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University of SAAD DAHLAB BLIDA

كلية التكنولوجيا
Faculty of Technology

قسم الإلكترونيك
Electronics Department



Memoir of Master

Option Electronic
speciality instrumentation

Presented by

KAB Djabri

&

M'ZIANE Mohamed

Control of separately excited DC motor using H-bridge

Supervised by

Mister ALIAN Ahcene

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Chapter 1

Introduction to DC motors

Chapter 2

Power Supply Unit

Chapter 3

Control Unit

Chapter 4

Practical issues and reflections on a test bench realization

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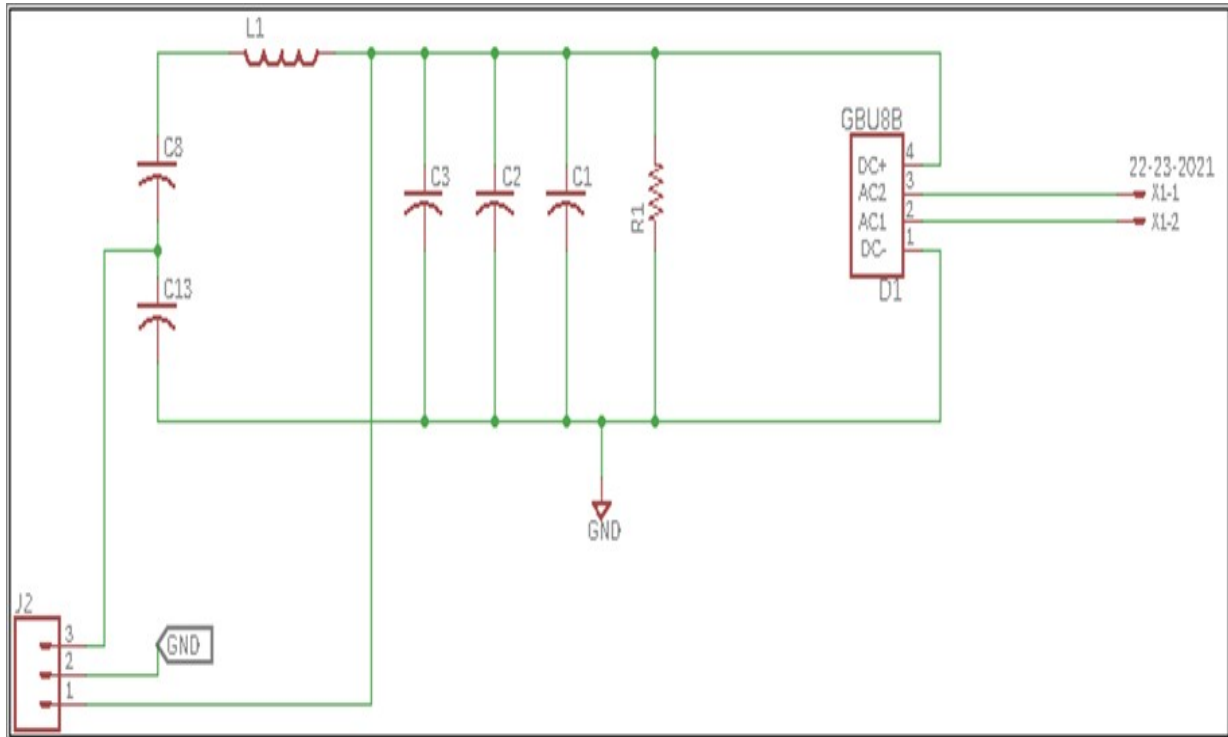
List of Acronyms and Abbreviations

General Introduction

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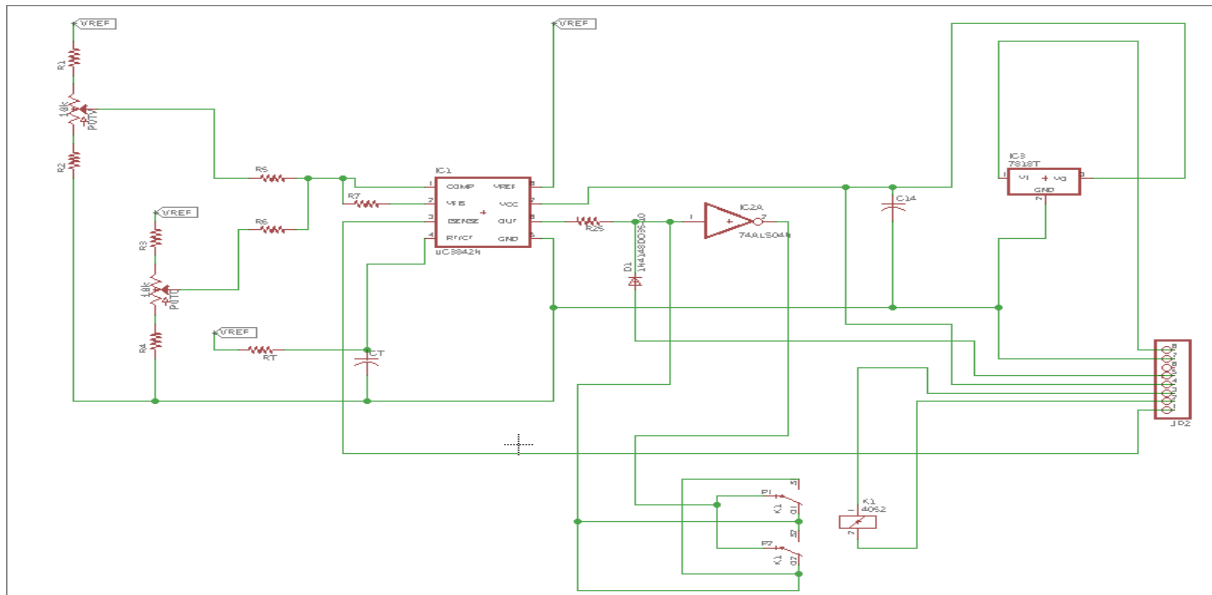
Diagrams and results of the circuits that make the drives



-Supply Circuit Diagram-

Components	Value or reference
Bridge Diode	GBU 12J
Resistor R1	10k Ω
Filter Capacitor	550 μ F
Capacitor C8	100 μ F
Capacitor C13	11 μ F

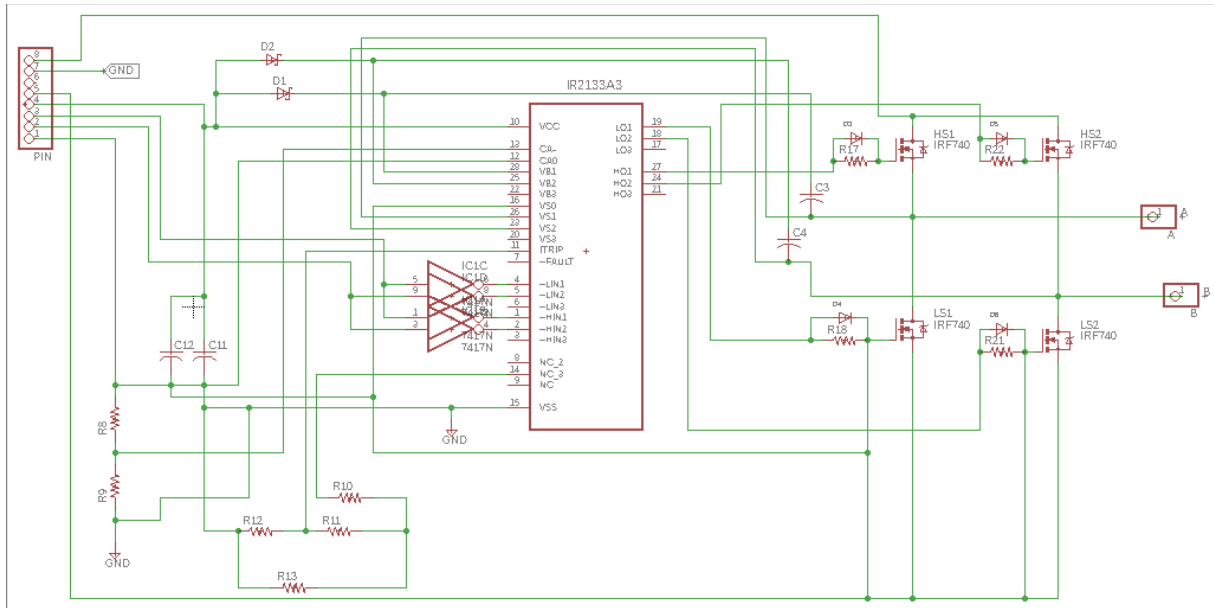
-Table of values and components-



-Control Circuit Diagram-

Components	Value or reference
Speed Potentiometre	10k Ω
Torque Potentiometre	10k Ω
R1	63,9k Ω
R2	8k Ω
R3	8,2k Ω
R4	15k Ω
R5	100k Ω
R6	100k Ω
R7	100k Ω
R _T	12k Ω
C _T	150nF
Integrated circuit	UC3842a
Relay	12v 2RT

-Table of values and Components-



-Power Circuit Diagram-

Components	Value or reference
Integrated circuit	IR2133
Integrated circuit	LS7404
Transistor	IRFP350
R8	10k Ω
R9	10k Ω
R10	5k Ω
R11	13k Ω
R12	100k Ω
R13	0,05 Ω

-Table of values and components-

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- [9] IRFP350, International Rectifier.
- [10] IR2133, International Rectifier.

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In this project, we studied the control of DC motor by taking a previous model of the motor obtained last year and making some modifications, basically in having the possibility to change the sense of speed and also the possibility of braking.

For that purpose, the main work was to calculate and realise three circuits: a PWM card controller based on UC3842A, an H-Bridge and its driver the IR2133 and a power supply mainly based on a full bridge AC-DC converter and a filter capacitor that feed the H-Bridge and the motor.

We basely used the H-bridge that give us the aptitude to rotate the motor in the both direction and the braking also. The choice of the H-Bridge switching elements was made after calculations well specify in chapter two, where we arrived to choose the MOSFETs (IRFP350) among available transistors in the market.

Despite the big harmonic contents of bipolar PWM control and its CEM noises, we've been forced to use it, basically for its great simplicity, the control was based on the integrated circuit UC3842 studied in chapter three and who is a current PWM controller. The fact that the control was based on current variation, was at first very difficult to us to control the motor speed, because if currents ripple depends on duty cycle, then can be a parameter in control duty cycle and then speed, the problem is that this command is possible only if load Torque is constant. This problem led us to introduce another command which is related to the variation of the average armature current and so to the load torque, which mean that by its use the card can be used as speed command and in the same time as a load torque value indicator. At first we worked with a commutation frequency of 10 kHz, but in calculating transistors, we were forced to lower the frequency to 1 KHz, what has made current ripples very high and limited our speed control from $0,25T_{nom}$ to T_{nom} . At the end of chapter 3, we proposed a solution for this problem which consist of amplifying the average armature current when load torque is smaller than $0,25T_{nom}$, and to proportionally attenuate the ripples current by increasing commutation frequency of the PWM controller by a simple use of a double pole throw switch.

Finally, we proposed in chapter four a panel that will be used within our cards to make possible the realisation of a DC motor Bench test and discusses the procedures and elements needed (speed sensor, DC generator) which can help us to calibrate the controller card basing our calibration on a known nominal electromagnetic torque, a known nominal armature current and a known nominal speed. Indeed, this calibration is based on the hypothesis that

General Conclusion

motor parameters don't change¹, and hence the motor under test will be almost the studhorse. For the purpose of calibration, we tried to give procedures to do that, but we realised very quickly the complexity of the work and estimated that this work may be need another project of other students that will continue the work we have not finished.

¹ If it is not the case, because our motors are very old, calibration couldn't not be made without load Torque.

1

1.1 Introduction:

DC motors are widely used in automatic systems that require precise speed variation, in this chapter; we present the construction of the DC machine and their operating principle. It is a question of establishing the different types of DC motors. Then we will give the advantages and disadvantages and finally the field of use.

1.2 Definition:

Direct current machines are electromechanical energy converters:

Either they convert absorbed electrical energy into mechanical energy when they are able to provide sufficient mechanical power to start and then drive a moving load. It is then said that they have an engine operation.

Either they convert the mechanical energy received into electrical energy when they undergo the action of a driving load. Then we say they operate as a generator.

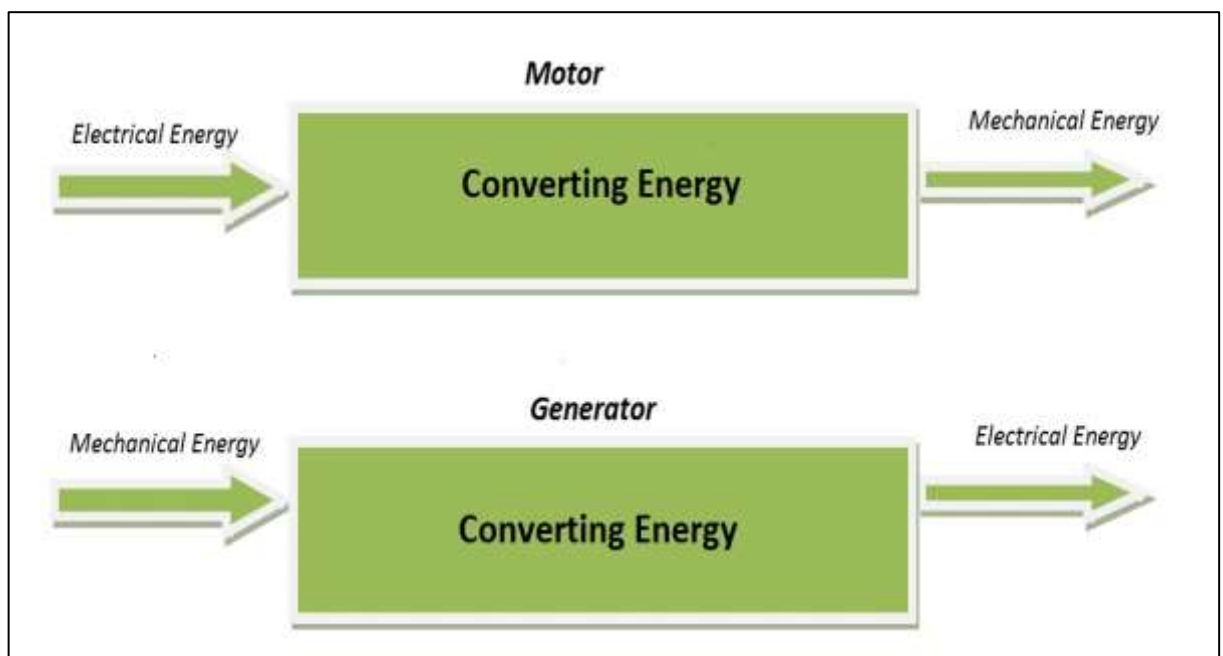


Figure 1.1: DC machine operating system.

1.3 DC motor setup:

The DC motor consists of three main parts:

- Stator
- Rotor
- The commutator/brush device

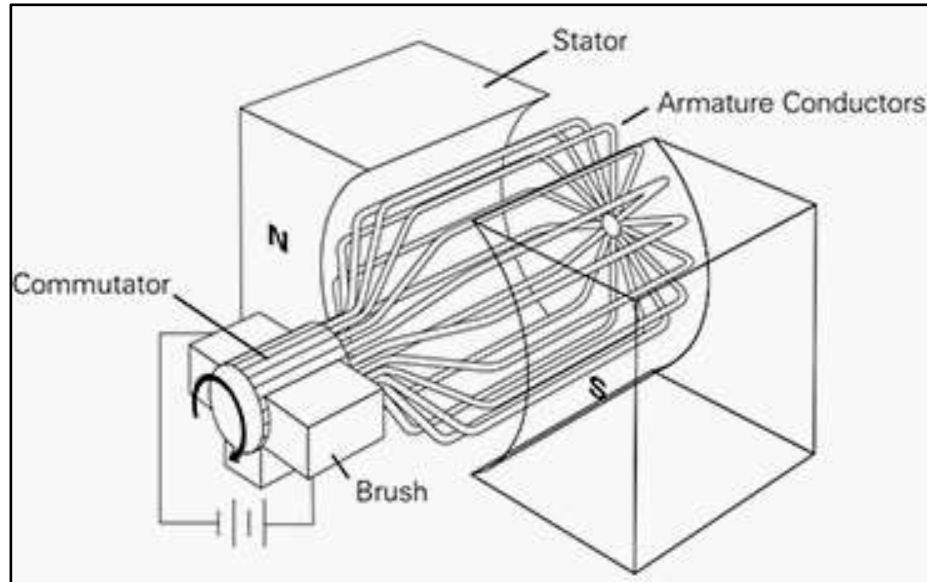


Figure 1.2: internal structure of a DC motor.

1.3.1 Stator:

The stator is the fixed part of the motor. It consists of a permanent magnet or an electromagnet powered by direct current excitation (I_e).

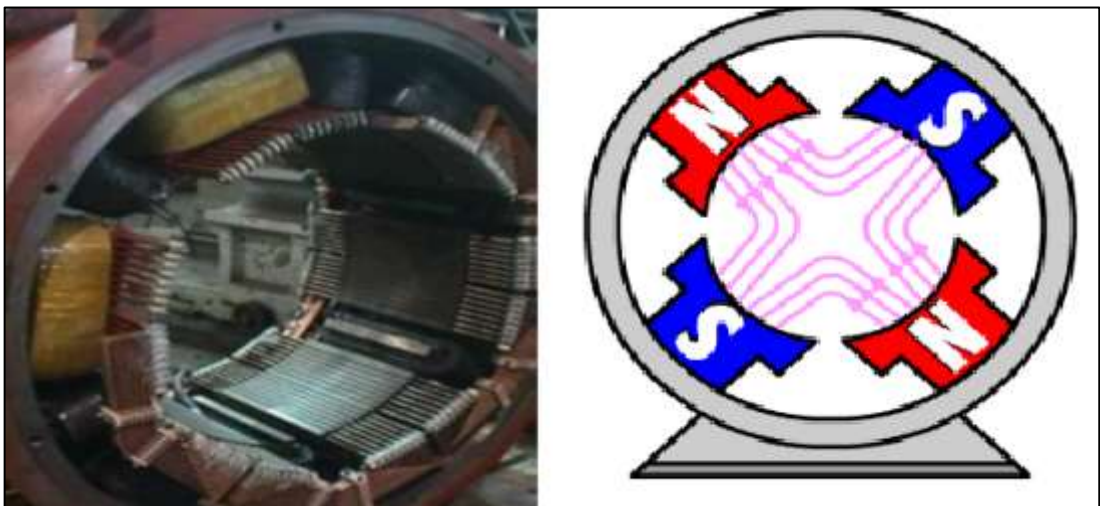


Figure 1.3: stator of DC motor.

1.3.2 Rotor:

The rotor consists of slots in which is wound a winding of conductors (N) supplied with direct current (I) via the commutator.

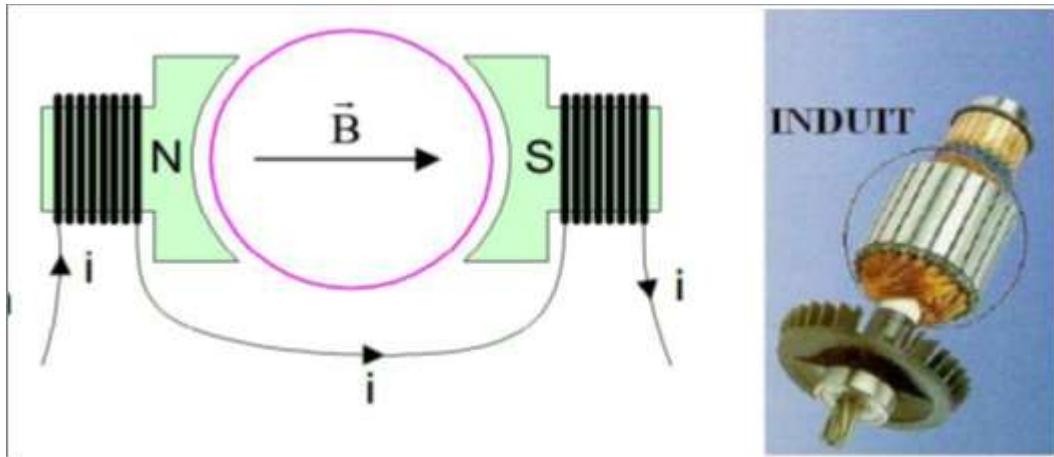


Figure 1.4: The rotor of DC motor.

1.3.3 The commutator and brushes:

To keep the torque on DC motors from reversing every time the coil moves through the plane perpendicular to the magnetic field a split-ring device called a commutator is used to reverse the current at that point.

The electrical contacts to the rotating ring are called brushes since copper brushes were used in early motors. Modern motors normally use spring-loaded carbon contacts, but the historical name for the brushes has persisted.

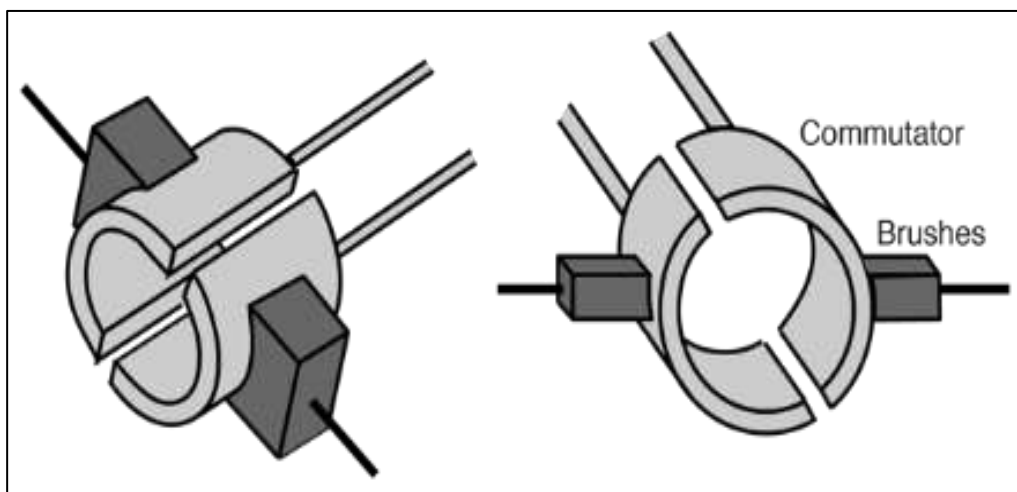


Figure 1.5: The assembly of commutator and brushes.

1.4 Working concept:

The functioning of the DC motor is based on the Laplace force principle:

A conductor with a Length (L), placed in a magnetic field and traversed by a current, is subjected to an electromagnetic force.

The field created by the inductor acts on the conductors of the armature: Each of the (N) conductors of lengths (L) placed in the field (B) and traversed by a current (I) is the seat of an electromagnetic force perpendicular to the conductor

$$F = B \cdot I \cdot L \cdot \sin \alpha \quad (1.1)$$

These Laplace forces exert a torque proportional to current (I) and flux (Φ) on the rotor. The motor starts to rotate at a speed proportional to the supply voltage (V) and inversely proportional to the flow (Φ).

When any armature conductor passes over the neutral line, the current flowing through it changes direction thanks to the collector. The motor maintains the same direction of rotation.

To reverse the direction of rotation of the motor, the direction of the field generated by the inductor with respect to the direction of the current flowing through the armature:

- Either we reverse the polarity of the supply voltage of the armature
- Either we reverse the supply polarity of the excitation circuit

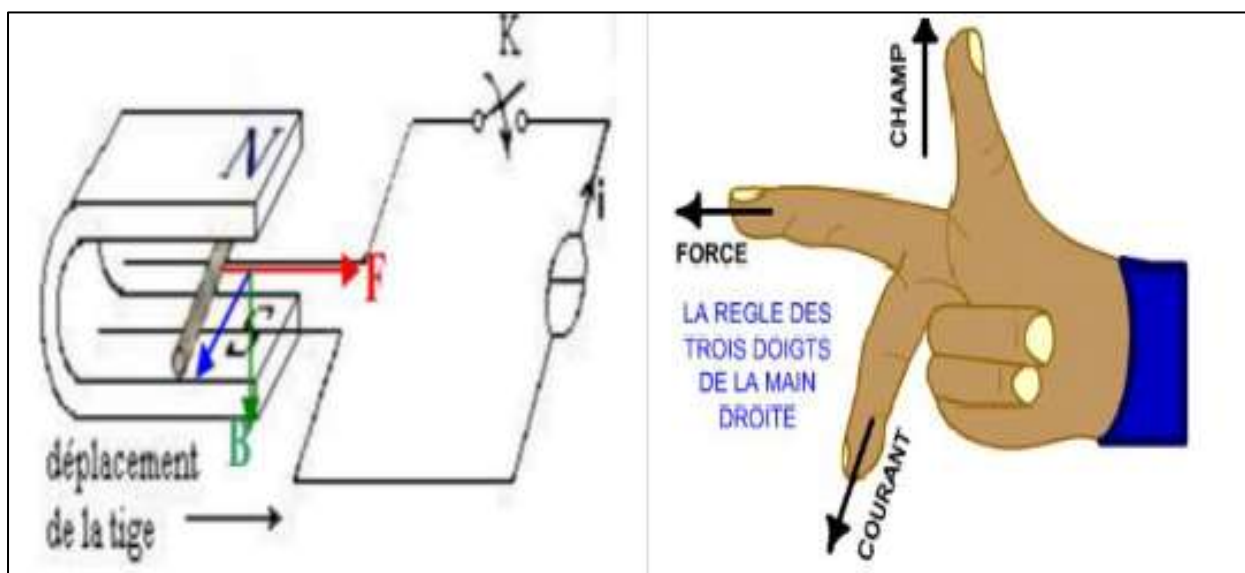


Figure 1.6: working concept of DC motor.

1.5 Counter-electromotive force:

The electromotive force (EMF) is the voltage produced by the rotor during its rotation in the magnetic flux produced by the fixed part. It depends on the construction elements of the machine.

$$EMF = \frac{P}{a} N \cdot n \cdot \Phi \quad (1.2)$$

P: number of pole pairs of the machine.

N: number of active conductors on the armature peripheries.

a: number of winding pairs between the two brushes.

n: armature rotation frequency (in t/s).

Φ : feed under a machine pole in Webers.

Finally we get

$$EMF = K \cdot \Omega \cdot \Phi \quad (1.3)$$

With

$$K = \frac{P}{2\pi a} N \quad (1.4)$$

1.6 Types of motors:

There are two types of DC motors:

1.6.1 Permanent magnet inductor motors:

There is no inductor circuit, the inductor flux is produced by a permanent magnet.

All low power DC motors and micromotors are permanent magnet motors. They now represent the majority of DC motors. They are very simple to use

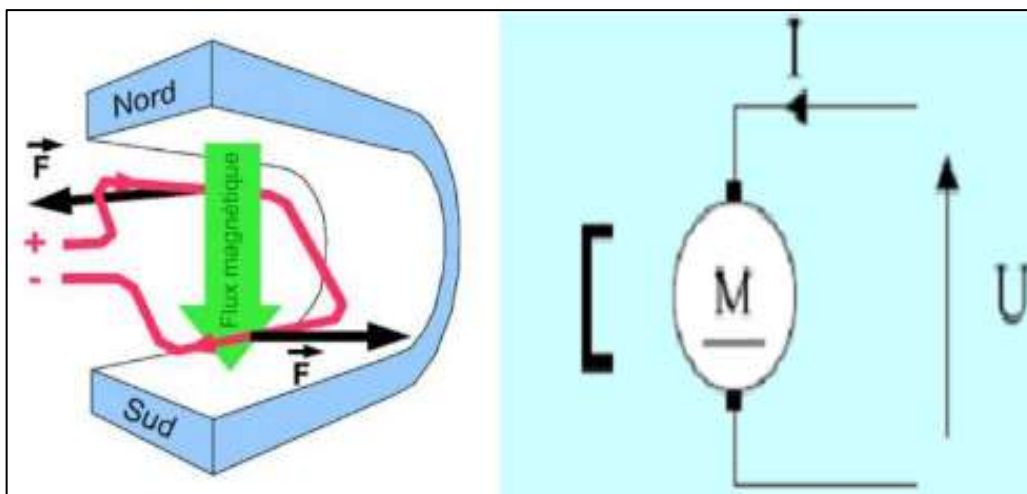


Figure 1.7: Permanent magnet inductor motors.

1.6.2 Inductor-wound motors:

There are 4 different types of electric motors that are classified according to the type of excitation that is used, which are:

- The shunt excitation motor.
- The series excitation motor.
- The compound excitation motor.
- The separately excited motor.

a. The shunt excitation motor:

In a shunt motor the field is connected in parallel (shunt) with the armature windings, the shunt-connected motor offers good speed regulation. The field winding can be separately excited or connected to the same source as the armature.

The shunt-connected motor offers simplified control for reversing; this is especially beneficial in regenerative drives

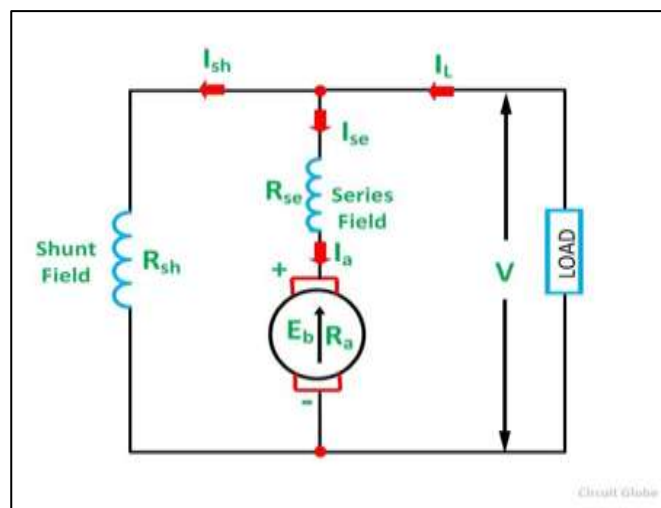


Figure 1.8: Shunt Wound DC Motor.

b. The series excitation motor:

In a series DC motor the field is connected in series with the armature, the field is wound with a few turns of large wire because it must carry the full armature current.

A characteristic of series motors is that the motor develops a large amount of starting torque. However, speed varies widely between no loads and fully loaded. Series motors cannot be used where a constant speed is required under varying loads.

Series connected motors generally are not suitable for use on most variable speed drive applications.

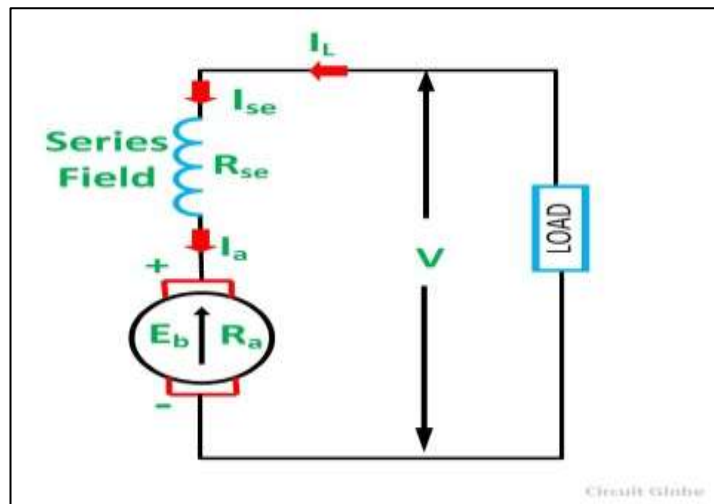


Figure 1.9: Series Wound Motor.

c. The compound excitation motor:

Compound motors have a field connected in series with the armature and a separately excited shunt field, the series field provides better starting torque and the shunt field provides better speed regulation

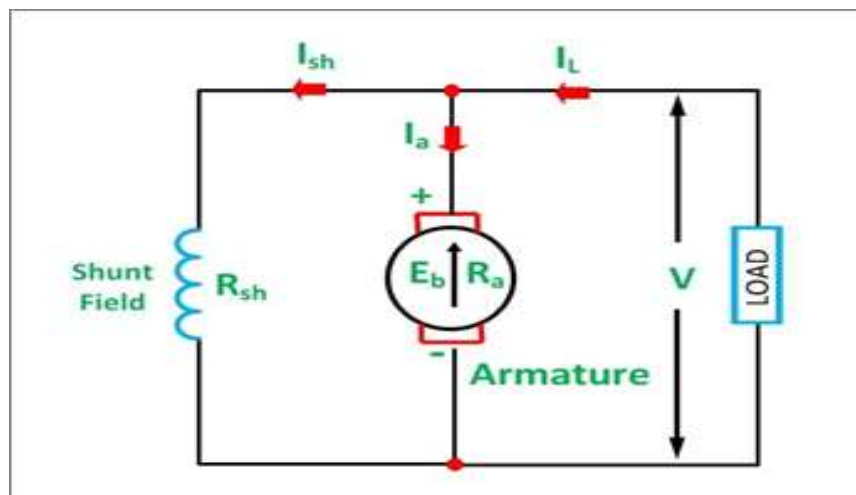


Figure 1.10: The compound excitation motor.

d. The separately excited motor:

In a separate or independent excitation motor, the excitation circuit is separated from the armature circuit. If the inductor is a permanent magnet, the flux (Φ) is constant. If the inductor is an electromagnet powered by an adjustable DC voltage source, the flux (Φ) depends only on the current in the inductor called the excitation current (I_e).

The current creates a field and a quantity of field through a coil gives a flow, if the voltage (V_e) is constant, the excitation current (I_e) is constant and the flux (Φ) is constant. Under these conditions, the counter-electromotive force (E) depends only on the frequency (n) of rotation.

Therefore the frequency of rotation (n) is proportional to the supply voltage (V) of the motor.

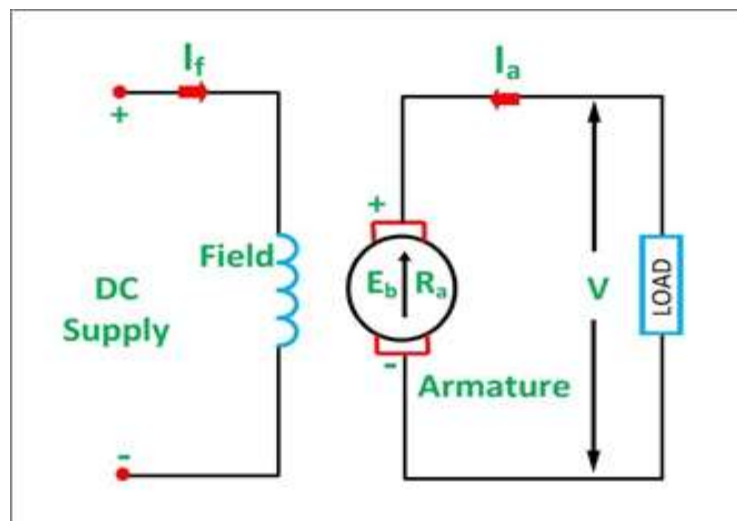


Figure 1.11: separately excited motor.

1.7 Speed variation:

To vary the speed of a DC motor, one can act on the voltage at the armature terminals; the armature voltage is directly proportional to the rotational speed. The power varies but the torque remains constant, this is called constant torque speed variation.

1.8 Power balance and Efficiency:**1.8.1 Power balance:**

The balance of the powers involved in a DC motor in rated operation can be represented by an arrow which shrinks as the power decreases.

The **figure 1.12** shows the powers and losses in the conversion of a DC motor.

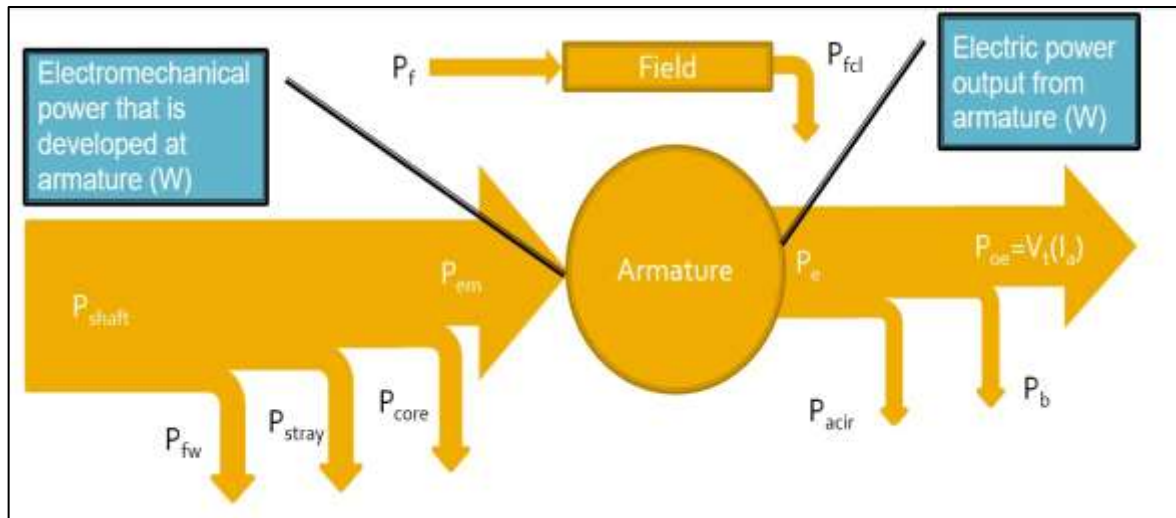


Figure 1.12: power balance representation.

- $P_{e,in}$ = electric power in at terminals (W)
- P_{acir} = armature circuit losses $I_a^2 (R_{acir})$ (W)
- P_b = losses due to brush drop $V_b (I_a)$ (W)
- P_e = electric power delivered to armature $(E_a) \cdot I_a$ circuit (W)
- P_{em} = Electromechanical power developed in armature $(T_d) \cdot \omega$ (W)
- P_{fw} = friction and windage losses (W) (from test)
- P_{stray} = stray load losses (W) (from test)
- P_{core} = core losses (W) (from test)
- $P_{shaft}(P_{out})$ = total mechanical power develop at the shaft (Rated hp)

In terms of losses

$$P_{losses} = P_{acir} + P_b + P_{fw} + P_{core} + P_{stray} \quad (1.5)$$

Input power

$$P_{in} = V \times I \quad (1.6)$$

With:

- I = current, measured in amperes (A)
- V = applied voltage, measured in volts (v)

Output power

$$P_{shaft} = P_{in} - P_{losses} \quad (1.7)$$

1.8.2 Efficiency:

Direct current motors consume part of the energy absorbed for their operation. The mechanical energy supplied will always be smaller than the electrical energy absorbed. The ratio between the energy supplied and the energy absorbed is the efficiency.

$$\eta = \frac{P_{out}}{P_{in}} \times 100\% \quad (1.5)$$

1.9 Conclusion:

This chapter has made it possible to recall the various elements that make up a direct current machine and the operating principle. After our study we found that the separately-excited motors with a perishing magnet are the most suitable for speed variation. In the following chapter, we will study the static converters.

2

2.1 Introduction :

The objective of this chapter is to identify and calculate the elements of the power supply unit of our project, which is constituted by an AC/DC converter (rectifier) and an H-bridge. In the end of this chapter we will be able to choose the elements that consists the H-bridge, and the elements that ensure its correctly and safely operating.

The **figure 2.1** shows a simple diagram bloc of our project,

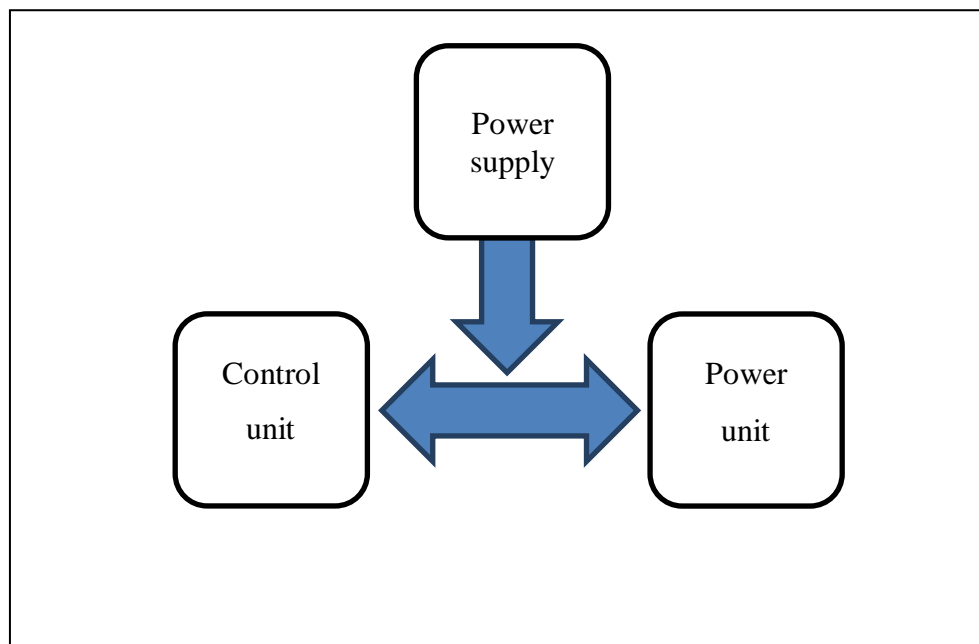


Figure 2.1: bloc diagram.

2.2 Supply calculating:

A power supply is an element that supplies power to at least one electric load. Typically, it converts one type of electrical power to another, the objective of the power supply in this project is to feed the motor by a DC voltage, and this purpose is realized by a simple full wave bridge rectifier with a filter capacitor.

2.2.1 Choosing the rectifier:

Depending on the characteristics measured from the nameplate of our motor, we can choose the rectifier that could suit our project.

The motor has a nominal voltage of 220V and an $I_{nominal}$ current of 6.8A, so we use a rectifier bridge with a V_{RRM} voltage higher than 310V to take account of power surges in the sector, and a maximum average current I_f higher than 6.8 A. The suitable bridge under these conditions is the rectifier **GBU 12M** characterized by a voltage $V_{RRM}=700V$ and a maximum average current $I_f=12A$.

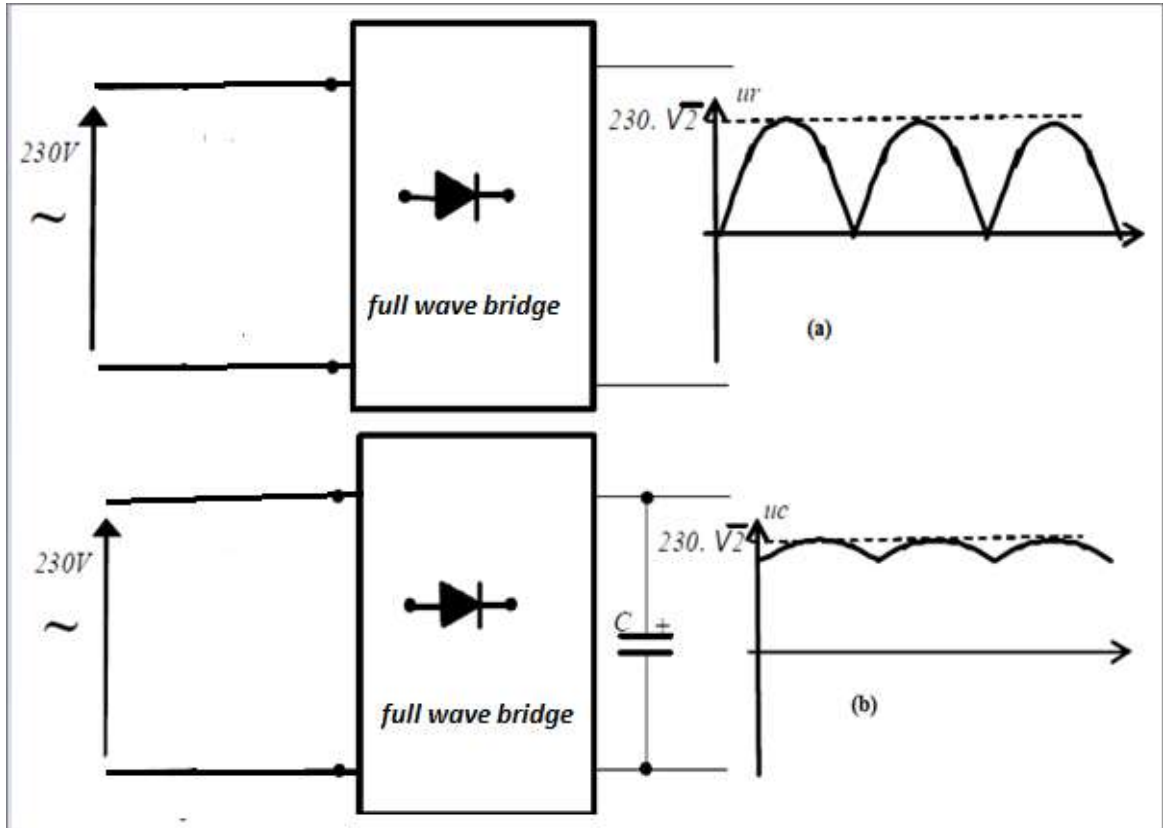


Figure 2.2(a): without filter capacitor. (b): with filter capacitor.

2.2.2 Calculating the filter capacitor:

The capacitor is associated with the rectifier, so that the input voltage varies between two values V_{dmin} and V_{dmax} , as shown in figure 2.3. Based on 220V and 6.8A motor characteristics, the power absorbed by the motor: $P = 220 \times 6.8 = 1496W \approx 1500W$.

If the ripple rate¹ is 25%, knowing that $V_{dmax}=310V$, we get: $V_{dmin} = V_{dmax} - 25\% V_{dmax}$, therefore $V_{dmin}=232,5 V$.

The voltage across the filter capacitor varies between $V_{dmax}=310V$ and $V_{dmin}=232.5V$, the energy stored, then evacuated to the motor is described by:

¹And with using a PWM controller that is almost needed in the motor command, we will demonstrate that ripples, no more will, influence the motor operation

² Because until the current drain has not reached its maximum value, the diode is still ON, because the

$$\Delta W = \frac{1}{2} C V_{dmax}^2 - \frac{1}{2} C V_{dmin}^2 \quad (2.1)$$

The average power supplied to the engine is:

$$P_{moy} = \frac{1}{2} C (V_{dmax}^2 - V_{dmin}^2) / \left(\frac{T}{2} - T_a\right) \quad (2.2)$$

With T_a is the charge time of the capacitor.

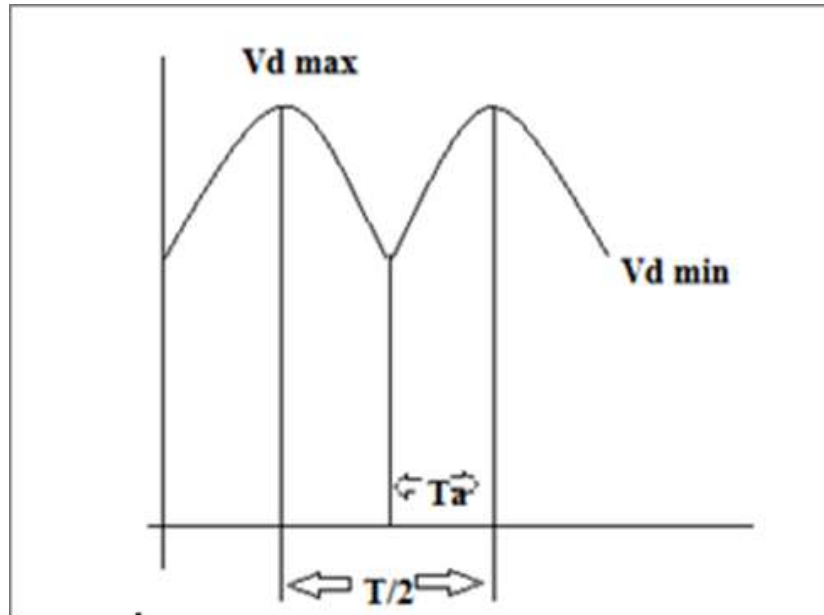


Figure 2.3: Chronogram of voltage ripples across the filter capacitor.

From the **figure 2.3** above:

$$V_{dmin} = V_{dmax} \sin\left(\frac{\pi}{2} - \frac{2\pi}{T}T_a\right) = V_{dmax} \cos\frac{2\pi}{T}T_a \quad (2.3)$$

Therefore:

$$\frac{2\pi}{T}T_a = \arccos\frac{V_{dmin}}{V_{dmax}} = \frac{232.5}{310} = 0.72 \text{ rad} \quad \text{so we get } T_a = 2,3\text{ms.}$$

Knowing that:

$$P_{moy} = 1500\text{W}, C = [2 \cdot P_{moy} / (V_{dmax}^2 - V_{dmin}^2)] \left(\frac{T}{2} - T_a\right) = 546\mu\text{F}, \text{ so } C = 550\mu\text{F.}$$

Our choice is a 550 μ F capacitor which should bear a voltage greater than 310V, we take C=550 μ F (400V).

2.3 The H-bridge :

In general an H-bridge is a rather simple circuit, containing four switching elements, with the load at the center, in an H-like configuration as shown the **figure 2.4**:

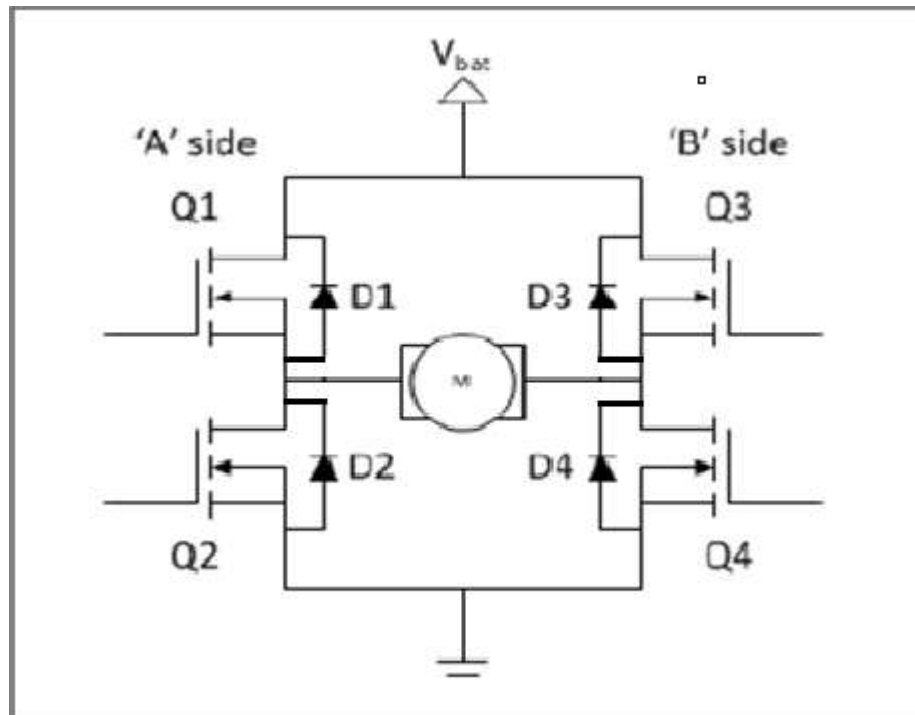


Figure 2.4: H-bridge configuration.

2.3.1 Circuit description:

The H-bridge is constituted by a four switching element (Q1, Q2, Q3 and Q4), these element are usually bi-polar or FET transistors, in some high-voltage applications or IGBTs (Insulated Gate Bipolar Transistor). In this project we are going to use the MOSFETs (Metal Oxide Semiconductor Field Effect Transistor) which have already body's diode, but we also can add diode in parallel to each one of the transistors to improve the switching.

The top-end of the bridge is connected to a power supply and the bottom-end is grounded and every switching element can be controlled independently.

2.3.2 Operating of the H-bridge:

The basic operating mode of an H-bridge is fairly simple: if Q1 and Q4 are turned on, the left lead of the motor will be connected to the power supply, while the right lead is connected to ground. Current starts flowing through the motor which energizes the motor in the forward direction and the motor shaft starts spinning. And if Q2 and Q3 are turned on, the reverse will happen, the motor gets energized in the reverse direction, and the shaft will start spinning backwards.

2.3.3 Controlling the H-bridge:

To control the h-bridge we need a driver that gives an order to the switching element to turn-on or off. In this work that driver will be the IR2133.

The IR2133 is a high voltage, high speed power MOSFET and IGBT driver with three independent high side and low side referenced output channels for 3-phase applications. In the case of our H-bridge we need two high side and low side drivers.

The driver IR2133 is characterized by a fixed dead time DT , which its minimum value is 100ns, maximum value is 400ns and the typical value is 250ns, which means that when one of the transistors (High Side or Low Side) is controlled to turn-on, it will be turn on only a dead time Dt after the other transistor of the same arm is controlled to turn off. Without this dead time precaution, it is impossible to operate the H-bridge, because any switching involves a short circuit in the S1 arm in the case of rotation in the positive direction and a short circuit in the S2 arm in the case of rotation of the motor in the negative direction.

2.3.4 Working concept:

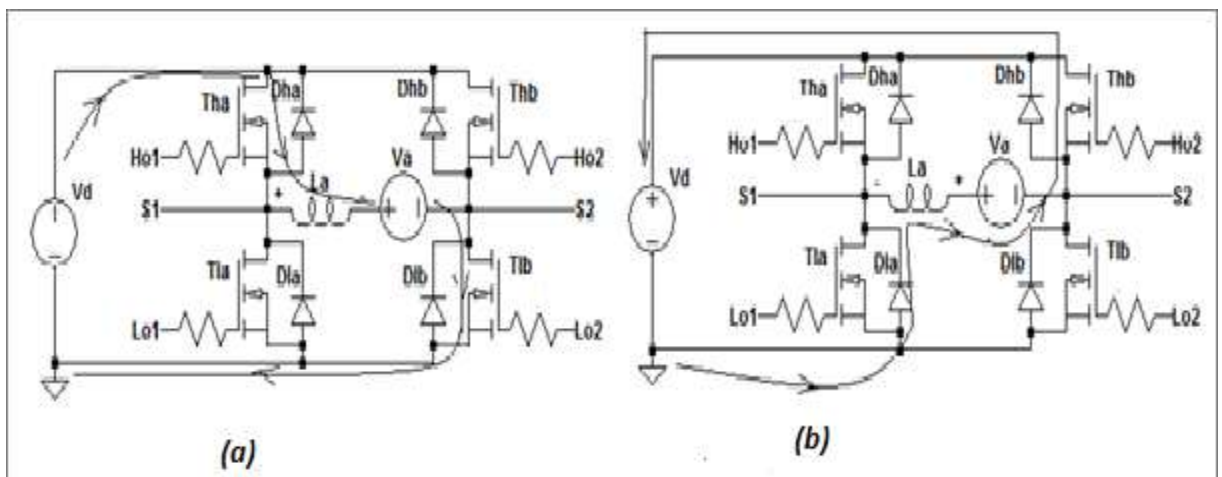


Figure 2.5: (a) charging path (b) discharge path.

The current path in the transistors and in the diodes parallel to them depends on the direction of the average current and this depends on the direction of rotation of the motor. Thus for a positive direction of rotation and therefore a positive average current flowing from + to - of the armature's motor, the load current of the armature flows through the transistors THS1 and TLS2 as shown in the **figure 2.5(a)**. On the other hand the discharge current flows through the diodes DLS1 and DHS2 as shown in the **figure 2.5(b)**, since the current is always positive, it cannot flow from the source to the drain in the transistors TLS1 and THS2 only in

the case of braking. For reverse rotation, however, the load current will flow through the THS2 and TLS1 transistors and the discharge current will flow through the DHS1 and DLS2 diodes.

2.3.5 Transistor and diode switching:

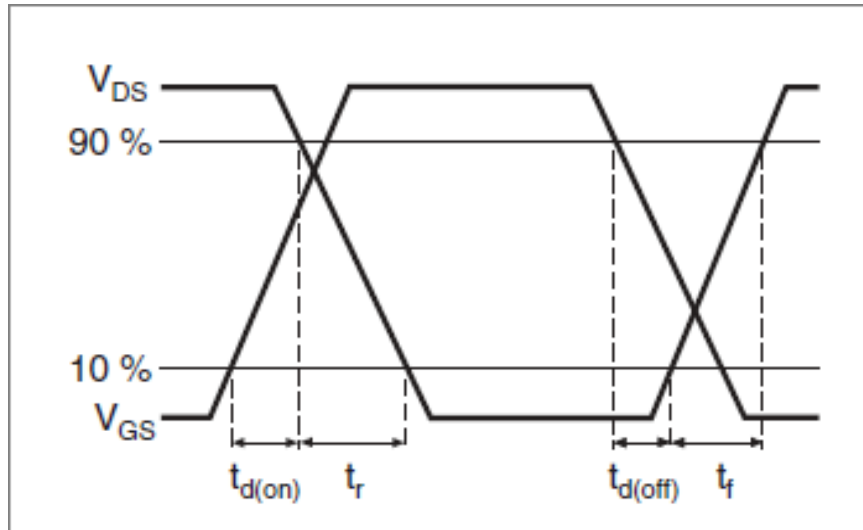


Figure 2.6: Switching Time Waveforms with resistive charge.

a. Mosfet turn on characteristic :

When the gate driver delivers the pulses to the gates of the high side Ho1 and low side Lo2 transistors, the transistors do not close instantly, but goes through different phases:

Delay time t_{di} : time duration which neither the voltage nor the current in the transistor changes. This time corresponds to the duration corresponding to the load of the capacity C_{gs} to reach the threshold voltage V_{gth} , corresponding to the opening of the inversion channel between drain and source.

Drain current rise time t_{ri} : Theoretically, this time is defined as the time required for the drain current to reach (90%) of the final value while V_{ds} decrease, but this is in the case of a resistive load. In our case, with an inductive charge things are a little bit more complicated. Moreover, in the literature the calculation of switching times is always undertaken by considering the driver as a voltage source and therefore the switching time is dictated by the charging of the capacitor C_{gs} through the gate resistor via a varying current. In our case, things are different, because the IR2133 driver generates a fixed current pulse of 0.2A for the ON state and a negative current pulse of -0.5A for the OFF state, which means

that the capacitor C_{gs} is charged under constant current. By neglecting the parasitic inductances of the connections between driver and gate and between gate and source, we can establish that the current in the drain reaches the charging current of the motor armature, as soon as the voltage V_{gs} reaches the voltage V_{gsp} given that: $I_g = C_{gs} \frac{dV_{gs}}{dt}$, so we have a linear evolution of the voltage V_{gs} ($V_{gs} = \frac{I_g}{C_{gs}} t$), we deduce that the t_{ri} can be described by:

$$t_{ri} = \frac{C_{gs} \times (V_{gsp} - V_{gth})}{I_{gon}} \quad (2.4)$$

$$\text{This gives us: } t_{ri} = \frac{(Q_{gsp} - Q_{gth})}{I_{gon}} \quad (2.5)$$

Note that this expression does not take into account the effective parasitic inductance L_s of the Mosfet and the much higher parasitic inductance [5] that can have the printed circuit track between pin S1 of the driver and the source of the high side transistor or that between pin **com** and the source of the low side transistor. As soon as the drain current increases, it creates an induced voltage at the level of the parasitic induction, thus decreasing the voltage V_{gs} and slowing down the current increase, which means that the t_{ri} is likely to be much greater than that given by the equation (2.5).

So we will analyse the effect that parasitic inductances can have on t_{ri} . The driver can be represented by a voltage source V_{cc} which generates a constant current $I_g = \frac{V_{drive}}{R_{drive}} = 0.2A$ which loads the transistor capacitance (C_{gs}). However, as soon as the voltage V_{gs} reaches V_{gth} , the drain current begins to flow and generates an induced voltage $L \frac{dI_D}{dt}$ which opposes the charging and acts as if the charging voltage becomes the sum of the voltage accumulated at the terminals of G and that at the terminals of the parasitic inductors. During the charging of the capacitance C_{gs} , and before that the capacitance C_{gs} reaches the value of V_{gsp} (voltage which will reach a plateau, when the current drain has reached its maximum value because the transistor is still in the active region and so proportional to the current), the transistor is still in its active region since the voltage V_{ds} has not yet started to decrease².

² Because until the current drain has not reached its maximum value, the diode is still ON, because the inductor current is almost constant within switching time, so V_{ds} cannot change until the diode is OFF.

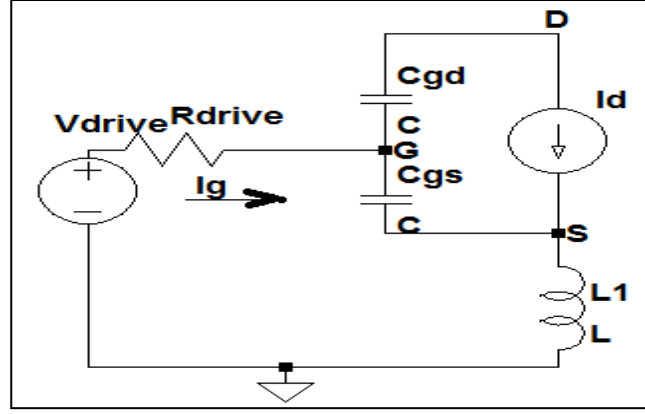


Figure 2.7: Equivalent circuit.

From the **figure2.7** above, we can write:

$$I_g = \frac{V_{drive}}{R_{drive}} - \frac{L_p}{R_{drive}} \frac{dI_D}{dt}, \text{ where } L_p = L_S + L_{parasitic} \quad (2.6)$$

We know that the capacitor C_{gs} is charged under the current I_g , therefore:

$$I_g = C_{gs} \frac{dV_{gs}}{dt}$$

With the transistor still in the active region, we have

$$I_D = K(V_{gs} - V_{gth})^2 \quad (2.7)$$

This gives us:

$$L_p \frac{dI_D}{dt} = 2K(V_{gs} - V_{gth}) \frac{dV_{gs}}{dt} \quad (2.8)$$

By replacing the equation (2.8) in (2.6), it gives us the following equation:

$$I_g = C_{gs} \frac{dV_{gs}}{dt} = \frac{V_{drive}}{R_{drive}} - \frac{2KL_p}{R_{drive}} (V_{gs} - V_{gth}) \frac{dV_{gs}}{dt} \quad (2.9)$$

And this gives us:

$$\frac{dV_{gs}}{dt} = \frac{\frac{V_{drive}}{R_{drive}}}{C_{gs} + \frac{2KL_p(V_{gs} - V_{gth})}{R_{drive}}} \quad (2.10)$$

so:

$$(R_{drive}C_{gs} + 2KL_p(V_{gs} - V_{gth}))dV_{gs} = V_{drive}dt \quad (2.11)$$

The drain current rise time t_{ri} , now, is the time when the voltage V_{gs} changes from the V_{gth} value to the V_{gsp} value, and finally by integrating the equation (2.11) we get:

$$\begin{aligned} t_{ri} &= \int_{V_{gth}}^{V_{gsp}} \frac{(R_{drive}C_{gs} + 2KL_p(V_{gs} - V_{gth}))dV_{gs}}{V_{drive}} \\ &= \frac{(R_{drive}C_{gs} - 2KL_p)}{V_{drive}} (V_{gsp} - V_{gth}) + \frac{KL_p(V_{gsp}^2 - V_{gth}^2)}{V_{drive}} \end{aligned} \quad (2.12)$$

Fall time t_{fv} : it is only at the end of the final opening of the diodes, that the voltage V_{ds} starts to decrease for the two transistors. The drain current having already reached its final value, the capacitor C_{gs} cannot be charged moreover because its voltage has reached a plateau voltage V_{gsp} , which means that the gate current, now charges the capacitor C_{gd} and this is what explains the decrease of the voltage V_{ds} . Any decrease in V_{ds} is expressed by a decrease in V_{dg} or vice versa of an increase in V_{gd} . Since the voltage V_{gs} does not change any more, this means that the decrease of V_{ds} is done at the rhythm of a charging of the capacitor C_{gd} which is done under constant current:

$$I_{gon} = \frac{V_{drive}}{R_{drive}} = 0,2A$$

Knowing the load corresponding to the plateau (given in the datasheet with Miller's load name), during which the voltage V_{gs} remains constant, thus load at the end of which the voltage V_{ds} passes from V_d to V_{dson} , we can calculate very easily the time t_{fv} :

$$t_{fv} = \frac{Q_{gmiller}}{I_{gon}} = \frac{Q_{gd}}{I_{gon}} \quad (2.13)$$

b. Mosfet turn off characteristic:

The opening time t_{doff} : When the transistors THS1 and TLS2 are controlled to switch off, the capacitance C_{gd} is discharged at constant current $I_{goff} = 0.5A$, so that the voltage V_g passes from V_{CC} to the board voltage V_{gsp} . This time is the opening delay time and corresponds to:

$$C_{gd} \frac{dV_{gd}}{dt} = I_{goff} = -0,5A \quad (2.14)$$

In this case, we deduce the evolution of $V_{gd} = \frac{I_{goff}}{C_{gd}} t$, hence:

$$t_{doff} = \frac{(V_{CC} - V_{gsp})}{I_{goff}} C_{gd} = \frac{Q_g - Q_{gd}}{I_{goff}} \quad (2.15)$$

Note in this equation that the C_{gd} value corresponds to the maximum value since the voltage V_{ds} is small and equal to V_{dson} .

The V_{ds} Voltage rise time t_{rv} : from the moment that the voltage V_{gd} reaches the V_{gsp} value, the discharge of the capacity C_{gd} continues, but this time the discharge is expressed by an increase of the drain voltage, until the moment when the voltage V_{ds} becomes equal to V_{gs} . When the capacitor C_{gd} will be completely discharged the presence of the reverse current I_{goff} is expressed by a negative ionization on the side of the gate and positive on the side of the

drain and the voltage V_{ds} will thus continue its increase until reaching the voltage V_d . we can calculate the t_{rv} by:

$$t_{rv} = \frac{Q_{gd}}{I_{goff}} \quad (2.16)$$

The fall time in the drain t_{fi} : t_{fi} is the time it takes the drain current to be cancelled.

And it can be calculated by:

$$t_{fi} = \frac{Q_{gsp} - Q_{gth}}{I_{goff}} \quad (2.17)$$

2.4 Choose and calculate the Mosfet:

The calculation of Mosfet transistors consists in calculating the switching times of Mosfet transistors during their turning ON and OFF. The purpose of this calculation is also to determinate the power dissipated P_{on} , and the switching powers in closing and opening, which make it possible to calculate the suitable heat sink that will ensure the safely operation of the transistors. Before to do this calculation, however, we must select the transistors carefully. For that, we will base the criteria of our choice on several factors:

- The maximum voltage that must be blocked by the transistor and must therefore be higher than 300V.
- The maximum DC current, which corresponds to the rated armature current of 7A of the motor.
- A low resistance R_{on} that ensures a low power P_{on} .
- A low reverse recovery time (t_{rr}) of the transistor body diode, which allows us to reduce current peaks, at the time of transistor closures.
- The price.

Of the Mosfet transistors we have found on the market, the best on terms of features is the IRFP350, that has a max $R_{ds(on)} = 0,3\Omega$, $V_{DSS} = 400V$ and $I_D = 16A$, knowing that the current of our motor is $I_D = 6,8A$, the IRFP350 can dissipate a minimum power $P_{ON} = 14W$.

2.4.1 Mosfet calculating

a. The time tri

Before we calculate this time, we have to find the coefficient value K , its value it can be calculated by the equation (2.7), by consulting the datasheet of IRFP350 transistor, we have a maximum $V_{gth} = 4V$, and a minimum $= 2V$, we will take the intermediate value of $3V$, also from the datasheet we have $V_{gsp} = 4,6V$, which gives:

$$k = \frac{I_D}{(V_{gsp} - V_{gth})^2} = \frac{6,8}{(4,6 - 3)^2} = 2,656A/V^2$$

Now by applying equation (2.12) and considering the inductance L_S of the IRFP350 (13nH) and assuming an additional parasitic inductance of the track L (12nH), so L_P (25nH):

$$t_{ri} = \frac{87 \times 2.3n - 2 \times 2.656 \times 25n}{17.4} 1.6 + \frac{2.656 \times 25n(4.6^2 - 9)}{17.4} = 53ns$$

When switching off, The diode body is characterized by a reverse recovery charge, that until it hasn't evacuated let the diode on during a time t_s , conducting a reverse current that reach a maximal value I_{rr} , so that $Q_{rr} = \frac{1}{2} I_{rr} t_s$ with I_{rr} is the maximum reverse current in the diode, and $I_{rr} = t_s \frac{di}{dt}$ where di/dt is the current rate imposed by the closing of the transistor, which gives $\frac{di}{dt} = \frac{I_{D0}}{t_{ri}} = \frac{6.8A}{53ns} = 128A/\mu s$. The Q_{rr} given in the datasheet corresponds to a di/dt of $100A/\mu s$ and a direct current of 16A. Therefore, that infer that natural recombination is less important than that corresponding to $\frac{di}{dt}$ of $100A/\mu s$ from the datasheet, but also that Q_{rr} is smaller than that given by the datasheet, because the charge which was accumulated in the diode is proportional to the current. Considering this, we can write that our $Q_{rr} = Q_{rr\text{datasheet}} \frac{6,8}{16} \times \frac{128}{100} = 2.56nC$ (and at most 3.84nC if we take in account the maximum value of Q_{rr} and not the typical value). Replacing the expression of t_s in the function of I_{rr} , we obtain:

$$I_{rr} = \sqrt{2Q_{rr} \frac{di}{dt}} = \sqrt{2 \times 2.56\mu C \times 128A/\mu s} = 26A, \text{ and at most } I_{rr} = 31,3A.$$

From the datasheet, we get also the maximal reverse recovery time:

$$t_{rr} = 280ns$$

Now, we get $I_{D\max} = 31.3 + 6.8 = 38.1A$, a small value in front of $I_{D\max}$ that can be supported by the IRFP350, which is 60A.

b. The time t_{fv} :

Considering the **figure 2.8** from the datasheet of IRFP350, that gives the gate charge (Q_g) versus the gate to source voltage (V_{gs}), we deduce the Q_{gd} value equal to 53nC,

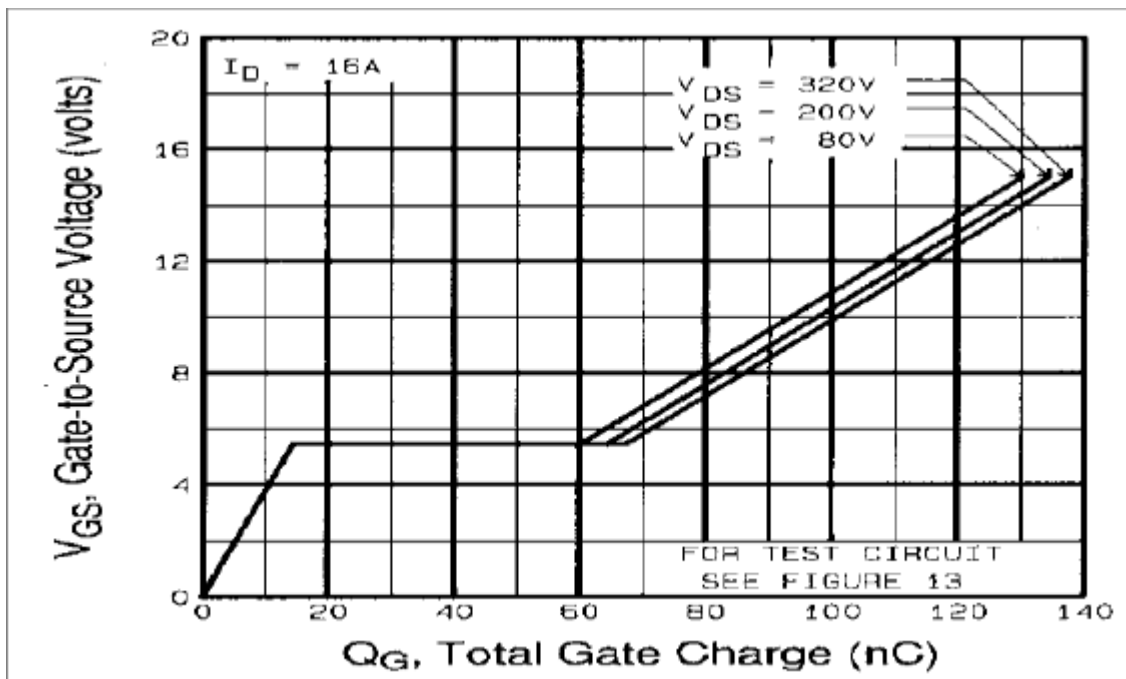


Figure 2.8: typical gate charges versus gate to source voltage.

By applying equation (2.13) we can easily get:

$$t_{fv} = \frac{Q_{gd}}{I_{gon}} = \frac{53n}{0,2} = 265ns$$

c. The time t_{rv}

It can be calculated by applying equation (2.16), which gives:

$$t_{rv} = \frac{Q_{gd}}{I_{goff}} = \frac{53n}{0,5} = 106ns$$

d. The time t_{fi}

By applying equation 2.17, we find:

$$t_{fi} = \frac{Q_{gsp} - Q_{gth}}{I_{goff}} = \frac{4,5n}{0,5} = 9ns$$

Notice that this duration is too short and implies a minimum of $\frac{di}{dt} = \frac{6,8}{9n} = 0,76A / ns$, and generates minimum voltage $L_p \frac{di}{dt} = 19Volt$ (with $L_p = L_D + L_{PB} = 25nH$, where L_D is the drain parasitic inductance and L_{PB} is the parasitic inductance of the printed path). This value is too high comparing to the value recommended by the driver manufacturer that is 5V to prevent the Boot voltage from being below the under lock bootstrap voltage (v_{BSUV}), from which the driver puts all transistors in the OFF state. So by using the application note [4] and

By choosing a maximum voltage of 1,5V and ensuring a very low parasitic inductance caused by the printed circuit, we get:

$$\frac{di}{dt} = \frac{1,5}{25n} = 60A/\mu s, \text{ which gives us } t_{fi} = \frac{I_{D0}}{\frac{di}{dt}} = \frac{6,8}{60} = 113,33ns$$

Now, to increase the t_{fi} , we have to insert a gate resistor in order to decrease the I_{goff} current for a safe operation of the driver. That current must be:

$$I_{goff} = \frac{Q_{gsp} - Q_{gth}}{t_{fi}} = \frac{4.5nC}{113,33ns} = 39,7mA,$$

Which gives us: $R_{tot} = \frac{V_{cc} - V_{gth}}{I_{goff}} = \frac{17,4-3}{39,7m} = 362\Omega$, knowing that $R_{drive}=34,8\Omega$, that gives $R_g = R_{tot} - R_{drive} = 330\Omega$, and finally $t_{rv} = \frac{Q_{gd}}{I_{goff}} = \frac{53n}{39,7m} = 1,335\mu s$

However, in order to prevent this resistor from having an effect on the transistor closing switching, we will have to provide a signal diode (such as 1N4148 which can stand a DC current of 0.3A) in parallel with the resistor R_g .

2.4.2 Heat sink calculation:

In order to calculate the radiator required for each transistor, we will have to calculate the power dissipated in the transistors. This consists of the P_{ON} power, the switching power in closing and that in opening.

$$P_{onmax} = R_{dson} I_{dson}^2 D_{max} = 13,9W \quad (2.18)$$

$$\begin{aligned} P_{closing} &= \frac{1}{2} f (V_d + V_F) I_{Dmax} (t_{ri} + t_{rr}) + \frac{1}{2} f (V_d + V_F) I_{D0} t_{fv} \\ &= \frac{1}{2} f (V_d + V_F) (I_{Dmax} (t_{ri} + t_{rr}) + I_{D0} t_{fv}) \\ &= \frac{1}{2} 10^3 \times 301 \times (26 \times (53n + 280n) + 6,8 \times 265n) = 13,03 + 2,71 \\ &= 1,574W \quad (2.19) \end{aligned}$$

Where I_{Dmax} is the spike current produced by the reverse recovery charge of the body diode of the transistor of the low side of the same arm (See figure 2.5, in the case of a rotation in a positive sense, current is alternatively flowing through transistors (T_{HS1} and T_{HS2}) and through diodes (D_{LS1} and D_{LS2})), I_{D0} is the average current flowing in motor, V_F is the Von voltage diode and f the frequency transistor's switching, which we choose to be equal to 1Khz.

$$\begin{aligned} P_{opening} &= \frac{1}{2} f (V_d + V_F) I_{D0} (t_{rv} + t_{fi}) \\ &= \frac{1}{2} 10^3 \times 300.6 \times 6,8 \times (1.06\mu + 90n) = 1,175W \quad (2.20) \end{aligned}$$

The total power to dissipate in the transistor is then:

$$P_{\text{tot}}=16.65\text{W}$$

By consulting the **Figure 2.9** from the IRFP350 datasheet that gives the maximum drain current versus the case temperature, we can see that with a maximum case temperature of 125°C it is guaranteed that the transistor can dissipate a current of 7A. So, we have to ensure a case temperature that is not above 125°C .

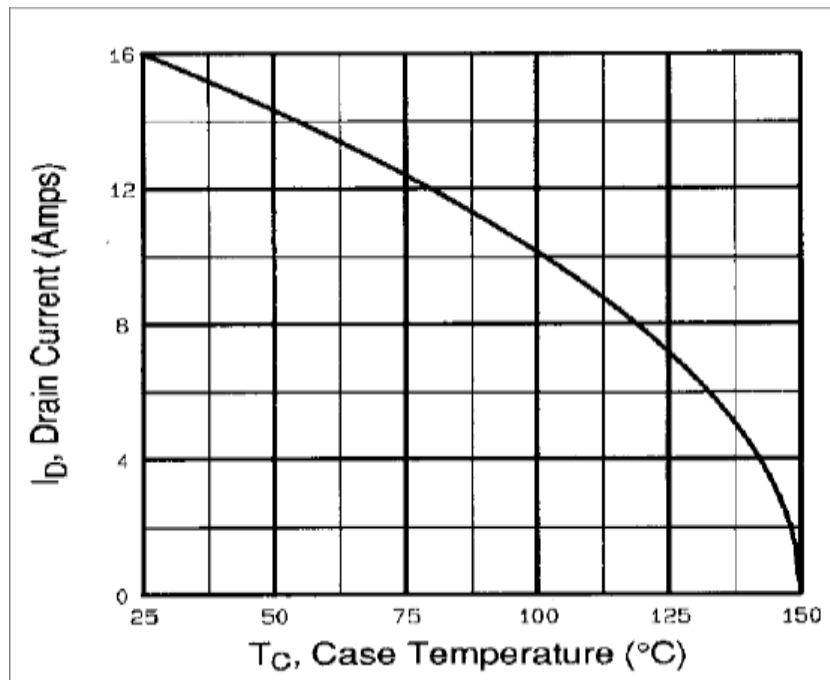


Figure 2.9: maximum drain current versus the case temperature.

Now, by consulting the radiator datasheet (avid standard heat sink datasheet) in order to determine which radiator may suit to us, the best choice that could ensure the case temperature that is not above 85°C is the 6400 heat sink.

2.5 Conclusion

In this chapter we made the choice of the different elements that we need in the power supply unit, the H-bridge and the heat sink(radiator) that ensure a safely operating of switching element of the H-bridge (Mosfet transistors). This choice is made after a specified calculation in the different states of our operating.

3

3.1 Introduction :

Varying motor speed is based on varying the amount of DC voltage supplied to the motor. In the past, the method used to do that was usually a linear one and changing of the amount of DC voltage supplied to motors was not possible without expensive hardware and large efficiency losses, but with the high density of integration of circuits in the nineties, other methods based on the pulse width modulation (PWM for short) control of power switching became more effective and more simple.

PWM control involves giving the motor an average voltage that varies from zero volts to the full amount, and we will see in the next section how this control is made by the varying of the duty cycle for each period of the PWM waveform.

In this chapter, we will focus on the design of the PWM control of the DC/DC converter studied in the second chapter. This control will be based on the integrated circuit UC3842 and on the calculation of external components.

However, before starting the calculation, we will start by specifying the basic principle over which the control will be made.

3.2 Bipolar PWM control:

3.2.1 Voltage Transfer functions of the control:

There is two kinds of PWM control: The bipolar PWM control and the unipolar control. The bipolar PWM control is simpler than the unipolar one. It's based, usually, on switching the high side transistor of an arm with the low side transistor of the other arm of the bridge (HS1 with LS2: On, for a positive average voltage applied on the motor, and HS2 and LS1: On for the negative voltage). This type of control has the advantage to be easy to make and doesn't need complicated circuit control, but it's inconvenient is that it induces in each period of commutation a bipolar voltage on the load, that goes from $-V_d$ to $+V_d$, creating a large harmonic voltages and currents in the motor and in the power line supplying the converter. On the other hand, the unipolar PWM control is a little bit more difficult to make but offers a very small harmonic content, by its aptitude to deliver a unipolar voltage pulses on the output of the converter.

The bipolar PWM control is based on the comparison of a DC Voltage control with an alternative saw tooth signal with a fixed frequency and fixed amplitude (**figure 3.1**). The comparison results make the switch HS1 and LS2 on or off, and inversely make the switches HS2 and LS1 off or on.

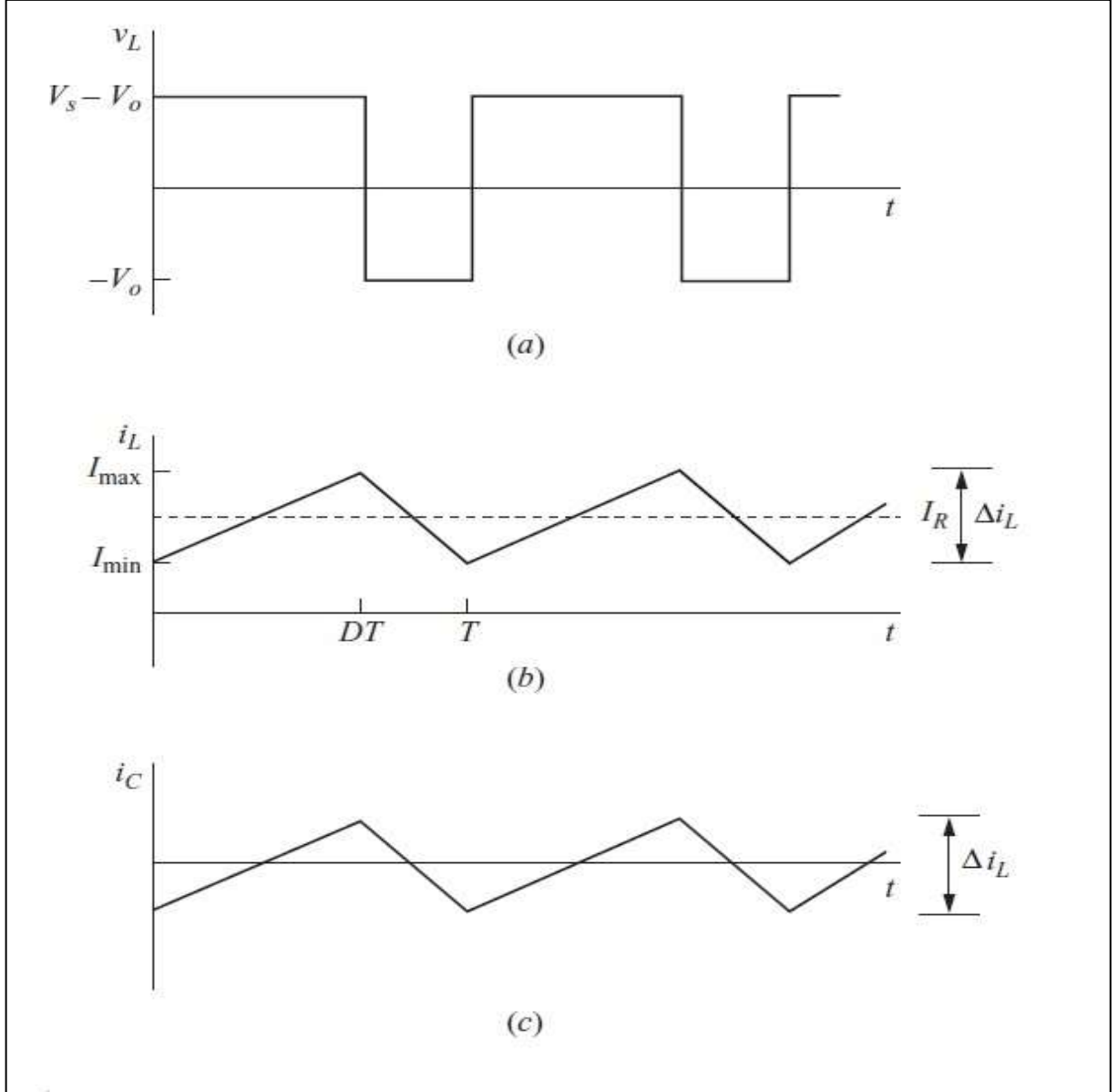


Figure 3.1: Full bridge dc/dc converter for positive armature current.

If we analyze the outputs of the points S1 and S2 (illustrated in figure) witch are connected to the armature input, we can see that:

$$\langle V_{S1} \rangle = \frac{1}{T} \int_0^{DT} V_a dt = DV_a \quad (3.1)$$

Where $D = \frac{t_{on}}{T}$ is the duty cycle corresponding to the closing of HS1 and LS2, t_{on} is the time of this closing and T is the period of the switching.

And

$$\langle V_{S2} \rangle = \frac{1}{T} \int_{DT}^T V_d dt = (1 - D)V_d \quad (3.2)$$

So the average voltage applied between S1 and S2, is:

$$\langle V_{S1S2} \rangle = \langle V_o \rangle = (2D-1)V_d \quad (3.3)$$

This means that voltage is positive When the duty cycle D is $> 0,5$ and negative when the when the duty cycle D is $< 0,5$.

3.2.2 Induced current in the DC motor:

To further design the PWM control of speed motor, we must understand the feedback of our control on the current, the speed and the EMF (electromotive forces) voltage, that armature exhibits.

The energy of each push is stored in the inertia of the heavy platform, which accelerates gradually with harder, more frequent, or longer-lasting pushes.

Pulse-width modulation (PWM), as it applies to motor control, is a way of delivering energy through a succession of pulses rather than a continuously varying analog signal. By increasing or decreasing pulse width, the controller regulates energy flow to the motor shaft. The motor's own inductance acts like a filter, storing energy during the "on" cycle while releasing it at a rate corresponding to the input or reference signal. So we can say that during one commutation, the EMF voltage that motor presents is almost constant, and write:

$$V_o(t) = R_a i_a + L_a \frac{di_a}{dt} + EMF \quad (3.4)$$

Where $EMF = K_E \Omega$, where K_E is a constant depending on the flux induced by stator input current and Ω is the motor speed, R_a : the armature resistor, L_a the armature inductor and I_a the armature current.

The inductance of the armatures being very large in the case of DC power motor: 34mH in the case of our motor, we can than neglect the transient current impact on the medium current I_a . The medium current I_a depend only on the load couple and while no load is applied on the motor, the current induced by the frictions is almost higher than the transient currents induced by L_a , so we can rewrite the equation 3.3 like this:

$$V_o(t) = \langle V_o(t) \rangle + L_a \frac{di_a(t)}{dt} \quad (3.5)$$

Knowing that in this PWM control HS1 and LS2 are ON during DT and are OFF during $(1-D)T$, we can than represent the DC motor and the full bridge within its PWM control with the equivalent schemes illustrated in **figure 3.3**.

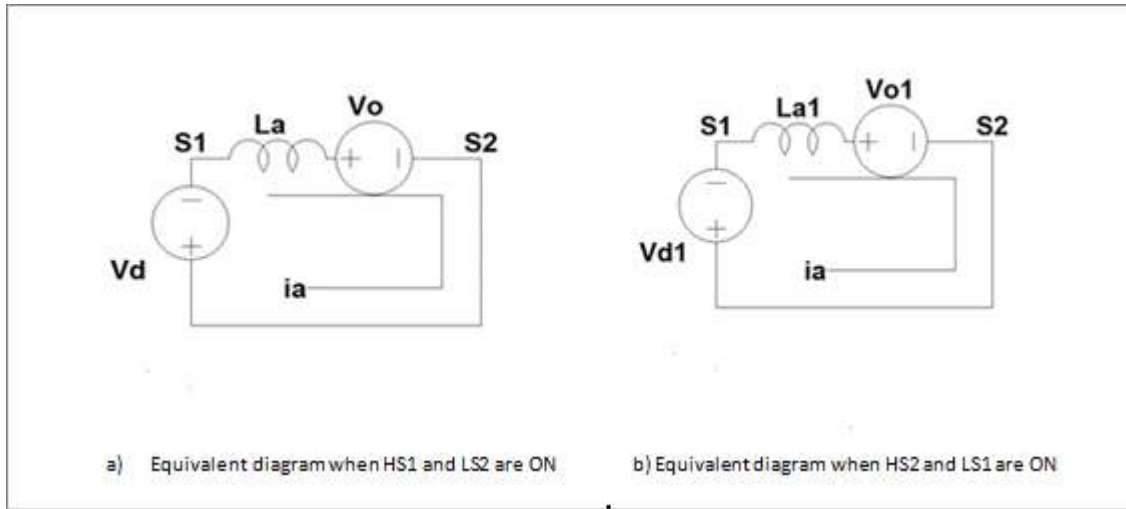


Figure 3.2: Equivalent diagram of DC motor.

We can notice that when HS1 and LS2 go off, the direction current doesn't change, only now the source voltage is applied through HS2 and LS1.

If we analyse these 2 schemes, we can easily find:

$$i_a(t) = \int \frac{V_d - V_o}{L_a} dt \text{ in the case a and } i_a(t) = - \int \frac{V_d + V_o}{L_a} dt$$

So, in each period, the current is growing in DT duration:

$$i_a(t) = \frac{V_d - V_o}{L_a} t + I_o \quad (3.6)$$

and decreasing in the rest of the period with:

$$i_a(t) = - \frac{V_{dmin} + V_o}{L_a} t + I_o \quad (3.7)$$

We can easily demonstrate that the current is almost the constant current I_o with a small ripple of:

$$\Delta I_L = \frac{V_{dmin} - V_o}{2L_a} DT = \frac{V_{dmin} + V_o}{2L_a} (1 - DT) \quad (3.8)$$

If we replace V_o by the relationship given by the equation 3.3, we find that the ripple is:

$$\Delta I_L = V_{dmin} \frac{1 - (2D - 1)}{2L_a} DT = \frac{V_{dmin} D(1 - D)T}{L_a} \quad (3.9)$$

By studying $\frac{d\Delta I_L}{dD}$ we can see that the ripple is maximum when D is equal to 0,5 which gives us the curve of figure 3.4. This curve can be used as a manner of controlling speed, because we notice that when D is growing from 0,5 to $D_{max}=0,95$ and V_o and speed grows from 0 to their nominal values, the current ripple decreases which means that the armature current $I_a - \Delta I_L$ increases when D increases. So according to the torque we will usually have a

growth of current that goes from $I_{amin} = I_a - \Delta I_L(D_{min})$ to $I_{amax} = I_a + \Delta I_L(D_{max})$ and we can notice that I_{amin} is maximum, when D is maximum and minimum when D is minimum.

$$\text{We have for } D_{mjn} = 0,5V, \Delta I_L(D_{min}) = \frac{V_{dmin} D_{min} (1 - D_{min}) T}{L_a} = 1,7A$$

$$\text{And for } D_{max} = 0,95, \Delta I_L(D_{max}) = \frac{V_{dmin} D_{max} (1 - D_{max}) T}{L_a} = 324mA$$

3.3 The PWM controller Design:

To achieve our PWM control, we will use the UC3842A circuit (**figure 3.3**), which is a PWM controller that can generate a duty cycle close to 1. In the case of the UC3842A, the duty cycle and thus the speed control and therefore the armature voltage control, is done by current control.

In fact, the control principle of the duty cycle ratio is based on the comparison of the control voltage V_c , defined by the user, with a voltage V_s proportional to the current flowing in the motor.

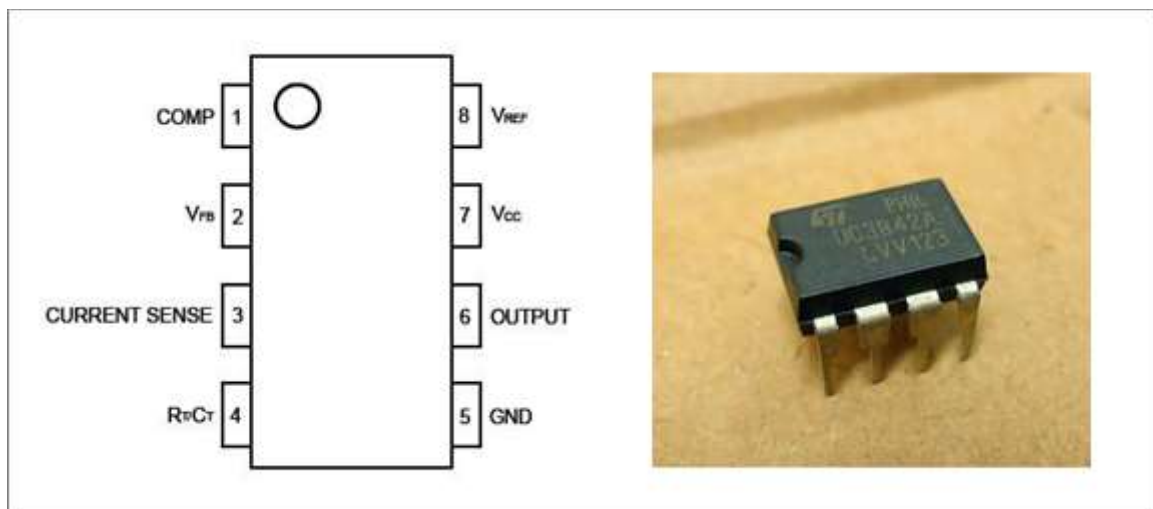


Figure 3.3 : UC3842a.

The comparison is such that for V_c greater than V_s , the internal oscillator of the UC3842 continues to produce a square signal of fixed frequency imposed by the external CT capacitance and of maximum constant duty cycle imposed by the external RT resistor. However, since the current in the motor increases linearly during the charge time (equation 3.6), this means that as soon as the detected voltage V_s is higher than the set-point voltage V_c , the comparator switches to the low state and sets, through an internal RS flip-flop, the output of the UC3842 power stage out to the low state, thus imposing the opening of the HS1 and

LS2 switches (or LS1 and HS2 when the converter produces a negative V_o voltage) before the t_{on} duration reaches its maximum value. This means that with the correct choice of voltage V_c , the duty cycle for closing the bridge power transistors can be varied from 0.5 to D_{max} for a speed variation from 0 to the rated motor speed Ω_{nom} (or from 0.5 to 0 for a speed variation from 0 to $-\Omega_{nom}$). Note, however, that at the beginning, when the motor is stopped, the detected voltage V_s is necessarily zero, since the armature current is then zero whatever is the torque, therefore for a V_c even slightly higher than 0, the PWM controller will generate a maximum duty cycle, so that the armature current evolves towards a current value limited by the value of the voltage V_c .

As long as the armature current has not reached the value such that the electromagnetic torque ($T_e = K_T I_a$) is higher than the load torque (the torque corresponding to friction must also be taken into account), the motor will remain at standstill keep the same state and therefore for an adjustment of the reference voltage, slightly higher than its minimum value V_{cmin} , the armature current will evolve during a transient regime imposed by the electrical time constant L_a/R_a towards the current I_o , whereas the PWM controller will continue to generate pulses whose duty cycle is equal to its maximum value, as long as the voltage V_s has not reached the voltage V_c . As soon as V_s exceeds V_c , the duty cycle of the PWM controller will then depend on the value of V_c and the current evolution in the armature. Thus, for V_c varying from V_{cmin} to V_{cmax} , the duty cycle will vary from 0.5 to D_{max} .

3.4 Changing motor rotation sense:

To rotate the motor in a positive direction, it will therefore be sufficient to control the THA and TLB switches by the output of the UC3842 with a duty cycle varying from 0.5 to D_{max} and to control the THB and TLA switches through an inverter, so that the closing duty cycle of these switches is $(1 - D)T$ (Figure 3.4.a).

The choice of reverse motor rotation can be done simply by a switch or by a relay, so that the UC3842 output controls the closing of THA and TLB in one case and THB and TLA in the other case (Figure 3.4.b)

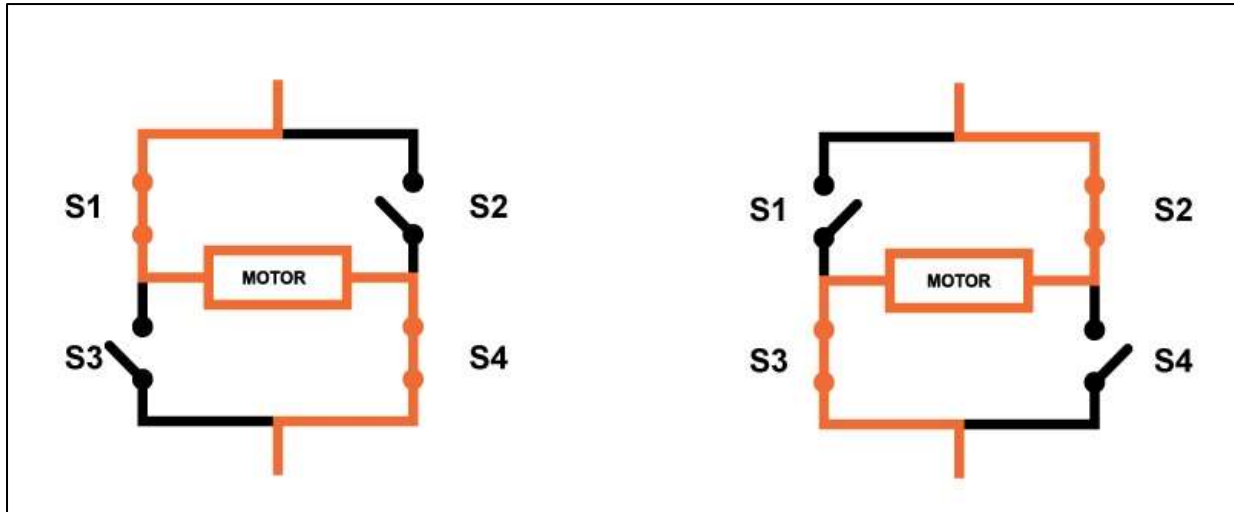


Figure 3.4: (a) Positive Sense. (b): Opposite Sense.

The minimum threshold voltage V_c , from which the motor starts to rotate (in either direction) is

$$V_{cmin} = R_s(I_{a0} - \Delta I_{Lmax}) \quad (3.10)$$

With ΔI_{Lmax} for $D=0.5$.

The maximum reference voltage, for which the armature voltage reaches its maximum value, is

$$V_{cmax} = R_s(I_{a0} - \Delta I_{Lmin}) \quad (3.11)$$

With ΔI_{Lmin} for $D = Dmax$.

The difficulty of control is however made difficult by the fact that the current in the motor varies with the load torque.

In order to simplify the control and to improve its accuracy, we will carry out this control voltage by means of two independent set points, one affecting the speed and the other reflecting the influence of the torque on the armature current, and this by means of voltage dividers consisting of two potentiometer Pot_{speed} and Pot_{torque} and 4 resistors $R1$, $R2$, $R3$ and $R4$.

3.5 Component calculations for the control Board:

3.5.1 R_T and C_T calculations:

The generation of the UC3842a oscillator square signals is based on internal wiring, which ensures the charging and discharging of the C_T capacitance.

As shown in **figure 3.4**, the C_T capacity is charged through the R_T resistor under constant current, from pin 8 where a voltage V_{Ref} (5V) is generated, and is discharged through pin 4.

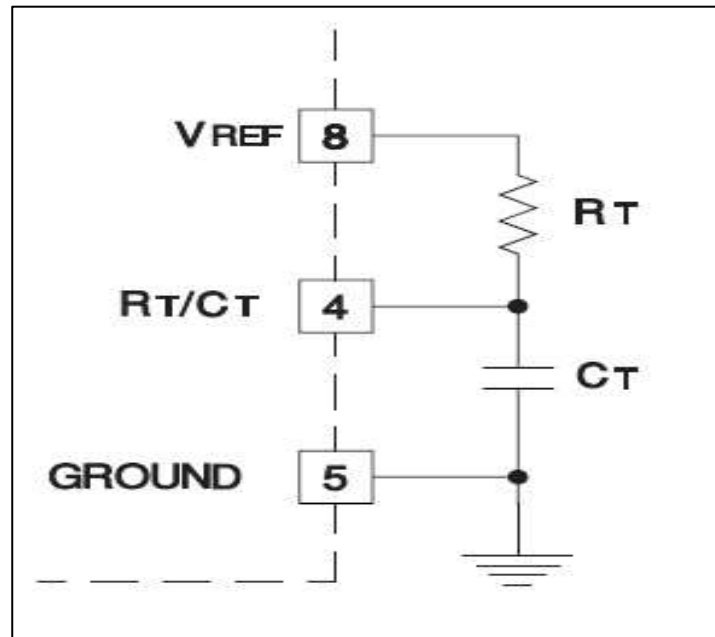


Figure 3.5: $R_T C_T$ Circuit.

The first step in selecting the oscillator components is to determine the dead time. **Figure 3.5** allows the C_T value to be derived from a given dead time. To calculate the dead time, it will be enough to define for our control a maximum duty cycle ratio, in the case of UC3842a the duty cycle ratio being limited to 100%, the relationship linking the dead time to the maximum duty cycle ratio is therefore:

$$T_{DEAD} = (1 - D_{max}) \times T \quad (3.12)$$

In our case, D_{max} being equal to 0.95 the dead time and the switching frequency is set at a value of 1 KHz, the period T is therefore 1ms and thus $T_{DEAD} = 50 \mu s$.

To determine R_T we need to consult the UC3842a Datasheet and take the chart giving the duty time as a function of R_T (Figure 3.5), so when the duty cycle is at maximum value (D_{max}) which is 95% we can see that $R_T = 12K\Omega$.

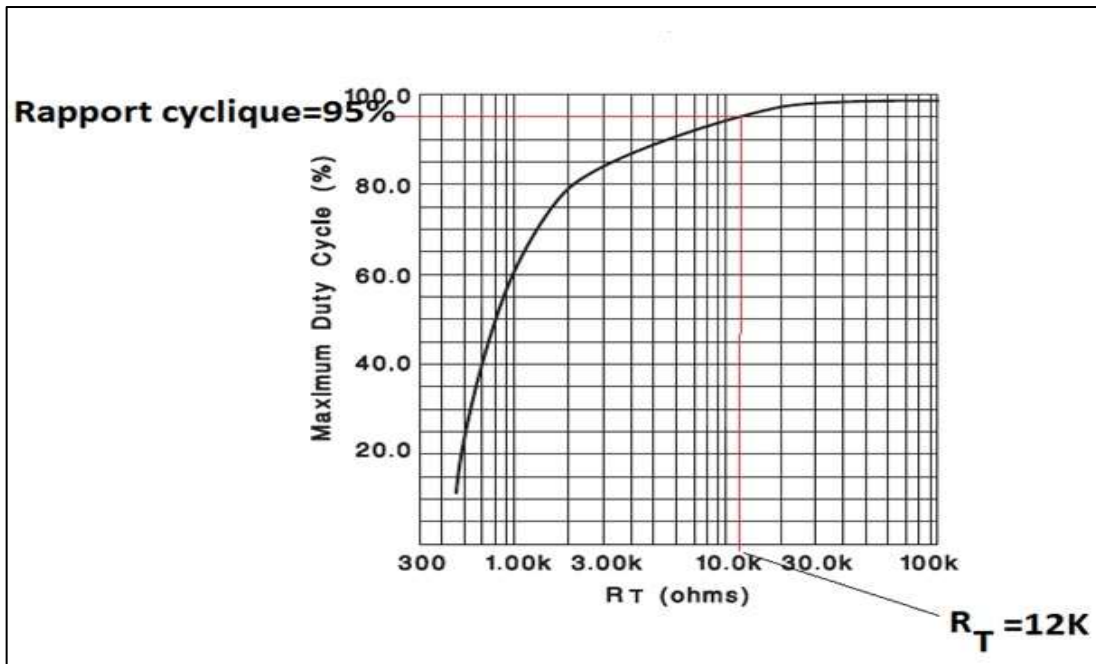


Figure 3.6: Maximum Duty Cycle versus Timing Resistor (determining R_T).

And as for the value of C_T we use the equations

$$f = \frac{1,72}{R_T C_T} \quad (3.13)$$

Knowing that the frequency is 1kHz and $R_T = 12k$ we find that the of the capacity is $C_T = 143.33$ nF, we take a normalised value of 150nF which mean that frequency will be a little bit lower than 1Khz.

3.5.2 Current sensing:

The current detection or sensing of the UC3842a is configured as shown in the **figure 3.6** below

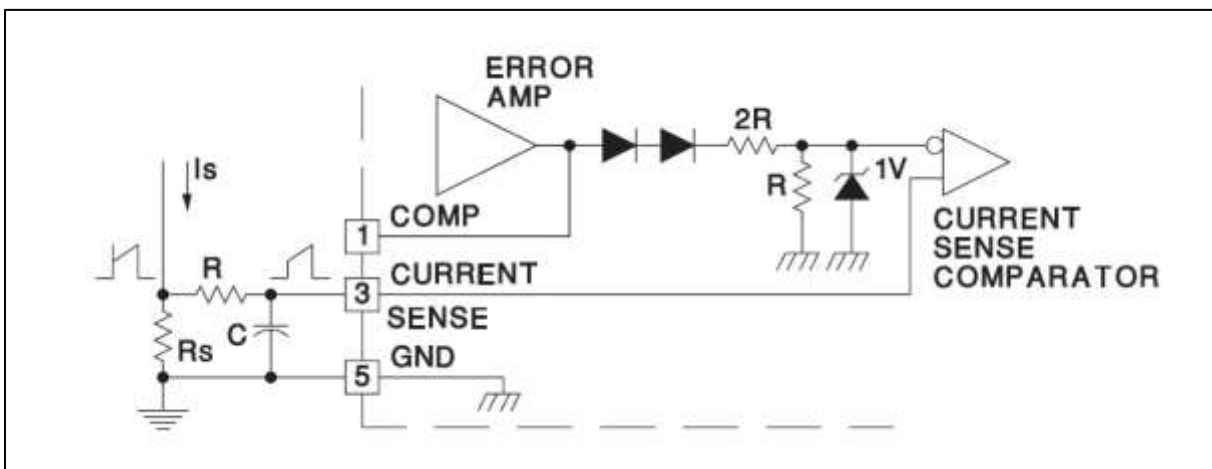


Figure 3.7 : Current Sense Circuit.

The current-voltage conversion is done externally by a resistor R_s connecting the source from the Mosfet to the ground.

The PWM control by current control will therefore be done by means of comparing the voltage U_{ref} with the voltage V_s .

Note however that the voltage U_{ref} is clipped by a 1V Zener diode, by therefore, the maximum voltage of U_{ref} , must in no case exceed the value of 1V.

a. Voltage variation range V_c of the excitation PWM controller and calculation of R_s :

Knowing that the current excitation of the inductor is proportional to the average voltage applied, therefore we can establish that $I_{smin}=0$ for $D = 0$. Knowing that nominal excitation current $I_{fnom}=220mA$ given on the name plate of the motor is maximum when D is maximum, we can write: $I_{smax}=0.22 A^1$ for D_{max} ,

We therefore deduce, from equation 3.3 that $U_{refmin}=0V$. In addition, the variation in the output voltage of the internal amplifier must not exceed a voltage of 1.1V around the virtual ground of 2.5V, because a variation higher than 1.1V below of the virtual mass implies a voltage U_{ref} lower than 0, which is not possible and mean that all the values that can be introduces by the user are similar to a null control voltage and then a null speed. A voltage variation higher than 1.1V, above virtual ground, can also cause the amplifier to not function in its linear region. The maximum value of the maximum voltage U_{refmax} is therefore:

$$U_{refmax} = \frac{1}{3}(U^+ - 2V_d + 1.1) = 0.73V \quad (3.14)$$

Where U^+ is the internal voltage of 2,5V introduced to not have a distortion of the amplifier.

The choice of V_c must be made so that it is lower than 0.73V. We choose

$U_{refmax} = 0.7 V$, therefore $V_{smax} = U_{refmax} = R_s \cdot I_{smax}(D_{max})$, from which we obtain :

$$R_s = \frac{V_{smax}}{I_{smax}} = \frac{0.7}{0.22} \text{ and finally we get } R_s = 3,18\Omega, \text{ we take a normalised value of } 3.3 \Omega.$$

b. Voltage variation range U_{ref} of the armature PWM controller and calculation of R_s :

We know that motor speed and armature average voltage are proportional to the duty cycle D , but we know also that the average current in the armature depends only on the load torque, so how can we make the control of the duty cycle based on current? The Response to this question is in the charge and discharge that inductance L_a induces: more are great the duty

¹Rated value of the excitation current of the inductor indicated on the motor nameplate.

cycle and thus the speed and more are great I_{smin} and V_c . The solution to make the duty cycle controlled by the current is then to have V_c maximum for D maximal and I_s maximum and inversely V_c minimal and I_s minimal for D minimal (than close to 0,5).

Based on this analysis, we can already establish:

$$V_c < 1v \quad \text{and} \quad \frac{V_{cmax} - V_{cmin}}{V_{cmax}} = \frac{I_{smax}(D_{max}) - I_{smin}(D=0,5)}{I_{smax}(D_{max})} \quad (3.15)$$

In the case that the motor drives a nominal load torque, the average value of the current is $I_a = 6.8A$ and we have:

- For D just above 0,5: $I_{smin} = I_a - V_{dmin} \frac{1-D_{min}}{L_A} D_{min} T = 5,1A$
- For $D = D_{max}$, $I_{smax} = I_a + V_{dmin} \frac{1-D_{max}}{L_A} D_{max} T = 7,12A$.

Using Equation 3.13, we can determine the relative range of variation

$$\frac{I_{smax} - I_{smin}}{I_{smax}} = 1 - \frac{U_{refmin}}{U_{refmax}} = 0.28$$

This gives: $\frac{U_{refmin}}{U_{refmax}} = 0,72$

Taking the choice U_{refmax} of 0.7V, we obtain $U_{refmin} = 0.504V \cong 0,5V$

We can deduce the sensing resistance R_s that we must take: $R_s = \frac{V_{cmax}}{I_{smax}} = \frac{V_{cmin}}{I_{smin}} = 0.1\Omega$. With nominal power greater than: $R_s I_s^2 = 0.1 \times 6.8^2 = 4.7W \approx 5W$.

3.5.3 Error Amplifier:

The role of the amplifier (**Figure 3.8**) is to convert the voltage reference V_{ref} introduced by the user through a potentiometer in a voltage control V_c .

The amplifier as shown in Figure 3.8 is used in reverse amplification and therefore at gain $-\frac{z_f}{z_i}$.

An internal 2.5V voltage source is applied to the non-reversing terminal of the amplifier, to ensure distortion-free amplification by making the virtual ground is not by ground and therefore the reference potential 0, but before that $\frac{V_{reff}}{2}$.

This way everything happens, as if the potential 0 feeding the $-V_{cc}$ of the amplifier becomes equal to -2.5V with respect to the virtual ground.

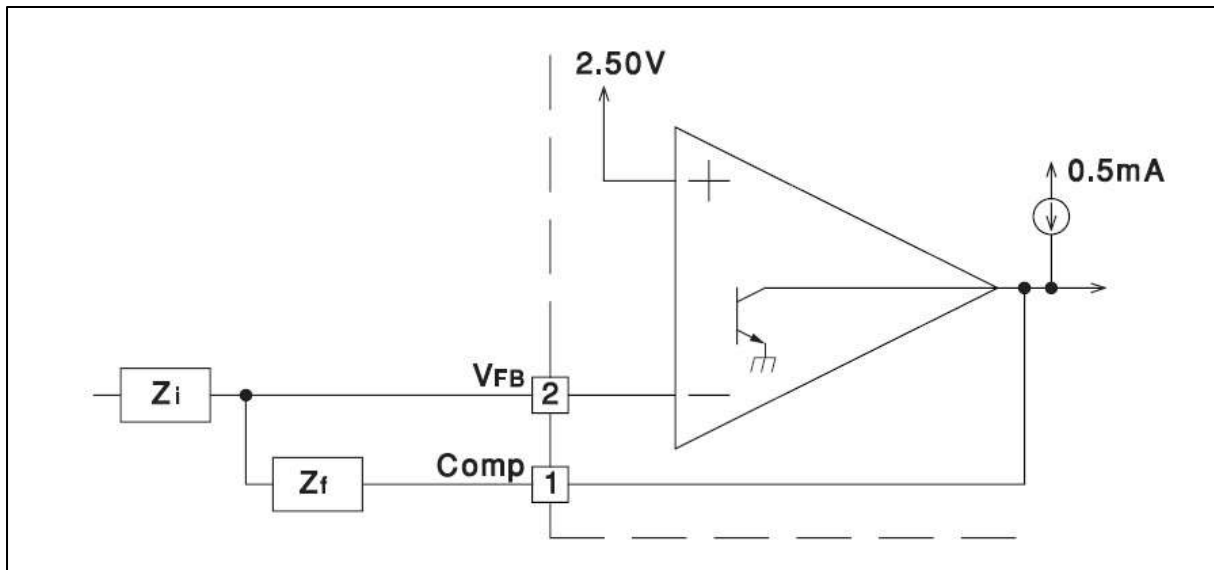


Figure 3.8: Error Amp Configuration.

To ensure a high output voltage V_{ref} , the R_f resistor must be selected so that the output current of the error amplifier is less than 0.5 mA.

In this case the minimum R_f value, must be such that

$$R_{fmin} = \frac{V_{outmax} - 2.5v}{0.5mA} \quad (3.16)$$

We deduce that $R_{fmin} = 5k\Omega$

This implies that the value of the resistors R_5 , R_6 and R_7 should be much higher than the value of R_{fmin} , so for the purpose of not disturbing the amplifier function and to ensure a unit gain, they must have the same value and this value must not be below $5k$, because this forces the O-amp to deliver a high current which implies a decrease of the output voltage of the internal O-amp (current limited to 2mA in the high state and 0.5mA in the low state), so for that we take

$$R_5 = R_6 = R_7 = 100K\Omega$$

3.5.4 R1, R2, R3, R4 and Potentiometers Calculations:

Until now, we have been able to establish for a nominal load torque a range of variation of voltage control V_{cmin} to V_{cmax} that ensure a variation of the duty cycle from 0,5 to 0,95, and then a variation of speed from 0 to Ω_{nom} , but what would happen if the load torque change? We know that reduction of load torque implies a proportional reduction of current, but we know also that armature current ripples do not depend on load torque, but depend only on duty cycle. Knowing this, we thought first to make 2 independents controls on V_c , one related to the reference speed introduced by the user through a potentiometer called Pot_{speed} which make the range of the variations of V_c equal to 0,2V, and the other one related to a

reference load torque introduced by the user through a potentiometer called Pot_{Torque} which reduce the voltage control proportional to the average armature current. But, in this way, the minimal control voltage will be 0V, so corresponding to an average current of $1,7A^2$, thus corresponding to only on a load torque equal to $\frac{1,7}{6,8}T_{nom} = 0,25T_{nom}$. This means that for load torques smaller than $T_{nom}/4$, the variation of speed is not possible because V_c cannot be smaller than V_s . So despite all the values of the speed potentiometer, speed will still be at its maximal value. To overcome this problem, When we reach this limit of V_c of 0V, we may act on V_s , by multiplying V_s by a factor of 4 and then changing the torque potentiometer to its maximum value, which make the range value of V_c another time between 0,5V and 0,7V and make speed variation possible in a new range of load torques between $T_{nom}/4$ and $T_{nom}/16$, and so on if the load torque is smaller than this minimal value, we can amplify V_s by a factor of 16.

-Establishing the equations defining the set point voltages:

The speed and torque set point voltage is fed to the UC3842 through the internal inverter amplifier which delivers a voltage

$$V_0 = V^+ \times \left(1 + \frac{R_5}{R_6}\right) - \frac{R_5}{R_6} (V_{speedsetpoint} + V_{torquesetpoint}) \quad (3.17)$$

V^+ being the internal voltage of 2.5V, which the circuit develops from the voltage of its reference voltage of 5V to avoid distortion of the set voltages. Looking at the internal diagram of UC3842 we can see that

$$V_c = \frac{R}{3R} (V_0 - 2V_d) \quad (3.18)$$

If we consider equations 3.15 and 3.16 we can notice that minimum values of voltage set points correspond to maximal value of V_c and vice versa maximum values of voltage set points correspond to minimum values of V_c .

We know that when V_c takes its minimum 0V value, Duty cycle goes to 0.5, so speed goes to 0 and the average armature current I_a equals to $I_{anom}/4$ and the minimal load torque will be $T_{nom}/4$. This is the case where V_{SSP} and V_{TSP} take their maximal values, so we can establish from equation 3.15 and 3.16 for $V_{cmin} = 0$: then $V_0 = 3V_{cmin} + 2V_d = 1,2V$ and hence for $R_6 = R_5$:

$$V_{SSPmax} + V_{TSPmax} = 3.8V \quad (3.19)$$

² $I_a - \Delta I_L(D_{min}) = 0$ thus $I_a = \Delta I_L(D_{min}) = 1,7A$

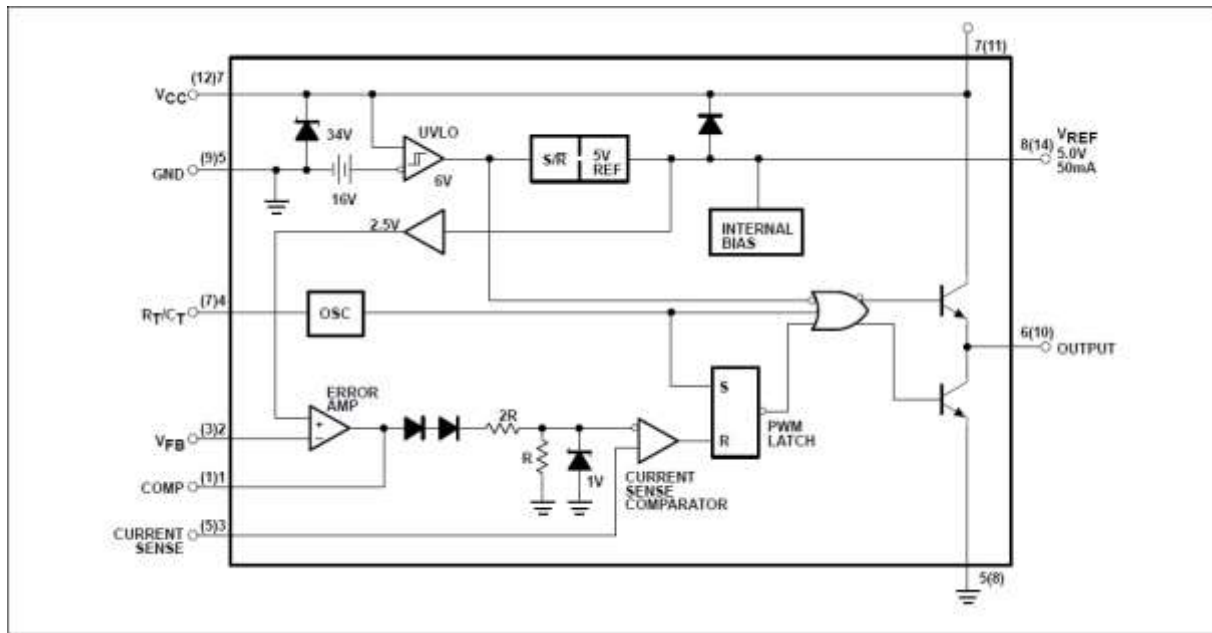


Figure 3.9: internal diagram of UC3842a.

For the choice of the voltage V_{cmax} , we will base the criteria of our choice on the operation of the O-amp in linear mode. By choosing a value of 3.3 Volts for the output voltage of the op amp, we will certainly avoid saturation of it, and therefore the value of V_{cmax} will be 0.7V. This is a very good choice since the control voltage is clipped by an internal 1Volt zener diode, we can control other DC motors, which nominal currents can reach 10A ($1V/R_S$). From this choice, we deduce:

$$(V_{speedsetpoint} + V_{torquesetpoint})_{min} = 5 - V_0 = 1,7Volt \quad (3.20)$$

To solve these 2 equations with 4 unknowns, 2 other independent equations must be found. If we take a look on the result of maximal excursion be done by Pot_S on voltage control V_c , we can say that independently of the position of Pot_T, a variation of Pot_S from V_{SSPmin} to V_{SSPmax} will usually implies a variation of $V_{cmax} - V_{cmin} = 0,2V$. If we consider result of variation of Pot_T on V_c , we can also show that independently of the position of Pot_S, excursion of Pot_T from V_{TSPmin} to V_{TSPmax} will implies a variation of $V_{cmax} - V_{cmin} = 0,68 - \frac{0,68}{4} = 0,51V$. So, we can establish a new independent equation by using:

$$\frac{V_{SSPmax} - V_{SSPmin}}{V_{TSPmax} - V_{TSPmin}} = \frac{0,2}{0,51} = 0,39$$

This third equation means that excursions of Pot_S represent 39% of excursions of Pot_T, so we can rewrite equations 3.17 and 3.18 as follow:

$$V_{SSPmax} + V_{TSPmax} = 1,39V_{TSPmax} = 3,8 \text{ so } V_{TSPmax} = 2,73V \text{ and } V_{SSPmax} = 1,07V$$

$$\text{and } V_{SSPmin} + V_{TSPmin} = 1,39V_{TSPmin} = 1,7V \text{ so } V_{TSPmin} = 1,22V \text{ and } V_{SSPmin} = 0,48V$$

If we consider the figure 3.8, we can assert the follow equations:

$$\frac{R_2 + Pot_{speed}}{R_2 + Pot_{speed} + R_1} V_{ref} = (V_{speedsetpoint})_{max} = 1,07V \quad \text{For } V_{ref} = 5V \quad (3.21)$$

$$\frac{R_2}{R_2 + Pot_{speed} + R_1} V_{ref} = (V_{speedsetpoint})_{min} = 0,48V \quad (3.22)$$

$$\frac{R_3 + Pot_{torque}}{R_3 + Pot_{torque} + R_4} V_{ref} = (V_{torquesetpoint})_{max} = 2,73V \quad (3.23)$$

$$\frac{R_3}{R_3 + Pot_{torque} + R_4} V_{ref} = (V_{torquesetpoint})_{min} = 1,22V \quad (3.24)$$

Having only 4 equations for 6 unknowns, we will have to choose 2 values and calculate the other four values. We will then choose the potentiometers which must be multi-turn potentiometers with 5% accuracy ($Pot_{speed} = Pot_{torque} = 10k\Omega$). We will however have to be very careful in our choice, because the application of our voltage dividers is only acceptable in the case where the impedance of the resistors R_5 and R_6 are at least 20 times greater than the output impedance of the dividers.

So we arrive at the next equations:

$$0,214R_1 - 0,7861R_2 = 0,786Pot_s \quad (3.25)$$

$$4,52R_2 - 0,48R_1 = 0,48Pot_s \quad (3.26)$$

$$2,73R_4 - 2,27R_3 = 2,27Pot_T \quad (3.27)$$

$$3,78R_3 - 1,22R_4 = 1,22Pot_T \quad (3.28)$$

We obtain

- $R_1 = 63,9k$
- $R_2 = 8K$
- $R_3 = 8,2k$
- $R_4 = 15k$

3.5.5 R_8 , R_9 , R_{10} , R_{11} , R_{12} and R_{13} Calculations:

The IR2133 is characterised by a V_{Itrip} , a fault output and resistors R_{11} , R_{12} and R_{13} that are very useful to ensure the protection of the H-bridge and the motor against any eventuality of a short circuit and thus currents exceeding the nominal current and the safe motor operation. Voltage at the terminals of the shunt resistor is used to deliver an I_{trip} voltage, which as soon as it exceeds the maximum value of 0.67Volt (see datasheet) lowers the power transistor control outputs and signals an excessive current through the brightness of a Led connected to the fault output. The use of resistors R_{11} and R_{12} gives more flexibility in

the detection of the fault. In fact, these two resistors placed in parallel with the shunt resistor allow to adjust the voltage V_{Itrip} of the IR2133 (this voltage V_{Itrip} is indicated on the datasheet with a minimum value, typical and a maximum value, respectively 470mV, 570mV and 670mV), so the calculation of R_{11} and R_{12} therefore consists only in making a voltage divider which can apply this voltage at the moment when a current higher than the rated current crosses the motor. In general, a current of the order of 2 times the rated current is taken to avoid too strong starting currents, so that voltage divider constituted by R_{11} and R_{12} , therefore $\frac{R_{12}}{R_{11}+R_{12}} V_{shuntmax}$ reaches the value $V_{Itrip} = 670\text{mV}$ (we take 670mV to be sure that whenever V_{Itrip} is in the range between $V_{Itripmin}$ and $V_{Itripmax}$, it will be guaranteed that fault will be detected).

The resistors R_{11} and R_{12} constitute a voltage divider, of which the delivered I_{trip} voltage must exceed 0.67 volt as soon as the current exceeds twice the nominal current. Their values must therefore be very large in relation to the shunt resistance and must be such that V_{Itrip} is equal to 0.67V. To make it possible to calculate, we fix one of the two resistances at a value of our choice and we calculate the other one, so we take $R_{12} = 100\text{k}$ and we calculate R_{11} using

$$\frac{R_{12}}{R_{11}+R_{12}} V_{shuntmax} = 0.57V \quad (3.29)$$

As regards the value of the shunt resistance, its value must be such that $2I_{anom} R_{13}$ is greater than or equal to 0.67V, therefore $R_{13} \geq \frac{0.67}{2 \times 6.8} = 0.05\Omega$ which must dissipate a power in nominal conditions greater than $0.05 \times I_a^2 = 2.3\text{W}$.

In addition to driving the 4 transistors of the converter and detecting current faults, The IR2133 circuit offers the possibility of adapting the shunt resistance to the detected voltage V_s needed in PWM controller. Indeed, the IR2133 has an amplifier that allows through resistors R_8 and R_9 (see **Figure 3.10**) to adapt the voltage proportional to the current that is across the shunt resistor R_{13} which is injected into the Ca^+ input (non-inverting terminal of the O-amp) so that it is adjusted according to the user's wishes by the correct choice of resistors R_8 and R_9 which adjust the gain of the non-inverting amp that provides the maximum voltage corresponding to the sensing current at the UC3842A Sensing input, which we had fixed at 0.7V, in these conditions the gain of the non-inverting amplifier is

$$V_s = V_{cmax} = 0.7 = \left(1 + \frac{R_8}{R_9}\right) V_{sh} \quad (3.30)$$

Once the shunt resistance is determined, then we will have:

$$\left(1 + \frac{R_8}{R_9}\right) = \frac{0.7}{R_{sh} \times I_{anom}} \cong \frac{0.1}{R_{sh}} = 2 \quad (3.31)$$

So, we have approximately $R_8 = R_9$ and we choose a value higher than 10K as already explained. Note that if we do not find the shunt value of 0.05Ω , we can choose R_8 different from R_9 , so that we have:

$$1 + \frac{R_8}{R_9} = \frac{0,7}{R_{shunt}} \quad (3.32)$$

For the value of R_{10} , it must be chosen so as to minimize the effect of the polarization currents at terminals Ca^- and Ca^+ on the output voltage. Because in addition to the offset voltage (around a few mV) at the output of the amplifiers, these are characterized by currents I_p at the terminals - and + and these currents risk generating an error voltage between the terminal - and +, which will be amplified and risk generating a high error voltage. One solution to minimize this problem is to ensure that the equivalent impedance present at the terminal - is equal to that presented at the + terminal, so that the V^+ due to the passage of the polarization current is equal to the V^- , which implies that $R_8/R_9 = R_{10}$.

a. Circuit adaptation to load torque changing:

We have seen in a previous section that our speed control is limited to a limit load torque of $0,25T_{nom}$, and under this load torque, speed change is no more possible. So as we have already proposed, a solution to this problem is to amplify the sensing voltage V_s when current goes under $\frac{I_{anom}}{4}$. The solution is then to use a switcher and resistors as illustrated in the figure, so have R_8 equal to R_9 when load torque is higher than $0,25T_{nom}$, R_8 equal to $7R_9$ (which make $V_{smax} = 8 \times R_{sh} \times I_s = 4I_s$) for load torque between $0,25T_{nom}$ and $\frac{1}{16}T_{nom}$, R_8 equal to $31R_9$

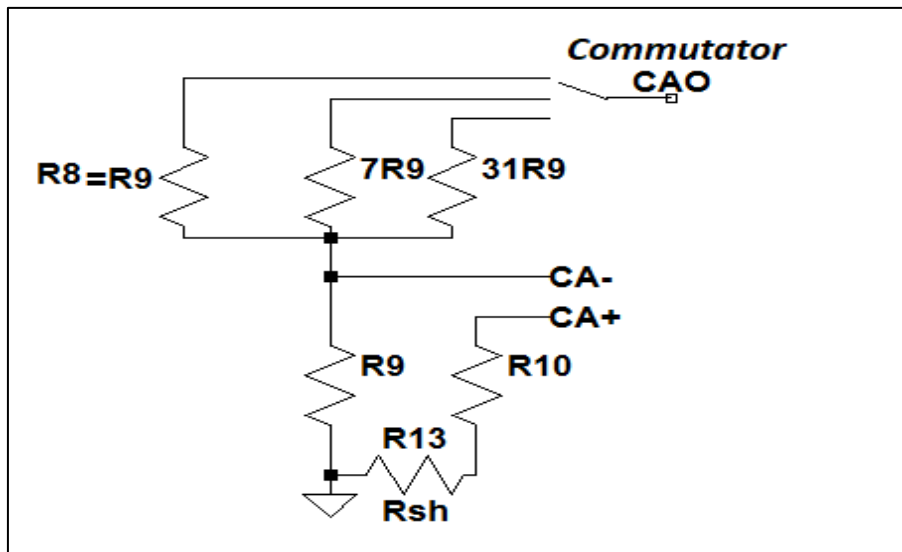


Figure 3.11: shunt resistance related diagram.

So at last we arrive at the values

- $R_8 = 10\text{k}\Omega$
- $R_9 = 10\text{k}\Omega$
- $R_{10} = 5\text{k}\Omega$
- $R_{11} = 13\text{k}\Omega$
- $R_{12} = 100\text{k}\Omega$
- $R_{13} = 0.05\Omega$

3.6 Conclusion:

In this chapter we touched the subject of control in our project by studying each part and component used to make our circuit board, from the UC3842a which ensures us, a good control of speed over our motor.

4 Chapter 4

4.1 Introduction

The goal of a bench test for DC motors is to offer all necessary accessories, facilitating mechanical and electrical measurements and the relationship between them, so as to measure Torque versus speed and armature voltage, power versus speed, etc.,. For that purpose, the bench must contain the motor that will be characterized, the speed variator that feed it by varying average armature voltage, the load torque that must be measured with a torque meter, an optical speed sensor that must measure its speed by converting it to pulse which can be measured directly either by a frequency meter or an oscilloscope, DC voltmeter and DC ampere meter to measure the average armature current and voltage, banana plugs for cabling, and fault protections circuit that must be efficient to avoid any dangerous motor operation. As we had explained in the last chapter, our speed variator is based in controlling average armature voltage, and so the speed, basis on 2 kind of current measurements:

- Ripple armature current measurement which is somewhat compared within a reference, the user introduce within a precision potentiometer we called Pot_{speed} or Pot_ξ that command duty cycle variation and then speed variation, from a minimum value to a maximum value.
- Average armature current measurement, which is a measure in some¹ way of the real load torque, and hence is compared to a reference, via a potentiometer we called Pot_{Torque} or Pot_T, the user adjust to make possible speed variation. This sensible adjustment of this potentiometer must be calibrated as we will try to explain further, to be a measure of the real load torque which can be very useful when we don't have torque meter.

So, the speed variator, we designed, one time calibrated will save the complicated presence of a torque meter. Finally, in this last chapter, after a presentation of our print board and some explanation of the software we used, we will present the panel, the circuit must have to facilitate measurements and finally we will try to give some guidelines for a further project of calibration of our realization.

¹ $I_a = K_T T_{em}$, where K_T is a constant that depends only on the intensity of current flowing in to the excitation windings, and T_{em} is equal to the friction Torque + load torque, so knowing the friction torque, we can say that the average armature current is also a measure of load torque.

4.2 Printed Board design:

When PCB was not discovered yet, all components were connected with wire externally, same as we connect all components on a breadboard. But this increases complexity and size. To solve this issue, PCBs were invented. Now to design a PCB, there are many tool and software available like EAGLE, Multisim, EasyEDA, Altium design, Proteus, OrCAD, and KiCAD. These softwares are widely used for PCB designing as well as for simulation of the circuit. In this project we have designed our PCB using EAGLE.

4.2.1 EAGLE

Most useful and common PCB software designing is EAGLE. EAGLE has a very simple, effective and easy interface and provides a library which contains a large number of electronics components. That is the main reason for being most popular PCB designing software among educationist and professionals. The free version of this software is available on website of AUTODESK, from where we already downloaded it.

EAGLE has two editors, schematic editor and PCB layout editor. The schematic editor is used to add all components and connect according to the circuit requirement. This schematic file has unique features like modular design block, multi-sheet schematic, electronic rule checking and real-time design synchronization. After this, schematic is directly converted into PCB layout editor, in which we can set components according to less complexity. This PCB layout editor also has some hood features like alignment tools, obstacle avoidance and routing engine. These features are available in free version. The big disadvantage in the free version of EAGLE is the limited size of board (10 centimetres \times 8 centimetres), which means that we cannot design and make a PCB boarder higher than 80cm². This disadvantage obliges us to design three different boards.

a. Power supply board

The power supply board required to feed the other two boards, one by a voltage of 310V, which is the H-bridge board, and the other by a voltage of 18V which feed the control board.

This board contains a **GBU 12** full bridge, two capacitor in parallel which gives a filter capacity of 550 μ F (value calculated previously in chapter 2), a resistance for the discharge of the capacitor for supplies a voltage of 310V for the H-bridge, and two other capacitor which form of a voltage divider in order to have a voltage of 31V to feed the 7818V regulator in the

control unit which we will discuss later in this chapter. **Figure 4.1** shows the circuit from the schematic editor in EAGLE.

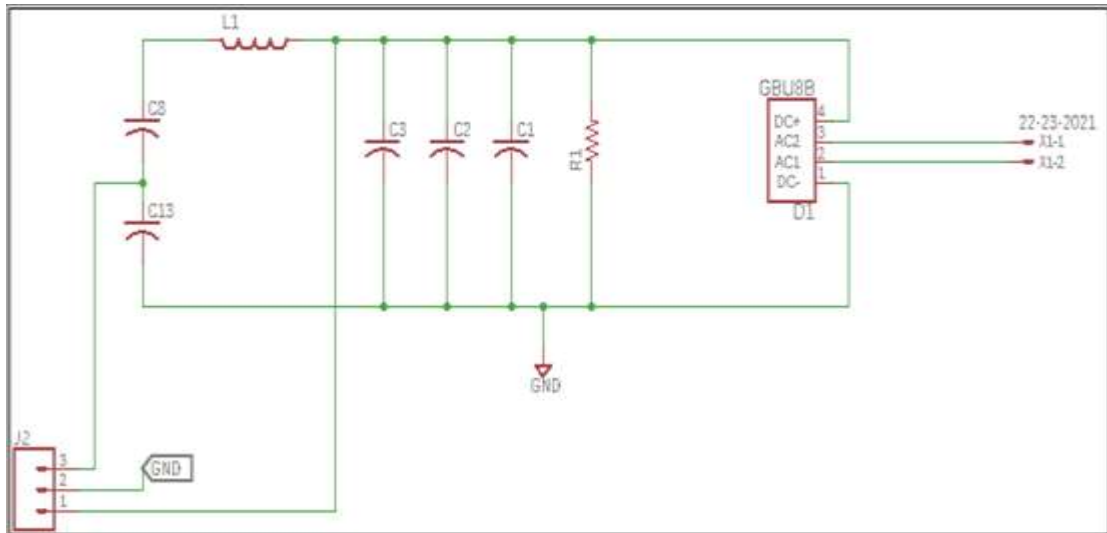


Figure 4.1: schematic of power supply board.

This schematic converted into PCB layout editor. After we set our components in where it's less complex we get a one side board as shown in the **figure 4.2**.

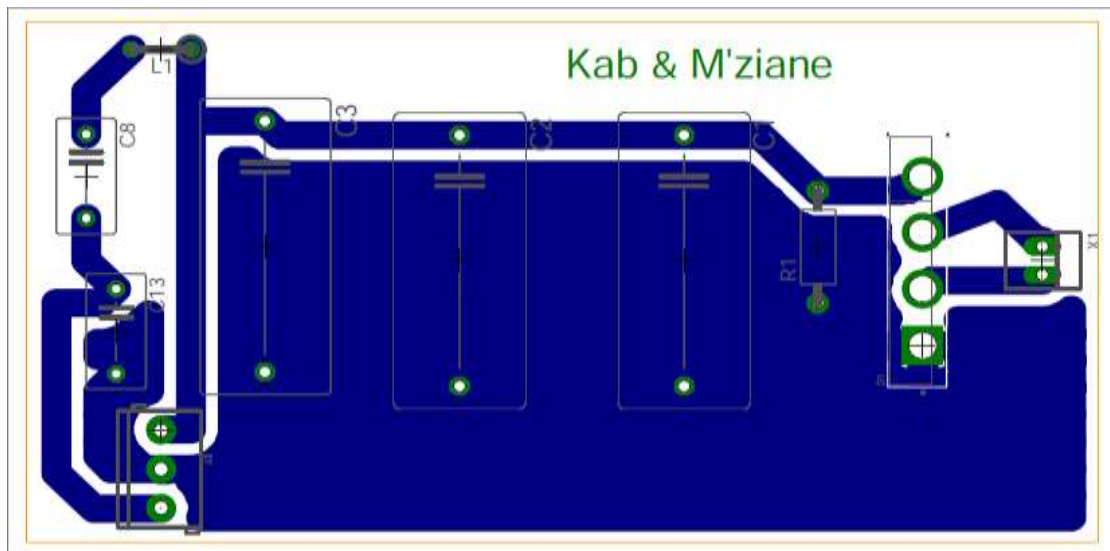


Figure 4.2: PCB layout of power supply board.

Our board size is 9,1cm width and 4,7cm height.

b. H-bridge board

The second board is constituted by the H-bridge and its driver which is the IR2133.

The IR2133 doesn't exist in EAGLE library, after searching on web but we didn't find it. Another driver was found, the IR2130, it has the same package with the IR2133, but

the pines are not in the same position. Fortunately, we found that EAGLE offers to us the opportunity to edit a library to make a new one. Indeed, that what we did for the IR2130 to get a new library for IR2133.

The **figure 4.3** shows the circuit from the schematic editor in EAGLE.

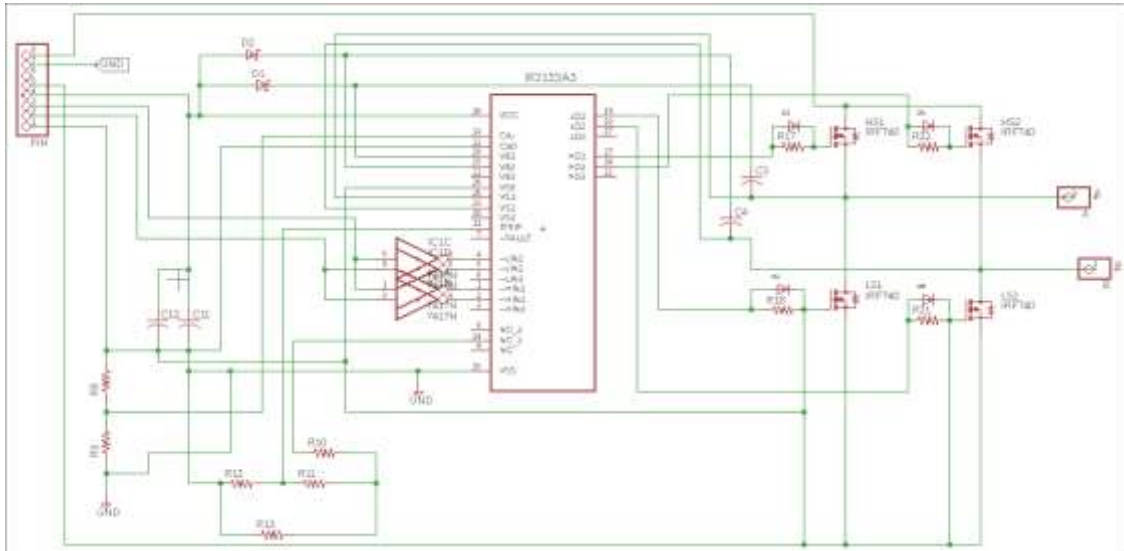


Figure 4.3: schematic of the second board.

It seems from this schema that the board will be a little more complicated than the power supply board, and because of the limited size (only 80cm), that forces us to design a two side board, and that what we get in the **figure4.4**. The blue connection is on the bottom side and red connection is on the top side.

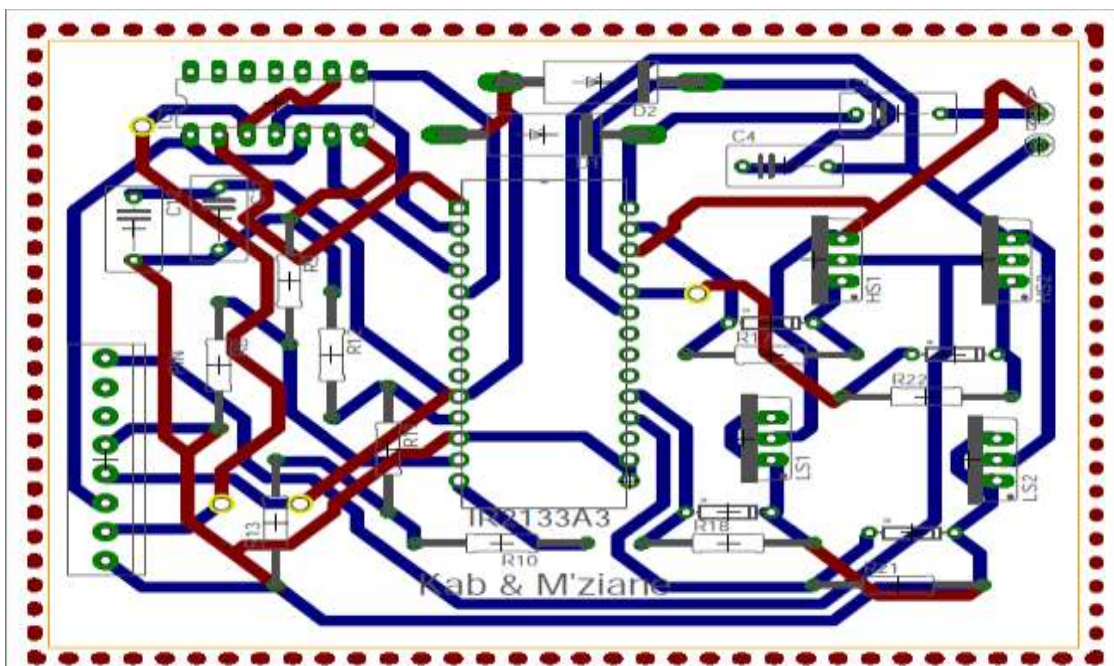


Figure 4.4: PCB layout of the second board.

c. The control unit board

This board is constituted by a regulator 7818, relay 12V and the control unit that we discussed previously in chapter three, it formed by an integrated circuit which is UC3842, resistors, capacitors and two potentiometers, as shown in the figures 4.5 from the schematic editor

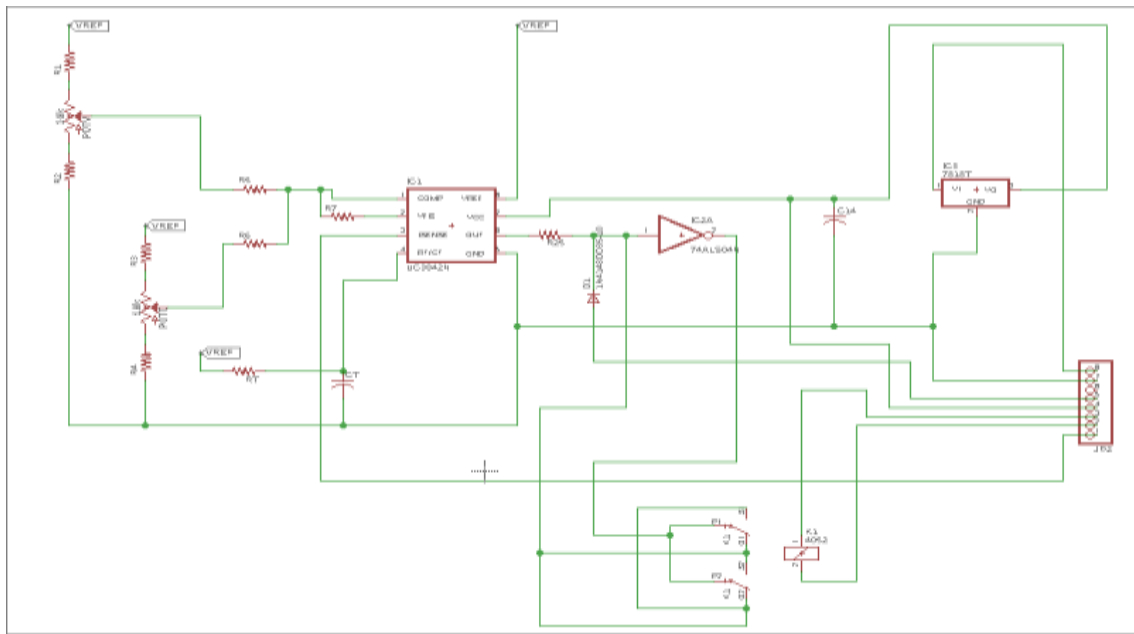


Figure 4.5: schematic of the control unit.

After set the component in reasonable places in the PCB layout editor we get a two side board as shown in figure 4.6.

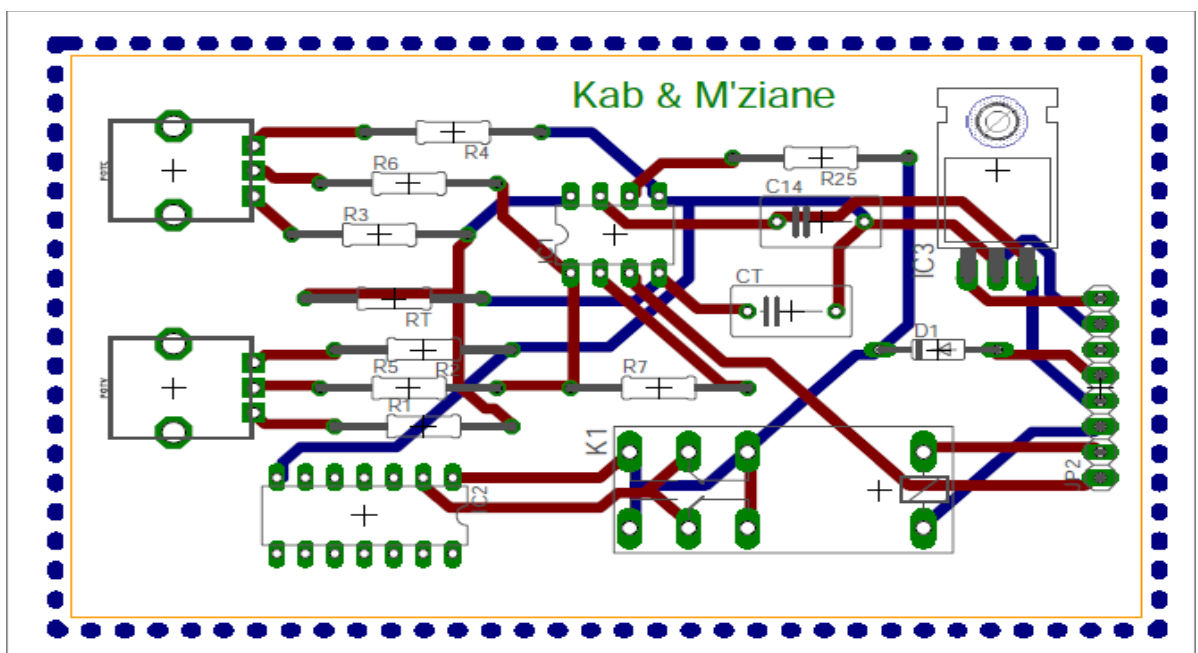


Figure 4.6: PCB layout of the control unit.

4.3 Panel design:

To facilitate measurements, the panel must present as illustrated in **figure 4.7**, the two precision multi-turns potentiometers Pot_S and Pot_T early studied in chapter 3, the double four pole through switch that was proposed to small load torque measurements, 2 banana plugs + and - linked respectively to the node S1 and to the node S2 for armature windings, 2 banana plugs + and - for excitation windings². Other banana plugs for armature average current and voltage measurement will be also useful. Banana plugs for average excitation current measurement will be useful to measurements of speed and adjustment of torque while excitation flux is varying. Finally 2 male BNC records for speed measurements within oscilloscope or frequency meter.

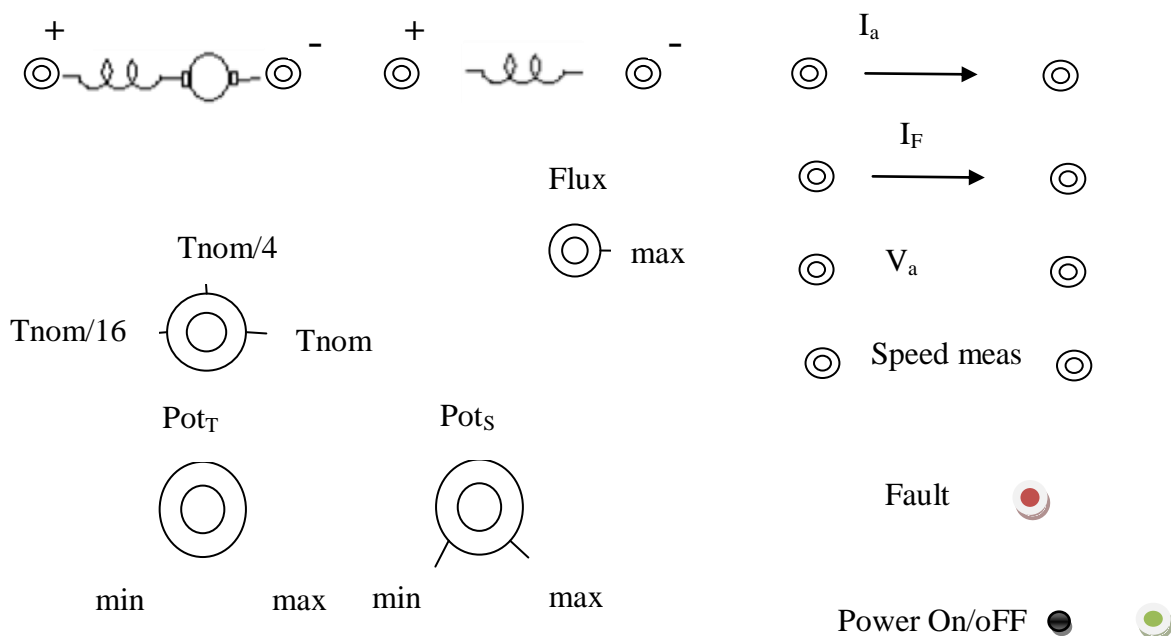


Figure 4.7: Panel.

4.4 Discussion of Calibration:

We will try in this section to:

- 1- explain the way that we will use to make speed variations while load torque is not known, and by varying speed, we will find how we can measure

² which will be powered by a control card realized before by other students to make variable excitation voltage

electromagnetic Torques, and thus to have speed versus electromagnetic torque.

- 2- Try to explore a way to calibrate our special speed variator.

We know that speed is usually related to load torque, so to varying the motor speed, we must apply a known load torque, and then measure the speed to be able to find the torque-speed-armature voltage curves. But how can we apply a load torque? In the practice, the method usually used, is to fix the engine shaft of the motor we want to characterise to the shaft of a dc generator that will be rotated and produce electrical energy proportional to the electrical load that is connected to its excitations windings, and measure the electromagnetic torque produced via a torque meter.

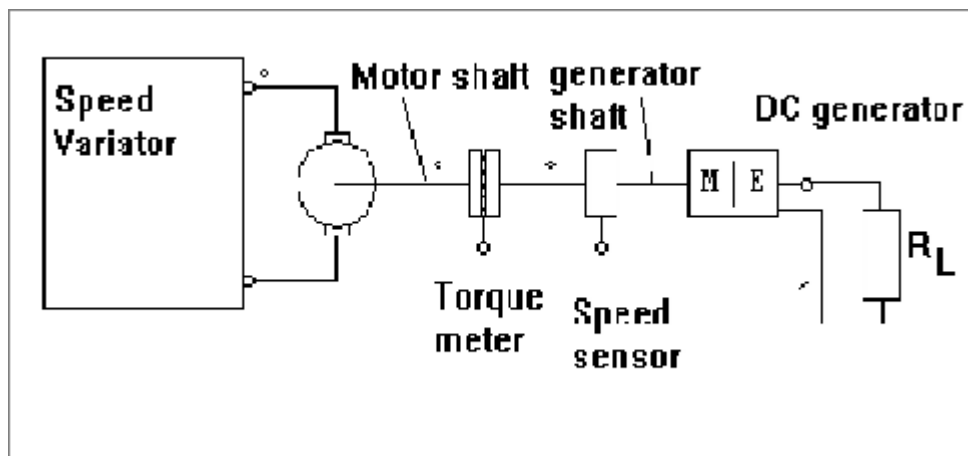


Figure 4.8: measuring chain of torque.

In our case, we have in automatic laboratory DC motors, DC generator and a Speed sensor, but still we don't have for the moment the torque meter. Knowing the nominal electromagnetic torque and the nominal current, we can use this in a first calibration. So, first we begin with the procedure A, where we will try to calibrate speed potentiometer in its minimal position which corresponds to 0 speeds.

- A) We turn the torque potentiometer so that it corresponds to the nominal value of the electromagnetic torque (Pot_T at maximal position corresponding to V_{TSPmin}) and turn speed potentiometer slightly close to 0 speed (Pot_S at its minimal position corresponding to V_{SSPmax}). At first time motor was not rotating, so there is no armature current yet present which means V_s is 0. Then PWM controller will produce maximal duty cycle, in each period of commutation, which imply that we will have a transient

time, where motor's EMF is null, which mean that a higher current will be developed, but as soon as the current reach the fault current value ($2 \times 6,8A$), the motor is protected by armature voltage suppression induced by the IR2133 during some commutation periods. So as long as transient time is not finished³, speed continues to grow even if motor's speed potentiometer was adjusted to its minimal value. Still we don't have increase the electrical load of generator, the armature current will then diminish (the transient time take only 1 to 2 seconds, a simulation with PSIM have demonstrated it) and once, we are in the permanent regime, we will have a maximal speed slightly higher than nominal speed, and ever we turn the speed potentiometer or no, nothing happens and speed doesn't change. So we let speed potentiometer slightly close to 0 speed, and then, we increase electrical load, and measure armature current, until it reaches the nominal value of $6,8A$. Notice that as soon as armature current reach the value of $5 A$, Duty cycle and also speed begin to diminish. When we notice that, we increase electrical load so that average armature current reach its nominal value. In this step, we must be care that motor doesn't stop rotating, then we must turn slightly speed potentiometer, so that motor continue to turn. Once, average armature current has reached its nominal value (in our case $6,8A$), this terminate the procedure A.

B) We verify if turning back the speed potentiometer in its minimal position will stop motor. If it is the case, we repeat the previous procedure A and begin procedure C, if not, that mean $V_{SSP_{max}}$ is too low, so we must adjust R_1 , slightly, until motor stop rotating and we repeat the previous procedure A. Procedure B terminate the calibration of the minimal speed and begin the procedure C of the calibrating of nominal speed.

C) We turn speed potentiometer until speed reaches its nominal value $1500rpm$. 2 cases can be presented:

1-If speed doesn't reach this value, that mean V_c voltage is lower than it must be, so 3 actions may be possible:

a- Either, $V_{SSP_{min}}$ is not too low, so we must diminish $V_{SSP_{min}}$, by changing the value of R_2 resistor⁴, so that speed reach the nominal value $1500rpm$.

³ Transient time is related to basically to the mechanical and electrical characteristics: Inertia moment, friction of the motor, R_a and L_a .

⁴ In this phase of calibration all resistors R_1 , R_2 , R_3 and R_4 must be adjustable.

b- Either, V_{TSPmin} is too high, so corresponding to a lower torque.

c- Either the more complicate case, where the two potentiometers are not calibrated.

For simplify the analysis, Let us choose the case a. Doing that and one time speed has reached the nominal speed value suppose that Torque potentiometer is supposed calibrated at its nominal position.

Choosing case a, we will now turn potentiometer Pot_T slightly, within changing the electrical load of generator and take care that speed do not change a lot (not exceeding too high the nominal value, and not goes too below the nominal value of speed, risking stop motor). We continue to turn Pot_T and lower electrical load until average current reaches $1,7A$ ($I_{anom}/4$),

and speed is a little bit higher than nominal speed: $\frac{Speed_{1/4Tnom}}{Speed_{Tnom}} = \frac{1 - \frac{R_a I_{anom}}{V_{anom}}}{1 - \frac{R_a I_{anom}}{4V_{anom}}} = 1,2$. If it is

the case, that mean that torque potentiometer is good calibrated, if not, that mean that case a, we've chosen, was wrong or not too right we have either the case b or c.

Finally, we think that we demonstrated the complexity of the calibration work, and may be other strategies must be used before applying voltage on motor. Another more simple method is to control the PWM card controller and verify the threshold voltage of Pot_S and Pot_T studied in chapter 3, within an oscilloscope, a DC alimentation and a voltage divider that will vary the sensing voltage before calibration.

4.5 Conclusion:

In this chapter, after showing the three PCBs that we made and then proposing a design of panel, we studied the calibration of the motor's speed and torque. We also discussed the different case that we might get and the explanations of each of these cases.

Lists of acronyms and abbreviations

EMF: Electromotive Force.

T_{ri} : drain current rise time.

T_{doff} : turn off delay time.

T_{don} : turn on delay time.

T_{rv} : discharging time le temps de la decharge de C_{gd} et la augmentation de V_{ds} .

T_{fi} : drain current fall time.

T_{rr} : reverse recovery time.

PWM: pulse width modulation.

DT: Dead Time.

HS: High Side.

LS: Low Side.

T_e : electromagnetic torque.

V_{ssp} : Voltage speed setpoint.

V_{tsp} : Voltage torque setpoint.

PCB: Printed Circuit Board.

Q_G : total gate to charge.

Q_{GS} : gate to source charge.

Q_{GD} : gate to drain.

Q_{rr} : reverse recovery charge.