Power Converters Design Report

“Advanced Electric Drives and Power Converter Systems” – Digital Automation Engineering

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1. Topology and input data

The objective of this project is to design a flyback converter (electrical insulated device), ensuring that it operates in Discontinuous Current Mode (DCM). Operating in DCM mode simplifies the control strategy, as the magnetizing inductance fully discharges before the next switching cycle begins. Additionally, in DCM, the diode naturally turns off when the inductor current reaches zero, avoiding the issues related to reverse recovery time, which can reduce switching losses and improve efficiency. The design process involves selecting key parameters such as magnetizing inductance, capacitance, primary-to-secondary turns ratio and appropriate magnetic components, in order to ensure the proper operation of the converter at the required power levels while maintaining DCM operation and a constant output voltage. Input specifications, including input voltage, output voltage, switching frequency and power requirements are listed below:

- Input voltage: $V_d = 15\ V$
- Output voltage: $V_o = 19\ V$
- Switching frequency: $f_s = 75\ kHz$
- Rated power: $P_{rat} = 50\ W$
- Maximum Duty cycle: $D_{max} = 0.45$

As additional specifications, the converter must ensure proper operation for a maximum output power $P_{max} = 2 *\ P_{rat}$ and maintain an output voltage ripple $\frac{\Delta V}{V_o} \leq 0.01$. A unit efficiency is assumed.

Once the design phase is completed, the converter will be simulated in *PLECS* to verify the correctness of the design and the validity of the assumptions. Additionally, the simulation will provide key parameters for the magnetic design (core selection, conductor sizing, and overall component dimensioning). The circuit topology equivalent to the flyback converter is shown in the figure below.

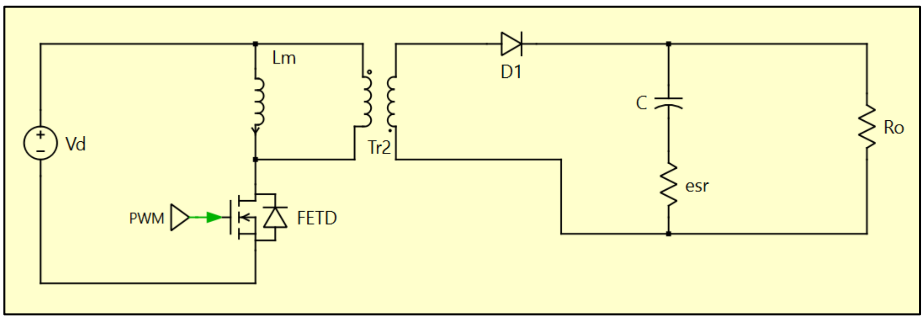


Figure 1. Flyback in PLECS - PWM Regulator to commutate a MOSFET with diode

2. Design phase: DCM operating mode

As a first step, it is necessary to calculate the values of the magnetizing inductance and the output capacitance. Assuming unit efficiency, the output currents respectively corresponding to rated and maximum output power can be calculated as:

$$I_{o_{rat}} = \frac{P_{rat}}{V_o} = \frac{50\ W}{19\ V} = 2.63\ A$$

$$I_{o_{max}} = \frac{P_{max}}{V_o} = \frac{100\ W}{19\ V} = 5.26\ A$$

The output capacitor was initially sized for the nominal operating point. However, its value was slightly increased to meet the required output voltage ripple specifications even at heavy load operating point, as the initial sizing based on nominal conditions did not fully satisfy the ripple requirements.

$$C = \frac{D_{max} * T_s}{R_o * \frac{\Delta V}{V_o}} = \frac{0.45 * (1.3 * 10^{-5}) s}{7.224\ \Omega * 0.01} = \ 1.662 * 10^{-4}\ F \rightarrow 2 * 10^{-4}\ F \qquad R_o = \frac{V_o}{I_{o_{rat}}}$$

To properly size the magnetizing inductance, it is essential to ensure that the converter operates in DCM even under the worst-case scenario. This condition occurs when the peak current is at its maximum, namely at maximum duty cycle and maximum output power.

$$I_{pk} = \frac{2 * \ P_{max}}{\eta * \ D_{max} * \ V_d} = \frac{2 * 100\ W}{1 * 0.45 * 15\ V} = 29.63\ A$$

Knowing peak current value and switching period it is possible to compute the critical (maximum) value of magnetic inductance above which DCM is not guaranteed anymore. Once the critical magnetizing inductance value is determined, a 20% safety margin can be applied to account for manufacturing tolerances. This results in a new, slightly lower inductance value.

$$L_{m_{crit}} = \frac{2 * P_{max} * T_s}{\eta * \ I_{pk}^{2}} = \frac{2 * 100\ W * (1.3 * 10^{-5})\ s}{1 * (29.63)^2 A} = \ 3.037\ \mu H$$

$$L_m = 0.8 * \ L_{m_{crit}} = \ 2.43\ \mu H$$

Consequently, it is possible to recalculate the new peak current, which, as expected, is higher than the initial estimation.

$$I_{pk_{target}} = \frac{V_d * \ D_{max} * \ T_s}{L_m} = \frac{15\ V * 0.45 * (1.3 * 10^{-5})\ s}{(2.43 * 10^{-6}) H} = 37.03\ A$$

As the final step in the electrical component design, the turns ratio of the transformer must be determined and it can be obtained through the following inequality:

$$\frac{N_1}{N_2} \geq \frac{V_d}{V_o} * \frac{D_{max}}{1 - D_{max}} = 0.64$$

The nearest ratio resulting from two integer numbers, greater than the one just obtained, is $n = \frac{2}{3}$.

3. Simulation and validation in PLECS

Once the values of magnetizing inductance, capacitance, and turns ratio have been determined, along with the peak current, the circuit can be simulated in *PLECS*. However, before proceeding with the simulation, it is necessary to calculate the duty cycle required to ensure DCM operation. Considering unit efficiency, the duty cycles required to respectively simulating rated and heavy load condition are computed as:

$$D = \frac{V_o}{V_d} * \sqrt{\frac{2 * L_m}{T_s * R_{rat}}} = 0.284 \qquad D' = \frac{V_o}{V_d} * \sqrt{\frac{2 * L_m}{T_s * R_{max}}} = 0.402$$

Both are lower than the maximum threshold D_{max} according to the fact that the converter has been designed to operate in discontinuous current mode.

3.1 Rated operating point

From the PLECS simulation at the nominal operating point, considering steady-state conditions, the following inductor current and transistor voltage waveforms were obtained. These results confirm operation in DCM, as the inductor current reaches zero for a non-negligible period before the end of the switching cycle.

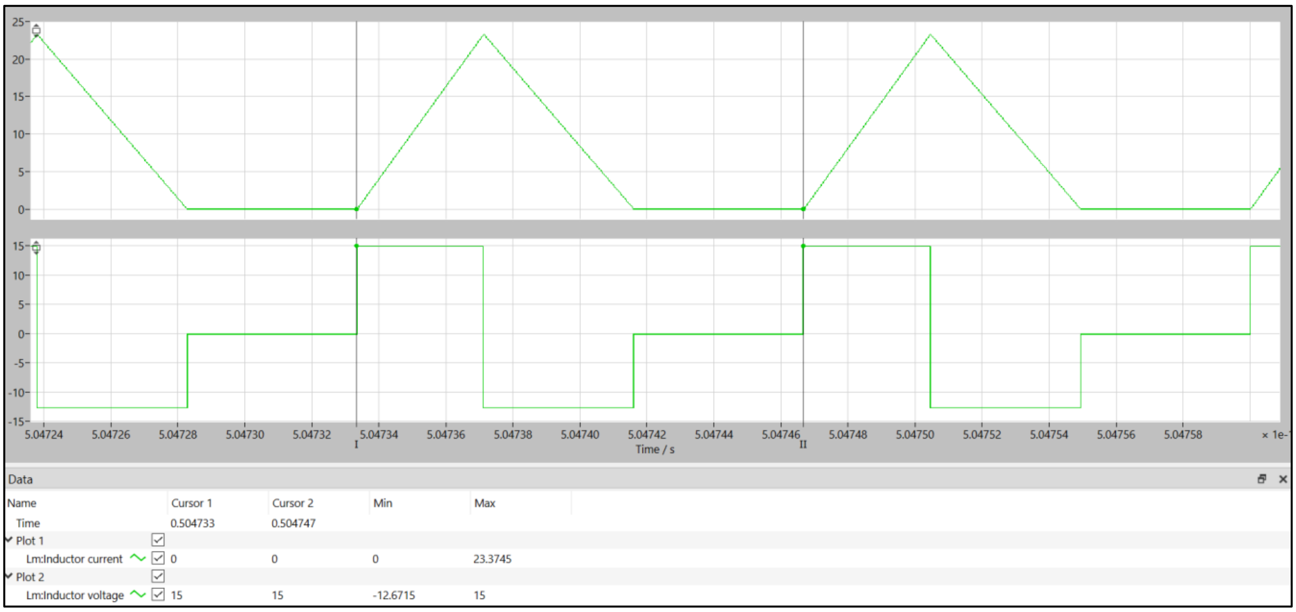


Figure 2. Inductor current and voltage

The peak current through the magnetizing inductance at the nominal operating point is 23.37 A. The voltage across the inductor is equal to the input voltage during the on-time, while during the off-time, it is equal to the reflected voltage from the secondary to the primary of the coupled inductor, with a negative magnitude.

At steady state condition, the required output voltage is met, and it remains nearly constant, with a ripple far below one per-cent.

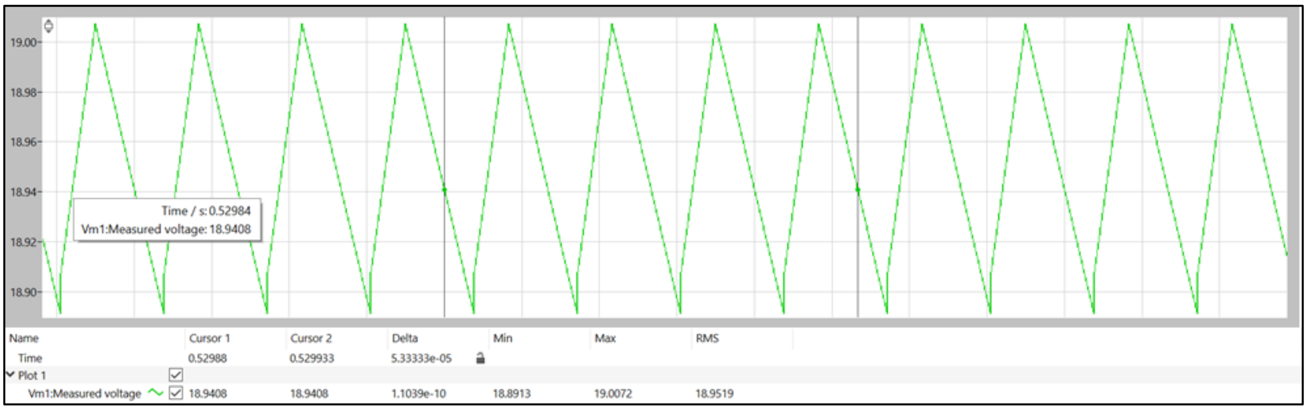


Figure 3. Output voltage at steady state

Another parameter of interest is the RMS current at the primary and secondary sides, which will later be used to determine the conductor area required. In the case of a flyback converter, these currents correspond to the current flowing through the MOSFET (primary side) and the diode (secondary side). Their respective waveforms are shown below:

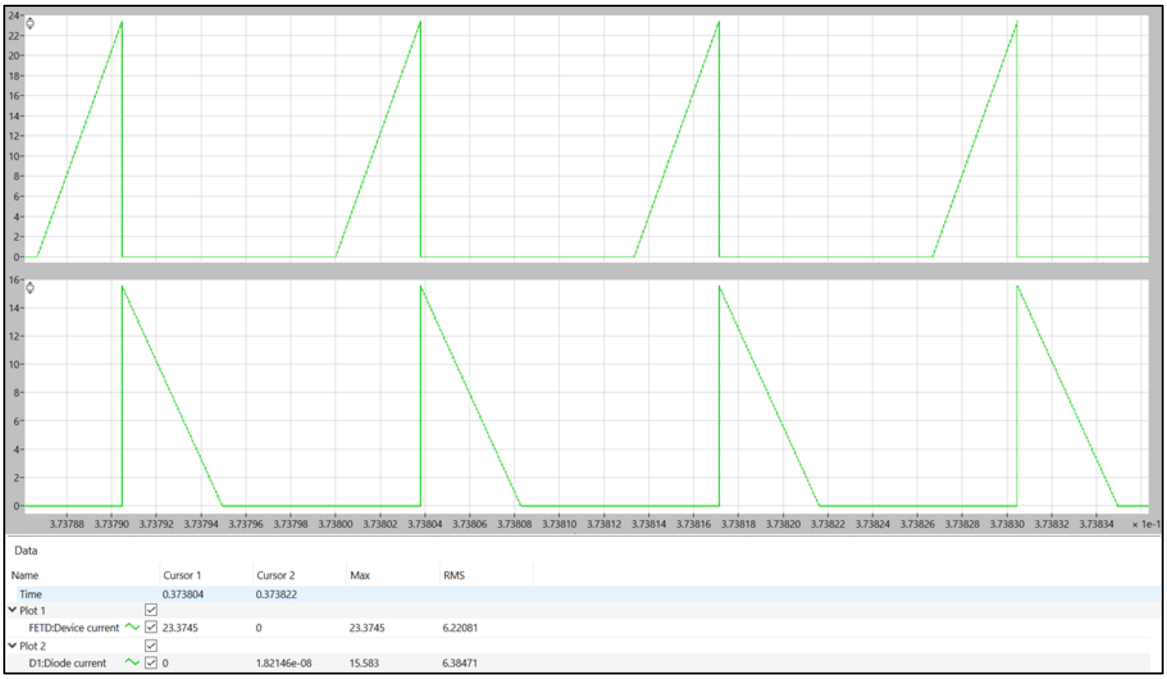


Figure 4. Currents on MOSFET and diode - Peak and RMS

As can be observed, the diode conducts during the off-time of the transistor, while during the on-time, the MOSFET is in a closed state, allowing energy to be stored in the inductor. The values of RMS current, respectively at primary and secondary are:

$$I_{RMS_1} = 6.22\ A \qquad\qquad I_{RMS_2} = 6.38$$

3.2 Maximum operating point

In the *PLECS* simulation under maximum load conditions, the inductor current waveform continues to confirm operation in Discontinuous Current Mode (DCM). The peak inductor current under these conditions is, as expected, higher than at nominal load.

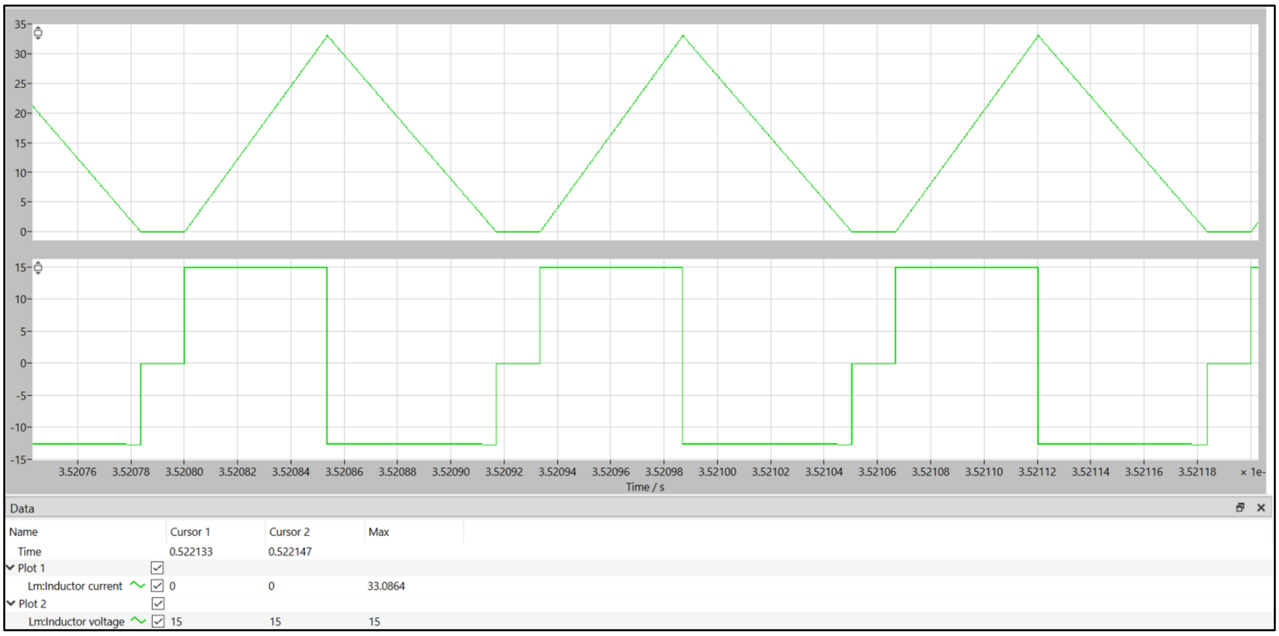


Figure 5. Current and voltage on the inductor at Pmax

Despite the increased load, the output voltage remains stable, and the ripple stays within acceptable limits, demonstrating the converter's effective performance even at heavy load operation. This may suggest that the capacitor size, previously increased in respect to the minimum value found, could be even slightly lowered.

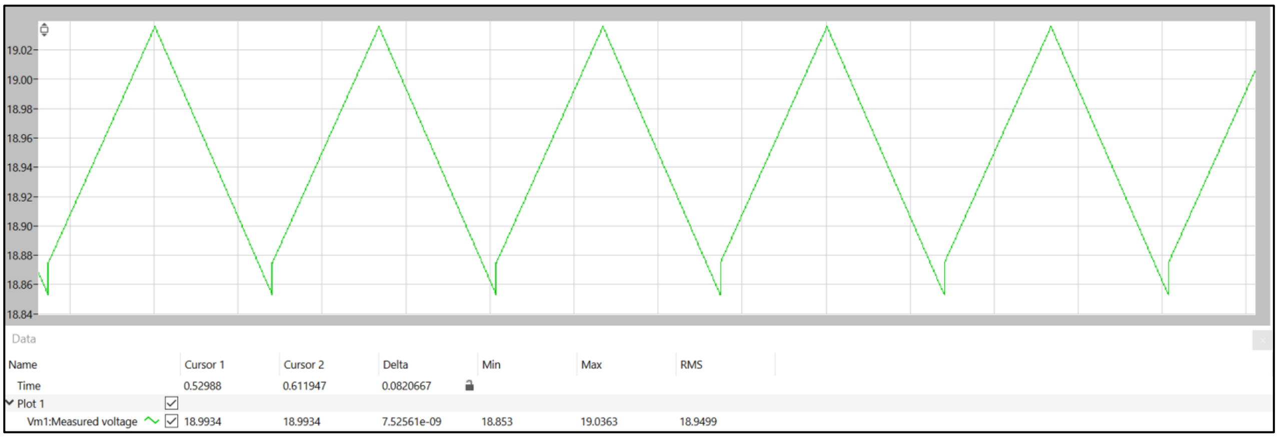


Figure 6. Output voltage and ripple at Pmax

The peak current on the magnetic inductance at maximum operating point will be matter of interest to properly addressing the magnetic design in order to avoid at all cost the saturation of the ferromagnetic material, even if most of the time the converter operates at rated power.

3.3 Electrical specifications for MOSFET and diode

It is necessary to define the electrical specifications for the transistor and diode. By analyzing the *PLECS* simulation under maximum load conditions, we can determine the voltage and current levels that these components must withstand. Therefore, it is essential to select a transistor and a diode capable of handling these values safely, even though the most of the time the converter operates at rated power. The respective waveform are shown:

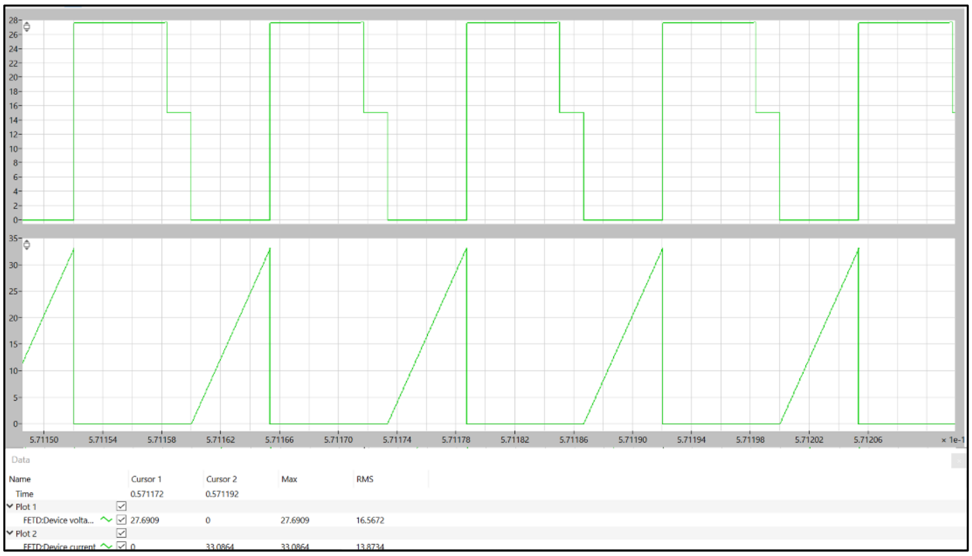


Figure 7. Withstand Voltage and current on the transistor at Pmax

The maximum $V_{DS} = 27.69\ V$ observed in the PLECS simulation represents the voltage the MOSFET must withstand when it is off. To ensure reliable operation, the selected MOSFET should have a V_{DS} rating at least 20% higher than the peak value observed in the simulation. The peak current is 22.05 A and the RMS is 8.78 A, a 20% safety margin will guarantee that the device will not break.

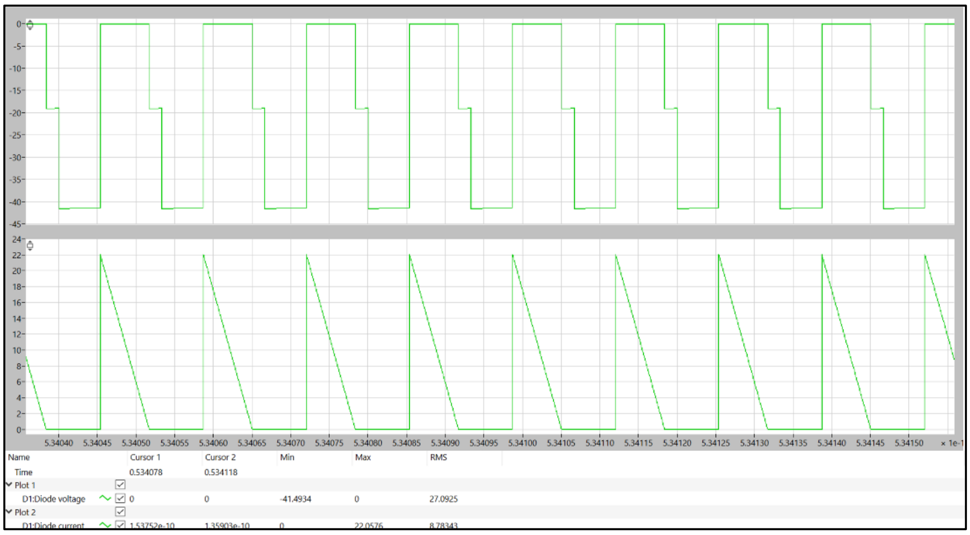


Figure 8. Withstand Voltage and current on the diode at Pmax

The voltage stress on the diode is determined by the sum of the output voltage and the reflected voltage from the transformer's secondary to the primary side. This value is $V_{diode} = 41.49\ V$. The current peak value is $I_{pk} = 22.05\ A$ with $I_{rms} = 8.78\ A$. A safety margin of 20% for what it concerns both current and voltage stress must be taken in to account.

4. Magnetic design: Core and wires selection

After obtaining the RMS currents for the primary and secondary windings, it is possible to proceed with the selection of the magnetic core and conductors. The first step is to determine the number of turns for both the primary and secondary windings for a certain core. To evaluate the number of turns at primary, the peak current on the magnetic inductance at maximum operating point, resulting from PLECS simulation, was taken into account.

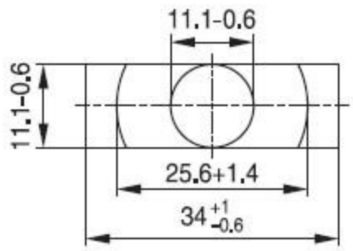
$$N_1 = \frac{L_m * I_{pk_{max}}}{B_{max} * A_{fe}} = 4\,, \quad N_2 = \frac{N_1}{n} = 6$$

Next, we calculate the theoretical required air gap based on the desired inductance. From this, we determine the mechanical gap and compare it with the specifications of the selected core.

$$l_o = \frac{N_1 * \mu_o * I_{pk_{max}}}{B_{max}}\,, \quad l_g = \frac{l_o}{2} \qquad\qquad \mu_o = 4\pi 10^{-7}$$

The core material must meet frequency and maximum value of inductance requirements to ensure efficient operation without saturation. After determining that the *ETD29* core was too small, the *ETD34* core was evaluated based on its specifications and found to be suitable for the design. The results of these computation and core's specifics of interest are shown in the tab below:

ETD-34	*Material N87*	
B_{max}	0.2	T
A_{fe}	97.1	mm^2
l_o	8.3	mm
l_g	4.15	mm
$\emptyset_{ext}$	25.6	mm
$\emptyset_{int}$	11.1	mm
h	11.8	mm



To evaluate whether the core saturates, it is possible to calculate the maximum magnetic flux using the following formula, and check if the corresponding flux inductance is above the desired threshold.

$$\Phi_{max} = \frac{L_m * I_{pk_{max}}}{N_1} \qquad\qquad B = \frac{\Phi_{max}}{A_{fe}} = 0.2 \leq 0.2\ \mathrm{T}$$

As previously said the peak current on primary at maximum operating point has been evaluated from *PLECS* simulation. The waveform and the peak value, obtained by mean of cursors, are shown below:

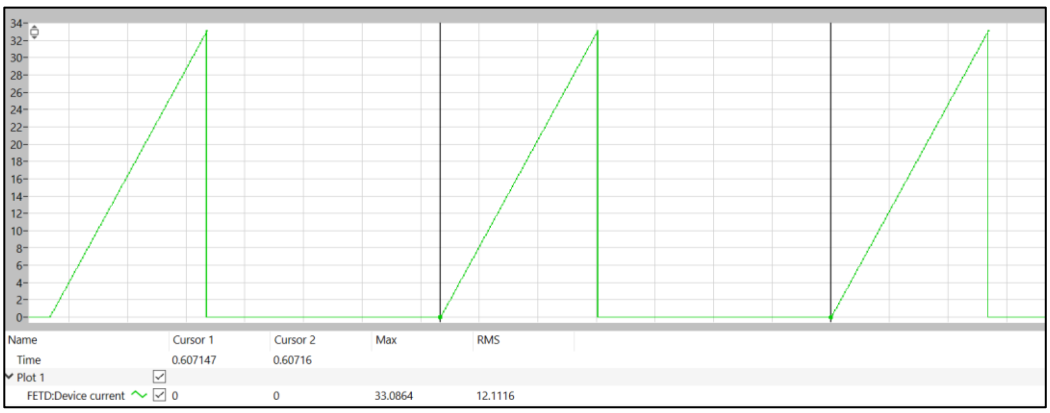


Figure 9. Peak current on primary winding (eg. MOSFET)

As the next step, an appropriate wire must be selected. The required conductor area for both the primary and secondary windings, based on the RMS currents obtained at nominal operating point, is given by:

$$A_{Cu_1} = \frac{I_{rms1}}{\mathcal{J}} = \frac{6.22\ A}{3\ A/mm^2} = 2.07\ mm^2$$

$$A_{Cu_2} = \frac{I_{rms2}}{\mathcal{J}} = \frac{6.38\ A}{3\ A/mm^2} = 2.13\ mm^2$$

where $\mathcal{J}$ is the assumed current density, set to 3 A/mm² for safety, considering air-cooling operation. Based on the *AWG* catalog, *AWG25* was selected:

AWG	Diameter [inches]	Diameter [mm]	Area [mm^2]	Resistance [Ohms / 1000 ft]	Resistance [Ohms / km]	Max Current [Amperes]	Max Frequency for 100% skin depth
25	0.0179	0.45466	0.162	32.37	106.1736	0.457	85 kHz

To compute the necessary number of wires at primary and secondary it is possible to compute the ratio between the area needed for the conductor wire and the area of the single wire:

$$n_{wires_1} = \frac{A_{Cu_1}}{A_{awg_{25}}} = 13 \qquad n_{wires_2} = \frac{A_{Cu_2}}{A_{awg_{25}}} = 14$$

Finally, it is necessary to verify that the selected wires fit within the available winding area. This is determined by applying a fill factor K_{fill} of 0.3 to the core winding window area.

$$A_w = \frac{\phi_{ext} - \phi_{int}}{2} * 2h = 171.1\ mm^2 \qquad A_{eff} = A_w * K_{fill} = 171.1 * 0.3 = 51.33\ mm^2$$

$$A_{Cu_{tot}} = N_1 * A_{Cu_1} + N_2 * A_{Cu_2} \leq\ A_{eff}$$

$$4 * 2.07 + 6 * 2.13 = 21.06\ mm^2 \leq 51.33\ mm^2$$

The total area needed for the windings for the chosen core is slightly below 50% of the total available area. In conclusion, both the electrical component design and the magnetic design appear to be feasible, and the converter operates in the intended mode as per our design objectives.