Emulating Full Load Testing of Air-cooled Nanocrystalline IHT at Zero Power

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Abstract— High-power high-frequency air-cooled induction heating transformers, mostly used as current multiplier and isolation purposes, are custom designed. For their reliable performance and long life, the thermal evaluation at rated load is necessary. Creating an equivalent load as test facility for reliability testing of such type of transformer is difficult. Characteristics of such loads drastically change after Curie temperature. Moreover, prolonged heating could increase the nearby ambient temperature and, more importantly, the traditional heat run test wastes large amount of energy. Whenever the coil is energized, windings of transformer draw respective rated current; even at no load condition the copper loss is at rated value. While both windings drawing rated current at desired frequency, using the concept of localized eddy current loss as well as excess eddy current loss, this article proposes a novel method to inject requisite core loss to the magnetic circuit to emulate the characteristics of full load condition but the power drawn from the transformer would be zero. The proposed idea would be validated where only 200W resonant inverter would be used to inject power loss equivalent to full-load condition of 30 kW transformer to emulate the heat run test.

Keywords— *Excess eddy current loss, induction heating transformer (IHT), magnetic loss, series resonant inverter (SRI), zero-voltage zero-current switching (ZVZCS).*

I. INTRODUCTION

HIGH frequency high power transformers [1]-[10] for use in various power electronics converters are not commercially available. They are used not only for voltage/current multiplication or isolation purposes, their characteristics, often, make significant influence in converter design [11], [12]. They are mostly custom designed and often manufactured inhouse. It becomes more important to verify the designed parameters (that includes reliability performance) for such transformers.

Optimal design of transformer involves maximal use of core and copper provided both are within their respective thermal limits [8], [9]; the power density should be large. Power loss in both core and copper should be minimized and heat conduction features of core and winding need to be improved. With the introduction of ribbon-structured low loss thermally conductive nanocrystalline soft magnetic material for core [5] and availability of thinner gauge litz-wire conductors the power loss in core and copper has been reduced significantly. High operating flux density of nanocrystalline cores help reduce the

volume of copper needed in windings, and, thereby the copper loss is also reduced. However, still, the heat removal features from cores and copper windings are not uniform [8]. There are multiple heat sources exist in the transformer. The prospect of generation of hot spots is prevalent.

High frequency power transformers are needed in high power induction heating (IH) applications. Traditionally, they are conduction-cooled [13]. Due to easy availability of ribbon type nanocrystalline toroidal cores with large window area as well as wide range litz wire conductors, several application areas in induction heating (e.g., where the density of transferred power (W/cm^2) on job surface is moderate) are now being covered by air-cooled controllers [14] where air-cooled induction heating transformer (IHT) is an integral part. High permeability and low leakage flux, in particular, make nanocrystalline toroidal cores ideally suitable for ZVZCS SRI based IHT [3], [7]. Due to windings placed all around the core, the heat removal features of each segment of core and copper could be different. Proper thermal evaluation is necessary. Organizing proper full load testing of induction heating system is difficult because it would increase the temperature of the ambience. Moreover, the load would mostly disappear once the job attains the Curie temperature and such type of heat run test wastes large amount of energy. It is known that, if loss characteristics as well as thermal conduction characteristics are modeled properly then the application of finite element method (FEM) [8] could be used to evaluate the temperature profile in core and copper accurately. Here, the traditional practical approach of thermal evaluation of transformer by loading both the core and copper to respective designed or rated capacity would be followed, but the power drawn from the system would be negligible. This article proposes a novel practical method of evaluating the thermal behavior of IHT without even applying any load to it. The structure of the article is as follows: Section II briefly discusses the IHT and approach for its optimal design. Section III discusses the loss characteristics of a high frequency transformer. Section IV details the utility of ZVZCS SRI for evaluation of functioning of IHT. It also discusses the practical and novel idea of equivalently loading the transformer even when the power drawn from coil head is negligible. Finally, Section IV proposes a novel idea of practically evaluating an IHT without even applying any load to it. The proposed method is more suitable for cores with good thermal conductivity.

II. AIR-COOLED INDUCTION HEATING TRANSFORMER

For necessary current multiplication, transformer is invariably used to feed requisite current to series resonant load. A typical SRI driven induction heating controller is shown in Fig. 1. It consists of two control loops. Phased locked loop is used to track the resonant frequency of tank circuit and chopper is used for power control at output. The design of medium to high frequency air-cooled IHT involves on deciding the number of primary and secondary turns, winding patterns, selection of litz wires of proper size, core materials, core geometry, thermal management, etc. For its optimal, the power loss in core and copper should be minimized and layout of windings should be able to remove the heat loss efficiently so as to keep the maximum temperature rise within the respective limits. For superior thermal conduction characteristics, the number of turns at primary (n_p) and at secondary (n_s) should be at respective minimum permissible value. For reduction of copper loss in windings, single layered windings are preferred, the impact of proximity effects on winding resistance becomes negligible. The number of primary turns n_p is decided by input conditions of transformer, as,

$$n_p = \frac{V_{pri}(FL)}{4B_m A_{core} f_s} \quad (1)$$

$V_{pri}(FL)$ is the primary input voltage at full load, B_m is peak flux density in core of effective area A_{core} and f_s is switching frequency of the inverter.

Fig. 1: Series resonant power controller for induction heating

A. Role of Topology on Transformer Optimization

When a square wave input voltage V_{pri} is applied at primary of a transformer feeding resonant tank circuit, both currents at primary i_p and secondary i_L are sinusoidal. Turns ratio n needed to feed rated power the load is expressed as,

$$n = \frac{n_p}{n_s} = 1.1 \frac{V_{pri}(FL)}{Z_{eq} i_L} \quad (2)$$

Z_{eq} is the tank circuit impedance, whose expression is,

$$Z_{eq} = (R_{eq} + esr + r_{ac}) + j \left(2\pi f_s L_1 - \frac{1}{2\pi f_s C_r} \right) \quad (3)$$

R_{eq} is effective load resistance, r_{ac} is ac coil resistance, esr is resistance of tank capacitor C_r and f_s is switching frequency of inverter. Large value of turns ratio n is desirable to reduce current in each primary side component. At ZVZCS, the tank

circuit impedance is minimum at $Z_{eq} = (r_{ac} + esr + R_{eq})$ where the value of n is maximum at n_{max} , at,

$$n_{max} = \frac{n_p}{n_s(min)} = 1.1 \frac{V_{pri}(max)}{(r_{ac} + esr + R_{eq}) i_L} \quad (4)$$

For constant coil current, the turns ratio at n_{max} minimizes the current in primary winding and the length of secondary winding is reduced, they together help minimize copper loss P_{cu} .

There are more reasons to choose the ZVZCS topology because it minimizes the power loss in the inverter by eliminating switching losses in IGBTs Q2-Q5. It also optimizes the conduction loss because there is no circulating current in inverter, the RMS current in capacitor C2 is also minimum. Moreover, it simplifies the equation of power delivery not only to IHT as well as to the output. For square wave voltage input V_{pri} at primary and with sinusoidal input current i_p the input power P_{in} could be expressed as,

$$P_{in} = 0.9 V_{pri} i_p \quad (5)$$

B. Core Geometry and Material on Inverter Topology

For balanced square wave primary V_{pri} (see Fig. 2), the primary current i_p consists of two components – the triangular wave magnetizing current i_m plus the reflected sinusoidal coil current i_L . In ZVZCS condition, i_L is in phase with V_{pri} . The peak of i_m i.e., I_m occurs when a pair of devices of inverter switching off. For ZCS operation of inverter switches, the value of I_m should be small. It is expressed as,

$$I_m = \frac{l_{mag} B_m}{\mu_0 n_p \mu_r} \quad (6)$$

Where l_{mag} is mean length of core having relative permeability μ_r . For minimum value of n_p (1), the core should have large value of μ_r (6) where high permeable toroidal cores are suitable.

Fig. 2: Magnetizing current i_m could make impact on ZCS operation

C. Core Material and its Geometry on Power Delivery

The power P_{OUT} transferred to load at coil current i_L is,

$$P_{OUT} = i_L^2 R_{eq} \quad (7)$$

When the coil is loaded, due to secondary effect of load the inductance L_1 drifts to, say, L_{eq} . Expressions of R_{eq} and L_{eq} are,

$$R_{eq} = \frac{\omega_s^2 M^2 R_{wp}}{R_{wp}^2 + \omega_s^2 L_{wp}^2} \text{ and } L_{eq} = L_1 - \frac{\omega_s^2 M^2 L_{wp}}{R_{wp}^2 + \omega_s^2 L_{wp}^2} \quad (8)$$

L_{wp} and R_{wp} respectively are inductance and resistance of workpiece, M is mutual inductance between coil and workpiece and $\omega_s = 2\pi f_s$ is angular switching frequency of inverter.

When the coil is loaded, the drifted resonant frequency f_{eq} is,

$$f_{eq} = \frac{1}{2\pi\sqrt{(L_{leak} + L_{eq})C_r}} \approx f_s \quad (9)$$

L_{leak} is leakage inductance of TR1 plus the stray inductance of extra cable in the circuit affecting the tank circuit. For efficient power transfer the loading of coil should make direct impact on f_s . Large value of L_{leak} would be a hindrance to effective power transfer. Here, as well, high permeable toroidal shaped cores are more suitable.

D. Testing of 30 kW IHT at Rated Load

To understand the functioning of IHT feeding a resonant tank circuit, an air-cooled 30 kW medium frequency induction heating controller was built. The complete system is shown in Fig. 3. Details of the experimental set up, including the detailed parameters of transformer and tank circuit, are listed in Table I. The resonant frequency of the tank circuit was 12.1 kHz. To understand the loss characteristics of IHT the system was tested at full load. To determine the value of μ_r of the core [16] accurately (6) at the operating value of B_m , the value of I_m was first determined by keeping the secondary open.

TABLE I
DETAILED PARAMETERS OF TANK CIRCUIT AND 30 kW IHT

Parameter	Value
Primary and secondary turns	9 and 3
Core material of IHT	Nanocrystalline
Steinmetz parameters K_s, α, β [16]	1.015, 1.4364, 1.638
Value of μ_r at $B_m=0.458$ T	29780
Ribbon thickness, μm	30
Peak magnetizing current I_m , A	0.42
Mean length of core l_{mag} , m	0.29
Total core area, cm^2	21.6
Total core weight, kg	4.64
Power input to IHT @460V, 75A, W	31000
Core loss P_{core} @0.458 T, 12.9 kHz, W	51.6
Total copper loss P_{cu} , W	94.3
Efficiency for sinusoidal excitation, %	99.53
Thermal conductivity of core, W/mK	10
Capacitor Cr, μF and its esr, m Ω	6.5, 0.3
Inductance of coil head L1, μH	26.5
Rated coil current, A	225
Current density in winding conductors, A/mm ²	3.80
Strand dia. of litz wires in both windings, mm	0.2

Fig. 4a shows the waveforms of the inverter at rated load of 30 kW. At rated output the value of V_{pri} was 460V where the calculated value of B_m was at 0.458T. Fig. 4b shows the waveforms of V_{pri} and i_p (or the magnetizing current i_m) when the secondary was kept open. At B_m of 0.458T, the value of I_m was 0.42A and the corresponding value of μ_r of core was 29780.

Fig. 3: Complete 30 kW air-cooled induction heating system, L1 not loaded

Fig. 4: a) Waveforms of V_{rip} and primary current of IHT at input power of 30.0 kW to IHT at frequency 12.9 kHz, and, b) the corresponding magnetizing current found with secondary open is linear at B_m of 0.458T with I_m of 0.42A.

III. LOSS CHARACTERISTICS OF IHT

The total power loss in a transformer, mostly, consists of two parts – the core loss P_{core} and the copper loss P_{cu} . Individual loss components are expressed as,

$$P_{core} = W_{core} K_s f_s^\alpha B_m^\beta \quad (10)$$

$$\text{and } P_{cu} = i_L^2 \left(\frac{R_{pri}}{n^2} + R_{sec} \right) \quad (11)$$

W_{core} is core weight. K_s, α and β are Steinmetz parameters. R_{pri} and R_{sec} respectively are the high frequency ac resistance of primary and secondary windings.

At full load condition, the power loss would increase the temperature of core, copper and insulation. The distribution of temperature on various parts of cores and windings depends on respective heat conduction characteristics. For complex transformer arrangement, finite element method (FEM) [8] is often used not only to determine the temperature distribution correctly, it is also used to find the hot spot temperature and its location. This approach is useful for transformer optimization.

Contrary to conventional transformers, windings of IHT are loaded to their respective rated current value because the coil is

kept energized to facilitate power transfer to job as and when needed. The magnitude of B_m depends on quantum of power extracted through coil L1. Here, P_{cu} is fixed at rated value, but P_{core} depends on quantum of power being transferred.

Ideally, power loss needs to be minimized. P_{core} depends on chosen soft magnetic material, waveform pattern of primary voltage, etc. whereas P_{cu} depends on strand size of litz wire, winding configuration, waveform pattern of current in windings, proximity effects and winding temperature. Using litz wire of proper strand size along with laying both windings in single layer would reduce the impact of high-frequency effects, P_{cu} is also reduced [4]. To simplify the thermal modeling and for better heat conduction, conductors in primary and secondary windings are placed in single layers [8].

A: Copper Loss Characteristics of IHT

Under ZVZCS conditions the current waveform in both windings is pure sinusoidal. Secondly, the impact of skin effect on copper loss is ignored because the skin depth is much larger than the diameter of litz wire strands. For such conductors the impact of proximity on copper loss could as well be neglected if each winding is laid in single layer [4].

In SRI fed IH system, even at no load, the value of P_{cu} could be at its rated value. Therefore, for energy efficiency, its value should be minimized first. Eqn. (11) could be re-written as,

$$P_{cu} = k_{pri} \frac{l_p}{A_p} \rho l_p i_p n_p + k_{sec} \frac{l_L}{A_s} l_p \rho n_s i_L \quad (12)$$

Where k_{pri} and k_{sec} respectively are ac resistance factor of primary and secondary conductors and A_p and A_s are their respective copper area, l_p is mean length per turn of winding and ρ is resistivity of copper. With both windings laid in single layers and with the designed ratio of diameter (=0.2mm) of strands of round litz wire conductors to skin depth (=0.571mm) at 0.35, the impact of skin and proximity effects on ac resistance value of windings would be negligible [4], [6]; the value of both k_{pri} and k_{sec} is considered at unity [4], [6]. Further, considering the same value of current density J for both the windings, (12) may be simplified as:

$$P_{cu} = J \rho l_p (i_p n_p + n_s i_L) = 2J \rho l_p n_s i_L \quad (13)$$

Cost factor needs optimal use of copper, if permitted by thermal considerations, the value of J should be large. For minimal copper volume, both $n_p l_p$ and $n_s l_p$ should be at respective minimum value. It helps IHT to have minimum number of layers in conductor layout to have reduced impact of high-frequency proximity effects on P_{cu} and also better features of heat removal. Laterally, the leakage inductance L_{leak} (9) is also reduced, inter-winding capacitance could be negligible.

B: Core Loss Characteristics of IHT

The expression of core loss in (10) is universally used for characterization of loss in soft magnetic materials where the Steinmetz parameters are derived from the loss curves (valid for sinusoidal excitation) given in the product datasheet. Conventionally, the core loss density P_c (W/kg) [17] consists of hysteresis loss P_h and eddy current loss P_{el} , as expressed below,

$$P_c = P_h + P_{el} = k_h f_s B_m^p + k_e f_s^2 B_m^2 \quad (14)$$

For a particular material, parameters k_h , k_e and p are considered as constant. However, there exists large difference between the measured core loss and the mathematically derived formula (14) which is derived with the consideration that B_m is uniformly distributed inside the core. However, even for perfect winding layouts, the skin effect influences the distribution of B_m and hence the P_c (14) in high permeable conductive nanocrystalline ribbon type cores. There is large difference between the measured and calculated core loss. The said difference is more for cores of larger ribbon thickness and/or at higher frequency of primary excitation [18]. To match the actual core loss a new component of magnetic loss i.e., the excess eddy current loss P_{ex} [17]-[21] has been introduced. It occurs due to non-uniform distribution of flux density in ribbons or core. The expression of P_{ex} is [19], [21],

$$P_{ex} = k_{ex} f_s^{1.5} B_m^{1.5} \quad (15)$$

The parameter k_{ex} depends on ribbon thickness, conductivity of material, winding pattern, etc. P_{ex} is more prominent for cores of higher thickness as well as at higher switching frequency.

Selection of core material plays an important role in optimizing the core loss. Nanocrystalline cores are suitable for reasons listed below [3], [5], [7], [15]:

1. Large value of saturation flux density B_{sat}
2. Small core loss density, W/kg
3. Large value of μ_r at operating flux density B_m helps reduce the value of I_m
4. Large value of I_m could introduce an error in sensing of phase angle between voltage V_{pri} and current i_p
5. Large thermal conductivity could avoid creation of any hot spot in the core and, importantly,
6. Robust parametric behavior; parameters of cores such as B_{sat} and μ_r , etc. are virtually insensitive to operating conditions of core and temperature

IV. EVALUATION OF IHT AT FULL LOAD

Reliability testing is critical aspect in power electronics product design. It is of particular interest for customized products such as air-cooled IHT, it has multiple heat sources with complex heat conduction characteristics. For reliable performance, accurate evaluation of temperature profile of core and two windings at rated load conditions is required. Finite element method [8] has been used for such purposes. Though, it is useful for complex model but accurate modeling of power loss and its distribution power loss as well as of dynamic heat conduction characteristics is essential. Ideally, to evaluate reliability parameters the transformer needs to be subjected to full load conditions. However, prolonged full load testing of high-power equipment is difficult because of following reasons:

1. The loading arrangement is application dependent, best suited at customer's site
2. Loads are functional till Curie temperature, otherwise,
3. For heat run test, it needs strong cooling arrangement that, as well, consumes extra energy
4. The arrangement of full load test is elaborate, and,
5. There is large waste of energy

Therefore, for heat run test of IHT, in particular, an alternate approach that consumes much less energy is desirable. It should be simple as well as not dependent on the actual job or load.

Fig. 5: Views of single-layer overlaid windings, such configuration has small value of leakage inductance and better distribution of B_m [22]

A. No Load Testing of Induction Heating System

To facilitate energy transfer by induction heating effect the coil L1 is kept energized to its rated capacity. Irrespective of loading condition the value of P_{cu} is at its rated value because each winding also draw respective rated current at resonant frequency of unloaded tank circuit. However, the rated copper loss alone cannot be used to validate the design of the copper circuit. The core loss and its distribution make large influence on thermal conduction characteristics of windings [8]. The optimized design of IHT, therefore, should not only minimize the value of P_{core} but also ensure its uniform distribution [22] in core. In that sense, the no load testing could indirectly be used to study the quality of design of IHT. Ideally, the value of resonant frequency f_r (16) of unloaded tank circuit should be maximum and the corresponding value of B_m or V_{pri} (1) should be minimum. Interestingly, the ZVZCS inverter is equipped to differentiate between different winding configurations and can decide what is best suited. Here, for best winding configuration, the value of V_{pri} at no load i.e., $V_{pri}(nl)$ would be minimum (i.e., B_m is minimum) and the phase locked loop would result maximum value of f_r . It means the leakage inductance L_{leak} (9) is at minimum value, because the expression of f_r is,

$$f_r = \frac{1}{2\pi\sqrt{(L_{leak}+L_1)C_r}} \quad (16)$$

The power input $P_{in}(nl)$ at no load to the transformer is,

$$P_{in}(nl) = 0.9V_{pri}(nl)i_p \quad (17)$$

Fig. 6: Waveforms of V_{pri} and i_p when the coil was not loaded, a) for perfectly overlaid windings where average value of V_{CH} was 12.3V and f_r was 12.15 kHz, and, b) when secondary was shifted by 30° where average V_{CH} was 13.2V and f_r was 11.9 kHz, but P_{core} injected was 60.7 W.

The no load waveforms of the actual system are shown in Fig. 6a. The measured value of $V_{pri}(nl)$ was approximately at 9.1V where the value of B_m was 9.6mT and the corresponding value of P_{core} was approximately 18mW/kg. Therefore, at no load, for

perfectly overlaid windings (see Fig. 5) the core loss P_{core} could be neglected.

Fig. 7: Single layer primary and secondary windings not laid one above the other, have non-uniform distribution of B_m in the core [22]

B. Injection of Rated Core Loss into the Core at No Load

The value of B_m at full load condition was noted at 0.458T (see Fig. 4a and Table I). Parametric details of nanocrystalline cores, used in IHT, are listed in Table I [16]. The corresponding value of P_{core} for rated sinusoidal excitation was 51.6 W. Can this power or the quantum of power equivalent to square wave excitation ($>51.6W$), with both windings drawing respective rated current, be injected into the core even when the coil is not loaded? The answer is yes. Though, using FEM method [22], it can be accurately predicted, here, a practical approach is followed to inject the requisite core loss.

It was detailed in the last sub-section that for perfectly overlaid windings (see Fig. 5) i.e., when single-layer secondary was laid over the single-layer primary the leakage flux was minimum and B_m was uniformly distributed on each ribbon throughout its periphery [22]. The prospect of localized eddy current loss was small and the value of P_{ex} was also negligible [17]-[21]. It is known that if one winding is shifted from the other then there is increase in leakage inductance by, say, l_1 and the distribution of flux density becomes less uniform [22]. The prospect of local eddy current loss is more and so is for P_{ex} . ZVZCS SRI is equipped to track both the changes accurately. The excess leakage would result in decrease in f_r to say f_{r1} , like,

$$f_{r1} = \frac{1}{2\pi\sqrt{(L_{leak}+l_1+L_1)C_r}} \quad (18)$$

Similarly, the extra core loss would be reflected in an increase in no load voltage primary voltage from $V_{pri}(nl)$ to *pseudo-load* condition, say, $V_{pri}(pl)$. They are related as,

$$0.9V_{pri}(pl)i_p - 0.9V_{pri}(nl)i_p = P_{ex} + \Delta P_{el} \quad (19)$$

$$\text{Or, } V_{pri}(pl) - V_{pri}(nl) = \Delta V_{pri}(pl) = \frac{(P_{ex} + \Delta P_{el})}{0.9i_p} \quad (20)$$

Table II details the situation, practically explored, when the angle between axis of secondary and that of primary is gradually increased from zero. It appears that there is loss in uniformity in B_m , it is gauged by an increase in leakage inductance l_1 , and the worst-case situation [22] occurs when the angle is 180° (see Fig. 7).

TABLE II
IMPACT OF SHIFTING SECONDARY FROM PRIMARY WINDING

Angle of shift	Frequency f_r , kHz	Injected P_{core} , W	Equivalent to operating B_m , T	Extra leakage l_1 , μH
0 degree	12.1	0.076	0.0096	0.0
30 deg.	11.85	60.7	0.545	1.2
90 deg.	11.60	106.9	0.784	2.4
180 deg.	11.50	115.0	0.826	2.9

Fig. 8: Waveforms of V_{pri} and i_p when the coil was not loaded, a) the shifted angle between primary and secondary was 90° where average value of V_{CH} was 13.8V and f_r was 11.6 kHz P_{core} and injected was 106.9 W, and, b) for 180° shift, V_{CH} was 14.0V, f_r was 11.5 kHz and P_{core} injected was 115.0 W.

Table II details, for the designed 30 kW IHT, the implication of shifting the secondary winding relative to the primary. Corresponding waveforms of V_{pri} , i_p and i_L are shown in Fig. 6b, Fig. 8a and Fig. 8b. When the secondary winding was shifted by 180° relative to the primary (see Fig. 7), then it resulted maximum nonuniformity of distribution, it injected more core loss (24.8 W/kg) and the value of l_1 was also more. In this application the calculated value of P_{core} for fully loaded IHT considering sinusoidal excitation was 51.6W. However, when square wave voltage was applied at primary, the core loss would be more [5]. In this set up, at rated value of i_p and i_L , it was practically found that by shifting secondary approximately 30 degree away from the primary the requisite core loss (here it was 60.7W) could be injected. The corresponding waveforms are shown in Fig. 6b. However, the extra loss occurred in the core in a nonuniform manner, a bit localized. Therefore, the cores under consideration should have large thermal conductivity where nanocrystalline material is suitable. The extra loss of 60.7W would get reflected in an increase in reading of V_{pri} i.e., ΔV_{Pri} . Therefore, at primary current of 75A, to inject core loss of 60.7W, the value of ΔV_{Pri} is,

$$\Delta V_{Pri} = \frac{60.7}{0.9i_p} = 0.89 \quad (21)$$

Above mentioned condition has been established experimentally, it is shown in Fig. 6b.

Now, the energized inverter was tested for 90 minutes. The enclosure (see Fig. 3) was forced air-cooled. Moreover, to avoid any loading of coil L1, during experimentation, it was kept away from any metallic object. The temperature reading was taken at multiple locations on each core segment as well as on copper windings. At injected core loss density of 13.1 W/kg and current density of 3.8A/mm² the maximum temperature noted on core was 81.3 °C and on windings it was 79.5 °C. The ambient temperature was 33 °C. In post weld heat treatment applications [14], for soaking of welded pipes at 750 °C, similar IHT delivers full capacity continuously for more than 10 hours.

V. HEAT RUN TEST OF IHT AT ZERO OUTPUT POWER

In the practical demonstration as detailed above the load drawn from IHT was not zero, rather it fed the power wasted in the coil L1 as well as in Cr. The power loss could have added error due to impact of high frequency and also of temperature caused by power loss in L1. In the proposed method, L1 is eliminated and the leakage inductance ($L_{Leak} + l_1$) of IHT is now

the only inductance in the tank circuit. Here, the full load performance of IHT would be evaluated even though the power drawn from it is zero because the secondary is shorted. Using suitable inverter, if following conditions are met then requisite core loss would be injected into the core and full load copper loss would remain unchanged:

1. Current i_L and i_p should be same as used for Fig. 6
2. The frequency of current must remain unchanged, and,
3. The angle of 30° between the windings remains same.

In the proposed evaluation a low power ZVZCS SRI, in half bridge configuration, was developed where IHT with shorted secondary constituted the tank circuit inductance as shown in Fig. 9. The primary and secondary currents were kept unchanged as used for results with waveforms shown in Fig. 5 and Fig. 7. To keep f_r of (18) unchanged, C_r has been accordingly modified to C_{r1} (see Table III), so that,

$$f_{new} = f_r = \frac{1}{2\pi\sqrt{(L_{leak} + l_1)C_{r1}}} \quad (22)$$

l_1 is the extra leakage inductance of IHT under test, contributed by shifting one winding relative to the other. The parametric details of the new low power inverter are listed in Table III. The power input to input of TR (see Fig. 9) is,

$$P_{TR} \approx 0.45V_{CH}i_r \quad (23)$$

Fig. 9: Schematic of low power half-bridge ZVZCS SRI to emulate as full load testing infrastructure for high power IHT

TABLE III
SYSTEM PARAMETERS TO EMULATE FULL LOAD TESTING OF IHT

Parameter	Value
Primary and secondary turns of TR	8 and 1
Rectified input supply DC voltage V_s , V	55.0
Capacitors C3, μF , Voltage, V	6000, 63
Core material of TR	Ferrite
IGBT Q6, Q7, Q8	IRFP4321
Diode D1	DSSK 2 X 015A
Value of L_{leak} and stray inductance together, μH	5.8
Capacitors C4, C5, C6, μF	100
Tank capacitor C_{r1} , μF	30
Primary and secondary current of IHT, A	75 and 225
Current density in winding conductors, A/mm ²	3.8
Ambient temperature, °C	33.0

To emulate the full load testing of high power IHT, one 200W low-voltage high-current ZVZCS SRI was developed. Transformer TR was used as current multiplier so that low

current Mosfets (Q6-Q8) could be used. The complete low power set up is shown in Fig. 10. The IHT shown here had secondary shifted by 30° from primary, it is the same one already used to generate results for Fig. 6b.

Fig. 11 shows two sets of waveforms. Waveforms in Fig. 11a belong to the condition when the two windings were in perfect overlaid condition. But when the secondary was shifted by 30 degree relative to the primary winding, the corresponding waveforms for shifted windings are shown in Fig. 11b. The power drawn by the transformer TR was 200W.

Fig. 10: Complete experimental set up with necessary instrumentation of 200W ZVZCS half-bridge SRI to emulate full load testing facility for 30 kW IHT

The inverter was load tested for 90 minutes. As shown in Fig. 10, the IHT was kept in open air under natural or free convection at ambient temperature of 33°C . Using both RTD and infra-red thermometers (Fluke make), the temperature at different segments of ribbons of nano-crystalline cores and windings were noted. For temperature reading of cores, number of ventilating ducts were available in plastic casing of cores. When RTD probes were used, to avoid high frequency effects the temperatures were recorded off line. The readings of both the thermometers were matching. The maximum temperature on ribbon of core close to primary winding was recorded high at 90.5°C and that of copper winding was 85.5°C . Due to large thermal conductivity of nanocrystalline core material, the temperature variation at different parts of core was within 4°C .

Fig. 11: Waveforms of V_{pri} and i_p when coil was not loaded, a) for perfectly overlaid windings, the value of V_{CH} was 35V and frequency f_r was 11.9 kHz, and, b) when secondary was shifted by nearly 30° where average value of V_{CH} was 48.1V and f_r was 11.6 kHz. The injected core loss was 56.8W.

However, when 4-inch (90 CFM) fan was added for superior air circulation of IHT the temperature of core was much lower value at 81°C and that of winding was 75°C . The temperature rise was less than normally considered value of 50°C .

VI. CONCLUSION

The proposed article analyzed that due to application specific load characteristics the prolonged full testing of induction heating transformer is complicated. This article also analyzed that, even in no load condition, the copper loss at primary and secondary windings is at their respective rated value because they always draw respective rated currents at rated frequency. This article, using the concept of excess eddy current loss in core, proposed a novel idea of injecting magnetic loss equivalent to the rated core loss value in the transformer. Initially, the idea was practically validated, under no load testing condition of the system, by shifting the axis of one winding relative to the other. Its output power was the high-frequency copper loss of the coil head. The article, through introduction of a low-power series resonant inverter, further proposed an idea of repeating the same when the secondary of IHT was shorted. The new concept was also practically validated where just 200W resonant inverter was used to emulate the full load testing 30 kW transformer. At least for type testing, the proposed approach could be applied for other high frequency applications. However, as it involves non-uniform power loss inside the core, the proposed concept is more suitable for core materials having large thermal conductivity, e.g., nanocrystalline core material.

REFERENCES

- [1] W. G. Hurley, W. H. Wolfe and J. G. Breslin, "Optimized transformer design: Inclusive of high-frequency effects," *IEEE Trans. Power Electr.*, vol. 13, no. 4, pp. 651-659, July 1998.
- [2] M. A. Bahmani, T. Thirigner and M. Khareji, "Design methodology and optimization of a medium-frequency transformer for high-power DC-DC applications," *IEEE Trans. Ind. Appl.*, vol. 52, no. 5, pp. 4225-4233, Sept/Oct 2016.
- [3] A. K. Paul, "ZVZCS SRI guides optimal use of copper and core for air-cooled nanocrystalline transformer for induction heating," *IEEE Trans. Ind. Appl.*, vol. 56, no. 2, pp. 970-978, Mar/Apr 2020.
- [4] R. Petkov, "Optimum design of a high-power high-frequency transformer," *IEEE Trans. Power Electr.*, vol. 11, no. 11, pp. 33-42, Jan 1996.
- [5] W. Shen, F. Wang, D. Boroyevich and C. Wesley, "High-density nanocrystalline core transformer for high-frequency resonant converter," *IEEE Trans. Ind. Appl.*, vol. 44, no. 1, pp. 213-222, Jan./Feb. 2008.
- [6] I. Villar, A. Garcia-Bediaga, U. Viscarret, I. Etxeberria-Otadui and A. Rufer, "Proposal and validation of medium frequency power transformer design methodology," in *Proc. IEEE ECCE 2011*, pp. 3792-3799, IEEE.
- [7] A. K. Paul, "Structured protection measures for better use of nanocrystalline cores in air-cooled medium frequency transformer for induction heating," *IEEE Trans. Ind. Electron.*, vol. 58, issue 5, pp. 3898-3905, May 2021.
- [8] M. Mogorovic and D. Dujic, "100 kW, 10 kHz medium-frequency transformer design optimization and experimental verification," *IEEE Trans. Power Electron.*, vol. 34, issue 2, pp. 1696-1708, Feb 2018.
- [9] W. G. Ondaal and J. A. Ferreira, "A Thermal model for high-frequency magnetic components," *IEEE Trans. Ind. Appl.*, vol. 35, no. 4, pp. 924-931, Jul./Aug. 2016.
- [10] Z. Ouyang, J. Zhang and W. G. Hurley, "Calculation of leakage inductance of high-frequency transformers," *IEEE Trans. Power Electron.*, vol. 30, no. 10, pp. 5769-5775, Oct 2015.
- [11] L. Shih, Y. Liu, and H. Chiu, "A novel hybrid mode control for a phase-shift full-bridge converter featuring high efficiency over a full-load range," *IEEE Trans. Power Electron.*, vol. 34, no. 3, pp. 2794-2804, Mar 2019.
- [12] J. -M. Choi, B. -J. Byen, Y. -J. Lee, D. -H. Han, H. -S. Kho and G. -H. Choe, "Design of leakage inductance in resonant DC-DC converter for

- electric vehicle charger," *IEEE Trans. Magnetics*, vol. 48, no. 11, pp. 4417-4420, Oct 2012.
- [13] O. Lucia, P. Maussion, E. J. Dede and J. M. Burdio, "Induction heating technology and its applications: Past developments, current technology, and future trends," *IEEE Trans. Ind. Electron.*, vol. 61, no. 5, pp. 2509 – 2520, May 2014.
 - [14] A. K. Paul, "Active-controlled passive distribution of power offers efficient heat treating solution for quality arc welding joints of steel pipes," *IEEE Trans. Ind. Appln.*, vol. 55, no. 5, pp. 4958-4966, Sep./Oct, 2018.
 - [15] A. Hilal, M. A. Raullet, C. Martin and F. Sixdenier, "A comparative study: Dynamic and thermal behavior of nanocrystalline and powder magnetic materials in a power converter application," *Journal of Electronic Materials*, vol. 44, no. 10, pp. 3768-3776, 2015.
 - [16] Advanced Technology and Materials Co., "Datasheet of core type 1G-1207030, part no. C.NA004.1200403-P062".
 - [17] G. Bertotti, "General properties of power losses in soft magnetic materials," *IEEE Trans. Mag.*, vol. 24, no. 1, pp. 621-630, Jan 1988.
 - [18] Z. Li, K. Yao, D. Li, X. Ni and Z. Lu, "Core loss analysis of Finemet type nanocrystalline alloy ribbon with different thickness," *Progress in Natural Science: Materials Intl.*, vol. 27, no. 5, pp. 588-592, 2017.
 - [19] T. Kauder and K. Hameyer, "Performance factor comparison of nanocrystalline, amorphous and crystalline soft-magnetic materials for medium frequency applications," *IEEE Trans. Magnetics*, vol. 53, no. 11, pp. 1-4, Nov 2017.
 - [20] F. Flohrer and G. Herzer, "Magnetization loss of nanocrystalline soft magnets," in *Intl. Conf. on Rapidly Quenched and Metastable Materials, Conf. Series 144(2009) 012075*, IOP Publishing, pp. 1-5, 2009.
 - [21] Y. Zhang, M. C. Cheng and P. Pillay, "Magnetic characteristics and excess eddy current losses," in *IEEE IAS Conf. Rec.*, pp. 1-5, 2009.
 - [22] M. S. Sanjarinia, S. Saadatmand, P. Shamsi and M. Ferdowsi, "Analysis of various transformer structures for high frequency isolation applications," in *North American Power Symp.*, pp. 1-6, IEEE, 2019.