Double-Sided Conduction: A Loss-Reduction Technique for High Frequency Transformers

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***Abstract*—Power converters are increasingly being operated in the high-frequency (HF) regime (3-30 MHz), where proximity-effect losses in transformers are difficult to contain. Litz wire does not scale well to multi-MHz operation where sub-skin-depth strands are prohibitive to manufacture. Interleaving is more scalable to HF and, in its best implementations, causes a current equal to the net current to flow on a single surface of a conductor or layer of conductors (e.g., a copper turn in a planar transformer), with no "negative" current or eddy currents. In this paper, we propose a field-shaping technique that results in the even distribution of current on *two* sides of each conductor layer, yielding two skin-depths worth of conduction. We present one method to achieve this by paralleling the first and last layer of a foil-wound transformer. We demonstrate this technique theoretically, through FEA simulation, and through an experimental prototype.**

***Index Terms*—High Frequency Magnetics, Eddy currents, flexible PCB, double sided conduction**

I. Introduction

Power converters are increasingly operated in the high frequency (HF, 3-30 MHz) regime to reduce converter energy storage requirements and passive component size. These gains are at risk if high-frequency losses place thermal constraints on component size, especially for magnetic components. Specifically, eddy current effects can drastically increase HF transformer conduction losses as operating frequency increases. Current tends to crowd into a very thin skin near regions of high magnetic (H-) fields. Furthermore, transformers often consist of multiple layers, which presents the opportunity for H-fields between each conductor to build up from layer to layer. Unfavorable distributions of H fields can result in orders-of-magnitude higher loss than the same net current flowing at dc [1]. This hinders the performance of transformers in high frequency power applications.

One current solution is the use of litz wire. However, litz wire does not scale well to the HF regime. The thinnest commonly-available litz strands (48 AWG) have a diameter equal to the skin depth in copper at only 4 MHz, at which point litz already can be ineffective or even counterproductive [2]. Even below this frequency, litz wire with many thin strands is often cost-prohibitive due to intensive and delicate manufacturing.

Another way to mitigate HF losses is by interleaving the transformer windings [1], [3]–[5]. What we might call "standard" full interleaving (Fig. 1b) results in current limited to one skin depth on one surface of each conductor, with none of the detrimental opposite-polarity current or discernible eddy currents.

In this paper, we propose "standard" interleaving is not the limit to performance. In [6], it is demonstrated that, if the H field can be of equal magnitude and opposite direction on either side of a conductor, current will flow on *both* surfaces yielding two skin depths worth of conduction, or *double-sided conduction.* In principle, this conduction pattern offers a factor of two improvement in ac copper loss over standard full interleaving in deeply skin-depth-limited applications.

Reference [6] demonstrated this method for inductors. Here we propose a technique to achieve double-sided conduction in transformers. The proposed approach applies .equally well whether the number of turns per layer is one (e.g., foil) or many [1] (e.g., helically wound round conductors). We validate this approach with FEA analysis and a hardware prototype which is assessed using both resonant and calorimetric means.

II. Conceptual Approach to Double-sided Conduction

A. Proximity Effect and Standard Interleaving

Eddy current mechanisms in transformers manifest as the skin effect and the proximity effect. Both effects are the result of magnetic fields generating eddy currents which oppose them according to Lenz's law. The term *skin effect* is used to refer to the phenomenon in which ac current tends to distribute itself near the surface of the conductor, resulting in the increase of the ac resistance with frequency. The term *proximity effect* is used to refer to eddy currents which can be attributed to magnetic fields generated by high frequency current carried by other conductors. Both effects ultimately derive from the same rule – current will crowd to the surfaces of conductors, and the magnitude of that current is determined by the magnitude and direction of the H field immediately outside the conductor. Current can be modeled as surface currents as long as the skin depth δ is much smaller than the conductor thickness, which we assume for the remainder of the paper.

These effects are illustrated in Fig. 1a, which depicts a conventional 2-winding foil transformer, with primary layers wrapped around the center post and the secondary layers wrapped around the primary. The center post of the high-permeability core would appear above the chart, the outer shell

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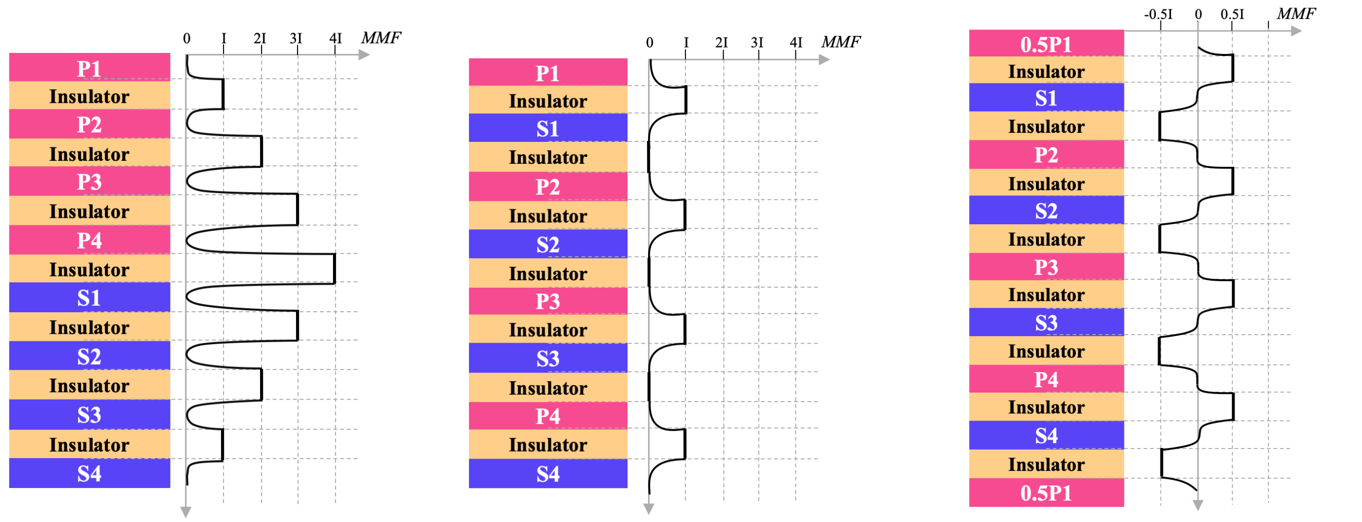

(a) Non-Interleaved Winding

(b) Traditional Interleaved Winding

(c) Proposed interleaved 'Double Sided Conduction' winding

Fig. 1: MMF Diagrams of Various Winding Strategies for 1:1 Planar or Foil Transformers, showing improvement in MMF distribution for the proposed strategy.

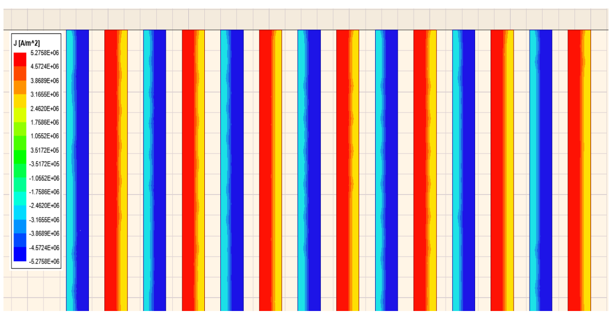

Fig. 2: Current distribution in a standard interleaved winding arrangement, turns ratio 7:7.

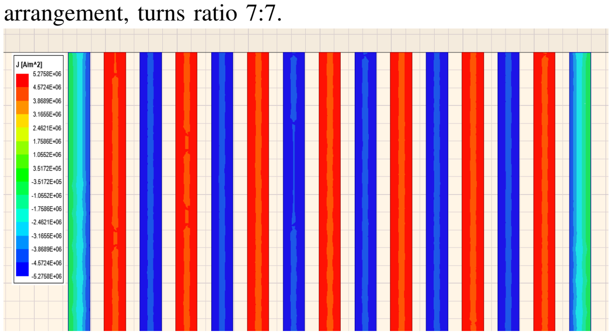

Fig. 3: Current distribution in *double-sided conduction* winding arrangement, turns ratio 7:7.

or "legs" would appear below the chart, and the (usually flat) top and bottom plates of the core would appear on the left and right sides of the chart. This arrangement creates a buildup of H-field between each conductor due to the consecutive stacking of same winding layers [P-P-P-P-S-S-S-S]. Current crowds towards regions with high H-fields and, because of the buildup of H fields, the surface currents on individual sides of a conductor can be many times the net current in the conductor. Current on one surface of each conductor flows in the direction of the net current while current on the other surface flows opposite the net current. Thus surface currents much larger than the net current can flow in each layer and cause large loss without contributing to the function of the transformer.

By alternating the primary and secondary layers of the transformer windings (Fig. 1b), the H field between layers does not build up with each consecutive layer. The current in each conductor still crowds to one side, but its magnitude is equal to the net current in the layer (there is no opposite-polarity current on the other surface), which is a significant improvement. Nevertheless, current only flows on one side of each conductor; for wide and thin conductors, this leaves room for improved conduction area by roughly a factor of two.

B. Balancing H-Fields and Double-Sided Conduction

We propose a field-shaping technique that manipulates the H field around conductors to achieve *double-sided conduction*, i.e. a conduction pattern in which the net current divides evenly into one skin depth on each side of a conductor or layer of conductors, thus increasing conduction area and decreasing copper loss by a factor of two. The H field pattern that yields the desired conduction is shown in Fig. 1c.

One way to achieve this is to split the first primary layer (for example) into two layers, electrically in parallel, with one parallel layer wrapped around the center post (as usual) and the second parallel layer positioned as the last layer in the stack. This simple modification to standard interleaving creates the structure in Fig. 1c. The MMF diagram shows equal-and-opposite MMFs on either side of every inner conductor, yielding two skin depths of conduction, aligned with the net current, with no negative/eddy currents. The first and last layer each have one skin depth worth of conduction and, taken together as they are in parallel, can be thought of as a single layer with double-sided conduction. This is volumetrically consistent, as inner layers would be designed to have a certain thickness, while the first and last (parallel) layers need only have half of that thickness in principle.

The authors are aware of only one instance in which such a structure has been proposed [3]. In [3], a structure like the one proposed here is introduced as the result of an extensive optimization search, with little clarity as to the source of the improved performance or its generality. The analysis presented here provides design intuition, broadly applicable design guidelines, and confidence in the generality of the proposed approach.

For comparison, Fig. 2 and Fig. 3 show current distributions for typical interleaved windings and windings using the double-sided conduction approach proposed here. These distributions were both extracted from ANSYS Maxwell with an excitation frequency of 3 MHz and both figures use the same color scale. With standard interleaving, current concentrates to one surface of each conductor as predicted theoretically. With the proposed approach, current evenly distributes itself on two sides of every inner conductor; the outer conductors each carry half of the primary current on one surface apiece and, taken together, constitute a single conductor with double-sided conduction.

We further note that it is possible to achieve double-sided conduction in a N:1 transformer, if the one-turn winding consists of $N + 1$ paralleled layers (including the first and last layers) interleaved with N series connected layers on the primary. In this arrangement, H fields will also distribute to guarantee double-sided conduction. However, if compared to an implementation with traditional interleaving (i.e. N paralleled layers and N series-connected layers interleaved together), the improvement is marginal, because H fields naturally balance to maximize conduction area in this case. Nevertheless, if one is using a PCB winding (as we do in this work), adding another paralleled turn does not increase complexity very much but does reduce loss.

C. Application Constraints

While double-sided conduction is broadly desirable, the proposed approach to achieving it is most suitable for transformer applications that do not use the magnetizing inductance for energy storage, i.e. those with high magnetizing inductance L_m (e.g., for most isolated PWM converters, series-resonant converters, dual-active-bridge converters, etc.). This condition enforces the primary current is in phase with the secondary current, scaled only by the turns ratio – a very common and sometimes unstated assumption in discussions of interleaving [1]. The proposed approach can achieve low leakage inductance L_{lk} through interleaving with closely-spaced layers. Nevertheless, designs which require energy to be deliberately stored in the leakage inductance (e.g., in series-resonant or DAB converters) can achieve higher L_{lk} by adding more inter-layer spacing, with the added benefit of reduced capacitance. For applications that deliberately store energy in the magnetizing inductance, other approaches that yield double-sided conduction are available [7].

III. Design of a HF transformer to achieve double-sided conduction

For this paper, a 1:1 transformer is designed to exhibit double-sided conduction. We achieve this with 7 turns in series on the primary (P) winding and 8 turns on the secondary (S) winding, with the 1st and 8th turn electrically paralleled and in series with the remaining 6 turns to give an interleaving structure of 0.5S-P-S-P-S-P-S-P-S-P-S-P-S-P-0.5S as discussed previously.

The double-sided conduction technique yields 2 skin depths worth of conduction per turn at high frequencies, hence, in designing the windings, a conductor thickness of ≈ 2 skin depths is desirable. In fact, the benefits of a double-sided conduction design become most apparent when the conductor is more than 2 skin depths thick, as will be shown in section IV. We note that thicker copper can be detrimental in non-interleaved designs by providing more space for eddy currents to flow – this penalty is not present in the proposed design or a standard interleaved design, and thermal considerations or practical limits could still recommend thicker layers.

The proposed structure can be wound using foil layers with protruding tabs for connecting to each other and to the circuit. Traditional foil windings can perform this task. We introduce a further manufacturing simplification by using a flexible PCB as the windings of the transformer. The flexible PCB has its dimensions and tabs designed to create the desired layers with the tabs aligned after rolling. This technique is interesting in its own right and, to our knowledge, also novel. The flexible PCB has great potential[1] as an alternative for winding materials as it is easily manufactured in large batches and readily assembled. Several manufacturers are able to print copper thicknesses up to 3oz ($105\,\mu\mathrm{m} = 2 \times \delta(100\,\mathrm{kHz})$) with up to 6 conductive layers on a flexible substrate. For the 1:1 transformer, only 2 layers are required to form the primary and secondary windings of the transformer. The design is shown in Fig. 4, where the primary is a solid plane across most of the PCB and the secondary is split into several separated planes to create the paralleled inner and outer secondary layers as well as the six inner secondary layers. The innermost and outermost layers are short copper pours with length equal to their expected circumference, while the inner turns are formed from a single copper plane. While the primary simply has two tabs (each representing a winding terminal), each separate pour of the secondary (i.e. the first layer, the last layer, and the inner layers) has two tabs so that inter-layer connections can be made. Construction is simple, with the PCB wound around a center bobbin and the tabs soldered together. The tabs have large copper pads and vias for convenient soldering – these can be simplified and miniaturized in mass production.

[1]One possible limitation in some applications is the minimum bend radius of the flexible PCB, which must be smaller than the center post radius of the core. This was not an impediment to the prototype example presented here.

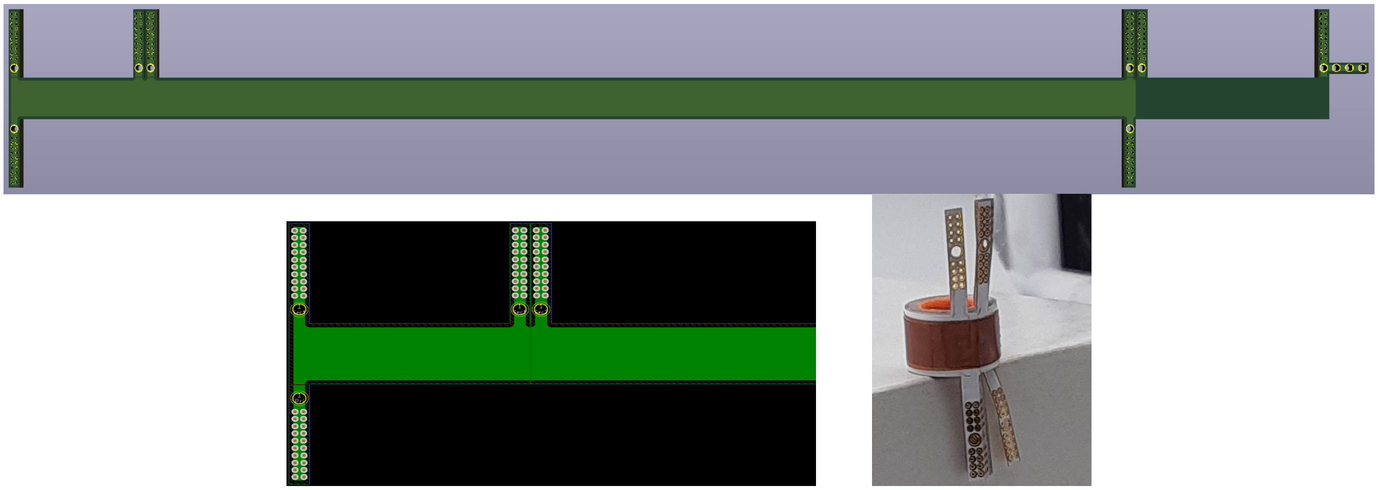

Fig. 4: *Top*, KiCAD 3D model of flex PCB showing the primary side with 7 turns in series. *Bottom left*, KiCAD layout view of the 1st parallel turn and 1st series turn of the secondary winding (green layer). *Bottom right*, wounded flex PCB on a core bobbin.

It is critical to ensure that the tabs connecting paralleled layers align once the structure is rolled. To aid in the design process, we laser cut thin plastic shimstock into the shape of the eventual PCB. Winding these sheets around the selected core (an RM core, to ensure enough room for the tabs) allowed us to properly align tabs prior to ordering the PCB.

We note that the closest analogue to this flexible PCB-based winding is a simple foil winding, which is possible but more difficult. The versatility of a PCB winding is apparent in both its ease of use and enabling potential for non-traditional winding structures. The archetypal double-sided conduction transformer we present has a 1:1 turns ratio, but one could imagine achieving any arbitrary N:1 turns ratio transformer using this method by paralleling all secondary turns (again, greatly simplified using the flexible PCB winding).

IV. FEA Simulation

An ANSYS Maxwell model of the transformer was created along with a circuit representation to specify the connection of the windings. The Eddy Current solution type is used to model the magnetic fields and current distribution at a single frequency.

A. Transformer Design Considerations

1) Geometry: The core shape used to build the actual transformer structure is an RM-7 core. To simplify analysis and simulation, the transformer model is analyzed as a 2D structure, cylindrically-symmetric about the Z-axis. Fig. 5 shows the geometry of the transformer, simulated for the double-sided conduction technique, with key dimensions highlighted.

2) Inter-layer spacing of windings: The two layer flexible PCB used has a total thickness of 0.265mm: the top and bottom insulators each having a thickness of $50\,\mu\text{m}$, the two copper layers each having a thickness of $70\,\mu\text{m}$, and the inner flexible insulator between the two copper layers having a thickness of $25\,\mu\text{m}$. Hence, the transformer prototype built has unevenly spaced turns with inter-layer distances of (50-25-100-25-100-...)μm, coming from the combined thicknesses of the insulating layer, when wound tightly about the core bobbin. FEA simulation, with the same frequency and drive level, was used (maintaining the same core geometry) to compare the cases for the even (Fig. 3) and uneven (Fig. 6) spacing sequences. The simulation results showed the case of the latter still achieved double-sided conduction, with losses within 5% of the ideally spaced case. This indicates that uneven spacing will not be a significant detriment to performance.

3) Bulk conductivity of copper windings: Ideally, the winding should be made out of high-conductivity copper, but this is not always guaranteed. Hence, to ensure the FEA simulation precisely matches those of the experimental, the bulk conductivity of the physical materials needs to be accounted for in the simulation. The bulk conductivity was calculated using the dc resistance of the flex PCB. This dc resistance is measured using the thermal calorimetric technique discussed in section V-A. The bulk conductivities of the physical materials were $\approx$ twice less than that of pure copper.

B. FEA Results

Fig. 9 shows the expected % loss reduction of the proposed structure (DsC) compared to a traditionally-interleaved counterpart with identical core and winding dimensions (SsC) across a range of frequencies. As frequency increases, the proposed method becomes more advantageous: At 1 MHz, the double-sided conduction method has roughly 10% less loss than the interleaved structure; at 3 MHz, that becomes a 24% loss reduction; at 8 MHz, 42%, approaching 50% as the conduction pattern more resembles two distinct "skins."

The copper thickness in the simulations is $70\,\mu\text{m}$, equal to two skin depths in copper at approximately 3 MHz. At frequencies lower than 3 MHz, then, we do not expect the full advantage of double-sided conduction to be evident, as is demonstrated by the simulation results. However, even at frequencies where the skin depth is less than half of a

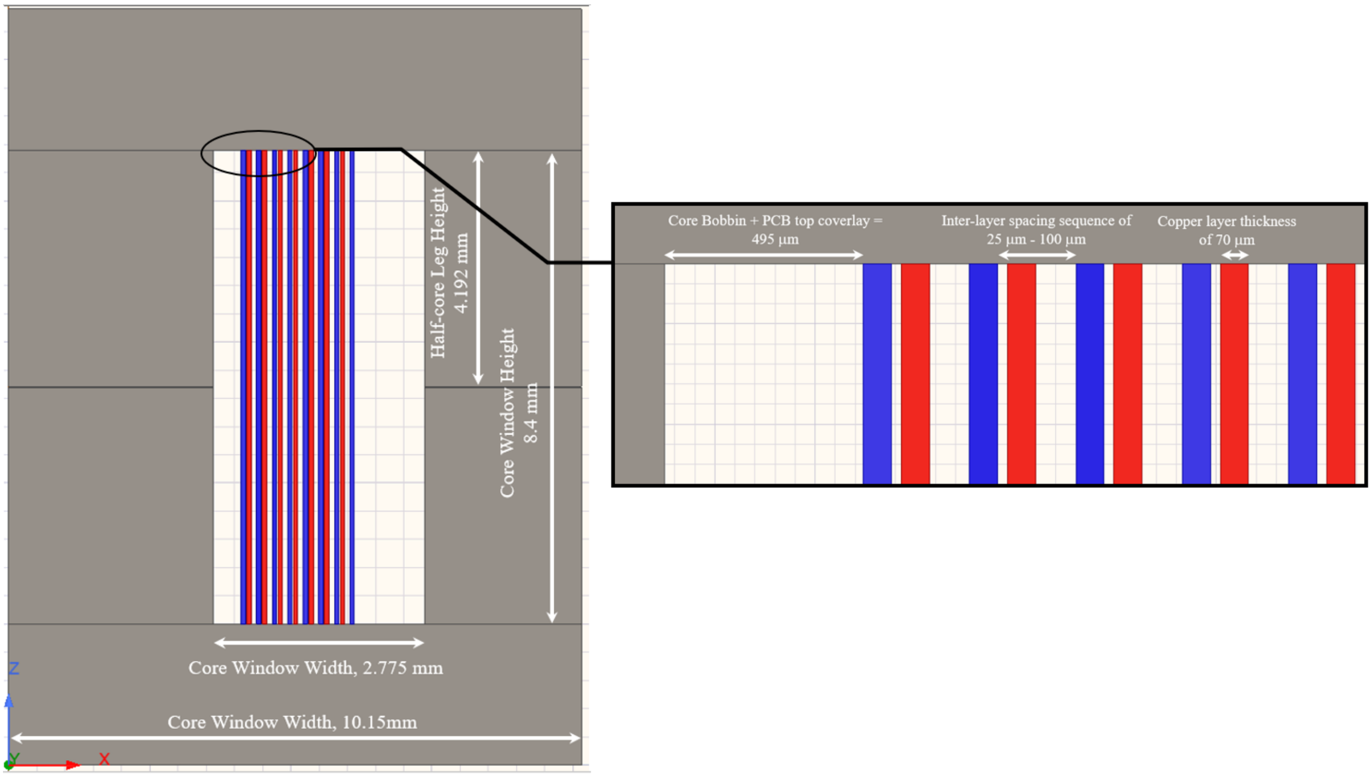

Fig. 5: Geometry of simulated transformer model with core window.

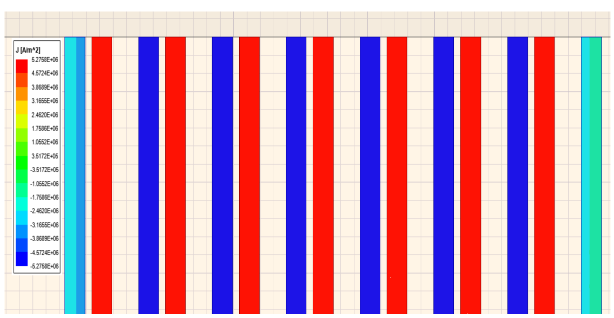

Fig. 6: Current distribution in *double-sided conduction* with uneven inter-layer spacing.

conductor thickness, current distribution may not form two or even one distinct 'skin'. As a result, there will be non-zero J distribution in the center portion of the conductor, leading SsC and DsC designs to have more similar loss than one might intuitively expect. This phenomenon is shown explicitly in Figs. 7 and 8. In Fig. 7, we observe that 2 distinct skins become more prominent when the skin depth is well below half the thickness of the layer. As such, the full advantage of the proposed method will not become fully apparent until operated at a high enough frequency to cause current to distribute into one or two true 'skins.'

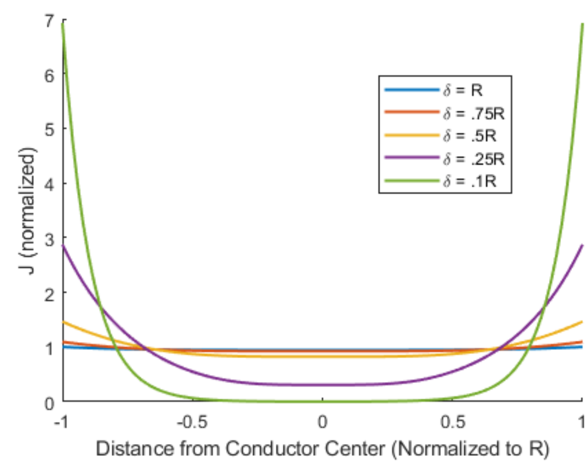

Fig. 7: Plot of normalized current density across conductor thickness ($2R$) with varying skin depths, showing that current does not form into distinct 'skins' until the skin depth is significantly smaller than conductor thickness.

V. Experimental Methodology for Loss Measurement in HF Transformers

It is difficult to directly assess the efficiency of high-frequency transformers. Direct I-V measurements require very high phase accuracy, which becomes difficult for high efficiencies and high frequencies. Placing the transformer in a power converter and measuring efficiency is also fraught, as one must accurately discern between multiple contributors to total loss (e.g. from switches, diodes, etc.). Here we use two methodologies to verify the loss characteristics of the structures being tested: calorimetry and a resonant electrical test, as shown in Fig. 13. These tests are individually defensible, and together represent strong evidence for the accuracy of the experimental

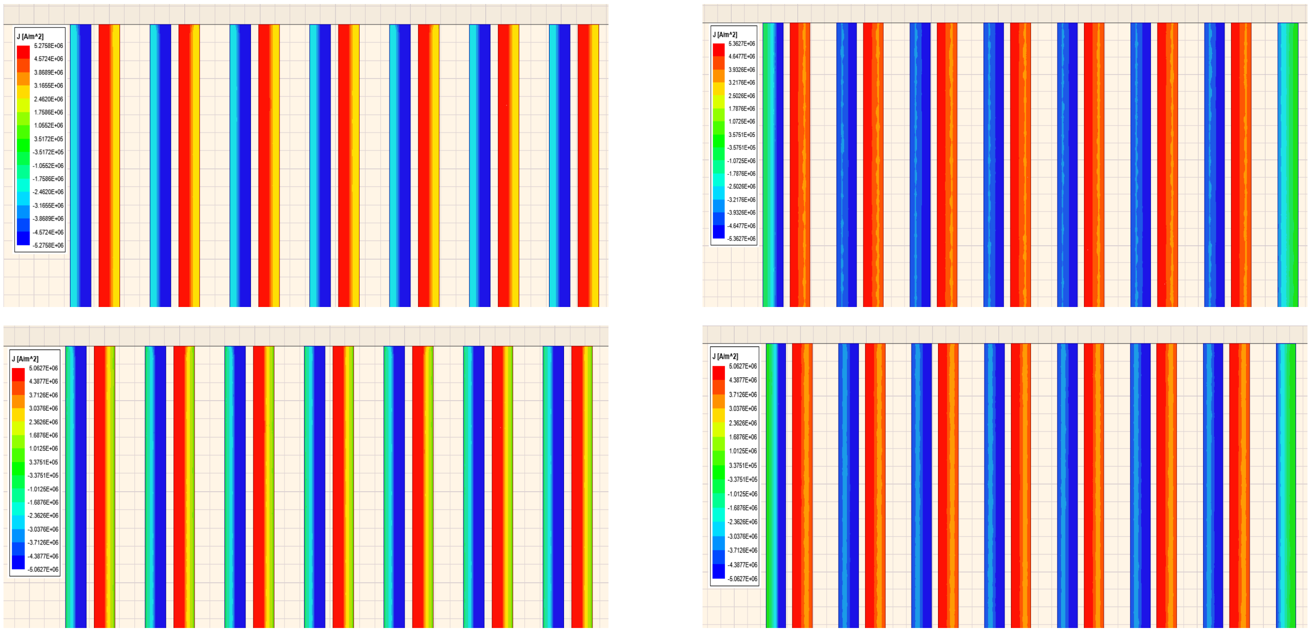

Fig. 8: Simulation results: Top, for SsC ($left$) and DsC ($right$) at 8 MHz; and $Bottom$, for SsC ($left$) and DsC ($right$) at 10 MHz.

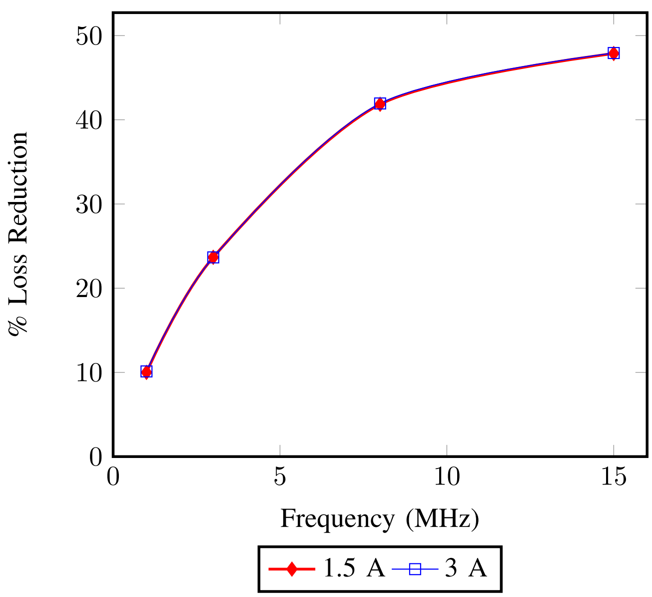

Fig. 9: Comparison of transformer power losses simulated across frequencies for two different current levels

results.

A. Calorimetry

The calorimetric method is an established method for accurate power loss measurements in HF transformers [8]–[10]. Calorimetry avoids the challenges with electrical measurements by more directly measuring the heat (power loss) in the transformer. In this paper, we use a calorimetric method similar to that demonstrated in [11], which implements a thermal camera to observe the temperature rise on the surface of the transformer. The calorimetric setup is calibrated by exciting the primary windings with dc current and accurately measuring the dc voltage and current. High precision can be achieved by using steady-state temperatures, with accelerated results available by comparing transient curves over shorter time periods. High-frequency tests are performed with the same transformer in the same physical orientation and the transformer's temperature is compared with the transformer's calibration curves (Fig. 10) to establish actual high-frequency loss.

B. Series Resonant Circuit

We further verify our loss predictions using a series resonant approach, frequently used in the characterization of magnetic materials and inductors [6], [12]. We implement an extended version of this approach (as described in [13]) which permits it to characterize transformers. In this approach, the device under test is placed in series with a capacitor and the LC circuit is driven at its resonant frequency. At resonance, the ratio of peak output (capacitor) voltage to input voltage is directly proportional to the quality factor, Q_L, of the inductive element under test. This enables indirect calculation of losses, since the excitation current is known from the voltage across the resonant capacitor. The ESR of the resonant capacitors and the parasitic capacitance of the board traces, although minimal sources of error, are also taken into account to ensure accurate measurements.

Since the transformer is designed to be excited with the primary and secondary currents in phase, with little magnetizing current, this test is conducted by exciting the primary with the secondary terminals shorted. The resonant capacitor is chosen to resonate with the combination of the primary and primary-referred secondary leakage inductances at the desired operating frequency. The transformer thus functions as it would in operation ($\omega L_m \gg (N_1/N_2)^2 R_{load}$) and its losses can be measured directly to assess the efficacy of the double-sided conduction technique.

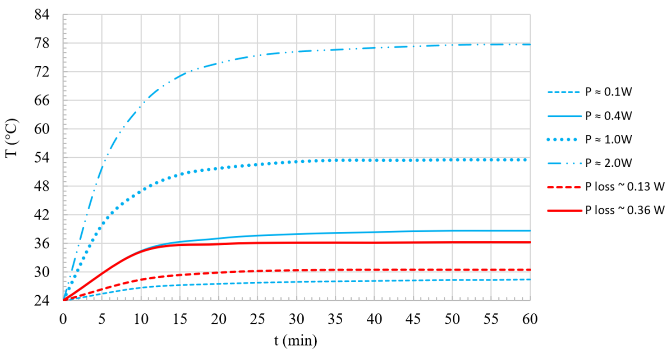

Fig. 10: Thermal measurement (red) of transformer at 2 current levels, obtained from Q-measurements and compared with the thermal calibration characteristics of the transformer (blue).

Figure 11 shows two sample measurements taken for the 1:1 double-sided conduction transformer at different drive levels. The results from the electrical (series resonant) measurements show good correlation with those of the thermal measurement, reinforcing confidence in the measurements.

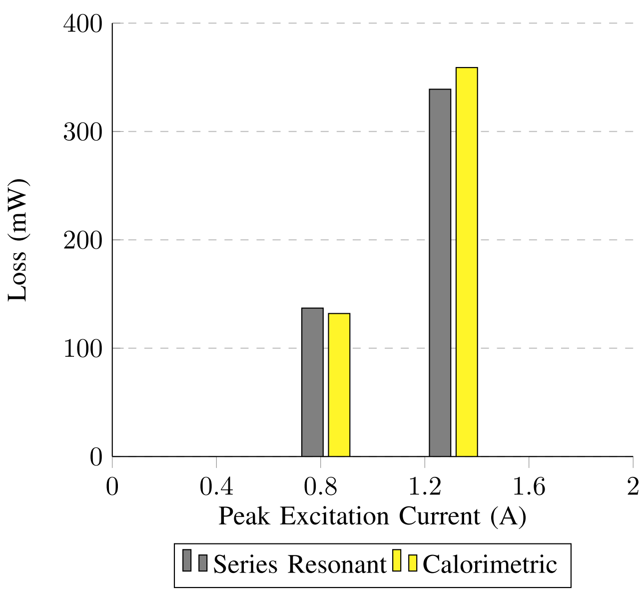

Fig. 11: Comparison of transformer power losses measured using the series resonant measurement approach and thermal measurements.

VI. Experimental Validation

To demonstrate the improved performance of transformers built with the proposed technique, a DsC transformer was built along with a second transformer with full standard interleaving, the state-of-the-art alternative to the proposed approach. This second transformer was built using the same core and using the same flexible PCB winding strategy, but does not experience double-sided conduction when operating, as illustrated in Fig. 1b and confirmed with FEA in Fig. 2.

Figure 12 shows the loss characteristics of the double sided conduction transformer prototype (DsC) and its traditionally-interleaved counterpart (SsC), measured using the series resonant method at 3 MHz. At an excitation current of 2 A, the proposed method results in a 13% reduction in loss compared to the interleaved structure; at 4 A we observe a 14.1% reduction in loss.

While experimental validation does show increased performance for the proposed design method, it does not show as large of a reduction in loss as FEA simulation predicts. As shown in Fig. 9, at 3 MHz ANSYS Maxwell predicts a roughly 25% reduction in losses when the double-sided method is used. It is unlikely that there is significant measurement error, as thermal and electrical based measurements indicate similar loss values. We attribute the discrepancy to several sources. There may be unexpected core losses, degrading the loss ratio. We used the method proposed in [14] to make an isolated core loss measurement. This test indicated that our core loss is on the order of nanowatts, and thus insignificant compared to the copper loss measured in the structure, as was intended in the design process. This indicates that core loss was not a significant contributor to the unexpected loss observed. One likely cause of the discrepancy between experimental and simulated loss values is the loss associated with the layer interconnects and board terminations. It is well documented that terminations can be a significant source of loss in planar transformer structures [15]. The proposed transformer not only has external terminations to the board where H fields are not well-controlled, but requires external interconnects (the tabs shown in Fig. 4) to make the paralleled inner- and outer-most turns that enable double-sided conduction. The effects of current crowding in these tabs is not modeled in the FEA simulations.

Nevertheless, the experimental improvement of DsC over SsC is nontrivial and not totally out of line with simulation results. With thicker copper or increased operating frequency (not attempted due to the unavailability of RM cores made of low-loss high-frequency structures), the benefit of double-sided conduction is expected to become more apparent.

VII. Conclusion

This paper presents a field-shaping approach to achieve double-sided-conduction in two-port transformers, specifically those that do not store energy in their cores. Double-sided conduction promises up to 50% improvement in copper losses in transformers wound with foil and solid wires, a scenario likely to be important as power converter operating frequencies exceed the useful range of litz wire. This technique has been validated using FEA simulation and an experimental prototype.

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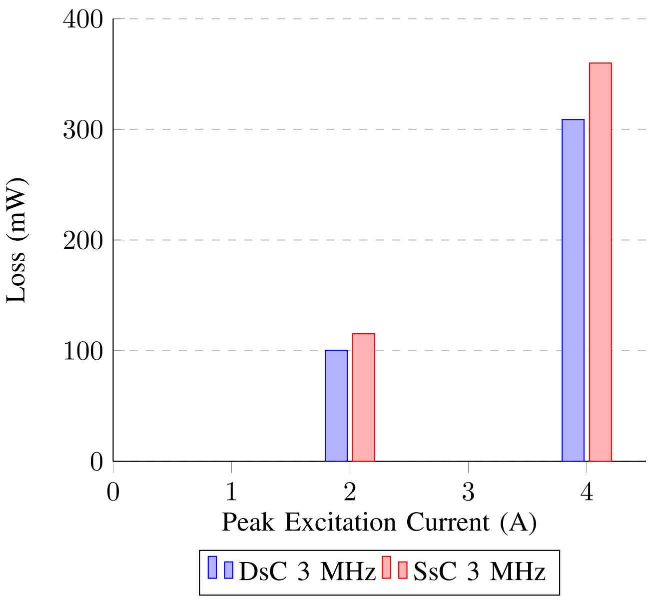

Fig. 12: Comparison of transformer power losses tested at 3MHz at two current levels.

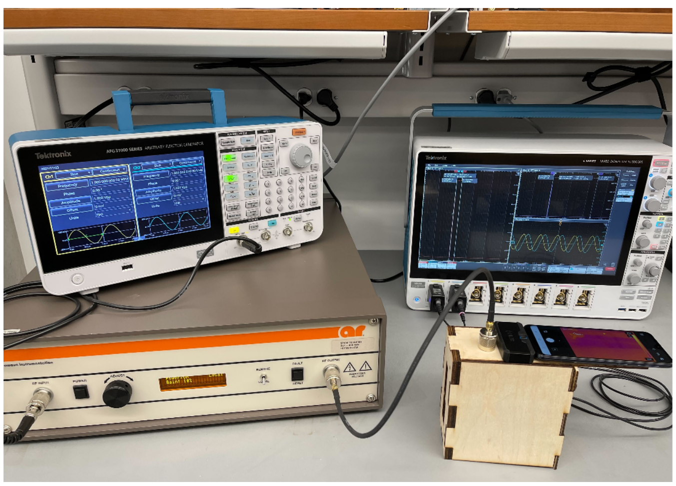

Fig. 13: Experimental test bench for series resonant and thermal measurements.

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