ADDITIONAL LOSSES IN HIGH FREQUENCY MAGNETICS DUE TO NON IDEAL FIELD DISTRIBUTIONS

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abstract- **Calculations based on Dowells analysis with extensions for non-sinusoidal current waveforms are widely used to estimate winding losses in high frequency magnetic components. This analysis, while extremely useful and informative, does not account for all losses. There can be additional losses due to gap fields, orthoginal fields at the ends of windings, shield eddy currents and proximity effects between windings with half wave currents. These additional losses can be significant and must either be minimized or at least taken into account in the design of the magnetic component.**

Determination of these losses using classical analysis techniques is usually not practical. CAD software employing finite element analysis is however, very practical. In addition to giving the values for the additional loss, finite element modeling can also display the fields and current distributions which are the sources of the losses. This paper discusses a number of loss mechanisms.

I. INTRODUCTION

Dowells analysis [1] and its various extensions [2,3] is widely used to calculate the power loss due to the currents in the windings of magnetic components. This loss calculation is especially important as switching frequency or power level is increased. These analyses all make a very basic assumption regarding the magnetic field distribution in the winding: the field is assumed to be the same as an infinite solenoid with all field components parallel to the winding layer. It is assumed that no orthoginal field components are present to introduce additional eddy currents which generate additional losses. While this assumption greatly facilitates the analysis and accounts for what is normally the major loss mechanism, in practice it is not complete. Orthoginal fields, arising from several causes, do exist in real world magnetic structures. In many cases it is necessary to take into account the additional losses due to the non-ideal field distributions.

In most situations the direct analytical calculation of the losses due to non-uniform fields is not practical. At least not for the average designer working under a deadline. However, finite element modeling software is readily available from a number of vendors [4,5] and is very practical means for solving this type of problem. In addition to calculating the loss due to a particular field distribution, the software allows the cause of the loss to be visualized. This permits the designer to modify the geometry to minimize the losses.

The work reported in this paper uses finite element modeling to visualize and quantify four loss mechanisms:

1) Leakage fields due to air gaps

2) Orthoginal fields due to the gap between the winding and the window

3) Interactions of windings with half wave currents

4) Shield eddy currents

II. AIR GAP FIELDS

Air gaps in the magnetic material are frequently used in inductors. This is particularly true when the operating frequency is sufficiently high that the only practical material with sufficiently low losses is a ferrite. The problem is worse when the component is an AC inductor as apposed to a DC filter inductor.

A gapped inductor structure is shown in figure 1 along with a field plot. It would appear from the field distribution that there is little or no fringing field due to the gap and that there is no problem. However, a plot of the current density along the inner edge of the winding (figure 2) shows that the essentially all of the current is concentrated in a thin strip adjacent to the gap. More detail of the current density in the winding near the air gap is shown in figure 3. The apparent lack of a gap leakage field is due to cancelation by the large eddy current in the winding. At DC the winding loss for this example would be 26 W. At 100 kHz the winding loss, in the absence of an air gap, would be 63 W. The power loss for the current distribution shown in figures 2 and 3 is 345 W. 5.5 times the loss predicted by normal proximity effect analysis! The most obvious way to reduce this loss is to move the winding away from the gap. The question is how far? Moving the winding away from the gap increases the window size which in turn increases the core size and core

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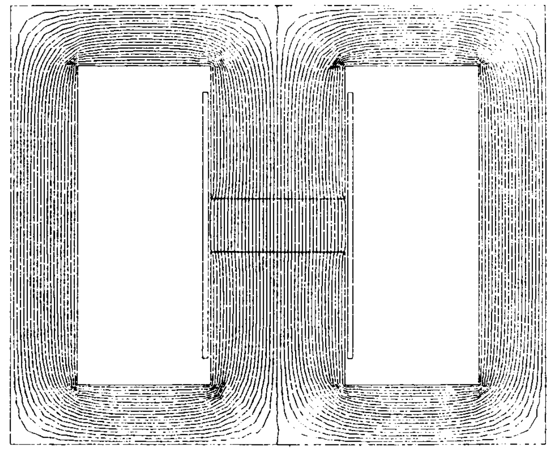

Figure 1. Gapped inductor field plot.

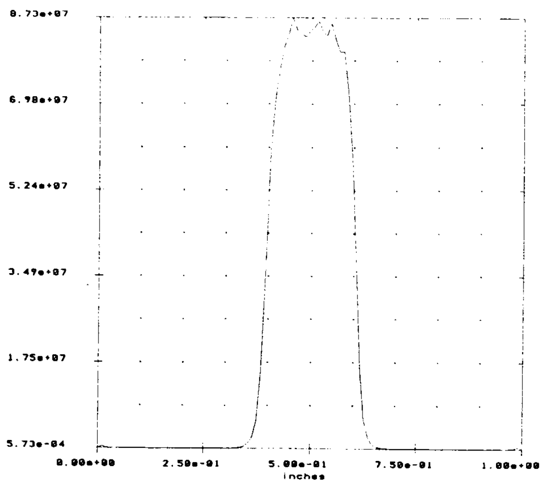

Figure 2. Current density on a vertical line just inside the winding surface of the inductor shown in figure 1.

losses. It is desirable to move the winding far enough to minimize the losses but not so far as to greatly enlarge the window. Figure 4 shows the field distribution for the geometry when the winding is moved away from the gap by a distance equal to twice the gap. The winding loss is now reduced to 78 W, which is only a 24% increase over the ideal. Moving the winding even further away would reduce the loss further but might result in excessive window size.

Another alternative would be to reduce the gap spacing by using multiple gaps such as shown in figure 5. Here the total gap is the same as in figure 1 but each gap is one fourth as wide. The winding is spaced a distance equal to twice the smaller gap. The winding loss in this example is 77 W, very nearly equal to the geometry in figure 4 but requiring a much smaller window.

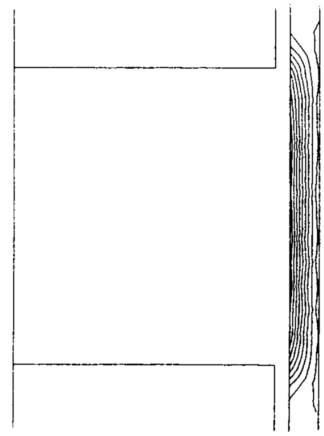

Figure 3. Equi-current density plot for the winding in the region of the gap.

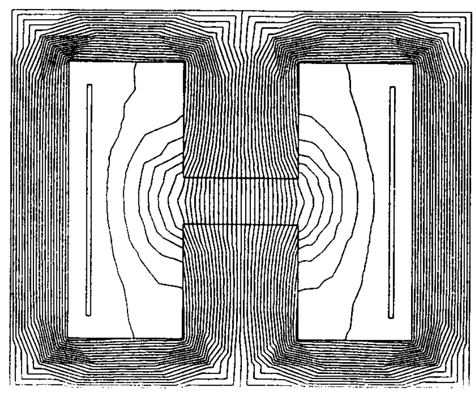

Figure 4. Field plot for the winding moved away from the gap by 2lg.

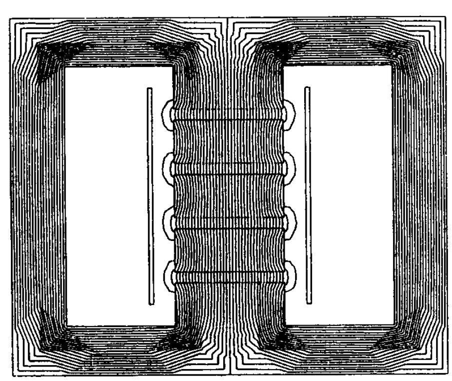

Figure 5. Example of a multigap structure.

Numerous other modeling exercises show that a useful rule of thumb for the spacing between a the innermost winding layer and the air gap is a distance of two to three times the gap length. With this spacing the additional loss due to the presence of the gap will be small.

A practical concern is how to implement a multigap structure in production. In a two or three gap configuration the multiple gaps can be implemented by modifying the winding bobbin to hold one or two ferrite disks as is shown in figure 6. The ferrite disks would be a press fit and/or held in place by a drop of adhesive. The placement of the disks will not be exact, there will be some manufacturing tolerances. It usually turns out that if the gap is large enough to require implementation with multiple gaps, small (in proportion to the gap) misplacements of the disks don't effect the inductance significantly. The use of multiple gaps is really called for only in large gaps.

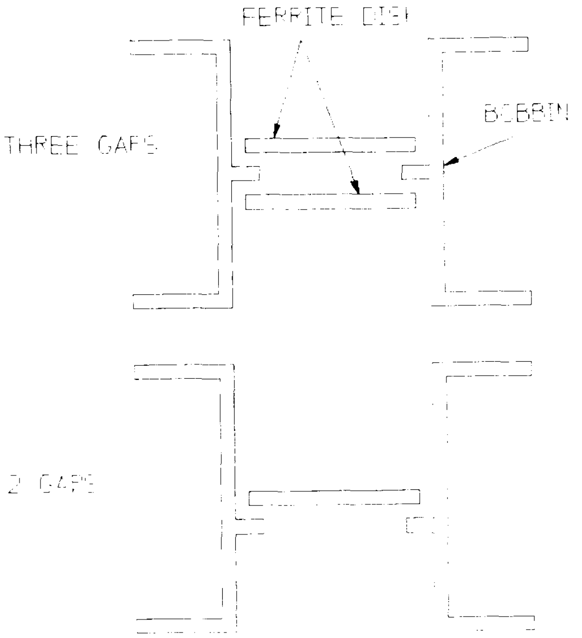

Figure 6. Practical means for implementing multiple gaps.

In some cases, particularly at high power levels, it is desirable to use more than three gaps. While such implementation are of questionable practicality they can be done using alternating ferrite and insulating disks. It should be pointed out that the insulating disk material must have low dielectric losses at the operating frequency. If it does not the insulating disks may melt or even ignite.

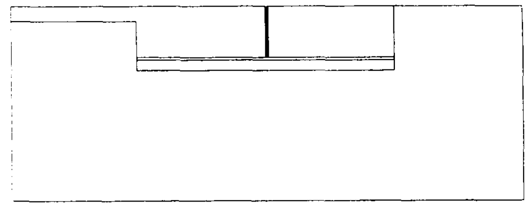

Figure 7. Magnetic structure example.

III. WINDING-WINDOW SPACING

In any practical structure there will be some finite space between the top and bottom of the winding and the window. The question is "does the distance between the winding and the core matter?" Figure 7 gives an example of a two turn winding in a gapped inductor. The winding has deliberately been moved to the middle of the window to avoid interaction with the gap or fields near the corner of the window. The picture represents one quarter of the cross section of a pot core structure. The upper half and the left side of the core is not shown but is included in the model. The two horizontal lines across the bottom of the window are regions which may be modeled either as core or window to determine if the spacing between the end of the winding and core matters. A plot of the field near the end of the winding is given in figure 8. The spacing is .001". It can be seen from the field plot that there is no obvious orthoginal field component. The calculated loss for this configuration is 2.05 W. If the spacing is increased to .026" the field is altered as shown in figure 9. Clearly there are orthoginal components to the field near the end of the winding. The calculated winding loss is now 2.25 W. An increase of 10%. While this is not an overwhelming loss it is significant and possibly could translate into hot spots at the ends of the windings.

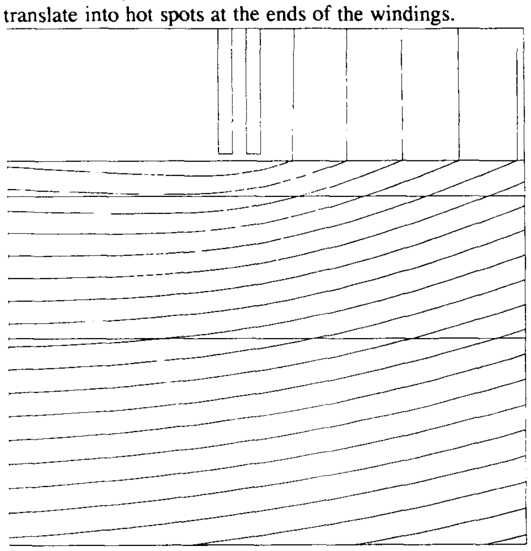

Figure 8. Field near the end of the winding for a .001" spacing.

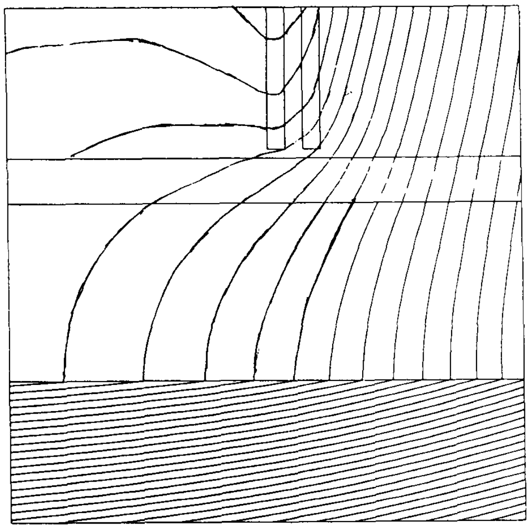

Figure 9. Field near the end of the winding for a .0026" spacing.

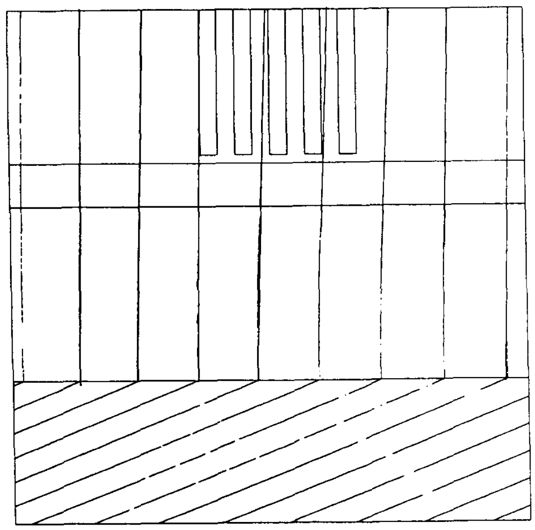

Figure 10. Field near the end of the winding with a 1" layer width and a spacing of .026".

If we increase the winding length by a factor of 5 and plot the field again we see the result shown in figure 10. Here for a spacing of .026" there is little distortion of the field lines and a loss calculation shows no difference between this spacing and .001".

A number of experimental structures have been analyzed. In general the short windings associated with low profile magnetics show the greatest loss due to end effect. In windings with more conventional aspect ratios the end effect losses are usually insignificant.

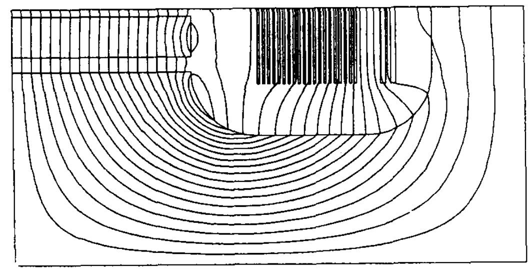

Figure 11. Field plot in a typical transformer-inductor.

An example of the effect of winding-window spacing in a transformer is given in figure 11. The bending of the field lines near the ends of the winding is clearly visible. Note that this transformer is also acting as an inductor and has air gaps in the core.

The quantitative effect of the field distortion will depend very much on the flux density in the windings. In the gapped structures shown in figures 7 and 11 the flux densities in the winding (for a given core flux density) will be higher because of the increase in reluctance of the main magnetic path. This implies that the loss due to field distortion at the ends of the winding will be higher in inductors than in normal transformers where the gaps are very small and the reluctance of the core small. In those structures where the gap is large and the winding window small the end losses can be substantial. This is the kind of structure often considered for low profile, high frequency, AC inductors.

IV. HALF WAVE WINDINGS

The fullwave center tapped rectifier connection shown in figure 12 is typically used for low voltage, high current outputs. In this circuit each of the secondaries (S1, S2) conducts alternately with a halfwave current in each side. As the frequency and power level are raised it is often possible to use a single turn for each of the secondaries. The usual proximity effect calculation would indicate that the optimum layer thickness for each of the windings is in the range of .8 to 1.8 skin depths at the frequency of operation depending on the current waveform.

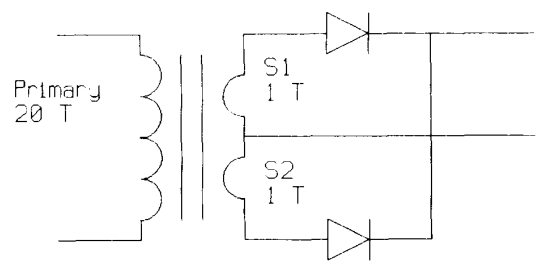

Figure 12. Transformer connections

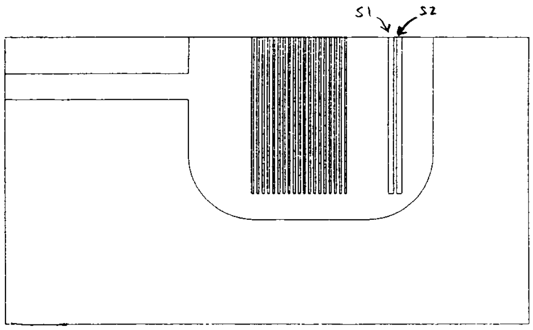

Figure 13. Winding arrangement.

An example of how a such a transformer might be wound is given in figure 13. The two outer turns (S1 and S2) are the low voltage, high current secondaries. S1 and S2 are each 1.2 skin depths thick. From the normal calculation one expect a winding loss of about .8 W but in practice the actual loss appeared to be much higher.

The reason for this can be determined by plotting the current densities in the windings. Figure 14 is a plot of the winding current density while S1 is conducting. The loss in S1 is .848 W and the loss in S2 very small. This agrees well with the predicted value. However, when S2 is conducting the current distribution is very different as shown in figure 14. The loss in S2 is .877 W (because it is the outside winding it is slightly longer), again as expected, but the loss in S1 is .6 W even though it has no average current flowing in it. The reason for this unexpected dissipation is the eddy current induced in S1 by the current in S2. This can be seen clearly in figure 15 where the eddy current in S1 is comparable to the normal conduction current.

The use of two concentric single layer windings is a very natural and practical arraignment for low voltage high current windings for use with a center tapped fullwave rectifier connection. Other winding geometries tend to be less convenient to wind. One way to eliminate this problem is to use either another circuit topology, such as a forward converter or to use an alternate rectifier connection which allows only a single turn to be used. A bridge rectifier could be used but this is normally not acceptable for low voltage outputs due the additional diode drop. Another approach is to use one of the two inductor voltage divider rectifier connections such as those shown in figure 16. These circuits have been in use for at least 70 years [6,7,8,9,10,11,12,13] but are for some reason are not widely known.

If the use of alternate topologies or rectifier connections is possible one can reduce the thickness of the inner turn. A compromise between increasing the conduction loss and reducing the eddy current loss can be made. It this particular example it was found that by increasing the thickness of S2 to .006" and reducing S1 to .004" the additional loss was cut in half. Some further reduction was probably attainable but was not explored.

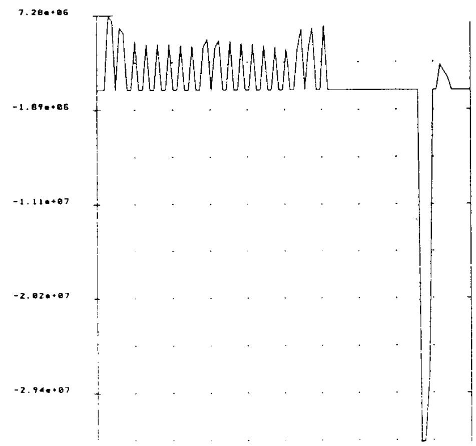

Figure 14. Winding current distribution with S1 conducting.

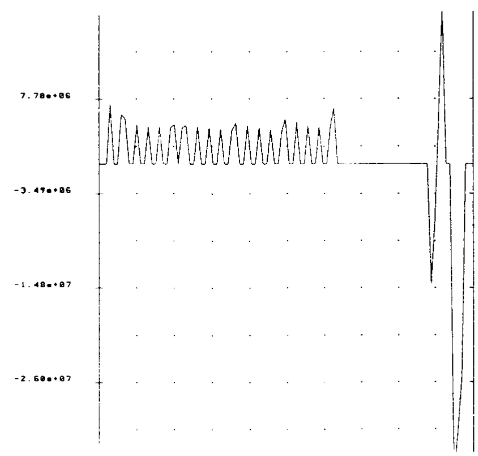

Figure 15. Winding current distribution with S2 conducting.

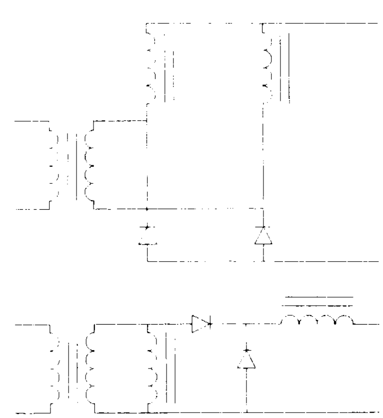

Figure 16. Single winding current multiplier rectifier connections.

V. SHIELD EDDY CURRENTS

If S1 in figure 13 is considered to be an open circuit shield rather than a conducting winding, there will still be an induced eddy current due to S1. The additional loss mechanism is the same. It is desirable to use as thin a foil as possible for the shield. In particular the foil should be much less than one skin depth at the operating frequency. This is particularly important if the winding currents are rich in harmonics as would be typical in switchmode and some resonant topologies. Fortunately the thickness of the shield is limited not by a conduction current but by what is practical to assemble. Thicknesses down to .0005" can be used in foils and if a metalized plastic film is used for the shield even thinner layers are possible. There is of course the problem of making a reliable ohmic contact to the metalized film. Given the thin foils that can be used foil eddy currents do not need to be a problem except at very high frequencies (MHz). The designer must however, be careful to use a foil thickness appropriate for the operating frequency.

VI. CONCLUSION

A finite element modeling effort has demonstrated that there are additional losses in high frequency windings beyond those included in the conventional analysis based on Dowells work. These losses may or may not be significant depending on the particular circumstances. In general it is necessary to model the geometry of interest to determine the additional losses.

Two useful rules of thumb appear: space windings at least a distance of twice the air gap length away from the gap and make shield foils thin compared to the skin depth.

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