A NEW HIGH SATURATION FERRITE INCREASES SWITCHMODE INDUCTOR ENERGY DENSITY

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Abstract

The energy storage capability of an inductor is a maximum when core flux and winding current limits are met simultaneously, and is proportional to the maximum core flux density. The core flux is limited by saturation at low frequencies or "dc", and by "hysteresis loss" at high frequencies.

MMG has recently developed a new high saturation ferrite to help fill the performance gap between ferrites, with low hysteresis losses and a moderate saturation flux density, and various "distributed gap" materials, which have a higher flux capability but much higher losses.

Introduction

The low cost, high efficiency and high power density of switchmode power supplies (SMPS) has caused them to become the predominant power conversion technology in electronic equipment today. Nonetheless, there remains an ongoing pressure to increase power densities yet further (as well as to raise efficiency and, of course, reduce costs), particularly in portable applications.

Filter inductors or "smoothing chokes" are typically the largest component in non-isolated power converters, and may be second in volume only to transformers in isolated converters and the bulk energy storage capacitors in off-line power supplies.

It is important to make filter inductors as small as possible in the drive for higher power densities. This may be accomplished in either of two ways: maximizing energy density with an optimal design on the one hand, and reducing the *required* inductance (and energy stored) on the other. The second approach involves either operating at higher switching frequencies, and/or allowing higher ripple currents in the inductor, either or both of which reduce filter inductance.

Maximizing Inductor Energy Density

Inductors alternately store and return energy in the form of magnetic fields to function. This energy may be located in several physical regions, as sketched in Fig.1:

1) A ferromagnetic core;
2) One or more "air gaps" in the core;
3) Magnetic fields within the winding;
4) "Fringe fields" near any core air gaps.

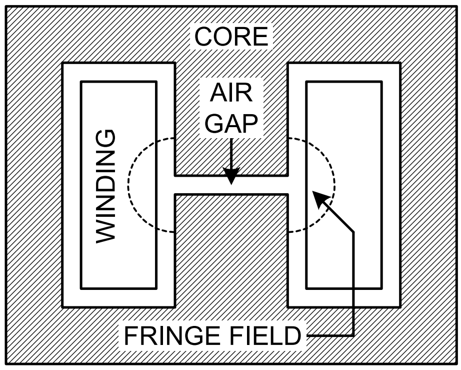


Fig. 1 INDUCTIVE ENERGY STORAGE LOCATIONS

The relative contributions of the various fields to the total energy stored vary considerably with the size of any air gaps, which will be discussed further below.

I have noted some confusion among non-specialists about two common measures of "magnetic field" strength; the magnetomotive force (MMF) field H, and the flux density or B field. An electrical analogy may help. The H field is measured in A/m (amps/meter) along a magnetic path, and the resultant flux density unit is the Tesla. The H field is analogous to a V/m (volts/meter) field, and the B field is analogous to the resultant current density (A/m^2) when the voltage field is applied to a conductive medium.

The energy density W in a magnetic field is proportional to the product of the flux density B and the magneto motive force (MMF) field H, or:

$$W = B \times H \qquad (1)$$

The electrical analog is the power loss (in W/m^3) in the conductive medium, which depends on the product of the voltage gradient and the current density.

The relative magnetic permeability μ of a medium is given by:

$$\mu = B/H \qquad (2)$$

where μ is analogous to electrical conductivity. This expression may be rewritten as:

$$H = B/\mu \qquad (3)$$

Combining equations (1) and (3) yields another useful expression for energy density:

$$W = B^2/\mu \qquad (4)$$

In other words, *the energy stored in a magnetic field is proportional to the square of the flux density divided by the permeability*. (Dimensionless units are used in the above for simplicity.)

Thus, <u>for a given flux density B,</u> the energy density is relatively low in a high permeability material like ferrite (and ferrite will be the assumed ferromagnetic core material hereafter) which typically has a μ several thousand times higher than that of 'air' or other "non-magnetic" material.

Inductance L (in Henrys) is a measure of the energy (in Joules) stored in magnetic fields due to current flow:

$$W = ½ I^2 L \qquad (5)$$

An inductor with a solid ferrite core has a high inductance, but stores little energy by the time the core begins to saturate (at several hundred mT) with a relatively low current in the winding around the core.

When a small air gap is placed in the core, the flux density in the gap is essentially the same as that in the core, but the energy density is several thousand times higher because of the low permeability. Thus even a small air gap will increase the energy storage significantly before the core saturates; ideally, the energy storage capability increases in proportion to the length of the air gap.

In practice, the flux "fringes out" from the core near an air gap due to the 'MMF drop' across the high reluctance (low μ) air gap, as illustrated in Fig. 2. (Magnetic reluctance is analogous to electrical resistance.)

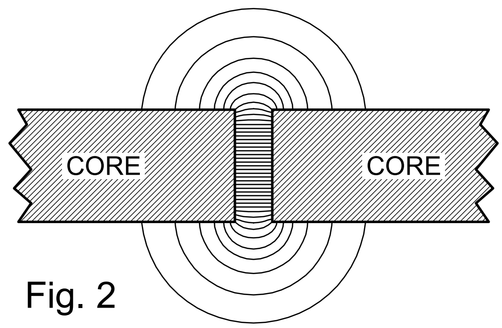


Fig. 2

FRINGE FIELD NEAR AN AIR GAP

The effective flux 'area' is increased at the gap, lowering flux density in the air gap and the magnetic reluctance of the gap. The 'effective' air gap length is thus less than the physical gap length. This effect increases with gap length until the gap is about ¼ of the winding width (in an E-E cored inductor), after which the gap field becomes more uniform again.

For a constant flux density in the core, the energy stored in the air gap increases with gap length while the energy in the core remains essentially constant. As the air gap length becomes greater than about 0.1% of the magnetic (core) path length "le", there is progressively more energy stored in the air gap than in the core itself. This effect is shown in Fig. 3 for a typical inductor constructed with E-E cores and a single, centrally located air gap in the center leg.

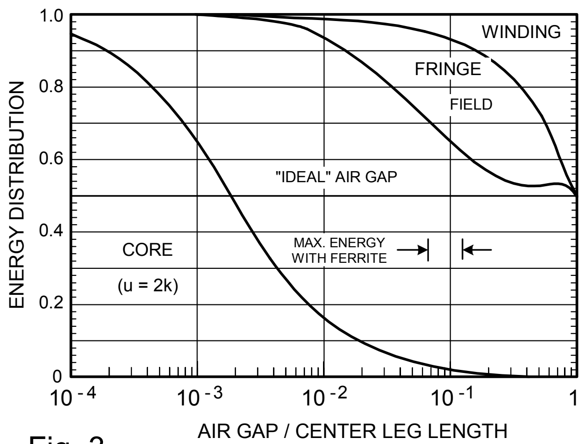


Fig. 3

RELATIVE ENERGY
DISTRIBUTION v. AIR GAP

A significant portion of the total energy also becomes stored in the fringe field as the gap becomes larger than a few percent of the core length. This effect disappears as the gap length approaches the width of the winding, but the relative energy in the winding field is increased. With an air gap equal to the winding width, roughly half the total energy is stored in the winding and half in the 'air gap'.

This increase in energy storage with an air gap is why core air gaps are used. In practice the energy cannot be increased arbitrarily with larger air gaps, however. The winding turns (and/or current) must be increased to maintain the flux density in the core as the air gap is increased, which increases the winding I^2R losses. At some air gap length the winding temperature will reach the maximum allowable, and the turns (or current) can no longer be increased. With still larger air gaps the flux density (and energy stored) will now decrease with gap length, as shown in Fig. 4.

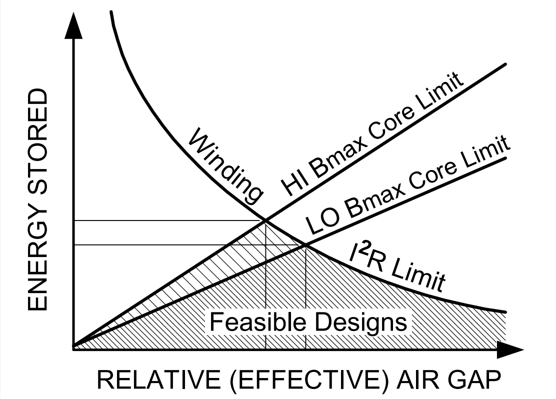


Fig. 4 INDUCTOR ENERGY LIMITS v. AIR GAP

It can be seen from Fig. 4 that a maximum of energy is stored in an inductor when limits on core flux and winding current are reached simultaneously [1]. Core flux may be limited by the saturation flux density Bsat (as in the case of "dc" filter inductors where ac ripple currents are relatively low) or by hysteresis losses with large or exclusively ac current (as in resonant L-C circuit inductors). Winding currents are *always* limited by I^2R losses which may, however, include various eddy current losses at high frequencies.

For ferrite cores used in 'dc' filter inductors, the energy limit occurs with an air gap of roughly 3% of the magnetic path length (or about 10% of the center leg (and winding) width with E-E cores), as noted in Fig. 3. With this "maximum energy" air gap it can be seen that over 90% of the energy is stored in the magnetic field of the gap and nearby fringe field; only 1% to 3% is stored in the core, and 3% to 10% within the magnetic field of the winding.

With an air gap smaller than 'optimum' the maximum core flux is reached while the winding is relatively cool, while a larger than optimum gap causes the winding to overheat before maximum core flux is reached.

The benefits of a higher core flux density capability are also illustrated in Fig. 4. For a given winding and I^2R limited current, a higher core flux allows a correspondingly smaller air gap. A smaller gap increases inductance, and hence energy storage. For dc filter inductors the energy density is proportional to the saturation flux density.

Historically most power ferrites have been limited to a Bsat of about 500 mT at room temperature, decreasing to less than 400 mT at 100 ºC and 300 mT at 150 ºC.

The saturation flux density of the new F49 material is on the order of 100 mT higher than conventional ferrites. This is shown in Fig. 5, where Bsat for F49 is compared to that of a typical power ferrite at an H field of 1200 A/m (15 Oe). The Bsat of F49 with lower H fields for several temperatures is given in Fig. 6.

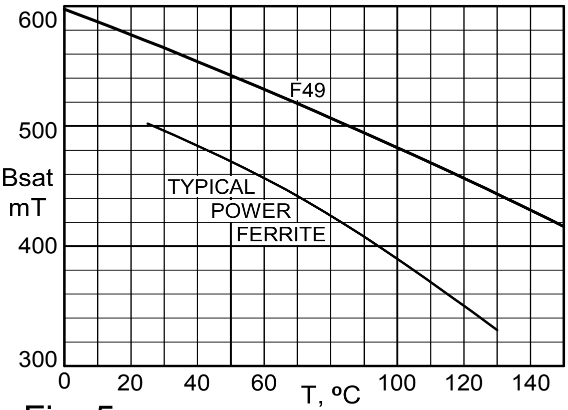


Fig. 5 FERRITE Bsat v. TEMP. COMPARISON @ 1200 A/m

For a given ratio of maximum operating to saturation flux density, a dc inductor using F49 core material can store about 16% more energy at 60 ºC, 23% more at 100 ºC, and nearly 40% more at 140 ºC than with a typical power ferrite.

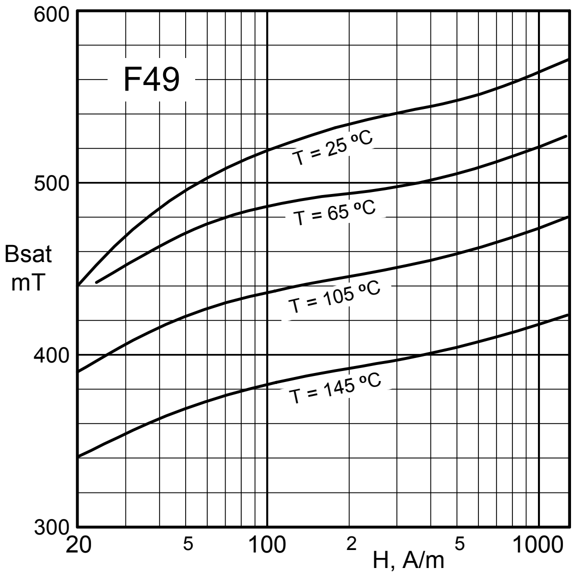


Fig. 6 Bsat v. H and T

Alternatively, the physical size can reduced correspondingly for the same inductance and energy storage.

Basic dc Inductor Design

The following is my personal approach to designing a dc inductor, a simple if iterative process. First, assuming the design "knowns" of:

L = Desired Inductance (Henrys)

I = Maximum Current (Amps)

B = Maximum Core Flux Density, (Tesla)

Then for a "candidate" core where:

A = Area of Core (m^2)

The required number of turns N is simply:

$$N = I\,L\,/\,B\,A \qquad (6)$$

A "safety factor" is applied to Bsat to obtain the working maximum flux density, depending on the degree of control of the worst case current on the one hand, and the detrimental effects of core saturation on the other. With peak current mode control (essentially instantaneous cycle-by-cycle current limiting) I feel it is safe to operate to 75% to 80% of Bsat at the maximum operating temperature. At a core temperature of 100 ºC, this is about 300 mT for a typical power ferrite, or 380 mT for the F49 material. Other designers are more comfortable with a slightly lower value, perhaps 70% of Bsat.

The next step is to check the winding losses. The "best case" dc winding resistance R, for magnet wire, can be calculated from:

$$R = 2.2 \times 10^{-4}\,(N^2)(lmt)(K1)(K2)/(Aw) \qquad (7)$$

Where: lmt = length of mean turn (m)

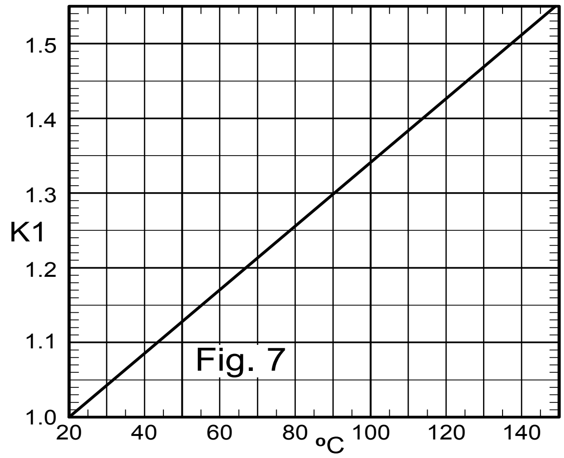
Aw = Winding area (m^2)

K1 = Temperature Correction Factor

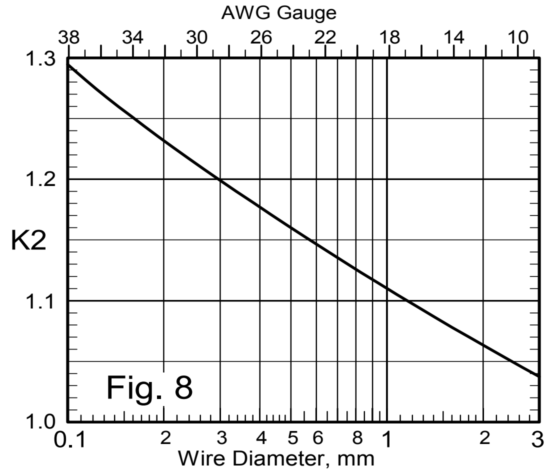
K2 = Insulation Correction Factor

This formula assumes a perfect square array of round wire (exact fit, no partial layers, etc.). The values for lmt and Aw can often be found in data books for core bobbins, and K1 and K2 are given in Figs. 7 & 8.

The power dissipation in the winding can now be calculated from the resistance and the dc current. If the dissipation is too high a larger core must be tried, or a smaller core can be used if the winding temperature rise is well below the maximum allowable.



TEMPERATURE CORRECTION FACTOR



INSULATION CORRECTION FACTOR

The expected temperature rise for a given dissipation/area with natural convection cooling is given in Fig. 9 for assistance. Whether only the exposed winding area or the total core and winding area can be used for heat dissipation depends on how well the core and winding are thermally coupled.

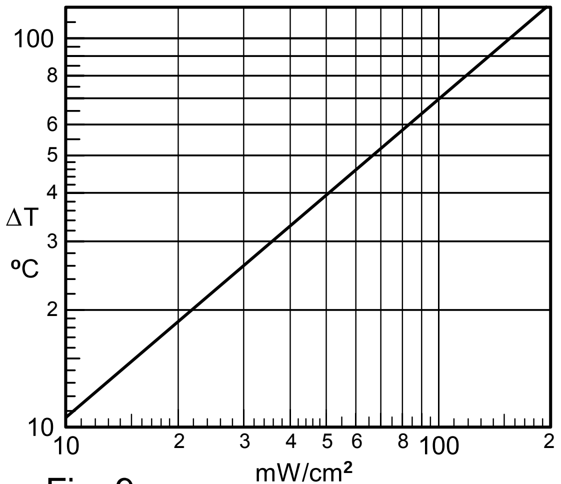


Fig. 9 TEMPERATURE RISE WITH NATURAL CONVECTION COOLING

Once all design criteria are met, the core is gapped to provide the design inductance. For typical power inductors the theoretical air gap "a" may be calculated as a guide from:

$$a = 1.257 \times 10^{-4} N^2 A/L \qquad (8)$$

If this gap is used, it will be found that the inductance is typically 20% to 40% higher than calculated. This is not a "free" inductance, but results in a correspondingly higher flux density in the core. For a single air gap, the physical air gap will typically need to be on the order of 40% to 50% larger than the theoretical gap to compensate for fringe field effects. The actual air gap to use is best determined empirically.

Design of ac Inductors

The design of "ac only" inductors is similar to that for dc inductors except that the peak flux becomes limited by loss above some frequency, typically around 10 - 20 KHz. As a rough guide, a core loss of around 300 KW/m^3 (or mW/cm^3) is allowable for small inductors (a few cm on a side), dropping to 100 KW/m^3 or so as dimensions reach the 10 cm range. Peak flux may be calculated from:

$$B = I L / N A \qquad (9)$$

Where "I" is the peak current, or from the sine wave voltage Erms and frequency F (in Hz) by using the "transformer" design equation:

$$N = Erms / 4.44 F A B \qquad (10)$$

If the applied voltage is not a sine wave, use the average voltage Eave and change the 4.44 in the denominator to 4.

Curves of core loss as a function of frequency and flux density are provided for all "power" ferrites. A partial "sample" plot for F49 at 65 ºC is given in Fig. 10, with characteristic plots of loss v. temperature in Fig. 11. The higher Bsat of this new ferrite will only be of benefit in "ac only" inductors below about 10 to 15 KHz. (More detailed plots of these and other F49 characteristics will be available at the website www.magdev.co.uk.)

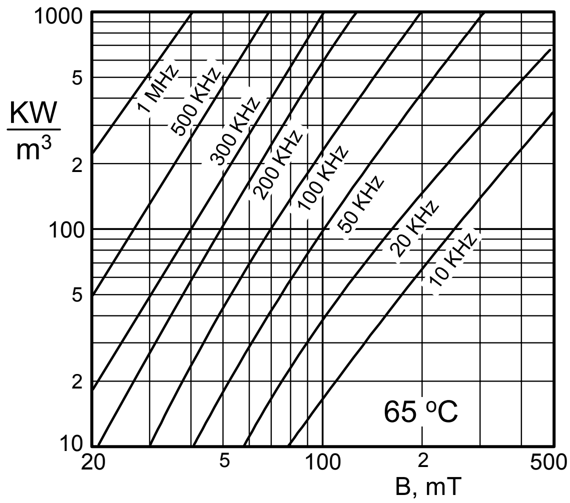


Fig. 10 F49 CORE LOSS

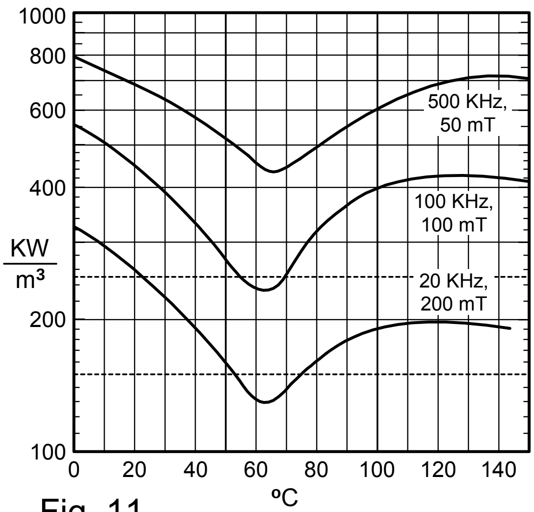


Fig. 11 F49 LOSS v. TEMPERATURE

Minimizing Filter Inductance

As noted earlier, increasing the switching frequency while maintaining a constant ripple current, or increasing ripple current at a given frequency, reduces the inductance and energy storage requirements but increase the ac hysteresis loss in the core. (Eddy current losses in the winding are also increased, but that is beyond the scope of this paper.) At some point the increasing core losses begin to limit core flux density, and an optimal design must take this into account.

It is generally assumed that the presence of dc flux in a ferrite core has no impact on the ac losses. Unfortunately, this is only approximately true for moderate dc flux densities, where the dc flux is about ¼ of Bsat. At higher dc flux the ac losses in ferrites begin to increase significantly, and I am unaware of any useful information or data published on this effect.

(Single-ended forward and flyback converter transformers also normally operate with a dc flux component, typically half the peak-peak ac flux swing. However, the ac core losses usually limit flux density to levels where the dc component does not raise the ac losses significantly.)

In dc filter inductors, however, the dc flux is pushed well beyond ½ of Bsat, and the hysteresis losses due to ripple current may be much higher than expected, which must be taken into account in design.

The effect of a high dc bias flux on ac losses have been measured for MMG's new high Bsat F49 material. A sample of the results is presented in Fig. 12 for 25 KHz and 60 ºC, and in Fig. 13 for 200 KHz and 60 ºC. As noted, the complete test results will be available on their website.

Design of High Ripple dc Inductors

The design of high ripple current dc filter inductors is similar to that for conventional inductors, using a peak current equal to the maximum dc current *plus* half the worst case peak-to-peak ripple current in equation (6) to calculate the peak flux for saturation avoidance.

The ac flux in the core is then calculated from half the peak-to-peak (p-p) ripple current using equation (9). Half the p-p ripple current is used, as core losses are given as function of the peak ac flux, which is half the p-p flux swing. The same formula provides the dc flux when the dc or 'average' current is used.

The ac core loss is then determined from the core volume and plots such as those in Figs. 12 and 13. Winding losses are added to core losses to obtain total power dissipation, and the temperature rise is estimated from the exposed cooling area (using Fig. 9 for natural convection cooling).

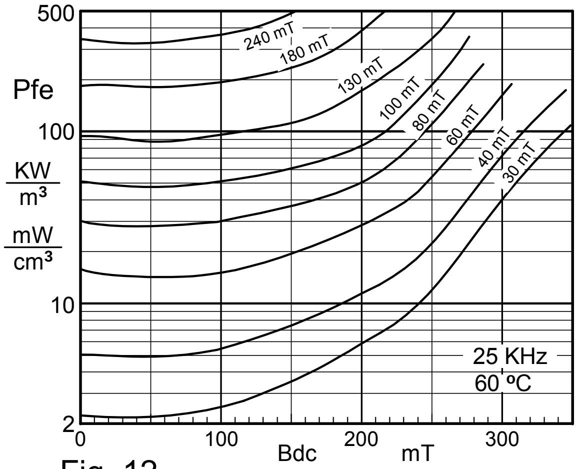


Fig. 12

F49 Core Loss v. dc Flux

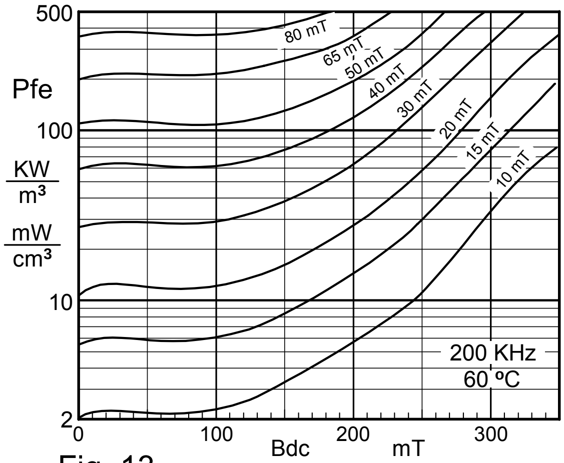


Fig. 13

F49 Core Loss v. dc Flux

As before, if the temperature rise is too high, a larger core is generally required for the next iteration, while a smaller core may be attempted if the temperature rise is significantly lower than that allowed.

Note that the F49 ferrite was designed as a high Bsat ferrite, and not as a minimum loss ferrite. If the ac core losses limit the peak flux to significantly less than about 70% of Bsat, a lower loss "power" ferrite such as MMG's F48 material should be considered, which has losses from ½ to 10% of those for F49, depending on frequency, flux density and temperature. Unfortunately, the required information on ac losses under high dc flux bias is not yet available for most power ferrites.

When ac core losses from ripple currents are high enough to be a design problem, it is probably also necessary to consider skin and proximity effects in the windings, and additional eddy current losses induced in the winding from the fringe field near air gaps [2]. As a guide, these losses begin to become significant when the wire diameter becomes greater than about one skin depth for a single layer winding, dropping to about 0.4 skin depths in diameter for a 10 layer winding. The skin depth δ (in m) for copper at 100 ºC at frequency F (Hz) is given by:

$$\delta = 7.6 \text{ x } 10\text{-}2 / \sqrt{F} \qquad (11)$$

Litz wire or foil windings may have to be used to keep the wire strand diameter (or foil thickness) small enough to minimize skin and proximity effects in the windings. Litz wire is generally preferred when a larger number of turns (more than about 20 or 30) are required, while foil windings are suitable for a relatively low number of turns with high currents. Multiple discrete gaps under the winding can also be used to minimize the excess winding losses due to fringe fields near air gaps [2].

Conclusions

A new high Bsat ferrite has been developed by MMG which can significantly reduce the size of dc filter inductors. The ferrite may also prove advantageous for low frequency “resonant” inductors, and for high ripple dc inductors when the ripple currents are not too high. When ac core losses limit the peak flux to well below saturation, however, a lower loss ferrite may be the better choice.

Disclaimer

Information on the F49 ferrite characteristics presented here was measured by the author (based on samples provided by MMG), who is solely responsible for their accuracy.

The hysteresis loss measurement technique used was originally designed for ungapped ferrite cores [3]. In essence, a ‘hyper-efficient’ ZVS (zero voltage switching) half bridge ‘chopper’ is used to drive the core, such that core losses can be largely measured as the dc input power to the chopper. In this application the drive winding I^2R (and other spurious) losses included in the measurement are an order of magnitude less than the core losses being measured, and relatively simple corrections can be made to improve the accuracy of the loss measurement to well less than 5%.

However, it is necessary to place one or more air gaps in cores to be tested with a known dc flux in order to establish a known relationship between the dc H and B fields. If this is not done, the variation of core μ with H and temperature, and any residual core flux, make the measurements meaningless.

Unfortunately, the introduction of core air gaps also increases the magnetizing current and associated losses appreciably. First order corrections were made for this and other known spurious losses of the technique, but measured losses are still likely to err on the high side to an unknown extent.

Improved techniques are under research and development for a more accurate measurement of ferrite ac losses in the presence of dc flux. Parties interested in the measurement of such losses should contact the author for more information.

References

[1] B. Carsten, tutorial “Design Considerations for High Frequency Linear Magnetics”, presented at conferences and other venues.

[2] B. Carsten, tutorial “High Frequency Conductor Losses in Switchmode Magnetics”, presented at conferences and other venues.

[3] B. Carsten, “Why the Magnetics Designer should Measure Core Loss; with a Survey of Loss Measurement Techniques and a Low Cost, High Accuracy Alternative”; Proc. of HFPC’95 Conference, pp. 103-119, May 7-11, 1995, San Jose, CA; Proc. Of PCIM’95, pp. 163-180, June 20-22, Nurnberg, Germany.