MIDDLE EAST TECHNICAL UNIVERSITY

ELECTRICAL & ELECTRONICS ENGINEERING

EE464 - STATIC POWER CONVERSION II
PROJECT II

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1 Analytical Calculations

1.1 a)

To obtain the transfer function of the push-pull converter, we should be properly understand its operation first. The schematic in Figure 1.1.1 is to be used while investigating the topology.

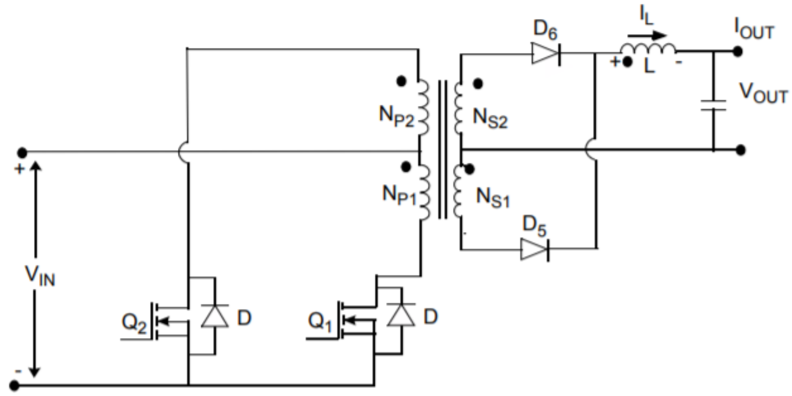

Figure 1.1.1: Push Pull Converter Topology

Switches Q_1 and Q_2 are turned in half cycles of the overall switching time, T_s. That is to say that they will be in ON state for DT_s period and they will turn on with $T_s/2$ time difference. It should be noted that D should strictly be smaller than **0.5**. A wise decision would be to keep a margin for the magnetizing current to flow through the body diodes as well during the OFF states of switches.

In the first DT_s period, Q_1 is ON. First primary winding sees input voltage on it. N_{S2} winding forward biases the D_6 diode and inductor charges up. Inductor current during this period is obtained to be

$$V_L = nV_s - V_O$$

where n is the turns ratio, $\frac{N_S}{N_P}$

After that time, for a -ideally- $T_s/2 - DT_s$ period, both of the switches stay OFF. Inductor current splits to two in the secondary side. Assuming the turn numbers are equal in secondary side windings, secondary side voltage on transformer becomes 0. During this period,

$$V_L = -V_O$$

Same chain of events happen in the remaining half cycle, only the switches are changed. If the voltage-seconds rule for the inductor is applied, the transfer function of the converter can be derived.

$$2(nV_S - V_O)DT_s - V_O(1-2D) = 0$$
$$2nV_S D - 2V_O D - V_O + 2V_O D = 0$$

Transfer function can now be obtained as

$$\frac{V_O}{V_S} = 2nD$$

1.2 b)

The output capacitor never feeds the load by itself during operation. Therefore we should use the inductor current ripple to find the voltage ripple on the capacitor.

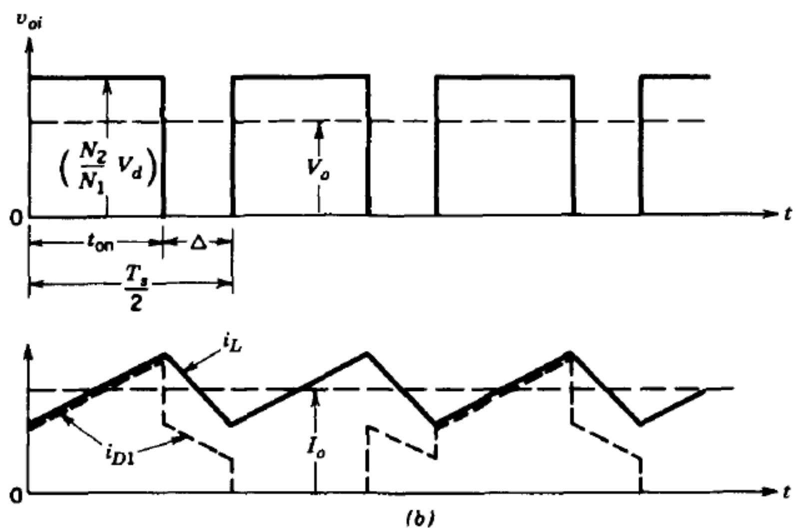

Figure 1.2.1: Inductor Current in Push Pull COnverter

During OFF times of the switches, inductor voltage equals $-V_{out}$ and inductor current drops. Let us start with this period. We know

$$\Delta i_L = \frac{V_L \Delta t}{L}$$

Therefore,

$$\Delta i_L = \frac{-V_O(0.5-D)T_s}{L}$$

.

Assuming inductor current changes linearly, half of this peak-to-peak ripple charges the capacitor for $0.25T_s$. We are talking about the small triangle on top of average line in Figure 1.2.1. We also know that

$$\Delta Q = \Delta VC = \frac{\Delta i_L}{2} 0.25T_s$$

Insert what is found so far for the ripple equation now to obtain the output voltage ripple as follows.

$$\Delta V_O = \frac{V_O(0.5 - D)T_s^2}{8LC}$$

2 Circuit Parameters

2.1 a)

We will use the output voltage equation found in Part 1.1. $V_S = 48V$ and $V_O = 12V$ in our design. $D = 0.25$ is chosen. Insert all in the following formula to find turns ratio, n.

$$\frac{V_O}{V_S} = 2nD \implies n = \frac{V_O}{2DV_S} = \frac{12V}{48V * 2 * 0.25} = 0.5$$

We will need a center-tapped, step-down transformer with turns ratio **0.5**.

2.2 b)

Inductor ripple current expression was found theoretically in Part 1.2 as follows.

$$\Delta i_L = \frac{V_O(0.5 - D)T_s}{L}$$

We aim for a converter that can supply $96W$ under $V_O = 12V$. This means an average output current value of $8A$. It can easily be seen that average inductor current is also equal to that value. We can then deduce

$$\Delta i_L = 0.8A$$

To find L_{min}, we can alter the ripple formula. We also know that $D = 0.25$ and $T_s = 1/40kHz = 25\mu s$. Then insert all in the formula.

$$L_{min} = \frac{12V * 0.25 * 25\mu s}{0.8A} = 93.75\mu H$$

Filter inductor should be **at least** $93.75\mu H$ to guarantee 10% inductor current ripple.

2.3 c)

It is aimed for 1% output voltage ripple. This corresponds to $\Delta V_O = 0.12V$. The ripple formula found in Part 1.2 is to be used as follows.

$$C_{min} = \frac{V_O(0.5 - D)T_s^2}{8\Delta V_O L} = \frac{12V * 0.25 * 625ps}{8 * 0.12V * 93.75\mu H} = 20.8\mu F$$

Output capacitor should be **at least** $20.8\mu F$ to guarantee 1% output voltage ripple.

3 Magnetic Design

3.1 a)

A ferrite core from *Magnetics Inc.* is chosen for the transformer. "R Material" ferrite is a medium frequency, multi purpose magnetic material. It has a saturation density 0.47T and permeabilities do not change till 1Mhz. It suites well for our application.

At the beginning, "Transformer Design" note from 2013 Catalog of *Magnetics Inc.* is used. But the given method there is not used for calculations. Different cores are put into test from the "Power Handling Capability" table in that document and at last **0R42530EC** is chosen.

To find the turns number, Faraday's Law is used. We can find the primary side turn number as follows.

$$N_{p1} = \frac{V_s D T_s}{A_e B_{max}}$$

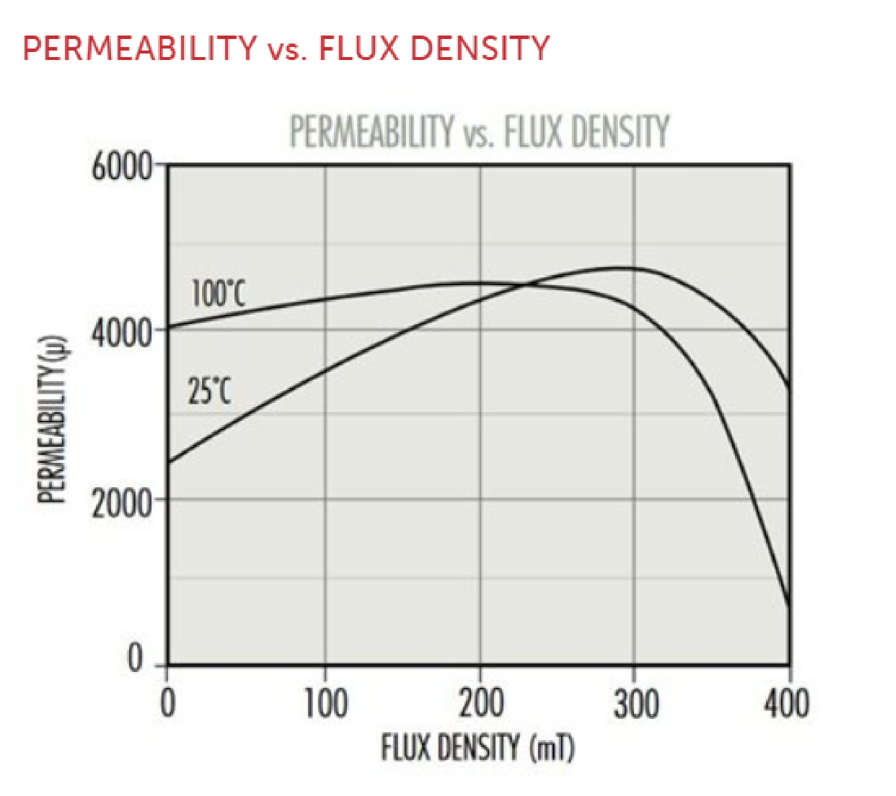

Figure 3.1.1: R Material Flux Density vs Permeability Graph

We restrict our B_{max} to **0.25 T**. It is reasonable choice as can be seen from Figure 3.1.1. A_e is given in the datasheet as $80.2mm^2$. Inserting these numbers in the formula and adding 2 more turns to ensure safety we obtain the turn numbers as

$$N_{p1} = N_{p2} = 18$$

$$N_{s1} = N_{s2} = 9$$

Magnetizing branch inductance can be calculated by looking at the A_l value of the core. This design gives the L_m value as **0.74mH**, which is a reasonable value.

3.2 b)

We have 40kHz of switching frequency. In order to not suffer from the skin depth problems, we should choose our cables accordingly. We will also stick to $4A/mm^2$ current density in the cables.

At the input, we have average $I_s = 2A$. This means we need a conductor area around $0.5mm^2$, at least. Cables having that cross section area are suffering from skin depth problems. To avoid this, we can use 2 of **AWG#22** in parallel. They have no skin depth problems till 43kHz.

At the output, we need at least $2mm^2$ of conductor area. Accomplishing this with the previous choice means using 7 of them in parallel. It is not a good practice to use this much. Instead, we choose 4 of **AWG#20** in parallel while increasing AC resistance of the cables a bit. We can compensate for it since 40 kHz is not that high of a switching frequency.

A small table can be provided about our decisions.

Primary	AWG#22 x 2 turns	3.05 A/mm^2
Secondary	AWG#20 x 4 turns	3.85 A/mm^2

In primary side, there will be 36 turns consisting of 72 turns of AWG#22 cables. This means a conductor area of $\mathbf{23.54mm^2}$ is needed. Similarly in secondary side, there will be 18 turns consisting of 72 turns of AWG#20 cables. This adds up to a conductor area of $\mathbf{37.36mm^2}$. Window area of the chosen core is calculated from the dimensions given in datasheet as $\mathbf{161.8mm^2}$. Fill factor in this case is

$$K_f = \frac{Conductor\ area}{Window\ area} = \frac{23.54 + 37.36}{161.8} = 37\%$$

Lastly, the winding resistances are to be calculated. The mean turn length is again estimated from the dimensions. To be on the safe side, a margin is added as well. DCR values per length is available from AWG tables. Since we have used them in parallel, they will be decreased. Also, it should be noted that primary cables are not suffering from skin depth problems, therefeore their AC resistances are same as their DC resistances. For secondary side, AC resistance is presented. The calculations are omitted here since they are trivial. Skin effect on the secondary side is also taken into considerations. Results are obtained as

$$R_p = 58m\Omega$$

$$R_s = 8.1m\Omega$$

Copper losses are easy to find now. Take $I_s = 2A$ and $I_o = 8A$ and use the renowned power formula. Result is **0.75W** of total copper loss in transformer.

3.3 c)

For the core losses, the Excel file provided by the *Magnetics Inc.* is used. For R material operated at $f_s = 40kHz$ and reaching up to $B_{peak} = 0.25T$, table provides a loss of $242.38mW/cm^3$. The screenshot of the table can be seen in Figure 3.3.1.

MAGNETICS ®

FERRITE MATERIAL
CORE LOSS CALCULATOR

$$P_{CL} = \frac{a f^x B^y L(T)}{1000}$$

		Enter Values Here	Units
P_{CL}	Core Loss (mW/cm^3)		
f	Frequency (Hertz)*	40,000	*Hertz*
B	Peak Flux Density (Tesla)	0.25	*Tesla*
T	Temperature (°C)	60	*(°C)*
$L(T)$=	b-c(T)+d(T^2) For All Other Temperatures		
$L(T)$ =	1 for 100°C		

Material	Frequency Range	a	x	y	b	c	d	Core Loss P_{CL} (mW/cm3)
R Material	20kHz-150kHz	3.53	1.420	2.880	1.970000000	0.022260000	0.0001250000	242.08
	150kHz-400kHz	5.88E-04	2.120	2.700	2.160000000	0.023270000	0.0001170000	

Figure 3.3.1: Core Loss Calculator

Multiplying this number with the core volume give us the core loss total as **1.43W**.

All the calculations are conducted on an Excel file, which can be found on the GitHub repo for this project. A screenshot from this Excel file is also provided in Figure 3.3.2.

D	0.25		Primary		Secondary		0R42530EC			
Ts	0.000025		AWG22 x2		AWG20 x4		Core Loss	Hacim	Window A	A_l
V_in	48		0.327		0.519		2.42E-04	5.90E+03	161.28	2.31E-06
B	0.25	N_pri	1.80E+01	N_sec	9					
A_e	8.02E-05									
			Alan_pri		Alan_sec	Total Alan				
			23.544		37.368	60.912		Fill	Core Loss	
Wire Lengt	61.08		2198.88		1099.44			0.377679	1.43E+00	
R_per km			26.45		8.3225					Total Loss
										2.18E+00
Total R			0.05816		0.00818			L_m	R loss	
Loss			0.232642		0.52352			7.47E-04	0.756162	

Figure 3.3.2: Magnetic Design for Transformer Excel File

Total transformer losses add up to **2.18W**, which is reasonable. Peak flux value inside the core can be lowered a bit to compensate for the core losses, but then winding turns should be increased and thus copper losses. It also would require a bigger core probably. Fill factor and losses seem perfectly well within the boundaries.

3.4 d)

We can use a toroid for the inductor core. Considering the availability and economics, we will stick to the ferrites. But keeping the peak flux density low enough, we need more turns and this can end up in over design. Our ripple will be lower at the end of the day, so no worries.

Once again an R material core from *Magnetics Inc.* is chosen. **ZR41809TC** is my choice for the output filter inductor. A_l value for the core is given as $2810nH/T^2$. It requires **6 turns**, but we will wind **14 turns** around it. We have a higher inductance than required but lower peak flux densities, lower core losses and safer operation.

Inductor voltage will be 12V at its maximum value. With 14 turns, it corresponds to **0.124 T** peak flux density inside the core. Considering that R material starts to saturate around 0.25T values -check Figure 3.1.1- we can be confident that core will not saturate.

Resulting actual inductance is **0.55 mH**, which deals with the current ripple way better than required inductance value for 10%.

3.5 e)

We can use the same cable set with the secondary winding here as well, 4 turns of AWG#20 cables in parallel. Window area of the core is calculated from the dimensions. With 14 turns of 4xAWG#20 cables, the fill factor is **38.9%**, which is a reasonable value.

Cable length to be is used, $410mm$. It corresponds to the following values for AC and DC resistances. Paralleling the cables is not considered here.

DC Resistance	13.6 mΩ
AC Resistance	13.9 mΩ

The difference is not that dramatic as can be seen. We are paralleling the cables also, so the effective resistance decreases further. Considering it will carry 8A continuously, the copper loss in the windings is calculated as **0.22 W**.

3.6 f)

Core loss of the material is again to be calculated using the Excel file provided by *Magnetics Inc.*. Peak flux density in our application is $B_{max} = 0.124T$. For this value overall core loss is calculated is **57.3 mW**. "R" is a low loss material and peak density is low as well, so this is expected.

Overall loss in inductor is around **0.28W** and a great portion of it is due to copper loss. This value is satisfactory enough.

The excel file used in this inductor design can also be found in Github. It looks like Figure 3.6.1

D	0.25		Cable 1 tu	2.076	Length tur	29.25
V	12		Conductor	29.064	Tot leng	4.10E+02
T_s	0.000025		Radius	4.875	res	0.003478
A_e	4.31E-05		Window	74.66191		
A_l	2.81E-06		Fill	0.389275		
L	1.00E-04		Volume	1783		
N	1.40E+01		Loss	3.21E-05		
B	1.24E-01		CoreLoss	5.73E-02		
L_act	5.51E-04		Winding lo	0.22256		
			Total Loss	2.80E-01		

Figure 3.6.1: Magnetic Design for Inductor Excel File

4 Simulation

4.1 a)

We have already designed the magnetic components. Semiconductors and capacitor will be chosen.

Diodes

Diodes carry 8A maximum and should withstand 48V when they are OFF. Although the current average is around 4A for them, let us stay on the safe side.

We can use model **SBRT10U60D1Q-13** from *Diodes Inc.*. It is Schottky diode with 60V-10A ratings. Forward voltage is 0.52 V, which is reasonable.Datasheet can be found here.

MOSFETs

Switching elements can be utilized using MOSFETs. They should be able to handle 96V at when OFF and will be peak currents go up to around 4A.

Chosen component is **Si7846DP** from *Vishay Siliconix*. It can withstand 150 V and carry 5A, continuously. ON state resistance is realtively low also, around $52m\Omega$. Dataheet can be found here.

Output Capacitor

Output capacitor should be holding at least 12 V and capacitance should be greater then $20.8\mu F$, as stated earlies in Part 2.3. A ceramic capacitor can be used in order to have low ESR values.

Chosen component is model **FG26X5R1E476MRT00** from *TDK Corporation*. It can withstand 25 V and has a capacitance of $47\mu F$. Datasheet can be found here.

4.2 b)

With non-idealities taken into account, system is working fine and requirements are satisfied with some expected discrepancies.

With $D = 0.25$ operation, output voltage is 11V. Ripple is in microvolts ranges. Result can be seen in Figure 4.2.1.

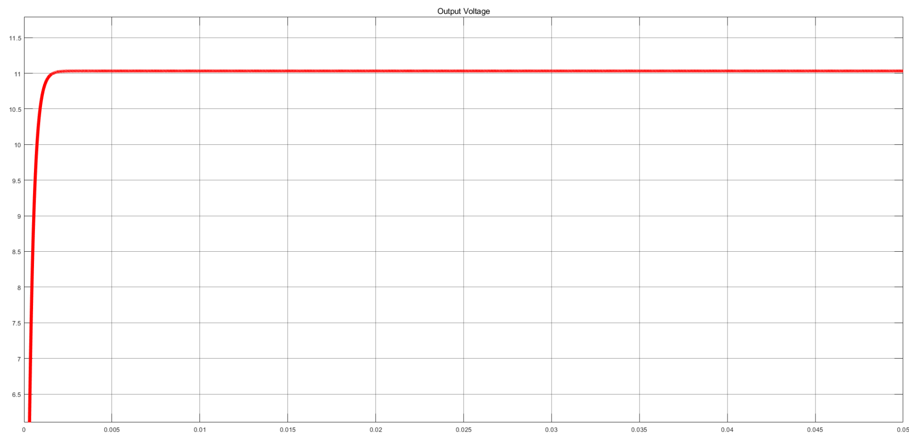

Figure 4.2.1: Output Voltage

V_x voltage is 24V in square wave form, can be seen in Figure 4.2.2.

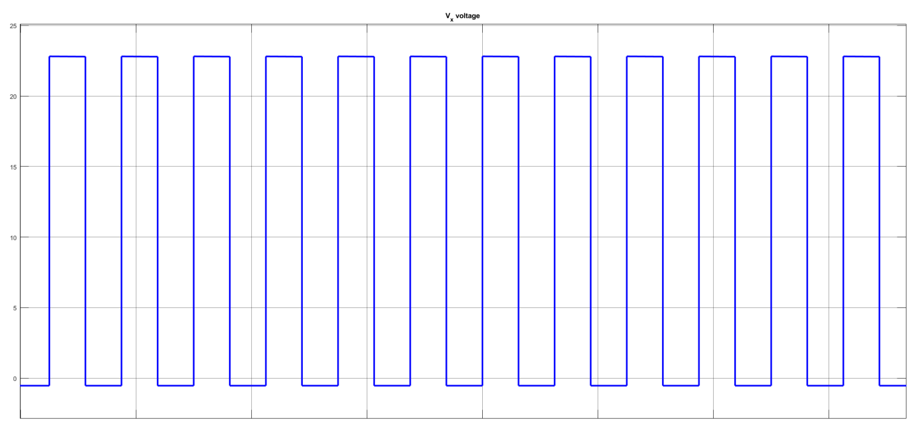

Figure 4.2.2: $\mathrm{V}_x Voltage$

Inductor voltage is swinging between 12V and -12V. Inductor current has an average of 7.5A and ripple value is 0.13A, corresponds to 1.7%, much better than what we aimed for. Voltage and current waveforms of inductor can be seen in Figure 4.2.3.

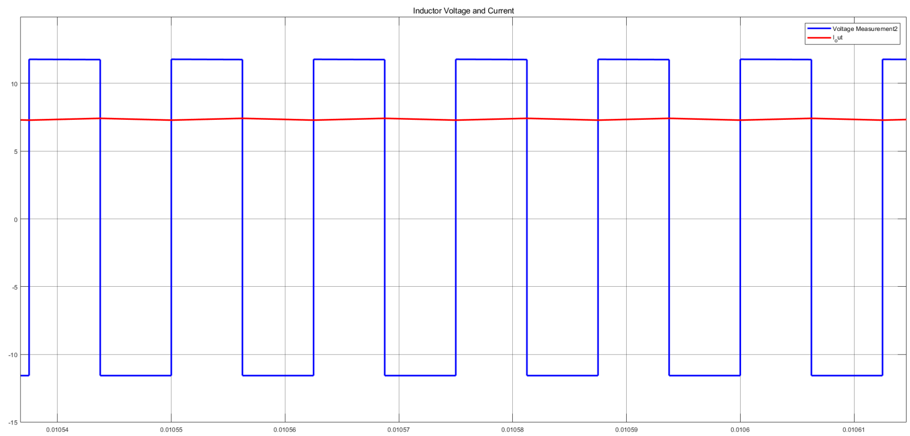

Figure 4.2.3: Inductor Voltage and Current Waveforms

Switching device current and voltage vaweforms can be seen in Figure 4.2.4.

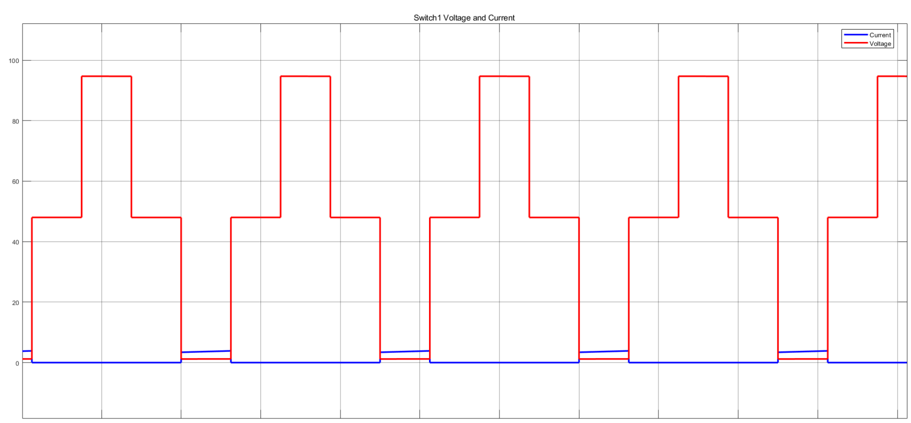

Figure 4.2.4: Switch Voltage and Current Waveforms

4.3 c)

Leakage in the primary side causes large spikes at the output current and voltage waveforms. After some time, simulation gives errors due to open circuited elements in the circuit. Resulting waveforms can be seen in the following figures.

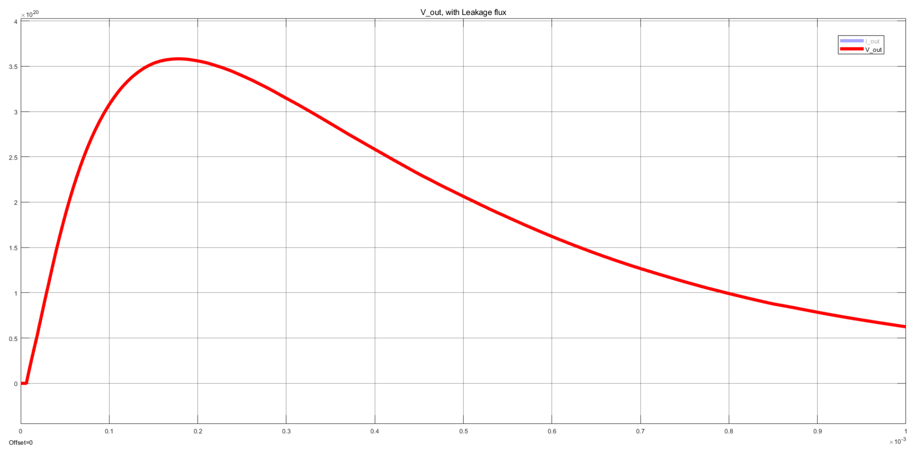

Figure 4.3.1: Output Voltage with Leakage

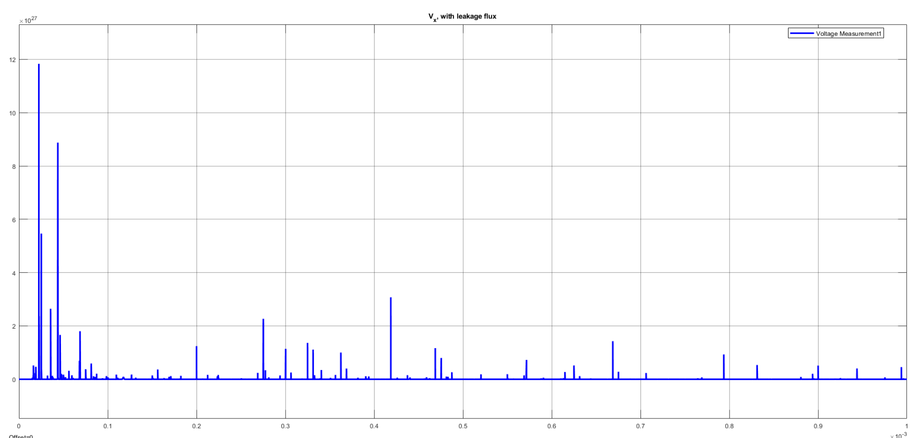

Figure 4.3.2: V$_x withLeakage$

This situation causes a great instability in the system. It can be due to the energy stored in the leakage inductances cannot find a path to flow once the switches are turned off together during the OFF times. A free-wheeling path or a snubber circuitry added to the primary side can solve this problem.

4.4 d)

There are some expected discrepancies stemming from the non-idealities in the circuit but they are not vital for the operation. The voltage drop at the output is around 1 V only and that can be easily solved by adjusting the duty cycle to compensate for this operation. **D=0.27** operation actually does the job and it is a reasonable adjustment.

4.5 e)

During two different half cycles, the output is to be fed by a different voltage due to this inequality. It will result in unbalanced ripples and most importantly, output voltage can be affected.

To compensate for this situation, different duty cycles for the switches can be arranged. A shorter ON duration for the higher turn number pair can solve this problem. But this may trap the magnetizing flux inside the core and end up saturating it. To avoid this, a freewheeling winding can be added to the topology.

Or, primary windings can be changed as well to allow magnetizing flux decrease or increase properly within the limitations.

5 Time Spent

I have spent approximately **20 hours** for this homework, no kidding. For better or worse, now it is like a child to me and I hope you take good care of it...