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# INDUCTOR DESIGNS FOR HIGH FREQUENCIES

## Powdered Iron "Flux Paths" can Eliminate Eddy Current 'Gap Effect' Winding Losses

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## INTRODUCTION

There are two inductor constructions in principle use for switchmode converters; those which used gapped ferrite cores (typically “E” cores), and those with toroidal cores of ‘distributed gap’ materials. Both types work well for general DC filtering. Gapped ferrite cores yield a more stable permeability and have a wider range of available core shapes, while distributed gap cores can tolerate a higher flux and have a soft saturation characteristic. ‘Powdered iron’ toroidal cores are particularly favored for cost sensitive applications. However, problems arise with either with the presence of significant HF AC current.

The toroidal cored inductor is often thought to have zero external magnetic field, but this is only theoretically true for a perfectly uniform winding covering the whole core. In practice, a high permeability core with a discrete air gap underneath the winding will typically have a much lower external field, and thus lower propensity for creating EMI. (However, an external air gap in a ferrite core, such as produced by spacing ungapped “E” cores or other “open” constructions, creates a very large external magnetic field.) The hysteresis losses in distributed gap materials are also significantly higher than in ferrites, which thus appear to have the edge.

The gapped ferrite core has an Achilles heel, however; the ‘fringe field’ around the air gap induces excess eddy current losses in the winding, which can exceed all other losses combined in some cases. The problem is related to the external field created in the toroidal inductor with partial or non-uniform winding coverage; there is a geometrical mismatch between the amp-turn or H-field generator (the winding) and the amp-turn or H-field adsorber (the high reluctance portion of the core). This mismatch creates a non-uniform H-field which results in excess winding losses and/or flux external to the inductor.

The same effect occurs in transformers when the primary winding (amp-turn generator) and the secondary winding (amp-turn absorber) do not cover the same portion of the core; distorted fields increase eddy current losses and leakage inductance. In effect, the high reluctance portion of an inductor's core (the discrete or distributed air gaps) serves an identical function to the secondary in a transformer. The lowest external fields and losses occur when the amp-turn generator and absorber cover the same portion of the core.



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## LOW PERMEABILITY FLUX PATHS

The ideal inductor with a solenoidal winding would have a uniform high reluctance underneath the winding, and a low reluctance (high permeability) core in the rest of the magnetic path external to the winding [1], [2]. Whereas distributed gap cores have typical permeabilities in the range of 20 to 100, the solenoidal winding inductor requires a permeability of about 10 underneath the winding for maximum energy storage (see Appendix).

The beneficial effects on solenoidal winding flux of replacing a discrete air gap with wider sections of low permeability material are graphically illustrated in Figure 1. These figures were created by Ansoft Maxwell FEA software for a 30/19 pot core inductor. The winding was assumed to be a solid conductor for simplicity of modeling, conducting a low frequency AC current. The winding is much less than a skin depth thick, minimizing eddy current redistribution of the winding current. The pictures show the rectangular winding within the winding window, with the flux lines within the winding and window and in the nearby high reluctance portion of the core. The rest of the core and its flux is not shown.

Fig. 1a shows the highly non-uniform flux distribution and concentration near a discrete air gap which is 1/10 of the winding in width. In Figs. 1b - 1f the high reluctance flux path width “a” is increased in steps up to the winding width “W”, while increasing the permeability by the same factor to maintain a theoretically constant reluctance. The flux becomes increasingly more uniform and lower in intensity, until it is nearly axial with only a little divergence at the ends of the winding. In Fig. 1g the flux path was widened to the full width of the window, which causes the flux to diverge further away from the core near the ends of the winding.

## WINDING LOSS BENEFITS

Although the winding is modeled as a single solid conductor, the relative induced losses from the core gap in other winding configurations can be estimated by the excess eddy current losses due to the high reluctance flux path. Eddy current resistance (Rec) is found by subtracting the DC (Rdc) resistance from the AC resistance (Rac). This is then compared to the Rec for a uniform flux (found by placing the ferrite return path on the ends of the winding with no magnetic material inside the winding) to obtain the induced “gap loss” resistance (Rgl).

Although this information was available from the higher frequency modeling done for the cores of Fig. 1, the inductance dropped from 176 nH for the narrow air gap to 145 nH for the full width flux path due to flux fringing. It was felt desirable to hold the inductance constant by adjusting the high reluctance flux path width to obtain a more meaningful loss comparison.

A second set of calculations were made where the inductance was held to 145 nH (±0.25%) as the flux path permeability was changed. The resistivity of the solid winding was set to 3.16 micro-ohm-cm, to give a DC resistance equivalent to that of a square array of insulated round wire of the same winding cross section operating at 100 deg. C. Winding AC resistance was modeled for frequencies of 163, 320, 650 and 1,280 Hz, for which the winding thickness/skin depth ratios were 0.5, 0.7, 1.0 and 1.4 respectively. Eddy current losses could not be accurately calculated at lower frequencies, as the loss decreases as the square of frequency. At higher frequencies the current distribution is becoming increasingly distorted in the solid winding, and it was felt the results would be of decreasing relevance to other windings.

The gap induced eddy current losses Rgl (relative to uniform flux) and flux path permeability are plotted in Figure 2 as a function of flux path/winding width ratio a/W. It can be seen that the excess loss ratio is similar for the four frequencies, decreasing dramatically as the flux path approaches the winding width.

Somewhat surprisingly (at least initially to me) the excess loss actually becomes somewhat negative, and continues to decrease as the flux path becomes wider than the winding width. The reason for this becomes evident on inspection of Fig. 1f and 1g; the flux density becomes lower on the ends of the winding than occurs with uniform flux.

There appears to be a discrepancy between this result and a earlier statement that the inductor's high reluctance flux path has the same function as a transformer secondary. Transformer winding losses are a minimum when primary and secondary are of equal width, implying that the inductor winding losses should be a minimum when the flux path has the same width as the winding. However, the flux path does not have the equivalent of secondary winding eddy current losses to increase with excess width, so the inductor winding losses apparently actually decrease with a slightly wider flux path. This is likely true for real wire windings, but I would expect it not to hold for foil windings where a maximally straight flux (with a/W = 1) should give the minimum loss (see also [2]).

The computer modeling of the winding resistance assumed lossless magnetic materials, which is not the case in reality. As noted, distributed gap materials have significantly higher losses than ferrites, and this is particularly true of the higher permeability powdered iron materials commonly used for power inductors. However, unannealed carbonyl iron materials with permeabilities from about 4 to 14 have significantly reduced hysteresis losses, comparable to or lower than other distributed gap materials, and this is just the permeability range of interest for the application.

## CONCLUSIONS

There is a significant potential to reduce the eddy current winding losses in gapped core inductors by replacing the air gap with a wider flux path of a distributed gap material. This construction has been approximated in the past by using multiple discrete gaps in a ferrite core with excellent results, However, the fabrication costs of cutting and reassembling the ferrite are quite high for cost sensitive applications. The use of a powdered iron flux path beneath the winding is an attractive alternative, but even low permeability powdered iron losses are greater than in ferrite; the issue becomes whether the reduced winding losses offset the increased core losses, and if so is the reduction worthwhile. This is still an open question, but initial results seem to show a definite, and sometimes dramatic, loss reduction. (Part of the problem is that magnetic losses in these low permeability materials is difficult to measure accurately at high frequencies.) The cost increment must then be compared to other alternatives, such as using litz wire or a larger inductor to dissipate the additional heat.

My own practical experience is limited to inductors operating at 1 MHz with a single turn copper strap winding. With a 90 A P-P ripple current (26 A RMS) the loss in the powdered iron flux path was about 0.6 W, whereas the excess winding loss with a discrete gap was about 2.4 W. When combined with the DC winding loss of 0.37 W, the total loss decreased by nearly 65% with the powdered iron flux path.

BIBLIOGRAPHY

[1] B. Carsten, Tutorial; “Design Considerations for High Frequency Linear Magnetics”.

[2] N.H. Kutkut, D.M. Divan; “OPTIMAL AIR GAP DESIGN IN HIGH FREQUENCY FOIL WINDINGS”; Proceedings Of APEC'97, Vol. 1, p. 381-387; Feb. 23-27, 1997, Atlanta, GA.

APPENDIX

MAXIMIZING ENERGY STORAGE IN AN INDUCTOR

Maximum energy is stored in an inductor when maximum current density in the winding and maximum flux density in the core occur simultaneously [1]. Maximum current in the winding is always a thermal loss limitation, while core flux may be either loss or saturation limited.

Maximum inductive energy storage with an air-gapped core occurs at a unique or 'optimum' gap length, which depends somewhat on operating conditions. A smaller than optimum gap causes the core to saturate (or overheat) before the winding thermal limit is reached, while a larger than optimum gap causes winding overheating before maximum core flux is reached. Unlike most optima, the optimum gap length is very sharp; an air gap 20% too large or small causes about a 20% reduction in energy storage.

For an inductor wound on a 'distributed gap' core material (such as 'powdered iron') there would be a similar equivalent optimum permeability for maximum energy storage were it not for complicating factors. First, core 'saturation' is only a very gradual decrease in permeability with flux density, which does not place a wall defined limit on flux, and this property changes with core permeability. Second, core hysteresis loss can also vary with permeability, and not always in a simple manner. Nonetheless the maximum energy storage usually occurs with a core permeability between 20 and 100, and usually between 30 and 60.

For an inductor with a gapped high permeability core it is convenient to define an 'equivalent permeability' for the core which is the ratio of the winding width to the width of the theoretical air gap (without fringing) required to obtain the required inductance. I have found that the optimum equivalent core permeability for this arrangement is typically about 10, rather less than for a toroidal core and winding.

As an illustration, the outline of the design of an inductor wound on a gapped 30/19 pot core is used. The surface area of the core is about 32 square cm. For a 40 deg. C temperature rise, the allowed dissipation per square cm can be about 50 mW for natural convection, or 87 mW for natural convection plus black body radiation. The total dissipation thus lies between 1.6 W and 2.8 W.

For a loss limited core flux the core and winding losses would be made about equal, so the winding loss might have to be as low as 0.8 W for an inductor with significant high frequency AC. At the other extreme, a DC inductor with no core loss could have a winding loss as high as the 2.8 W limit. A third winding loss limit of 1.6 W is taken as an intermediate case.

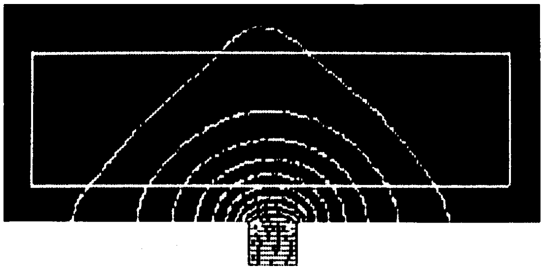
At 100 deg. C a normalized one turn winding would have a resistance of about 46 micro-ohms, allowing the (1T normalized) winding current to be 132 A, 186 A and 247 A for the low, medium and high allowed dissipation cases respectively.

If the core air gap is to limit the flux to 2000 gauss for a ferrite core operating at higher temperatures, then the (theoretical) air gap lengths are 0.086 cm, 0.117 cm and 0.155 cm respectively. For a winding width of 1.2 cm, the calculated effective permeabilities become 13.74, 9.75 and 7.34, which are neatly grouped around the value of 10.

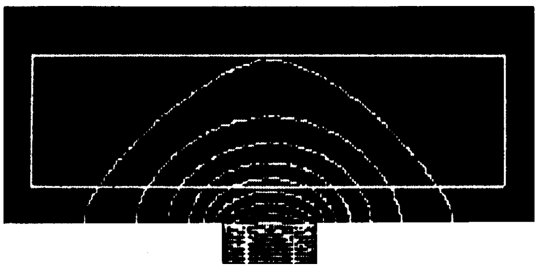
These effective permeabilities are lower than many engineers expect, and the question arises as to how realistic or representative they are. The answer is that even lower effective permeabilities are often required.

If the core flux in the above example were loss limited to less than 2000 gauss, the effective permeability would have to be lower (ie, a larger air gap). This is also true if the inductor current waveform has a higher peak/RMS ratio than a steady state DC or sinusoidal AC current. Finally, realistic scaling laws [1] show that for larger inductors the required effective core permeability will vary as the inverse square root of the linear dimensions for a saturation limited core, and about as the inverse of the linear dimensions for hysteresis loss limited cores. Thus, larger inductors require lower effective core permeabilities (or relatively larger air gaps).

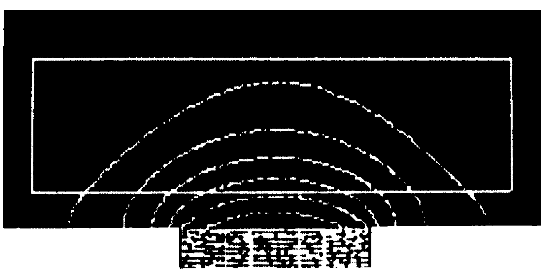
Only two situations would raise the optimum effective permeability: an operating core flux density higher than 2000 gauss, or an inductor significantly smaller than a 30/19 pot core.



(a) a/W = 0.1 μ = 1



(b) a/W = 0.2 μ = 2



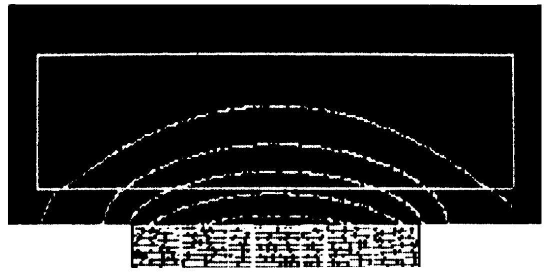
(c) a/W = 0.4 μ = 4

a = Flux Path Width
W = Winding Width
μ = Flux Path Permeability

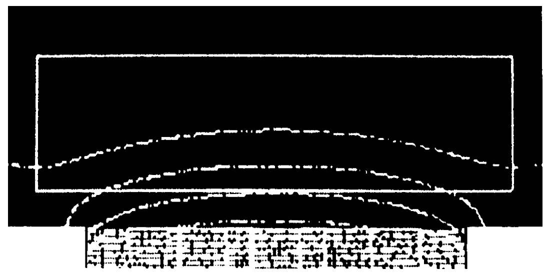
Figure 1

Winding Flux Distribution

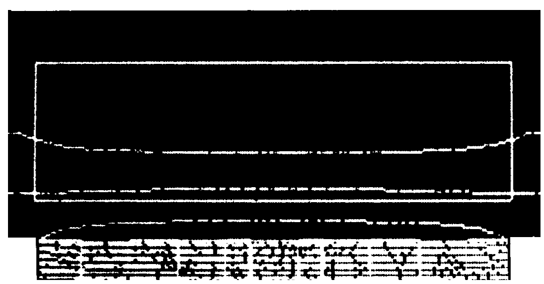
vs. High Reluctance Flux Path Width



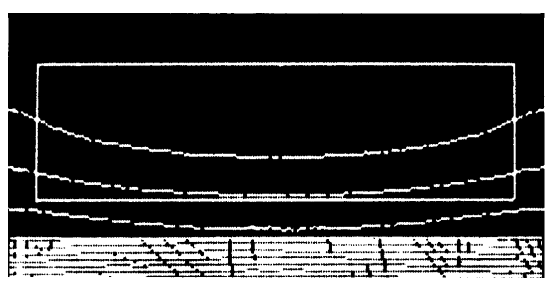
(d) a/W = 0.6 μ = 6



(e) a/W = 0.8 μ = 8



(f) a/W = 1.0 μ = 10



(g) a/W = 1.13 μ = 11.3

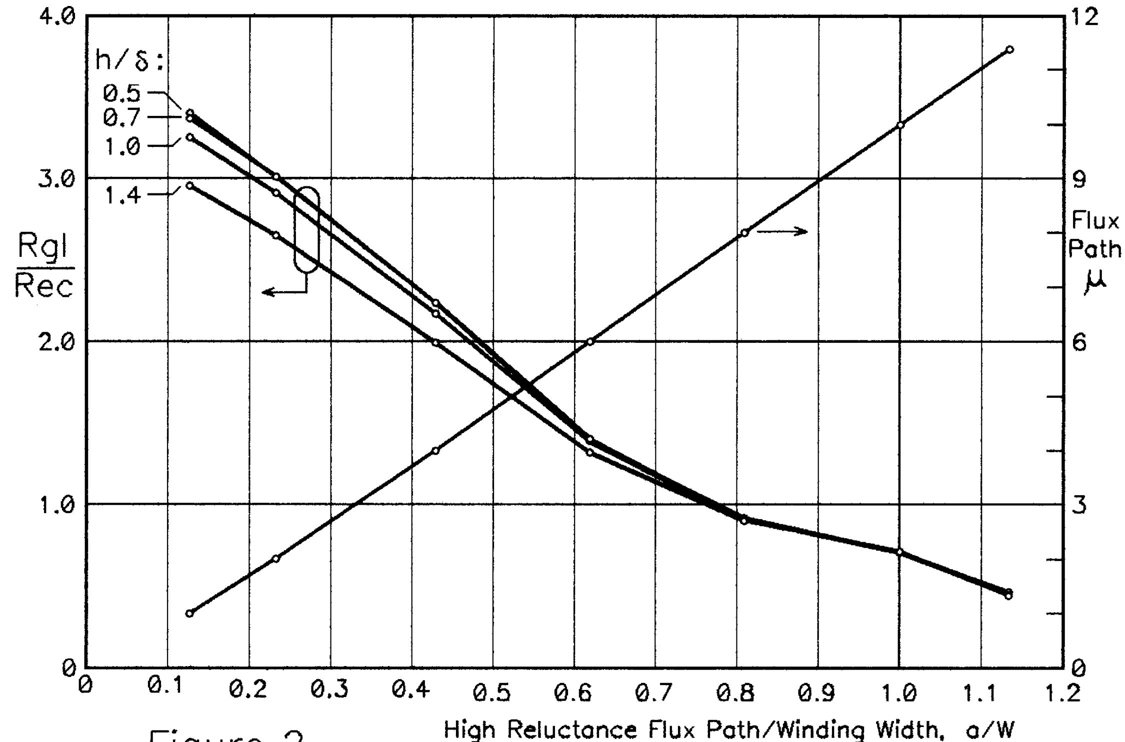



Figure 2

Induced Gap Losses (Rgl) Decrease with Increasing a/W