Modelling and Design of HF 200 kHz Transformers for Hard- and Soft-Switching Application

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Introduction

The paper deals with modelling and design of HF 200 kHz (and more) low–power transformers with determinated leakage inductances for both hard– and soft–switching applications. The first group features by small value of the leakage inductance contrary to second one demanding that value substantially higher.

The further aims of the design are minimized weights and dimensions of the transformers. Results of numerical simulations and experimental verifications of designed transformers operated in converter circuits are given at the end of the paper.

Modelling of transformers using OrCad Environment

OrCad environment is a powerful tool that offers all the advantages of modern software for development electronic devices. It contains many complex features for simulation, modeling, PCB design and at not least the editor of magnetic elements. HF inductance and HF transformer modeling is a basic step in preparing simulations of converters used in SMPS or related applications. Transformer model can be compiled on the basis of various considerations (Electric properties representation, Magnetic field radiation, EMC predictions, Heat transfer and others) and evaluate it according to various criteria. Criteria according to model evaluation of the transformer may include: accuracy, fidelity, stability, speed and many other.

Simple linear model

For simple and fast implementation, we compiled a model of HF transformer which is depicted in Fig. 1 based on the equivalent electrical circuit. The model consists of "analog components" (Resistance, Inductance, Capacitance) and linear magnetic coupling. These replacements of elements are trying to represent (at least in part) the real parameters of the HF transformer. As "all know" HF transformer is very complex device. It features many of nonlin-

earities, parasites and other properties. This model is not able to model the complex properties listed above, as we tried to keep it simple and linearized.
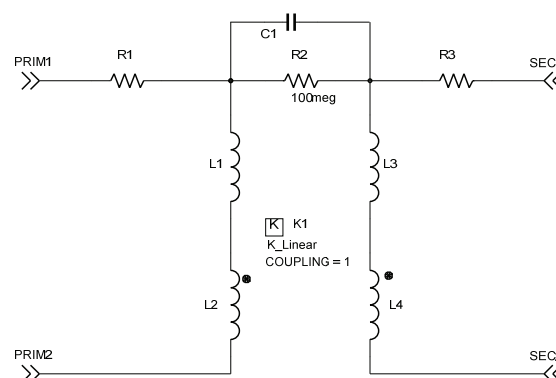

Fig. 1. Simple model of HF transformer with linear magnetic coupling

Meaning of each component in Fig. 1 is the following: R1 – primary winding serial resistance, L1 – primary leakage inductance, L2 – primary magnetizing inductance, R3 – secondary winding serial resistance, L3 – secondary leakage inductance, L4 – secondary magnetizing inductance, C1 – inter–windings capacitance, R2 – isolation resistance between windings, K1 – (K_linear) linear magnetic coupling between L2 and L4.

Basic features of the model in Fig. 1. can be summarized in the following lines. It is a simple model that is suitable for verifying the functionality of the initial simulation, in which can still be deleted elements C1 and R2. Since it is based on linear magnetic link magnetization inductances are constant depending on the current changes. This does not allow us to see the status of core by the saturation of transformer, effect of magnetic flux doubling and other phenomena influenced by the magnetic circuit of the transformer. Nevertheless, it is also possible to monitor the transformer from the perspective of the electric circuit. This means the slope of voltage, slope of current, voltage peaks, as well as Joule losses in the windings on various

loads of the transformer. The data obtained can then be used for the design of switching elements.

K_linear element is the most important component at simulation model, as it creates a magnetic coupling between the selected inductance (defined in the property editor). The parameter (coupling) of this element should be added that sets coupling efficiency (this mean coil overlap). This element can link up together up to 6 elements. The best results are given with the parameter settings COUPLING = 1, if set to less than 0.95 resulting waveforms shave unexpected shapes. The parameters of circuit elements of transformer can be determined by measurement, calculation or estimation. For identifying the parameters can be used following relations

$$L_{leak} \approx (0.001 \div 0.05) L_{mag} \ (H), \tag{1}$$

where L_{mag} is magnetizing inductance of individual windings and L_{leak} is their leakage inductance. Furthermore, the transformer ratio expressed by the number of turns is valid

$$n = \frac{n_s}{n_p}, \tag{2}$$

where n is the transformation ratio, n_p – number of primary turns, n_s – number of secondary turns, L_p – primary winding inductance, L_s – inductance of secondary winding.

If it is necessary to determine the transformer ratio of unknown transformer and we don't know the number of turns. So we can measure windings inductance, then can use the following equation

$$L = A_L \cdot N^2 \ (H) \Leftrightarrow N = \sqrt{L/A_L} \ (-), \tag{3}$$

where L – inductance, A_L – inductance per square turn, N– number of turns in general. Substituting equation 3 into equation 2 we get

$$n = \frac{n_s}{n_p} = \frac{\sqrt{L_s/A_L}}{\sqrt{L_p/A_L}} = \sqrt{\frac{L_s}{L_p}} \ (-), \tag{4}$$

where L_p – primary winding inductance, L_s – inductance of secondary winding.

Simple nonlinear model

Model including the characteristics of the core material in Fig. 3, consists of the same electrical circuit as the previous model. The main difference is in using component Kbreak, which defines the geometry of the transformer core (core area, length of magnetic path), the material (with hysteresis loops, initial permeability), number of turns for each inductance in magnetic link. To create a model to be followed the next procedure. In the model editor we have to create a new library. In the library you select a new model and a choice of MAGNETIC CORE. In the definition table, set the initial value of relative permeability. In our case for material 3F3 is the value of the 2000. Furthermore, it is necessary to write a few points from the first quadrant BH characteristics of material, just select few points from the lower part of the BH loop.

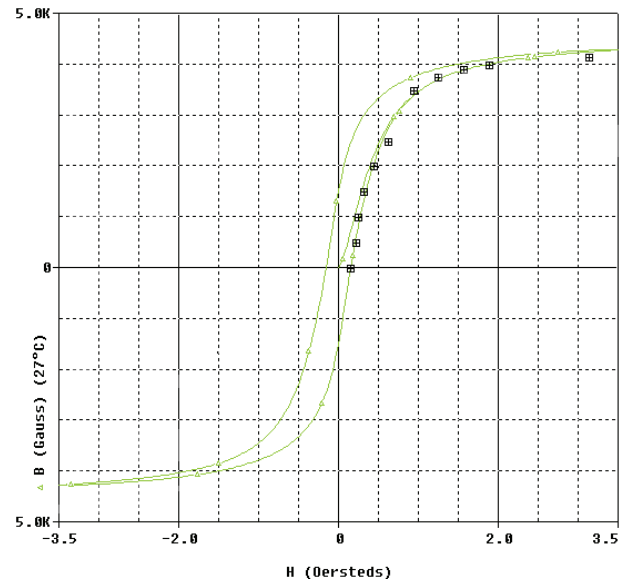Model editor automatically approximates the hysteresis loop and draws it in the first and third quadrant see Fig. 2.

Fig. 2. BH loop created by model editor for 3F3 material grade

All values needed for a model are obtained from datasheets of producer core material. Units placed in the datasheets are mostly in SI units. Since OrCad working with B in Gauss and H in Oersteds (Fig. 2.) so must be converted between physical units. After finished with the core model, we add a element Kbreak to equivalent circuit of the transformer in Capture tool. Further, it is necessary to edit the properties for element Kbreak and set its value to the model core inductors and names that will be tied to these elements. If we wanted to extend the model we can add more winding parasitic capacity C2 and C3 see Fig. 3. The values of these parameters can only be obtained by measuring with the resonance method with subsequent calculation or with Qmeter. If Transformer works on lower frequencies we can neglect these parasites.
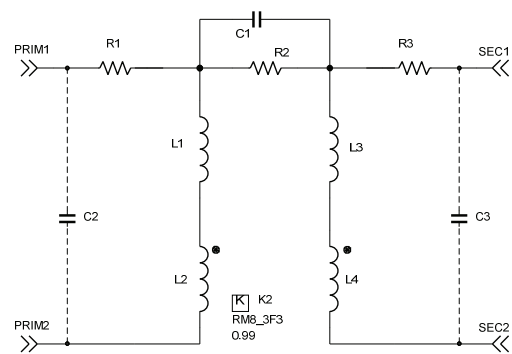

Fig. 3. Simple model of HF transformer with nonlinear magnetic coupling

Meaning of each component in Fig. 3 is the same as in Fig. 1. Elements of C2, C3, are parasitic capacity of individual windings.

What relates to mentioned mutual capacity exists between all the windings of the transformer. If modelled transformer with multiple windings it can be included into

the simulation model. For interesting add here Table 1. with the capacity values for different types of transformers [5].

Table 1. Measured values interwinding capacitance of HF transformer for various converter type

CORE	Type of converter	Interwinding capacitance
EFD20	Auxiliary suply – flyback	30 pF
PQ32	PFC Choke	69 pF
PQ32	Power supply – halfbridge	150 pF
E32/6/20	Planar transformer	200 pF

Design of HF 200 kHz transformer with determined ratio of leakage and magnetizing inductances

For proper design of transformer with integrated leakage inductance it is necessary to determine electrical parameters as well as target application of converter. We have focused on front–end converter for telecom servers, whereby main criterion is to operate at wide input range. Analysis of LLC converter is done by methematical simplification. The filtering action of the resonant network allows the classical fundamental approximation to obtain the voltage gain of the resonant converter, which assumes that only the fundamental component of the square–wave voltage input to the resonant network contributes to the power transfer to the output. Because the rectifier circuit in the secondary side acts as an impedance transformer, the equivalent load resistance is different from actual load resistance, whereby for its calculation, next equation is valid

$$R_{ac} = \frac{U_{RO}^F}{I_{ac}} = \frac{8}{\pi^2}\frac{U_O}{I_O} = \frac{8}{\pi^2}R_O, \qquad (5)$$

where R_{ac} is equivalent load resistance, U_{RO}^F is reflected output voltage, U_O is output voltage, I_O is output current and R_O is output load resistance.

Based on this mathematical simplification, we can start with first step of design. Main converter's parameters are: U_{IN} = 415 V (PFC output), f_{sw} = 200 kHz, P_O = 1500 W, U_O = 60 V, I_O = 25 A, efficiency = 0.94.

Estimated efficiency, input voltage (min, max – values):
$P = \frac{P}{E} = \frac{1500}{0.94} = 1595W$, $\quad U_{in}^{max}=U_{o.PFC}=415V$,

$U^- = \sqrt{U^- - \frac{2.P.T}{C}} = \sqrt{415 - \frac{2.1500.20e-3}{940e-6}} = 325V$ (where P_{in} is input power, U_{in}^{max} is maximum input voltage, U_{in}^{min} is minimum input voltage and $C_{PFC.OUT}$ is PFC output capacitance value).

Determination of maximum and minimum voltage gains of the resonant network:
$M^- = \frac{U^-}{U^-/2} = \sqrt{\frac{m}{m-1}} = \sqrt{\frac{6}{6-1}} = 1,1 M^- = \frac{U^-}{U^-}\cdot M = \frac{400}{306}\cdot 1,1 = 1,34$
(where M^{min} and M^{max} are minimum and maximum gains of resonant tank, m is ratio between primary inductance L_p

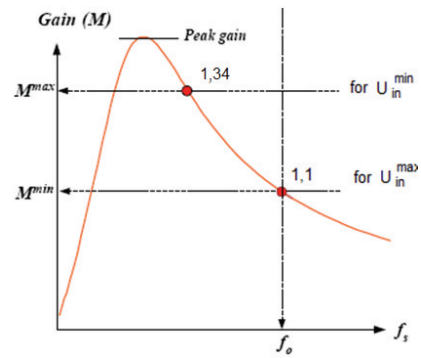and resonant inductance L_r, U_{RO} is reflected output voltage).

Fig. 4. Voltage gain characteristic with maximum and minimum gains

Determinations of transformer turns ratio: $n = \frac{N}{N}$

$= \frac{U^-}{2.(U^- + U^-)} = \frac{415}{2.(60+0,6)}.1,1 = 3,63 \cong 4$ (where U_O is rectified output voltage and U_f is voltage drop of output diodes).

Calculation of equivalent load resistance:
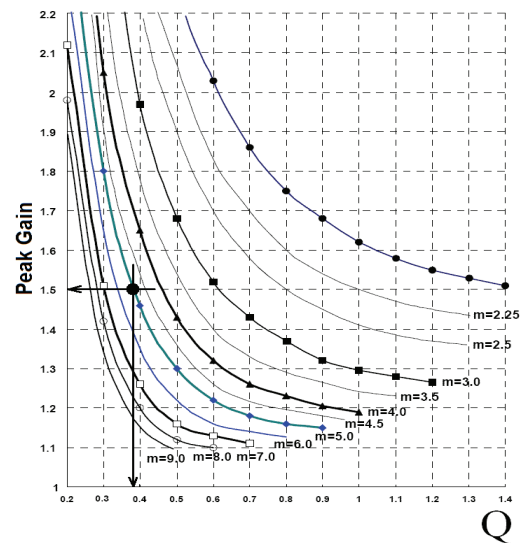$R_{ac} = \frac{8.n^2.U_O^2}{\pi^2.P_O} = \frac{8.4^2.60^2}{\pi^2.1500} = 26,66$.

Fig. 5. Voltage gain with maximum and minimum gains

Design of resonant network: for this purpose we have to establish maximum peak gain which is set to 15% higher than M^{max}=1,54. With this peak gain and for m = 6 the value of quality factor Q = 0,33 has been chosen. By selecting resonant frequency as 200 kHz, the resonant components are determined as:
$C_{-} = \frac{1}{2.\pi.Q.f_o.R_{ac}} = \frac{1}{2.\pi.0,3.200e3.25,67} = 100nF, L_{-} = \frac{1}{(2.\pi.f_o)^2.C_{-}} =$

$= \frac{1}{(2.\pi.200e3)^2.100e-9} = 6,3uH, L_- = m.L_- = 5.6,3e-6 = 37,5uH$
(where Lr and Cr are resonant inductor and capacitor, Lp is

primary inductance, Q is loaded quality factor, f_O is resonant frequency of Lr and Cr).

These five steps are key points of main components determination for simulation analysis and verification of design accuracy. Also region, in which converter will operate have to be established (Fig. 5). Therefore DC–gain characteristic have been made, whereby next equation is valid

$$M = \left| \frac{\left(\frac{\omega}{\omega_0}\right)^2 . \sqrt{m.(m-1)}}{\left(\frac{\omega^2}{\omega_p^2} - 1\right) + j.\left(\frac{\omega}{\omega_0}\right).\left(\frac{\omega^2}{\omega_0^2} - 1\right).(m-1).Q} \right|. \quad (6)$$
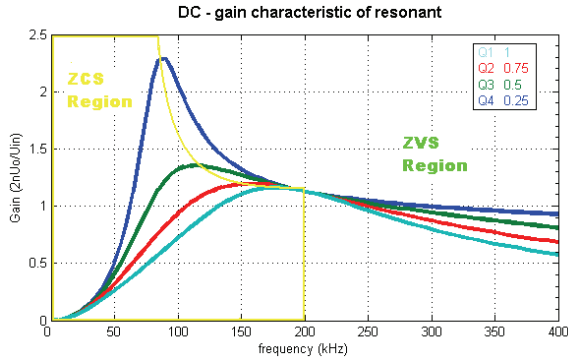

Fig. 7. DC gain characteristic of LLC converter for different Q values

Design of high frequency transformer. The worst case for transformer operation occurs at minimal switching frequency, what is valid for minimal voltage operation. Thus for Mmax = 1,34, fswmin = 120 kHz. Therefore transformer core selection have to be made for this frequency. For computation of core's area, next relationship was used $S = 25.\sqrt{\frac{P[W]}{f_{SW}[Hz]}} = 25.\sqrt{\frac{1500}{120e3}} = 2,79 cm^2$.

Selected core is ETD49, material 3C90, whose maximum width of flux density is $\Delta B = 0,4T$. Effective area $A_e = 220 mm^2$. Next we have determined minimal number of primary turns $N = \frac{n.(V + V)}{2.f.\Delta B.M.A} = 12$.

Then $N_P = n.N_S = 4.3 = 12 > N_{P min}$ and so secondary turns are: $N_S = 3$

Next figure is showing wiring diagram. By utilization of this configuration we have reached requested value of leakage inductance, so no additional inductance was necessary [1–3, 4].
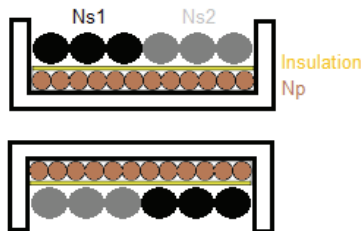

Fig. 8. Windings arrangement

Design of HF 200 kHz transformer with minimized lea–kage inductance

For a considerable range is not possible to show all calculation of the HF transformer, so we decided to accelerate the design using Viper Flyback converter design (VIPer). This program is designed for a series of intelligent power switches VIPER from STMicroelectronics company. As the name implies is designed directly for the design of "transformers" of Flyback type. Transformer is not an accurate term, because in this type of converter we can talk more about the two (or several) inductance with the common magnetic circuit. Nevertheless exhibit many of the transformer. Generally, we can noted that even well designed transformer may be because of its poor quality or poor technology of windings cause low efficiency of the whole system. The main reason comes in the leakage inductance. This ultimately causes a reduction of hardness of the transformer, formation of voltage spikes, increasing of switching losses in transistors and diodes. Also, all this reflected in EMC. Optimization done on the basis of several arguments and simplification. We think that the principle of leakage inductance is the same for a classic as well as the Flyback–type transformer. Neglecting air gap magnetic circuit. Simulating of the energies of magnetic fields made for a steady state in magnetostatic mode, skin and proximity effect is neglected.

After defining the requests for converter in program VIPer, we obtain the basic parameters of the transformer:

Input voltage range 88–264 VAC, Switching frequency 200 kHz, output 18 V/ 2,5 A – 45 W, total efficiency 91%. Output losses 1,9W, Clamper losses 0,71W, Power switch losses 1,7W, transformer losses 0,29W, $L_p = 166uH$, Primary leakage inductance 1,7uH, $I_{max} = 1,8A$ with $B_{max} = 181mT$.

Core Material: 3F3 – Ferroxcube, Geometry: RM8, Air gap 0,31mm, $n_p = 26$ turn – 0,63mm, $n_{s2} = 4$ turn – 0,063mm, $n_{s1} = 6$ turn – 1,4mm.

From the obtained data, we draw a section of a transformer in the program FEMM. Cross–section captures the view a distribution of the windings in transformer bobbins (used dimensions of winding wires see. previous). The dimensions of the effective core area were obtained from the datasheet. Due to the optimization we conducted a study of 3 basic winding arrangements. Whereby we watched change of energy of magnetic field dispersed in a winding and around the core (in the "air"). We assume that the energy of magnetic circuit (ferrite) is unchanged. In our considerations we rely on formula of energy of inductor, which is given

$$W_L = \frac{1}{2} L.I^2 \ (J). \quad (7)$$

There are in Fig. 9 and Fig. 10 shown simulation results of cross–section of transformer in different winding configurations. Simulations were carried for an immediate state of current density of winding, we chose 6A/mm2. The value 6A/mm2 corresponds to a magnitude of primary current at full power. Each arrangement corresponds to different values of energy of dispersed magnetic field, see Table 2. These values are obtained by integrating over the

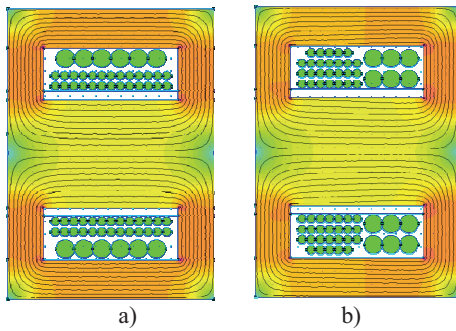selected area (winding, non–magnetic winding area and vicinity of the core) in the FEMM tool.

a)   b)
Fig. 9. FEMM simulated windings configurations
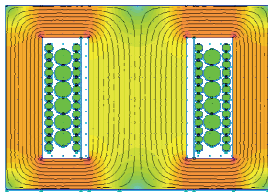

Fig. 10. FEMM simulated winding configuration called "sandwich"

Table 2. Results based on FEMM simulations

Winding type	Magnetic field energy of core	Leakage field energy
Fig. 9a	2,47e–6 J	5,79e–7 J
Fig. 9b	2,47e–6 J	2,88e–6 J
Fig. 10	2,47e–6 J	2,72e–7 J

If we compare the values with respect to equation 7, we can confirm that the transformer on Fig. 10. has the lowest value of leakage inductance and vice versa highest value of leakage inductance is transformer on Fig. 9b. These facts are known, but we want to say that with FEMM can obtain relatively accurate values energy of magnetic fields in the transformer. From energy fields we calculate the magnetization and leakage inductance using current density and the equation "7 ". The condition is good geometric replacement (exact cross–section) to maintain the shape of winding section and equivalent area of the core. The equivalent cross–section means so area which to allow even distribution of magnetic induction and subsequent exact calculations of energy. The above conditions must be observed as FEMM counts magnetic energy per unit volume of element. The base of element is drawing geometric replacement. We performed simulations (Fig. 9 and Fig. 10.) the above facts do not respect because it was not necessary to obtain accurate values, but only representative size for comparison [3, 5].

Simulation of transformers included into LLC converter circuit and comparison with measurement

For comparison purposes were compiled two simulation models LLC converter. The first used a transformer with linear magnetic coupling, second with nonlinear cou-
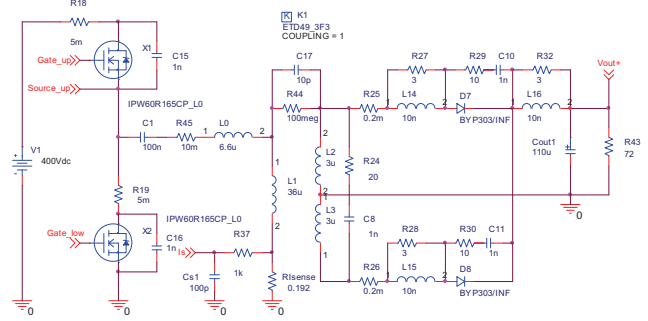
pling Fig. 11. A control part not included.

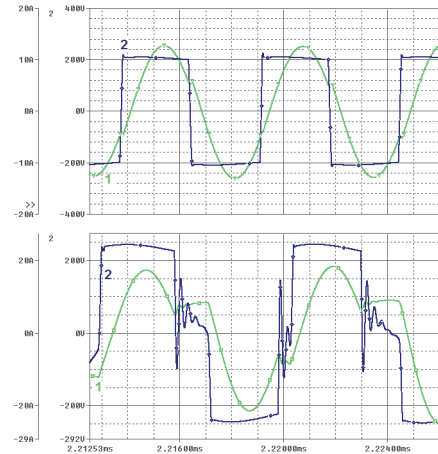**Fig. 11.** Simulation circuity of LLC converter (nonlinear transformer model)

Fig. 12. Comparison waveforms on primary winding LLC transformer (upper using K_linear, down Kbreak)
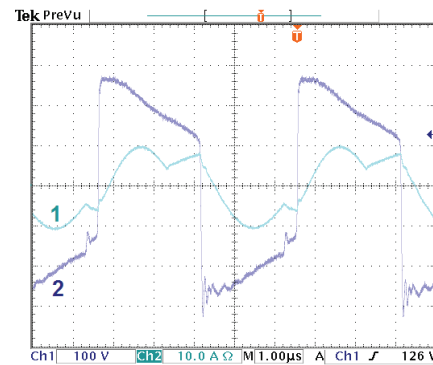

Fig. 13. Measured waveforms on primary winding LLC: 1 – current through primary winding, 2 – the voltage on the primary winding of the transformer

The aforementioned models were included in the LLC converter circuit and then simulations were performed. The resulting comparison waveform simulation is on Fig. 12. Waveform denoted "2" is primary winding voltage, "1" – current through primary winding. If we compare these waveforms is a significant difference between them. Waveforms in Fig. 12 upper are close to the theoretical waveforms LLC converter, while waveforms on Fig. 12 are down closer to the measured data on Fig. 13. Here you can see the major impact of nonlinear magnetic coupling, which positively influenced the result of the simulation.

11

Conclusions

This paper has demonstrated the way how to create linear and nonlinear simplified model of HF transformer in OrCad environment. It was shown by the simple nonlinear model has a relatively strong impact on simulation results. In this fact is necessary to create models which are more close to real behavior as possible. As previously mentioned one of the strategic parameter of most HF transformers is leakage inductance, because it may cause poor efficiency despite the design of transformer is correct. Using FEMM can evaluate the effect of winding arranged to leakage inductance and possibly with good accuracy to establish its value. In future works, we would like to focus on creation realistic HF transformer models. Also we want to improve in the use of Final Element Method Magnetic (FEMM) in order to achieve accurate results.

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R. Radvan, B. Dobrucky, M. Frivaldsky, P. Rafajdus. Minkštai ir kietai persijungiančio transformatoriaus HF 200 kHz modeliavimas ir projektavimas // Elektronika ir elektrotechnika. – Kaunas: Technologija, 2011. – Nr. 4(110). – P. 7–12.

Nagrinėjama modeliuojamo ir projektuojamo mažos galios transformatoriaus HF 200kHz pašalinės indukcijos įtaka minkštam ir kietam persijungimams. Siekiama sumažinti projektuojamo transformatoriaus svorį ir matmenis. Modeliavimo rezultatai patikrinti eksperimentiškai. Il. 13, bibl. 5, lent. 2 (anglų kalba; santraukos anglų ir lietuvių k.).