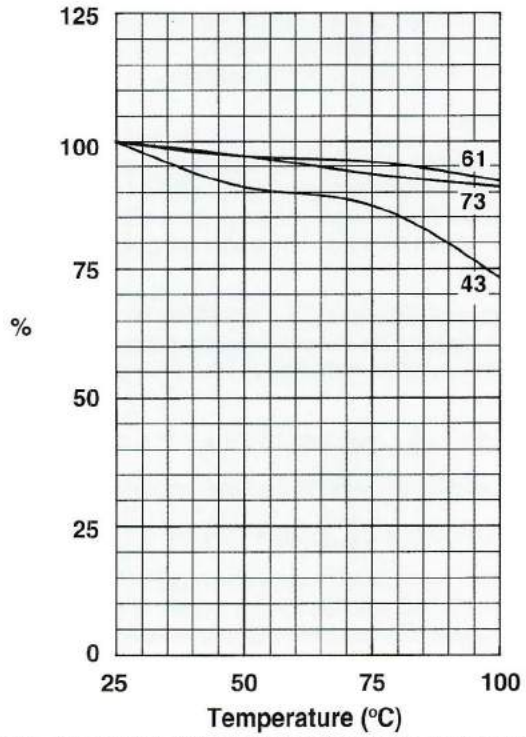


Figure 18 Percent of Original Impedance vs. Temperature at a frequency of 100 MHz for Fair-Rite's three major suppressor materials. Measured on 26--000801 using HP 4193A.

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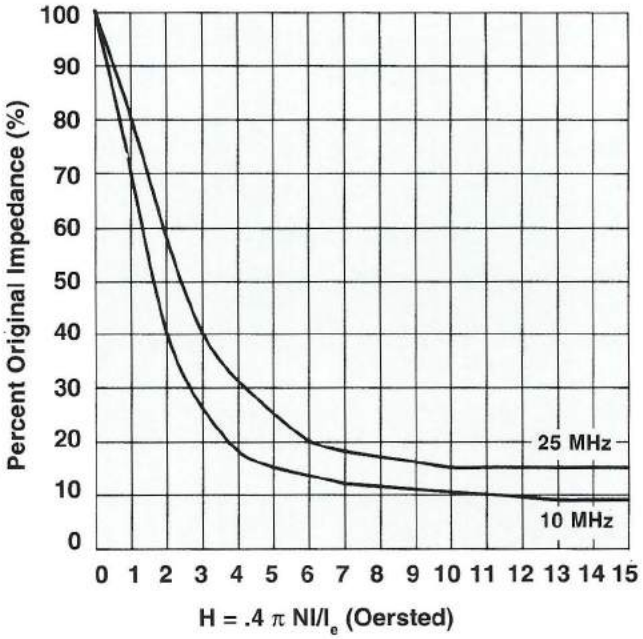


Figure 19 Impedance vs. Field Strength, H, in oersted, for 73 material, for two frequencies. Measurements made using HP 4193A.

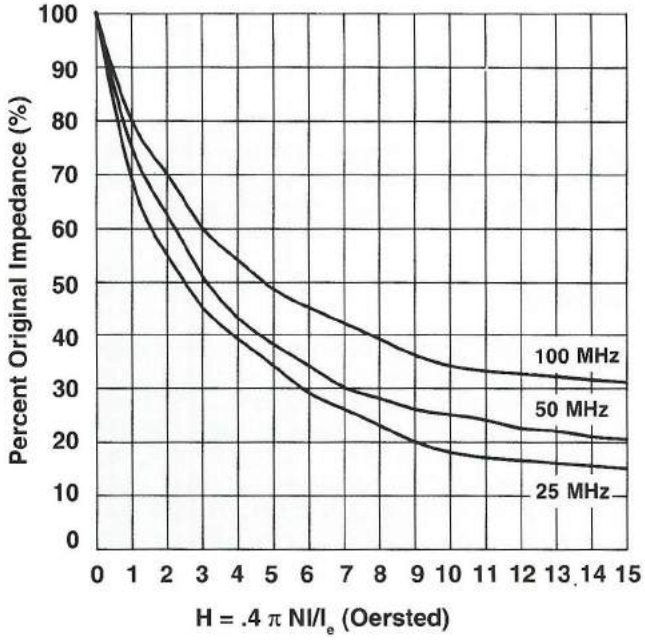


Figure 20 Impedance vs. Field Strength, H, in oersted, for 43 material for three frequencies. Measurements made using HP 4193A.

Biases

As in the case of temperature, dc and 50 or 60 Hz power current will also affect the same intrinsic ferrite characteristics which in turn will result in lowering of the impedance of the suppressor core.

Figure 20 illustrates the effect of dc bias on the initial impedance of 43 material. The curves depict the degradation in impedance vs. field strength, H in oersted, for frequencies of 25, 50 and 100 MHz.

In the design example cited earlier, the user chose a 43 material ferrite core with a 5 mm outside diameter, .8 mm inside diameter and a 10 mm length. Suppose this core will be exposed to 1 amp dc, and the resulting impedance must be determined.

In order to calculate the field strength, H, the path length of the core must be established. This is a lengthy procedure that involves first calculating the core constants, C_1 and C_2, then using these to find the path length.

These calculations reduce to:

path length $le = 2\pi\ [\log_e(r_2/r_1)]/(1/r_1 - 1/r_2) = .55$ cm eq.[10]

and $H = .4\ \pi\ NI/le = 2.28$ oersted eq.[11]

where: $r_1 = \frac{ID}{2}$ $r_2 = \frac{OD}{2}$

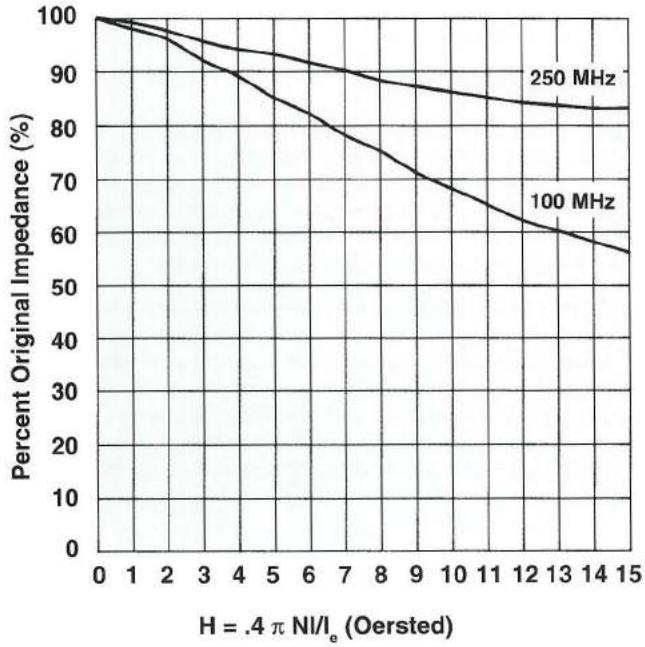


Figure 21 Impedance vs. Field Strength, H, in oersted, for 61 material, for two frequencies. Measurements made using HP 4193A (100 MHz) and HP 4191A (250 MHz).

The curves confirm that the greatest degradation in impedance occurs at the lower frequency, 25 MHz, and for 2.28 oersted the resulting impedance will be approximately 52% of the original, in this case, 66.6 ohm (.52 x 128.1 ohm).

Similar plots are made for 73 material in Figure 19, and 61 material in Figure 21. Two frequencies are plotted for 73 material, 10 and 25 MHz, and two for 61 material, 100 and 250 MHz.

Resistivity

The dc resistivities of ferrites can range from 10 ohm-cm to more than 10^9 ohm-cm. Generally, high permeability manganese zinc ferrites have low resistivities, below 10^2 ohm-cm, while nickel zinc ferrites have resistivities that range from 10^5 - 10^{12} ohm-cm. In most instances this parameter is measured using low voltage values, and since a ferrite's resistivity is voltage sensitive, these values may not be valid if the body of the core itself is exposed to high dc or ac values. There are ferrite materials available that have resistivity of 10^9 ohm-cm measured with 1000 volts. Fair-Rite products offers 44 material for such applications. If the frequency that must be attenuated is lower than 25 MHz, requiring a manganese zinc ferrite, coating the core with Parylene, epoxy or polyurethane varnish or using insulated wires, may be the only solution.

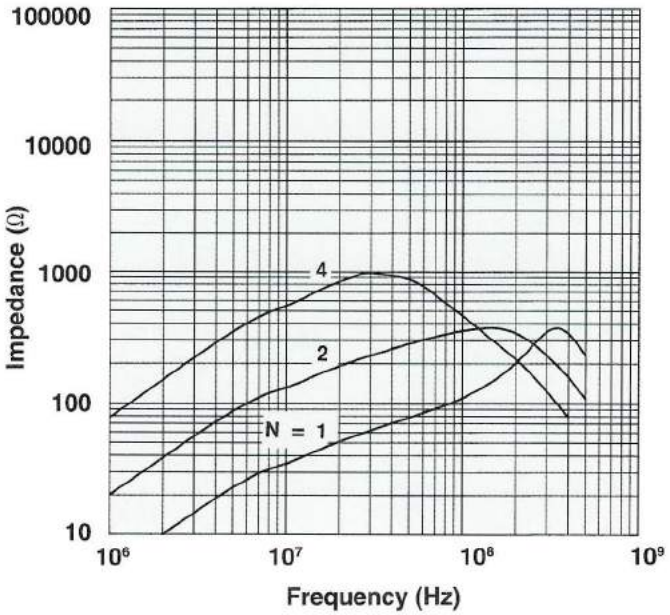


Figure 22 Impedance vs. Frequency for Fair-Rite large bead, PN 2643803802, with one, two and four turns.

Helpful Hints.....

Increasing Impedance

Impedance can be increased significantly by adding turns to a bead or a coil on a slug. Figure 22 is a graph of impedance vs. frequency for a large toroidal core with one, two and four turns. The impedance will increase in direct proportion to the turns squared, but the frequency at which the maximum impedance is reached is lowered due to the additional capacitive effects.

Figure 23 is a graph of different plots of impedance for the same size slug, same number of turns, in different materials. The impedances gained by this type of configuration can be significant. Again there is a tradeoff; increasing the permeability lowers the resonant frequency at which the impedance becomes maximum, therefore narrowing the effective frequency band. The lower the initial permeability of a material, the higher the frequency where this occurs.

Increasing DC Handling

Introducing an airgap in the cores path length can decrease the degrading effects of dc bias. The larger the gap the less effect the bias will have on the impedance. Gaps vary from finishing gaps in mated parts to the gaps found in open magnetic circuits, for example, a slug. Figure 24 shows the effect of dc bias on a flat ribbon cable core with different currents and several size airgaps. Note that the higher the frequency the less the effect of the gap and the current.

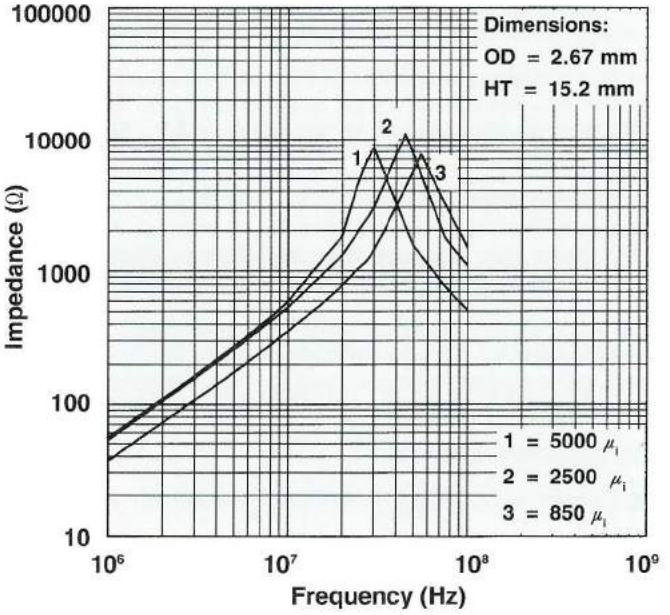


Figure 23 Impedance vs. Frequency for slug type core in a coil varying material.

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Use of Ferrites in Broadband Transformers

Introduction

Most of the magnetic information in this catalog is data obtained from cores wound with a single multi-turn winding which forms an inductor. When a second winding is added on the core, the inductor becomes a transformer. Depending on the requirements, transformers can be designed to provide dc isolation, impedance matching and specific current or voltage ratios. Transformers designed for power, broadband, pulse, or impedance matching can often be used over a broad frequency spectrum.

In many transformer designs ferrites are used as the core material. This article will address the properties of ferrite materials and core geometries which are of concern in the design of low power broadband transformers.

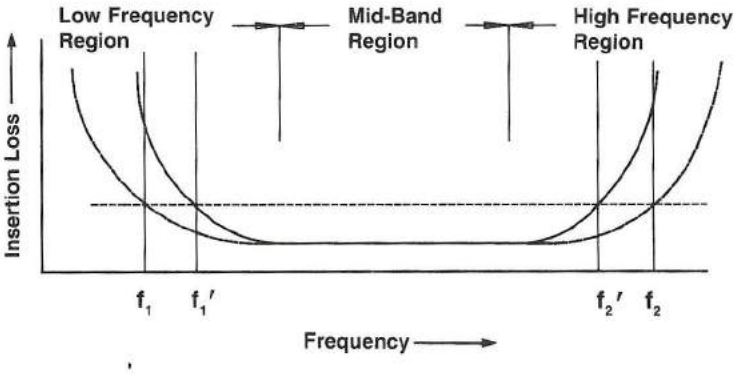


Figure 1 Typical Characteristic Curve of Insertion Loss vs. Frequency for a broadband transformer.

Brief Theory

Broadband transformers are wound magnetic devices that are designed to transfer energy over a wide frequency range. Most applications for broadband transformers are in telecommunication equipment where they are extensively used at low power levels.

Figure 1 shows a typical performance curve of insertion loss as a function of frequency for a broadband transformer. The bandwidth of a broadband transformer is the frequency difference between f_2 and f_1, or between f_2' and f_1', and is a function of the specified insertion loss and the transformer roll-off characteristics.

It can be seen that the bandwidth is narrower for transformers with a steep roll-off (f_2'-f_1') than for those with a more gradual roll-off (f_2-f_1). Also in figure 1, the three frequency regions are identified.

The cutoff frequencies are determined by the requirements of the individual broadband transformer design. Therefore, f_1 can be greater than 10 MHz or less than 300 Hz. Bandwidths also can vary from a few hundred hertz to hundreds of MHz. A typical broadband transformer design will specify for the mid frequency range a maximum insertion loss and for the cutoff frequencies, f_1 and f_2 maximum allowable losses. Figure 2 is a schematic diagram of the lumped element equivalent circuit of a transformer, separating the circuit into an ideal transformer, its components and equivalent parasitic resistances and reactances. The secondary components, parasitics and the load resistance have been transferred to the primary side and are identified with a prime.

To simplify this circuit, the primary and secondary circuit elements have been combined and the equivalent reduced circuit is as shown in Figure 3. The physical significance of the parameters are listed below the equivalent circuits. In the low frequency region the roll-off in transmission characteristic is due a lowering of the shunt impedance. The shunt impedance decreases when the frequency is reduced, which results in the increased level of attenuation. The impedance is mainly a function of the primary

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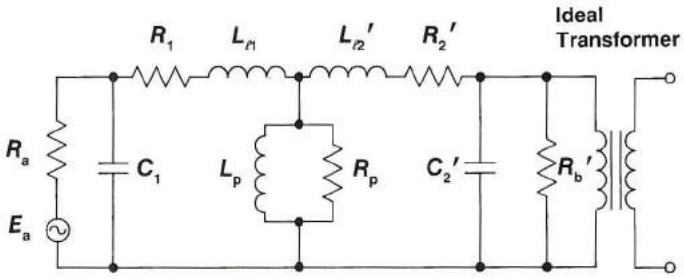


Figure 2 Lumped equivalent circuit of a transformer.

E_a = source EMF
R_a = source resistance
C_1 = primary winding capacitance
R_1 = resistance of primary winding
L_{f1} = primary leakage inductance
L_p = open circuit inductance of primary winding
R_p = shunt resistance that represents loss in core

Secondary parameters reflected to the primary side.

C_2' = secondary winding capacitance
R_2' = resistance of secondary winding
L_{f2}' = secondary leakage inductance
R_b' = load resistance

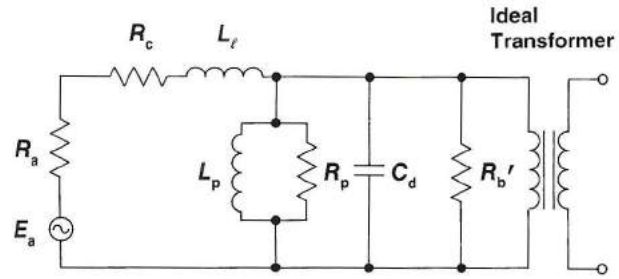


Figure 3 Simplified equivalent transformer circuit.

$C_d = C_1 + C_2'$
$R_c = R_1 + R_2'$
$L_f = L_{f1} + L_{f2}'$

For other circuit parameters see Figure 2.

reactance X_{LP}, with a negligible contribution of the equivalent shunt loss resistance R_p. The insertion loss may therefore be expressed in terms of the shunt inductance:

$$A_i = 10 \log_{10} \left(1 + \left(\frac{R}{\omega L_p} \right)^2 \right) \text{ dB}$$

Where $R = R_a \times R_b' / R_a + R_b'$

For most ferrite broadband transformer designs, the only elements that are likely to effect the transmission at the mid-band frequency range are the winding resistances. The insertion loss for the mid-band frequency region due to the winding resistance may be expressed as:

$$A_i = 20 \log_{10} \left(1 + \frac{R_c}{R_a + R_b'} \right) \text{ dB}$$

Where $R_c = R_1 + R_2'$

In the higher frequency region the transmission characteristics are mainly a function of the leakage inductance or the shunt capacitance. It is often necessary to consider the effect of both of these reactances, depending upon the circuit impedance. In a low impedance circuit the high frequency droop due to leakage inductance is:

$$A_i = 10 \log_{10} \left(1 + \left(\frac{\omega L_f}{R_a + R_b'} \right)^2 \right) \text{ dB}$$

This high frequency droop in a high impedance circuit, due to the shunt capacitance, is as follows:

$$A_i = 10 \log_{10} \left(1 + \left(\omega CR \right)^2 \right) \text{ dB}$$

Reviewing the insertion loss characteristics for the three frequency regions, it can be concluded that the selection of ferrite material and core shape should result in a transformer design that yields the highest inductance per turn at the low frequency cutoff f_1. This will result in the required shunt inductance for the low frequency region with the least number of turns. The low number of turns are desirable for low insertion loss at the mid-band region and also for low winding parasitics needed for good response at the high frequency cutoff f_2.

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Low and Medium Frequency Broadband Transformers

For broadband transformer applications the optimum ferrite is the material that has the highest initial permeability at the lower cutoff frequency f_1. Manganese zinc ferrites, such as Fair-Rite 77 or 73 material, are very suitable for low and medium frequency broadband transformers designs. As stated before, the transformer parameter that is most critical is the shunt reactance (ωL), which will increase with frequency as long as the material permeability is constant or diminishing at a rate less than the increase in frequency. This holds true even if a transformer is designed using a manganese zinc ferrite where f_1 is at the higher end of the flat portion of the permeability vs. frequency curve. Although the whole bandpass lies in the area where the initial permeability is decreasing, yet the bandpass characteristics will be virtually unaffected. For broadband transformers that use a manganese zinc ferrite material the core geometry should be such as to minimize the R_{dc}/L ratio. In other words, the ratio of dc resistance to the inductance for a single turn should be a minimum. The range of pot cores, standardized by the International Electrotechnical Commission in document IEC 133, has been designed for this minimum R_{dc}/L ratio.* Other core shapes such as the EP cores and PQ cores can also be used in the design of these broadband transformers. Often the final core selection will also be influenced by such considerations as ease of winding, terminating and other mechanical design constraints of the transformer.

Broadband Transformers with a Superimposed Static Field

In transformer designs that have a superimposed direct current, gapped cores can be employed to overcome the decrease in the shunt inductance. Hanna curves can be used to aid in the design of inductive devices that carry a direct current. For more information see section "The Effect of Direct Current on the Inductance of a Ferrite Core" on page 26.

High Frequency Broadband Transformers.

Although there is no clear division between the frequency regions, for this article it is assumed that the high frequency broadband transformer designs use nickel zinc ferrites as the preferred core material. This will typically occur for transformer designs where the bandpass lies wholly above 500 kHz. At these higher operating frequencies it becomes more important to consider the complex magnetic parameters of the core material, rather than use the simple core constants, such as A_L, recommended for low frequency designs.

Another important consideration is that high frequency transformers are generally used in low impedance circuits, which means that these designs require low shunt impedances. This can often be accomplished with a few turns, hence winding resistances are no longer an issue, and the design concept of minimizing R_{dc}/L is no longer required. The design will instead become focused on core shape and material for the required shunt impedance at f_1 along with reducing leakage inductance of the winding. Since the material characteristics permeability and losses affect the shunt impedance these parameters need to be considered in high frequency broadband transformer designs. Figures 4, 5 and 6 are typical curves of impedance Z, equivalent parallel reactance X_p and equivalent parallel loss resistance R_p as a function of frequency. They are measured on the same multi-aperture core 28--002402, in 73, 43, 61 and 65 material, wound with a single turn through both holes. For high frequency broadband transformers the toroidal core shape becomes an attractive core geometry. The few turns that are often required can easily be wound on the toroid. However, windings that require only a few turns may give rise to problems in obtaining the desired impedance ratios. To minimize leakage inductance it is suggested that the primary and secondary windings be tightly coupled and where possible a bifilar winding be used.

An improvement in core performance over toroids can be obtained by the use of multi-aperture cores, which can be considered as two toroidal cores side by side. This core shape has a lower single turn winding length than the equivalent toroidal core with the same core constant C_1, and will result in a wider bandwidth of the transformer design. Many broadband transformers have been designed utilizing nickel zinc ferrite toroids with good results. If bandwidth requirements cannot be met using toroids, multi-aperture nickel zinc cores should be considered.

The multi-aperture cores listed in this catalog on page 86, are available in the nickel zinc ferrite materials 65, 61 and 43 as well in the manganese zinc ferrite 73 material. Intrinsic material and part performance data are shown in Figures 7 through 32 for all these ferrite materials.

***Footnote:** *E. C. Snelling, Soft Ferrites, Properties and Applications. 2nd Edition 1988, Butterworths Publishing, Stoneham, MA.*

Technical Information

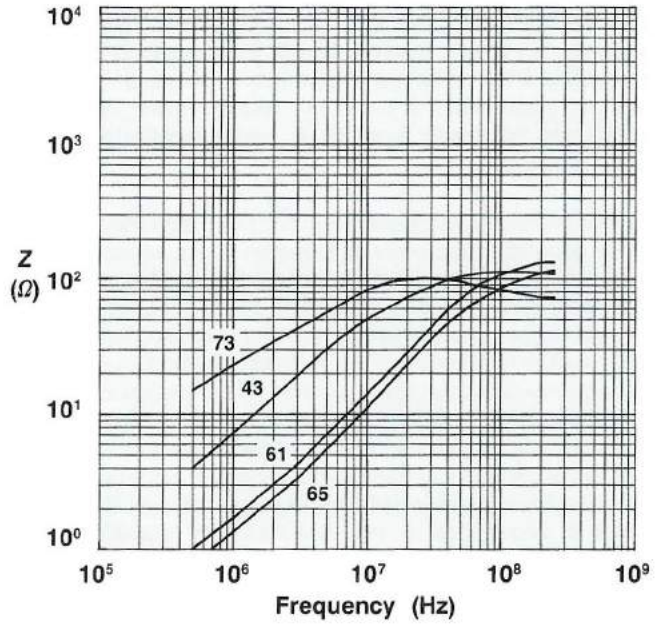


Figure 4 Impedance vs. Frequency for part number 28 - - 002402 in several materials.

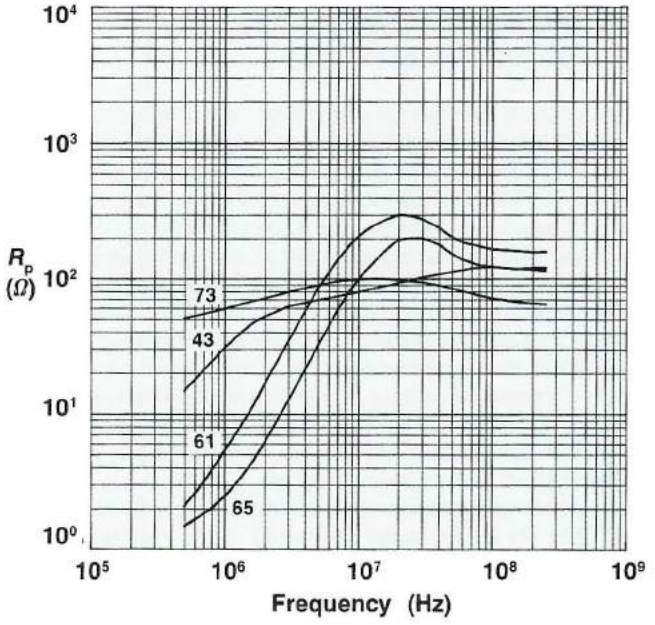


Figure 6 Parallel Resistance vs. Frequency for part number 28 - - 002402 in several materials.

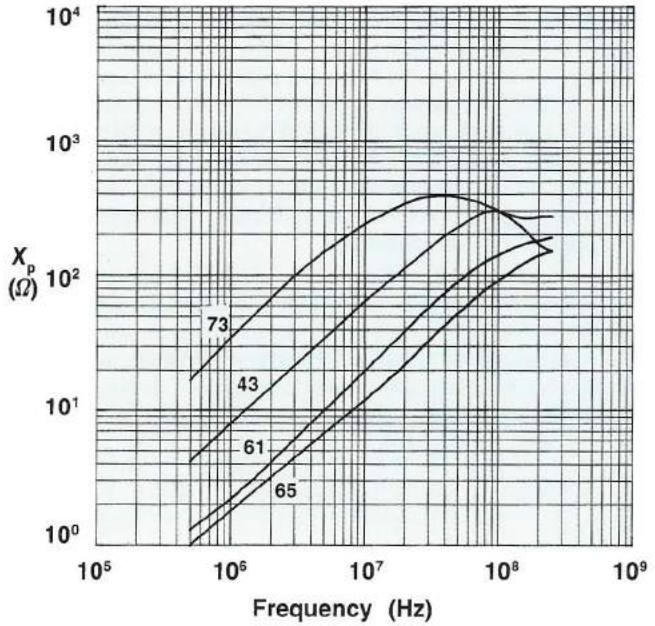


Figure 5 Parallel Reactance vs. Frequency for part number 28 - - 002402 in several materials.

Summary

The low cutoff frequency f_1 is the single most important factor in the ferrite material selection. The material with the highest initial permeability at f_1 is the recommended choice.

Manganese zinc ferrites, 73 and 77, can be used to a cutoff frequency f_1 of 500 kHz. Above this frequency use a nickel zinc ferrite, again depending upon the frequency f_1, select 65, 61 or 43 material.

For low and medium frequency transformers the optimum core shape should provide the lowest dc resistance per unit of inductance. If there is a superimposed dc present the use of gapped cores and Hanna curves is suggested. For high frequency designs, use nickel zinc ferrite. The toroidal and multi-aperture cores are the recommended core configurations.

The number of turns should be kept to a minimum to reduce leakage inductance and self-capacitance of the windings. Wind primary and secondary windings tightly coupled or as bifilar windings to lower leakage inductance.

The Bead, Balun and Broadband Kit II, part number 0199000011, contains a variety of components suited for broadband transformer design evaluations, see page 84.

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Typical Performance Data

Material 65

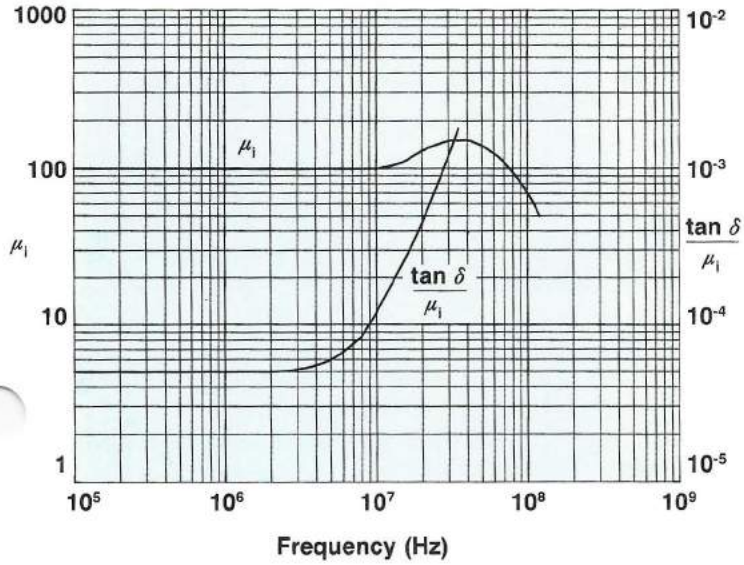


Figure 7 Initial Permeability and Loss Factor vs. Frequency.

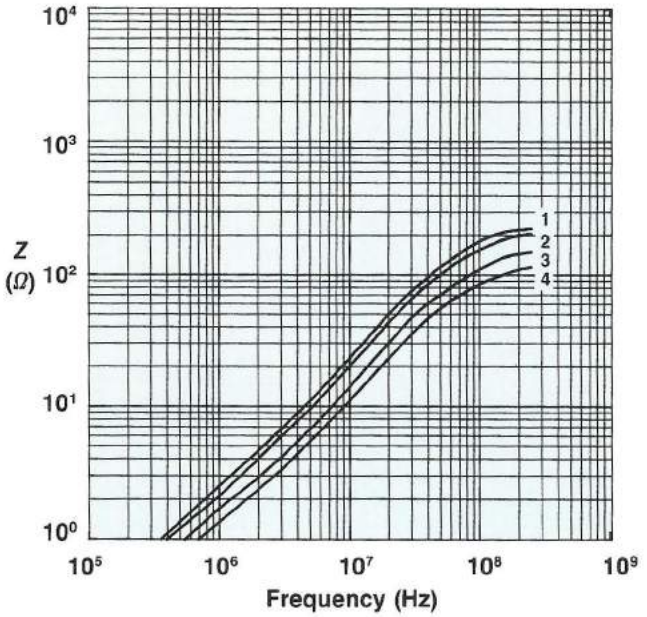


Figure 8 Impedance vs. Frequency of multi-aperture cores in 65 material.

1 2865003102 3 2865003002
2 2865000202 4 2865002402

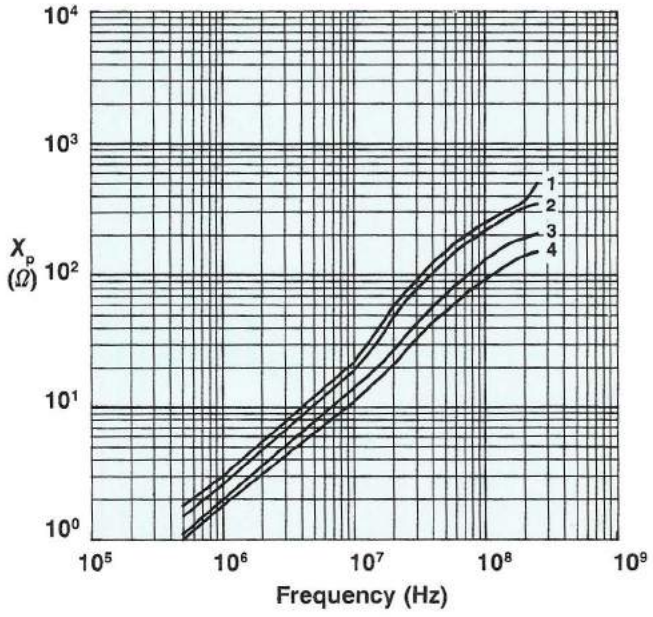


Figure 9 Parallel Reactance vs. Frequency of multi-aperture cores in 65 material.

1 2865003102 3 2865003002
2 2865000202 4 2865002402

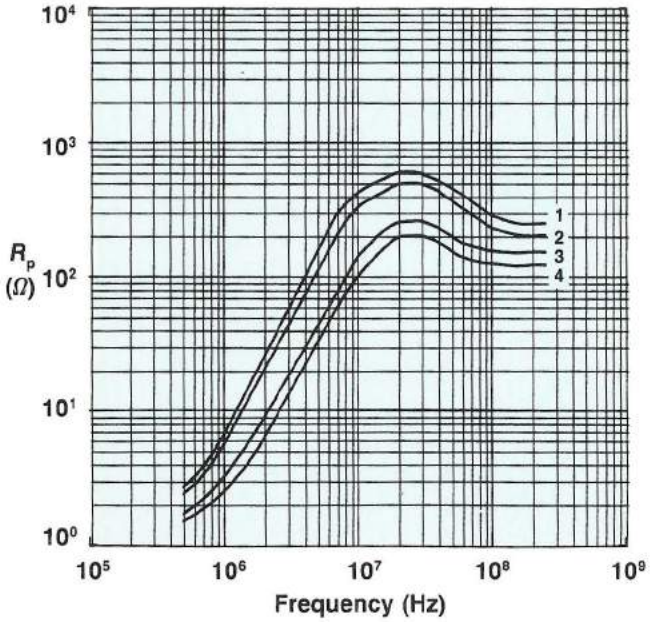


Figure 10 Parallel Resistance vs. Frequency of multi-aperture cores in 65 material.

1 2865003102 3 2865003002
2 2865000202 4 2865002402

Typical Performance Data

Material 61

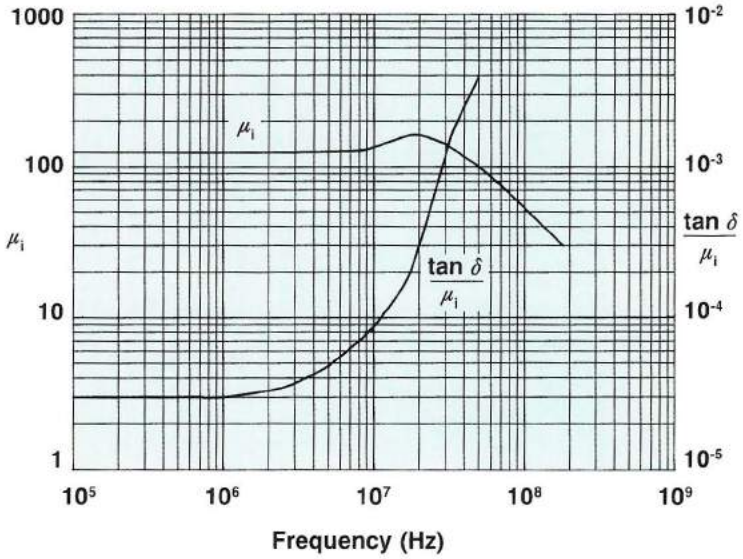


Figure 11 Initial Permeability and Loss Factor vs. Frequency.

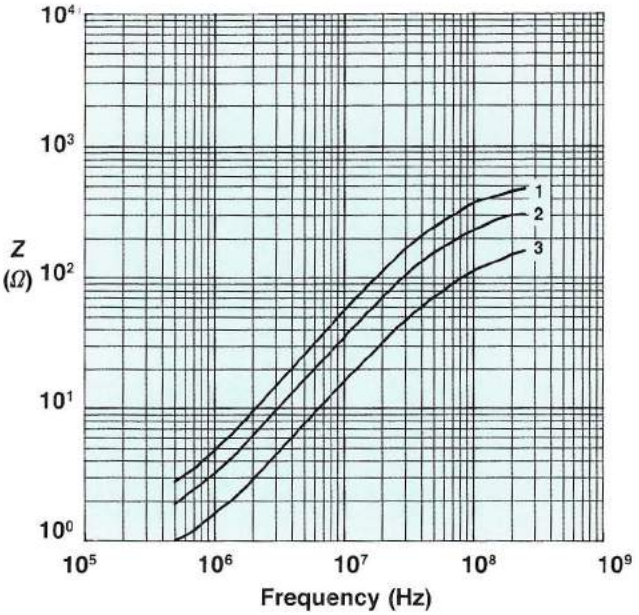


Figure 12 Impedance vs. Frequency of multi-aperture cores in 61 material.

1 2861006802
2 2861001702
3 2861001802

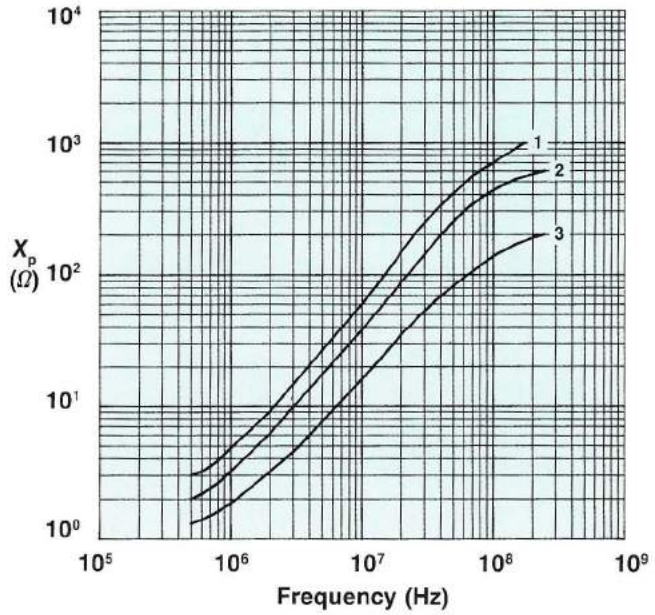


Figure 13 Parallel Reactance vs. Frequency of multi-aperture cores in 61 material.

1 2861006802
2 2861001702
3 2861001802

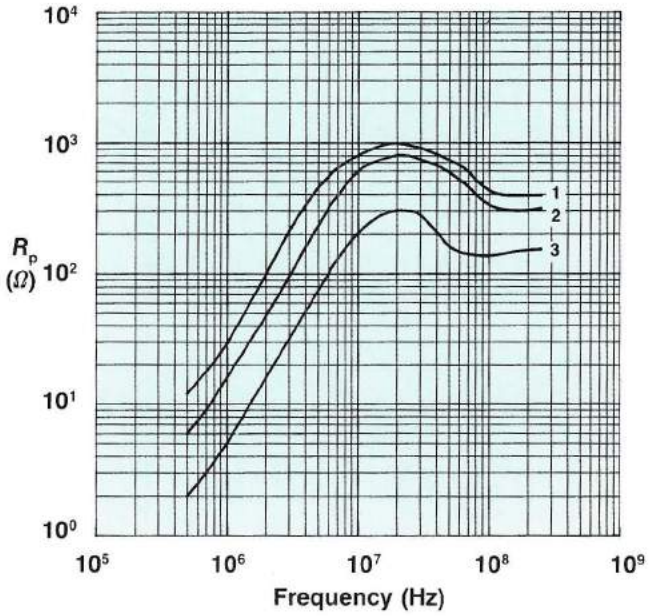


Figure 14 Parallel Resistance vs. Frequency of multi-aperture cores in 61 material.

1 2861006802
2 2861001702
3 2861001802

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Typical Performance Data

Material 61

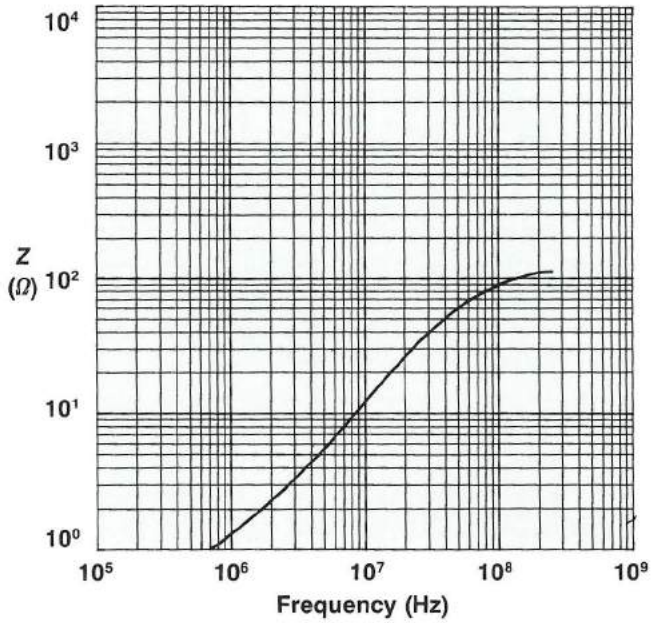


Figure 15 Impedance vs. Frequency of the **12.7 mm** OD shield bead 2661801102.

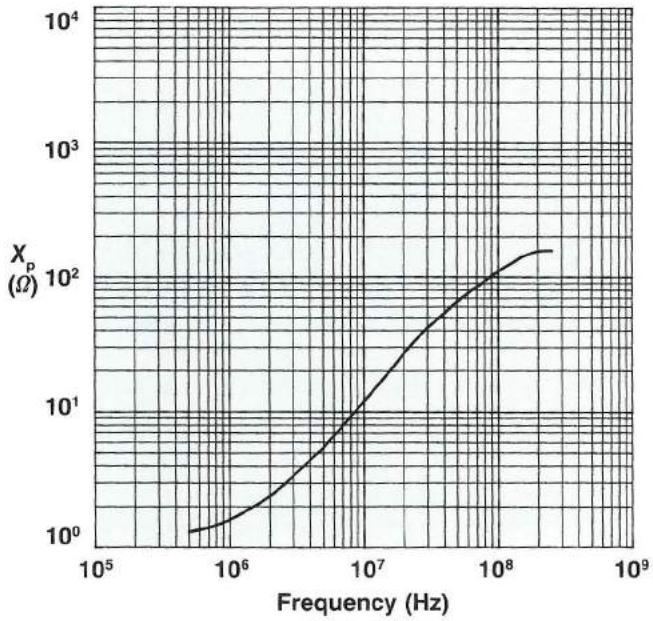


Figure 16 Parallel Reactance vs. Frequency of the **12.7 mm** OD shield bead 2661801102.

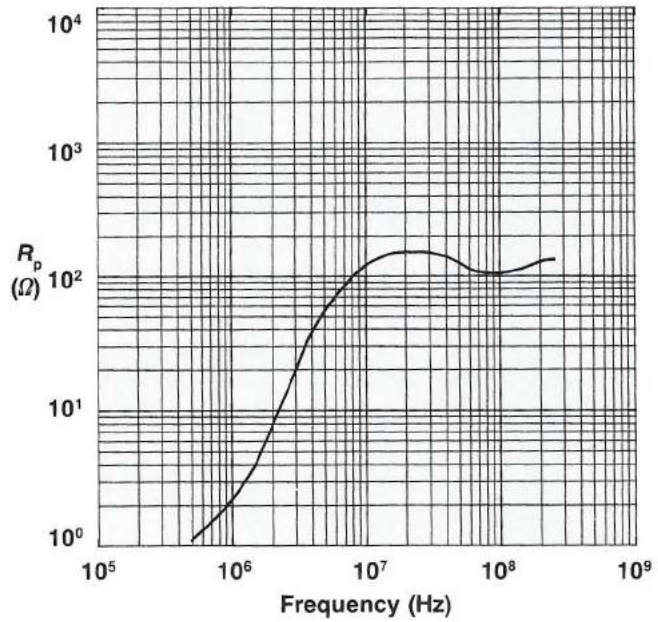


Figure 17 Parallel Resistance vs. Frequency of the **12.7 mm** OD shield bead 2661801102.



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Typical Performance Data

Material 43

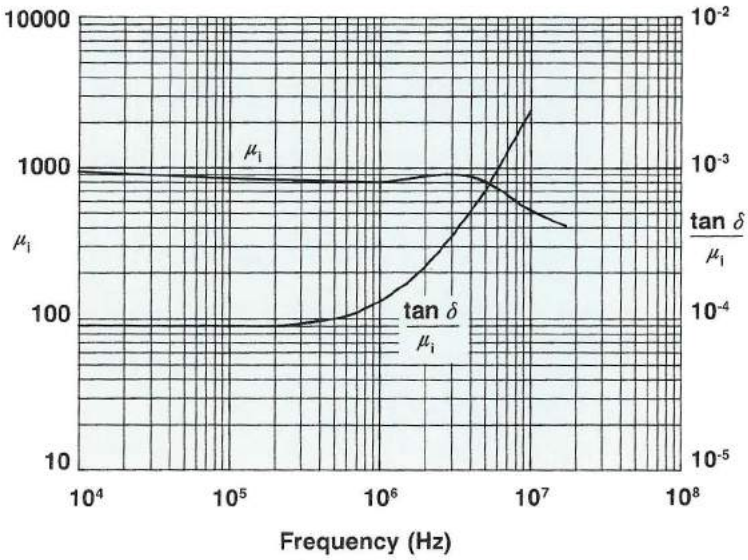


Figure 18 Initial Permeability and Loss Factor vs. Frequency.

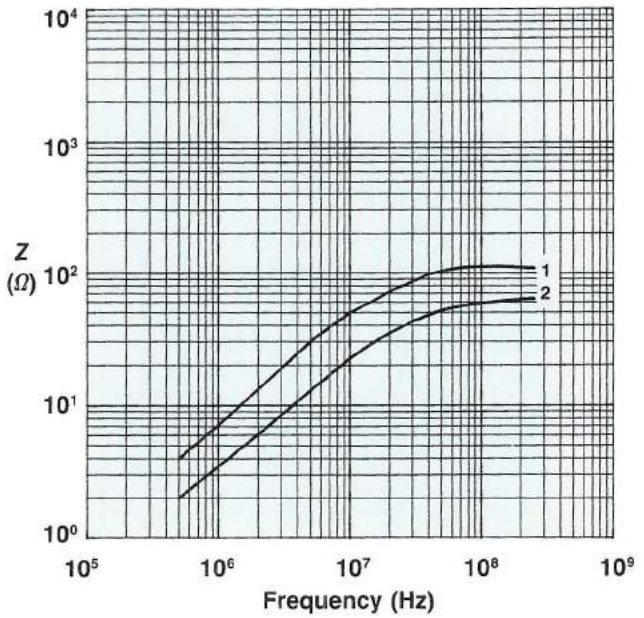


Figure 19 Impedance vs. Frequency of multi-aperture cores in 43 material.

1 2843002402 2 2843002302

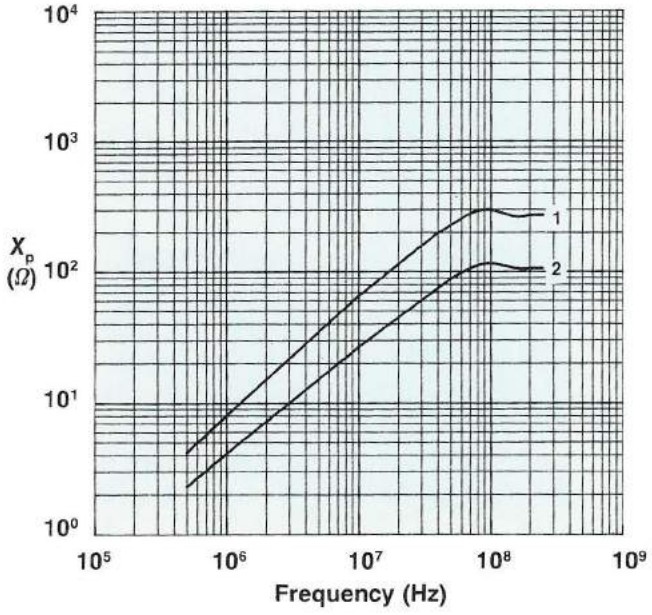


Figure 20 Parallel Reactance vs. Frequency of multi-aperture cores in 43 material.

1 2843002402 2 2843002302

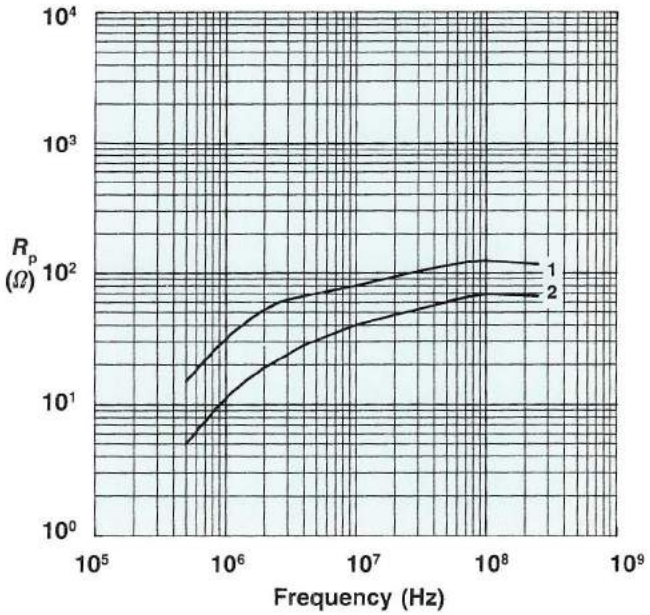


Figure 21 Parallel Resistance vs. Frequency of multi-aperture cores in 43 material.

1 2843002402 2 2843002302

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Typical Performance Data

Material 43

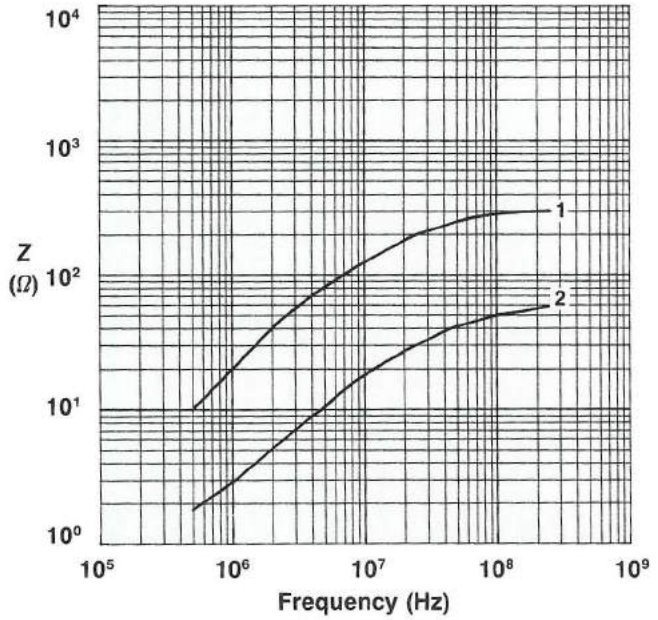


Figure 22 Impedance vs. Frequency of large shield beads in 43 material.

1 2643540002 2 2643002402

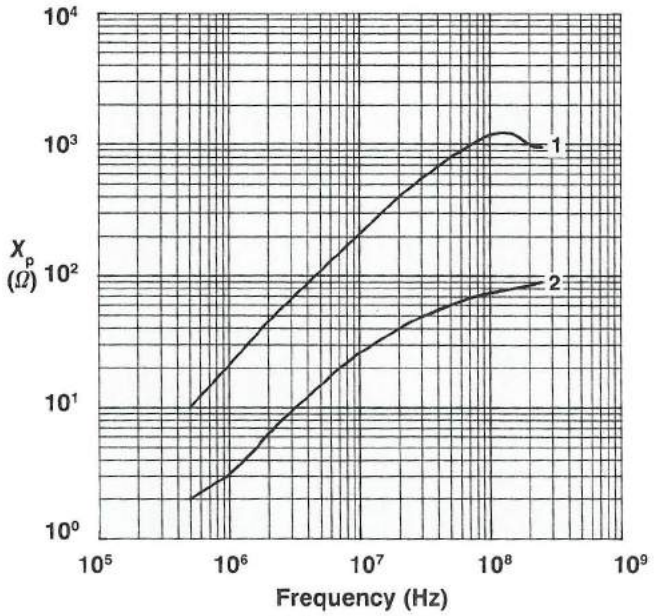


Figure 23 Parallel Reactance vs. Frequency of large shield beads in 43 material.

1 2643540002 2 2643002402

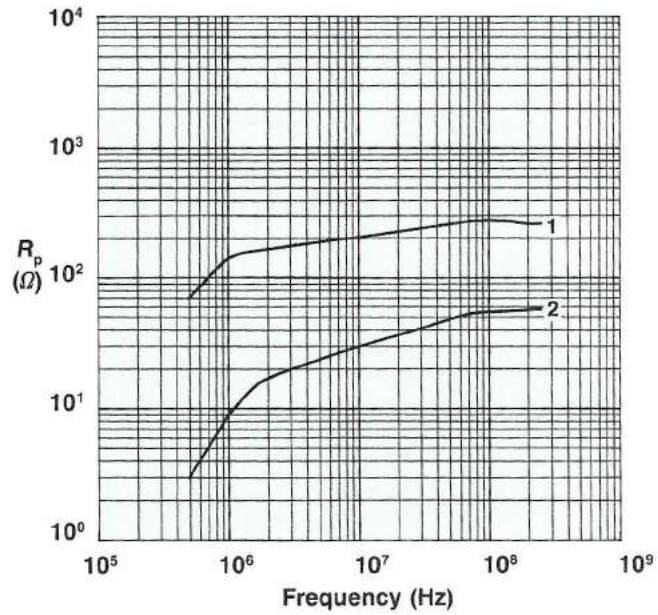


Figure 24 Parallel Resistance vs. Frequency of large shield beads in 43 material.

1 2643540002 2 2643002402



Typical Performance Data

Material 73

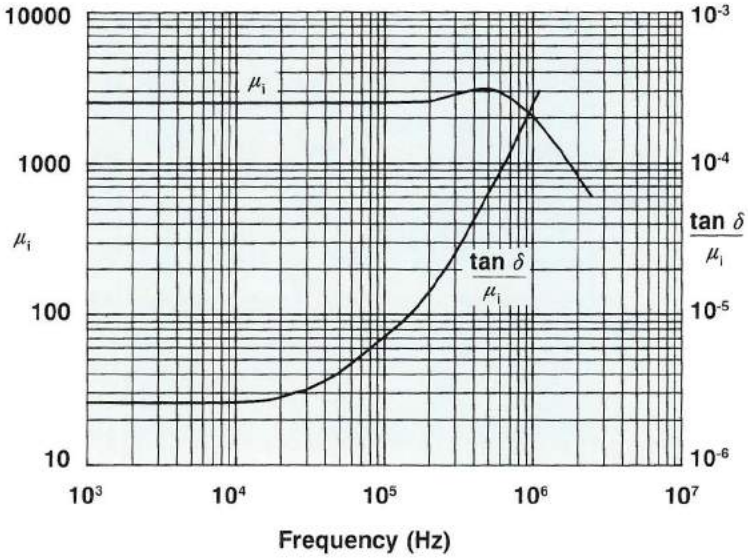


Figure 25 Initial Permeability and Loss Factor vs. Frequency.

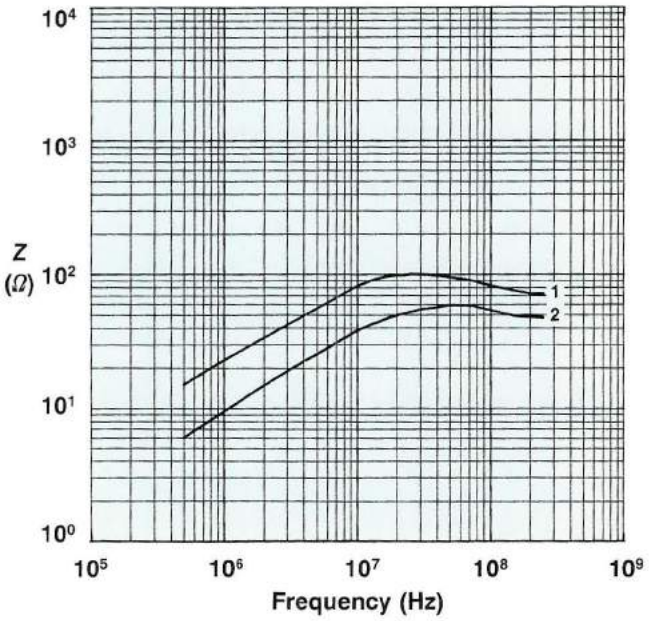


Figure 26 Impedance vs. Frequency of multi-aperture cores in 73 material.

1 2873002402 2 2873002302

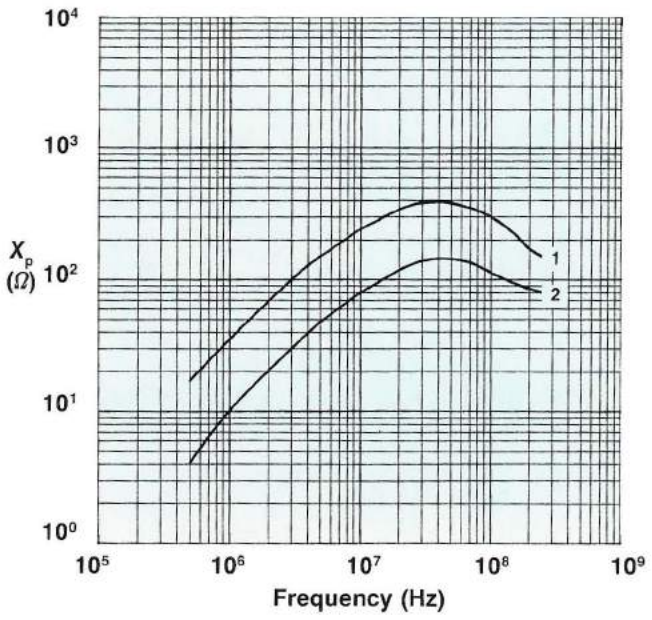


Figure 27 Parallel Reactance vs. Frequency of multi-aperture cores in 73 material.

1 2873002402 2 2873002302

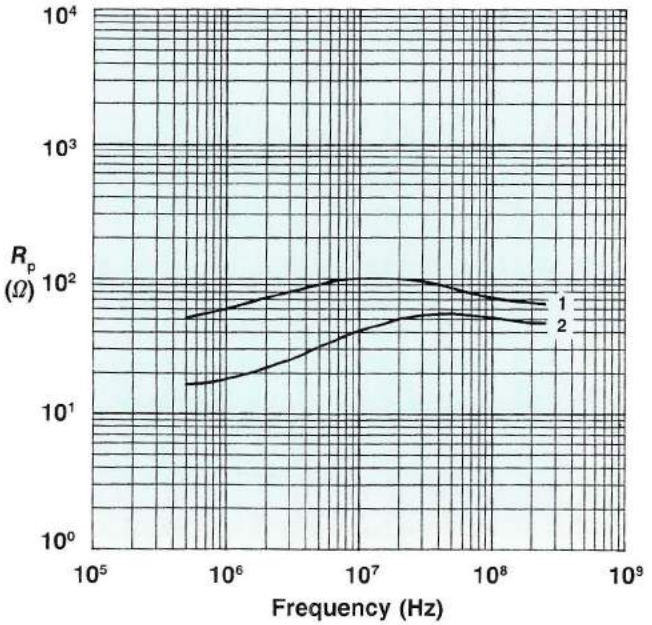


Figure 28 Parallel Resistance vs. Frequency of multi-aperture cores in 73 material.

1 2873002402 2 2873002302

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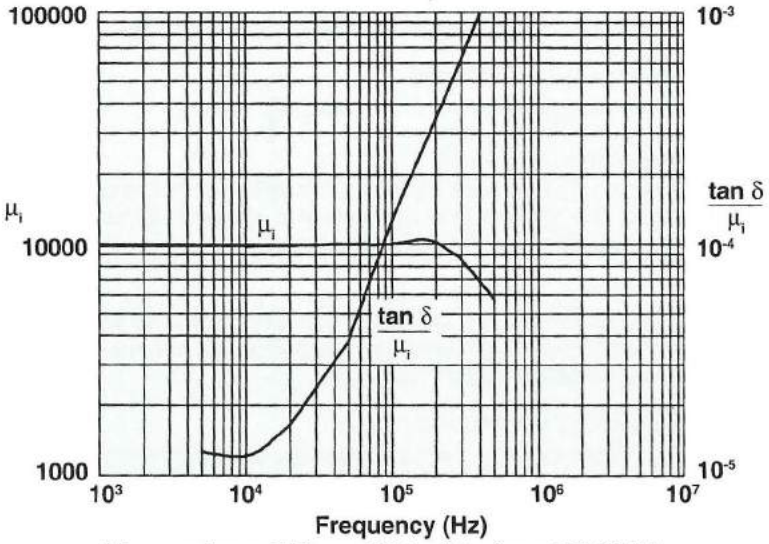
76 Material

Primary Characteristics
Very high initial permeability
Low core loss
Acceptable Curie temperature

Applications
Pulse and broadband transformers
Ground-fault interrupters
LAN Transformers

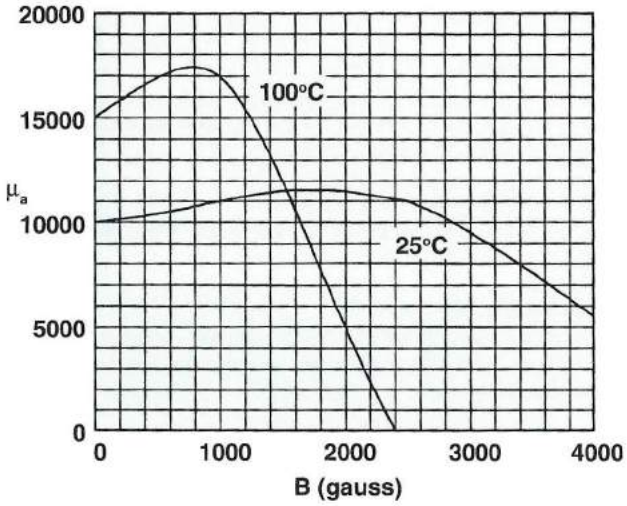
Available Core Shapes
Toroids

Initial Permeability & Loss Factor vs. Frequency



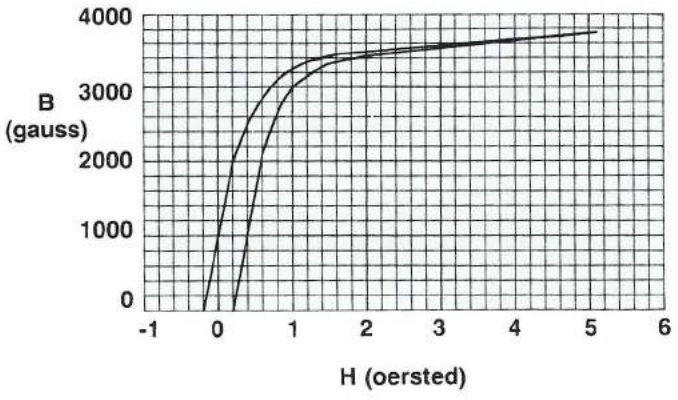
Measured on a **25.4mm** OD toroid using a HP 4275A.

Amplitude Permeability vs. Flux Density



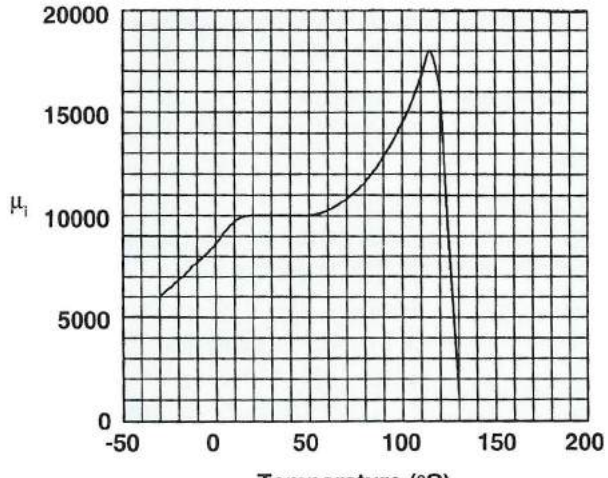
Measured on a **25.4mm** OD toroid at 10 kHz.

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 10 kHz using a HP 4275A.

Glossary of Terms

Air Core Inductance - L_o (H)
The inductance that would be measured if the core had unity permeability and the flux distribution remained unaltered.

Coercive Force - H_c (oersted)
The magnetizing field strength required to bring the magnetic flux density of a magnetized material to zero.

Core Constant - C_1 (cm^{-1})
The summation of the magnetic path lengths of each section of a magnetic circuit divided by the corresponding magnetic area of the same section.

Core Constant - C_2 (cm^{-3})
The summation of the magnetic path lengths of each section of a magnetic circuit divided by the square of the corresponding magnetic area of the same section.

Curie Temperature - T_c (°C)
The transition temperature above which a ferrite loses its ferrimagnetic properties.

Disaccommodation - *D*
The proportional decrease of permeability after a disturbance of a magnetic material, measured at constant temperature, over a given time interval.

Disaccommodation Factor - *DF*
The disaccommodation factor is the disaccommodation after magnetic conditioning divided by the permeability of the first measurement times $\log_{10}$ of the ratio of time intervals.

Effective Dimensions of a Magnetic Circuit -
Area A_e (cm^2), Path Length ℓ_e (cm) and Volume V_e (cm^3)
For a magnetic core of given geometry, the magnetic path length, the cross-sectional area and the volume that a hypothetical toroidal core of the same material properties should possess to be the magnetic equivalent to the given core.

Field Strength - *H* (oersted)
The parameter characterizing the amplitude of alternating field strength.

Flux Density - *B* (gauss)
The corresponding parameter for the induced magnetic field in an area perpendicular to the flux path.

Flux Density, saturation - B_s (gauss)
The maximum intrinsic induction possible in a material.

Inductance Factor - A_L (nH)
Inductance of a coil on a specified core divided by the square of the number of turns. (Unless otherwise specified the inductance test conditions for inductance factor are at a flux density <10 gauss).

Loss Factor - $\tan \delta/\mu_i$
The phase displacement between the fundamental components of the flux density and the field strength divided by the initial permeability.

Magnetic Constant - μ_o
The permeability of free space.

Magnetic Hysteresis
In a magnetic material, the irreversible variation of the flux density or magnetization which is associated with the change of magnetic field strength and is independent of the rate of change.

Magnetically Soft Material
A magnetic material with a low coercivity.

Permeability, amplitude - μ_a
The quotient of the peak value of the flux density and the peak value of the applied field strength at a stated amplitude of either, with no static field present.

Permeability, effective - μ_e
For a magnetic circuit constructed with an air gap or air gaps, the permeability of a hypothetical homogeneous material which would provide the same reluctance.

Permeability, incremental - μ_Δ
Under stated conditions the permeability obtained from the ratio of the flux density and the applied field strength of an alternating field and a superimposed static field.

Permeability, initial - μ_i
The permeability obtained from the ratio of the flux density, kept at <10 gauss, and the required applied field strength. Material initially in a specified neutralized state.

Power Loss Density - *P* (mW/cm^3)
The power absorbed by a body of ferrimagnetic material and dissipated as heat, when the body is subjected to an alternating field which results in a measurable temperature rise. The total loss is divided by the volume of the body.

Remanence - B_r (gauss)
The flux density remaining in a magnetic material when the applied magnetic field strength is reduced to zero.

Temperature Coefficient - *TC*
The relative change of the quantity considered, divided by the difference in temperatures producing it.

Temperature Factor - *TF*
The fractional change in initial permeability over a temperature range, divided by the initial permeability.

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Magnetic Design Formulas

Effective Core Parameters

$C_1 = \Sigma \ell/A$ (cm^{-1}) $\quad C_2 = \Sigma \ell/A^2$ (cm^{-3})

$\ell_e = C_1^2/C_2$ (cm), $\quad A_e = C_1/C_2$ (cm^2) and $\quad V_e = C_1^3/C_2^2$ (cm^3)

Flux Density (Peak)

$$\hat{B} = \frac{E\ 10^8}{4.44\ f\ N\ A_e}\ ^{*} \text{ (gauss)}$$

Field Strength (Peak)

$$\hat{H} = \frac{.4\ \pi\ N\ I_p}{\ell_e} \text{ (oersted)}$$

Where E = RMS sine wave voltage (V)
f = Frequency (Hz)
A_e = Effective cross-sectional area (cm^2)
ℓ_e = Effective path length (cm)
I_p = Peak current (A)
N = Number of turns

* *To check for maximum peak flux density in a non-uniform core set substitute A_{min} for A_e.*

Air Core Inductance

$$L_o = \frac{4\ \pi\ N^2\ 10^{-9}}{C_1} \text{ (H)}$$

C_1 in cm^{-1}

Number of Turns

$$N = \sqrt{\frac{L\ 10^9}{A_L}} \quad L \text{ in H}$$

Inductance

$L = N^2\ A_L$ (nH)

$L = \mu_i \frac{4\ \pi\ N^2}{C_1} 10^{-9}$ (H)

$L = \mu_e \frac{4\ \pi\ N^2}{C_1} 10^{-9}$ (H)

} C_1 in cm^{-1}

Effective Permeability

$$\mu_e = \frac{\ell_e}{\ell_e/\mu_i + \ell}$$

Where ℓ_e = Effective path length
ℓ = Air gap length

Attenuation

$$20 \log_{10} \frac{|Z_s + Z_L + Z_{sc}|}{|Z_s + Z_L|} \text{ (dB)}$$

Where Z_s = Source impedance
Z_L = Load impedance
Z_{sc} = Suppression core impedance

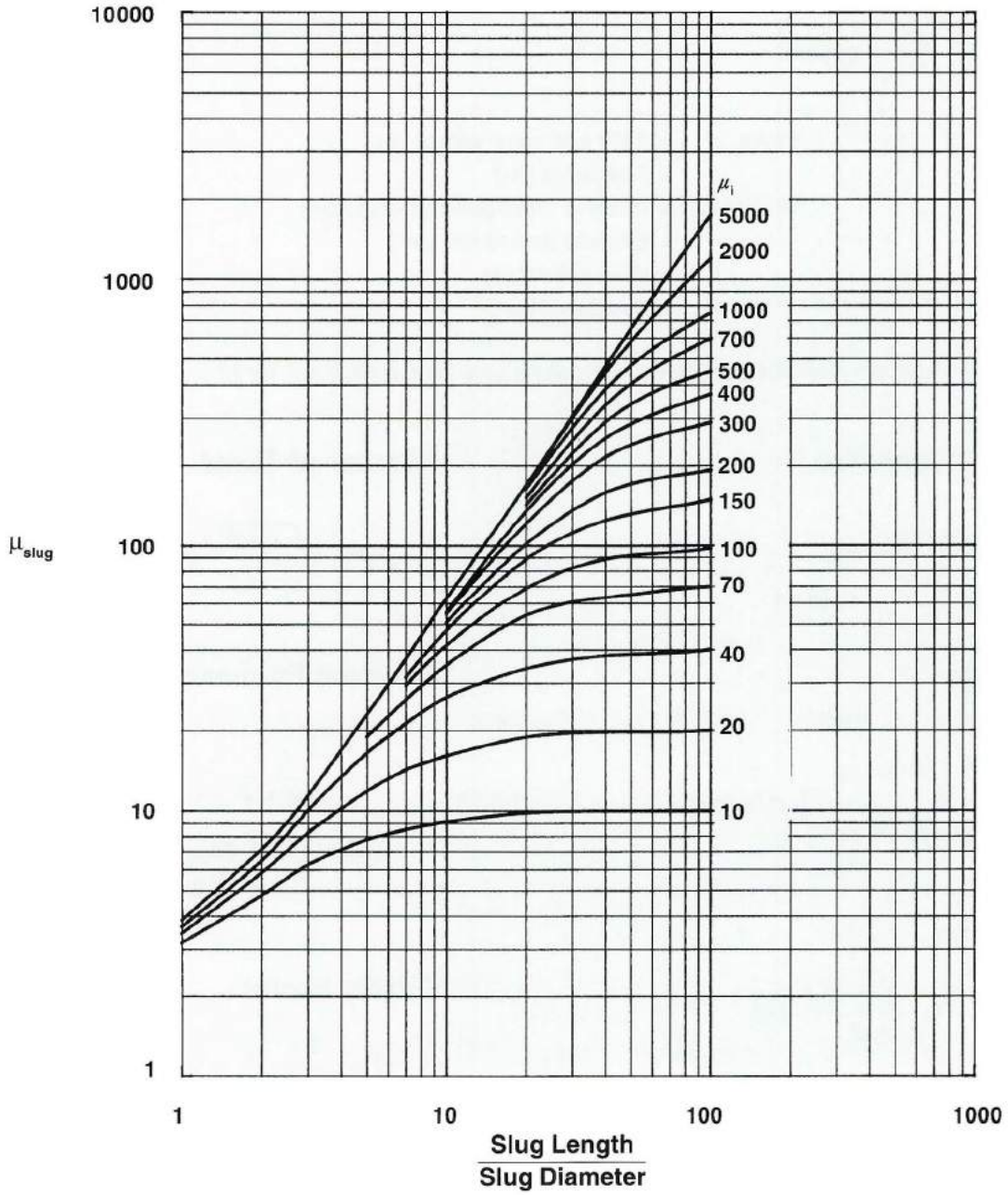
Quality Factor

$$Q = \frac{2\ \pi\ f\ L_s}{R_s} = \frac{R_p}{2\ \pi\ f\ L_p}$$

Effective Permeability of Ferrite Slugs

This family of curves shows the value of the effective permeability of a ferrite slug as a function of its length to diameter ratio, as well as a function of the material permeability of the slug. It illustrates that generally, a great difference exists between the material permeability and the effective permeability of a slug. It also illustrates how, in some instances, the effective permeability of a slug can be influenced by changing its mechanical dimensions more than by changing its material permeability, while in other cases, the reverse is true.

Slug Permeability vs. Slug Length divided by Slug Diameter



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Wire Table of Copper Magnet Wire

AWG & B&S Gauge	Diameter (Inch)	Cross-Sectional Area		Feet per Ohm (20°C)	Ohms per 1000 ft (20°C)	Amperes for 1mA/cir mil	Turns per Inch²
		(Inch²)	(cir mils)				
10	.1019	.00815	10380	1001	1.00	10.4	92
11	.0907	.00647	8234	794	1.26	8.25	118
12	.0808	.00513	6530	630	1.59	6.54	146
13	.0719	.00407	5178	499	2.00	5.18	180
14	.0641	.00322	4107	396	2.53	4.11	231
15	.0571	.00256	3257	314	3.18	3.26	275
16	.0508	.00203	2583	249	4.02	2.59	346
17	.0453	.00161	2048	198	5.06	2.05	432
18	.0403	.00127	1624	157	6.39	1.62	544
19	.0359	.00101	1288	124	8.05	1.29	679
20	.0320	.000804	1022	98.5	10.2	1.03	854
21	.0285	.000638	810.1	78.1	12.8	.81	1065
22	.0254	.000505	642.4	62.0	16.1	.64	1345
23	.0226	.000400	509.5	49.1	20.4	.51	1675
24	.0201	.000317	404.0	39.0	25.7	.40	2095
25	.0179	.000252	320.4	30.9	32.4	.321	2630
26	.0159	.000200	254.1	24.5	40.8	.255	3325
27	.0142	.000158	201.5	19.4	51.4	.201	4110
28	.0126	.000126	159.8	15.4	64.9	.160	5210
29	.0113	.000100	126.7	12.2	81.9	.128	6385
30	.0100	.0000785	100.5	9.7	103.1	.100	8145
31	.0089	.0000622	79.7	7.7	130.1	.079	10,097
32	.0080	.0000503	63.2	6.1	163	.064	12,270
33	.0071	.0000396	50.1	4.8	206	.050	15,615
34	.0063	.0000312	39.8	3.83	261	.040	19,655
35	.0056	.0000248	31.5	3.04	330	.0316	25,530
36	.0050	.0000196	25.0	2.41	415	.0250	31,405
37	.0045	.0000159	19.8	1.91	524	.0203	39,570
38	.0040	.0000126	15.7	1.52	670	.0160	49,070
39	.0035	.00000962	12.5	1.20	832	.0122	65,790
40	.0031	.00000755	9.89	0.953	1049	.0098	82,180
41	.0028	.00000616	7.84	0.756	1323	.0079	98,860
42	.0025	.00000491	6.20	0.598	1672	.0062	121,175
43	.0022	.00000380	4.93	0.476	2101	.0048	158,245
44	.0020	.00000314	3.88	0.374	2674	.0039	205,515
45	.0018	.00000254	3.10	0.299	3344	.0032	249,855
46	.0016	.00000201	2.46	0.238	4202	.0025	310,205

The Effect of Direct Current on the Inductance of a Ferrite Core

Introduction

If ferrite cores are used in the design of transformers, chokes or filters, which are required to carry direct current, it is necessary to predict the degree of inductance degradation caused by the static field. When dc flows through the winding of a ferromagnetic device, it tends to pre-magnetize the core and reduce its inductance. The permeability of a ferrite material measured with superimposed dc might increase slightly for very low values of dc ampere-turns, but then it progressively decreases as the dc field is increased and the core approaches saturation. This permeability is referred to as the incremental permeability μ_Δ. If an air gap is introduced into the magnetic path of a core, the reluctance is increased hence the inductance is decreased. However, the core's capacity for dc ampere-turns without a degradation in inductance is significantly improved, albeit at the expense of a lower effective permeability.

DC Bias in Gapped Cores

The use of graphs such as the Hanna* curves has simplified the tedious trial and error methods often employed when designing inductors with superimposed dc. A Hanna curve is created by measuring the inductance vs. dc bias of various core sizes and gap lengths of the same material grade. The measured data is used to create curves such as those plotted in Figure 1 (this curve is specific for a set of 9477015002 E cores). A line is drawn connecting the individual curves through the point of tangency. The graphs are then normalized by dividing the vertical scale of Figure 1 by the effective core volume V_e and the horizontal scale and the gap lengths by the effective path length ℓ_e of the core set. The individual curves, once normalized, overlay creating the Hanna curve. Figure 2 is such a curve for Fair-Rite 77 material and can be used for all core sets in that material.

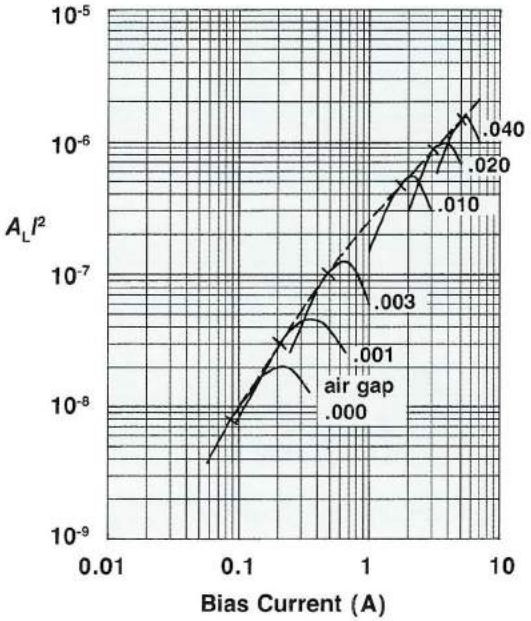


Figure 1 Product Inductance Factor and Current Squared vs. DC Current for a pair of 9477015002 E Cores.

Design Example

For a typical output choke application, the designer knows a number of design criteria such as the required inductance, the direct current, alternating ripple current and allowable dc resistance. He will also have requirements for core size, ambient temperature an often a preference for a particular core geometry.

***Footnote:** C.R. Hanna presented a paper "Design of Reactances and Transformers which Carry Direct Current" at the 1927 Winter Convention of AIEE. The paper provided a method of calculating the air gap that will yield the maximum inductance for a given number of turns, with a specified amount of dc, for a particular material.*

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The following example illustrates the use of the Hanna curve in the design of an inductor.

Inductor specifications:

Minimum inductance	L	= 1 mH
Direct current	I_{dc}	= 1 A
Alternating ripple current	I_{ac}	= 0.2 A
Maximum dc resistance	R_{dc}	< 0.2Ω

Step 1. Initial Core Selection.
Using the Hanna curve for 77 material of Figure 2, select a value for LI^2/V_e approximately mid range on the vertical axis, that is between 10^{-4} and 10^{-3}. Any value greater than 10^{-3} will work the ferrite too hard and the dc resistance is apt to be high. Anything lower than 10^{-4} will result in a conservative design and the dc resistance will be quite low.

Select therefore $LI^2/V_e = 3.5\ 10^{-4}$

Calculate V_e from:

$V_e = LI^2/3.5\ 10^{-4}$
L_{min} = 1 mH, design for $L = 1.1\ 10^{-3}$ H
$I = I_{dc} + I_{ac}/2 = 1 + 0.2/2 = 1.1$ A
$V_e = 1.1\ 10^{-3} \times 1.1^2/3.5\ 10^{-4} = 3.8\ cm^3$

Select E core (preferred core shape), based upon the calculated core volume of 3.8 cm³ from the catalog, pages 106 and 107. Two Fair-Rite E cores are considered:

9477015002 $V_e = 1.95\ cm^3$ and
9477014002 $V_e = 3.92\ cm^3$.

The 9477014002 is closest and will be used in this inductor design. The core parameters for this E core set are:

$\ell_e = 4.9$ cm, $A_e = .80\ cm^2$ and $V_e = 3.92\ cm^3$.

Recalculate
$LI^2/V_e = 1.1\ 10^{-3} \times 1.1^2/3.92 = 3.4\ 10^{-4}$.

Step 2. Number of Turns, Wire Size and Wire Fit.
From Figure 2, a $LI^2/V_e = 3.4\ 10^{-4}$ yields a H value of 17 oersted.

Calculate turns N from the formula $H = .4\ \pi\ NI/\ell_e$ oersted.
$N = 17 \times 4.9/.4 \times \pi \times 1.1 = 60.3$ or 61 turns.

From the core dimensions, the core winding area can be calculated, see Table 1.

Winding area for a set of E cores 9477014002 is:

A_w = D (F-E) in inch².
$A_w = .255\ (.740\text{-}.250) = .125$ inch².

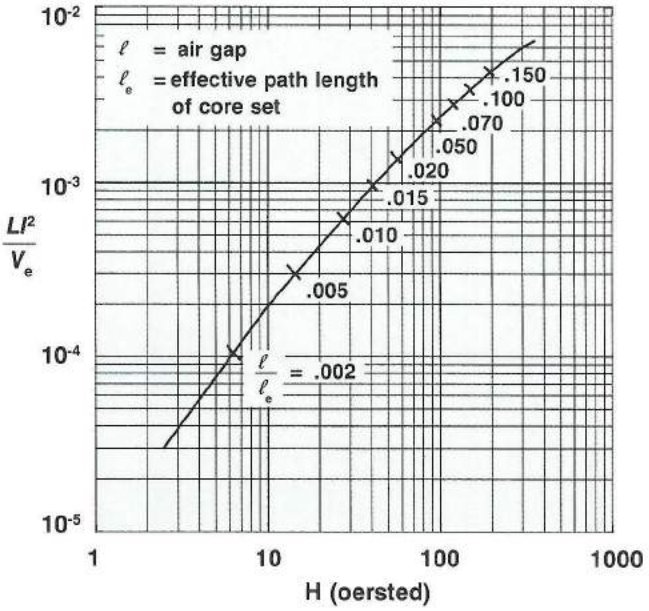


Figure 2 Hanna Curve for Core Sets in Fair-Rite 77 material.

Table 1 Core Winding Area (inch²)	
E Cores	D(F-E)
ETD Cores	D(F-E)
PQ Cores	D(F-E)
Pot Cores	F(B-C)
EP Cores	F(B-C)

Since the winding area of the appropriate bobbin is smaller than the core winding area, a correction factor F_c has to be used to determine the bobbin winding area. Figure 3 gives this correction-factor F_c as a function of the calculated core winding area A_w. A set of E cores 9477014002 has a $A_w = .125$ inch², from Figure 3 can be determined that the $F_c = .55$, therefore the bobbin winding area is .55 x .125 = .069 inch². Using a conservative current density of 1 mA per circular mil or 1275 A per inch², an initial wire size selection of 20 AWG can be made from the Wire Table on page 25. To determine the dc resistance of the winding, first find the average length of turn from Table 2.

Table 2 Mean Length of Turn (inch)	
E Cores	2 (C+F)
ETD Cores	.5 π (E+F)
PQ Cores	.5 π (E+F)
Pot Cores	.5 π (B+C)
EP Cores	.5 π (B+C)

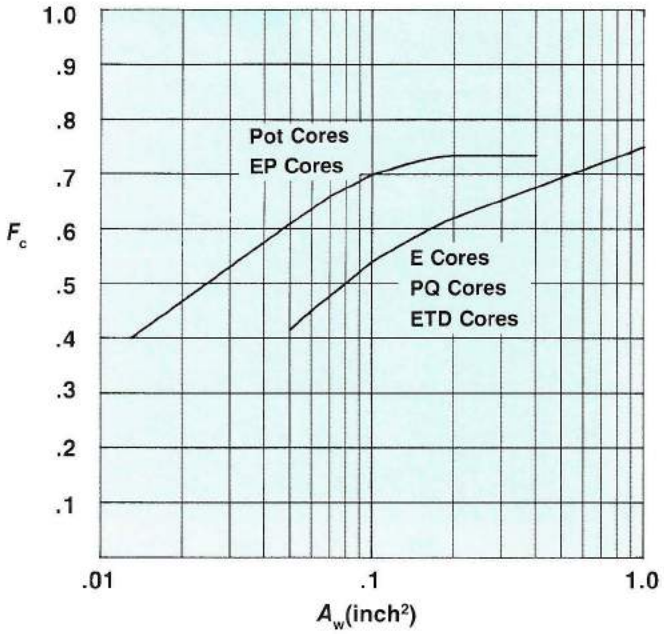


Figure 3 Correction Factor F_c vs. Core Winding Area A_w.

Average length of turn for E 9477014002 is:

$\ell_{avg.}$ = 2 (C+F)

$\ell_{avg.}$ = 2 (.500 + .740) = 2.48 inch.

R_{dc} = 2.48 x 61 x 10.2/12000 = 0.13 Ω

(From the Wire Table, 1000 ft of 20 AWG has a resistance of 10.2 Ω)

To check for winding fit, multiply the number of turns per square inch for 20 AWG from the Wire Table with the bobbin winding area of .069 inch². For 20 AWG, the bobbin winding area can accommodate 854 x .069 = 58.9 turns. This is too close to the calculated turns for an easily manufactured magnetic design. Use 21 AWG wire instead.

R_{dc} = 2.48 x 61 x 12.8/12000 = 0.16 Ω.

Winding fit for 21 AWG:

N = 1065 x .069 = 73.5, well above the require 61 turns.

Step 3. Air gap.

Going back to Figure 2, for $LI^2/V_e = 3.4\ 10^{-4}$ and a H = 17 oersted, a ℓ/ℓ_e ratio of approximately .006 is found.

The gap length = .006 x ℓ_e.

ℓ = .006 x 4.9/2.54 = .012 inch.

To summarize:

E core 9477014002	N = 61 turns
Wire size 21 AWG	Gap length .012 inch

The graphs in Figures 4 through 8 show the inductance factors or A_L values as a function of the air gaps for the different core types and sizes.The air gap determined in the design example and the air gaps shown in Figures 4 through 8 represent the total air gap. The most practical way to obtain this air gap is to grind this gap into the center leg of one of the core halves. Non-metallic shims can also be used to obtain the desired air gap. This is usually done by placing shims between the outer legs or outside rims of the core halves. In cores with a uniform cross-sectional area, the A_L value or inductance index will be the same whether the core is gapped or shims are used that have a thickness half the total air gap. For cores that have a non-uniform cross-sectional area the shim thickness can be calculated from:

$$\text{Shim thickness} = \text{total air gap} \times \frac{\text{center mating area}}{\text{total mating area}}$$

The above example of the E core 9477014002, a core with a uniform cross-sectional area, can therefore use .006 inch shims between the outer legs.

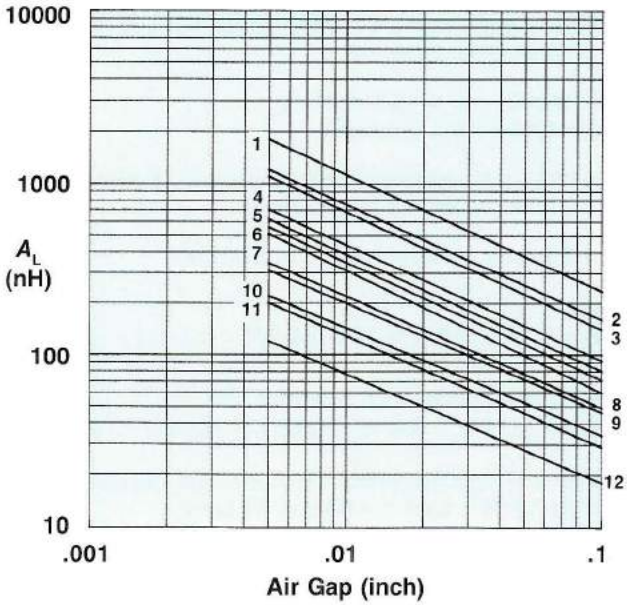


Figure 4 A_L vs. Gap for E Cores in 77 and 78 material.

1 94 - - 625002	**5** 94 - - 375002	**9** 94 - - 015002
2 94 - - 500002	**6** 94 - - 014002	**10** 94 - - 016002
3 94 - - 036002	**7** 94 - - 018002	**11** 94 - - 020002
4 94 - - 017002	**8** 94 - - 012002	**12** 94 - - 019002

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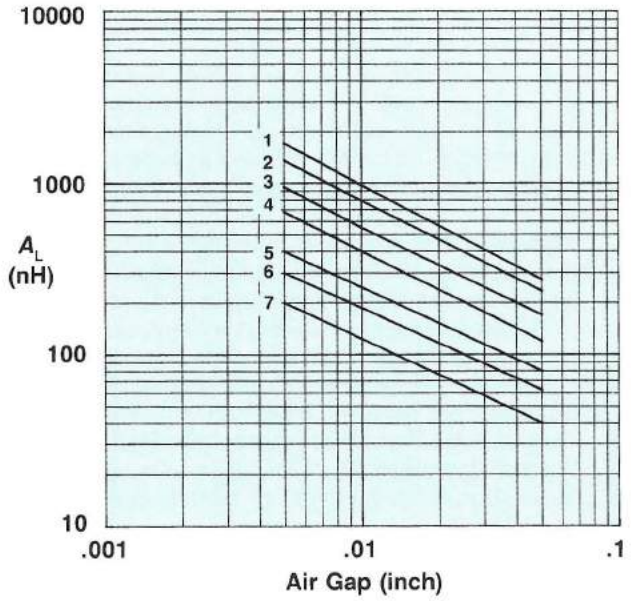


Figure 5 A_L vs. Gap for Pot Cores in 77 and 78 material.

1 56 - - 422921
2 56 - - 362221
3 56 - - 301921
4 56 - - 261621
5 56 - - 221321
6 56 - - 181121
7 56 - - 140821

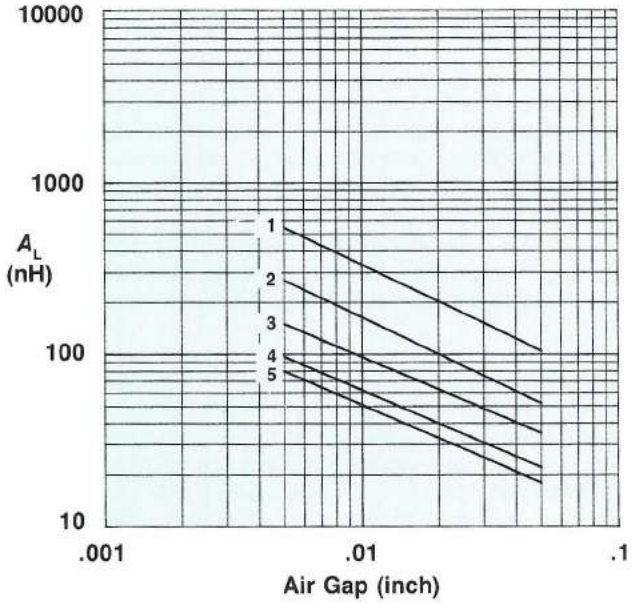


Figure 6 A_L vs. Gap for EP Cores in 77 and 78 material.

1 65 - - 202021
2 65 - - 171721
3 65 - - 131321
4 65 - - 101021
5 65 - - 070721

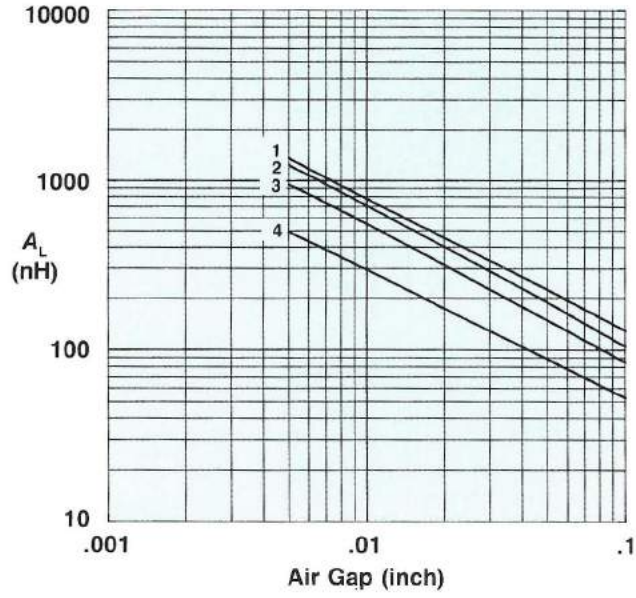


Figure 7 A_L vs. Gap for PQ Cores in 77 and 78 material.

1 66 - - 404021
66 - - 353521
2 66 - - 323021
66 - - 322021
3 66 - - 262521
66 - - 262021
4 66 - - 202021
66 - - 201621

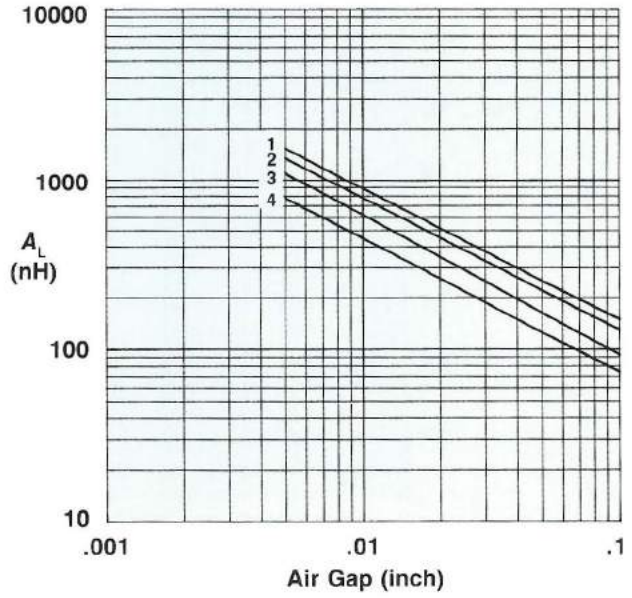


Figure 8 A_L vs. Gap for ETD Cores in 77 and 78 material.

1 95 - - 490002
2 95 - - 440002
3 95 - - 390002
4 95 - - 340002

DC Bias in Open Magnetic Cores

The discussion so far has been on core types that have a closed magnetic path, in which a small air gap has been inserted by either a ground gap or the use of shims. An open magnetic core can be thought of as a core with a very large fixed air gap. Since the air gap is determined by the core geometry and cannot be changed, the Hanna curves can not be used for these types of cores. Such cores as rods, slugs and bobbins can be used quite successfully in inductor designs that have relative low inductance values and can accommodate significant amounts of static currents.

The large air gap will forestall the saturation of this type of core, hence the inductance will not as easily decrease as a function of the dc ampere-turns. The Fair-Rite bobbins, listed on the pages 110 and 111 of the catalog, are specified to an inductance index or A_L with a tolerance of ± 10% and also by a *NI* product of dc ampere-turns, which might reduce the A_L value but not more than 5%. For an inductor design the number of turns can be calculated from the required inductance *L* and the inductance index of the bobbin. $N = \sqrt{L/A_L}$, (*L* in nH). The turns *N* times the direct current *I* will give the *NI* product, which should be less than the value quoted for the bobbin. For winding fit and dc resistance check, the same procedure is used as outlined in the example above, except here the W_a of the bobbin is the total available winding area. The graphs of Figure 9 show the effect of temperature on the inductance factor vs. dc bias characteristics of the 9677242409 bobbin. As can be seen from these curves, the decrease in inductance increases with temperature. The *NI* values listed in the catalog are at room temperature, and must be derated when operating at elevated temperatures. Open magnetic cores, rods, slugs and bobbins are used and designed into SCR and triac controls, speaker crossover networks and differential-mode input filters. They are also utilized for EMI suppression applications where relative large direct currents are present and for output chokes in switched-mode power supplies.

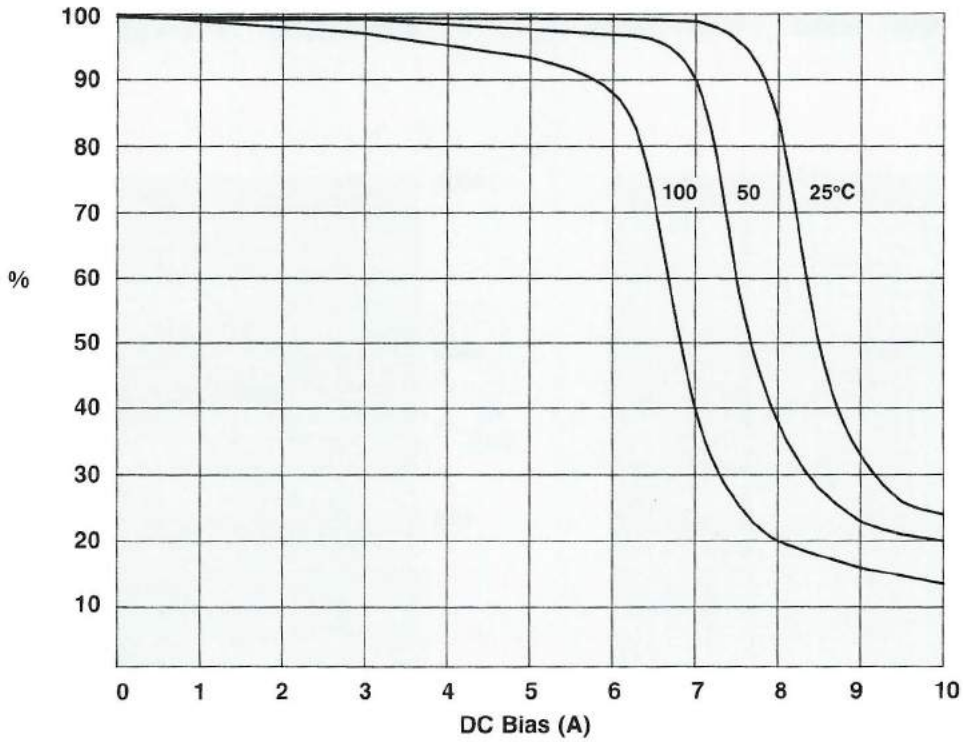


Figure 9 Percent of Original Inductance Factor vs. DC Bias and Temperature.

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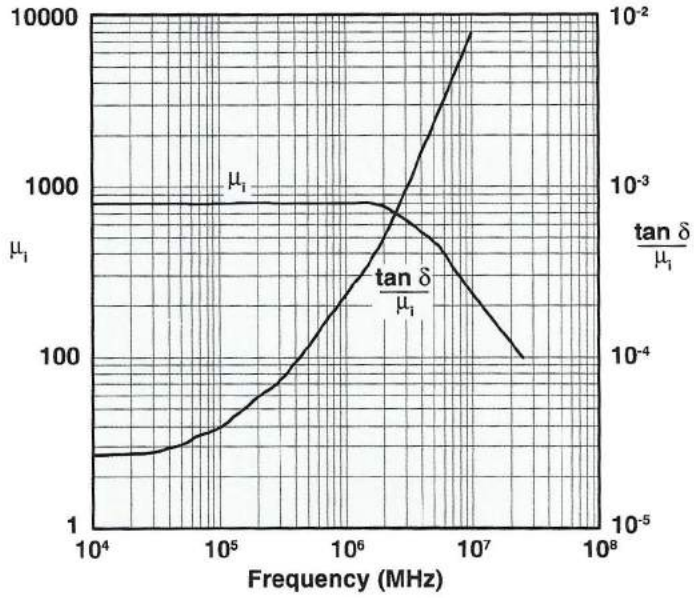
85 Material

Primary Characteristics
Square hysteresis loop over a wide temperature range
Squareness ratio Br/B > 0.85 at 23°C
Curie temperature > 200°C

Applications
Converters and inverters
Output regulators for switch-mode power supplies
Spike Suppression

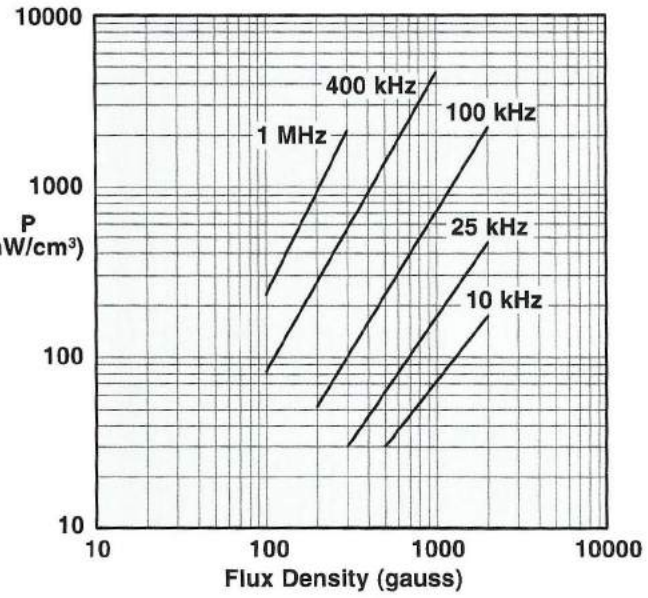
Available Core Shapes
Toroids

Initial Permeability & Loss Factor vs. Frequency



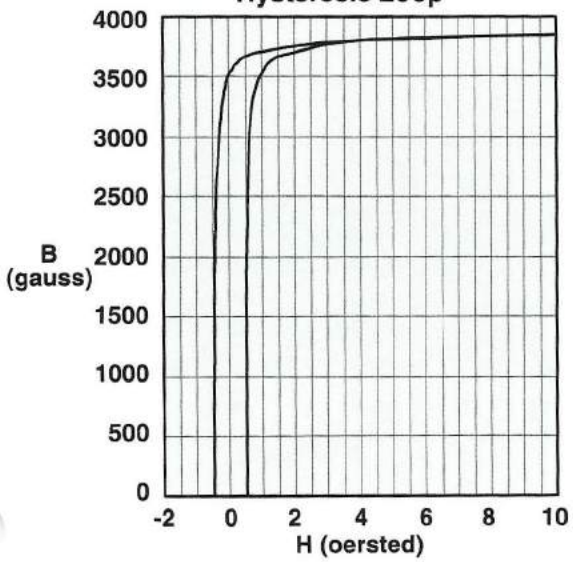
Measured on a **13mm** OD toroid using HP 4284A and HP 4285A

Power Loss Density vs. Flux Density



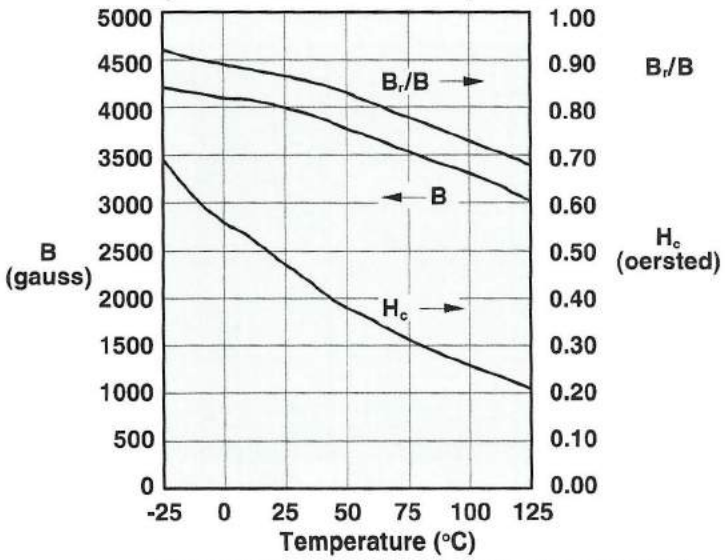
Measured on a **13mm** OD toroid using a Clarke-Hess 258 VAW.

Hysteresis Loop



Measured on a **13mm** OD toroid.

Flux Density, Coercive Force and Squareness Ratio vs. Temperature



Measured on a **13mm** OD toroid, B is measured at H=10 oersted.

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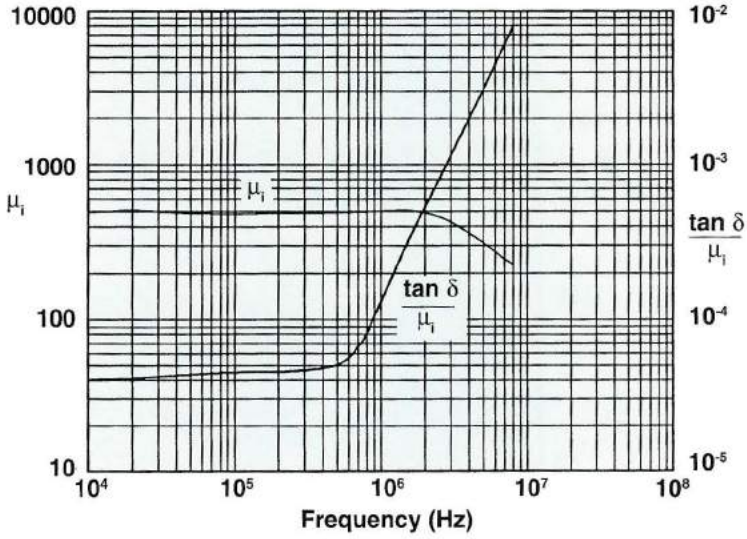
44 Material

Primary Characteristics
High impedance
Very high volume resistivity

Applications
Designs which require a very high ferrite resistivity. Typical applications are when copper conductors make contact with the ferrite material, eg in connector suppression plates.
Suppression material for use over the 25 to 200 MHz frequency range.

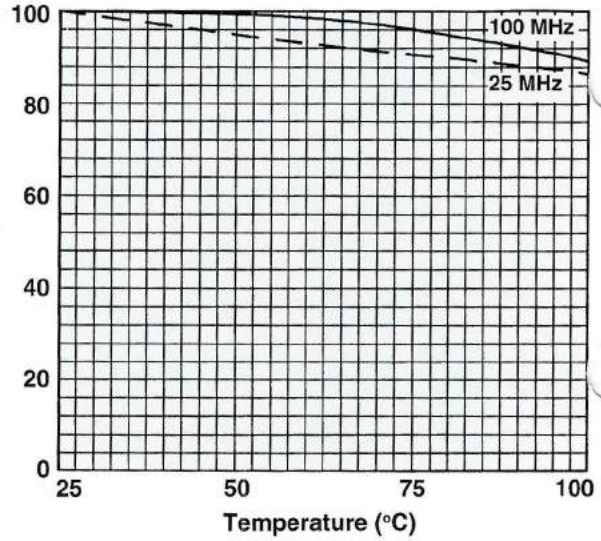
Available Core Shapes
Round cable suppression cores, SM beads, connector suppression plates

Initial Permeability & Loss Factor vs. Frequency



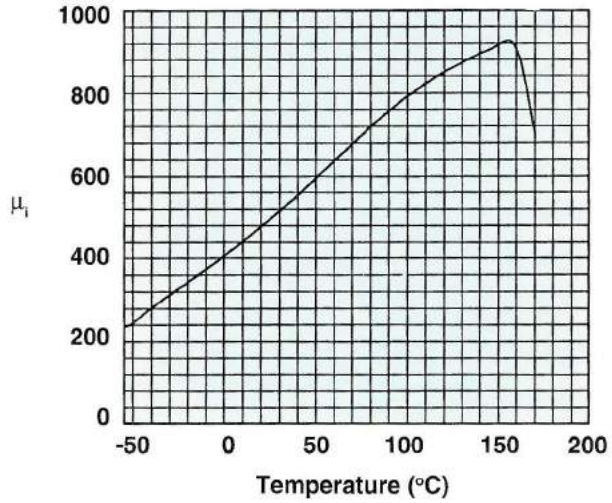
Measured on a **25.4mm** OD toroid using HP 4275A.

Percent of Original 25°C Impedance vs. Temperature



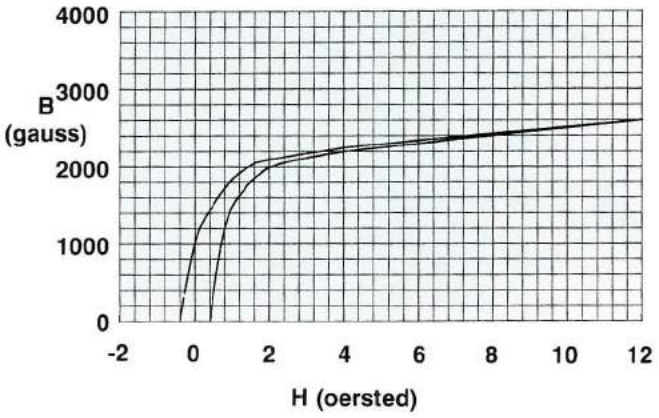
Measured on a **30mm** OD toroid using a HP 4193A.

Initial Permeability vs. Temperature



Measured on a **30mm** OD toroid at 100 kHz using a HP 4275A.

Hysteresis Loop



Measured on a **25.4mm** OD toroid at 10 kHz.

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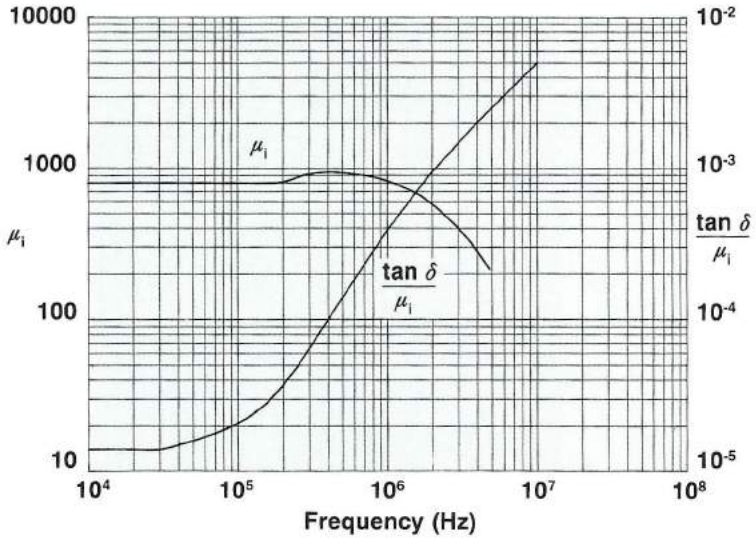
33 Material

Primary Characteristics
Frequency range up to 30 MHz
Medium permeability
Good temperature stability

Applications
Inductors at medium frequency.

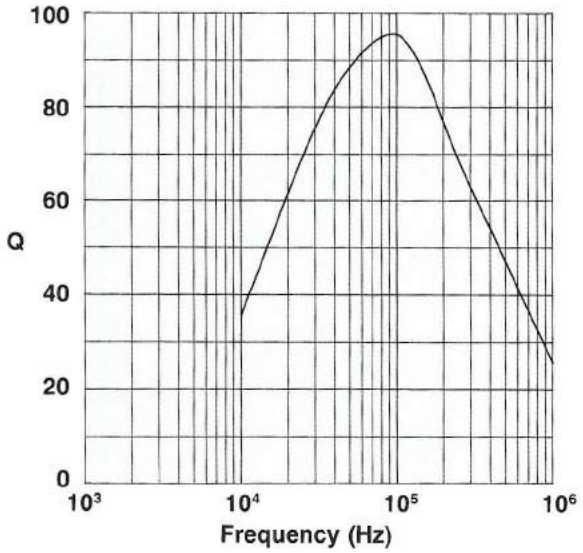
Available Core Shapes
Slugs

Initial Permeability & Loss Factor vs. Frequency



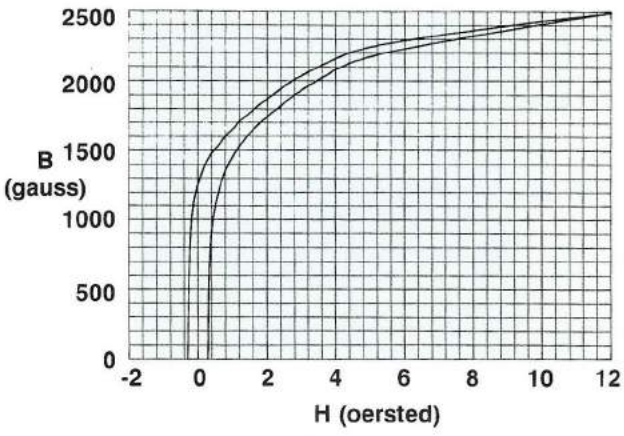
Measured on a **25.4mm** OD toroid using HP 4275A.

Quality Factor vs. Frequency



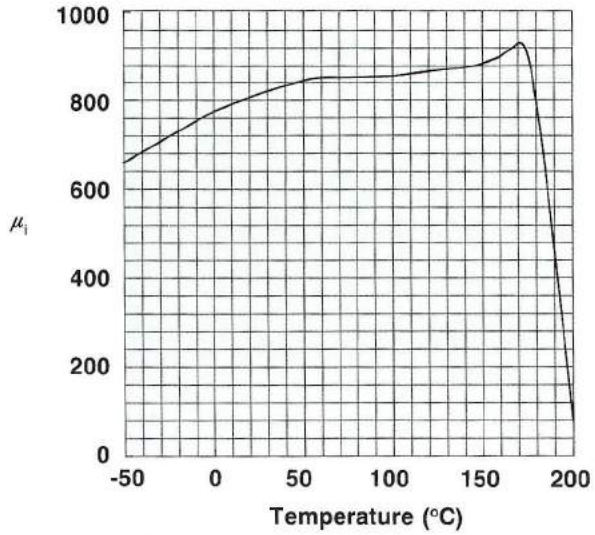
Measured on a **25.4mm** OD toroid using HP 4275A.

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

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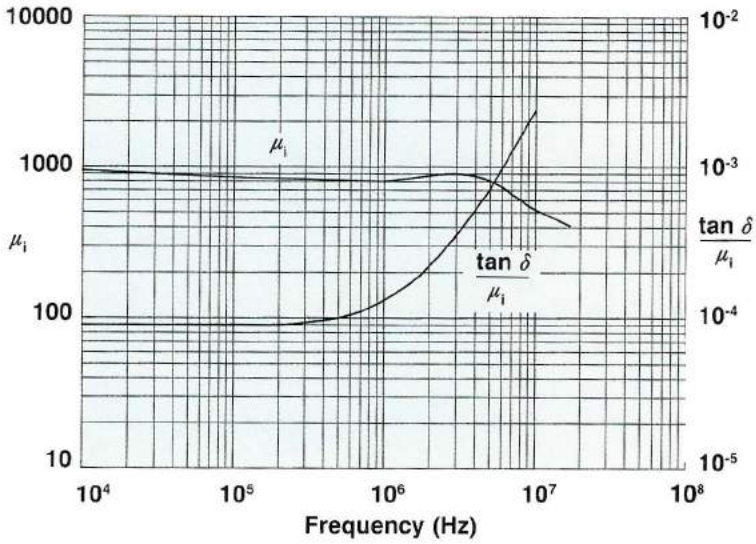
43 Material

Primary Characteristics
High impedance
High resistivity

Applications
Recommended suppression material of unwanted signals between 30 MHz and 200 MHz.

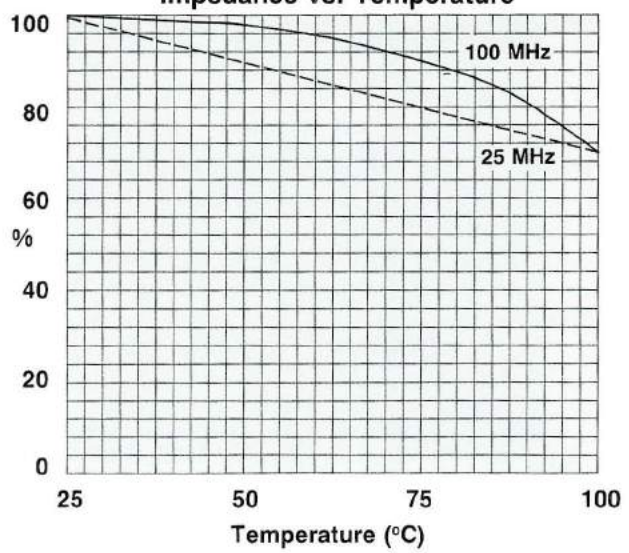
Available Core Shapes
Shield beads, SM beads, beads on leads, wound beads, PC beads, flat and round cable suppression cores, bobbins, toroids, multi-aperture cores

Initial Permeability & Loss Factor vs. Frequency



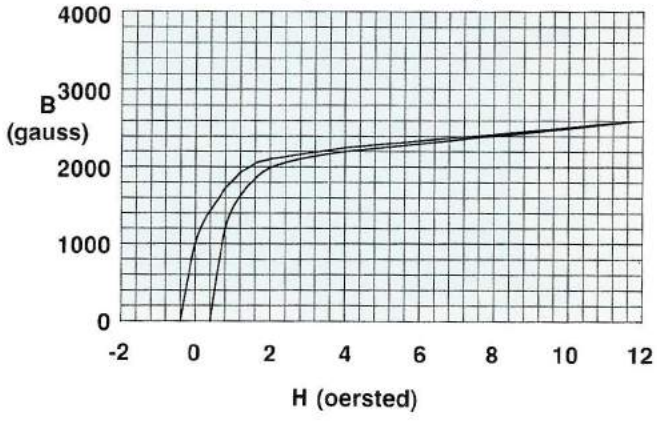
Measured on a **25.4mm** OD toroid using HP 4275A and HP 4191A.

Percent of Original 25°C Impedance vs. Temperature



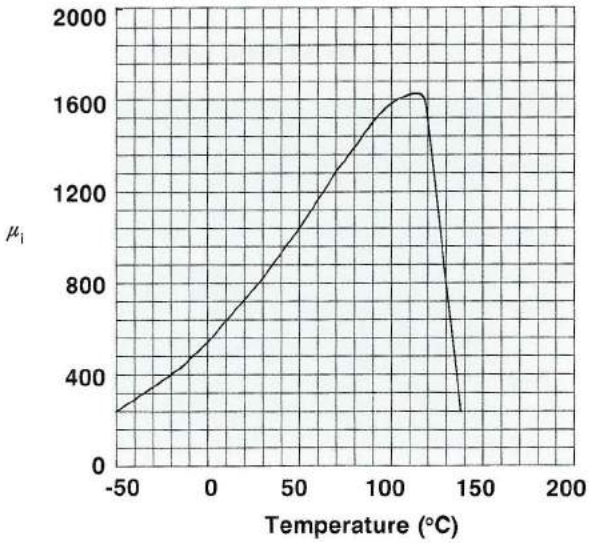
Measured on a **25.4mm** OD toroid using a HP 4191A.

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

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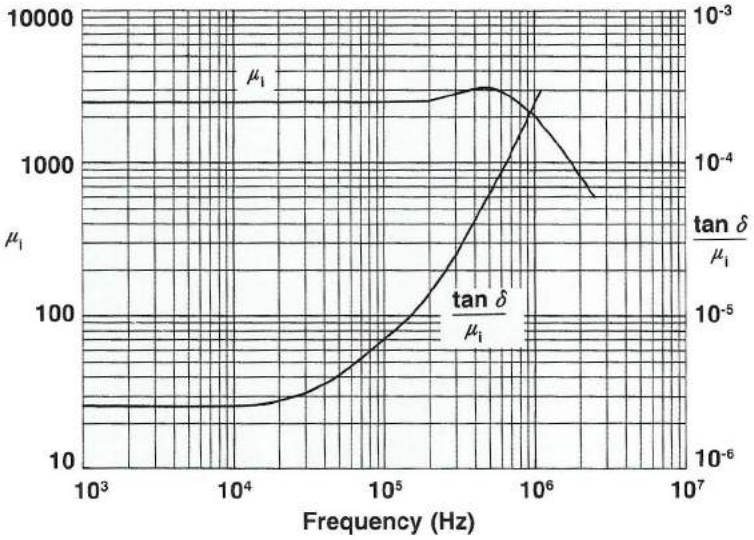
73 Material

Primary Characteristics
High impedance
Temperature stable impedance
High initial permeability

Applications
Optimum material for EMI suppression below 30 MHz
Broadband transformers

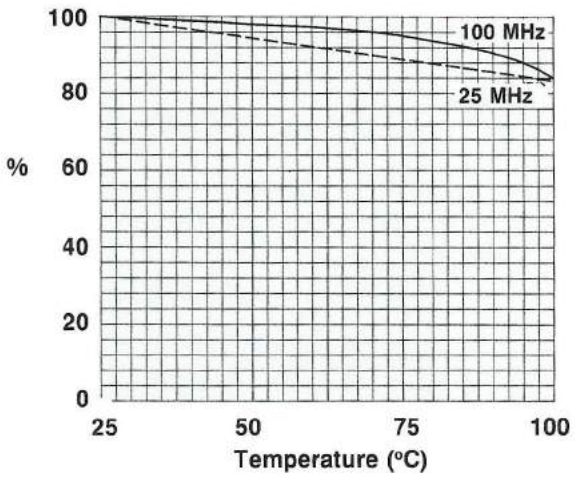
Available Core Shapes
Shield beads, beads on leads,
multi-aperture cores, SM beads

Initial Permeability & Loss Factor vs. Frequency



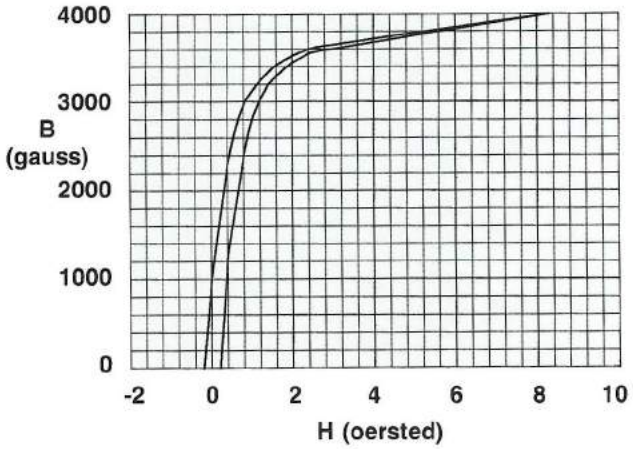
Measured on a **25.4mm** OD toroid using a HP 4275A.

Percent of Original 25°C Impedance vs. Temperature



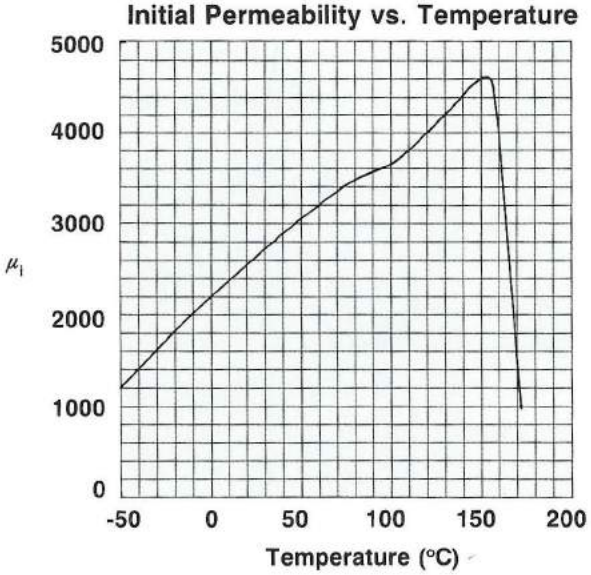
Measured on a **25.4mm** OD toroid using HP 4191A.

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

Initial Permeability vs. Temperature



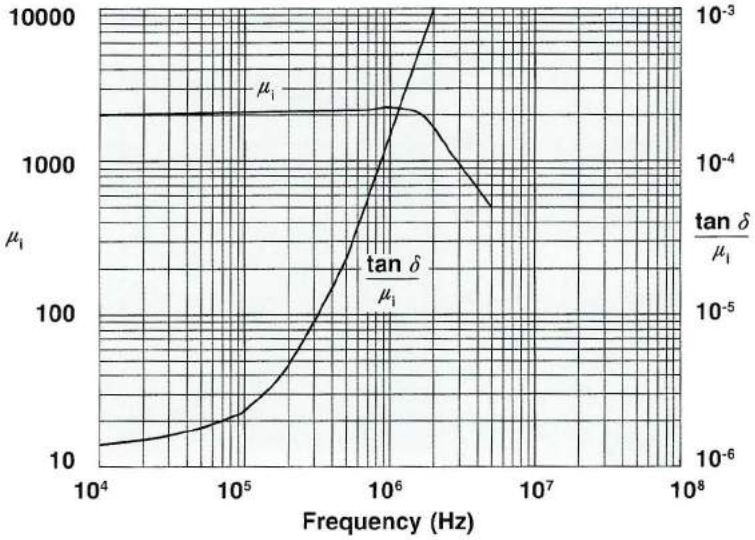
Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

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77 Material

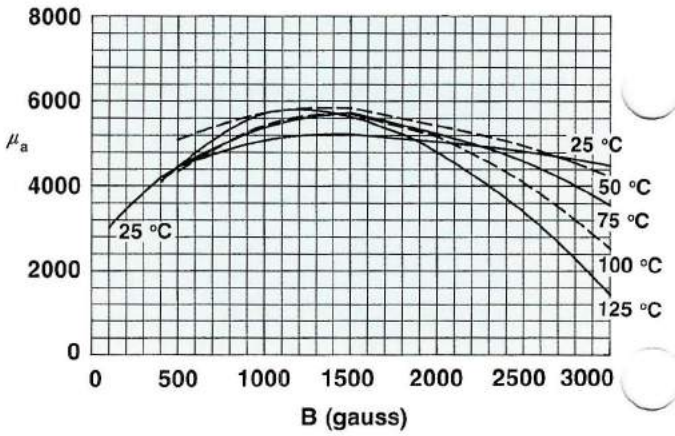
Primary Characteristics
High flux density at high temperatures.
Low core loss at high flux densities and high temperatures.
High amplitude permeabilities at high temperatures.
High Curie temperature.

Initial Permeability & Loss Factor vs. Frequency



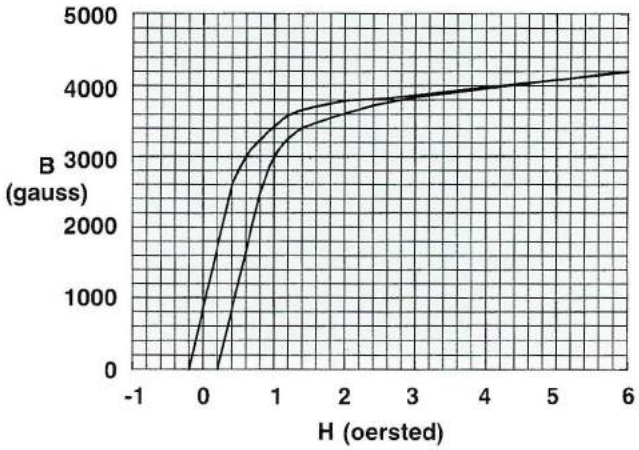
Measured on a **25.4mm** OD toroid using a HP 4275A.

Amplitude Permeability vs. Flux Density



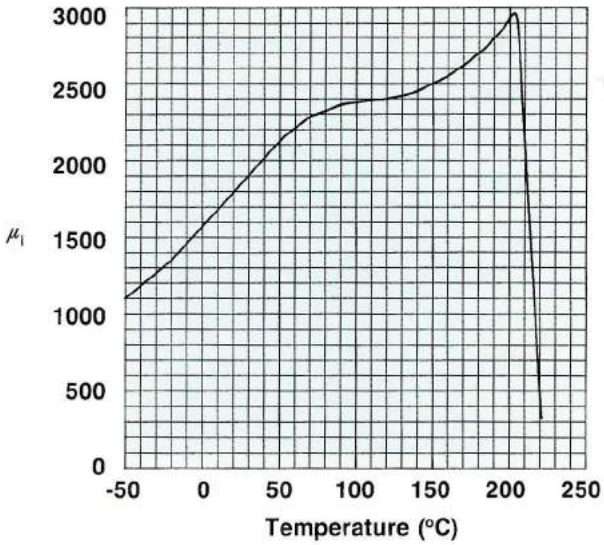
Measured on a **25.4mm** OD toroid at 10 kHz .

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

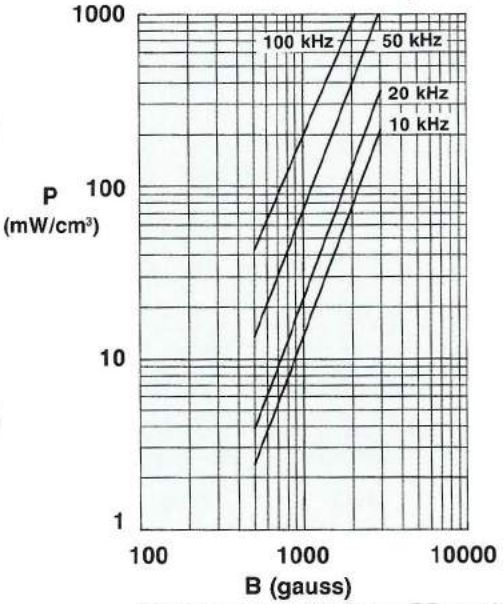
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77 Material

Applications
Transformers and inductors for switched-mode power supplies and DC-DC converters.
Power filters
Ignition coils
Broadband transformers

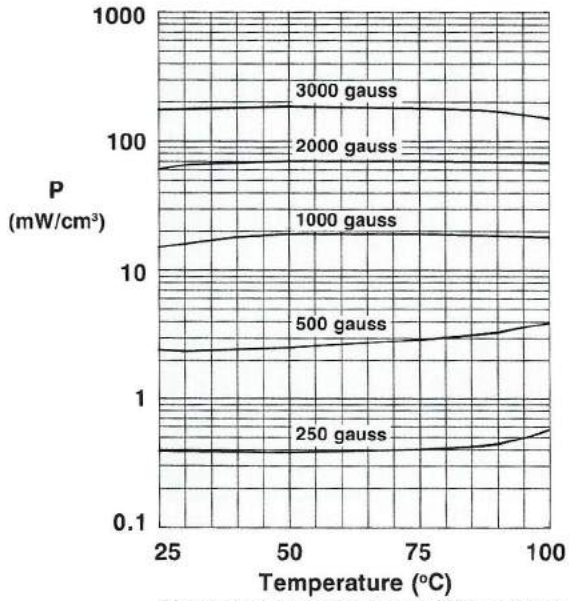
Available Core Shapes
Pot cores, EP cores, PQ cores, ETD cores, E & I cores, U cores, discs, slugs, toroids, bobbins

Power Loss Density vs. Flux Density



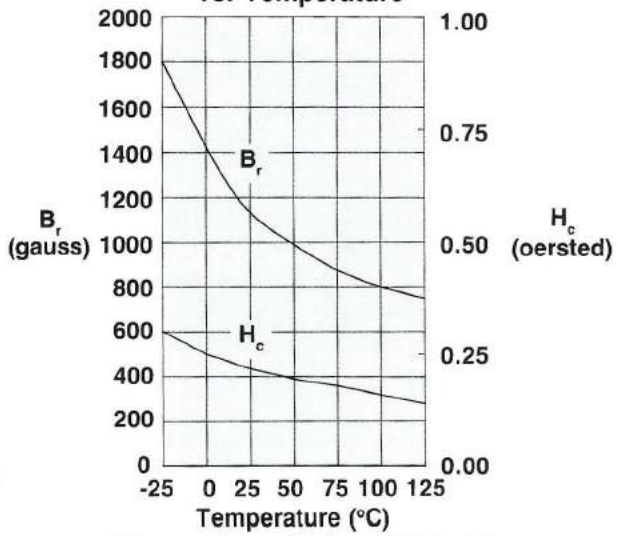
Measured on a **25.4mm** OD toroid using a Clarke Hess 255 VAW at 100°C.

Power Loss Density vs. Temperature



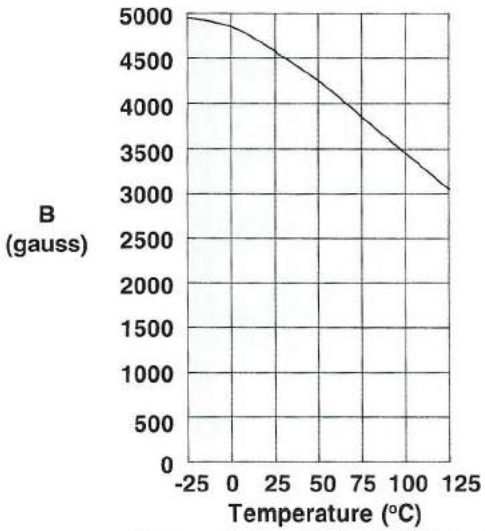
Measured on a **25.4mm** OD toroid at 10 kHz using a Clarke Hess 255 VAW.

Remanence & Coercive Force vs. Temperature



Measured on a **25.4mm** OD toroid.
H = 10 oersted.

Flux Density vs. Temperature



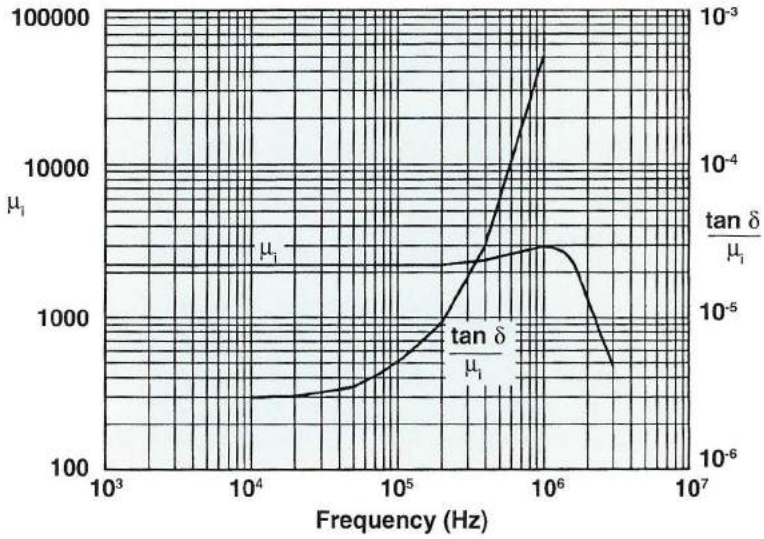
Measured on a **25.4mm** OD toroid.
H = 10 oersted.

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78 Material

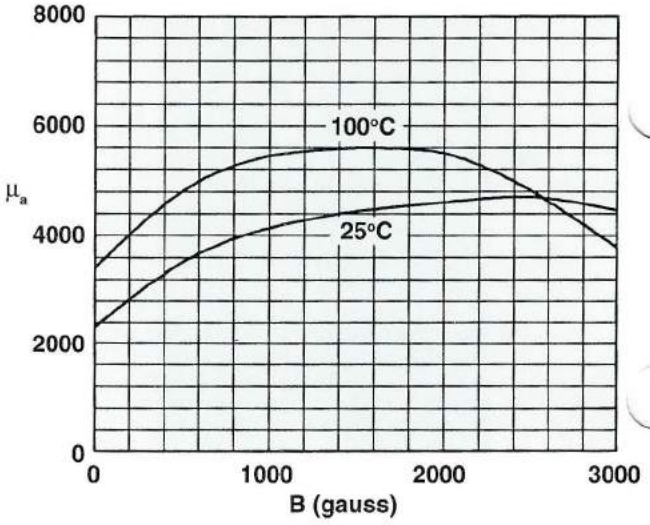
Primary Characteristics
Low core loss at high temperatures and flux densities for frequencies up to 200 kHz.
High amplitude permeability at high temperatures and excitation levels.
Curie temperature >200° C.

Initial Permeability & Loss Factor vs. Frequency



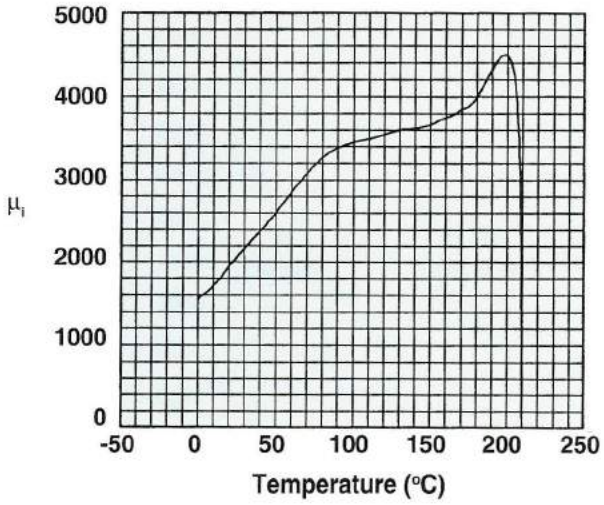
Measured on a **25.4mm** OD toroid using a HP 4275A.

Amplitude Permeability vs. Flux Density



Measured on a **25.4mm** OD toroid at 10 kHz.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

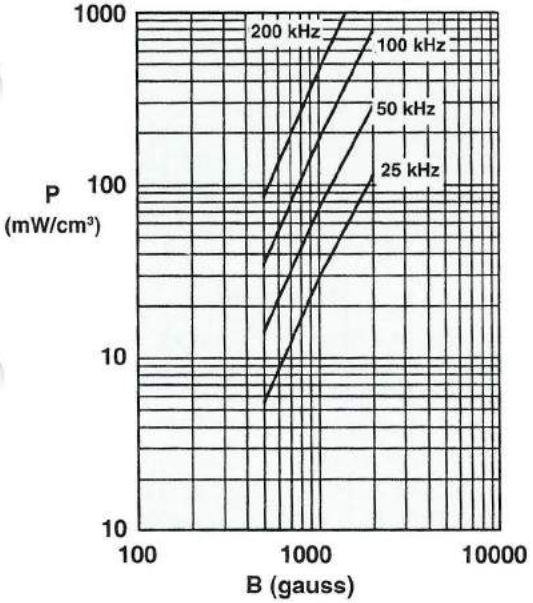
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78 Material

Applications
Transformers and inductors for high frequency switched-mode power supplies and DC converters.
High frequency electronic ballasts for fluorescent lighting.

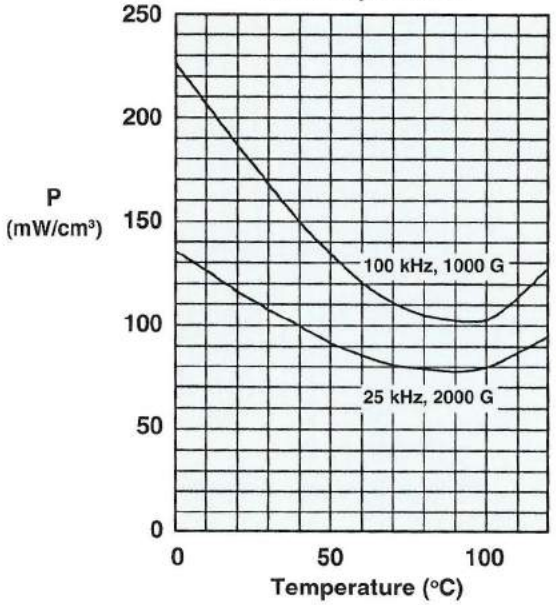
Available Core Shapes
Pot cores, EP cores, PQ cores, E & I cores, ETD cores

Power Loss Density vs. Flux Density



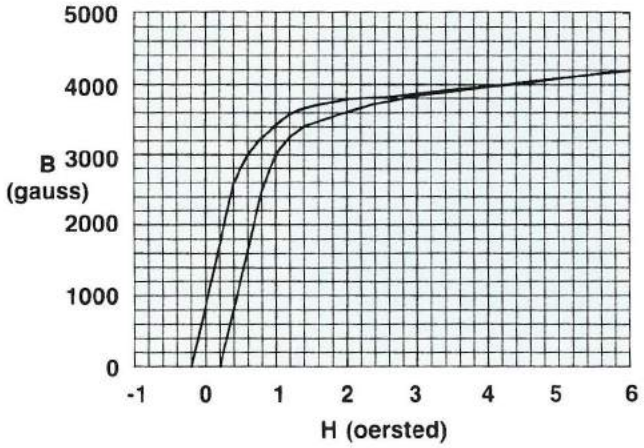
Measured on a **25.4mm** OD toroid using a Clarke Hess 258 VAW at 25° C.

Power Loss Density vs. Temperature



Measured on a **25.4mm** OD toroid using a Clarke Hess 258 VAW.

Hysteresis Loop



Measured on a **25.4mm** OD toroid.

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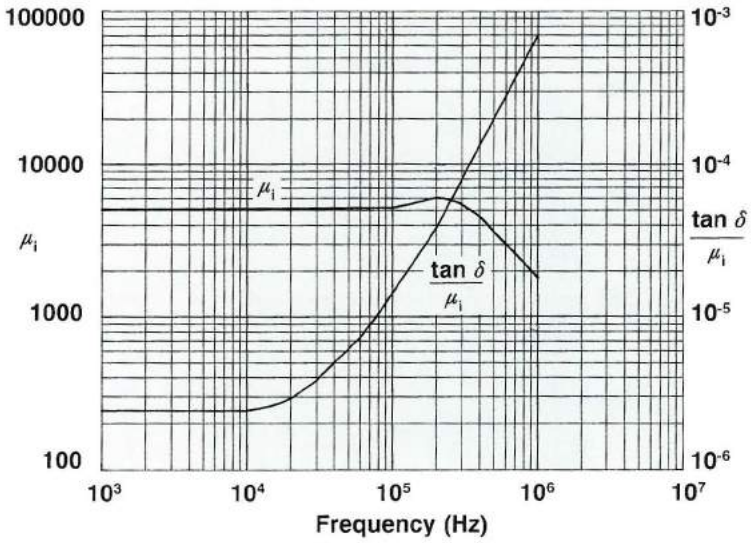
75 Material

Primary Characteristics
High initial permeability
Low core loss

Applications
Pulse and broadband transformers
Common-mode power chokes
Ground-fault interrupters

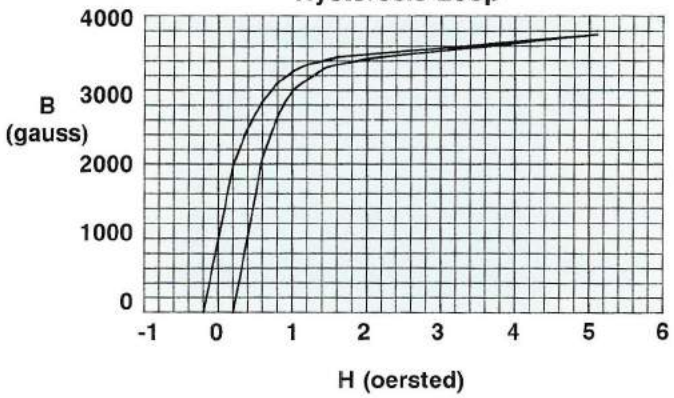
Available Core Shapes
Toroids, E & I cores, EP cores

Initial Permeability & Loss Factor vs. Frequency



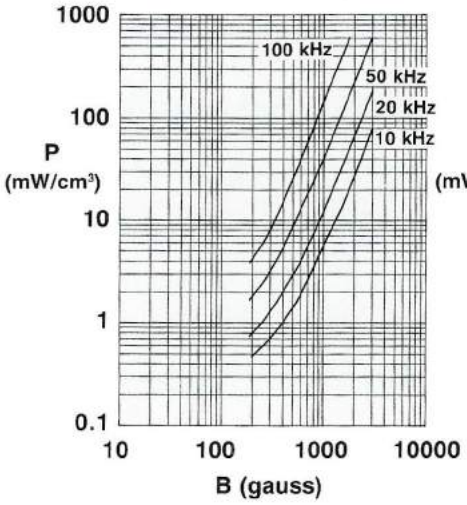
Measured on a **25.4mm** OD toroid using a HP 4275A.

Hysteresis Loop



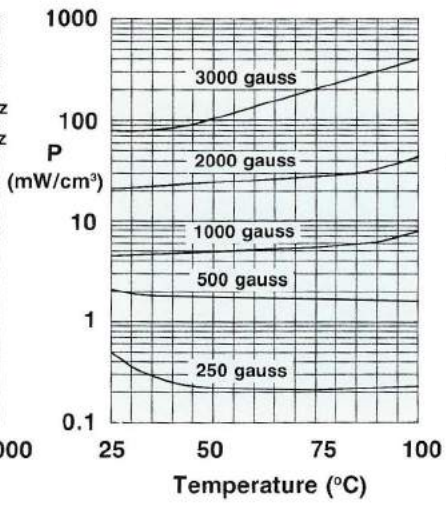
Measured on a **25.4mm** OD toroid.

Power Loss Density vs. Flux Density



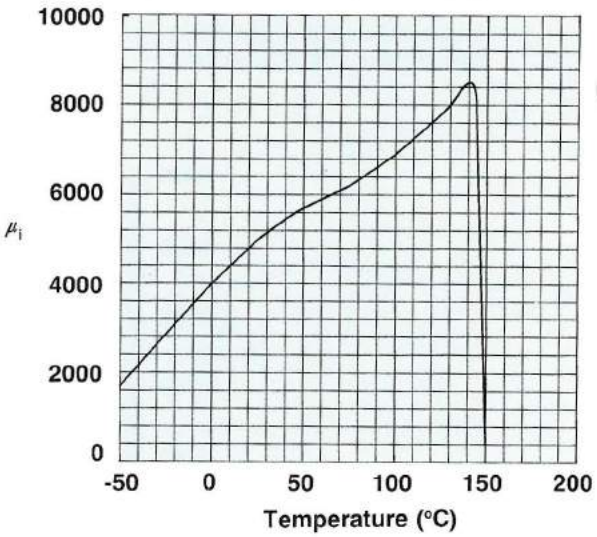
Measured on a **25.4mm** OD toroid using a Clarke Hess 255 VAW.

Power Loss Density vs. Temperature



Measured on a **25.4mm** OD toroid at 10 kHz using a Clarke Hess 255 VAW.

Initial Permeability vs. Temperature



Measured on a **25.4mm** OD toroid at 100 kHz using a HP 4275A.

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