

Type of Document Technical Report	Document ID EPFL-PEL – EE-365 2026 – Report 2	Status Final Version
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Title R2: DAB CONVERTER – DESIGN OF MAGNETIC COMPONENTS		
Course Name EE-365 Power Electronics		
Keywords DAB Converter, Magnetics		

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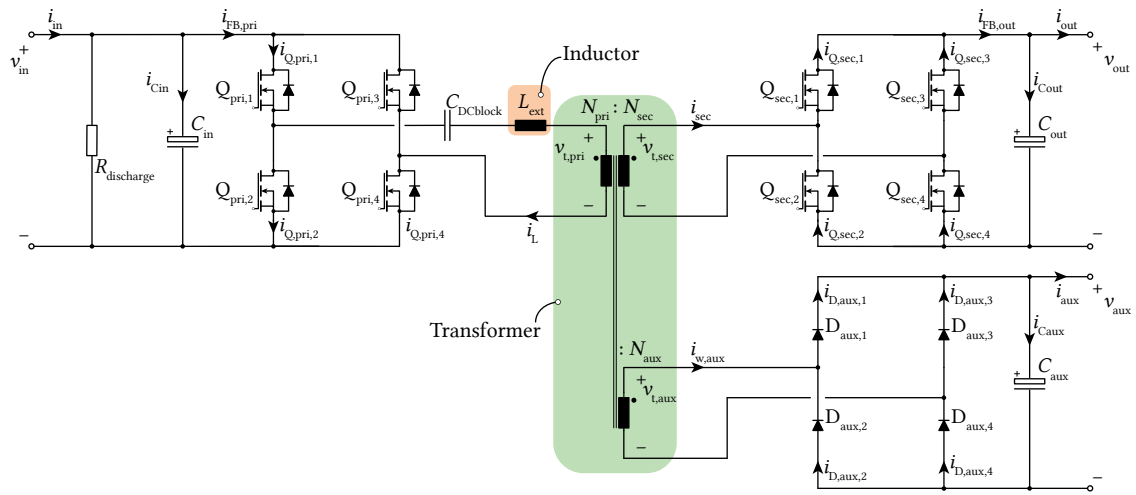


Figure 1 Schematics of the DAB converter for this report, highlighting the separation of the transformer and the inductor that will be designed separately in this report. Compared to Report 1, a winding of N_{aux} is also added to power the auxiliary circuits at the secondary side. More details about these aspects will come in the next reports.

TRANSFORMER – DESIGN SPECIFICATIONS AND CONSTRAINTS

The second part of the project deals with the design and realization of the magnetic components (i.e., the transformer and the inductor) of the DAB converter, according to the specifications and requirements from Report 1. The schematic to be considered for this report is shown in Figure 1. **Fill out Table 1** with the parameters assigned to your group and calculated in the previous Report.

One of the most important choices for the design of inductors and transformers is the selection of the material for the magnetic core. For all the groups in this class, the operating frequency of the converter are above 10 kHz. For such frequencies, materials known as ferrites are typically employed, while other materials are available for lower operating frequencies. Here, only the N87 ferrite material from TDK (datasheet pdf) is considered.

Another important guideline to follow for this project is that **the design of the transformer and the inductor are to be pursued separately**. In other words, **the transformer is designed to have maximum coupling** (i.e., with the lowest leakage and highest magnetizing inductance possible), while **the inductor is designed to obtain the desired inductance L** as calculated in Report 1. The transformer will be discussed first. In addition to what has been discussed in Report 1, the transformer will have to include an additional winding to power the secondary-side auxiliary circuitry. The auxiliary winding will be then connected in the next Report to a diode rectifier to generate a DC voltage $V_{aux,out}$ (reported in Table 1 together with the power requirement $P_{aux,out}$).

To successfully complete the second part of the project, it is **essential** to go through the corresponding lecture slides before starting to work on the report. To provide your answers and explanations, use the framed boxes below each question. For the questions where no box for explanations is provided, it is sufficient to write only the solution to the question without providing any additional reasoning steps. The design processes in this part of the project are **iterative**. You will (most likely) need to go back and rethink your previous decisions whenever you find out that some constraints are broken or yield operating conditions that are unfeasible or undesirable. You are asked **not** to report every iteration **nor** to provide a written account of all the iterations. **Report only** your final design choices taking care of checking that they comply with **all** the constraints given in the previous and following steps. Regarding the results of analytical calculations, please provide them with two digits after the decimal point in the unit provided in the solution box. Use the framed boxes below each question to insert your answers and explanations. Keep them concise and use the symbols indicated in Figure 1 (if a symbol is not present, define it).

Table 1 Given parameters and design requirements for the DAB converter.

Property	Value	Unit
$P_{\text{out,nom}}$	50	W
$V_{\text{in,nom}}$	50	V
$V_{\text{in,min}}$	30	V
$V_{\text{in,max}}$	60	V
V_{out}	24	V
$\Delta V_{\text{out,pp,rel}}$	2	%
f_{sw}	250	kHz
$V_{\text{aux,out}}$	18	V
$P_{\text{aux,out}}$	5	W

Q1: CALCULATION OF VOLT-SECOND STRESS

Volt-second stress (also known as peak flux-linkage) is the peak-to-peak value of the integral of the voltage applied to the transformer over a positive half cycle. This value is independent of the transformer design. You can consider constant input and output voltages during the positive half cycle. Evaluate the volt-second stress at the primary side under the worst-case condition (i.e., the highest value).

The magnetic flux linkage $\lambda(t)$ inside the core is defined by the following equation :

$$\Delta\lambda = \int v(t)dt$$

The Volt-Second Stress (λ_1) is the maximum accumulated area under this voltage-time curve during the positive half-cycle. This corresponds to $t_1 = \frac{T}{2} = \frac{1}{2} \times \frac{1}{f} = \frac{1}{2 \times 250'000} 2 \mu s$. Since we are considering the worst case scenario, we take $v = 60V$.

$$\lambda_1 = \int_0^{t_1} v(t) dt = 60 \times 2 \times 10^{-6} = 120 \times 10^{-6} Wb$$

$$\lambda_1 = 120 \times 10^{-6} Wb$$

/ 2 pt.

Q2: ESTIMATION OF LOSS COEFFICIENT

The Kg method used in this project requires the calculation of the core loss coefficient K_{fe} . The manufacturer of the core material (namely, TDK) provides the data reported in Table 2 that gives us a first estimate of the volumetric loss at 80 °C as

$$p_{vol} = K_b \cdot \left(\frac{f}{f_{base}} \right)^a \cdot \left(\frac{B_{pk}}{B_{base}} \right)^\beta \quad (1)$$

where f is the operating frequency, B_{pk} is the peak operating magnetic flux, f_{base} is the base frequency, B_{base} is the base peak magnetic flux, and a and β are the Steinmetz parameters.

Calculate the K_{fe} coefficient using the row of Table 2 with the nearest base frequency to your operating frequency.

Note: This relation assumes sinusoidal excitation on a toroidal core, but it is a reasonable estimation for your project.

Table 2 Core loss data from the manufacturer of material N87.

f_{base} (kHz)	a ()	β ()	K_b (kW/m ³)	B_{base} (mT)
25 kHz	1.44	2.93	67.48	200
50 kHz	1.72	2.66	23.51	100
100 kHz	1.72	2.68	61.15	100
200 kHz	1.839	2.57	30	50

We will be using the equation :

$$p_{vol} = K_{fe} \cdot (B_{pk})^\beta$$

By rearranging Equation 1, we can isolate K_{fe} as:

$$K_{fe} = K_b \cdot \left(\frac{f}{f_{base}} \right)^a \cdot \left(\frac{1}{B_{base}} \right)^\beta$$

The closest base frequency in Table 2 is 200 kHz. We extract the corresponding parameters:

- $f_{base} = 200$ kHz
- $a = 1.839$
- $\beta = 2.57$
- $K_b = 30 \text{ kW/m}^3 = 0.03 \text{ W/cm}^3$
- $B_{base} = 50 \text{ mT} = 0.05 \text{ T}$

Then we get:

$$K_{fe} = 0.03 \cdot \left(\frac{250}{200}\right)^{1.839} \cdot \left(\frac{1}{0.05}\right)^{2.57} = 99.766$$

$$K_{fe} = 99.77 \text{ W/(T}^\beta \text{ cm}^3)$$

/ 1 pt.

Q3: TOTAL CURRENT STRESS

Assuming a worst-case RMS current $I_{w,aux,RMS} = 0.8 \text{ A}$ in the auxiliary winding, calculate the total RMS current stress I_{tot} in the transformer (see the slides).

$$I_{tot} = 4.45 \text{ A}$$

/ 1 pt.

Q4: WINDOW UTILIZATION FACTOR

Besides the conductors in the windings, the core window area is also occupied by non-conductive materials. Make an estimation of the window utilization factor K_u , also known as *fill factor*, to have a starting point for your design. Be aware that this is only a first reference for yourself, and that once the core sample is wound, you will be able to improve this estimate for the next design iteration.

$$K_u = 0.6$$

/ 1 pt.

Q5: ESTIMATION OF LOSS BUDGET

Set a reasonable target efficiency for your design and calculate your allowed losses P_{tot} considering the nominal output power. Be aware that this is only a first estimate, and you will be able to improve it in the next iterations.

The allowed losses (P_{tot}) represent the maximum allowed power dissipation inside the transformer to meet our performance goals.

By setting an efficiency of $\eta = 90\%$ and with $P_{out,nom} = 50 \text{ W}$ as our target output power, we get :

$$P_{in} = \frac{P_{out,nom}}{\eta} = \frac{50}{0.90} = 55.56 \text{ W}$$

$$P_{tot} = P_{in} - P_{out,nom} = 55.56 - 50 = 5.56 \text{ W}$$

$$P_{tot} = 5.56 \text{ W}$$

/ 2 pt.

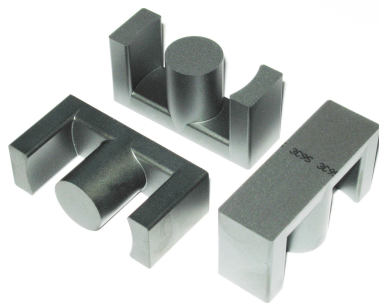
TRANSFORMER – CORE SELECTION WITH KG METHOD

As suggested in the slides, one of the most commonly used methods to design a transformer is the K_g -method modified for AC applications to take into account both core and copper losses and to minimize them. This method is usually referred to as the K_{gfe} method. More details can be found in the slides or in the book *Fundamentals of Power Electronics* by Erickson and Maksimovic, Springer.

Magnetic core manufacturers provide a vast variety of core shapes and sizes (examples are shown in Figure 2). In this course, a preselection of suitable cores is given in Table 3. The geometrical K_{gfe} factor for each of the available cores is also calculated, considering the geometrical dimensions given in the datasheets and $\beta = 2.7$ (average value for ferrites). You are requested to choose one of the available cores in Table 3 for your design.

Table 3 Available core materials and core shapes.

Core shape	N87 core	Coil former	Yoke Clamp	$K_{gfe} [cm^5]$
E20/10/6	B66311G0000X187	B66206B1110T001	B66206A2010X000	0.0019
E25/13/7	B66317G0000X187	B66208X1010T001	B66208A2010X000	0.0044
E30/15/7	B66319G0000X187	B66232B1114T001	B66232A2010X000	0.0049
E32/16/9	B66229G0000X187	B66230A1114T001	B66230A2010X000	0.0090
ETD34/17/11	B66361G0000X187	B66362X1014T001	B66362A2000X000	0.0120
ETD39/20/13	B66363G0000X187	B66364B1016T001	B66364A2000X000	0.0197
ETD44/22/15	B66365G0000X187	B66366B1018T001	B66366A2000X000	0.0290



(a) ETD shape,



(b) EE shape.

Figure 2 Exemplary core shapes.

Q6: MAXIMUM VALUE OF PEAK FLUX DENSITY

To avoid core saturation, the maximum peak flux density $B_{pk,max}$ must be constrained to a value much lower than the saturation value of flux density for the selected core material (here, N87) at any time during operation. The selection of this value is a trade-off between the use of the core (i.e., volume) and core losses. The magnetization curves of the available core material N87 are in this datasheet. To serve as a constraint in the following steps, select the maximum value of the peak flux density $B_{pk,max}$.

$$B_{pk,max} = 0.15 \text{ T}$$

/ 1 pt.

Q7: MINIMUM K_{GFE}

Copper resistivity can be considered constant at $\rho = 1.724 \cdot 10^{-6} \Omega \text{ cm}$. Considering again $\beta = 2.7$, calculate the minimum required K_g -factor $K_{gfe,min}$ for your specific conditions.

Note: while performing the calculations, be very careful to use the correct units.

We use the equation from the slides :

$$K_{gfe,min} = \frac{\rho \cdot \lambda_1^2 \cdot I_{tot}^2}{4 \cdot K_u \cdot P_{tot}} \left(\frac{K_{fe}}{P_{tot}} \right)^{\frac{2}{\beta}} \times 10^8$$

We have the following parameters :

- Copper resistivity: $\rho = 1.724 \times 10^{-6} \Omega \cdot \text{cm}$
- Volt-second stress: $\lambda_1 = 120 \times 10^{-6} \text{ V} \cdot \text{s}$
- Total current stress: $I_{tot} = 4.45 \text{ A}$
- Window utilization: $K_u = 0.60$
- Total allowed losses: $P_{tot} = 5.56 \text{ W}$
- Core loss coefficient: $K_{fe} = 99.77 \text{ W}/(\text{T}^\beta \text{cm}^3)$
- Steinmetz parameter: $\beta = 2.7 \implies \frac{2}{\beta} = 0.7407$

Substituting these values into the equation:

$$K_{gfe,min} = \frac{1.724 \cdot 10^{-6} \times (120 \cdot 10^{-6})^2 \times 4.45^2}{4 \times 0.60 \times 5.56} \left(\frac{99.77}{5.56} \right)^{0.7407} \times 10^8 = 3.13 \times 10^{-5} \text{ cm}^5$$

$$K_{gfe,min} = 3.13 \times 10^{-5} \text{ cm}^5$$

/ 3 pt.

Q8: SELECTION OF THE CORE

Indicate which core shape from Table 3 you want to use for your design, considering the calculated minimum K_{gfe} . As a first attempt, you can choose whatever core has a K_{gfe} bigger than the needed one. Bigger cores will reduce the power density of your converter, which is undesirable. Be aware that your core size also influences the current density and the losses, which will be analyzed in the following questions. Likely, you will need to iterate on your answer to meet all the constraints.

Core shape: E20/10/6

/ 3 pt.

Q9: VERIFICATION OF THE PEAK FLUX DENSITY

Using the formula presented in the slides, check that the expected ΔB (peak ac flux density, peak-to-average) in your operating conditions is not exceeding the $B_{pk,max}$ selected in Q6. This will ensure that saturation never occurs in nominal conditions. Information on the geometrical dimensions of the selected cores can be found in the provided files.

Looking at the datasheet of the chosen core, we get the following values :

- $MLT = 4.2 \text{ cm}$
- $A_c = 0.321 \text{ cm}^2$
- $W_A = 0.34 \text{ cm}^2$
- $I_m = 4.63 \text{ cm}$

We solve the equation to find ΔB :

$$\Delta B = \left[10^8 \times \frac{\rho \lambda_1^2 I_{tot}^2}{2 K_u} \frac{(MLT)}{W_A A_c^3 I_m} \frac{1}{\beta K_{fe}} \right]^{\left(\frac{1}{\beta+2} \right)} = 0.0901 \text{ T}$$

$$\Delta B: 0.09 \text{ T}$$

/ 2 pt.

Q10: REQUIRED NUMBER OF TURNS

Calculate the required turn ratio of the main and auxiliary windings and report it below. Then, using the formula presented in the slides, calculate the desired number of turns on the primary side to minimize total losses. Moreover, considering the transformer ratio required for your design, calculate the number of turns needed for the secondary and auxiliary windings. Make sure that each winding is composed of an integer number of turns. Then, calculate the **actual** resulting turn ratios and report them below, also checking that

this rounding does not make the turn ratio deviate significantly (anyhow, you will be asked to check this again later with a more detailed procedure).

Note: while performing the calculations, be careful to use the correct units.

$$N_1 = \frac{\lambda_1}{2 \times \Delta B \times A_c} \times 10^4 = 20.74$$

We compute the secondary and auxiliary windings :

$$N_{sec} = \frac{24}{50} \times n_{pri} = 9.95$$

$$N_{aux} = \frac{18}{50} \times n_{pri} = 7.47$$

As we chose $N_{pri} = 21$, the calculation for n_{aux} becomes : $\frac{18}{50} \times 21 = 7.56$, which is closer to 8.
With the actual windings, we compute the actual turn ratios :

$$n = \frac{N_{pri}}{N_{sec}} = 2.1, \quad n_{aux} = \frac{N_{pri}}{N_{aux}} = 2.625$$

$$\text{Calculated } n = \frac{50}{24} = 2.08$$

$$\text{Calculated } n_{aux} = \frac{50}{18} = 2.77$$

$$N_{pri} = 21$$

$$N_{sec} = 10$$

$$N_{aux} = 8$$

/ 3 pt.

$$\text{Actual } n = 2.1$$

$$\text{Actual } n_{aux} = 2.625$$

Q11: MAGNETIZING INDUCTANCE

The magnetizing inductance of the transformer (referred to the primary side) must be sufficiently high to avoid influencing the ideal DAB operation. Calculate the minimum magnetizing inductance $L_{m,MIN}$ to be 10 times higher than your DAB inductor L . Then, compute the expected value of the magnetizing inductance L_m . To do this, you can use the inverse of the formula for computing the number of turns. If L_m is smaller than $L_{m,MIN}$, you need to reconsider your choices and go back to the previous steps.

From Report 1, the DAB inductance is $L = 8.33 \mu\text{H}$, so the minimum magnetizing inductance is:

$$L_{m,MIN} = 10 \times L = 10 \times 8.33 = 83.3 \mu\text{H} = 0.083 \text{ mH}$$

The expected magnetizing inductance is computed by inverting the turns formula. Using the core permeability $\mu_r = 2200$ for N87 (from the datasheet : at 25°) and the E20/10/6 geometrical parameters ($A_c = 0.321 \text{ cm}^2$, $l_m = 4.63 \text{ cm}$):

$$L_m = \frac{N_{pri}^2 \cdot \mu_0 \cdot \mu_r \cdot A_c}{l_m} = \frac{21^2 \times 4\pi \times 10^{-7} \times 2200 \times 0.321 \times 10^{-4}}{0.0463} \approx 0.845 \text{ mH}$$

Since $L_m = 0.845 \text{ mH} \gg L_{m,MIN} = 0.083 \text{ mH}$, the constraint is satisfied and no design iteration is required.

$$L_{m,MIN} = 0.083 \text{ mH}$$

$$L_m = 0.845 \text{ mH}$$

/ 2 pt.

Q12: CHECK TURN RATIO FOR OUTPUT PORT

In Report 1, you were asked to plot the output power capability of the DAB as a function of the input voltage, considering the ideal turn ratio. The turn ratio that you obtained in the previous questions might deviate from the ideal one, since only integer numbers of turns can be realized. Therefore, you are tasked with plotting the output power capability (as in Report 1, in the answer box) considering the turn ratio that you intend to realize (Q10); provide also the minimum power transfer $P_{real,min}$ in the worst-case condition. To ensure that the converter has sufficient regulation capability, make sure that the realized turn ratio does not decrease the margin you took from Report 1 by more than 5 W.

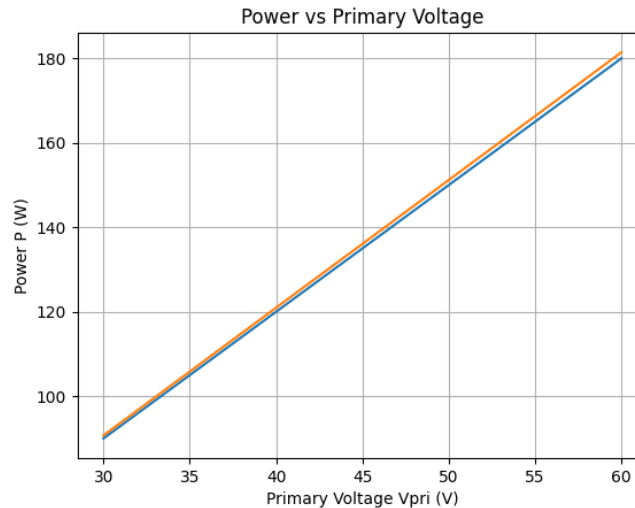


Figure 3

As seen in the picture above, the max $\Delta P = 1.44$ W and happens at P at $V_{pri} = 60$ V. We can also compute $P_{real,min}$ with the real turn ratio $n = 2.1$. This gives us :

$$P_{real,min} = \frac{n_{1,real} \times V_{pri,min} \times V_{sec}}{8 \times f_{sw} \times L} = 90.76 \text{ W}$$

This value can also be seen on the above plot. The difference with the ideal case of $n = 2.083$ (90.0 W) is around 0.76 W, which stays within the 5 W limit.

$P_{real,min} = 90.76 \text{ W}$

/ 3 pt.

Q13: CHECK TURN RATIO FOR AUXILIARY PORT

Considering the turn ratio that you intend to realize (Q10), calculate the voltage $V_{aux,true}$ of the auxiliary port. Now, compare this with the nominal value $V_{aux,out}$ and make sure the difference is below 15 %. If not, you need to reconsider your choices and improve the turn ratio.

Note: you can consider the nominal output voltage and ignore any non-ideality of the components of the converter (e.g., the voltage drop on the rectifier diodes).

$$V_{aux,real} = \frac{U_{nom}}{n_{aux}} = \frac{50}{2.625} = 19.05 \text{ V}$$

$$\frac{V_{aux,real} - V_{aux,out}}{V_{aux}} = \frac{19.05 - 18}{18} = 5.8 \%$$

The difference auxiliary voltage is $5.8 \% < 15 \%$.

$V_{aux,real} = 19.05 \text{ V}$

/ 1 pt.

TRANSFORMER – WIRE SELECTION

The following step in the transformer design is to select a suitable wire. Different aspects need to be considered in this choice. Increasing the wire cross section would reduce the resistance but, at the same time, requires a higher utilization factor (for the winding area) or a larger core. In addition, the high-frequency operation leads to a significant impact of the skin effect on the effective AC resistance. In the following tables, you will find a list of the available wires. You might consider to slightly revise the target fill factor (Q4) depending on your wire choice.

Table 4 Available magnet wires.

Wire number	M1	M2	M3	M4	M5	M6	M7
Conductor diameter (mm)	0.65	0.9	1.0	1.2	1.54	1.6	1.8

Table 5 Available Litz wires (the diameter of a single conductor is 0.2 mm).

Wire number	L1	L2	L3	L4
Number of strands	25	35	45	80
Total Conductive Area (mm ²)	0.78	1.10	1.41	2.51
External Diameter (mm)	1.53	1.8	2.06	2.8

Q14: SKIN DEPTH

To develop a sense of the specific weight of the skin effect on your design, evaluate the skin depth in millimeters.

From chapter 13 of the *Fundamentals of Power Electronics* book by Erickson and Maksimovic Ed. Springer, we get the following equation :

$$\delta = \sqrt{\frac{\rho}{\pi \mu_r \mu_0 f}}$$

with $\rho = 1.724 \times 10^{-8} \Omega \cdot m$ for soft-annealed copper at room temperature. This gives us a value of skin depth :

$$\delta = \sqrt{\frac{1.724 \times 10^{-8}}{\pi \times 1 \times (4\pi \times 10^{-7}) \times (250 \times 10^3)}} = 1.32 \times 10^{-4} \text{ m}$$

$$\delta = 0.13 \text{ mm}$$

/ 2 pt.

Q15: ALLOCATION OF THE WINDING WINDOW

Taking into account the RMS current expected in each winding, the number of turns, and the total current calculated in the first section, calculate the fraction α_w of the winding area that should be allocated to each of your windings.

From Q3, we know that $I_{tot} = 4.45 \text{ A}$ and we know that the fraction of area that should be allocated to the winding j is:

$$\alpha_{w,j} = \frac{N_j}{N_{pri}} \cdot \frac{I_j}{I_{tot}}$$

We get:

$$\alpha_{w,pri} = \frac{N_{pri}}{N_{pri}} \cdot \frac{I_{pri}}{I_{tot}} = \frac{21}{21} \cdot \frac{2.08}{4.45} = 0.47, \quad \alpha_{w,sec} = \frac{N_{sec}}{N_{pri}} \cdot \frac{I_{sec}}{I_{tot}} = \frac{10}{21} \cdot \frac{4.33}{4.45} = 0.46, \quad \alpha_{w,aux} = \frac{N_{aux}}{N_{pri}} \cdot \frac{I_{aux}}{I_{tot}} = \frac{8}{21} \cdot \frac{0.8}{4.45} = 0.069$$

$$\alpha_{w,pri} = 0.47$$

$$\alpha_{w,sec} = 0.46$$

$$\alpha_{w,aux} = 0.07$$

/ 2 pt.

Q16: WINDING AREA AND MEAN LENGTH OF TURNS

Considering the selected core, report the available winding area and the mean length of turns (MLT).

Note: to verify that you picked the correct data, you can calculate K_{gfe} and compare it to the one in Table 3.

$$A_w = 0.34 \text{ cm}^2$$

$$MLT = 4.2 \text{ cm}$$

/ 2 pt.

Q17: MAXIMUM WIRE CROSS-SECTION

Considering the dimensions of the selected core and the allocated fraction for each wire winding, calculate the maximum wire cross-section that would fit in the available space.

From chapter 15 of the book, we extract:

$$A_{w,j,max} = \frac{\alpha_{w,j} \times K_u \times W_A}{n_j}$$

We get:

$$A_{w,pri,max} = \frac{\alpha_{w,pri} \times K_u \times W_A}{n_{pri}} = \frac{0.47 \times 0.6 \times 0.34}{21} = 4.57 \times 10^{-3} \text{ cm}^2$$

$$A_{w,sec,max} = \frac{\alpha_{w,sec} \times K_u \times W_A}{n_{sec}} = \frac{0.46 \times 0.6 \times 0.34}{10} = 9.38 \times 10^{-3} \text{ cm}^2$$

$$A_{w,aux,max} = \frac{\alpha_{w,aux} \times K_u \times W_A}{n_{aux}} = \frac{0.07 \times 0.6 \times 0.34}{8} = 1.79 \times 10^{-3} \text{ cm}^2$$

$$A_{w,pri,max} = 4.57 \times 10^{-3} \text{ cm}^2$$

$$A_{w,sec,max} = 9.38 \times 10^{-3} \text{ cm}^2$$

$$A_{w,aux,max} = 1.79 \times 10^{-3} \text{ cm}^2$$

/ 2 pt.

Q18: TARGET CURRENT DENSITY

Densely packed multi-layer windings cannot dissipate heat as effectively as wires in free-air. In fact, there is a trade-off between copper losses, temperature rise, and power density. Considering the typical values provided in the slides, select a current density constraint as a starting point for your design. In the following steps of the design or prototyping process, you might adjust this constraint and re-iterate your design to change your design point with respect to the above-mentioned trade-off. Note that, due to discrete wire gauges, it will not be possible to exactly match the selected value.

$$J_w^* = 500 \text{ A/cm}^2$$

/ 1 pt.

Q19: REQUIRED CONDUCTOR AREA

Considering the target current density defined in Q18 and the RMS current expected in your windings according to the simulations performed in Report 1, evaluate the total conductive areas A_{TC} needed to satisfy your requirements. Compare this area with the maximum available in the core you selected (see the answer from previous question). If the available area is smaller than the needed one, go back and select a bigger core (Q8). Be aware that the core selection will also be influenced by the answers to the next questions. Note that if the needed area is only slightly bigger than the maximum available one, you can choose to proceed with your selected core; however, if the area is barely sufficient, the utilization factor that you will calculate in a following question (Q21) could determine that your design is not feasible.

From the slides, we get the equation:

$$A_{TC,j} = \frac{I_j}{J_w^*}$$

Computing this the different windings, we finally get:

$$A_{TC,pri} = \frac{I_{pri}}{J_w^*} = \frac{2.08 \text{ A}}{500 \text{ A/cm}^2} = 4.16 \times 10^{-3} \text{ cm}^2$$

$$A_{TC,sec} = \frac{I_{sec}}{J_w^*} = \frac{4.33 \text{ A}}{500 \text{ A/cm}^2} = 8.66 \times 10^{-3} \text{ cm}^2$$

$$A_{TC,aux} = \frac{I_{aux}}{J_w^*} = \frac{0.80 \text{ A}}{500 \text{ A/cm}^2} = 1.60 \times 10^{-3} \text{ cm}^2$$

Comparing these to the maximum available areas allocated in Q17 (4.57 , 9.38 , and $1.79 \times 10^{-3} \text{ cm}^2$ respectively), we see that every required area is strictly less than its maximum allocated space. Therefore, the E20/10/6 core is perfectly sized for our target current density.

Remark: Decreasing J_w^* to a value of 400 A/cm^2 would cause the primary and secondary to exceed their allocated areas. Hence a high current density is needed.

$$A_{TC,pri} = 4.16 \times 10^{-3} \text{ cm}^2$$

$$A_{TC,sec} = 8.66 \times 10^{-3} \text{ cm}^2$$

$$A_{TC,aux} = 1.60 \times 10^{-3} \text{ cm}^2$$

/ 3 pt.

Q20: WIRE SELECTION

Select the wires for your windings from the available wires in Table 4 and Table 5. In the answer box, indicate the wire of your choice with the label given in the tables and provide the current density you expect with that wire.

Notes:

- For Magnet Wires (Table 4), you can neglect the thickness of the isolation layer, i.e., you can assume that the diameter of the wire is the diameter of the conductor.
- For litz Wires (Table 5), the total conductive area is given in the Table.
- Your technology choice (magnet wire or litz wire) will affect your losses and the size of your transformer; you may want to reconsider your choice after calculating AC losses in a following question (Q23).
- Once again, consider the target current density from Q18 and the expected RMS current from Report 1, and use the expected total conductor area as a guideline.
- Due to the limited number of wires available, you may not be able to obtain exactly your target current density; choose a reasonably-close value considering the wire technology of your choice.
- Check that the area needed for your selected winding is smaller than the maximum available one; otherwise, reconsider your core choice (Q8).
- If you cannot satisfy the requirements with the available wires, you are allowed to parallel the windings. Only in this case, write the number of parallels and a letter x before the wire reference in the answer box. For example, "2xM7" means two wires of type M7 in parallel.

We must select discrete wires from the provided tables that respect the following :

1. They need match the A_{TC} requirements from Q19.
2. They need to have a diameter smaller than twice the skin depth δ found in Q14, if possible.
 - **Primary Winding** ($4.16 \times 10^{-3} \text{ cm}^2 = 0.416 \text{ mm}^2$): We select the solid magnet wire **M1** ($d = 0.65 \text{ mm}$).
 - Its conductive area is $A = \pi \times (0.65/2)^2 \approx 0.332 \text{ mm}^2 = 3.32 \times 10^{-3} \text{ cm}^2$.
 - Expected current density: $J_{pri} = 2.08 \text{ A}/0.00332 \text{ cm}^2 = 627 \text{ A/cm}^2$.
 - **Secondary Winding** ($8.66 \times 10^{-3} \text{ cm}^2 = 0.866 \text{ mm}^2$): Due to the higher current, Litz wire is highly recommended to mitigate AC resistance. We select **L1**.
 - Its conductive area is $A = 0.78 \text{ mm}^2 = 7.8 \times 10^{-3} \text{ cm}^2$.
 - Expected current density: $J_{sec} = 4.33 \text{ A}/0.0078 \text{ cm}^2 = 555 \text{ A/cm}^2$.
 - **Auxiliary Winding** ($1.60 \times 10^{-3} \text{ cm}^2 = 0.160 \text{ mm}^2$): The smallest available wire is **M1** (0.33 mm^2). We must use this.
 - Its conductive area is $A = \pi \times (0.65/2)^2 \approx 0.332 \text{ mm}^2 = 3.32 \times 10^{-3} \text{ cm}^2$.
 - Expected current density: $J_{aux} = 0.80 \text{ A}/0.00332 \text{ cm}^2 = 241 \text{ A/cm}^2$.

Selected wire 1: M1

$$J_{w,pri,selected} = 627 \text{ A/cm}^2$$

Selected wire 2: L1

$$J_{w,sec,selected} = 555 \text{ A/cm}^2$$

Selected wire aux: M1

$$J_{w,aux,selected} = 241 \text{ A/cm}^2$$

/ 4 pt.

Q21: WINDOW UTILIZATION FACTOR

Considering the core and wires you selected, calculate the current window utilization factor $K_{u,real}$ to verify whether the core window area is overly packed before proceeding to the actual realization of the transformer. Compare it with your initial estimate in Q4.

The real window utilization factor is calculated by summing the total bare copper area of all turns and dividing by the core's available window area ($W_A = 0.34 \text{ cm}^2$):

$$A_{copper,total} = (N_{pri} \times A_{pri}) + (N_{sec} \times A_{sec}) + (N_{aux} \times A_{aux})$$

$$A_{copper,total} = (21 \times 0.0033) + (10 \times 0.0078) + (8 \times 0.0033) = 0.17$$

$$K_{u,real} = \frac{A_{copper,total}}{W_A} = \frac{0.17}{0.34} \approx 0.5$$

Our calculated $K_{u,real} = 0.5$ is lower than our initial estimate of $K_u = 0.6$ from Q4. This is a good result: it guarantees that all of our selected discrete wires will fit into the E20/10/6 coil former while leaving enough margin for practical imperfect windings (gaps between wires).

$$K_{u,real} = 0.5$$

/ 2 pt.

TRANSFORMER – EVALUATION OF LOSSES

Q22: DC WINDING RESISTANCE

Give a reasonable expected temperature T_w for your windings. Calculate the DC winding resistances R_{DC} of the designed windings using the MLT (extracted from the datasheet), A_w (from selected wires, using the total conductive area), the resistivity of copper ($\rho = 1.724 \cdot 10^{-6} \Omega \text{ cm}$), and the temperature coefficient for copper $\alpha_T = 0.003862 \text{ K}^{-1}$.

A winding temperature of $T_w = 80^\circ \text{C}$ is chosen as a estimate transformer temperature. The DC resistance of each winding is (from chap 15.4 of the book):

$$R_{DC} = \frac{\rho_T \cdot N_i \cdot MLT}{A_w}$$

By adding a temperature dependence to the resistivity :

$$\rho_T = \rho_0 \cdot [1 + \alpha_T (T_w - 20)]$$

we finally get :

$$R_{DC} = \frac{\rho_0 \cdot N \cdot MLT}{A_w} \cdot [1 + \alpha_T (T_w - 20)]$$

with $\rho = 1.724 \times 10^{-6} \Omega \cdot \text{cm}$, $MLT = 4.2 \text{ cm}$, and temperature correction factor $1 + \alpha_T (T_w - 20) = 1 + 0.003862 \times (80 - 20) = 1.232$.

Primary (wire M1, $A_w = 3.32 \times 10^{-3} \text{ cm}^2$, $N_{pri} = 21$):

$$R_{DC,pri} = \frac{1.724 \times 10^{-6} \times 21 \times 4.2}{3.32 \times 10^{-3}} \times 1.232 = 0.056 \Omega$$

Secondary (wire L1, $A_w = 7.8 \times 10^{-3} \text{ cm}^2$, $N_{sec} = 10$):

$$R_{DC,sec} = \frac{1.724 \times 10^{-6} \times 10 \times 4.2}{7.8 \times 10^{-3}} \times 1.232 = 0.011 \Omega$$

Auxiliary (wire M1, $A_w = 3.32 \times 10^{-3} \text{ cm}^2$, $N_{aux} = 8$):

$$R_{DC,aux} = \frac{1.724 \times 10^{-6} \times 8 \times 4.2}{3.32 \times 10^{-3}} \times 1.232 = 0.021 \Omega$$

$$T_w = 80^\circ \text{C}$$

$$R_{DC,pri} = 0.056 \Omega$$

$$R_{DC,sec} = 0.011 \Omega$$

$$R_{DC,aux} = 0.021 \Omega$$

/ 3 pt.

Q23: AC WINDING RESISTANCE

Calculate the AC winding resistance R_{AC} considering only the skin effect (i.e., neglecting proximity effects and other phenomena). If your skin depth is lower than the diameter of your wire, the ratio between AC and DC resistance—known as F_R factor—can be approximated as

$$F_R \approx \frac{d_w}{4\delta} \quad (2)$$

(this formula can be found in the book "High frequency magnetic components" by M. K. Kazimierczuk, page 190 for the second edition). If this condition is not met, then you can reasonably assume a uniform current density (again, neglecting proximity effects). **Note:** in the case of a litz wire, you can consider the diameter of the strand.

The skin depth from Q14 is $\delta = 0.13 \text{ mm}$. We check whether each wire's diameter exceeds $2\delta = 0.26 \text{ mm}$:

Primary and auxiliary (M1, $d = 0.65 \text{ mm}$): Since $d = 0.65 \text{ mm} > 2\delta = 0.26 \text{ mm}$, the skin effect applies. Using the formula:

$$F_{R,M1} = \frac{d_w}{4\delta} = \frac{0.65}{4 \times 0.13} = 1.25$$

$$R_{AC,pri} = F_{R,M1} \times R_{DC,pri} = 1.25 \times 0.056 = 0.07 \Omega$$

$$R_{AC,aux} = F_{R,M_1} \times R_{DC,aux} = 1.25 \times 0.021 = 0.026 \Omega$$

Secondary (L1 Litz, strand $d = 0.20$ mm): Since the strand diameter $0.20 \text{ mm} < 2\delta = 0.26 \text{ mm}$, the current is uniformly distributed within each strand, so $F_{R,L_1} = 1$:

$$R_{AC,sec} = R_{DC,sec} = 0.011 \Omega$$

$$R_{AC,pri} = 0.07 \Omega$$

$$R_{AC,sec} = 0.011 \Omega$$

$$R_{AC,aux} = 0.026 \Omega$$

/ 4 pt.

Q24: DC AND AC COPPER LOSSES

Calculate the total DC copper losses $P_{Cu,DC}$ and AC copper losses $P_{Cu,AC}$ arising due to Joule effect in the windings. Perform the calculation considering the nominal power at both the minimum and maximum input voltage. Calculate the percentage of AC copper losses relative to the nominal output power. This gives you the expected efficiency drop due to the winding losses.

Copper losses are $P_{Cu} = \sum_j I_{j,RMS}^2 \cdot R_j$. The RMS currents at each operating point are taken from Report 1. The DC and AC resistances are taken from Q22 and Q23. The currents are taken from report 1.

- At $V_{in,min} = 30 \text{ V}$: $I_{pri} = 2.08 \text{ A}$, $I_{sec} = 4.33 \text{ A}$, $I_{aux} = 0.80 \text{ A}$.

$$P_{Cu,DC} = 2.08^2 \times 0.056 + 4.33^2 \times 0.011 + 0.80^2 \times 0.021 = 0.46 \text{ W}$$

$$P_{Cu,AC} = 2.08^2 \times 0.07 + 4.33^2 \times 0.011 + 0.80^2 \times 0.026 = 0.53 \text{ W}$$

$$\Rightarrow P_{Cu,AC}/P_{out} = 0.53/50 = 1.06\%$$

- At $V_{in,max} = 60 \text{ V}$: $I_{pri} = 1.17 \text{ A}$, $I_{sec} = 2.46 \text{ A}$, $I_{aux} = 0.80 \text{ A}$.

$$P_{Cu,DC} = 1.17^2 \times 0.056 + 2.46^2 \times 0.011 + 0.80^2 \times 0.021 = 0.16 \text{ W}$$

$$P_{Cu,AC} = 1.17^2 \times 0.07 + 2.46^2 \times 0.011 + 0.80^2 \times 0.026 = 0.18 \text{ W}$$

$$\Rightarrow P_{Cu,AC}/P_{out} = 0.18/50 = 0.36\%$$

The worst-case copper losses occur at minimum input voltage, where the larger RMS currents appears.

With minimum input voltage: $P_{Cu,DC} = 0.46 \text{ W}$ and $P_{Cu,AC} = 0.53 \text{ W}$, which corresponds to 1.06% of the output power.

With maximum input voltage: $P_{Cu,DC} = 0.16 \text{ W}$ and $P_{Cu,AC} = 0.18 \text{ W}$, which corresponds to 0.36% of the output power.

/ 3 pt.

Q25: ESTIMATION OF CORE LOSSES AND CORE TEMPERATURE

Using the plot of the loss density provided in the core material's data sheet (N87), estimate the core losses P_{core} of the designed transformer. The core volume can be found in the datasheet of the core shape you selected. Then, calculate the core losses as a percentage of the nominal output power. In addition, estimate the temperature rise of your transformer's core, considering an ambient temperature of 40°C . You can find values for the thermal resistance R_{th} and the temperature rise limit of different materials in the provided documents. Make sure that the temperature rise ΔT above ambient is reasonable and, more importantly, compatible with the material; if this is not the case, review your core and wire selection.

Using the Steinmetz-based core loss model from Q2 with $\Delta B = 0.09 \text{ T}$ and the plot of the power density in the datasheet, we can estimate P_V to be around $300 \frac{\text{kW}}{\text{m}^3} = 0.3 \frac{\text{W}}{\text{cm}^3}$ (in between 25°C and 100°C , but a little lower because we have only 90 mT)

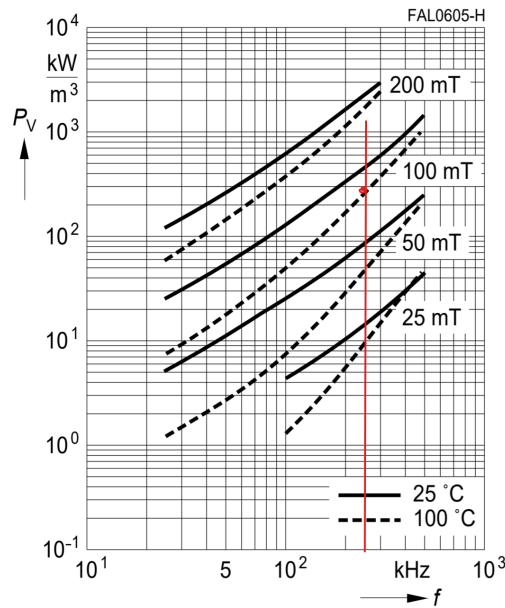


Figure 4 graph from the datasheet

The effective volume of the E20/10/6 core is $V_e = A_c \times l_m = 0.321 \times 4.63 = 1.49 \text{ cm}^3$, so:

$$P_{core} = P_V \times V_e = 0.3 \times 1.49 = 0.45 \text{ W}$$

$$\Rightarrow P_{core}/P_{out} = 0.45/50 = 0.9\%$$

From the provided thermal resistance document, $R_{th} = 46 \text{ K/W}$ for the chosen E20/10/6 core.

$$\Delta T = R_{th} \times P_{loss} = 46 \times 0.45 = 20.5 \text{ K}$$

The estimated core temperature is $T_{core} = 40 + 20.5 = 60.5^\circ\text{C}$, within the N87 operating range.

$P_{core} = 0.45 \text{ W}$ which corresponds to 0.9 % of the output power.

$\Delta T = 20.5 \text{ K}$

/ 3 pt.

Q26: OVERALL LOSSES

For the worst-case operating condition, calculate the total losses (copper and core) expected in your transformer and compare them with the loss budget you defined in Q5.

The worst case occurs at $V_{in,min} = 30 \text{ V}$, which produces the largest RMS currents and therefore the highest copper losses. Total transformer losses:

$$P_{loss} = P_{core} + P_{Cu,AC,worst} = 0.45 + 0.53 = 0.98 \text{ W}$$

Comparing with the loss budget from Q5: $P_{tot} = 5.56 \text{ W}$.

$$P_{loss} = 0.98 \text{ W} \ll P_{tot} = 5.56 \text{ W}$$

The design satisfies the loss constraints, with the actual losses representing only $0.98/5.56 = 17.6\%$ of the allowed budget. This confirms that the E20/10/6 core with the selected wires is a valid design.

$P_{loss} = 0.98 \text{ W}$

/ 3 pt.

TRANSFORMER – DESIGN SUMMARY

Q27: DAB TRANSFORMER DESIGN

Fill Table 6 with your design results. Do not forget to include the appropriate units.

Table 6 Design summary of the transformer

Electrical Specifications		
Primary current	$I_{w,pri,RMS}$	2.08 A
Secondary current	$I_{w,sec,RMS}$	4.33 A
Auxiliary current	$I_{w,aux,RMS}$	0.80 A
Magnetizing inductance	L_m	0.845 mH
Core specifications		
Core shape	-	E20/10/6
Coil former	-	B66206B1110T001
Core material	-	N87
Max flux density	ΔB	0.09 T
Winding Specifications		
Primary wire diameter	$\varnothing$	0.65 mm (M1)
Number of turns for primary	N_{pri}	21
Secondary wire diameter	$\varnothing$	1.53 mm (L1)
Number of turns for secondary wire	N_{sec}	10
Auxiliary wire diameter	$\varnothing$	0.65 mm (M1)
Number of turns for auxiliary wire	N_{aux}	8
Losses and temperature rise (worst-case)		
Core losses	P_{core}	0.45 W
Winding losses	$P_{Cu,AC}$	0.53 W
Total transformer losses	P_{loss}/P_{out}	0.98/50 = 1.96 %
Expected Temperature rise	ΔT	20.5K

/ 1 pt.

Q28: OVERALL EFFICIENCY OF THE CONVERTER

Taking into account the results of the Loss Tool of Report 1, estimate the best-case and worst-case efficiency of your DAB converter for the nominal output power considering the transformer losses estimated theoretically in the previous questions.

The total DAB efficiency is estimated by adding the transformer losses to the semiconductor losses from the Loss Tool of Report 1. We will compute the DAB efficiency for the three cases of 30V, 50V and 60V.

- $V_{in,min} = 30$ V:

$$\eta_{DAB,worst} = \frac{P_{out}}{P_{out} + P_{loss,trans,30V} + P_{semi,30V}} = \frac{50}{50 + 0.98 + 7.08} \times 100\% = 86.12\%$$

- $V_{in,max} = 60$ V:

$$\eta_{DAB,60V} = \frac{P_{out}}{P_{out} + P_{loss,trans,60V} + P_{semi,60V}} = \frac{50}{50 + 0.63 + 3} \times 100 = 93.23\%$$

- At $V_{in,nom} = 50$ V: We compute the transformer losses to assess the case of $V_{in,pri} = 50$ V since the transistor losses are minimal at this input value.

$$I_{pri} = 1.09 \text{ A}, I_{sec} = 2.26 \text{ A}, I_{aux} = 0.80 \text{ A}.$$

$$P_{Cu,DC} = 1.09^2 \times 0.056 + 2.26^2 \times 0.011 + 0.80^2 \times 0.021 = 0.136 \text{ W}$$

$$P_{Cu,AC} = 1.09^2 \times 0.07 + 2.26^2 \times 0.011 + 0.80^2 \times 0.026 = 0.156 \text{ W}$$

$$P_{loss,50V} = P_{Cu,AC} + P_{core} = 0.156 + 0.45 \approx 0.6 \text{ W}$$

$$\eta_{DAB,50V} = \frac{P_{out}}{P_{out} + P_{loss,trans,50V} + P_{semi,50V}} = \frac{50}{50 + 0.6 + 2.2} \times 100\% = 94.70\%$$

⇒ This case give the best overall efficiency.

$$\eta_{DAB,best} = 94.70 \%$$

$$\eta_{DAB,worst} = 86.12 \%$$

/ 3 pt.

TRANSFORMER – BUILDING PROCEDURE

Q29: WIRING OF THE TRANSFORMER

Based on the theoretical calculations done above, choose:

- 1x Coil former ! **Careful: Pins are easy to bend and break! The main structure is made of plastic, don't push clamps too hard!**
- 2x Core halves ! **Careful: Very fragile! Do not drop or scratch!**
- 2x Yokes

Then, follow these steps:

- Proceed with these items to the assembly station, and prepare the coil former.
- Carefully consider the following notes before proceeding:
 - Since an inductor will be added in series later on to obtain the desired inductance value for the DAB, the transformer should have the smallest leakage inductance, which is determined by the relative position (and geometry) of the windings. Closer windings will have lower leakage inductance.
 - Therefore, the windings can be interleaved to mitigate the proximity effects, i.e., reducing the winding losses. This is suggested to avoid significant overheating of the transformer (remember that proximity effects have been neglected in the previous calculations, so the temperature increase will be higher than predicted).
 - While the insulating varnish of the windings is sufficient for the operating voltage of the converter, you can consider adding some layers of kapton tape; this will ensure that the isolation is guaranteed even if you accidentally scratch the isolating varnish, but your transformer will run much hotter because the tape will hinder the heat exchange.
- Plan the length, the direction, and the layers of the windings before starting to wind. Consider that interleaving means wiring all three windings at the same time. Leave some margin at the end of each side of the windings to make the following steps easier. Mark the beginning and end of each of your windings to help you connecting them appropriately.
- Wind the appropriate number of turns for your primary and your two secondaries using the selected wire.
- Insert a single core half into the coil former, making sure not to scratch the insulating layer of the wires.
- Insert the second core half into the other side of the coil former, making sure not to scratch the insulating layer of the wires.
- Use the yokes to hold the transformer together. It might be required to use a bit of force; use the provided tools and be careful not to break the cores.

When the assembly is completed, add a picture of the fully assembled transformer in the answer box.

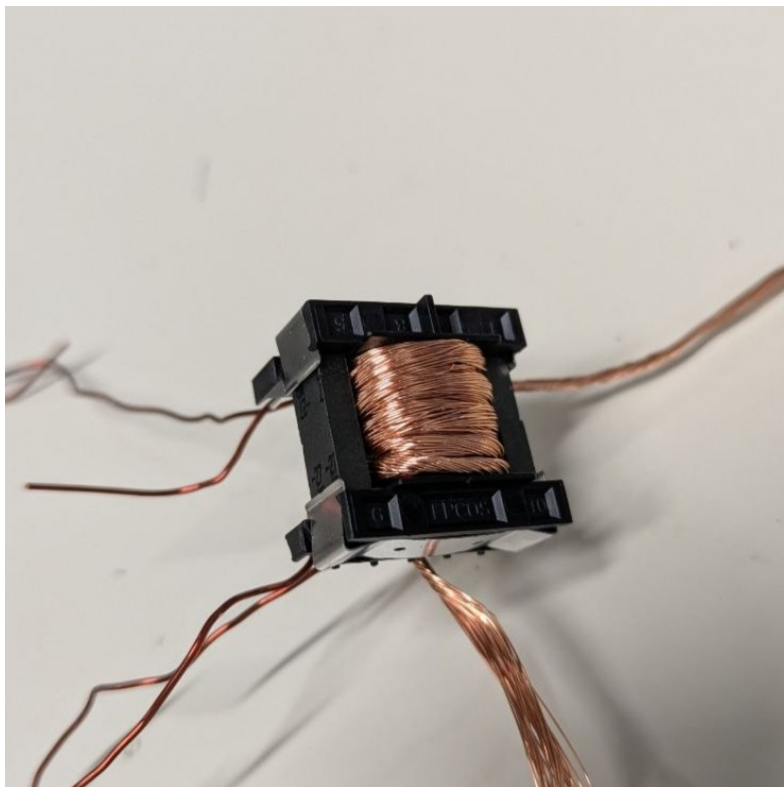


Figure 5 Close picture of the fully assembled transformer

Q30: VERIFYING THE WIRING DIRECTIONS

Using the LCR meter as a supply connected to the terminals of the primary winding, verify the winding direction of the secondary windings by following the steps below:

- Remove the insulation layer from the wire ends. If you are using bare wire, you need to scratch the varnish with sandpaper or scissors' blades. If you are using Litz wires, you need to terminate them using a soldering bath (keep the wire inside the bath for 5 to 20 seconds according to the size of your wire).
- Set the frequency on the LCR meter to 10 kHz;
- Mark the terminal of the primary winding connected to the high potential of the LCR;
- Measure the voltage across the secondary winding;
- If the voltage is in phase with the primary voltage, mark it as the positive terminal; otherwise, mark it as the negative terminal.
- Repeat the procedure in the previous two points for the auxiliary winding.

Provide a picture of the transformer with the markings on the windings.

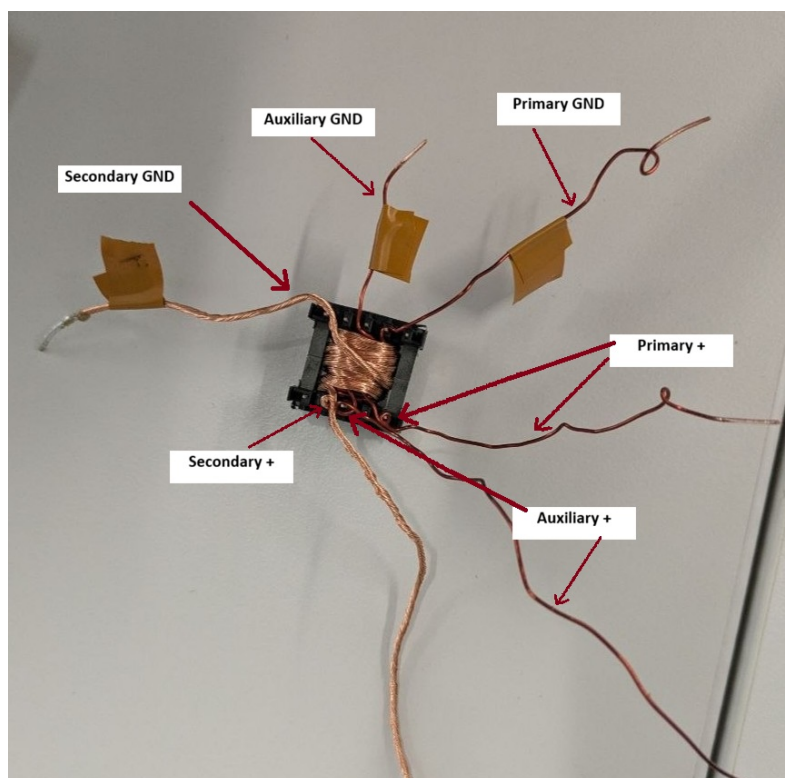


Figure 6 Transformer with winding markings

The primary side is positioned on the right side, the auxiliary side in the middle and the secondary on the left side of the transformer pins (See Fig. 6). The minus side is characterized by a piece of tape.

Q31: SOLDERING

Solder the windings to the respective pins of the coil former and provide a picture of the final soldered transformer.

Note: Make sure to solder the winding terminals to the intended pins (a suggestion is provided in Figure 7).

Note: You need to remove the insulation layer from the wire ends before soldering. If you are using bare wire, you need to scratch the varnish with sandpaper or scissors' blades. If you are using Litz wires, you need to terminate them using a soldering bath (keep the wire inside the bath for 5 to 20 seconds according to the size of your wire).

Be aware that the wire can get very hot. **Do not burn your fingers!!!**

Add a picture in the answer box after completing the soldering process.

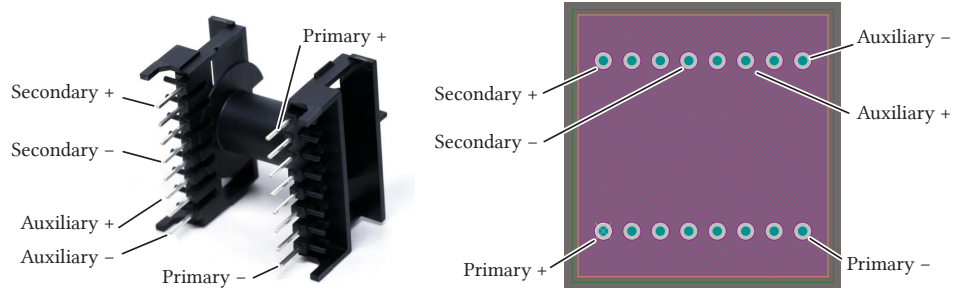


Figure 7 Suggested exemplary connection of the windings on the coil former (on the left) and the corresponding PCB footprint (on the right). This is only an example and, therefore, your coil former might be different.

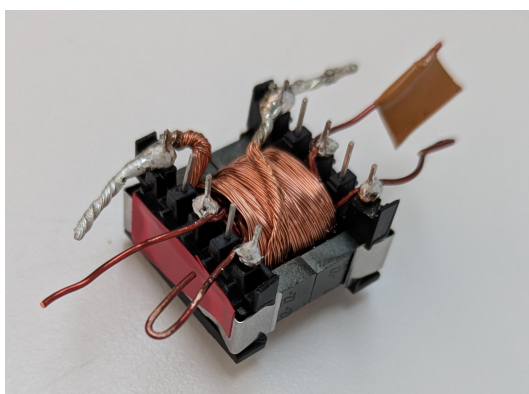


Figure 8 Soldered transformer

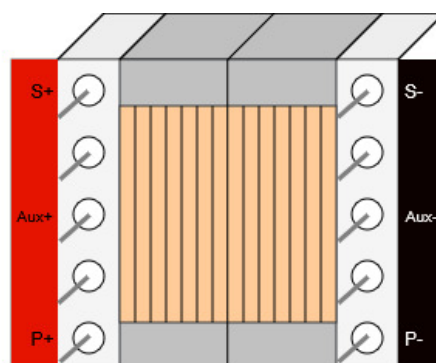


Figure 9 Schematics (same orientation as the left picture)

We left a bit of wire in case we need to do more testing in the future, it will ease of access. The positive side of the windings are marked with a red piece of tape (all on the same pin row: The red tape seen in the picture). The negative are marked by a black tape (not seen in the picture). The secondary windings are the Litz wire, the auxiliary are in the middle of the pin row and the primary are at the other end of the Litz wires.

TRANSFORMER – PROTOTYPE MEASUREMENT

Q32: OPEN-CIRCUIT TEST WITH LCR METER

To estimate the magnetizing inductance, perform an open-circuit test using the LCR Meter. Keep both secondary windings open and measure the inductance on the primary winding (using L_s mode). Show a picture of your measurement setup in the answer box and report the inductance value with the appropriate unit.

We performed the open circuit test with the LCR meter with three different frequencies:

- 10kHz: $L_m = 0.609$ mH
- 100 kHz: $L_m = 0.567$ mH
- 250 khz: $L_m = 0.58$ mH

Even though the three values are close, we choose the one measured at 10 kHz, as this is the standard LCR measurement frequency at which the instrument is most accurate and parasitic probe effects are negligible. Indeed, as L_m is a property of the core, it should be measured at a frequency at which the device is most precise and accurate. The measured value of 0.609 mH is consistent with the theoretical one of 0.845 mH (-28% difference).

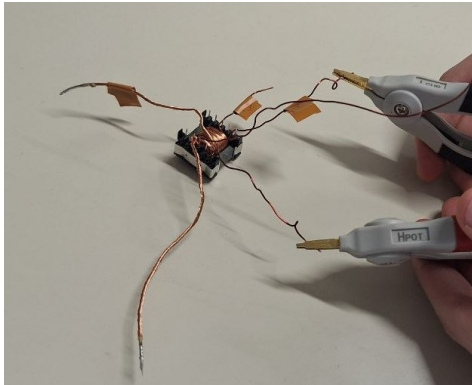


Figure 10 Setup for the open-circuit test with the LCR meter



Figure 11 Oscilloscope readings for the test

$L_m = 0.609$ mH at $f = 10$ kHz

/ 2 pt.

Q33: SHORT-CIRCUIT TEST WITH LCR METER

To estimate your leakage inductance and the total winding resistance, perform a short-circuit test using the LCR Meter. Short circuit both secondary windings using wires or clips, and measure the inductance and resistance (L_s - R_s mode) on the primary winding. Show a picture of your measurement setup in the answer box. Compare the inductance value with your desired DAB inductance to ensure that your leakage inductance is smaller. If the constraint is not fulfilled, rewind your transformer by keeping the windings closer to each other. Then, compare the measured resistance with the one you estimated in Q25. **Note:** The value you measure is the combined effect of your primary-side winding resistance and all the other winding resistances reflected at the primary side.

As in Q32, we performed the short circuit test with the LCR meter with three different frequencies:

- 10kHz: $L_{\text{leak,tot,pri}} = 3.06$ μ H and $R_{s,\text{tot,pri}} = 0.11$ Ω
- 100 kHz: $L_{\text{leak,tot,pri}} = 2.7$ μ H and $R_{s,\text{tot,pri}} = 0.35$ Ω
- 250 khz: $L_{\text{leak,tot,pri}} = 2.44$ μ H and $R_{s,\text{tot,pri}} = 0.75$ Ω

Since the operating frequency of our DAB converter is 250 kHz, we choose to report the measured values at that frequency. The leakage inductance of 2.44 μ H < 8.33 μ H satisfies the design constraint. The resistance of 0.75 Ω is significantly higher than the theoretical $R_{AC} = 0.07$ Ω (Q23), reflecting proximity effects that were intentionally neglected in the analytical model. This confirms that interleaving the windings during construction was important to limit these effects, something that we were not able to do due to the small size of the chosen core shape.

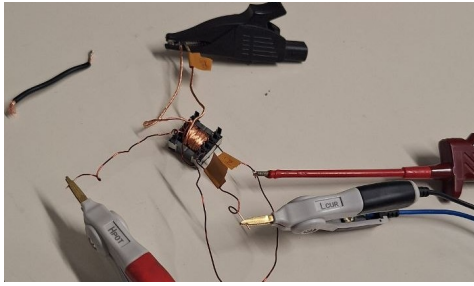


Figure 12 Setup for the short-circuit test with the LCR meter



Figure 13 Oscilloscope readings for the test

$$L_{\text{leak,tot,pri}} = 2.44 \mu\text{H}$$

$$R_{\text{s,tot,pri}} = 0.75 \Omega \text{ at } f = 250 \text{ kHz}$$

/ 3 pt.

Q34: OPEN-CIRCUIT TEST WITH A FREQUENCY RESPONSE ANALYZER

Validate the measurements in the previous open-circuit test by using the provided frequency response analyzer (Omicron Bode 100). Keep both secondary windings open and measure the inductance (Impedance Analysis/One-Port/Ls) on the primary winding. Provide a screenshot showing the measured inductance for a frequency range from 100 Hz to 1 MHz.

By looking at the value of the inductance measured in open circuit with the LCR meter (10kHz: $L_m = 0.609 \text{ mH}$, 100 kHz: $L_m = 0.567 \text{ mH}$ and 250 kHz: $L_m = 0.58 \text{ mH}$), we observe a similar trend : a slight drop around 150 kHz (maybe due to some resonance effects). However, we can see the presence of an offset of around $100 \mu\text{H}$. It is likely due to the calibration differences and different connection methods (coaxial impedance adapter vs Kelvin clips) between the two devices, leading to slightly different impedance at the measurement terminals.

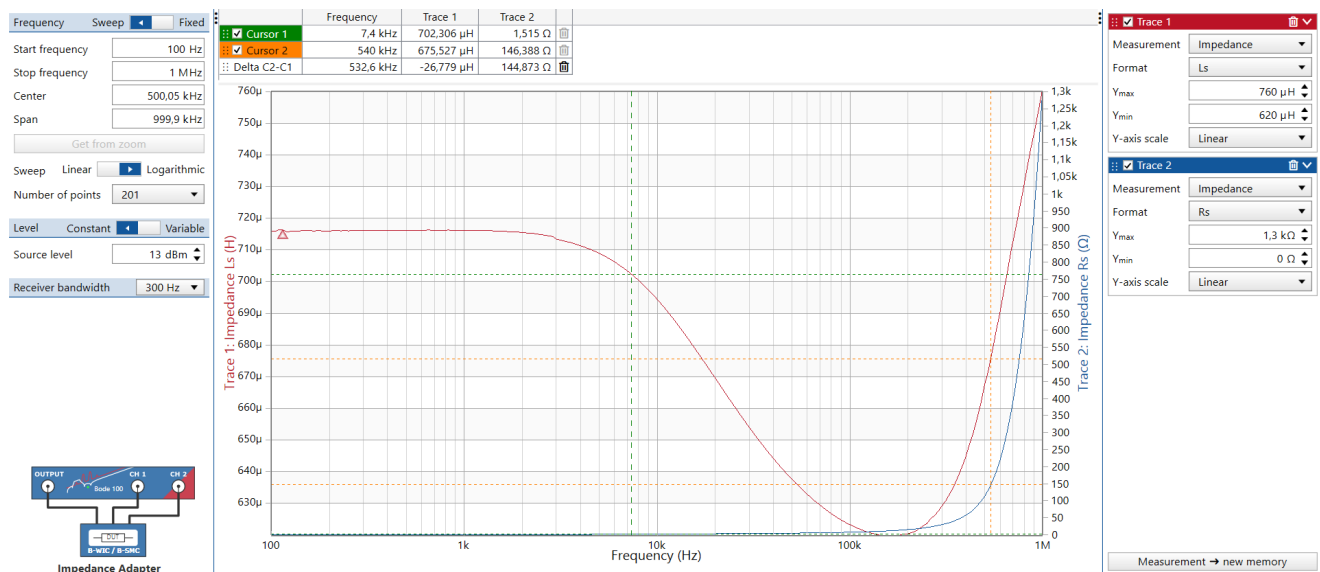


Figure 14 Open-Circuit test: Inductance and Resistance values for frequencies $\in [100 ; 10^6] \text{ Hz}$

/ 3 pt.

Q35: SHORT-CIRCUIT TEST WITH FREQUENCY RESPONSE ANALYZER

Validate the measurements in the previous short-circuit test by using the provided frequency response analyzer (Omicron Bode 100). Short Circuit both secondary windings and measure the inductance and resistance (Impedance Analysis/One-Port/Ls+Rs) on the primary winding. Provide a screenshot showing both the measured inductance and the measured resistance for a frequency range from 100 Hz to 1 MHz. Note the effect of frequency on your resistance.

The results of the short-circuit test with the response analyzer are shown in Fig 15.

Compared to the results gotten from the LCR meter (Q33), the inductance measurements (red curve) are matching the previous results (10kHz: $L_{\text{leak,tot,pri}} = 3.06 \mu\text{H}$, 100 kHz: $L_{\text{leak,tot,pri}} = 2.7 \mu\text{H}$ and 250 kHz: $L_{\text{leak,tot,pri}} = 2.44 \mu\text{H}$). We observe a slight decrease in inductance with increasing frequency.

The resistance R_s (blue curve, right axis) increases steadily with frequency, from approximately 123 mΩ at 7.4 kHz up to 1.47 Ω at 540 kHz, with a value of approximately 0.75 Ω at 250 kHz consistent with the LCR measurement in Q33. This increase is mainly caused by proximity effects. The theoretical $R_{AC} = 0.077 \Omega$ from Q23 accounts only for skin effect and is therefore roughly 10 times lower than measured, caused by not interleaving the windings.

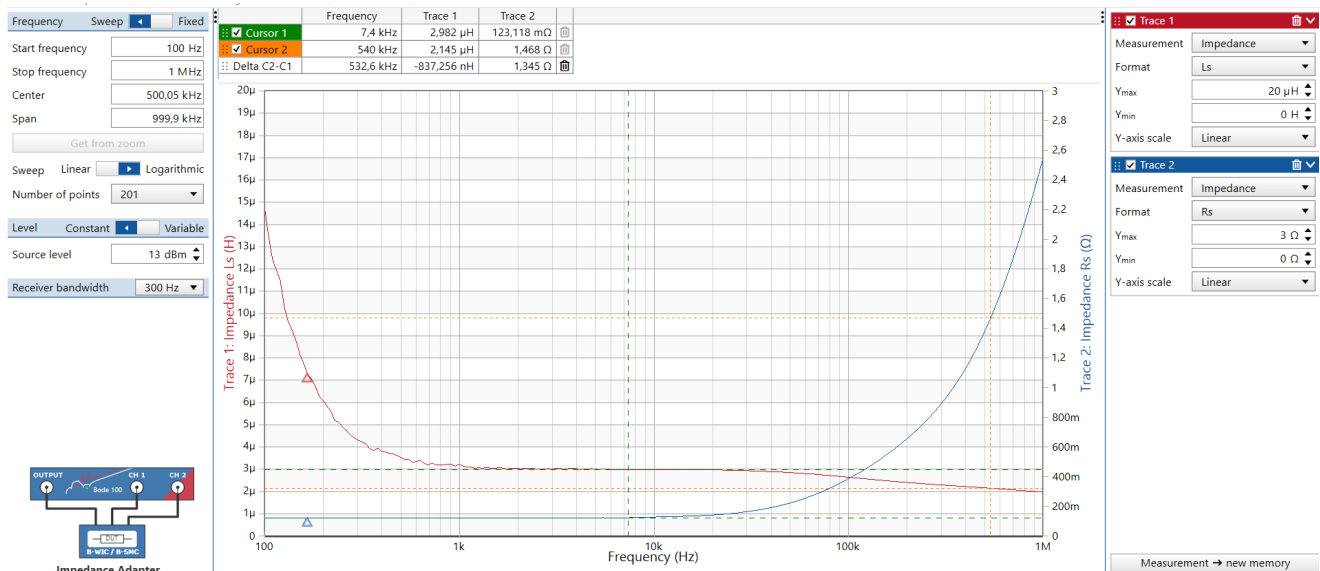


Figure 15 Short-Circuit test: Inductance and Resistance values for frequencies $\in [100 ; 10^6]$ Hz

/ 3 pt.

Q36: SATURATION MEASUREMENT

Using the power choke meter (DPG10-1000B), execute the saturation test and provide the results below. Follow these steps:

- Keep the secondary windings open.
- Connect both the Force and Sense cables to the power choke meter and to the terminals of the primary.
- Once connected, cover with the provided Plexiglas box.
- **DO NOT TOUCH IT until the end of the measuring process. You could get seriously injured!**
- In the control software:
 - Set the maximum voltage to 60 V;
 - Set the maximum pulse duration to 200 μs (increase it if necessary);
 - Set the maximum current to the maximum magnetizing current rounded **up** to the nearest integer.
- Start the measurement.

Verify that your core is not saturating, write the maximum value of the magnetizing current $I_{m,peak}$ below, and provide the curves of both the incremental and the secant inductance in the answer box, along with a picture of the measuring setup. In addition, provide a concise reason for why you believe the core is not saturating.

From Chapter 15 and the slides, we get the formula for the magnetizing current:

$$i_{M,pk} = \frac{\lambda_1}{2 \times L_m} = \frac{120 \times 10^{-6}}{2 \times 0.845 \times 10^{-3}} = 0.071 \text{ A}$$

For the saturation measurements, since we round up to the nearest integer, we choose to put $i_{M,pk} = 1 \text{ A}$.

For the test settings we also set:

- Maximum voltage to 60 V.

- Maximum pulse duration to 200 μs .
- Maximum current to $i_{M,pk} = 1 \text{ A}$

The results are shown in Fig. 17.

They show that at a value of 0.07 A of the magnetizing current ($\approx$ the computed value of 0.071 A), $L_{sec} = 693 \mu\text{H}$. The computed theoretical prediction was $L_m = 0.845 \text{ mH} = 845 \mu\text{H}$. This represents a difference of around 18 % which is totally normal and which can be explained by the fact that the datasheet value of $\mu_r = 2200$ is for a gapless toroidal core at 25°C and low field. However, the assembled E-core has two mechanical joints between the halves, each introducing a small effective air gap that reduces the net permeability thus leading to a slightly different result. Furthermore, $L_m = 690 \mu\text{H} \gg L_{m,min} = 83 \mu\text{H}$ (Q11) making this measured value valid.

Core Saturation: The core is not saturating around the operation point. Indeed, both curves show very small variations between $I = 0.03 \text{ A}$ and $I = 0.17 \text{ A}$, an interval containing the operating magnetizing current of 0.071 A.

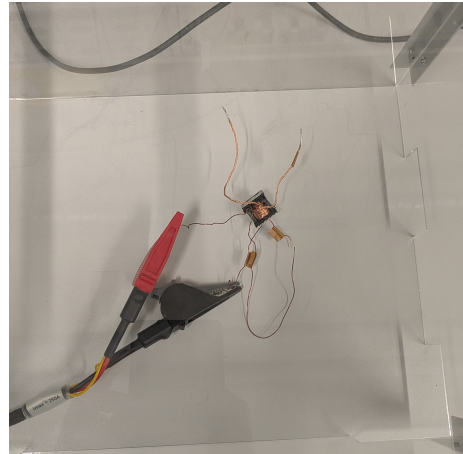


Figure 16 Setup of choke test

L_{inc} only begins dropping drastically around 0.2 A, indicating that saturation starts at roughly 2.8x above the operating point. The operating flux density $\Delta B = 0.09 \text{ T}$ is well below the N87 saturation flux density of 0.41 T at 80°C (0.49 T at 25°C and 0.38 T at 100°C). The core is not saturating.

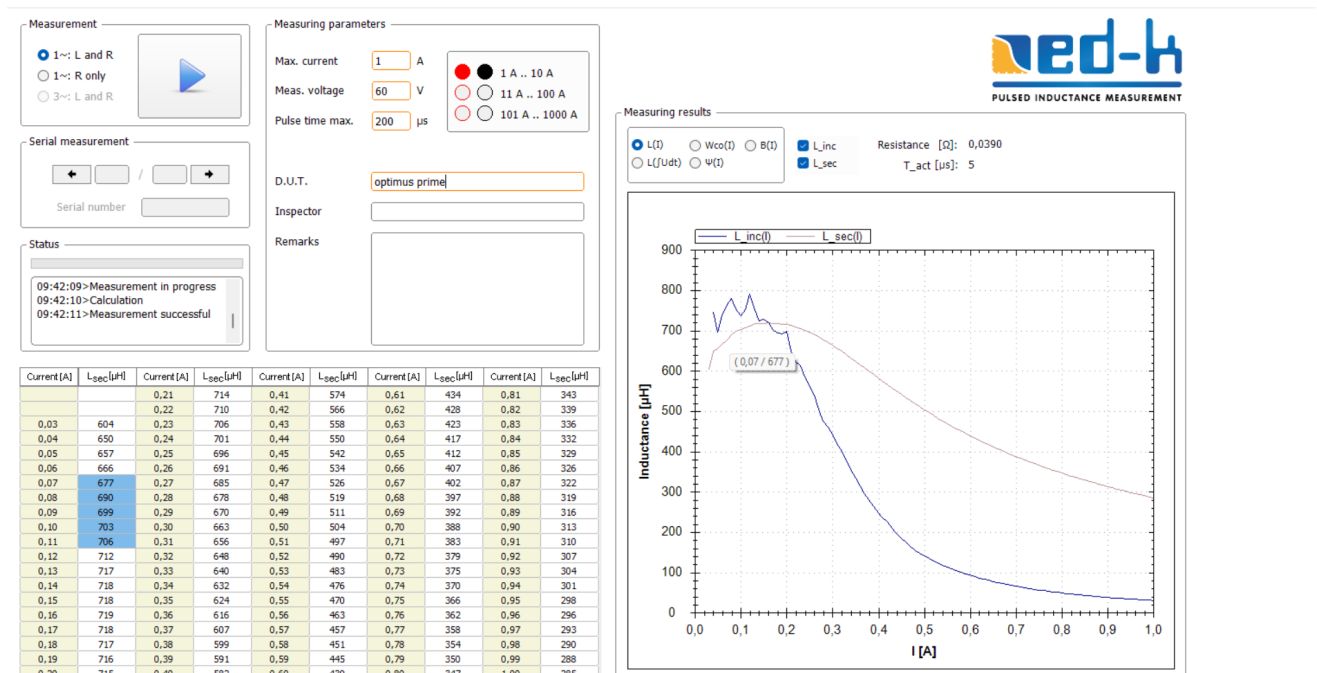


Figure 17 Incremental and the secant inductances values as a function of the current.

$$I_{m,peak} = 0.071 \text{ A}$$

/ 4 pt.

Q37: MAGNETIZING CURRENT AND VOLTAGE RATIO CHECK

Using the provided hardware setup and assisted by a TA, you will excite your transformer at the operating frequency and maximum output voltage. Report the oscilloscope acquisition of primary voltage, secondary voltages, and magnetizing current in the box below, along with a photo of the test setup. Check that the voltage ratios from the primary to the secondaries are correct. **Note:** keep the test as short as possible.

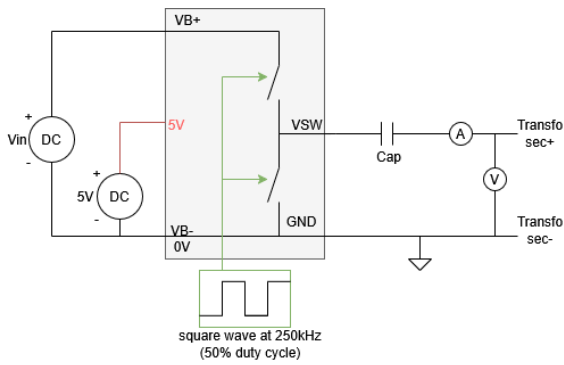


Figure 18 Setup diagram

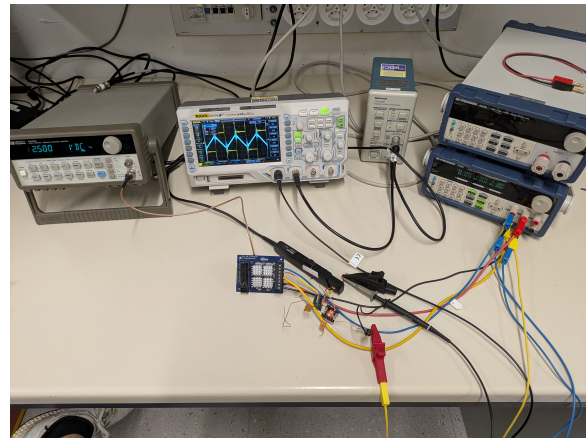


Figure 19 Picture of the setup



Figure 20 Primary voltage with 24V in the secondary



Figure 21 Secondary voltage with 24V in the secondary

From those results, we can estimate again the voltage ratio. In the secondary, we observe a voltage of around $5 \times 4.5 = 22.5$ V between each phases. In the primary, this voltage is around $10 \times 4.8 = 48$ V. This give us a ratio of $n = \frac{48}{22.5} = 2.13$, which is indeed quite close (considering the measuring uncertainty) to the researched value of 2.083.

By looking at the current, we can see that the magnetizing current we calculated before ($I_{m,peak} = 0.071$ A) is indeed really close to the measured values of around $I_{per,div} \times \frac{N_{div,peak,to,peak}}{2} = 50 \text{ mA} \times \frac{2.8}{2} = 70 \text{ mA}$ (the $\times 2$ compensation factor comes from the fact that the input dc voltage was two times lower than the nominal value (because of the half bridge), and the $/2$ is there to compensate the fact that we want the peak value instead of the peak to peak current).

Another thing we can calculate with this measurement is the value of the magnetizing inductance.

We can use the relation $u = L \frac{di}{dt}$. First we estimate $\frac{di}{dt} = \frac{50 \text{ mA} \times 2.8}{4 \mu \text{ s}} = 35 \text{ kA/s}$. Then $L = \frac{u}{\frac{di}{dt}} = \frac{22.5}{35} = 0.642 \text{ mH}$, which is relatively close to the calculated $L_m = 0.845 \text{ mH}$

/ 4 pt.

Q38: CORE HEAT-RUN TEST

You are tasked with performing a heat-run test using the setup used for the preceding question. The objective is to validate your design with respect to core losses. Follow these steps:

- Wait for the core to return to room temperature from your last test.
- Take a thermal photo of the core at room temperature.
- Start the excitation of the core and closely monitor the temperature closely for at least 10 minutes. Stop the test immediately if the temperature is too high and fix your design.
- After thermal steady-state is reached, take another thermal photo of the core.

Report the thermal images of the test in the box below.

The core heat-run test was conducted. The result is shown in the figures below. The initial core temperature was $T_0 = 25.2^\circ \text{C}$ and after 10 minutes of excitation it reached $T_{10} = 32.3^\circ \text{C}$, giving $\Delta T = 7.1 \text{ K}$, which is below the predicted 20.5 K from Q25. This makes sense because the steady-state might not be fully reached after 10 minutes of test and that the datasheet graph

reading likely overestimates P_{core} slightly.

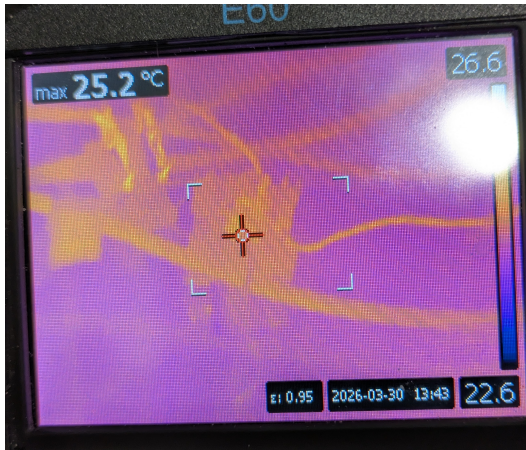


Figure 22 Thermal image before the test

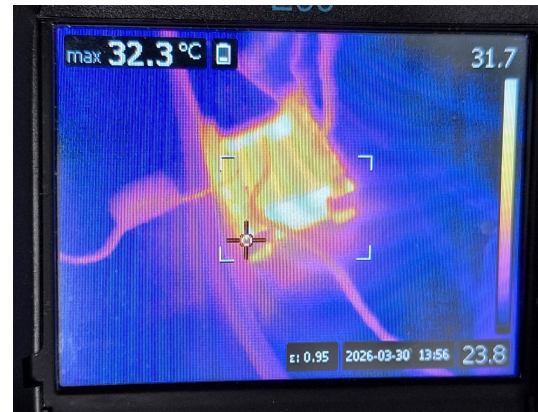


Figure 23 Thermal image after 10 minutes of test

/ 4 pt.

Q39: CONDUCTION HEAT-RUN TEST

You are tasked with executing a heat-run test following these steps:

- Take a DC power supply, connect it to your primary windings, and set it up in current mode with the current limit set to your expected worst-case primary RMS current increased by at least 10% (to account for part of the inevitable frequency effects).
- Repeat the same procedure for the secondary windings. **Make sure** that the cables you use for the connection can withstand your secondary current and are well connected to the power supply; otherwise, the cables could overheat and **burn** (use at least 1 mm^2 cross-section for 10 A; use at least 1.5 mm^2 cross-section for 15 A). If any current limit exceeds 9 A, contact a TA.
- Repeat the same procedure for the auxiliary windings. In this case, you can avoid increasing the RMS value since the provided one for the design is already very conservative.
- Take an initial thermal image of the transformer.
- Activate the power supplies and keep monitoring your transformer closely.

Provide a picture of the test-setup as well as two thermal images: one prior to the test and one after 15 minutes. Indicate the maximum temperature in the answer boxes below, together with the value of the currents in the test.

We conducted the test on all three windings. Every time, the tested winding was connected to the DC supply while the two others remained open. For each test, we let the core get back to room temperature then performed the test for 15 minutes. The current limit was set by taking the worst case RMS current increased by a 10% margin. This corresponds to :

- $I_{pri,test} = 2.08 \times 1.1 \approx 2.3 \text{ A}$
- $I_{sec,test} = 4.33 \times 1.1 \approx 4.76 \text{ A}$
- $I_{aux,test} = 0.8 \text{ A}$ (We were asked not to add margin)

The results are shown below in Fig. 25 to Fig. 28.

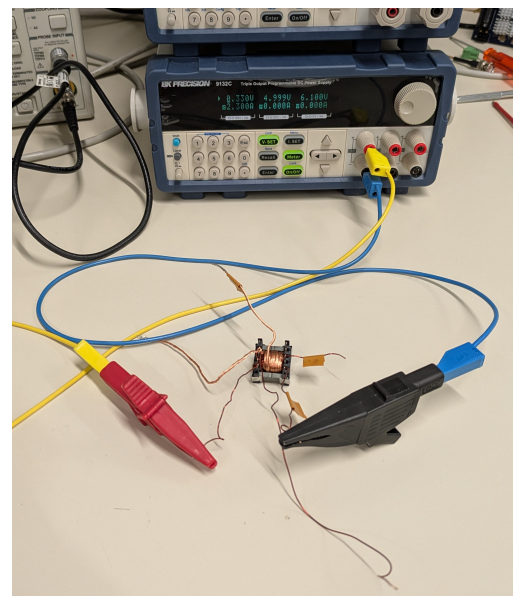


Figure 24 Test bench setup (for primary windings in this case)



Figure 25 Initial temperature

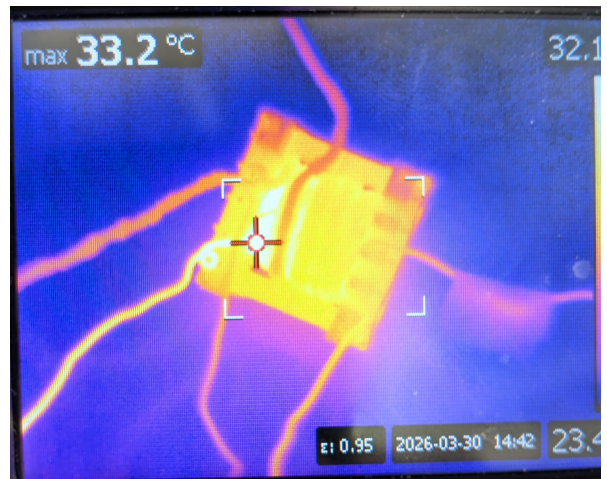


Figure 26 Temperature after 15min test on the primary windings

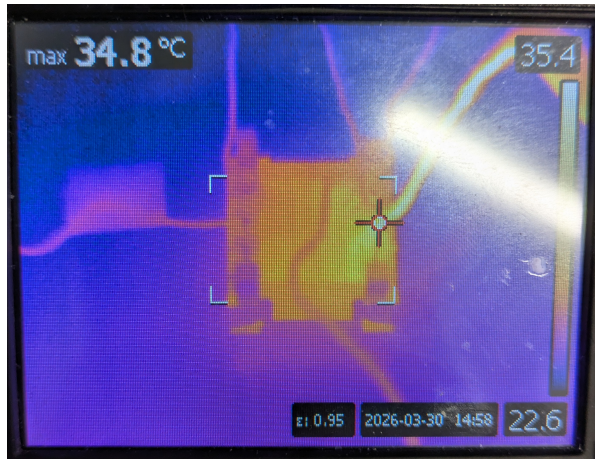


Figure 27 Temperature after 15min test on the secondary windings

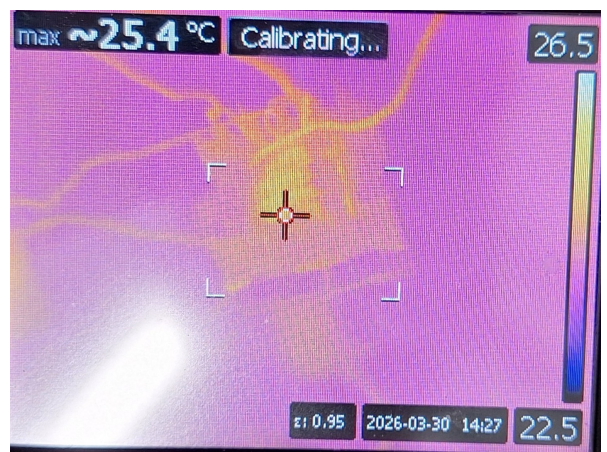


Figure 28 Temperature after 15min test on the auxiliary windings

$$T_0 = 24.5\text{ }^{\circ}\text{C}$$

$$T_{15\text{min}} = 34.8\text{ }^{\circ}\text{C}$$

$$I_{\text{pri,test}} = 2.3\text{ A (10 \% margin)}$$

$$I_{\text{sec,test}} = 4.76\text{ A (10 \% margin)}$$

$$I_{\text{aux,test}} = 0.8\text{ A (0 \% margin)}$$

$$\text{ / 3 pt.}$$

INDUCTOR DESIGN

The aim of this last section is to design and build the external inductor to integrate the obtained leakage inductance and reach the desired value for your DAB. This process is also iterative; be aware that you might need to reconsider your choices. The questions in this section represent a (mostly) independent iterative loop with respect to the one of the transformer (it is only influenced by the measured leakage inductance). To wind your inductor, choose the same wire you used for the primary side of your transformer. The core shape is a toroidal core. This shapes leads to simpler design at the cost of a higher volume. Be aware that it may not be possible to realize exactly the value of your choice and that **reasonable** deviations are acceptable.

Q40: EVALUATE THE EXTERNAL INDUCTANCE

Based on the measured leakage inductance (total, referred to the primary side), calculate the external inductance L_{ext} you need to add in order to reach the desired inductance value L .

$$L_{\text{ext}} = L_{\text{DAB}} - L_{\text{leak,tot,pri}} = 8.33 - 2.44 = 5.89 \mu\text{H}$$

$$L_{\text{ext}} = 5.89 \mu\text{H}$$

/ 2 pt.

Q41: DESIGN OF THE EXTERNAL INDUCTANCE

Choose one of the cores in the Table 7. Evaluate the integer number of turns needed to reach the value L_s as close as possible to the desired L_{ext} . **Note:** Indicate "2x" in your selected core if you need to use two cores in parallel.

The number of turns needed is given by: $N = \sqrt{L_{\text{ext}}/A_L}$ which gives us :

- **T94-2:** $N = 26.48 \Rightarrow N \in \{26; 27\}$ and $L_s \in \{5.68; 6.12\}$
- **2X T94-2:** $N = 18.73 \Rightarrow N \in \{18; 19\}$ and $L_s \in \{5.44; 6.06\}$
- **T130-2:** $N = 23.14 \Rightarrow N \in \{23; 24\}$ and $L_s \in \{5.82; 6.34\}$
- **T157-2:** $N = 20.51 \Rightarrow N \in \{20; 21\}$ and $L_s \in \{5.60; 6.17\}$
- **T184-2:** $N = 15.67 \Rightarrow N \in \{15; 16\}$ and $L_s \in \{5.40; 6.14\}$

If we take:

- **Core A** ($N = 26$): $L_s = 26^2 \times 8.4 \approx 5.68 \mu\text{H}$, which is $0.21 \mu\text{H}$ below target, representing a 3.57% deviation, within acceptable tolerance. This core weights 11 grams.
- **Core B** ($N = 23$): $L_s = 23^2 \times 11 = 5.82 \mu\text{H}$, which is $0.07 \mu\text{H}$ below target, representing a 1.2% deviation, within acceptable tolerance. The B core weights 29 grams.

Core A leads to slightly larger deviation in the inductance value. However, in order to have the smallest and lightest design possible, we choose to use core A. The practical inductance value will be tuned, later, by changing slightly the number of windings and tuning the winding distribution across the core to get the desired value.

Selected core = T94-2

$N_{L,s} = 26$

$L_s = 5.68 \mu\text{H}$

/ 4 pt.

Table 7 Available toroids for realizing the external inductor.

Ref.	Part Number	Material	A_L (nH/N ²)	μ_r	H_{max} (A/m)	Datasheet
A	T94-2	2	8.4	10	7150	Link
B	T130-2	2	11 (corrected)	10	7150	Link
C	T157-2	2	14	10	7150	Link
D	T184-2	2	24	10	7150	Link

Q42: MAGNETIC FIELD AND FLUX DENSITY

Calculate the peak magnetic field $H_{L,peak}$ for the inductor. To avoid saturation, the magnetic field should be lower than the maximum one reported in Table 7.

Then, evaluate the magnetic flux value in your inductor considering your peak primary current and the number of turn you selected in Q43. To limit the losses, the flux should be lower than 30 mT.

If you cannot fulfill this requirement, try to use another available core. If even this last solution is not enough, you can parallel two cores. Be aware that you need back to the previous question and recalculate the number of turns.

The peak primary current is taken from Report 1 (Q5) at the worst-case operating point ($V_{in,min} = 30$ V): $I_{pri,peak} = 3.6$ A.

Core A (T94-2, $N = 26$):

From the T94-2 datasheet, the mean magnetic path length is $l_e = 5.97$ cm = 0.0597 m.

$$H_{L,peak} = \frac{N \cdot I_{pri,peak}}{l_e} = \frac{26 \times 3.6}{0.0597} = 1568 \text{ A/m}$$

Since $H_{L,peak} = 1568$ A/m $\ll H_{max} = 7150$ A/m, the saturation constraint is satisfied.

The peak flux density, using $\mu_r = 10$ for Material 2:

$$B_{L,peak} = \mu_0 \cdot \mu_r \cdot H_{L,peak} = 4\pi \times 10^{-7} \times 10 \times 1568 = 19.7 \text{ mT}$$

Since $B_{L,peak} = 19.7$ mT < 30 mT, the flux density constraint is also satisfied. No core change is required.

$$H_{L,peak} = 1568 \text{ A/m}$$

$$B_{L,peak} = 19.7 \text{ mT}$$

/ 3 pt.

Q43: WINDING AND MEASUREMENT OF THE INDUCTOR

Using the same wire you selected for the primary of the transformer, wind the calculated number of turns on the selected core. Terminate your wires using the same procedure as for your transformer. Then, using the LCR meter to evaluate L_s and R_s .

Note: The toroidal cores are also fragile. Handle them with caution!

Note: Distribute your turns uniformly around the core to obtain a result closer to the one calculated analytically. You can also adjust the position of your turns to tailor your inductor value to the desired one. Once you find the optimal position, fix the two winding ends to the core with a small amount of glue or tape (e.g., use a glue gun).

Show a picture of your inductor in the answer box.

We use the M1 wire to wind our inductor ($d = 0.65$ mm, $A_w = 3.32 \times 10^{-3}$ cm²), the same wire as the transformer primary. The theoretical resistance values are computed at $T_w = 80$ °C with $F_R = 1.25$ (skin effect, Q23).

From the T94-2 datasheet, $MLT \approx 4.57$ cm.

$$R_{DC} = \frac{\rho_0 \cdot N \cdot MLT}{A_w} \times 1.232 = \frac{1.724 \times 10^{-6} \times 26 \times 4.57}{3.32 \times 10^{-3}} \times 1.232 = 0.076 \Omega$$

$$R_{AC} = F_R \times R_{DC} = 1.25 \times 0.076 = 0.095 \Omega$$

Figure 29 shows a picture of the wound inductance.

The measurement is performed with the LCR meter in L_s - R_s mode at $f = 250$ kHz (operating frequency). We adjusted the turn distribution across the core to get an inductance value as close as possible to the theoretical value $L_{ext} = 5.89$ μ H (Q40).

Results are shown below, in Fig. 30. The measured resistance is $R_s = 0.084$, Ω and the measured inductance $L_s = 5.73$ μ H

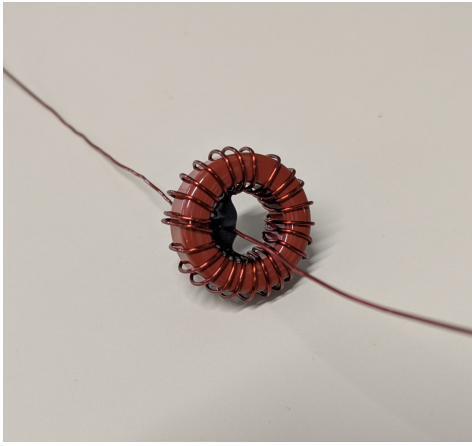


Figure 29 Final inductor

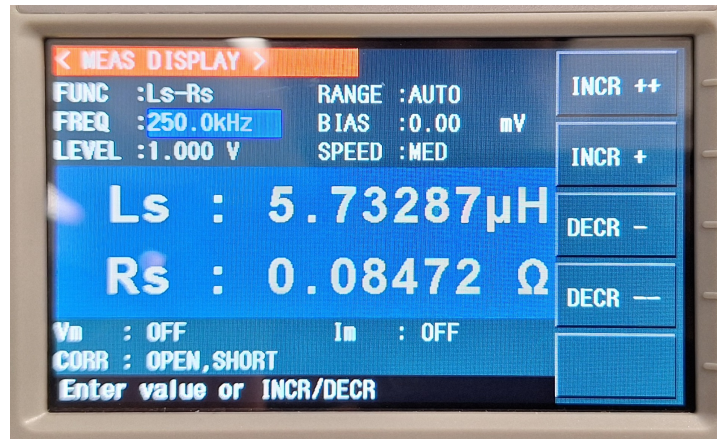


Figure 30 Inductor measurements

$L_s = 5.73 \mu\text{H}$ and $R_s = 0.084 \Omega$ at $f = 250 \text{ kHz}$

/ 5 pt.

Q44: CHECK COPPER LOSSES IN THE INDUCTOR

Taking into account the primary-side RMS current and the R_s you just measured, evaluate the copper losses expected in your inductor and calculate them as a percentage of the nominal output power. This is the estimated efficiency drop due to the losses in your inductor.

Note: It is suggested not to allow more than 1 W for the smaller core and 2 W for the larger cores to avoid overheating. If this happens, change the wire to reduce the resistance (you are eventually allowed to parallel two wires if needed).

The copper losses in the inductor are computed using the measured R_s at 250 kHz and the worst-case primary RMS current $I_{pri,RMS} = 2.08 \text{ A}$ (at $V_{in,min} = 30 \text{ V}$, from Report 1):

$$P_{Cu,L} = I_{pri,RMS}^2 \times R_s = 2.08^2 \times 0.084 = 0.36 \text{ W}$$

$$\frac{P_{Cu,L}}{P_{out}} = \frac{0.36}{50} = 0.72 \%$$

The 1 W limit for smaller cores is satisfied ($0.36 \text{ W} < 1 \text{ W}$). No wire change is required.

$P_{Cu,L} = 0.36 \text{ W}$ which corresponds to 0.72% of output power

/ 2 pt.

Q45: SATURATION MEASUREMENT

Repeat the saturation test that you previously performed on the transformer. Using the power choke meter, provide the results following these steps:

- Keep the secondary windings open.
- Connect both the Force and Sense cables to the power choke meter and to the terminals of the primary.
- Once connected, cover with the provided Plexiglas box.
- **DO NOT TOUCH IT until the end of the measuring process. You could get seriously injured!**
- In the control software:
 - Set the maximum voltage to 60 V;
 - Set the maximum pulse duration to 200 μs (increase it later if necessary);
 - Set the maximum current to **at least twice** the maximum input current rounded up to the nearest integer.
- Start the measurement.

Verify that your core is not saturating, and provide the curves of both the incremental and the secant inductance in the answer box, along with a picture of the measuring setup. In addition, provide a concise reason for why you believe the core is not saturating.

The instrument settings follow from the worst-case peak primary current $I_{pri,peak} = 3.6\text{ A}$ (Report 1, Q5). The maximum current is set to at least $2 \times 3.6 = 7.2\text{ A}$, rounded up to 8 A . The theoretical saturation current margin is:

$$I_{sat} = \frac{H_{max} \times l_e}{N} = \frac{7150 \times 0.0597}{26} = 16.4\text{ A}$$

Measurements:

We tried performing the test with 60 V , 8 A and $200\text{ }\mu\text{s}$, but we got an error message (See Fig.31). Therefore, we reduced the voltage to 10 V (30 V still gave the same error) and did the test again. The results are shown in Fig 33.

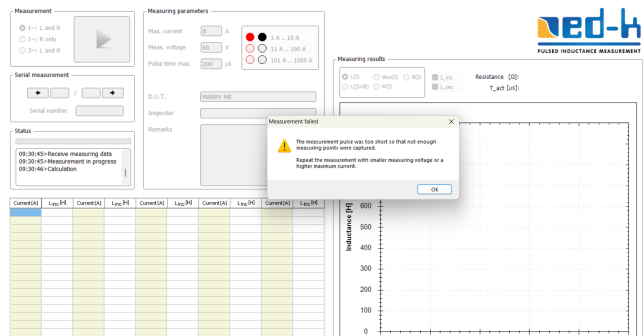


Figure 31 Inductor saturation test with 60 V (Error message)

We can see from Fig. 33 that the inductance value remains flat at approximately $5.77\text{ }\mu\text{H}$ across the full $0\text{--}8\text{ A}$ range, confirming that the core is not saturating. The measured winding resistance is approximately $R = 0.047\text{ }\Omega$, consistent with the theoretical DC value at room temperature. The operating point at 3.6 A represents only $3.6/16.4 \approx 22\%$ of the theoretical saturation current.

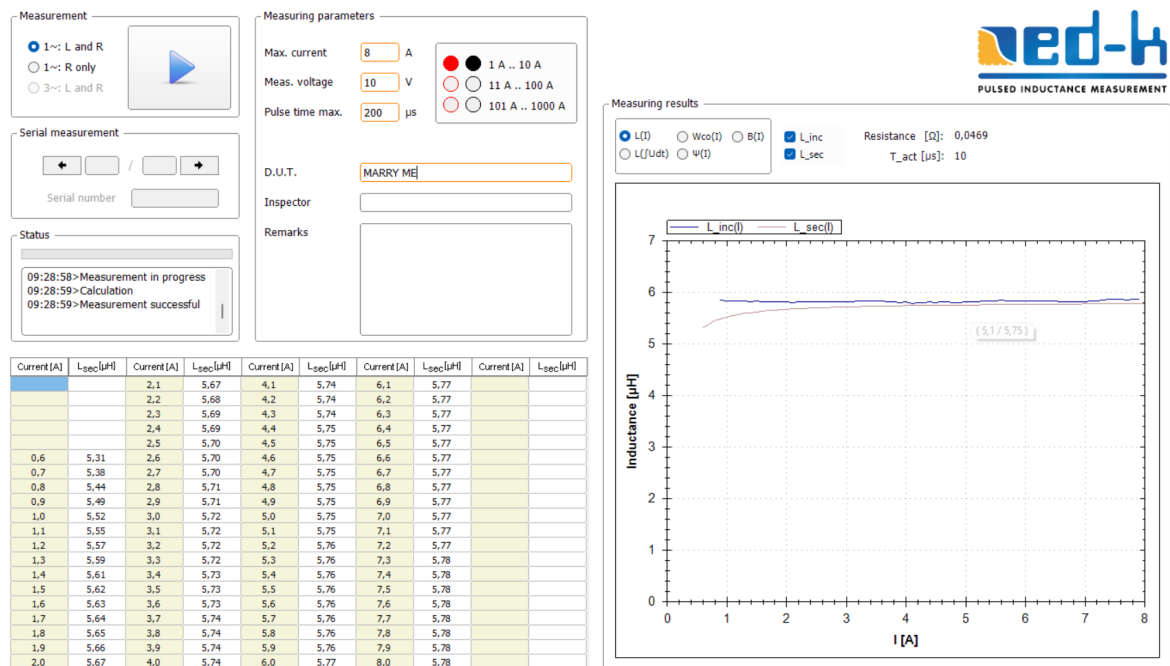


Figure 33 Inductor saturation test with 10 V

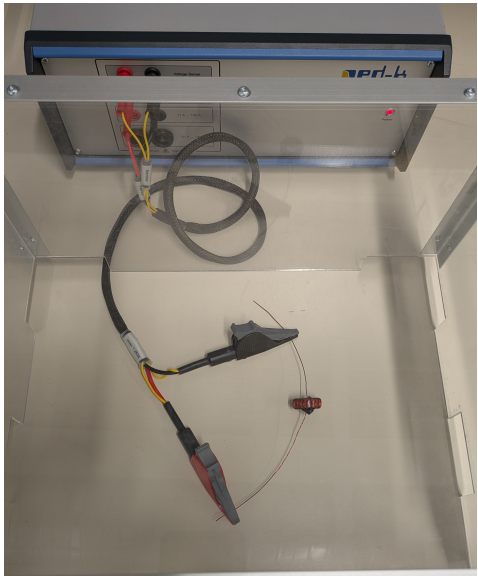


Figure 32 Picture of the test setup