MIDDLE EAST TECHNICAL UNIVERSITY

ELECTRICAL & ELECTRONICS ENGINEERING

EE464 – STATIC POWER CONVERSION 2

TERM PROJECT – FINAL REPORT

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Introduction

The main purpose of this project is converting 12-18VDC range input voltage to 48VDC output by using isolated converter. The power requirement for this design is 48W at output. This design must be regulated, and line regulation as well as load regulation should not decrease below 3%. Considering all these requirements, the correct isolated converter topology selection process is discussed in the report. The transformer parameters of the isolated converter and magnetic design process are discussed. Last but not least, the experimental results are included in the report and the compliance with the requirements are discussed. The component selection for the project is included. Furthermore, the loss calculations are done for MOSFET, diode and transformer of the design. The PCB design and further design decisions due to not having the ideal components are discussed in detail. Finally, in the further discussion & analysis part, the causes of the differences between the simulations and the practical parameters are discussed.

1. Topology Discussion

In this project, an isolated DC-DC converter was designed with the specifications as can be seen from Table 1.

$V_{in,min}$	12V
$V_{in,max}$	18V
V_{out}	48V
P_{out}	48W
$V_{out,pp}$ Ripple	3%
Line Regulation	3%
Load Regulation	3%

Table 1. Specifications of the Project

There were 5 candidates for this isolated DC-DC converter as follows:

- Flyback Converter
- Forward Converter
- Push-Pull Converter
- Half-Bridge Converter
- Full-Bridge Converter

The advantages and disadvantages of each topology are considered when choosing the most ideal topology for this project. Flyback topology is decided after in-depth analysis of each topology. The potential of smaller size and easier control were the main reasons for the selection of the flyback converter. In the flyback converter, there are a smaller number of switches which means less switching losses and thus increase in efficiency, which is yet another reason for the selection of this topology. Although other topologies have advantages over the flyback, their disadvantages were considered greater than advantages, for example the control of switches is deemed problematic as turning on both switches at the same time for push-pull, half-bridge and full-bridge converters may blow up the converter. Last but not least, forward converter is avoided due to the duty cycle limitation to avoid the saturation of the core.

Topology Name	Advantages	Disadvantages
Flyback Converter	- Only one switch is needed, so it is easier to control. - Comparatively smaller in size because no additional inductor is used. - Only one switch, so smaller switching losses.	- The peak current and voltage through the switch are high. - High output current ripple.
Forward Converter	- Power is transferred immediately. - Low output current ripple. - If a third winding topology is used, the energy is transferred back to the supply. - Only one switch, so smaller switching losses.	- RCD snubber circuit or a third winding on the transformer is necessary. This causes heat or worse fill factor. - One inductor is used, slightly bigger size. - Maximum duty cycle is limited to $\frac{1}{(1+\frac{N_3}{N_1})}$, otherwise core will saturate.

Push-Pull Converter	- Better use of core, 2 quadrant operation (1^{st} and 3^{rd} Quadrants).	- Two switches need to work together, hard to control. - Center tapped transformer will mean worse fill factor in the transformer. - One inductor, two diodes, two switches cause bigger size and switching losses.
Half-Bridge Converter	- Small output current ripple.	- One inductor, two diodes, two switches cause bigger size. - Small variances in the capacitances in the primary may cause unbalanced voltage division among them.
Full-Bridge Converter	- Small output current ripple. - Higher gain with the same turns ratio and duty cycle.	- 4 switches working synchronously, worse switching losses.

Table 2. Advantages and disadvantages of each topology

The specifications in Table 1 suggests that the converter will need to take input between 12V and 18V. Considering the gain formula for the flyback converter is $\frac{V_O}{V_{in}} = \frac{D}{1-D} * \left(\frac{N_2}{N1}\right)$, the maximum duty cycle was selected as 0.5 and turns ratio N2:N1 is selected as 4:1 to obtain 48V output voltage while having 12V input voltage. When the input voltage becomes 18V, the duty cycle gets reduced to 0.4 and this is the minimum duty cycle value. Hence, the converter should be able to operate at a duty cycle range of 0.4 to 0.5.

2. Magnetic Design

2.1 Power Calculations

The design of the transformer is one of the most critical parts of flyback topology. The maximum duty cycle for this design is 0.5 and the flyback operates at CCM. To design transformer and choose core, the worst-case Lm can be calculated from the Equation M1 firstly [M1].

$$L_m = \frac{(V_{min}D_{\max})}{2P_{in}f_sK_{RF}} \tag{M1}$$

In the equation, Vmin is given as 12V and Dmax is specified before as 0.5. The desired efficiency is 85%. Therefore, from output power requirements, input power can be found as around 56.5W. The Lm can be found as 13.3uH from equation M1. In the next step to find turns ratio, peak and rms current value of primary side must be found. The peak current and rms can be found from the equation M2 and M3 respectively.

$$I_p = \frac{2P_{in_{max}}}{V_{min}D_{max}} \tag{M2}$$

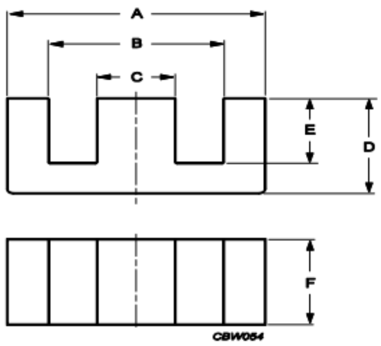
$$I_{p_{rms}} = I_P\sqrt{\frac{D_{max}}{3}} \tag{M3}$$

From equation M2 and M3 peak and rms current can be found as 18.8A and 7.68A respectively. After these parameters are calculated, the saturation of flux density must be specified. Since there is no limitation for this design, Bmax can be selected as 0.3T. With these parameters the core for transformer was selected.

2.2 Core Selection

There are two important parts of core selection. One of them is core type and another one is dimensions and magnetic parameters of the core. In this design, two types of cores were considered because of other cores' prices. These are powdered core and ferrite core. Even if powdered core store more energy than ferrite core, the ferrite core has higher magnetic permeability for high frequencies. Since the switching frequency of the design is 60kHz, ferrite will be a more reasonable option. This low permeability can cause more leakage inductance for powdered core. Because of these reasons, ferrite core was chosen as core type.

Two ferrite core option was simulated with all calculations. These are 00K3515E090 and E42/21/20-3C94. The effective area of 00K3515E090 is much smaller than E42/21/20-3C94. This is good for core loss and compactness. However, this cause more turns for winding and the transformer will be handmade to change easily for alternative solutions. Therefore, choosing larger effective areas can provide easier production process for transformer. Therefore, E42/21/20-3C94 was selected as core of the flyback transformer. The dimensions and mechanical drawing of the core can be seen in Figure 1.



Effective parameters			
	Parameter	**Value**	**Unit**
Σ(I/A)	*core factor (C1)*	0.417	mm⁻¹
Ve	*effective volume*	22700	mm³
Le	*effective length*	97	mm
Ae	*effective area*	233	mm²
Amin	*minimum area*	233	mm²
m	*E42/21/20*	≈ 56	g/pcs

Dimensions for product: E42/21/20						
	Nom	**Tol +**	**Tol -**	**Max**	**Min**	**Unit**
A	43.00	0.00	1.70	43.00	41.30	mm
B	29.50	1.40	0.00	30.90	29.50	mm
C	12.20	0.00	0.50	12.20	11.70	mm
D	21.00	0.20	0.20	21.20	20.80	mm
E	14.80	0.60	0.00	15.40	14.80	mm
F	20.00	0.00	0.80	20.00	19.20	mm

Figure 1 Dimensions and Parameters of E42/21/20

2.3 Transformer Characteristics and Calculations

Once the core has been determined, the first step will be to decide on the turns ratio of transformer. To decide on this, equation M4 can be used.

$$n = \frac{Vin_{min}D_{max}}{(Vout+V_D)(1-D_{max})} \quad \text{(M4)}$$

From M4, turns ratio was found as 0.244. Then the turns number in primary side was found from Equation M5. It is 3.57. Since selecting it as an integer will make easier handmade production, it was rounded to4. Therefore, the turns number of secondary side is 16.

$$N_p = \frac{L_m I_p}{B_{sat}A_e} x10^6 \quad \text{(M5)}$$

Before cable selection, rms and peak current of the secondary side should be chosen. These can be chosen according to M6 and M7 respectively. From these equations peak and rms current was found as 4.70 A and 1.92 Arms.

$$I_{sec} = I_p \frac{Np}{Ns} \quad \text{(M6)}$$

$$I_{sec,rms} = I_{sec}\sqrt{\frac{(1-D_{max})}{3}} \quad \text{(M7)}$$

2.4 Cable Selection

To make correct cable selection, firstly area of cable should be determined. To find this areas, current density should be selected. For long cables, current density can be taken as J=5A/mm^2. From this value, copper area primary and secondary current can be calculated with respect to M8.

$$A_{copper} = \frac{I_{rms}}{J} \tag{M8}$$

The copper areas of primary and secondary side can be found as 1.53mm^2 and 0.38mm^2 respectively. To choose appropriate cables for the design, American Wire Gauge Table should be checked. From the Table-1, AWG15 cable for primary and AWG 21 can be chosen for secondary side winding. However, the skin depth of the cables should be considered. According to this table, AWG 15 reaches its skin depth limit at 8250 Hz and AWG-21 reaches it at 33 kHz. However, system switching frequency is 60 kHz. So, smaller but parallel cables should be chosen. The AWG-24 is suitable for this design because it reaches its limit at 68 kHz. Finally, AWG-24 was chosen, and it will be paralleled 8 times at primary and 2 times at secondary.

G gauge	Conductor Diameter Inches	Conductor Diameter mm	Conductor cross section in mm^2	Ohms per 1000 ft.	Ohms per km	Maximum amps for chassis wiring	Maximum amps for power transmission	Maximum frequency for 100% skin depth for solid conductor copper	Breaking force Soft Annealed Cu 37000 PSI
15	0.0571	**1.45034**	**1.65**	3.184	10.44352	28	4.7	8250 Hz	94 lbs
16	0.0508	**1.29032**	**1.31**	4.016	13.17248	22	3.7	11 k Hz	75 lbs
17	0.0453	**1.15062**	**1.04**	5.064	16.60992	19	2.9	13 k Hz	59 lbs
18	0.0403	**1.02362**	**0.823**	6.385	20.9428	16	2.3	17 kHz	47 lbs
19	0.0359	**0.91186**	**0.653**	8.051	26.40728	14	1.8	21 kHz	37 lbs
20	0.032	**0.8128**	**0.519**	10.15	33.292	11	1.5	27 kHz	29 lbs
21	0.0285	**0.7239**	**0.412**	12.8	41.984	9	1.2	33 kHz	23 lbs
22	0.0253	**0.64516**	**0.327**	16.14	52.9392	7	0.92	42 kHz	18 lbs
23	0.0226	**0.57404**	**0.259**	20.36	66.7808	4.7	0.729	53 kHz	14.5 lbs
24	0.0201	**0.51054**	**0.205**	25.67	84.1976	3.5	0.577	68 kHz	11.5 lbs
25	0.0179	**0.45466**	**0.162**	32.37	106.1736	2.7	0.457	85 kHz	9 lbs
26	0.0159	**0.40386**	**0.128**	40.81	133.8568	2.2	0.361	107 kHz	7.2 lbs

Table *3*- American Wire Gauge (AWG) Conductor Table

2.5 Air Gap

The core is ferrite core. To prevent saturation of the core, an air gap is necessary for this core. The air gap of the core can be calculated by using Equation M9.

$$Air\ gap = \frac{A_e\,\mu}{A_L} \tag{M9}$$

From the equation, airgap for the core can be found as 4.5mm. However, E42/21/20-3C94 is gapless core. Therefore, 1.5mm gap will be okey since all three legs of core will have this gap.

2.6 Core and Copper Loss

To design transformer of flyback, all parameters is considered. However, the performance of the transformer should be calculated to see efficiency of overall system. The core loss can be found by using Steinmetz' Equation.

$$P_{core} = af^{x}B^{y}V_{core} \tag{M9}$$

The a, x and y values can be found at the core's datasheet. Ferroxcube provides a excel sheet for these coefficients. From this sheet, a, x and y are 3.530102481, 1.419999968 and 2.884999936 respectively. The volume of the core can be calculated from dimensions of core and it is 5130mm^3. When these parameters are put to equation M9, core loss can be found as 2.52W.

$$P_{core} = (3.53)(60000)^{1.42}0.3^{2.88}(5130 * 10^{3}) = 2.52W$$

The other loss is copper loss of the transformer. To find cupper loss, resistivity of coppers should be found firstly. The resistivity can be found with respect to the equation M10.

$$R = \frac{N*MLT}{(\#\ of\ parallel\ cables)} \tag{M10}$$

From this equation, resistivity of primary and secondary side can be found as 2.47mΩ and 39.5mΩ respectively. The cupper loss can be calculated by using equation M11.

$$P_{copper} = P_{cu,pri} + P_{cu,sec} \tag{M11}$$

$$P_{cu} = (7.68Arms)^{2}(2.47m\Omega) + (1.92Arms)^{2}(39.5m\Omega) = 0.29W$$

Output Voltage Ripple Calculation for Worst Case (Minimum Input Voltage, Maximum Load)

$$\Delta V = 48 * 0.03 = 1.44V \quad \Delta Q = C\,\Delta V$$

$$\frac{I_{out}\,\Delta t}{C} = \frac{I_{out}\,D\,T_s}{C} = \Delta V$$

$$C_{out} = \frac{1 * 0.5}{60 * 10^{3} * 1.44} = 5.78\ \mu F$$

Note: To keep the voltage less reactive during sudden load changes, we will be using an output capacitor $C_{out} = 47\ uF$. Choosing these values ensures that we both satisfy the output voltage ripple constraint and regulation constraint.

RCD Snubber Calculation

Let $V_{snub} = 2.5\ V_{in} = 30V$

$$I_{pri\_max} = I_{pri\_avg} + \frac{1}{2}I_{ripple} = \frac{(Vout + V_f)I_{out}}{V_{in}\,f_s\,D\,T_s} + \frac{V_{in}\,D\,T_s}{2\,L_{pri}}$$
$$= \frac{(48+0.7)*1}{12 * 60 * 10^{3} * 0.5 * \frac{1}{60 * 10^{3}}} + \frac{12 * 0.5}{2 * 13 * 10^{-6} * 60 * 10^{3}} = 12\ A$$

Let $\Delta V_{snub} = 0.1\ V_{snub} = 3V$

$$\Delta V_{snub} = \frac{V_{snub}}{C_{snub} * R_{snub} * f_s}$$

$$3 = \frac{30}{C_{snub} * R_{snub} * 60 * 10^3} \quad => \quad C_{snub} = \frac{30}{3 * 480 * 60 * 10^3} = 347\ nF$$

$$R_{sn} = \frac{V_{snub}^2}{\frac{1}{2} L_{leak} * i_{peak}^2 * \frac{V_{snub}}{V_{snub} - V_{in}} * f_s}$$

$$R_{snub} = \frac{V_{snub}^2}{\frac{1}{2} 13 * 10^{-6} * 0.02 * 12^2 * \frac{30}{30 - 12} * 60 * 10^3}$$

$$= 480\ \Omega\ (Assumed\ leakage\ inductance\ is\ \%2)$$

RC Snubber Calculation

In calculating the RC snubber for the secondary side diode, procedure in [2] was followed.

Oscillation frequency: $f_0 = 10MGhz$

Shifted oscillation frequency $f_1 = 4MGhz$ with parallel capacitor $C_1 = 1000pF$

Snubber resistor $R_{snub} = 83.56\Omega$, snubber capacitor $C_{snub} = 571.43pF$

Note: In order to optimize the losses and reduce the peak oscillation voltage and frequency to a desired value, the final values of the components were changed as follows:

$R_{snub} = 100\Omega$, $C_{snub} = 330pF$

2.7 Production of Transformer

As the transformer is calculated theoretically, the primary side is paralleled as 8 windings. For safety, length of the cables was taken with respect to 5 turns. The secondary side was prepared with respect to 2 parallels and 18 turns. To obtain the desired air gap (1.5mm for each leg) A4 paper pieces were added between two cores. At the end of the production of the transformer primary side is measured as 8.57uH. Secondary side is 150uH and leakage inductance of the transformer is observed as 0.37uH. The primary and leakage inductance measurements can be seen in Figure 2 and Figure 3 respectively. According to these results, leakage inductance can be found as 6.5%. This is higher than the simulation results. This leakage inductance can cause higher ripple on circuit and higher noises. Because of that, the snubber design of the circuit should be changed again. As a result, it was decided to produce the transformer one more time.

Figure 2 Primary Side Inductance of First Transformer

Figure 3 Leakage Inductance of First Transformer

In the second try, sandwich type winding is decided to use. The first two turns of the primary side were winding first. Then, the secondary side was winding to the coil former. Finally, the last two turns of the primary side were finished. After this operation, the primary side of the transformer was measured as 12.3uH and the secondary side was 196uH. The desired value for this transformer was 13uH. Therefore, the gap of the transformer is adjusted more properly than before. The leakage inductance of the transformer was measured as 0.2uH. This means that the leakage inductance is only 1.6%. Therefore, this design was chosen as the final transformer design. The primary, secondary and leakage inductance measurements can be seen in Figure 4, Figure 5 and Figure 6 respectively.

Figure 4 Primary Side Inductance of Second Transformer

Figure 5 Secondary Side Inductance of Second Transformer

Figure 6 Leakage Inductance of Second Transformer

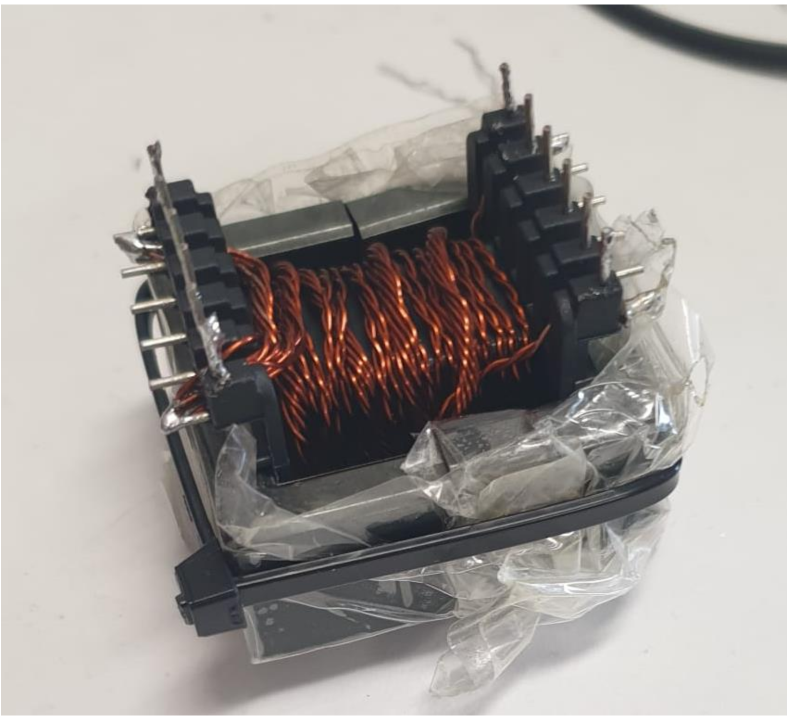


Figure 7 Final form of the Transformer

During the production process, one of the magnetic cores of the transformer has fallen to the ground and shattered into 3 pieces while trying to adjust the air gap required for the transformer. As a practical solution, we've proposed to join the pieces of the core using tape and a cable tie due to time constraints and material logistic insufficiencies. The final form of the transformer can be seen in Fig. 7. Unexpectedly, our new core had better leakage properties (2% leakage) and higher inductance compared to our first transformer. In the end, the transformer was working properly and possibly didn't saturate.

3. Design decisions

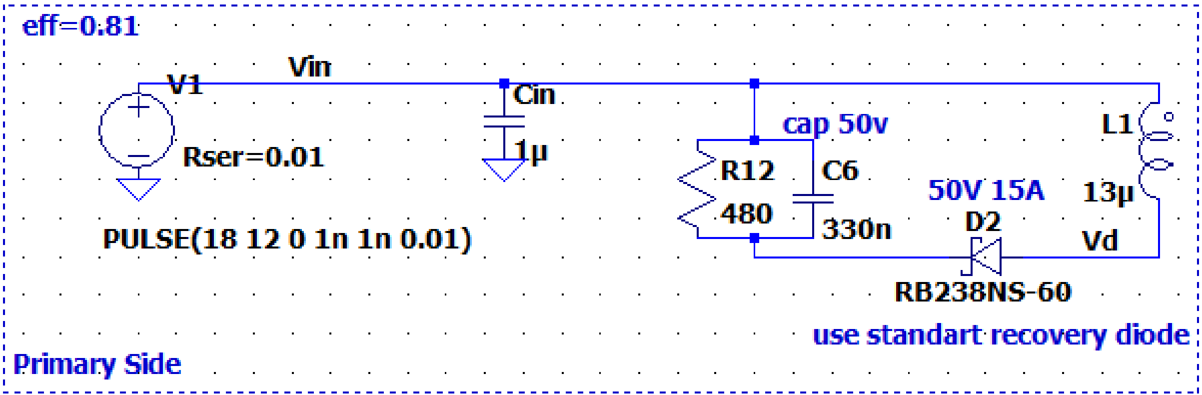


Figure 8 Primary Side of the Flyback Converter

-In the simulations, we used $C_{in} = 1\ \mu F$, however in real life applications, we saw that this value for an input capacitor is not enough. Therefore we chose to use a $C_{in} = 1000\ \mu F$ capacitor.

-The RCD snubber components are chosen regarding the section before.

-The snubber diode is chosen so that it can supply the average snubber current and peak oscillation currents.

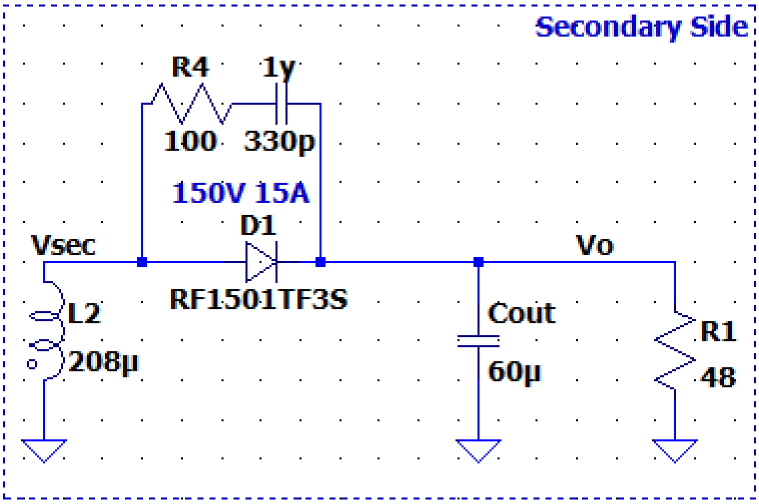


Figure 9 Secondary Side of the Flyback Converter

-Since there are no 60 uF capacitors available commercially, we opted to include an output capacitor $C_{out} = 47\ uF$.

- The RC snubber components are chosen regarding the section before.

-The output diode will be able to withstand peak currents of the oscillations and the average output current.

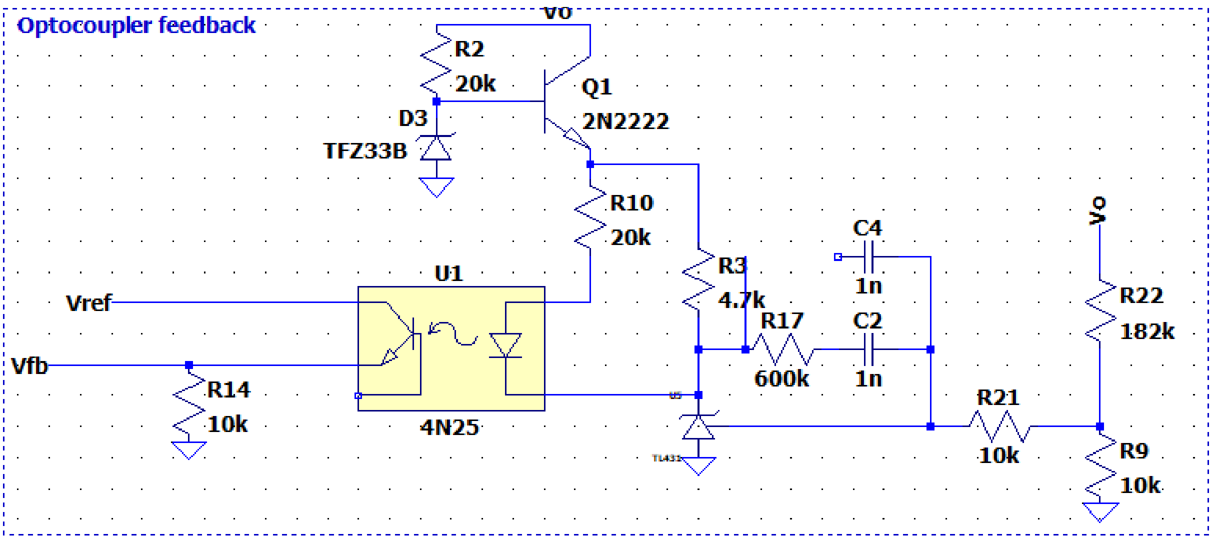


Figure 10 Optocoupler Feedback Configuration

-TL431 Adjustable Voltage Reference is used to provide feedback from the output voltage. Since it can achieve 2.5V at the reference pin, we can reflect that voltage to the output using a voltage divider network. Since the maximum operating voltage of TL431 is 37V, a linear regulator is implemented with a 30 Zener diode to protect the component. Because this stage is not a power stage, the power loss from the linear regulator is negligible.

-4N25 optocoupler is used because it is included in the LTSpice library. The resistors are connected to the optocoupler anode adjusts the controller parameters, so it is chosen such that is slows down the response. A 4.7k resistor is connected in order to protect the component.

-TL431 component is used as a PI controller in this stage. The parameters of the controller is adjusted so that the regulation constraints are satisfied. In this resistor and capacitor configuration, the converter satisfied these requirements. If required, a compensation capacitor can be adde to the controller to slow down the response.

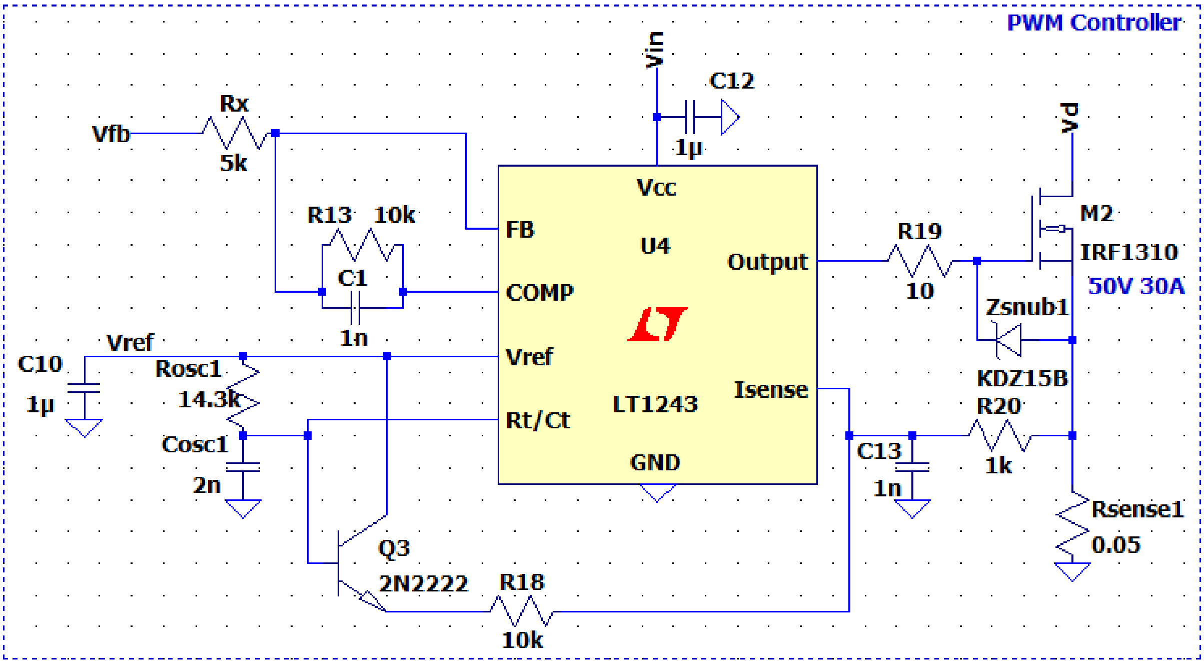


Figure 11 PWM Controller Configuration

-For our controller IC, we opted to use UC 3843 Current Mode PWM Controller (LT1243 is pin to pin equivalent of this IC in LTSpice) because of its current mode abilities. It includes a gate driver for the MOSFET and can operate up to %100 duty cycle (Although our maximum duty cycle is %50).

-Since we are operating at 60kHz frequency, our R_{osc} and C_{osc} values were chosen accordingly to set the oscillator frequency inside the PWM controller.

-A protection Zener diode is used to protect the MOSFET V_{GS} from voltage spikes. Additionally, A 10R resistor is used to limit the charging currents of the MOSFET capacitors.

-1µF bypass capacitors are used to supply a reasonably constant input and reference voltage to and from the IC.

-To use the current mode capability of the IC, we will be using a 50mΩ sense resistor to measure the primary current. To supply a noise-free voltage I_{sense} pin of the controller, a low-pass filter is implemented on the sensed voltage. Additionally, a slope compensation network is implemented to reduce the voltage spikes in the waveform.

-The controller parameters of the controller are determined by trial and error and the resistor and capacitor values are chosen to provide adequate Kp and Ki values.

3.1 PCB Design

The PCB design of flyback converter are divided into three parts. These are the controller part, primary side, and secondary side. Since the flyback is an isolated converter, the ground of primary and secondary must be separated. The ground of the controller is connected to primary side of the converter.

The initial schematics of the converter can be seen in Figure 1. To decrease the complexity of the circuit, voltage protection of optocoupler is design with only Zener diode. The input and output capacitors are divided into different capacitor values to eliminate noises at different frequencies. There are capacitors at the Vcc pin of the IC to clear noises at the supply voltage of it.

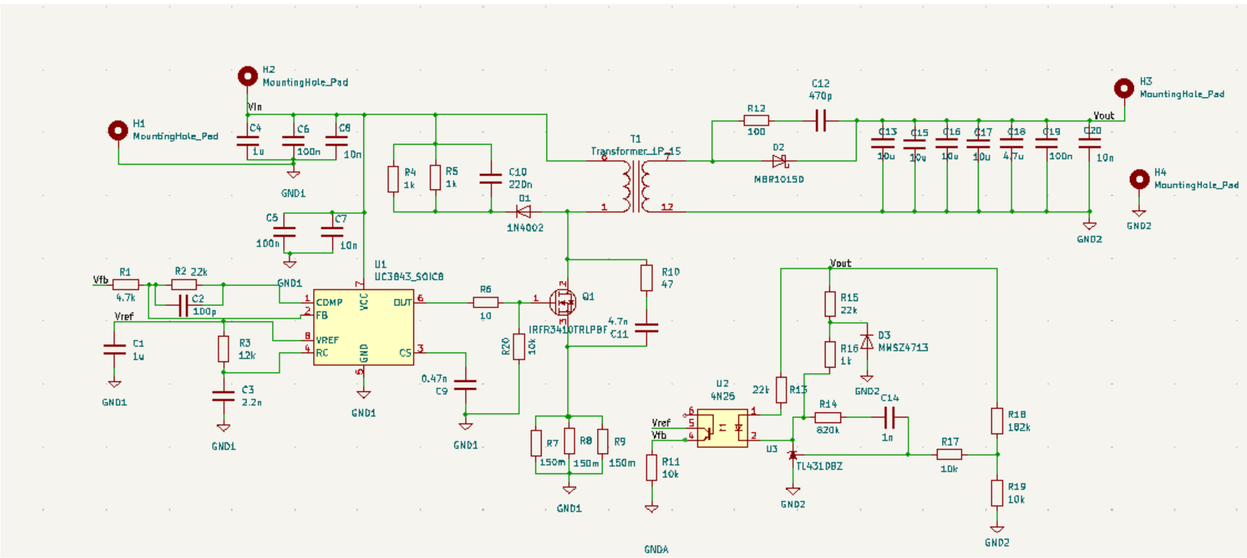


Figure 12 Initial Schematic of Flyback Converter

Top and bottom layer of PCB can be seen in Figure 2 and Figure 3 respectively. As can be seen in Figure 2, controller circuit is placed left side of the PCB. The components are placed as close as possible. The switching node is also drawn as small as possible to decrease stray inductances since the switching frequency is high. The near of optocoupler, protection and feedback circuit are placed. Feedback line at primary side and anode pin of optocoupler is drawn at bottom since there is not enough place.

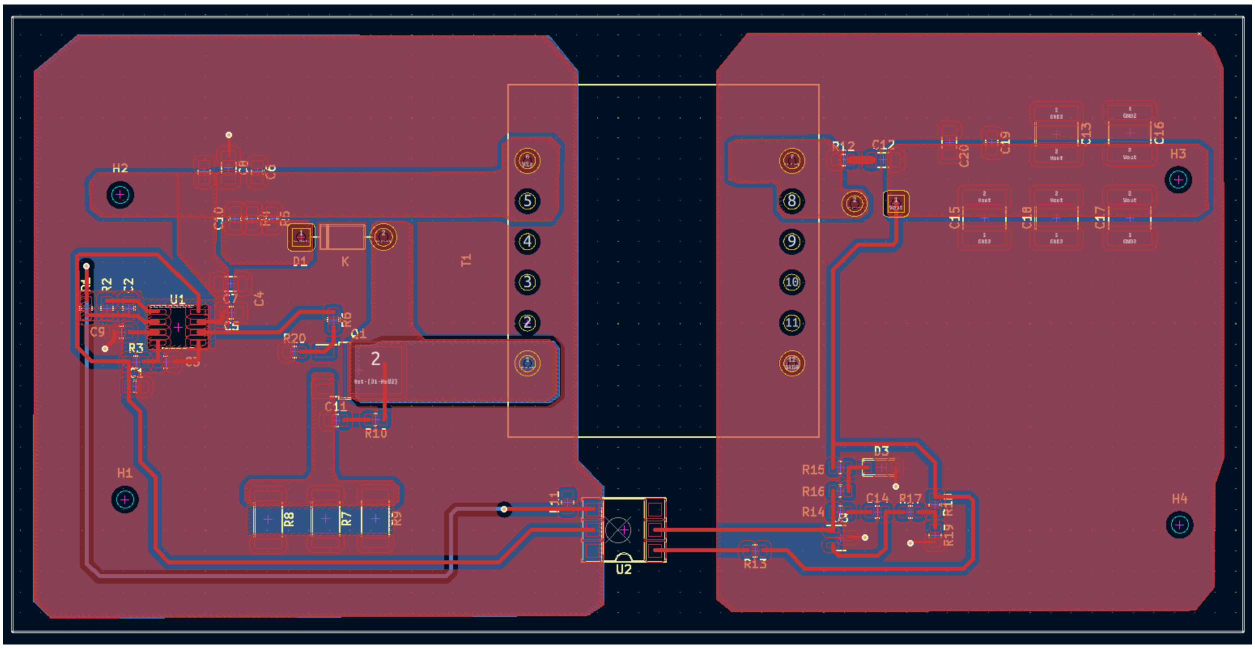


Figure 13. Front Side of Flyback Converter PCB

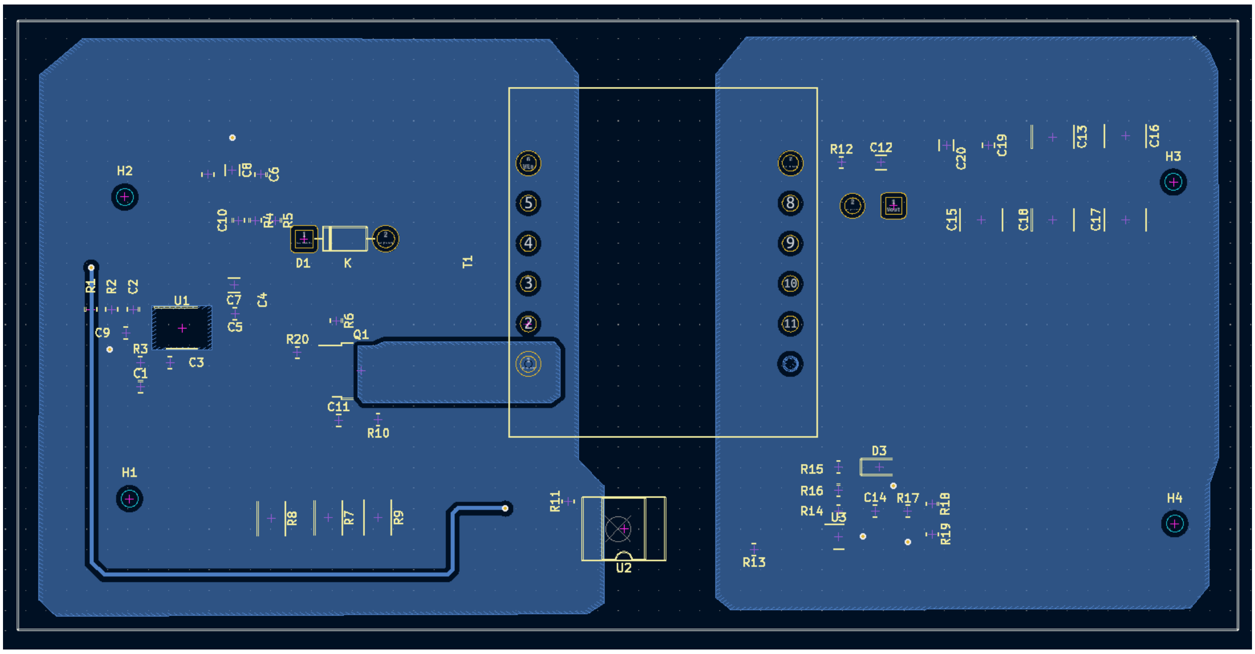


Figure 14. Back Side of Flyback Converter PCB

After experimental tests of the circuit, PCB is updated with some jumpers since there is no time to produce again. Firstly, just upper side of optocoupler, at secondary side, the protection circuit with BJT is placed again. Current sense connection is completed with 1kΩ through hole resistance to filter current sense output. To compensate inrush current at the beginning of transient, a higher value electrolytic capacitor added at the input. The value of it is 2.2mF 50V. Finally, to decrease output ripple, two electrolytic capacitors, 470uF and 330uF are added at the output of the circuit. The final circuit can be seen in Figure 4. The input capacitor is added after this photo is taken.

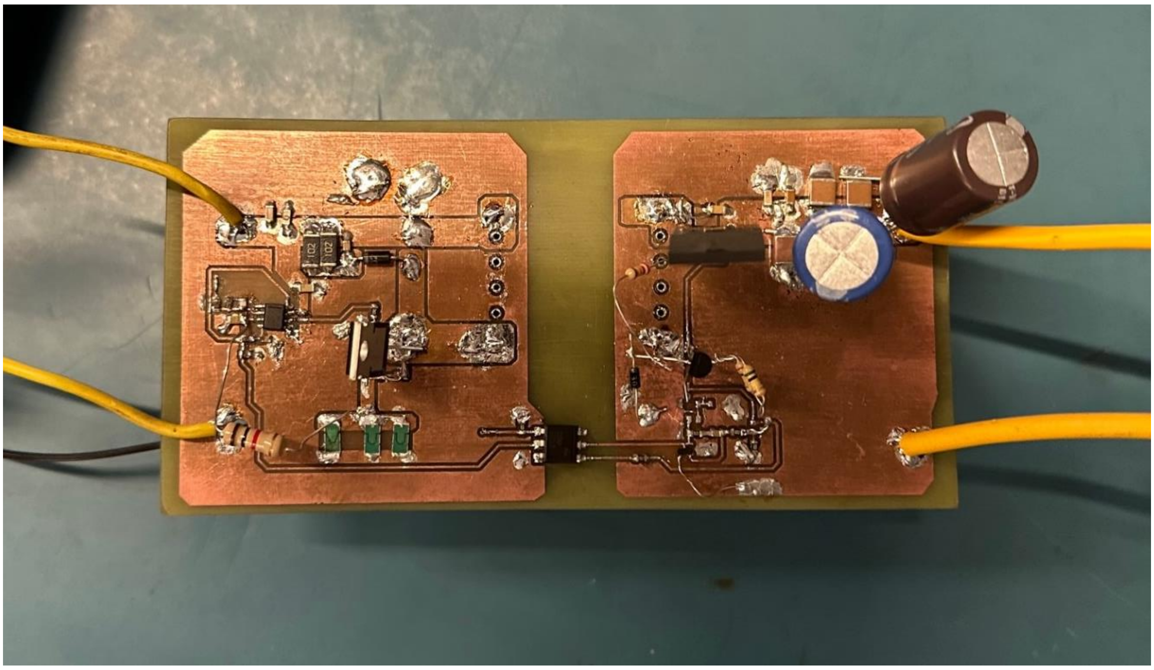


Figure 15. Final PCB Design

4. Practical Results

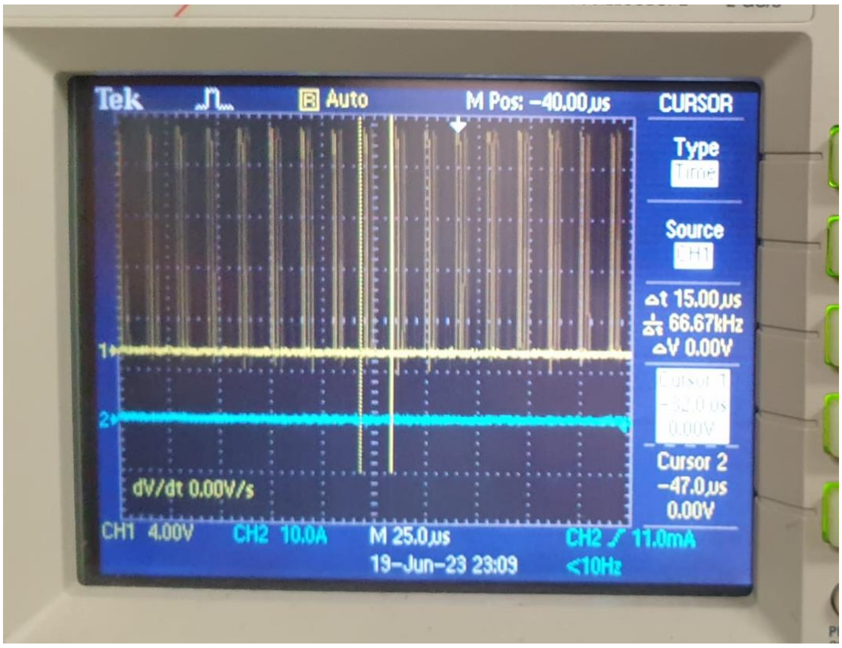


Figure 16. Gate-Source voltage of the MOSFET during %30 load operation

Figure 17. Output voltage, current and power for %30 load operation

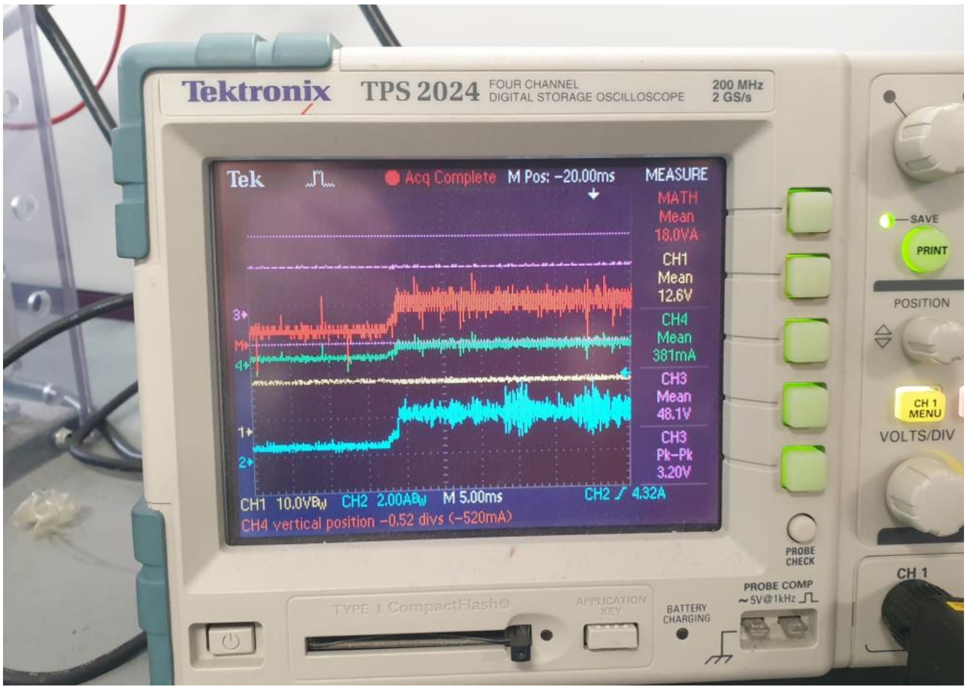


Figure 18. Response time for the 30% load operation

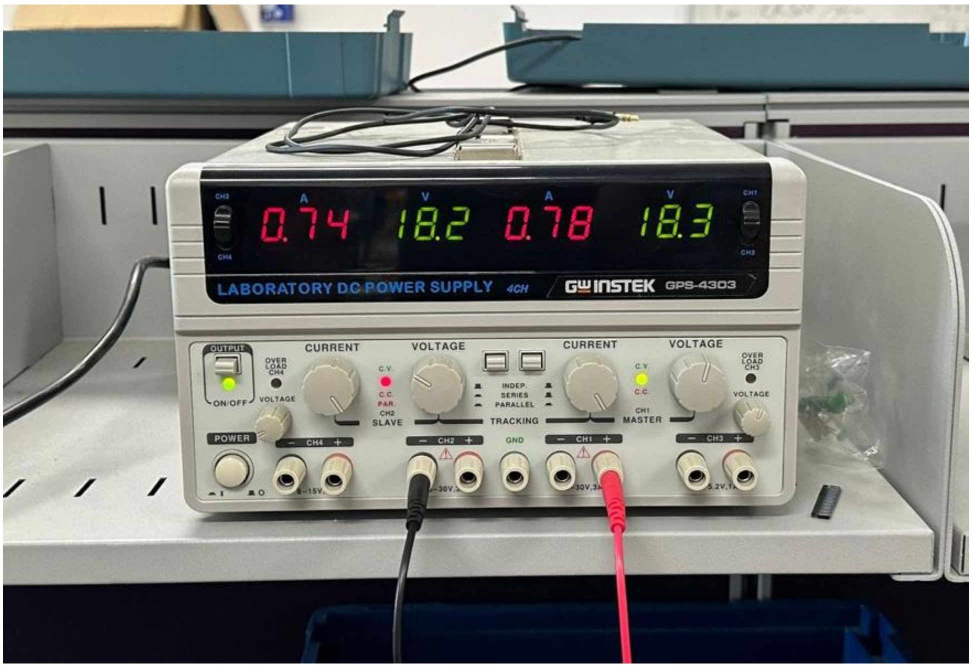


Figure 19. Input voltage and current for %50 load operation

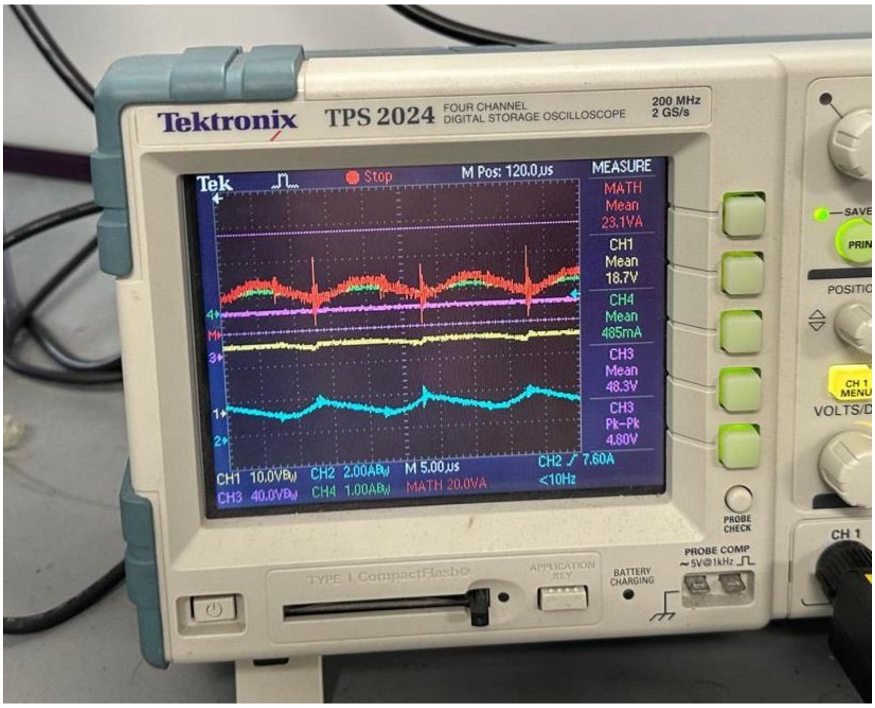


Figure 20. Output waveforms for %50 load operation

Comparing these two figures we can see that our efficiency in 50% load operation is 86%.

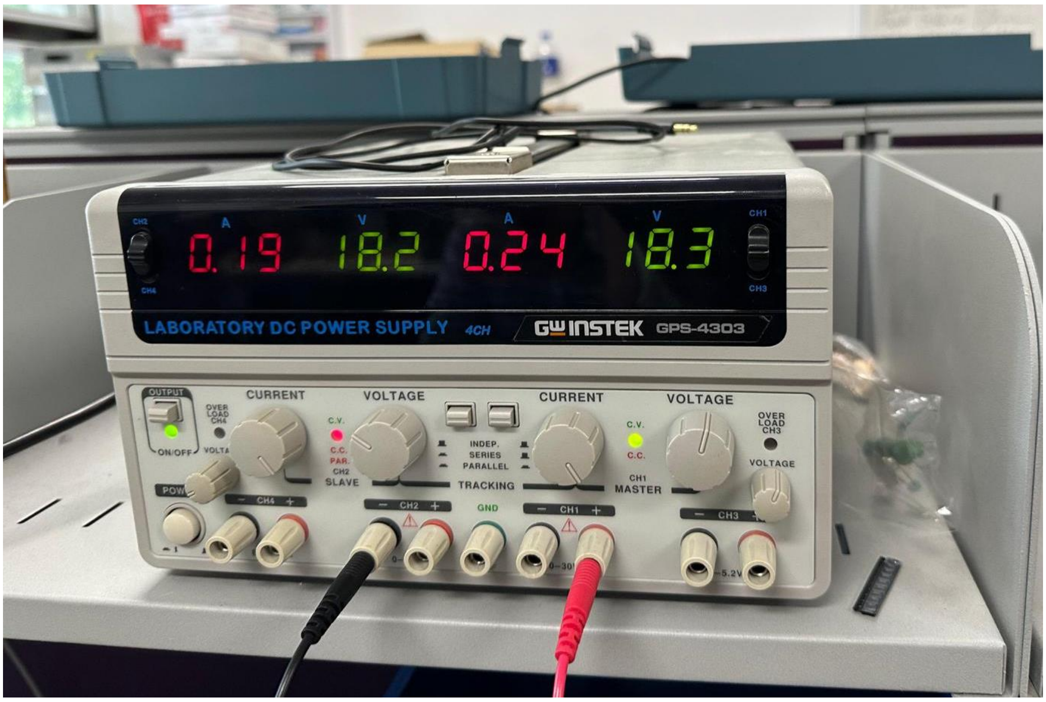


Figure 21. Input voltage and current for %10 load operation

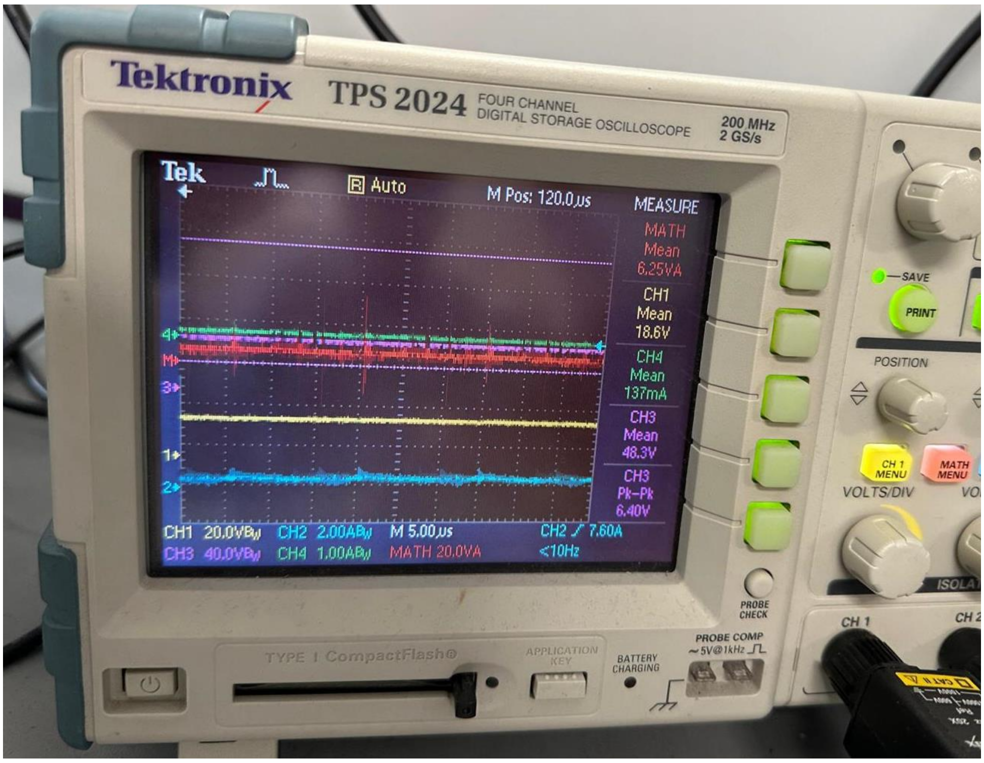


Figure 22. Output waveforms for %10 load operation

Comparing these two figures we can see that our efficiency in 10% load operation is 85%.

Figure 23. Input voltage and current for full load operation (Converter breaks down and doesn't function properly)

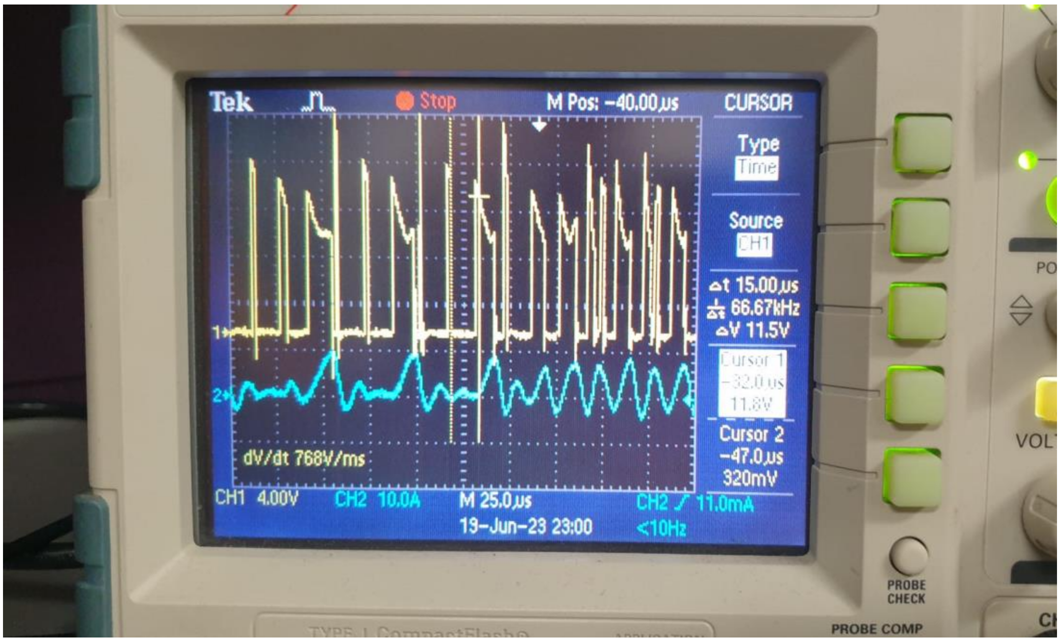


Figure 24. Gate-Source voltage of the MOSFET during full load operation

5. Component Selection

In the flyback converter there are two critical component selections. These components are diode and switch (MOSFET). While choosing the components, we have decided to stay in the safe zone at first such that our converter would work as intended. After constructing the circuit, we are planning to change the components with narrower ratings to achieve a more efficient converter. While choosing the switch, there are two considerations which are the peak switch current and peak switch voltage.

$$V_{SW,peak} = \frac{V_d}{1-D} = \frac{18}{0.4} = 45V$$

$$I_{SW,peak} = \frac{1}{(1-D)} * \frac{N1}{N2} * I_O + \frac{N1}{N2} * \frac{(1-D)}{2Lm * f_s} V_O = 5.01A$$

In the simulations, it is seen that Vds peak is around 60V, however we will be going for a safer option and use a 200V, 9A rated IRF630 MOSFET.

From the LTSpice simulations, it is also seen that a diode with about 150V and 15A ratings should be selected to achieve safe working conditions. We have decided to select MUR1615CT diode to meet our requirements. This diode has 150V and 16A ratings.

For the RCD snubber circuit, we have decided to use RB238NS-60 Schottky barrier diode, which has 60V and 40A ratings.

6. Further Discussion & Analysis

During our operational tests, we've encountered some problems, which are:

-Due to the unexpected behavior of the circuit, the MOSFET in the primary side and the diode in the secondary side broke down. Therefore we had to change these components with one switch higher ratings for the continuity of the project. The reason for this behavior could be the insufficiency of the snubber parameters for these components.

-In our PCB design, we've omitted some of the components and routes. For this reason, we had to bypass some of the PCB parts using through hole components and cables. This evidently caused a decrease in the EMI/EMC compatibility because of the extra paths caused by the bypassed components. This solution might have caused some of the components to malfunction unexpectedly, for instance, our PWM controller UC3843 broke down several times without a defined reason.

-At the end, our converter function properly until 50% load. However, the output current didn't achieve 1A at full load case. The main reason for this could have been the current limit of the power supply. In our operational tests, we've observed that with a low current limit setting on the power supply, our circuit didn't function properly, and the input current would be stuck in the maximum current limit. After some trials, we've observed that shorting the current sense resistor to ground would be enough to start our circuit, which eventually forced us to increase the current limit on the power supply. In our simulations however, the input current ripple for the converter was ~14A, which is well above the current limit of 6A of the power supply. For this reason, the power supply couldn't provide us with that much current and our converter didn't function properly at full load case.

7. Conclusion

In this project, we have designed an isolated DC-DC converter with certain specs and included our design process and results in this report. This report includes the design process, topology selection, magnetic design, practical results, and PCB design. For meeting the project requirements, the flyback topology was selected and while the load is low, our converter works properly, however when the load becomes high the converter stops functioning properly. The analysis of the working of the converter is discussed in the further discussions & analysis part. The main reasons for the choosing of flyback converter topology is its relative easiness to control and small amount of switching losses due to having only one switch. Another reason is the small size of the flyback converter relative to other topologies due to having less components and inductors. The magnetic design process consists of iterative design to reach the most ideal transformer design for the transformer which fits the project requirements. To sum up, our simulations suggested that our design was compatible with the project requirements, however in practice, this proved not to be the actual case. The possible causes for the shortcomings of our operational tests are included in the further discussion & analysis part. In this project, we have learned so much about the design process of a flyback converter and surely this project will be immensely helpful in our professional lives.

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