ACKNOWLEDGMENTS

We express our recognition to all people that by their listening, their advices, their kindness and their support made possible the fulfilment of this project. Thanks to them the work was realised in pleasant conditions and we never forgot this enriching experience.

First we thank **Dr PETRIDIS** our supervisor. Despite a busy timetable he always help us and give us information to well advance on each part of the project. He introduced us to a lot of people to make our integration easier. Moreover his good mood made the work very pleasant.

Mr HATZAKIS for all his devices concerning the electronic parts. He checked our design giving us a lot of information to improve the circuits although he had few time to devote us. Moreover he helped us to weld the boards.

Ms KOUROUTAKI who give us all the details concerning the administrative process. She also provide us the mail of others student to ease our integration.

Mr ZERVOUDAKIS , one of the TEI technician who settle the network connexion on our computers and provide us all the required software.

Finally **Dr TATARAKIS,** the electronic department's responsible who welcomed us and accepted to fund our project.

We more generally thank all the staff and the students of the TEI, especially Hassan and Eugene. We shared a good time in their company.

Table of contents

Introduction ____ 4

I : The semiconductor diode lasers ____ 5

1) Applications, why to use diode lasers ---- 5
2) Physics bases needed ---- 5
a) Absorption and emissions 5
b) Semiconductor structures 7
3) Principe of a laser diode ---- 12

II : ECDL System ____ 17

1) Introduction : brief history and advantages of ECDL system ---- 17
2) Principe ---- 17
a) The feedback effect 17
b) Structure of an ECDL system 19
3) The different components of our ECDL ---- 21
a) The laser diode 21
b) The anti-reflexion coating 23
c) The lens 23
d) The grating 24
4) The design of our ECDL system ---- 27
a) Bases of mechanical 27
b) General description 28
c) The diode mount 31
d) The rail and the grating mount 32
e) The cover 34

III : Design of the current source for the laser diode system ____ 36

1) Specification and constraints ---- 36
2) Description of the current source schematics ---- 37
a) First circuit 37
b) Second circuit 38
c) Description of the selected circuit 42
d) Critical points of the selected circuit 45
e) Modifications of the original schematic 48
f) Choice of the components 52
α) choice of the operational amplifier 52
β) choice of the current amplifier 53
χ) choice of the voltage reference 53
δ) choice of the resistances 55
g) The sources of current instability 60
h) Construction and test of the circuit 61

IV : Design of the temperature controller ________________ 63

1) Specification and constraint --63
2) Theory of the Peltier effect--63
a) The different heat transfer modes 64
α) Radiation 64
β) Convection 64
χ) Conduction 65
b) The physical equations of the Peltier effect 66
α) case of one conductor (thermo electrical coupling) 66
β) case of the TEC structure 68
3) The different electric architecture to control the Peltier element ---------71
a) First structure : the proportional control..71
b) Second structure : the PI control 79
c) Third structure : the PD control...81
d) Fourth structure : the PID control..83
e) Description of the PID electronic circuit..84
α) Error detection circuit 85
β) PID structure 91
-Proportional stage 93
-Integrator stage 94
-Derivation stage 95
χ) The conversion voltage/current 96
δ) Displaying ports 97
ε) The critical points of the temperature controller 98
-characteristics of the resistances of the detection bridge 100
-choice of the voltage reference 100
-considerations concerning the instrumentation amplifier 101
-considerations of the others points of the circuit 102
f) Adjustment method of the PID parameters.......................................103

V : The Fabry Perot interferometer ______________________ 103

1) Principe -- 103
2) Design--- 107
3) The photodiode circuit-- 109

Conclusion____ 114

Bibliography ____ 115

Annexes____ 117

1) - Autocad--- 117
2) ECDL Dimensions-- 120

Introduction

Our project consist in the development and the characterisation of an External Cavity Diode Laser system. This work include the design of the External Cavity, the development of the electronic circuits which drive the laser diode and finally the construction of one Fabry Perot interferometer.

The goal of the External Cavity is to improve the selectivity of the laser diode and to make a tuneable laser system for spectroscopic applications.

The electronic part include two cards, one which control the current and another one which control the temperature. The laser diode is sensitive of the variations of the current and the temperature thus the current source and the temperature controller have to be as accurate as possible to keep stable the laser performance. The current source can also make the laser diode tuneable.

The aim of the Fabry Perot interferometer is to characterise the system by analysing the output beam spectrum.

I : The semiconductor diode lasers

1) Applications, why to use diode lasers

Semiconductor Diode Laser was first developed more than forty years ago and is the most famous of all lasers. It has a wide range of applications and can be found in a lot of different devices because it emits coherent light that can contain information or provide energy.

Diodes Lasers have many advantages like a wide tunability, a low power consumption, an easy modulation and short dimensions. Moreover the noise is quite low and the hermetic sealing of the package assures high reliability.

Here are some examples which show the importance of this component :

The diode laser is used to read and etch optical Cds. The component produces the laser which is focalised on the surface of the cd composed of reflective metal cover by a photosensitive colorant. With a high power the heat of the laser can make holes in the colorant (etching) and with a lower power the beam is reflected and is analysed by a photodiode (reading).

The laser diode can be found in optical telecommunications. The light is used like a communication mount, the fibre optic ships the beam and multicoated filters, lens and mirror net add the beams with different wavelength or separate them.

Tuneable Laser diode is useful for precise spectroscopy measurements. The beam goes through a liquid or gas and the output light is analysed. The tunability allows to scan the absorption in a large wavelength range. The obtained spectrum indicates the composition of the analysed sample.

2) Physics bases needed

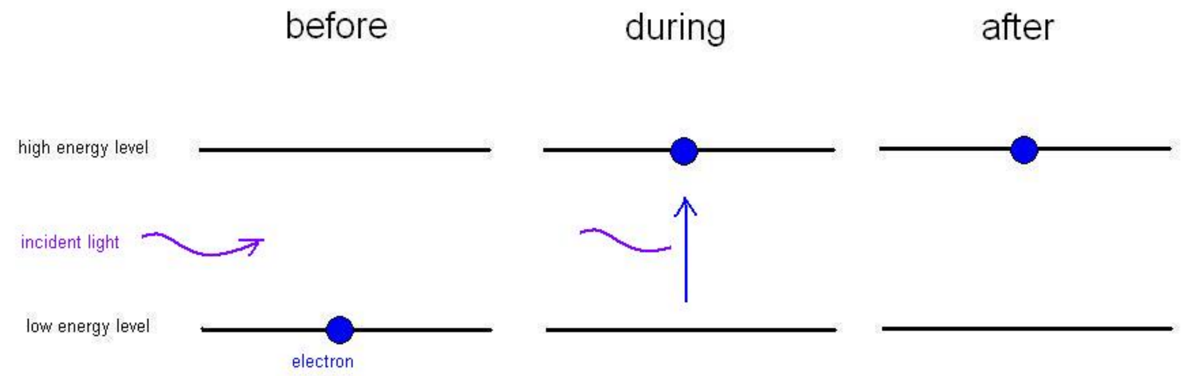
a) Absorption and emissions

Any electron in an atom has a specific energy level. When an electron change of energy state the atom radiates by light emission with a frequency $\nu = \frac{Ef - Ei}{h}$ with the Planck's constant h = 6.62.10-34 J.s., Ei the initial Energy level and Ef the final one. The time the electron takes to change of level is call lifetime

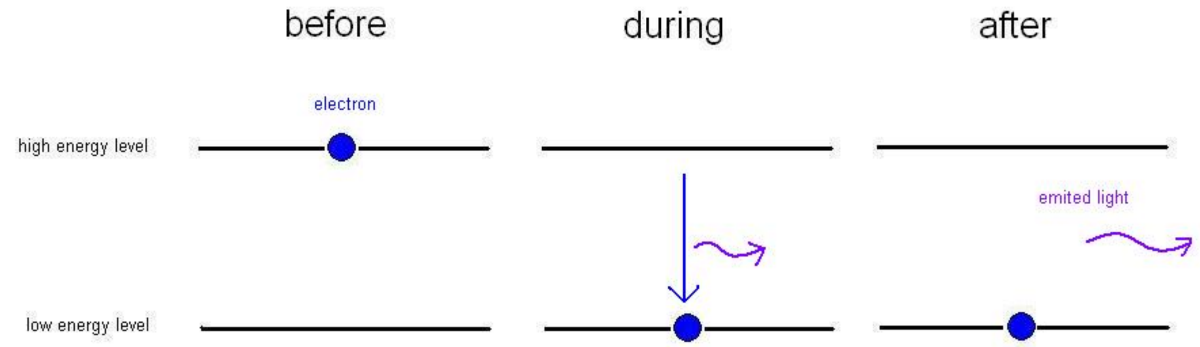
The electron can pass from a stationary state to another with 3 different fundamental radiations :

- resonant absorption
- spontaneous emission
- stimulated emission

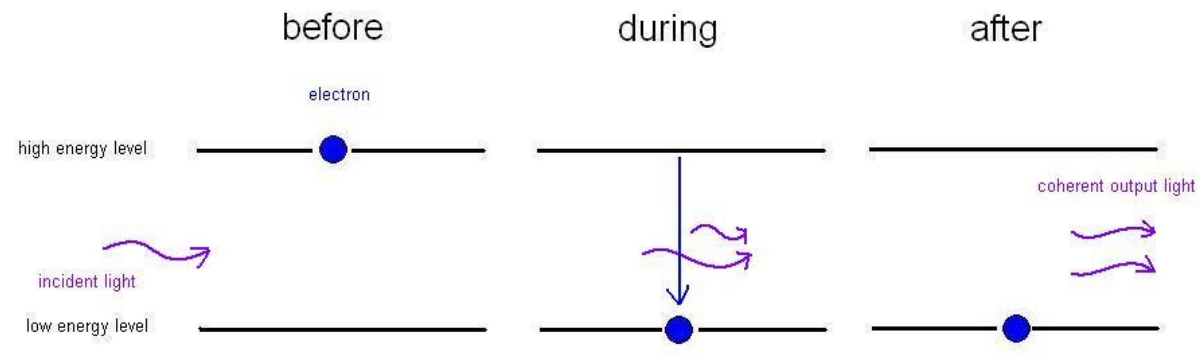
In resonant absorption the electron is first in a low level energy. We send an incident light with a frequency corresponding to the difference between two energy states. The electron can absorb this light energy to pass from the low level to the high level.



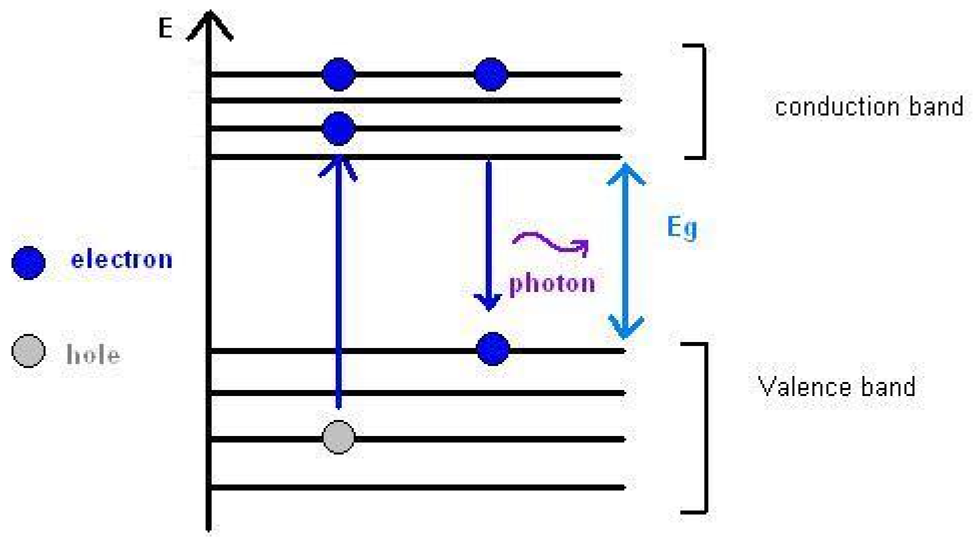
In spontaneous emission the electron has already jumped in a high level. It tends to go to the lower stable level,. To compensate the difference of energy the electron emits light whose frequency is $n = Ef - Ei$.



In stimulated emission the electron is first in the high level energy, we stimulate the transition between high level and low level sending incident light. A new wave is generated and added to the incident light. The output light has the same direction and phase that the incident light, it’s a coherent light.



In a material there are more than two energy levels, the density of state g(E) indicate if the energy levels are closed are not.



In a stable configuration most of the electrons are in the lowest bands called Valence band. When energy is provided into the material the electrons can pass into the highest bands called Conduction bands. But this energy must be superior to the gap between the Valence and the Conduction bands to permit the excitation of the electrons, the value of this gap is called Eg. Eg is small for a conductor, high for an isolator, between the two for a semiconductor. As the electron has left the Valence band, an hole is created.

After about 1ps, electrons in the conduction band drop to the lowest unoccupied levels. The electrons which pass from the Conduction band to the Valence band combine with the holes. These combinations produce photons.

b) Semiconductor structures

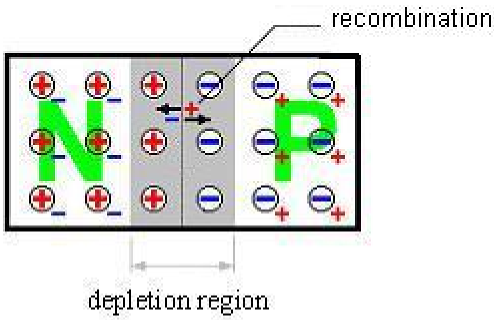
Two important conditions are needed to produce laser :

- More electrons must be in the higher states than in the lower. This case is called population inversion.
- We need more stimulated emissions than spontaneous emissions.

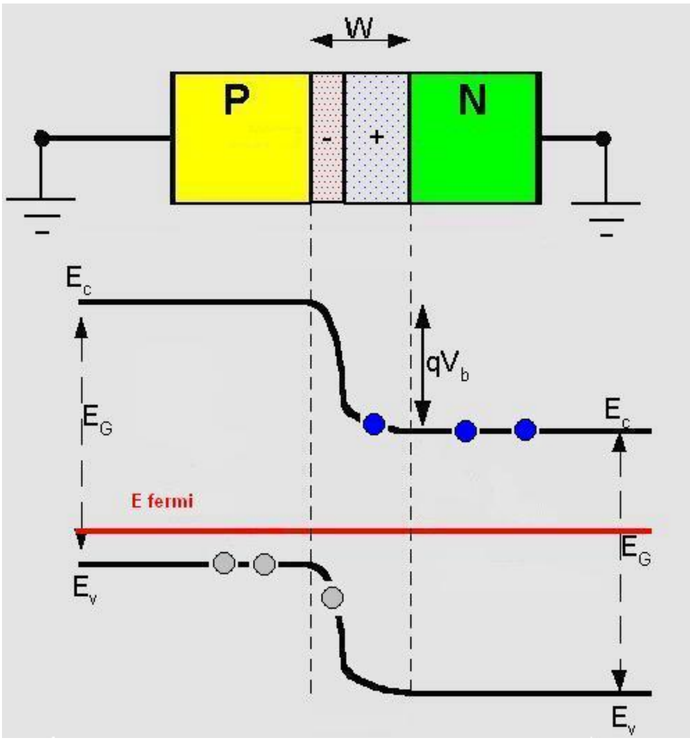
There are three main kinds of structure for the semiconductor diode laser :

- Homojunction structure
- Double-heterostructure
- Quantum well structure

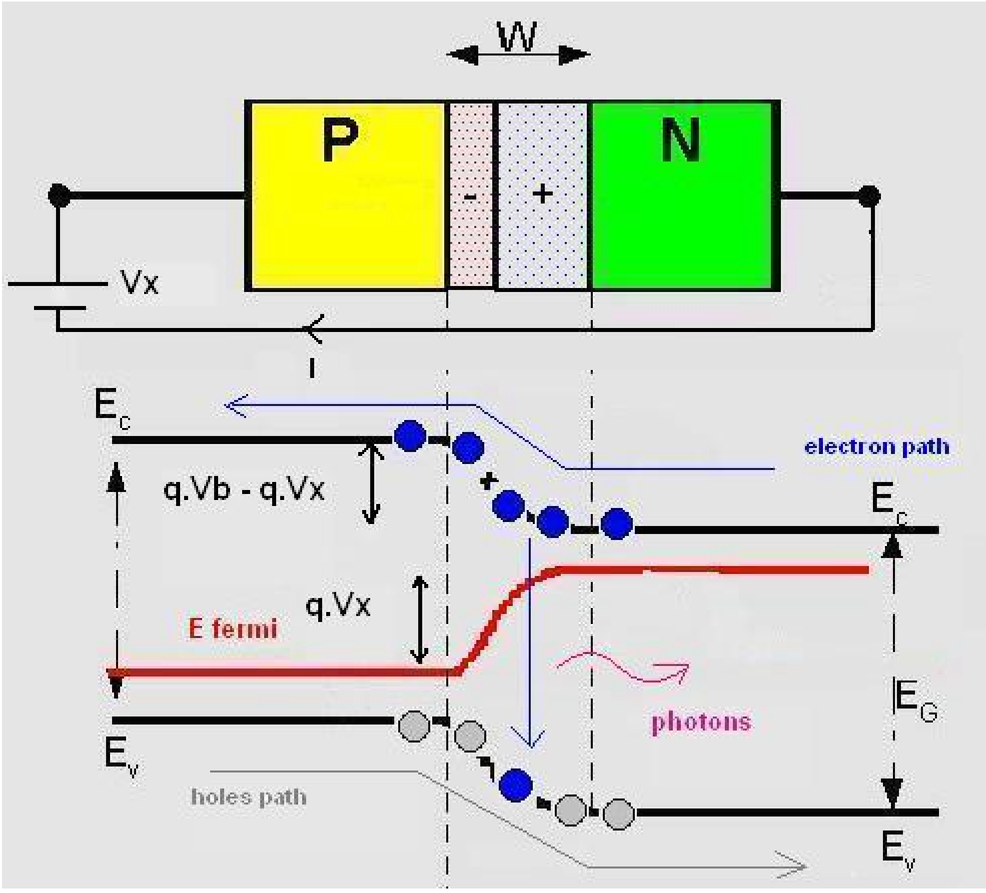
In Homojunction structure two materials are used (a p-type and a n-type) with the same band gap Eg. When the two materials are put in contact, at the boundary some electrons of the n-type part combine with the holes in the p-type part. Therefore, positive charges appear in the n-type side and negative charges appear in the p-type side. A carrier-depletion region (W) is thus created.



The charges densities on each side of the depletion region induce an electric field $E = q.V_b$ with V_b the barrier potential. The field avoids all the electrons of the n-type part to go toward the p-type part. So the system becomes stable in few times.

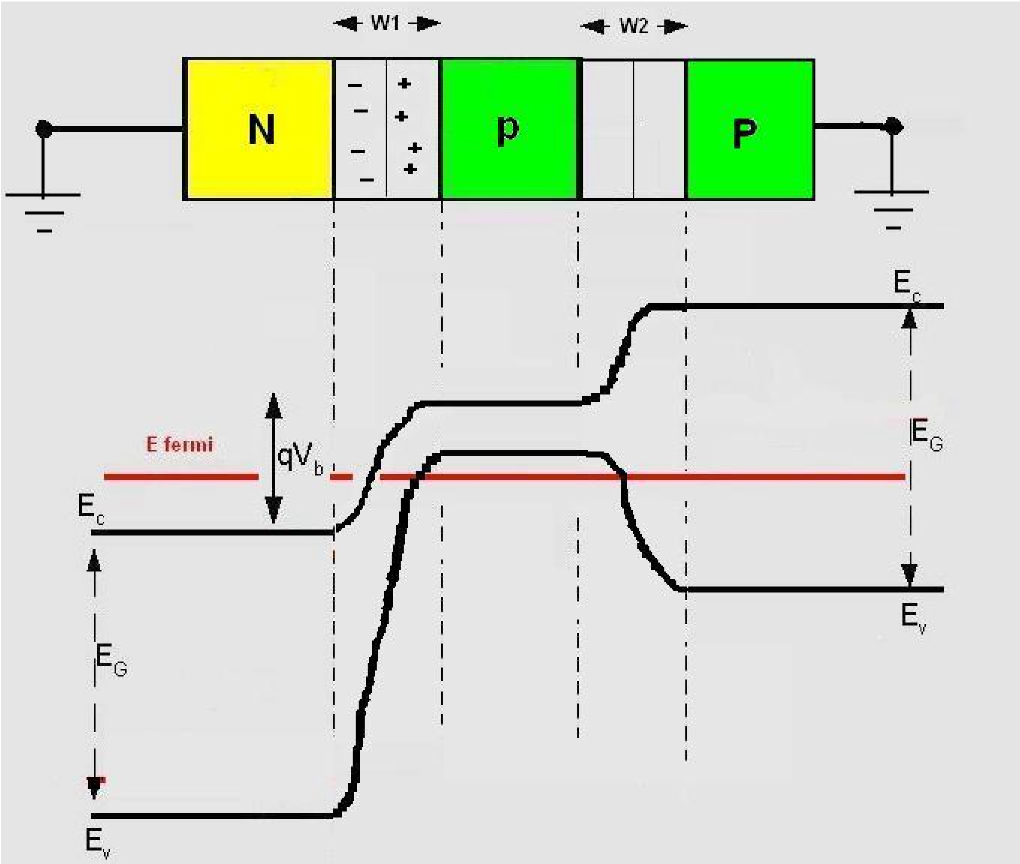


If we apply bias current in the system, the balance is broken and most of the carriers can pass the depletion region. Many of electrons combine with the holes and generate spontaneous emission.

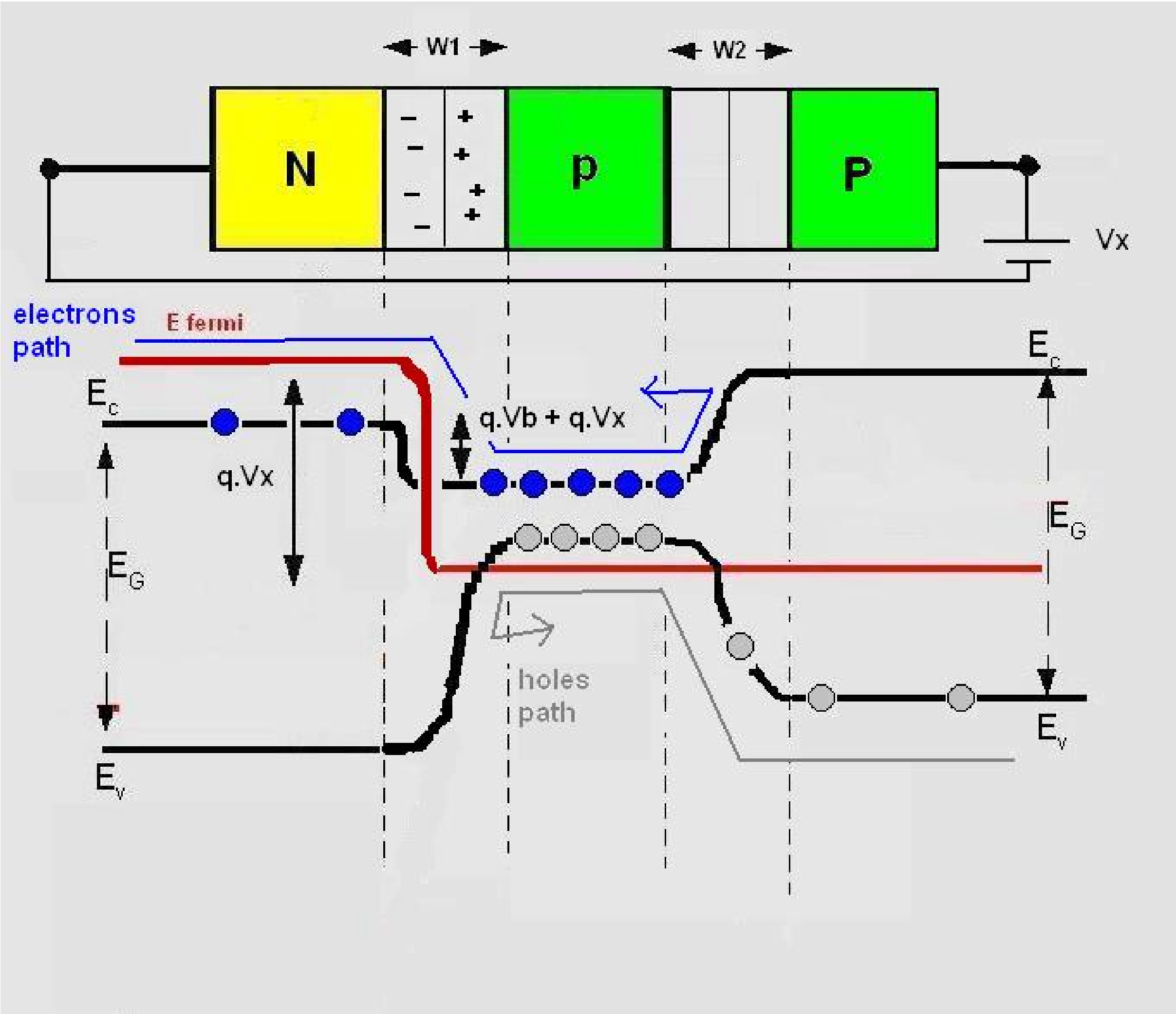


For lasing we need a high electrons and holes densities into the conduction and valence bands. The problem is that with a single material the carriers quickly diffuse away from the junction so we need an extremely high current at room temperature (more than 1000 A) which is impossible for practical use.

In Double-heterostructure we have 3 parts (N-p-P for example) with three different band gaps, the band gap of the middle layer is smaller than the band gaps of the others. When the layers are put in contact, two depletion regions appear at each side of the middle layer.



If we apply bias current, the electrons are injected into the small band gap active region from the n doped cladding layer whereas the holes come from the p cladding layer. The injected electrons and holes are confined in the active medium because they cannot climb the high band gap of the cladding layers.



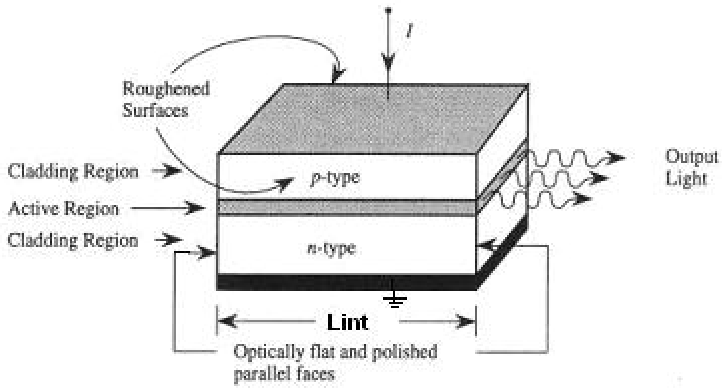
In the middle layer there is a population inversion, a lot of combination produce the lasing. Moreover the change of index of refraction between the active layer and the claddings layer provides a guide for the laser beam like in the fibre optics. These properties decrease the current needed and make the laser useful for practical use.

The Quantum well structure look like the Double-heterostructure, the main difference is that the active layer dimension is at an atomic scale (few nanometres). Quantum effects occur only in one dimension. Fewer states have to be filled to reach population inversion and we need fewer electrons which allow reducing the current. Moreover the spectral width of this laser can be smaller than double-heterostructure.

In Multi-Quantum Well structures there are more than one active layer so the features of quantum well structure are improved.

3) Principe of a laser diode

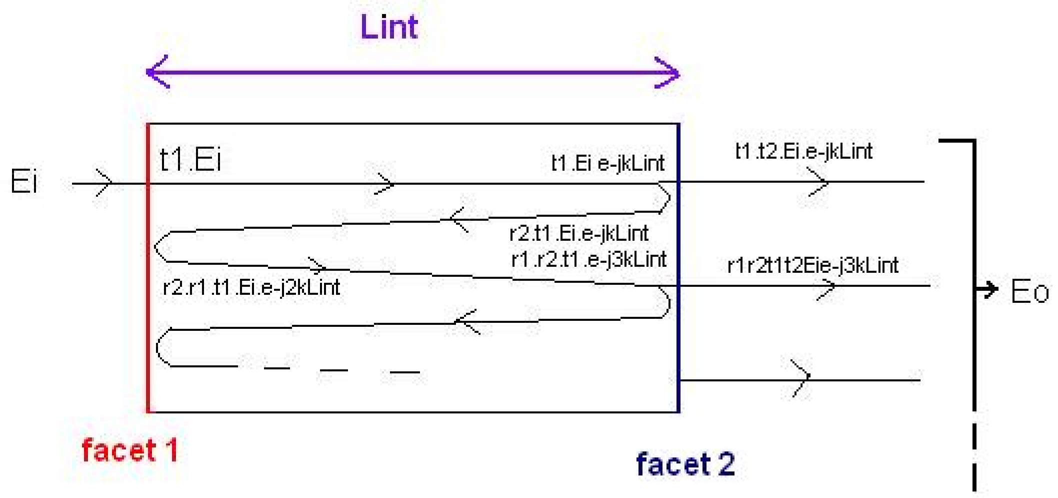
Here is a schema of the laser diode :



The stimulated emissions occur into an active region between two or more cladding layers. A pumping of atoms system does the population inversion. In our case the pumping consist in sending current through the active region of the diode between the n- and p-type cladding layers. The wavelength of the emitted light depends of the material band gap ($\nu = Eg / h$).

A resonant system is needed to produce a feedback that increase enough the output power of the laser. In the Fabry Perot configuration the amplifier middle is surrounded by two mirrors. The mirrors are less than 100% reflective to allow the beam to go outside. To change the output power we can apply optical coating on the diode facets.

Now suppose that we emit an incident light at the left side of the laser cavity of index of refraction n. This light is a plane wave with a propagation constant k and an amplitude E_i.



When the wave just pass the first facet, the amplitude become $t_1.E_i$ with t_1 the ratio of transmitted light of the left facet. If Lint is the length of the cavity, just before the second facet we have an amplitude of $t_1.E_i\, e^{-jkLint}$ (if we considerate the wave independent of time).
This wave is divided in two waves by the second facet, one of them is reflected with the amplitude $r_2.t_1.Ei.e^{-jkLint}$ and the other pass the facet with the amplitude $t_1.t_2.E_i.e^{-jkLint}$ (with r_1 the reflective index of the first facet).
The wave which come back arrive at the first facet with an amplitude $r_2.t_1.Ei.e^{-j2kLint}$ and is reflected with the amplitude $r_2.r_1.t_1.Ei.e^{-j2kLint}$. This wave arrive at the second facet with an amplitude $r_1.r_2.t_1.e^{-j3kLint}$ and a part of it go out with the amplitude $r_1r_2t_1t_2E_ie^{-j3kLint}$.The wave reflected by the facet come back towards the first facet and so on.

The sum of all the amplitude waves at the output is :

$$Eo = Ei \times \frac{t_1.t_2.e^{-jkLint}}{1 - r_1.r_2.e^{-j2kLint}}$$

To create oscillation, we must have Eo different of 0 when Ei = 0, it is obtained only if the denominator equals 0.

Hence $r1.r2.e^{-j2kLint} = 1$

Moreover k = b + j(G . g – a) with a the internal attenuation per unit length, b the propagation coefficient equal to (2pn) / l, g is the gain per unit length and G the confinement factor.

The confinement factor depends on the size of the active medium and on the refractive index of the different layers. The ideal confinement is obtained when G = 1, the worst when G = 0.

So the equation become $r1.r2.e^{-j2bLint}.e^{2(Gg-a)Lint} = 1$

So r1.r2.e$^{2(Gg-a)Lint}$ must be upper than 1 that correspond to a gain g upper than $\frac{\alpha + \frac{1}{2d} \times \ln(\frac{1}{r_1 r_2})}{\Gamma} = \frac{\alpha + \alpha_m}{\Gamma}$ and 2.b.L_{int} must be equal to 2.p.q (a_m are the mirror losses).

So in order to create a coherent light we have two conditions :

- the gain in a round trip must be upper than the losses (g >= $\frac{\alpha + \alpha_m}{\Gamma}$) .
- The phase in a round trip must be equal to **2.p.q** where q is an integer

2.b.L_{int} = 4.p.n.L_{int} / l = 2.p.q

So q = 2.n.Lint / l

Dq = - (2.n.Lint / l^2) .Dl

If Dq = 1 then $\Delta\lambda = -\frac{\lambda^2}{2 \times n \times L_{\text{int}}}$

We have a quantification of the wavelength.

As the frequency n = c/l we can deduce Dn = - (c / l^2).Dl with Dl = -l^2/(2nL_{int})

Therefore $\Delta\upsilon = \frac{c}{2 \times n \times L_{\text{int}}}$

The space between two possible frequencies that can produce coherent output beam is called **F**ree **S**pectral **R**ange (FSR).

FSR = Dn = $n_{q+1} - n_q$ = c / (2nLint)

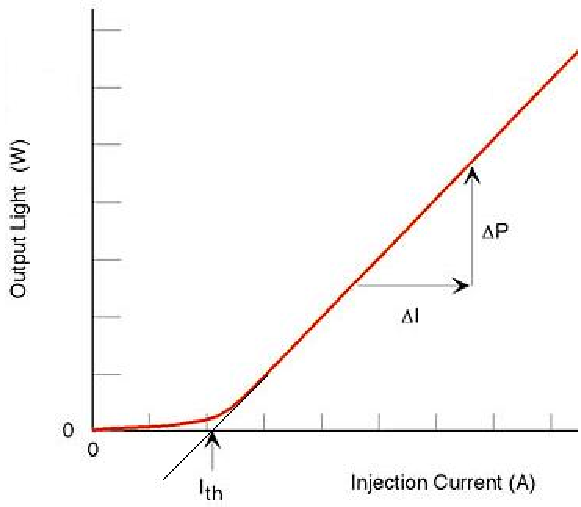
So to have a high FSR we have to choose a diode with a little amplifier middle length, but the power needed constraint the lower limit of this dimension.

The gain as a function of wave length is very difficult to determine, the equation is :

$$g(\lambda_\mu) = \frac{\alpha \times N\left[1 - \left(2 \times \frac{\lambda(t) - \lambda_\mu}{\delta \times \lambda_g}\right)^2\right] - \alpha \times N_o}{1 + \varepsilon \sum_{\mu=1}^{M} P_\mu}$$

The shape of the curve is a flat parabola.

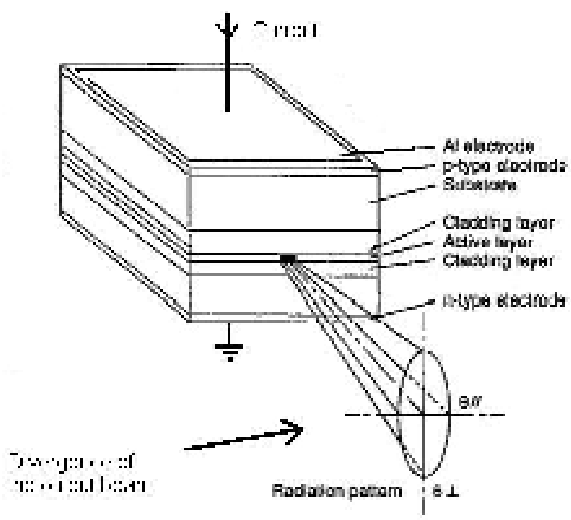
Another important parameter is the output power of the diode. Below is the L-I curve which represent the output power as function of current and temperature.



When the current is low, the diode operate like a LED with a majority of spontaneous emission. If the current is high enough we enter in laser mode, population inversion occur and power quickly increase with intensity for a little range of intensity. If the current is high enough the power increase proportionally with the current. The limit I_{th} which separate the two behaviours is called threshold current and is determined by extrapolation of the linear part.

If the temperature change for a same current the output power can be different, that is one of the reason why to control the diode temperature is important.

The tiny dimensions of the diode give a diffraction effect at the beam, it results a divergence of the emitted ray.



The vertical and horizontal divergences (θ_{vert} and θ_{par}) are in general different. An optical system can correct this problem.

Another problem is the astigmatism of the laser due to the imperfect surfaces of the materials. This phenomenon provoke a misalignment of the different rays and it result a fuzzy effect. The magnitude of the astigmatism distance depends on the radius of curvature of the wave front, the wavelength, and the width of the beam near the output mirror.

A diode not only provide one harmonic because the fabrication process is not ideal, in fact the spectrum contains some picks near the expected wavelength. The mechanical vibrations and the quantum fluctuations explain the broadening of the laser spectrum. The line width of the laser is given by $Dn = \frac{R}{4 \times \pi \times N}$ with R the total spontaneous emission rate and N the total number of photons. The line width can be measured with a Fabry Perrot Interferometer. This feature in some applications is too bad and avoid to only use a diode laser to provide the laser beam. Other components are needed to decrease the line width, we will describe it in the next part.

II : ECDL System

1) Introduction : brief history and advantages of ECDL system

An ideal tunable single mode laser would be one than can be tuned continuously over the gain bandwidth without any reduction in the output power and which produce an infinite narrow line width. All the diode laser have not such a good features, it is possible to change the composition and shape of the diode to make it tunable but the process is often too difficult for many applications.

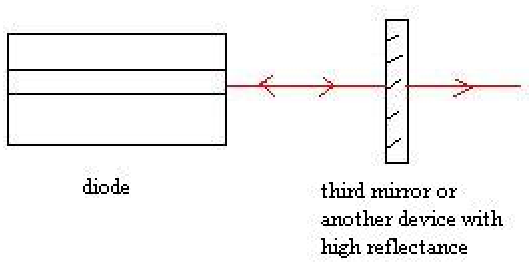
To tend toward an ideal case we have the possibility to add an External Cavity Diode Laser (ECDL) system that is a quite simple device to design and provide proper results. ECDL was first used in 1964, and studied since the 70s. Intensive researches have been done in the 80s to use ECDL as a transmitter and local oscillator in the optical telecommunications. In the early 90s ECDL systems were for the first time used for spectroscopic measurement and for commercials fibbers. Nowadays this device is mainly use for telecommunications applications and research for micro-electromechanical system.

Typically the line width we can obtain with only a laser diode is 10-15 MHz, with an ECDL system it is possible to achieve less than 1MHz. The resolution of the output beam depends of the laser diode, the cavity length and the ECDL filter features. The FSR of the external cavity Dn_{ext} is in general about Dn_{int} divided by 100.

2) Principe

a) The feedback effect

The ECDL uses the sensitivity of the laser to optical feedback. We add a third mirror in the system which reflects a part of the output beam toward the laser medium.

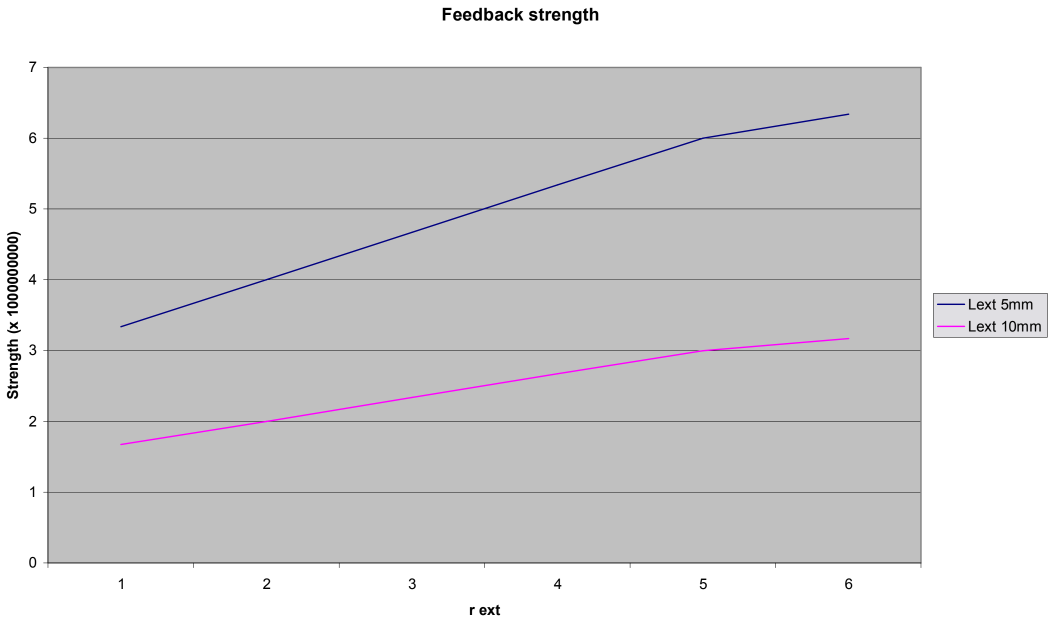


When photons come back into the gain medium of the laser diode they generate more carriers which increase the net gain. If we well choose the feedback strength we can make the line width narrower.

The feedback level is **20.log r_{ext}(l)** with r_{ext}(l) the reflectivity coefficient of the external device that produces the photons backlash. The feedback strength is

$$\xi = \frac{(1 - r_2) \times c \times r_{ext}}{2 \times L_{ext} \times r_2}$$

Below is the curve of feedback strength as function of the reflectance r_{ext} for different cavity length L_{ext} with a typical value of 0,31 for r_2.



The feedback has five regimes depending on its strength, for each regime we have a different impact on the diode laser system and so on the output beam.

Regime I is the lowest regime of feedback with a level less than –90dB. In this situation frequency stability is improved but we can observe a narrowing or a broadening of the output beam, it depends of the phase of the feedback.

At the second regime, we can observe a splitting of the emission line depending on the distance of the external cavity and of the strength of the feedback.

Regime III deletes the mode hop effect, and the laser is on a single narrow line. The problem is that the regime is between –45dB and –39dB so the reflectivity has to be well controlled.

The fourth regime between –39dB and –10dB has a broad laser line width because satellite modes appear. The effect is independent of the phase.

Regime V at a level upper than –10dB offer a narrow line width. In general it is necessary to coat the laser facet to reach this level. With a high frequency selectivity in the laser the line width stays narrow for all the phases of the feedback. Moreover this level is insensitive to other reflections.

For our application the most important is to produce a line width as narrow as possible and not depending of the phase, so we can choose between the third and the fifth regime. The fifth is probably the best because we have not to well control the reflectivity. To reach the fifth regime we have to keep L_{ext} at a low value and to find a device with a high reflectivity.

Some features of the diode can change with the external feedback :

- The threshold current of the diode is influenced by the feedback, for a feedback F the difference is :

$$\Delta I_{th} = -I_{th} \times \frac{\ln(1 + \frac{1 - R_2}{R_2} \times F)}{2\alpha L + \ln(\frac{1}{R_1 R_2})}$$

- In the gain equation we must replace the reflectivity index r_2 by r_{2eff}(l) which takes in consideration the reflectance of the third mirror r_{ext}.

$$g = \frac{\alpha + \frac{1}{2d} \times \ln(\frac{1}{r_1 r_{2eff}(\lambda)})}{\Gamma} \quad \text{With} \quad r_{2eff}(\lambda) = \frac{r_2 + r_{ext}(\lambda) \times \exp(\frac{i \times 4\pi \times L_{ext}}{\lambda})}{1 + r2 \times r_{ext}(\lambda) \times \exp(\frac{i \times 4\pi \times L_{ext}}{\lambda})}$$

- The phase condition become $\omega \times \tau_L + \phi(\omega) = 2 \times m \times \pi$ with $\phi(\omega)$ a phase shift produced by the feedback and depending on r_{ext} and r_2.

$$\phi(\omega) = \tan^{-1}(\frac{r_{ext}(1 - r_2^{\ 2}) \times \sin(\frac{2\omega L_{ext}}{c})}{r_2(1 + r_{ext}^{\ 2}) + r_{ext}(1 + r_2^{\ 2}) \times \cos(\frac{2\omega L_{ext}}{c})})$$

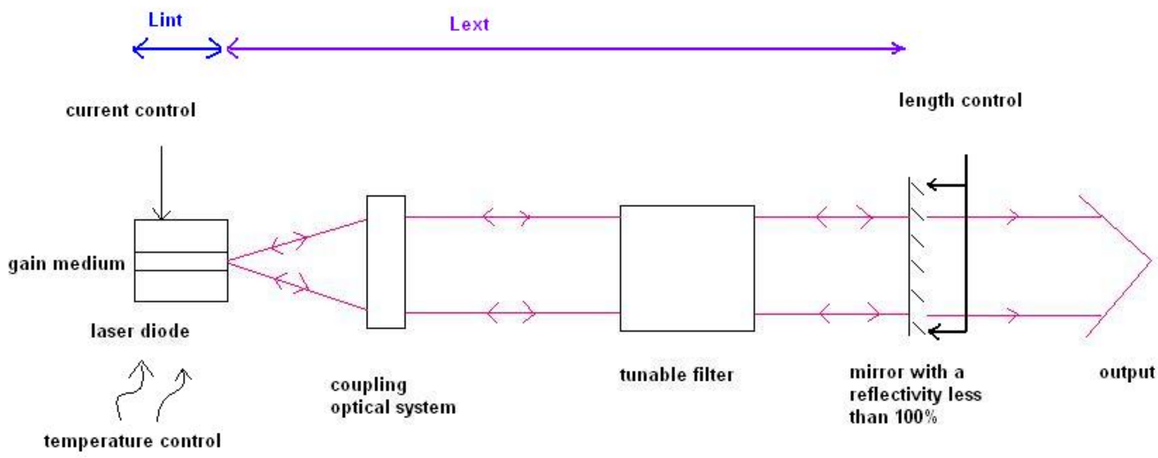
We have seen the impact of the feedback on the laser features, now we are going to analyse the different parts to make an ECDL system.

b) Structure of an ECDL system

The ECDL has tree principal roles :

- To collimate the beam that is to say make it horizontal to avoid a too high divergence.
- To filter the laser to provide a narrow line width.
- To provide a feedback thanks to a high reflectivity device.

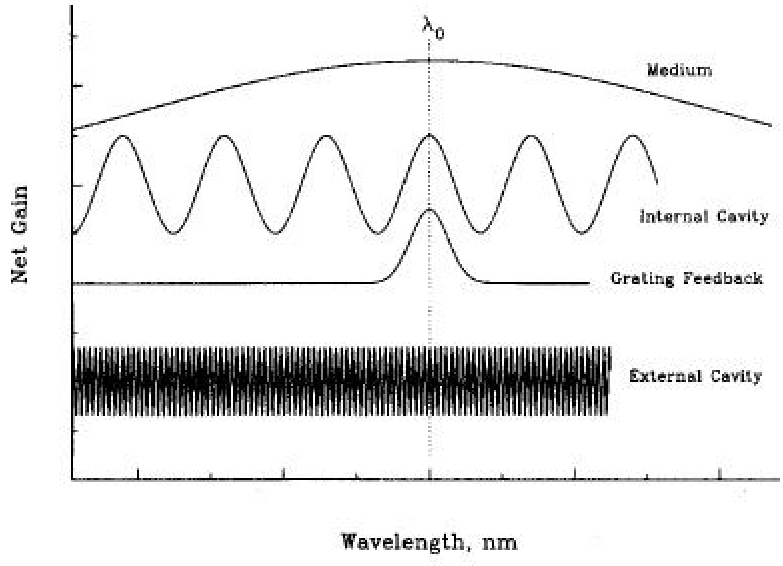
Below is an example of ECDL system architecture :



An optical system (one or more lens and eventually beam expander system) make the collimation and then a filter is placed into the light trajectory to make the mode selection (to choose the frequency of the output beam). The mirror reflects a part of the beam to make the feedback and the other part is used as output beam.

The tunability is made by the filter and the control of the external cavity length. The current and the temperature of the diode can be controlled in order to have a more important tunability or to keep them constant.

The aim to have good output features is to align the gain of the different parts. Thus we can produce a high power output beam at a definite wavelength with a narrow line width.



We can see the parabolic shape of the medium gain with a maximum at a definite wavelength. The external cavity has a FSR inferior than the internal cavity because its length is much more high, it offers a lot of possibilities to align external and internal cavity modes. The filter has only one narrow peak to select the wavelength.

The FSR of the external cavity is given by $\frac{c}{2 \times (n \times L_{\text{int}} + L_{ext})}$.

Here is a table of the FSR for different length L_{ext} :

Length of the cavity L_{ext} (mm)	**FSR (GHz)**
2	49.7
5	24.9
8	16.6
10	13.6
11	12.5
12	11.5
13	10.7
14	10
15	9.4
16	8.8
17	8.3
18	7.9
19	7.5
20	7.1

The line width of the laser is proportional to the photon lifetime, so if we increase the length of the cavity we improve the finesse of the laser. It also depends on the feedback power and on the phase inside the cavity.

The mirror losses of the ECDL system are : $\alpha_{mir} = \frac{1}{L_{\text{int}}} \times \ln(\frac{1}{r_1 \times r_{ext}(\lambda)})$. α_{mir} as a function of wavelength is a sinusoidal ripple. A minimum of a_{mir} correspond to a maxima of r_{ext}. So with the aim to have a strong feedback we will try to minimise these losses.

3) The different components of our ECDL

a) The laser diode

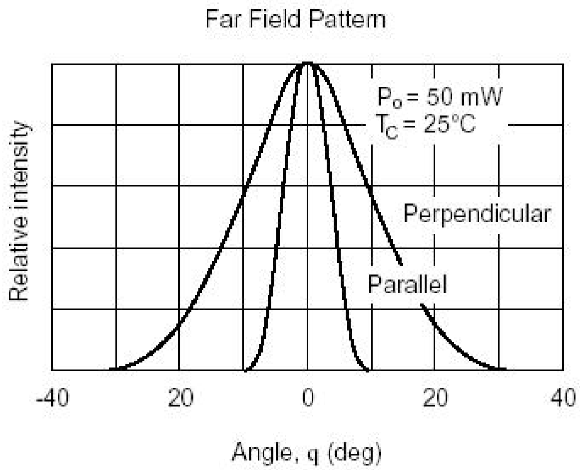
We had bought the HL7851G GaAlAs diode laser whose main wavelength is l = 785 nm ($\nu = \frac{c}{\lambda} = 3{,}82.10^{14}\,Hz$) and the index of refraction is n = 3.4. Its structure is Multi Quantum Well (MQW).

The length of the amplifier middle is Lint = 300 mm , so we can calculate the FSR which is c / (2.n.Lint) = 3.10^{8} / (2 x 3,4 x 300.10^{-6}) = 147 GHz.

The divergence is $\theta_{horizontal} = 9.5°$ and $\theta_{vertical} = 23°$.

Here is the curve of the beam intensity according to the angle which shows the importance to have a filter well aligned with the diode.

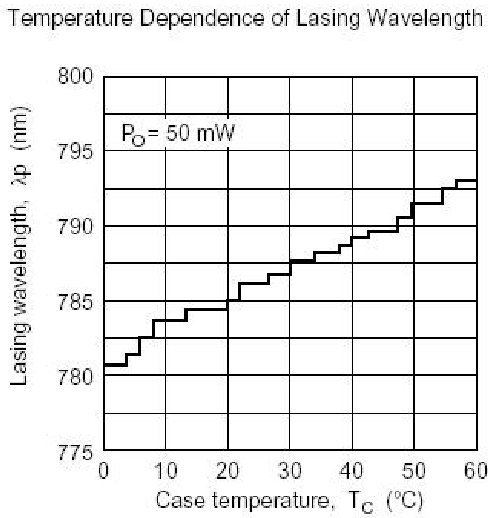
For only 2 or 3 degrees we can lose 1/5 of the beam power.



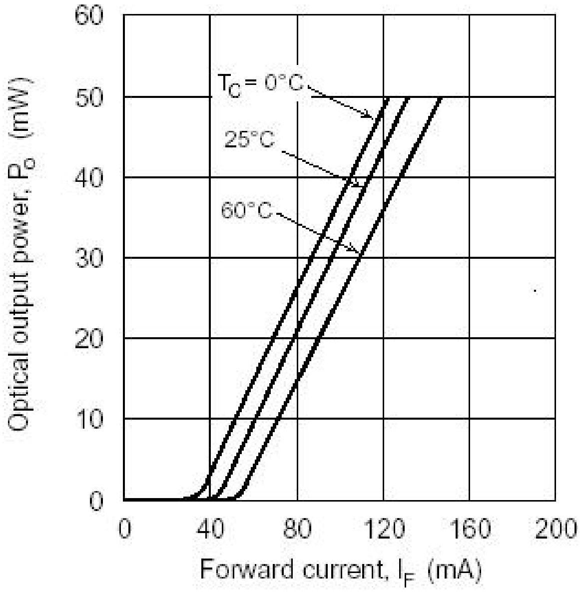
The temperature range into which the diode can operate is : -10°C to 60°C. For our application we will try to keep the temperature at 25°C as accurately as possible.

If we increase the temperature, the peak of both medium and internal cavity gain shift toward longer wavelength but not at the same value, it results mode hops so the diode cannot be properly tuned to any arbitrary wavelength.

Here is the curve of our diode which depicts the phenomenon :



Below is the L-I curve of the diode which defines the threshold current for different temperatures :



The maximum for Pout is 50 mW. At 25° the threshold current is about 50 mA (this value will be affected by the feedback of the ECDL system).

b) The anti-reflexion coating

To avoid tuning nonlinearities and axial mode instabilities the condition is : $r^2_{ext} >> r_2^2$ with r_2 the reflectivity of the diode output facet and r_{ext} the reflectivity coefficient of the external device that produces the feedback. In general we choose r^2_{ext} between 0,1 and 0,3 and r_2^2 is 10^{-3} or less.

It is possible to apply an optical coating on the laser output facet. It has for effect to reduce the reflectance and can improve the performance of the ECDL system decreasing the coupled cavity effect and making stronger the feedback effect.

The refractive index of the coating must be $n_{coat} > n_{amp}$ with n_{amp} the refractive index of the gain medium of the diode. The thickness must be $t_{coat} > l / (4.n_{coat})$

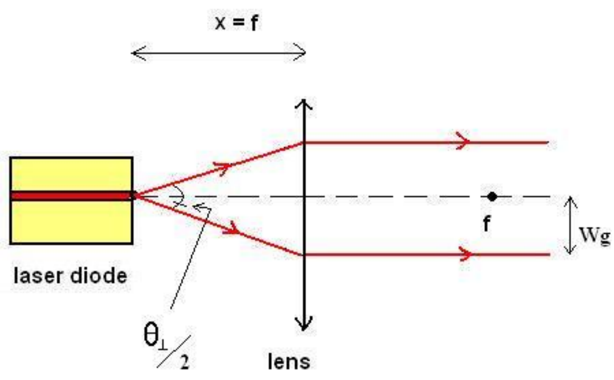
c) The lens

The lens is made of glass, it is better than plastic to keep low wave front distortion.
To well collect the light provided by the diode the lens should have a numerical aperture

$$NA >= \sin(1.7 \times \theta_{vert}/2) = 0.33.$$

We have chosen a C230TM lens whose numerical aperture is 0.55 and focal value is 4.5 mm.

The diode and the lens are both inside a collimation tube (LT230A). The distance x between the emission point of the diode and the lens is 4.52 mm (about the value of the focal in order to avoid divergence), so the ray coming from the diode with a divergence after the lens becomes horizontal :

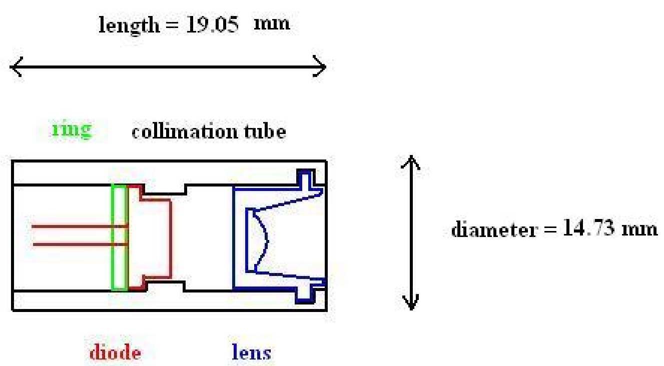


We can estimate the radius of the laser beam using trigonometry formula :

$Wg = x \cdot \tan(\theta_{vertical} / 2)$

In our case Wg = 1.8392 mm.

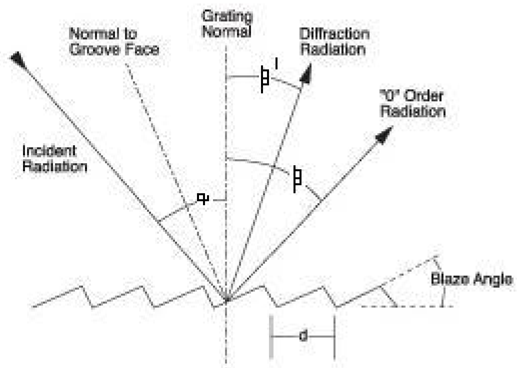
Both the diode and the lens are placed into a LT230A collimation tube



The advantage of using a collimation tube is that it well fixes the distance between the diode and the lens. Moreover it is more convenient to integrate it into the ECDL design because it can be placed into only one plate making one hole.

d) The grating

The grating acts like a filter in the structure. It consists of a periodic variation of the thickness of a medium at a definite refractive index or a constant thickness with a periodic change of refractive index. The GH13-18U grating that we use is in the first grade.



GRATING ANGLES & TERMS

The distance d of the groove spacing is called the period of the grating, it must be about the same than the laser wavelength.

The input beam come on the surface at an a angle with the grating normal. The diffraction occurs at a different angle for each wavelength that the input polychromatic light contains. Therefore the output beam at a definite angle is more pure than the input beam.

The grating equation shows that the angles of the diffracted orders only depend on the grooves' period, and not on their shape. By controlling the cross-sectional profile of the grooves, it is possible to concentrate most of the diffracted energy in a particular order for a given wavelength. A triangular profile is commonly used. This technique is called blazing. The incident angle and wavelength for which the diffraction is most efficient are often called blazing angle and blazing wavelength. The efficiency of a grating may also depend on the polarization of the incident light but in our case we don't need any polarization.

The grating equation is : $q.l = d(\sin a + \sin b)$

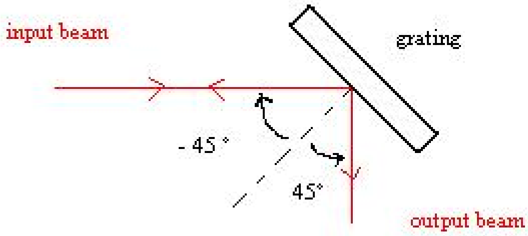
The beam that corresponds to direct transmission is called the zero order, it corresponds to q = 0. The other orders correspond to diffraction angles which are represented by non-zero integers q.

In the Littrow configuration the first order beam (q = 1) come back in the laser cavity, the zero order beam is used as useful output beam. In Littrow configuration we have a = b which simplifies the grating equation :

$q.l = 2.d.\sin a$

So we can deduce the good angle to use for our application which is :

$a = \sin^{-1} (q.l /(2d)) = 44.95°$ for q = 1 (= - 45° for q = 0)



The line width of the feedback (q = 1) depends of the wavelength of the input beam, the formula is : Dn = n / d the value is usually between a dozen and hundreds of GHz.

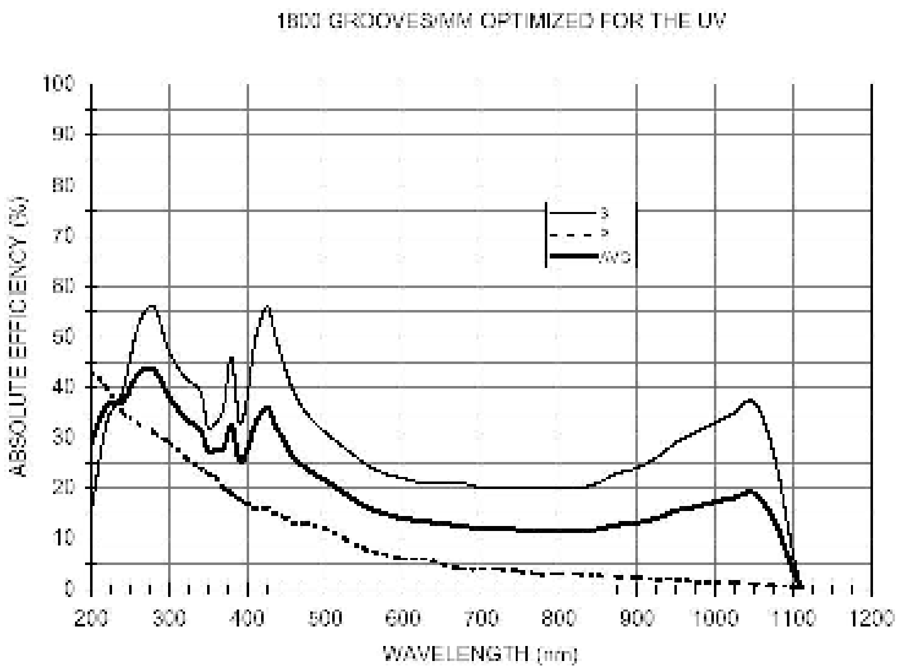
The resolution of the filter can be given by : $\Delta\nu = \dfrac{c}{2 \times \pi \times Wg \times \tan(\alpha)}$

In our case Dn = 1,8 GHz

The resolving power of the grating directly depends of the number N of groove that are lighted up :

$R = \dfrac{N \times 2 \times d \times (\sin\alpha)}{\lambda}$ This value represent the grating ability to separate adjacent wavelengths so the better is to light a large amount of grooves.

The efficiency of the grating is the ratio of the output beam power to the input beam power. It depends of the wavelength and on the polarization (S = horizontally aligned, P vertically, AVG = average) , here is the curve of our component :



The quality of this grating is that around the 785 nm the efficiency is constant so the output power will not change a lot when we will tune the laser.

The last feature to know is the angular dispersion D which depict the shift of wavelength when the angle changes.

$$D = \frac{2 \times \tan\alpha}{\lambda} = \frac{d\alpha}{d\lambda} = 4126815$$

So if we change the angle of 1 degree we change the wavelength of 242 nm (1,24 10^{15} Hz). This value show that we need to change accurately the angle if we want to change the wavelength.

4) The design of our ECDL system

a) Bases of mechanical

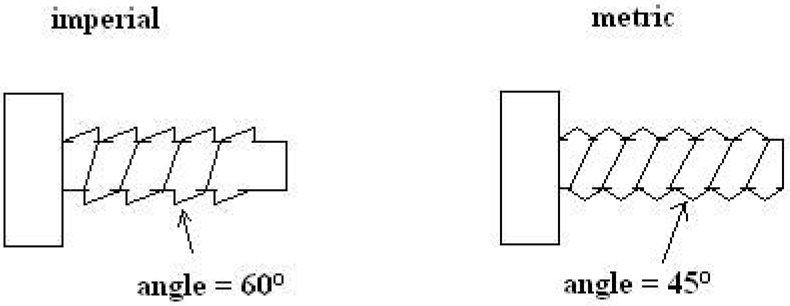
For the measures and some tools it exists two different systems : the imperial system mainly use in England and USA and the metric system.

Here are the length conversions between the two systems :

Imperial		Metric	
Mile	1760 yards	1.609 344	kilometres
Yard	3 feet	91.44	centimetres
Foot	12 inches	30.48	centimetres
Inch		2.54	centimetres

It is very important to precise which kind of dimensioning we use, in our case this is the metric system.

For the screw the difference resides in the shape of the threaded part :



A screw has a name which begin by MX where X is a number which depict the diameter of the screw. After that in our schematic we put T that means threaded or NT for no threaded.

For our application we need thermal conductor and insulator materials. It is important to well choose the materials because it can have an influence on the laser features because of the vibrations or a non constant temperature.

In the thermal conductors we mainly find metals and for the insulators plastics.

Here is a non exhaustive list :

<u>Insulators :</u>

- natural materials : wood, wool, straw, ...
- industrial materials : fiberglass, mineral wools...
- polymer : plastic, rubber...

<u>Conductors :</u>

- Metals : aluminium, steel, copper,…
- Fluids : water, oil, …

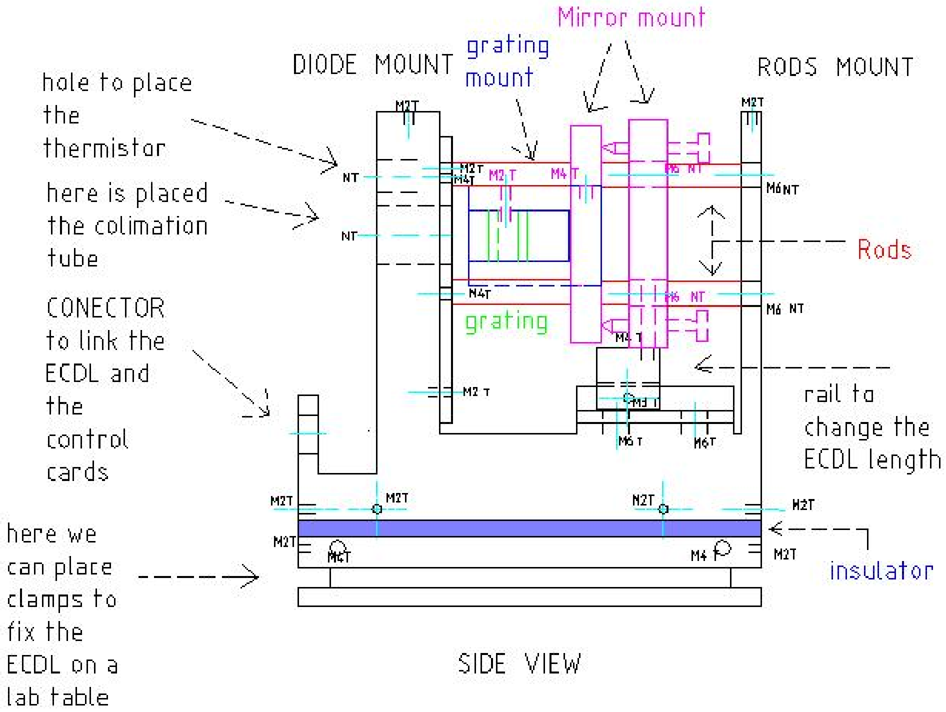
Another important element to take in consideration is the vibration of the system. If we want to avoid the bad effects of this we need a specific material, for example sorbothane. Sorbothane was first sell in 1970, it’s a polymer whose shape is like a sheet of 3mm thick. We can cut it with simple scissors, the texture look likes rubber. It can absorb about 95% of the vibrations.

b) General description

The design had been developed with the software Autocad. In the annexe is a quick description and some bases to use this software.

The ECDL must be composed of a mount which contain the collimation tube, a mount which contain the grating, a mount to connect the external electronic cards and a mount to well fix all the parts of the system on a lab table.

On the next page is the schematic of the ECDL side view which depict the main parts of the design :



The system is mainly made of aluminium. This material is lighter than steel and its solidity and mechanical properties make it useful for a lot of applications. Moreover to mill this material is very easy (for example it can be easily folded). The inconvenient is its high price.

The heights at the level of the grating mount and the diode mount have been calculated to place the grating in front of the diode. The grating is protected from dust by his mount on the top and the bottom.

Four stainless steel rods keep the system stable, they are screwed at the back of the diode mount and also in the rods mount. Four M6 holes in the mirror mount allow it to slide along the rods.

To improve the stability of the system and to well fix the distance between the diode and the grating the mirror mount it screwed on a rail. Thanks to the rail the strain on the rods are less important and we can select the length of the external cavity L_{ext} and fix it with a M3 screw.

At few mm before the diode mount we make another mount which contain two connectors commonly used in the serial bus RS232, the DB9 connectors. Into these 9 pins connectors we solder the different wires coming from the laser diode, the thermistor, the PZT and the TEC. On the other side of the connectors will be plug the wires towards the different control cards (temperature and current controllers).

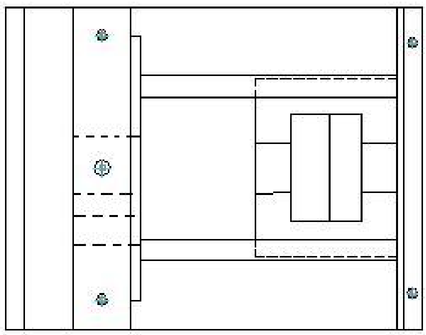
Another plate fixed under the base plate have a hole of 5 mm height and 5 mm depth to support the clamps at each side which maintain the system stable on a lab table.

The two plates are separated by a thermo isolator which contain a TEC of the Marlow company. The TEC is put into a hole whose dimensions are 40x44x4 mm to control the temperature of the base plate. It is included inside the insulator thus the TEC can transfer the heat without a backlash of the heat towards the diode and grating mounts.

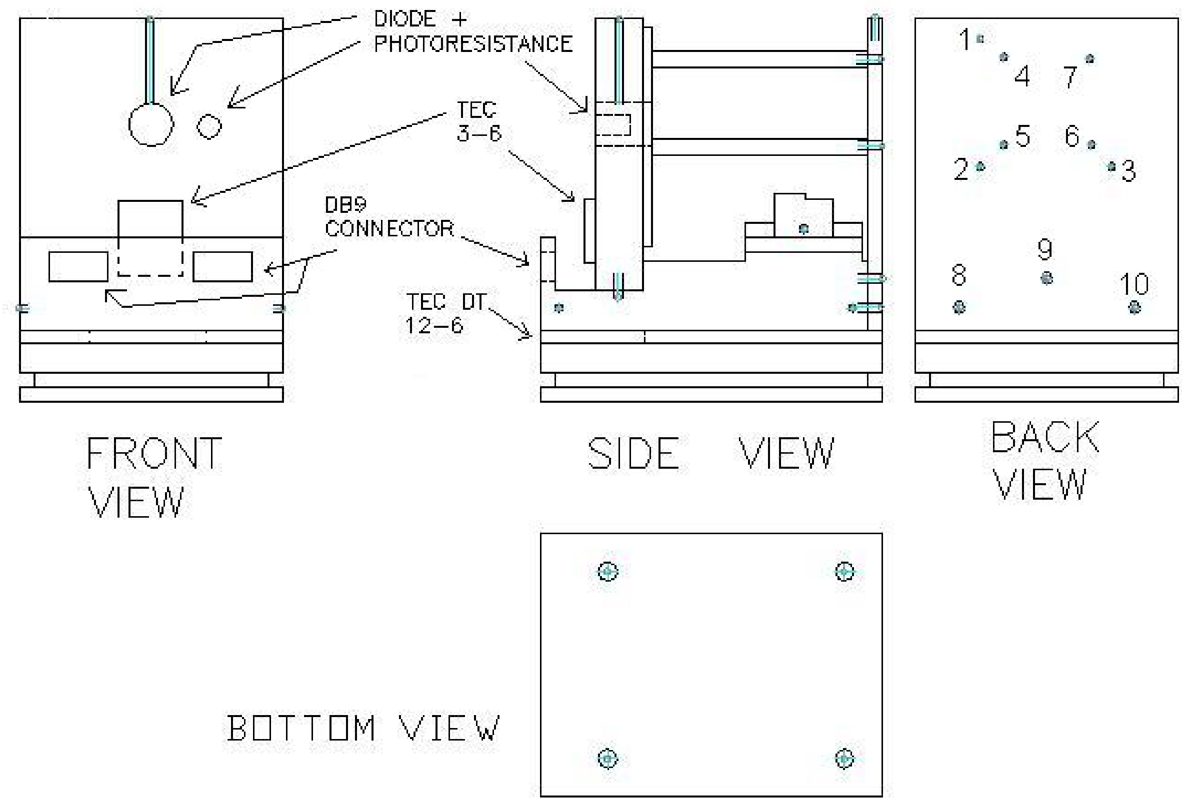
Some features have been changed to improve the design :

- The first idea was to do all the base plate in one part, but to make it more easy to mill the more simple is to separate the rods mount and the base and link them with some screws.
- To link the different base plates (for the clamps, the insulator and the main plate) we decided to put four screw at the bottom of the device (plastic screws to not conduct heat). These screws are represented on the next figure in the bottom view.
- A screw has been added on the top of the diode mount to fix the collimation tube, and the hole position for the thermoresistance has been changed (not above but at the right of the collimation tube at 2 mm).
- To fix a cover we only will use two screws on each side and five at the top.

Here is the design of the final ECDL for the different views :



TOP VIEW



All the dimensions can be found in the annexe.

The rods mount contains 3 holes aligned with the screws of the mirror mount (1,2,3). The aim is to pass a screwdriver, to circumvent the screws which has for effect to change the angle of the mirror mount and thus to change the grating angle.

The 4,5,6 and 7 holes permit to screw the four rods on the mount.

The 8,9 and 10 screws fix the rod mount and the baseplate.

c) The diode mount

This mount must contain the collimation tube, the thermistor and the TEC.

One of the most important thing is to well control the temperature of the diode, the TEC must be placed near the collimation tube and quickly change the temperature of the diode. The TEC will be soldered on the front surface using low melting point In-Sn eutectic solder. It needs a 30 mm square place.

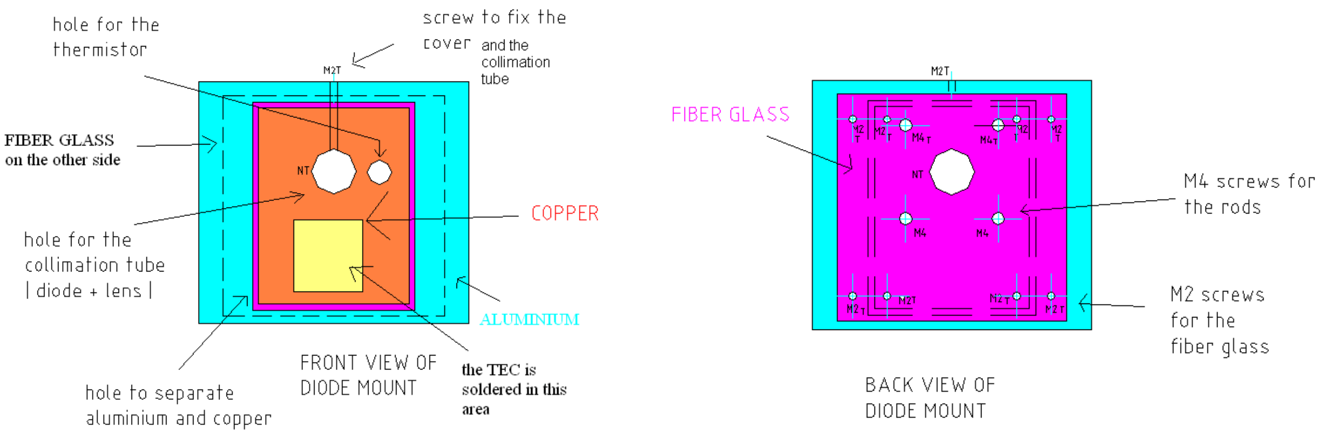
A way to improve the control of the temperature is to place the tube and the thermistor into a little copper plate. The advantage of copper is that this material conduct the heat with a high efficiency so if the TEC works the temperature of the plate quickly decrease and cool the diode.

We isolate this system using a fiber glass plate which separate the diode, the thermistance and the TEC of the rest of the assembly. So the grating will not be affected by the change of temperature near the diode. The fibber glass acts like an isolator for the heat and is screwed at the alimunium plate and at the copper plate by eight plastic screws at the back side to ensure a good isolation. The advantages of fiber glass are the low weight and a high resistivity to shocks.

The diameter of the colimation tube is 14.73 mm and its lenght is 19.05 mm so we need to mill a cylindrical hole inside the aluminium plate and the fiber glass with a diameter of 14.80 mm to well fix the tube. And to well include the tube into the thickness of the mount we need to choose a total thickness (Al + fibber glass) equals to 19.05.

The thermistor will be hold into a special mount which thermicaly isolate each of its wire, we have to mill a cylindrical hole of diameter 8 mm to place it inside the diode mount.

The holes at the top and the bottom are used to fix the diode mount to the cover and the base plate.

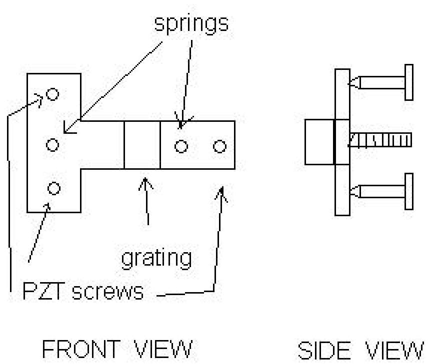


When we received the milled ECDL we noticed that the manufacturer forgot to make the copper plate and make all in aluminium. It will decrease a little the performance of the temperature control but not significantly. Moreover the hole for the thermistance was not made.

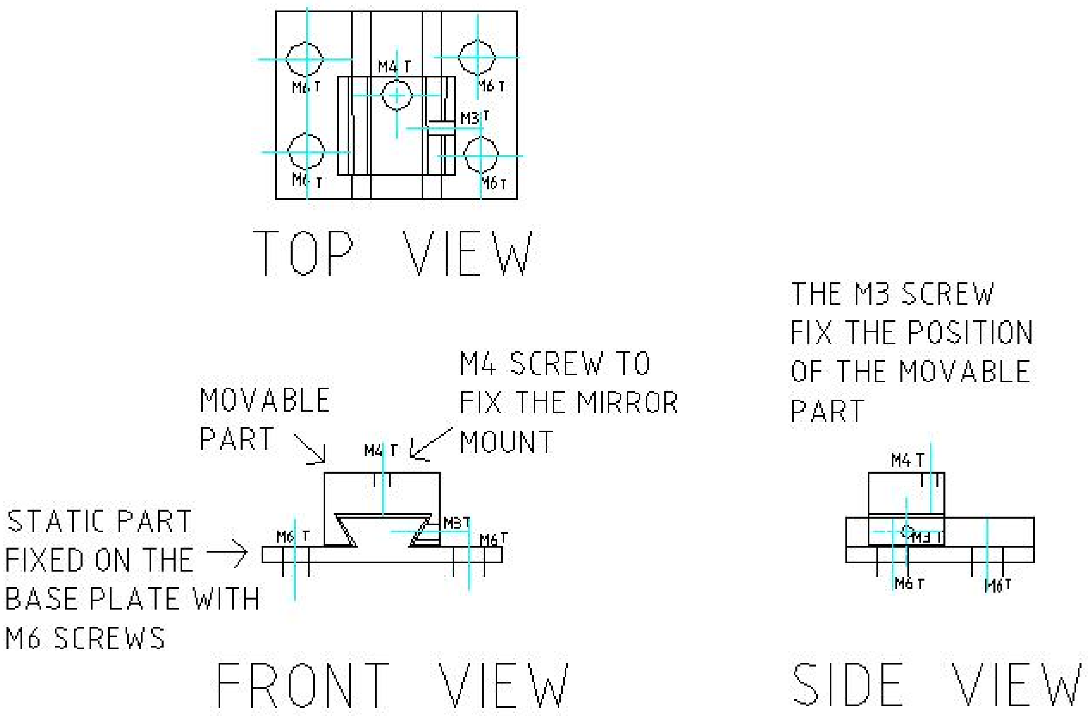
d) The rail and the grating mount

To have a well tunable cavity we must be able to control the distance between the diode and the grating and also be able to change accurately the angle of the grating.

The first idea was to fix the grating on a T shape aluminium plate and to use PZT screw and springs to control the position of the plate. But this idea was rejected because of the vibrations due to the springs.



Another idea to change the distance is to mill a rail, below is his schematic :



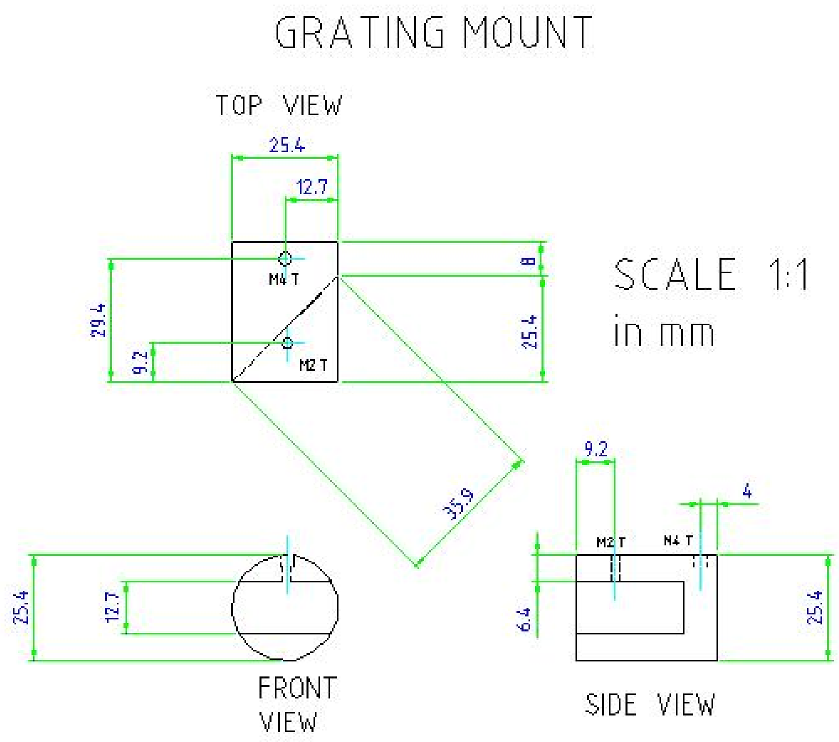
Four M6 screws fix the rail to the base plate, the M3 screw is tightened at the distance we choose to make the mount stable. The M4 screw is used to fix a mirror mount which contain the grating.

This mirror mount is useful to change the angle of the grating. Here is a picture of this device :

The screws of this mount (on the back in the picture) allow a coarse adjustment of the grating angle. A PZT element can be added between the two plates to make the fine adjustment. This actuator converts electric energy into mechanical energy. This is a very high precision device that can be useful to settle the grating angle. When you apply a voltage the length of the PZT changes so the angle of the mirror mount plate accurately changes. In our case we use the AE0203D04 PZT whose dimensions are 4.5 by 3.5 by 5 cm.

Another piece is milled to maintain the grating vertically in front of the collimation tube. This piece is cylindrical to be well fixed into the mirror mount and the front part is triangular to fix the angle of the grating at 45 degrees.

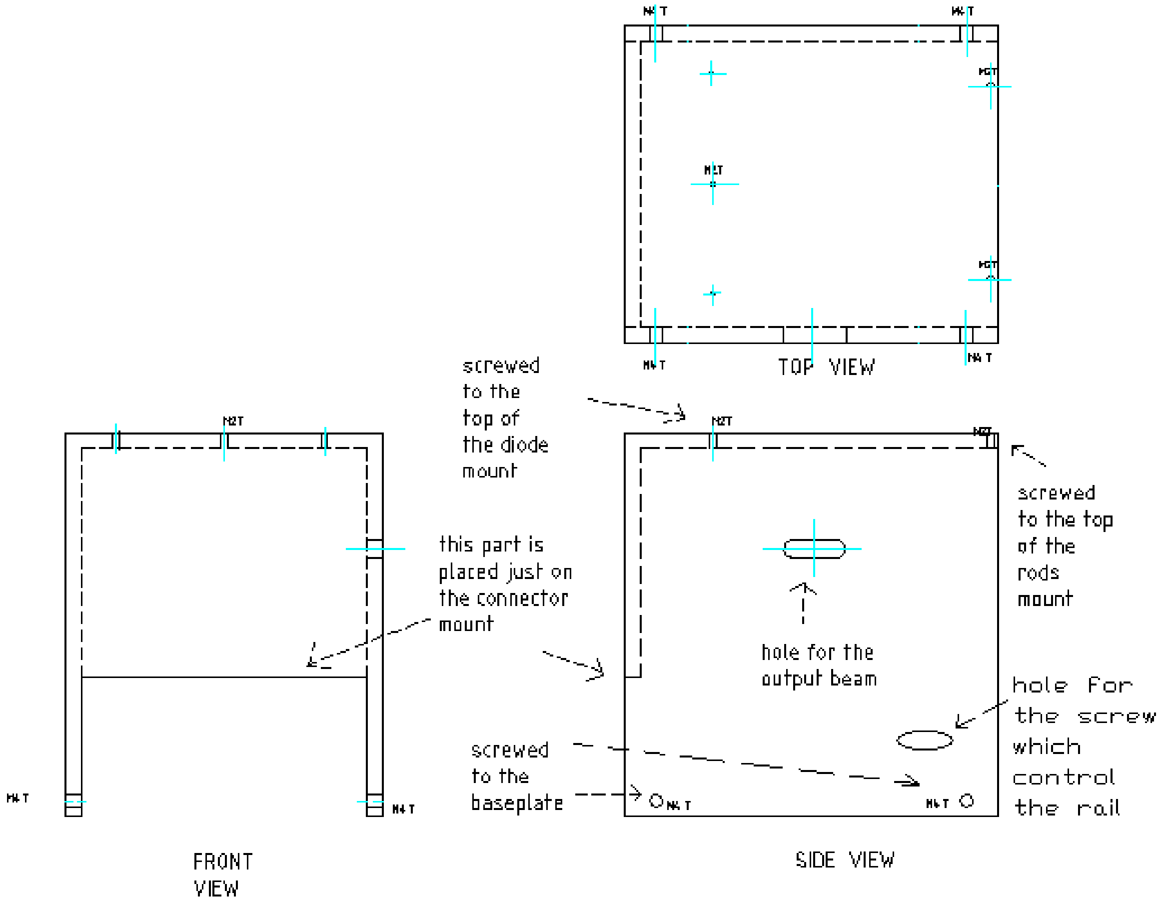
A M4 screw makes the link between the mirror mount and this grating mount and a M2 screw will fix the grating on the triangular surface. Here are all the dimensions of the grating mount :



e) The cover

We still need a cover to isolate the system of the environment. This part has to well coat the entire system without affecting the laser beam, the driving current and the temperature.

Here is the schematic of the side view, the front view and the top view :



The cover will be fix by 9 screws , 2 at each side and 6 at the top. It is made in one piece with an hole on one side to permit the beam to go out after had been reflected by the grating. This hole must have a long enough size because the output beam position depends on the position of the mirror mount on the rail. The cover has 4 facet, 1 at the top, 2 for each side (one with the hole for the output beam and the other without any hole), 1 which stop just above the connector mount. We doesn't need to cover the rods mount because the mount seal enough the side. We add another hole to fix the position of the rail with a screw.

III : Design of the current source for the laser diode system

1) Specification and constraints

The current source is aimed to supply the laser diode we use in our external cavity. As we said previously, the working wavelength of laser diode is very sensitive over the variation of the supply current (about 10 nm/mA). That implies that the circuit has to provide a very stable current if we want that our laser system be selective. Besides, since the laser system needs to work at a stable wavelengh, the current source also need to have a poor drift in long terme (about 10 hours).

The laser diode we use is the Hitachi HL7851G and the characteristics we need are the followings:

-the operation current is 140 mA
-the maximal current that can flow in the laser diode is 170 mA
-the sensitivity is about 0.33 nm/mA

In a spectroscopic application, a typical specification for the current source is to have short-term fluctuations less than about 8 µA and a drift less than 2 µA in a long-term application (10 hours). For the used laser diode, a drift of 2 µA implies a wavelengh's shift of 6.6 nm.

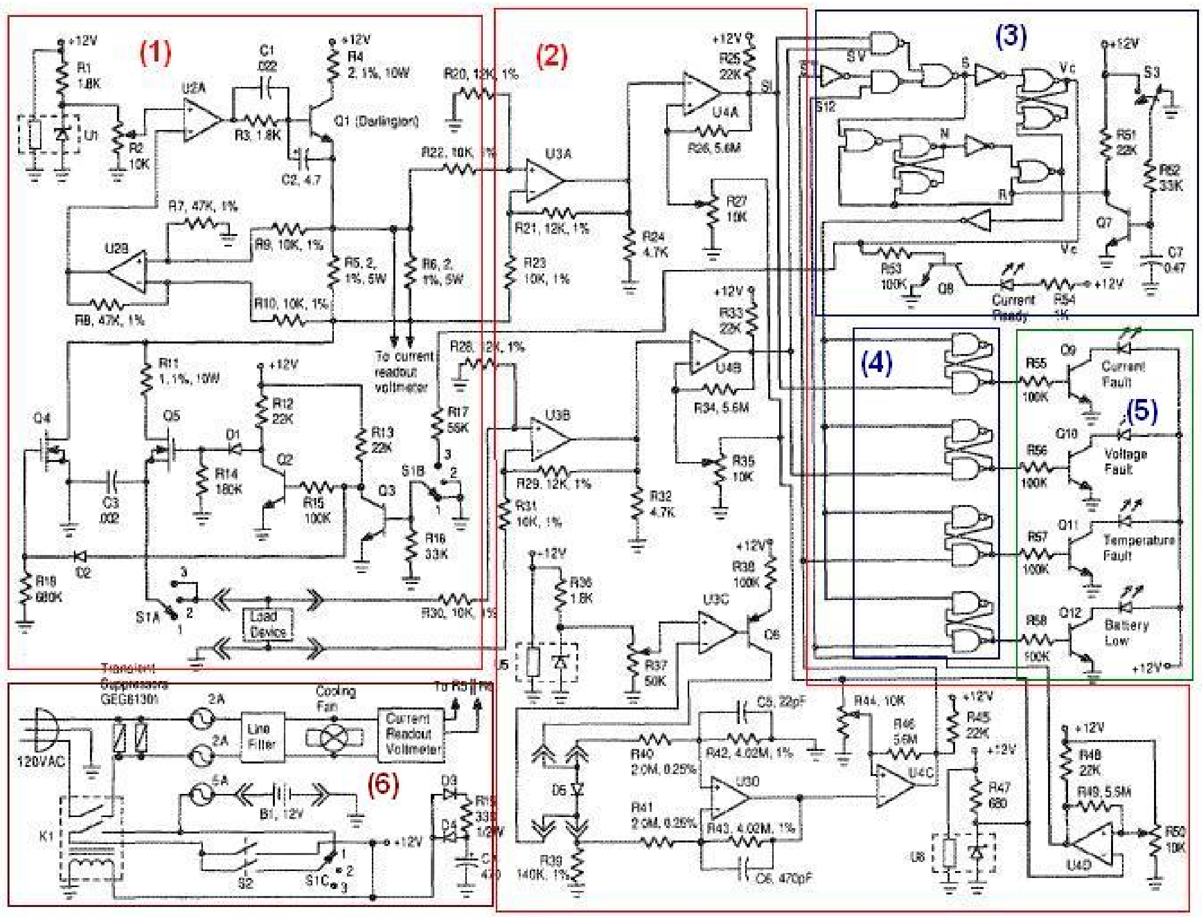
We know the main specifications that should have our current driver. In the scientific reviews, some circuit architectures are described for our kind of application. We are going to explain quickly some of them and see which of them fits better for our application.

2) Description of the current source schematics

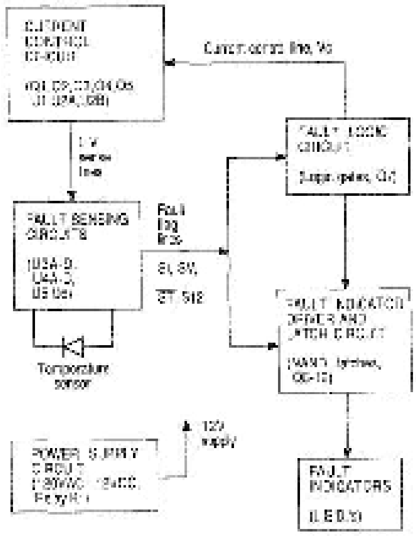
We can find a lot of applications of the diode laser systems and a large quantity whose topic concerns the elaboration of an external cavity for spectroscopic analysis. In most of these articles, we can find the description of current source to drive the laser diode.

a) First circuit

This circuit was described in 1989 by Michael S. Cafferty in the *review of scientific instrumentation (**ref 17**)*. The schematics is shown in the following figure:



The circuit is divided into six specific blocs whose links are summerized in the following bloc schematics:



The part (1) on the complet schematics is the current control circuit, where the current level is settled and the input voltage is converted into the output current.

The part (2) is the fault sensing detector aimed to detect current variations, voltage variations, batterie level to be treated by a logic control.

The faults detected by the part two is treated by the fault logic circuit (part 3), and are indicated by LEDs (part (4): fault indicator and part 5: LEDS)

The part 6 is the supply voltage circuit.

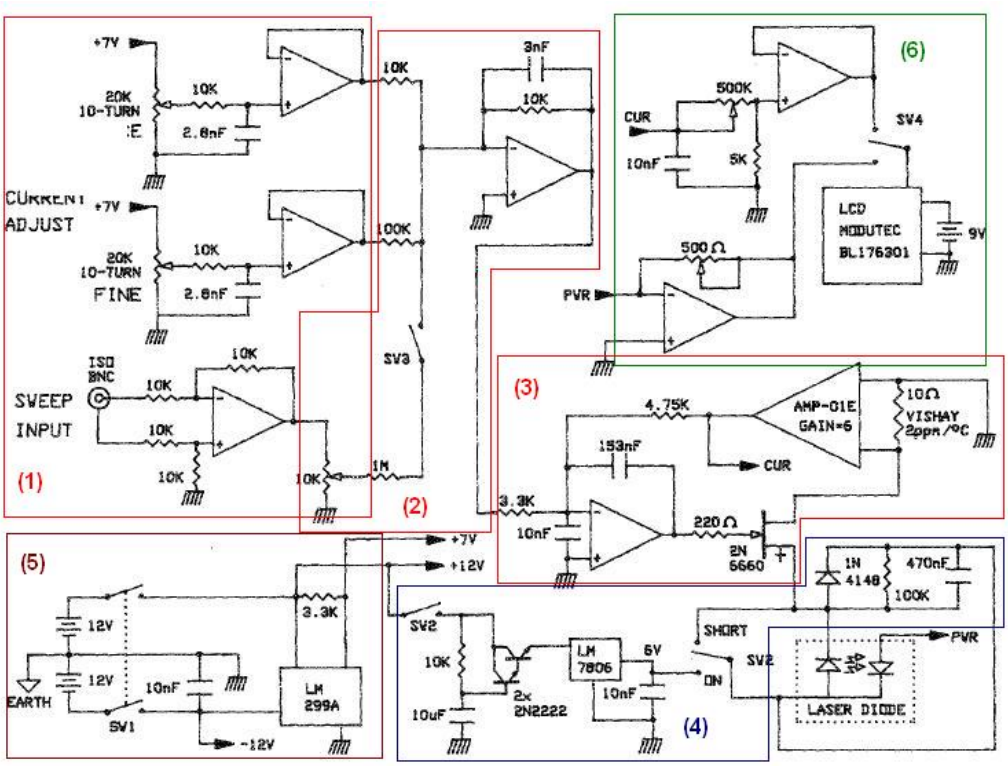
We do not describe this circuit further because, on one hand the specifications made for this circuit are not very near from ours because it had to produce 1A current and moreover it is difficult to compare the performance of this circuit with the others we found. Indeed, in the article, the current stability is studied in 24 hours wherease in the others articles it studied in 10 hours. In this article, the mesured drift was 37 µA in 24 hours. On the other hand, we found simplier circuit than this one which realizes the wished operation with the excelent performance.

We described more the following architecture because the structure is near of that we used.

b) Second circuit

The architecture of this circuit, described by C.C. Bradley, J. Chen and Randall G. Hulet in the *review of scientific instruments* ***(ref 16)*** is quit easier than the previous one for the same work and for the same performances. It is moreover almost a similar architecture that we used to build our current driver. In fact the begining of the circuit is the same but the what change is the end of the architecture with the final current controle loop. We are going to make a quik description of this schematics before to see why we have chosen it.

The schematics is shown in the following figure:



The structure can be devided into six blocs, each of them having his own specific work, as shown in the previous figure.

The role of the bloc (1) is to settle the output current following three ways:

-two ways for the current adjustment (coarse and fine)

-one way (sweep input) linked to a BNC connector to be driven by computer or digital counter for spectroscopic operation.

The coarse and fine current adjustments are, each of them, realized by one potentiometer which proportionally settle the input voltage following a reference of 7V. (As we used this kind of current adjustment in our circuit, we will describe this adjustment way more closely later). A basic first-order low-pass filter follows the potentiometer (for the fine and coarse adjustement) in order to eliminate eventual high frequency parasites that can appear.

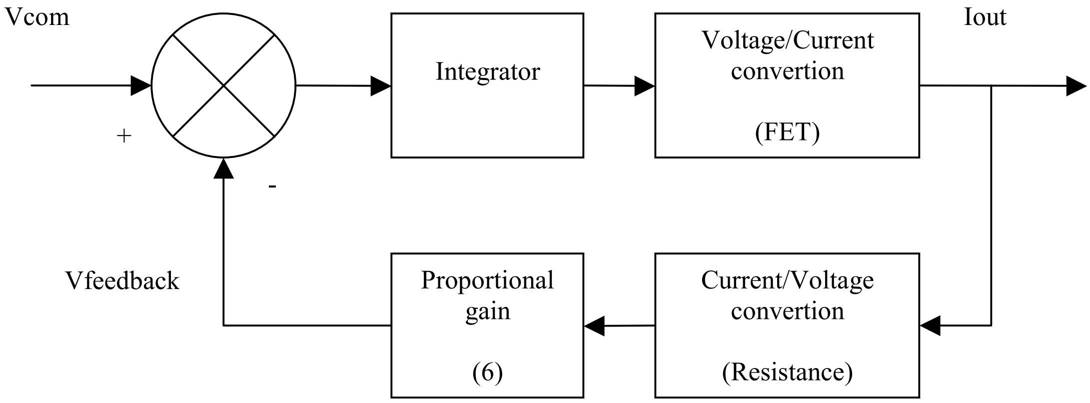
The sweep input is composed by one substracter that does the difference between the two BNC-connector's electrodes.

The bloc (2) is composed by one inverter-adder/low-pass filter that adds and ponderates the three settled voltages comming from the coarse and the fine adjustment and the sweep input. The difference between the coarse and fine adjustment is made by ponderation coefficients of the adder. In the schematics the static gain for the coarse input is 1 and that for the fine gain is 0.1, so for the same angular movement of the coarse and fine potentiometers, the coarse voltage component variates ten times faster than the fine voltage.

Concerning the sweep input, the gain is adjustable by a potentiometer in a voltage-divider configuration. It allows to select the range of the sweep operation.

The bloc (3) is aimed to produce the current providing the laser diode according to the addition of the coarse, fine and sweep settled voltage. The convertion voltage/current is made by a FET transistor, driven by the difference between the settled voltage and a feedback voltage, proportional to the output current. The difference is integrated before the FET transistor to make the system faster. The feedback voltage, proportional to the output current, is optained by a precision 10Ω resistance connected to the two inputs of an instrumentation amplifier settled to have a gain of 6.

So, the structure of the bloc (3) can be sumerized by the following bloc-schematics:



The bloc (4) is a protection bloc agains current and voltage spikes and transcient that can occure when the supply power is turning on or when voltage from the current setting change quickly. Indeed, a laser diode is very sensitive and can easily be destroyed by current or voltage out of its specification range, even on a short period. Moreover, the use of protection features is as justified as the bloc (3) incorporate an integrator inside the loop, allowing to improve the rise-time of the response to correct current drift and to cancel steady-state error but degrading the transient response.

The voltage spikes and transients could arise from the supply voltage and the settled voltage bloc.

To protect the laser diode against power supply spikes, one switch allows to short the laser diode, setting the voltage and the current inside the laser diode at zero. In this way, the laser diode is protected when the supply voltage is turned on or turned off, i. e. in the moment when spikes and transient can appear. Moreover, between the supply block and the laser diode, a transistor in a darlington configuration, a 10 µF capacitor allows to slow voltage variations, decreasing the slope when the supply voltage source is turned on.

To protect the laser diode during the normal work, a reverse biased diode is put in parallele with a capacitor and a resistor in order to avoid fast start-up currents.

The power supply (bloc 5) is composed by two 12V-batteries and a precision voltage reference (from the same familly we use) allowing to have a references of 7V and 12V. The use of batteries instead of a classic supply allow to minimize the noise in the power supply lines.

The last bloc (6), allows to display the value of the current flowing in the laser diode or the optic power. A swich allow the selection between the both. To display the value of the current, the voltage, proportional to the current, just after the instrumentation is used and a potentiometer in a voltage divider allow to select the good gain to have a tention whose value is that of the current. This voltage is finally buffurized to be displayed. The same principle is used to display the optic power.

Performances of the circuit:

To reach the specified performances, the components must be accurate. In the description, the used resistors had a temperature coefficient of 20 ppm/°C, the voltage reference (the most critical) have a temperature coefficient of 0.2 ppm/°C and the operational amplifiers and instrumentation amplifier were chosen for their low offset voltage, offset and bias current and low sensitivity over temperature.

With this configuration, current variation did not exceed 1.5 μA in one hour.

Consequently, this architecture, simple to realize, fit to our applications. However, we found another structure, almost similar, which reach almost the same performances without using integrator, reducing the risk having spikes and transient.

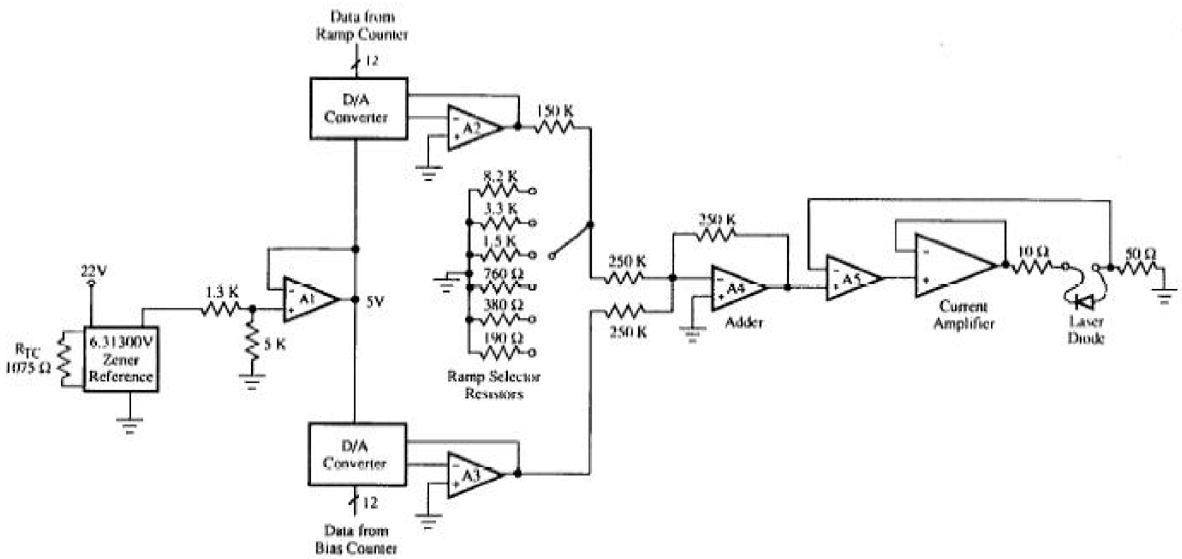
c) Description of the selected circuit

This architecture was found in the review of scientific instrument, *Vol 63, No 4, of April 1992* ***(ref 15)*** and seemed to be convenient for our application :

Indeed, this current source is aimed to drive an external cavity diode laser system used for spectroscopic applications supplying an operation current of 100 mA with a stability lower than ± 5 µA and a drift over temperature of 1.2 uA in over 10 hours.

The contrainsts described in the review are about the same than ours, except the value of the operation current, but it is easy to modify the schematic to reach the 140 mA (± 20 mA) needed in our case.

The next figure shows the basic schematics found in the *review of scientific instrumentation*:



The architecture is very easy to understand:

The part (1) of the circuit is composed by a stable voltage reference followed by a voltage divider to supply the two digital to analog converter (Part 2 and 3) in a voltage of 5V.

$$5V = \frac{R_2}{R_1 + R_2} \times V_{ref}$$

Then the circuit is divided in two almost symetrical branchs that contains the two converters (part 2 and 3).

The part two, composed by one digital counter and one D/A converter is aimed to supply the operation current (140 mA in our case). The counter allows to set the value of the bias current (from 0 to 140 mA) to which we want to work, and the D/A converter turns the digital value of the counter into the corresponding analog value following the 5V reference. That is to say that the analog output value is put into the scale of the voltage reference (5V) and is proportionnal to the value of the counter. A maximal bias current (140mA) correspond to a value of 5V at the end of the converter (part 2). A buffer is connected to the output of the converter in order to isolate the commands from the rest of the circuit.

$$V_{bias} = \alpha_{bias} \times 5$$

α_{bias} is the proportionnal coefficient given by the value of the counter (value between 0 and 1).

The part 3 is quite similar to the second one except the presence of one selective voltage divider after the buffer. This branch is used to adjuste more closely the operation current or to realise a spectroscopic scan around working frequency determined by the value of the bias current. Indeed, since the wavelength of the laser diode is proportionnal to the variation of the supply current, we can change continiously the current from the operation current in order to carry out a spectroscopic operation. The selective voltage divider is used to select the appropriate ramp current range (for scaning operation or fine adjustement) thanks to a swich: Different resistances determining the range are connected together in parallele to the ground and the switch allow to select one of them at their other side.

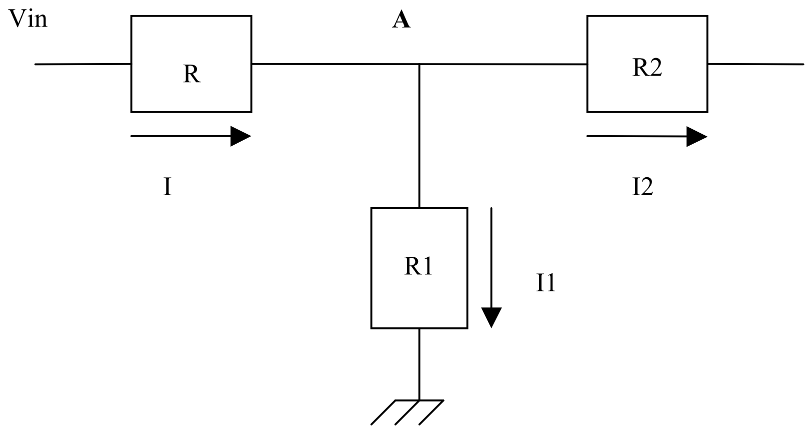
$$V_{ramp} = \frac{R_i}{R + R_i} \times \alpha_{ramp} \times 5$$

where R_i are the resistances in parallele that can be selected by the switch and R is the fixed resistance of the voltage divider. α_{ramp} is the coefficient taken its values between 0 and 1 according to the value of the counter.

The selected resistance R_i is big, bigger is the range of the ramp current. We will describe how we have chosen the value of these resistances in the final circuit later.

The part number 4 is composed by an reverser/adder that adds the bias voltage with the ramp voltage. The gain of the adder is determined by the values of the resistances. For our application, it is better that the gain of the adder be -1 in order to avoid to amplifie the voltage drift that occurs upstream. That means that the value of each resistances that compose the adder be the same and have to be at least 10 times bigger than the biggest value of the resistance of the ramp current selection voltage.

Indeed, if we considere the following schematics, we express the voltage in A as follow:



$V_A = R_1 I_1 = R_1 (I - I_2)$ and $I = \dfrac{V_{in} - V_A}{R}$ implies that $V_A = \dfrac{R_1}{R + R_1} V_{in} + \dfrac{R_1 R}{R + R_1} I_2$

If $R_2 >> R_1$ then $I_1 >> I_2$ and consequently $V_A \approx \dfrac{R_1}{R + R_1} V_{in}$. That implies that if the impedance that follows the voltage divider, the voltage A is determined by the voltage divider and doesn't undergo the influence of the current that flows downstream.

In the basis schematics, the biggest resistance in the voltage divider is 8.2K that mean that the value of the resistances of the adder should have a value bigger than 82K (In the case of the basis schematics, their value is 250K).

The expression of the voltage V_4 just after the reverser/adder can be expressed as follow:

$$V_4 = V_{bias} + V_{ramp} = -V_{ref} \left(\alpha_{bias} + \frac{R_i}{R + R_i} \alpha_{ramp} \right) \qquad V_{ref} = 5V$$

The last part of the circuit is composed by one buffer, one current amplifier and the laser diode. The buffer enclose the current amplifier and the laser diode in a loop in order to fixe the potentiel at a fifty-ohm resistor. This potential determines the value of the current that foster the laser diode because of the Ohm law: $I = \dfrac{U}{R} = \dfrac{V_4}{50}$

The current amplifier is chosen to be able to supply a current bigger than the operation current of the diode (100 mA in the case of the review). If the laser diode would be directely connected to the output of the buffer, the operation amplifier working as a buffer would produce heat if its output features don't allow it to support such a current.

The ten ohm resistance just before the laser diode is not very important in the design because its value is not present in the expression of the current flowing through the laser diode but it is aimed to protect the laser diode and to avoid a big potential difference at its pins.

Finally, the current supplying the laser diode can be expresssed as follow:

$$I_{diode} = -\frac{V_{ref}}{50}\left(\alpha_{bias} + \frac{R_i}{R + R_i}\alpha_{ramp}\right)$$

We just have broadly seen how works the circuit and expressed the analytic expressions of the main points of the circuit. Now we will discuss about the critical points of the design.

d) Critical points of the selected circuit

If we look at the equation of the current, the value, and of course the stability depends on the voltage reference, on the noise introduced by the D/A converters, on the resistances of the voltage divider and finally on the fifty-ohm resistance.

The most significant point for the stability of the output current is the quality of the voltage reference because the current is directly proportional to Vref. That implies that we have to choose an as stable voltage reference as possible if we want to have a good accurate current. Just to have an idea, we need to have short-term fluctuations less than $8\,\mu A$. If we work at the operation current of $100mA$, i.e. $\alpha_{bias} = 1$ and $\alpha_{ramp} = 0$, that suppose: $\Delta V_{ref} = 50\Delta I = 0.4mV$ only if we don't considere the other sources of fluctuations. So we have to find a very stable reference to have the best stability for the output current.

The other main critical point of the circuit concerning the stability is the fifty-ohm resistance at the end, because the current that flows through the laser diode is determined by the voltage at this resistance (Ohm law). If the tolerance and the drift over temperature of the fifty-ohm resistance are not enough low, first, we are not sure to reach the wished current (in the case of a bad tolerance), and secondly, the stability of the output current over temperature is not assured (If the drift of the resistance is high). Consequently, we need to have an as accurate resistance as possible at the end of the circuit (low drift and good tolerance).

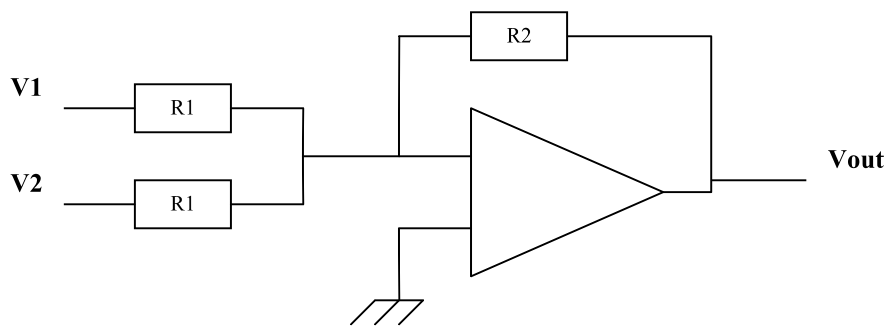
More over, we have to be carreful to the power dissipation of this resistance, because if the component dissipate a power superior to its nominal power dissipation, it would generate heat that will generate a shift of the resistance's value and concequently the output current could be drifted. In the case in which we work with a operation current of 100 mA, the output resistance have to dissipate a power equal to $R \times I^2 = 50 \times 0.1^2 = 0.5\,W$. In this example, it is easy to find an accurate resistor that can dissipate 500 mW. But if we consider our case, we have to reach a maximal current of 160 mA, in other word, the output resistance needs to dissipate a power of 1.28 W. It is problematic because it is impossible to find an accurate resistance that is able to dissipate such a power without generating heat. We will discuss later about the different way to bridge this problem.

The resistances that compose the voltage divider of the ramp current setting appear in the formula of the output current. Consequetly, these resistances are also critical (but a little bit less than the fifty-ohm resistance because, in the current equation, these values intervene in a ratio, so the total drift generated by the voltage divider is lower than the drift of each resistance). To have a current as stable as possible it is better to use as accurate resistances as possible.

Finally, we have to considerate the case of the operationnal amplifiers. Since each of them generate a drift at their input, we have to know if there are critical for the stability of the current or not. All of them, except the current amplifier belong to a direct chain. As a consequence, a drift appearing in each of them affect the the value of the potential at the fifty-ohm resistance, in other words, the output current. More over, the consequence are biger if we considere the case that one of the operational amplifier has a gain superior to one. In this architecture, all the operational amplifier, except the adder, are in a follower configuration, so they have a gain aqual to one. For the adder (see the following figure), since the output voltage is given by the following expression:

$$V_{out} = -\frac{R_2}{R_1}(V_1 + V_2)$$

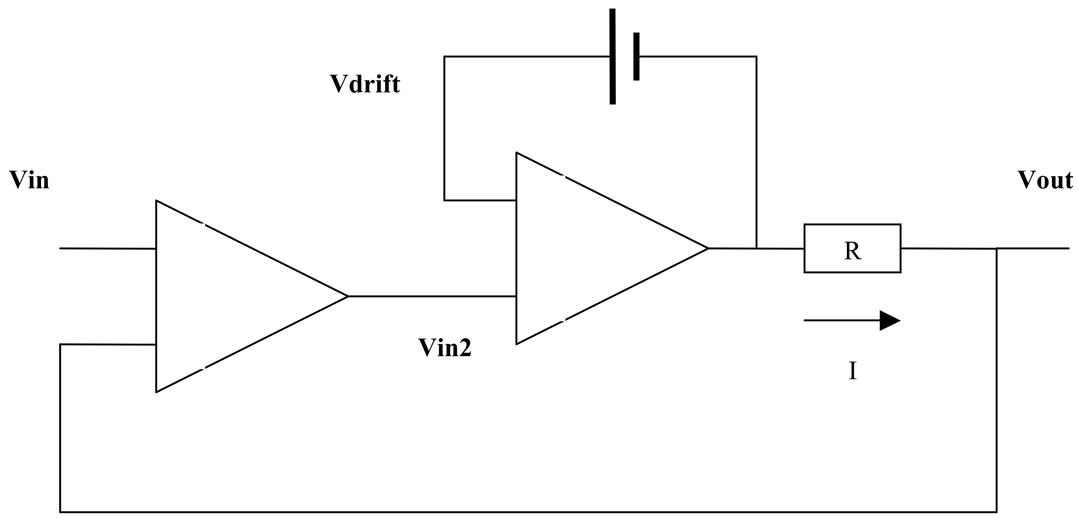
the resistance R_1 and R_2 must have the same value in order to have a gain of one.



According to what precedes, the operational amplifiers, except the current amplifier must ha a good precision, I mean a very low drift over temperature and a low noise.

For the current amplifier, the situation is different, because it is encosed in the loop of the previous follower that fixe the potential at the output resistance. Indeed, if we consider that the current amplifier has an open-loop gain of **A** and the previous operational amplifier used as a follower a gain of **B**, we can show that the defaults of the current amplifiers and the resistance protecting the laser diode are bridged by the gain of the follower.

The part of the schematics is shown in the following figure:



We have $V_{out} = A.(V_{in2} - V_{drift}) - RI$ and $V_{in2} = B.(V_{in} - V_{out})$ so:

$$V_{out} = A(B.V_{in} - B.V_{out} - V_{drift}) - RI$$

and finally:

$$V_{out} = \frac{A.B}{1 + A.B} V_{in} - \frac{A}{1 + A.B} V_{drift} - \frac{R}{1 + A.B} I$$

Since usually the operational amplifier have a very huge gain ($A, B >> 1$), the effects of the drift and the ten-ohm resistance are negligeable, as shows it the final expression:

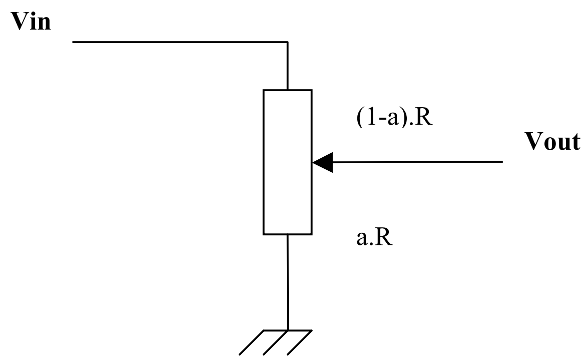
$$V_{out} \approx V_{in} - \frac{1}{B} V_{drift} - \frac{R}{A.B} I \approx V_{in}$$

So all the physical characteristics of the components enclosed in the loop are divided by the gain of the follower and since this gain is about 10^6, the defaults, for exemple the drift, of the current amplifier and the ten-ohm resistance protecting the laser diode are negligeable. Consequently, the current amplifier and the ten-ohm resistance are not critical.

e) Modifications of the original schematics

Fist change:

We decided first to replace the digital control of the current by an analog one. A potentiometer can play the role of the digital counter (or microcontroller) like it is shown in the following figure:



This structure behaves exactelly like a voltage divider and the output voltage can be expressed as follows:

$$V_{out} = \frac{a.R}{(1-a).R + a.R} V_{in} = a.V_{in}$$

where “a” stand for the position of the floating pin of the potentiometer (so the value of a is between 0 and 1). As we can see, the analog way give the same result than the digital one.

The two ways have advantages and disadvantages: First of all, the analog implementation is easier to implement because we just need to use one potentiometer, wherease in the case of the digital adjustment, one counter or one microcontroller connected to a computer and one D/A converter are necessary. The main advantage of the digital way is given by the fact that the bias and the ramp command can easely be interfaced by a computer without introducing instability. But the backward is that the D/A converter is sensitive to the variation of the temperature and generate a drift of about 5 ppm/°C. As opposite, the analog adjustment, in theory, does not add drift over temperature, or at least a negligeable one, because of the ratio. Indeed, if we suppose that the drift introduced by the temperature (ΔR) is the same everywhere in the potentiometer, we can write that:

$$V_{out} = \frac{a.(R + \Delta R)}{(1-a).(R + \Delta R) + a.(R + \Delta R)} V_{in} = a.V_{in}$$

So as we can see, in the ideal case, the drift doesn’t affect the value of the input voltage that determine the value of the current. But in the reality, the value of the drift is not really the same everywhere in the potentiometer. Consequently, we have to choose a potentiometer whose characteristics are linear and drift over temperature is as low as possible. If the drift doesn’t change a lot according to the position in the potentiometer, the effect of it is bridged by the ratio and the more the linearity of the potentiometer is good, the more the effect of the drift is perceptible. We will discuss about it more closely later.

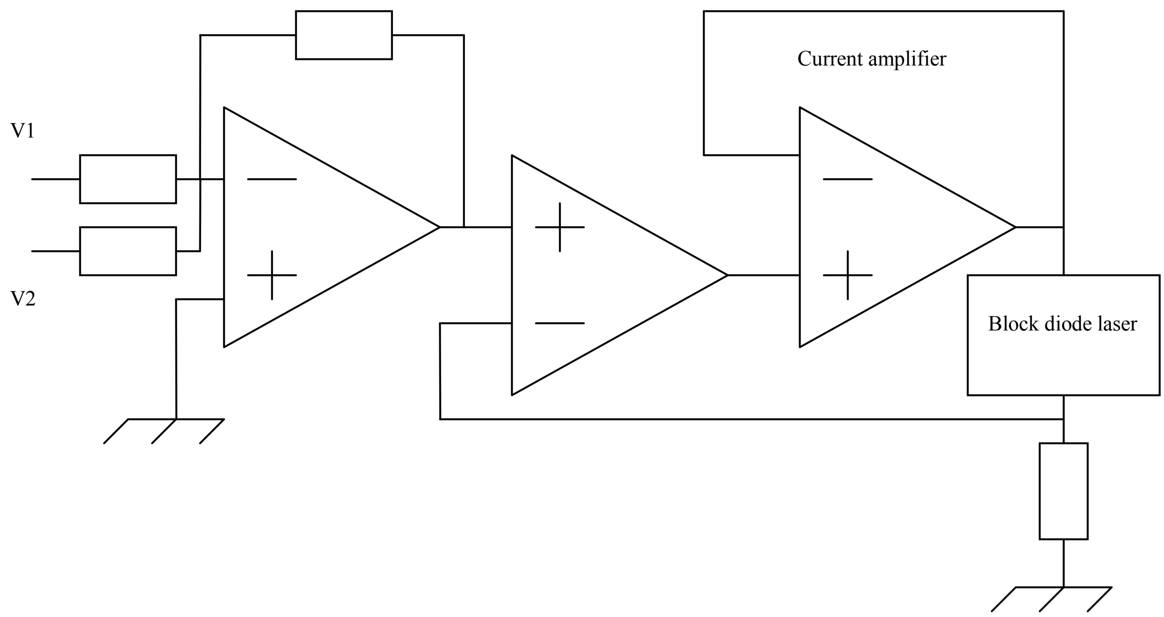
The reason why we choose this analog solution to set the current in the laser diode instead of the digital one is that with accurate potentiometers we expect to generate a drift lower than 5 ppm. But if the circuit behave good, it would be possible in future improvements to remove the potentiometers and connect to our current driver a digital setting, to allow accurate spectroscopic applications.

Second change:

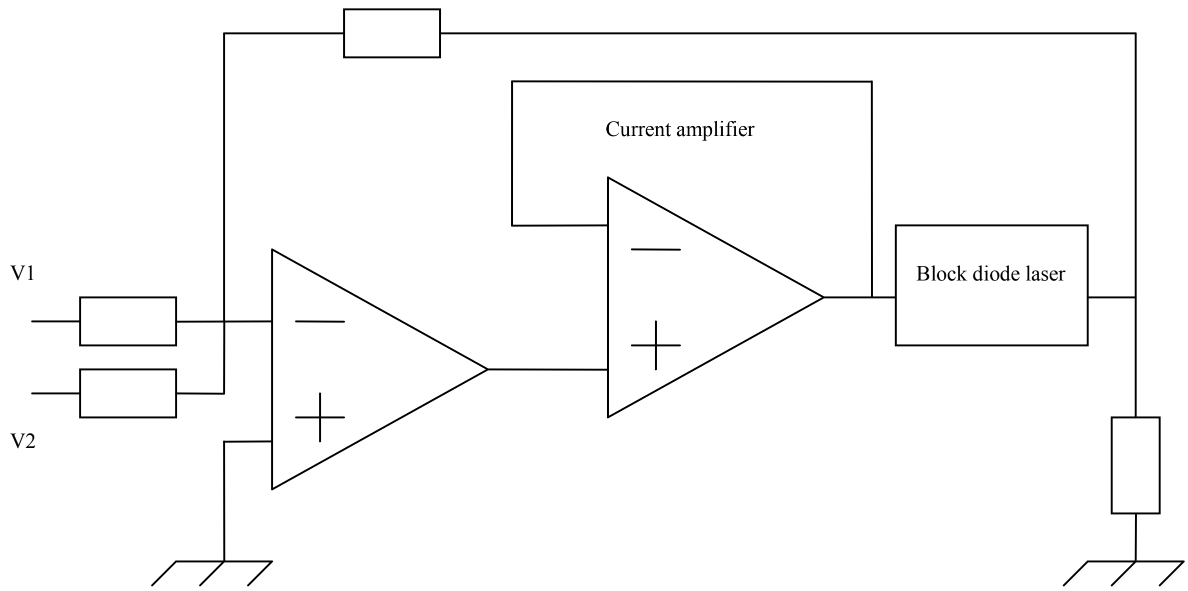
The second modification consisted in evaluating the usefulness of the operational amplifiers in order to know if it is possible to remove those which are not necessary. The first one is used to isolate the voltage reference from the others parts of the circuit. Likewith, the operational amplifiers buffering the two commands (bias setting and ramp setting) play the same role than the first. And the operational amplifier that works as an adder is of course useful. So all these previous operational amplifiers are important for our design. But we could wander if the operational amplifier working as a follower just before the current source is really necessary. It would be preferable to remove it if it is unuseful and to use the adder instead to enclose the current amplifier and the laser diode in a loop.

The two possible structures corresponding to the end of the circuit are shown in the following figures:

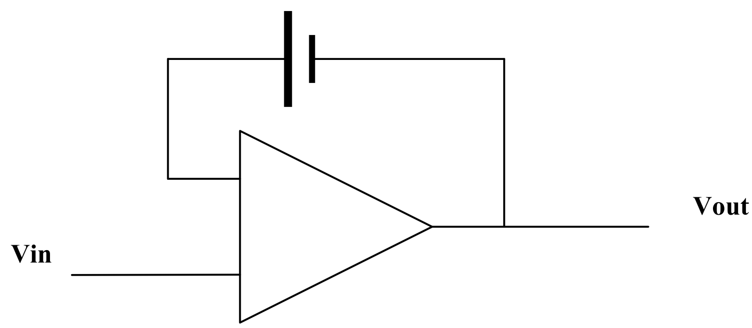
Figure 1: The original structure



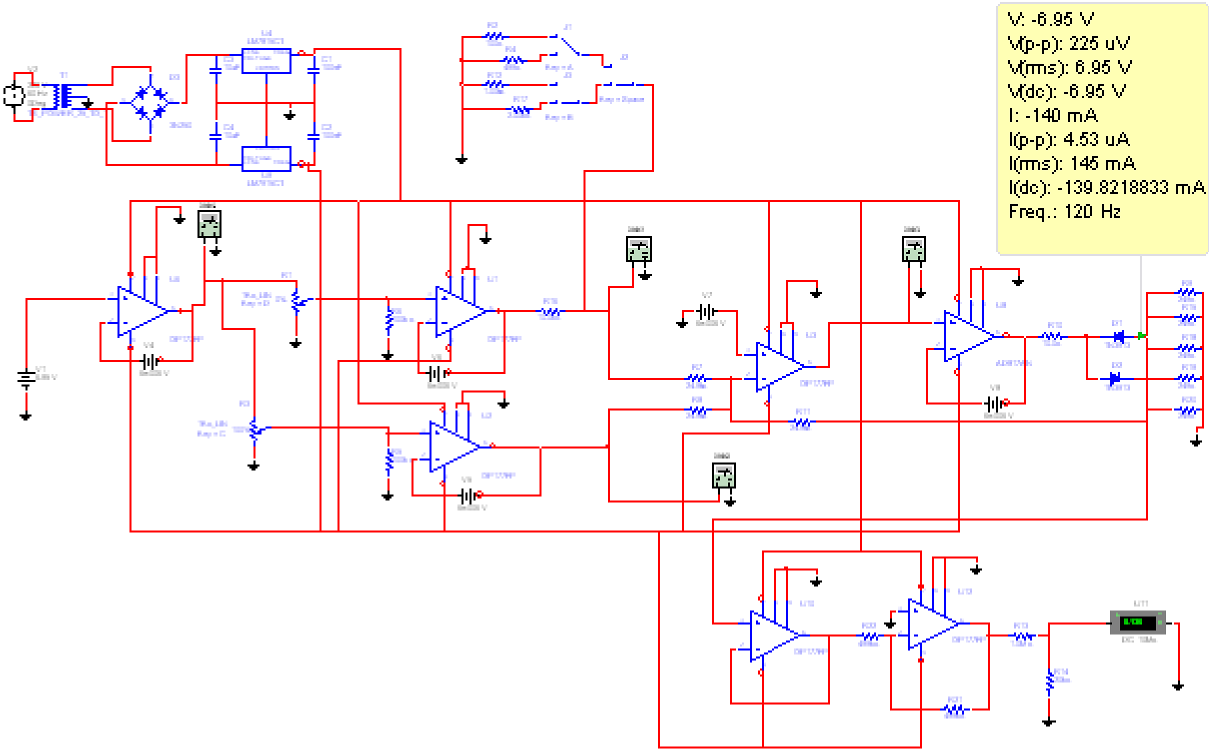
<u>Figure 2:</u> The modified structure



To check if it is possible to remove the last follower, we have to simulate the two versions with a spice model. For that, we use *Multisim*, to simulate the sensitivity of the two schematics to the drifts appearing in the inputs of the operational amplifiers. The drift over temperature is modelized by a DC voltage source at one of the two inputs of the operational amplifier. That stands for the shift between the inverter input and the non-inverter input. If the input voltage of the follower is in the non-inverter input, we have to set a positive DC voltage source at the minus input. In the opposit case, the DC voltage source has to be negative. (See the following figure that shows the modelization of a drift)



The following figure shows the total view of the schematics under simulation on *Multisim*:



Results of the simulations:

We realized simulations for different values of the drift over temperature that can occur at the inputs of the operational amplifiers by applying an DC voltage between the two inputs for the two schematics. The purpose is to compare the output current of the two architectures and see in which of them the current is more affected. The results are displayed in the following array, where the structure 1 is the original one and the structure 2 the modified architecture:

Value of the drift in each op amp	Current structure 1	Current structure 2	Comments
0 ppm	139.5574637 mA	139.8217893 mA	initial temperature
0.3 ppm	139.5574744 mA	139.8217939 mA	variation of 1°C of the temperature
6 ppm	139.5576782 mA	139.8218833 mA	variation of 20°C of the temperature
9 ppm	139.5577855 mA	139.8219303 mA	variation of 30°C of the temperature

As we can see, for the values of the drift simulated, corresponding to a temperature variation less than 30°C, that the changes of the output current are not significant for our application in the both structure. Indeed, a drift of 9 μV has for consequence to change the output current of only few hundren nV (0.32 μV for the structure 1 and 0.141 μV for the structure 2). In our circuit, we use operational amplifiers whose temperature coefficient is less than 0.3 $\mu V / °C$. Consequently, if the temperature of the environment of the operational amplifiers increase from twenty degrees, the drift at the inputs of each operational amplifier is about 6 μV, so it would not affect a lot the value of the output current, in the two structures. But the modified structure performs a little bit better according to the simulations and moreover, it allows us to remove one unusefull operational amplifier from the original schematics.

However, in these simulations, we didn't consider the drift of the others components, like the resistances. But a change in the value of a resistance causes a change in the value of the voltage in the input of a operational amplifier that follows the resistance. As a consequence we can consider that a drift that appear at a resistance is simulated at the same time than the previous effected simulations.

To conclude this part, the simulations has shown us that it is possible to use the structure 2 instead the original one. It allows us remove a unuseful amplifier, that will ease the design of the PCB.

We have determined the structure with which we will work, but now we have to determine and choose the most suitable components for our application.

f) Choice of the components

As we saw earlier, the choice of the component are critical because, except the current amplifier and the ten-ohm resistance, all of them must be very accurate.

a) The choice of the operational amplifier

In the original schematics, the operational amplifier used (except for the current source) is the AD510, chosen for its low temperature coefficient (0.5 ppm/°C). But it is an old component difficult to find today and moreover we can purchase one which fits better for our application.

As we saw erlier, the critical parameters that determines the choice of the operational amplifier in our application are their temperature coefficient that should be less than 0.3 ppm/°C, their low offset voltage and offset current. The OPA277 has the caracteristics that fits very well to our application because it has a maximal drift of 0.15 ppm/°C, a low input offset voltage of 20 μV and a low bias current of 1 nA. More over its other caracteristics are quite very good (I mean for exemple its excelent common mode rejection of 140 dB).

Consequently, for its good precision caracteristics, we used the OPA277U in our design.

b) The choice of the current amplifier

The current amplifier does not need to have very good performances, I mean a very low temperature coefficient, because it is enclosed in a loop and as we explained earlier, the bad behaviour of the device is bridged by the huge gain of the operational amplifier that starts the loop.

But we need to choose an operational amplifier that is able to produce an output current at least bigger thant 160 mA (i. e. the maximal current provided to the laser diode). If the output currant of the amplifier does not have this specification, it will heat and will probably be destroyed.

The LM7372 fits to our application because it is a dual operational amplifier where each part is able to supply an out put current of 150 mA. So, by connecting the two operational amplifier of the package LM7372 in parallele, the global structure provides a maximal output current of 300 mA.

This device matchs to the specification we need, but in this kind of device it is necessary to check how the temperature increases according to the power dissipation. This parameter is discribed by the thermal resistance of the package which is about 59°C/W in the 8-Pin PSOP and since the maximal power we have to dissipate is about 0.25W, the temperature of the device increase less than 15°C. Consequently, for our application, the power dissipation is not really a problem.

c) The choice of the voltage reference

The voltage reference is the most critical component of the circuit because the stability of the output current is directely proportional to the stability of the voltage reference. So we have to use a voltage reference whose temperature coefficient and noise are as low as possible. We have seen ealier that our application needs to have current long-term-stability less than $2\mu A$. If we just consider the coarse adjustment, that means that the voltage reference fluctuation should be about $50\times\Delta I_{diode}=50\times 2.10^{-6}=0.1mV$. If we choose a voltage reference arround 7V (very common), that imply that we must have a maximal drift less than 0.0014%. If we imagine than the temperature variation is about 50°C, the temperature coefficient of the voltage reference must be smaller than 0.00028%/°C.

With a maximal temperature coefficient of 0.00005 %/°C and a maximal noise of 20 μV, the LM199AH is perfectelly adapted for this task. But unfortunatly, this component was difficult to find in the commerce, so we used an other componant, the LM399H, which belong to the same familly than the LM199AH. Although the LM399H has less good performances than the LM199AH (maximal temperature coefficient: 0.0002%/°C and maximal noise: 20 μV), this device fits nevertheless to our application.

Indeed, if we consider that the device has the maximal temperature coefficient of 0.0002%/°C and if the increase about 10°C from the initial temperature, the voltage reference increase of 0.002% of its typical value. The component's family of LM399H provides a typical voltage of 6.95V. Consequently, if the temperature increase about 10°C, the voltage reference increase from 6.95 to 6.950139V, in other word the drift is 139 μV. With the formula of the output current, we calculate the drift of the current source du to the voltage reference:

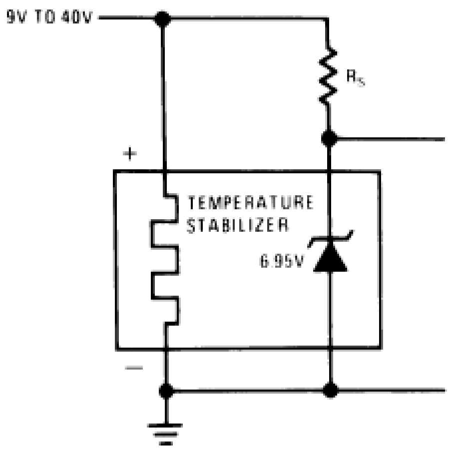
We suppose that α_{bias} =1 and α_{ramp} =0

$$Drift\_I_{diode}(10°C) = -\frac{Drift\_V_{ref}(10°C)}{50} = \frac{139 \times 10^{-6}}{50} = 2.78\mu A$$

So in the worst case and with a difference of temperature about 10°C, the contribution of the voltage reference give us a drift of the output current of $2.78\mu A$. This results just give us an idea about the scale of the drift value we can expect. In this example we considered the maximal value of the temperature coefficient, but the typical value is much lower since it is 0.00003%/°C. With this typical we calculate a drift of the output current of $0.42\mu A$. So with lower value of the temperature coefficient than the max one, the contribution of the voltage reference concerning the drift correspond to the specifications.

If during the test of the current source, the temperature coefficient is near the maximal value, it would better to change the component or to replace it by the LM199AH if it is possible to find it.

Now, we have to speak about the connection of the voltage reference. According to the datasheet, a supply voltage of 15V is enough, what is very convenient for us because we use a 15V voltage to supply all the operational amplifier. A protection resistance of about 7.5 $K\Omega$ is required as shown in the following figure.



The problem is that this resistance has to be accurate in order to not disturb the work of the voltage reference. I mean that we have to find a resistance which has a very low temperature coefficient (about 0.2 ppm/°C). But it is not easy to find such a value with so low temperature coefficient. The following array give the list of the 0.2 ppm temperature coefficient resistance we can find:

10 Ω	350 Ω	2K Ω	10K Ω
100 Ω	500 Ω	2.5K Ω	20K Ω
250 Ω	1K Ω	5K Ω	25K Ω

To have a resistance near the specified value, we can use two resistances in parallel. One resistance of $10\,K\Omega$ in parallele with one resistance of $25\ K\Omega$ gives us an equivalent resistance of $7142\ K\Omega$, acceptable in this case.

d) The choice of the resistances

-The output resistance

The output resistance is very critical as we explained earlier, so this resistance must be very accurate (temperature coefficient of 0.2ppm/°C).

In the original schematics, a resistance of $50\,\Omega$ is used and in this application the circuit produced a maximal current of 100 mA. So, with these specifications the 50-ohm resistance had to dissipate a power of $50 \times 0.1^2 = 0.5W$. In this case, it is very common to find a half-Watt fifty-ohm resistance.

But in our application, the maximal current we provide is 160 mA and if use one fifty-ohm resistance like in the original application, the power dissipation is about $50 \times 0.16^2 = 1.28W$. It is impossible to find a resistance that is able to dissipate such a power without heating. To bridge this problem, we have two solutions:

-The first one is to decrease the value of the output resistance to reduce the power disipation. The amongh of resistance we decrease in the output resistance has to be added to the protection resistance (initially ten-ohm). But the disadvantage of this solution is that it increase the sensitivity of the output current following the variation in the adjustment. For example, if we use in the output a ten-ohm resistance instead of a fifty-ohm resistance, the power disspation is $10 \times 0.16^2 = 0.254W$, but if the output voltage changes about 0.1 mV, we have a variation of the output current of about $10\,\mu A$ instead of $2\,\mu A$ in the case of a fifty-ohm resistance. So, although the power dissipation is convenient, the fact that the sensitivity of the current source increase is not acceptable for our application. It is better to use the second solution which is the following:

-The second solution is to put in parallele some resistances in order to have an equivalent resistance value of $50\,\Omega$. In this way, if we add n resistances in parallele, the value of these resistance are $n \times 50\,\Omega$ and at the same time, the value of the current that flows through each of these resistances is divided by n. Consequently, in this configuration, the power dissipation is given by $n \times 50 \times \left(\frac{I_{diode}}{n}\right)^2 = \frac{50 \times I_{diode}^2}{n}$, in other words, the power dissipated by each resistance in parallele is equal to the power dissipated by one fifty-ohm resistance divided by the number of the resistances puted in parallele. In this way, if we want a power dissipation about one quarter Watt, we need to have five resistances of $250\,\Omega$ in parallele. More over, it is easy to find precision 250-ohm resistances with only 0.2 ppm/°C able to dissipate until 300 mW without heating. Consequently, we used this configuration for our application.

Now, we are going to speak about the effect of a drift in the value of the equivalent fifty-ohm resistance on the value of the output current.

If we take the same example of an increasing of temperature of 10°C than for the voltage reference, since the 250-ohm resistances have all a temperature coefficient about 0.2ppm/°C, it means that the value of the resistances change about 2 $\mu\Omega$. If we are in the case where α_{bias} =1 and α_{ramp} =0, the initial output currant is equal to

$$I_{diode}(\Delta T = 0°C) = -\frac{V_{ref}}{50} = \frac{6.95}{50} = 139mA$$

And with a drift of the equivalent fifty-ohm resistance, we have:

$$I_{diode}(\Delta T = 10°C) = -\frac{V_{ref}}{R + \Delta R} = \frac{6.95}{50 + 2.10^{-6}} = 138.9999944mA$$

So it gives us a drift of $5.56.10^{-9} A$ concerning the output, in other words, it is negligeable compared to the effect of the voltage reference.

What we studied until now are maily the choice of the components used to fixe the current with the coarse adjustment. But as we could see previously, with a voltage reference of 6.95V and an output resistance of 50 Ω, we can provide a maximal current of 139 mA. So, to reach the typical operation current of the laser diode of 140 mA or to be able to increase the current until 160 mA, we need the fine adjustement branch. Now we are going to study how to determine the value of the resistances used for the fine adjustment.

-The fine adjustment resistance:

For the fine adjustement, we decided to have six ranges of ajustment. The first range would allow a very fine adjustment from 0 to 0.5 mA and the last range would be from 0to 20 mA to allow reaching a maximal current of 160 mA with the combination of the coarse setting. The following array gives for each fine adjustment mode the ranges we wished to have.

Modes of the fine adjustment	**Range of the mode**
Mode 1	0 to 0.5 mA
Mode 2	0 to 1 mA
Mode 3	0 to 4 mA
Mode 4	0 to 8 mA
Mode 5	0 to 12 mA
Mode 6	0 to 20 mA

But since the value of these resistances appear in the formula of the output current, they need to be accurate. That means resistance whose temperature coefficient is about 0.2 ppm/°C. Because of that fact, it would be difficult to find the exact value of the resistance allowing the previous range selection. So we calculated the needed resistance value, and if it was not possible to find it, we took the one whose value is near the calculated one.

To calculate the value of each resistance, we use the output current formula in the case where the current from the coarse adjustment is null and where the current from the fine adjustment is maximal (α_{bias} =0 and α_{ramp} =1) for each mode:

$$I_{diode} = -\frac{V_{ref}}{50}\left(\frac{R_i}{R+R_i}\right)$$

We arbitrarily fix the value of R at $10\,K\Omega$, and the calculation of the resistance R_i for each mode is given by:

$$R_i = \frac{50\,R\,I_{mode\_i\_max}}{V_{ref} - 50\,I_{mode\_i\_max}}$$

The following array give us the value of the calculated resistance for each mode:

Modes of the fine adjustment	Value of the resistance Ri
Mode 1	36
Mode 2	72
Mode 3	296
Mode 4	610
Mode 5	945
Mode 6	1680

Since for all of these values, it is impossible to find the 0.2 ppm/°C-temperature coefficient resistance, we have to take the nearest value of his avalaible kind of resistance. The following array gives the final used value of resistance and the corresponding range of current:

Modes of the fine adjustment	Resistance value	Range of the mode
Mode 1	$50\,\Omega$ *	0 to 0.691mA
Mode 2	$100\,\Omega$	0 to 1.38mA
Mode 3	$250\,\Omega$	0 to 3.39mA
Mode 4	$500\,\Omega$	0 to 6.61mA
Mode 5	$1\,K\Omega$	0 to 12.64mA
Mode 6	$2\,K\Omega$	0 to 23.17mA

**: Since no 0.2 ppm/°C temperature coefficient resistance are no avalable for the value of 50Ω, we use two 100-ohm resistances in parallele.*

With these settings, the maximal output current we can reach is:

$$I_{max} = I_{bias\_max} + I_{mode6\_max} = 139 + 23.17 = 162.17\,mA$$

Now we have to consider the effect of the drift introduced by the sensitivity of the fine adjustment resistance over temperature. As we know, the output current is given by:

$$I_{diode} = -\frac{V_{ref}}{50}\left(\alpha_{bias} + \frac{R_i}{R+R_i}\alpha_{ramp}\right)$$

If, for the mode i, we consider only a variation of the resistances R_i and R, the corresponding drift of the current is given by:

$$\Delta I = \frac{\delta I_{diode}}{\delta R}dR + \frac{\delta I_{diode}}{\delta R_i}dR_i$$

$$\Delta I_{diode}(\delta R, \delta R_i) = -\frac{V_{ref}}{50}\frac{\alpha_{ramp}}{(R+R_i)^2}(R\,\Delta R - R_i\,\Delta R_i)$$

As we can see, the variation of the current is proportional to ponderated difference of the drift of the resistance R_i and R. Consequently, the effect of the drift of these resistances is bridge. If take the previous example of a temperature shift of 10°C in each of these resistances, as there are both a temperature coefficient of 0.2 ppm/°C, the drift of R and R_i is about $2\,\mu\Omega$ (In the reallity, the temperature at each of these resistances are not exactly the same given that they are not at the same position).

If $\alpha_{ramp} = 1$ and if we consider the mode 6, the equivalent current drift given by the previous formula is:

$$\Delta I_{diode}(\delta R, \delta R_i) = -\frac{6.95}{50}\frac{10000-2000}{(10000+2000)^2}\times 2.10^{-6} = 1.54.10^{-11}\,A$$

This calculation gives us an idea about the fact that the contribution of the adjustment resistance in the current drift is negligeable compared to the other sources of contributions.

The following array gives the current drift du to each mode and for $\alpha_{ramp} = 1$:

Modes of the fine adjustment	**Corresponding current drift**
Mode 1	$2.74\times10^{-11}\,A$
Mode 2	$2.70\times10^{-11}\,A$
Mode 3	$2.58\times10^{-11}\,A$
Mode 4	$2.39\times10^{-11}\,A$
Mode 5	$2.07\times10^{-11}\,A$
Mode 6	$1.54\times10^{-11}\,A$

The selection of the fine adjustment mode is made by one six-track-circular selector.

-The choice of the potentiometer:

As we saw ealier, the potentiometers allow to adjust the magnitude of the output current (the first one for the coarse adjustment, the second one for the fine adjustment). In the formula of the output current, the magnitude is only determined by the position of the floating pin of the potentiometers (α_{bias} and α_{ramp}) and not by the value of the maximal resistance:

$$V_{out} = \frac{\alpha.R}{(1-\alpha).R + \alpha.R} V_{in} = \alpha.V_{ref}$$

So, as we saw ealier, in the ideal case, the potentiometer has no linearity default, and thus a variation of the resistance does not affect the current selection.

In the reality, the potentiometer has non linearity defaults that can occure. So the drift of the resistance over temperature is not the same along all the potentiometer. If, to simplifiy, we suppose that the average drift of the resistance between the fixed first pin of the potentiometer and its floating one is ΔR_1 and that between the second fixed pin and the floating one is ΔR_2, we have:

$$V_{out} = \frac{\alpha.(R + \Delta R_1)}{(1-\alpha).(R + \Delta R_2) + \alpha.(R + \Delta R_1)} V_{ref} = \frac{\alpha.(R + \Delta R_1)}{R + (1-\alpha).\Delta R_2 + \alpha.\Delta R_1}.V_{ref}$$

Consequently, we have to choose a potentiometer whose linearity defaults and temperature coefficient are as low as possible in order to minimize uncontrolled current setting du to these characteristics.

We fortunatly found a very high precision potentiometer with a linearity of 0.3% and a temperature coefficient of 20 ppm/°C.

To have an idea about the scale of the drift corresponding to the potentiometer sensitivity over temperature, if, as previously, we consider a temperature drift of 10°C, and with a potentiometer which has a temperature coefficient of 20 ppm/°C and a linearity of 0.3%, we can imagine that $\Delta R_1 = 200\mu\Omega$ and $\Delta R_2 = (1 + 0.003)\Delta R_1 = 200.6\mu\Omega$. In this case, if we consider that the potentiometer is settled in the middle ($\alpha = 0.5$), we have:

$$V_{out} = \frac{0.5 \times (1000 + 200 \times 10^{-6})}{1000 + 0.5 \times 200.6 \times 10^{-6} + 0.5 \times 200 \times 10^{-6}}.V_{ref} = 0.4999999999...V_{ref}$$

instead of $0.5V_{ref}$

In the case of the coarse current adjustment ($\alpha_{ramp} = 0$), we expect an output current of $I_{diode} = -\frac{0.5.V_{ref}}{50} = \frac{0.5 \times 6.95}{50} = 69.5\,mA$ and we have also a current of about $I_{diode} = -\frac{0.49999999999... \times 6.95}{50} = 69.49999999...mA$, in other words, the current shift due to the potentiometer default is negligeable (with the precision potentiometer we have chosen)

More over, since potentiometer resistance value is adjusted by a screw, mechanical instabilities could occure, but it is bridged by the high number of turns of the potentiometer. Indeed, the potentiometer we have chosen has 10 turns. Consequently, if we considere the one who works for the coarse adjustment (the worst case), the current range is 139 mA, so one turn correspond to 13.9 mA. If we suppose that the mechanical instabilities give an uncertainty of 0.05° about the angular position of the potentiometer screw, it means an uncertainty of about $\frac{0.05}{360} \times 13.9 \times 10^{-3} = 1.93 \mu A$. The resulting current drift is not negligeable but respect the specifications (we expect to have current fluctuation less than $2 \mu A$). We do not know the real value of the angular uncertainty introduced by the mechanical instabilities but we can expect to have angular uncertainty less than 0.05° given that the potentiometer is a precision.

In the case where the potentiometers would produce a biggest instability as expected, maybe it would be usefull to consider a digital way to adjust the current as it is discribed in the original circuit. On our circuit it would be easy to replace the potentiometers by digital adjustment system because the trimmers are linked to our board by one connector and it is possible to build on an other interface board the digital selection circuit to be connected to the same board.

-The choice of the others resistances:

The others resistances are not critical. So we do not need to use precision resistances.

For the laser diode protection ten-ohm resistance, the reason is that it i enclosed in a loop whose input gain is huge, so as explaining downstream, all default of the ten-ohm resistance are divided by the gain.

We discribed the way how we chosed the component of the circuit and for each the drift introduced in the output current for a temperature shift of 10°C to have an idea about the components which generate a predominating drift and those whose generated drift is negligeable.

Now, we will make the synthese of the total sources of current drift with the selected components.

g) The sources of current instability

As we know, the formula of the output current is defined as follow, for the mode i selected:

$$I_{diode} = -\frac{V_{ref}}{R_{out}}\left(\alpha_{bias} + \frac{R_i}{R + R_i}\alpha_{ramp}\right)$$

So the total current drift is difined by the sum of all the conributions as follow:

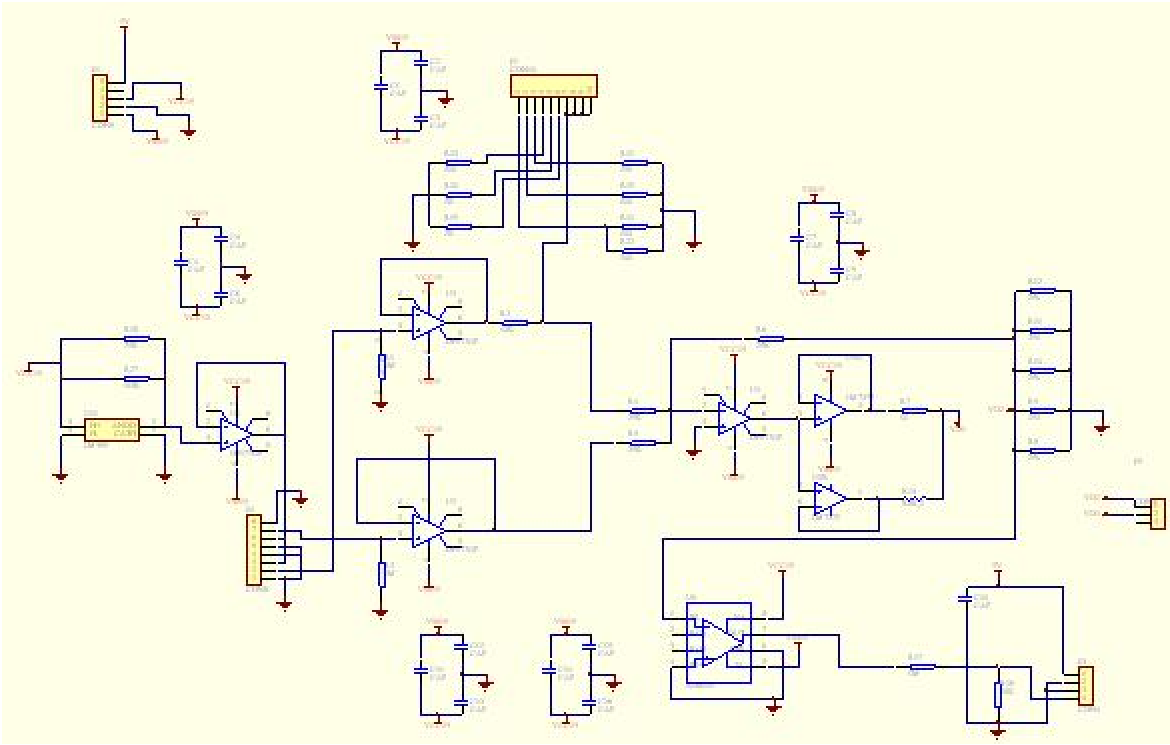
$$\Delta I_{diode} = \frac{\delta I_{diode}}{\delta V_{ref}} dV_{ref} + \frac{\delta I_{diode}}{\delta R_{out}} dR_{out} + \frac{\delta I_{diode}}{\delta \alpha_{bias}} d\alpha_{bias} + \frac{\delta I_{diode}}{\delta \alpha_{ramp}} d\alpha_{ramp} + \frac{\delta I_{diode}}{\delta R} dR + \frac{\delta I_{diode}}{\delta R_i} dR_i$$

$$\Delta I_{diode} = -\frac{1}{R_{out}}\left(\alpha_{bias} + \frac{R_i}{R+R_i}\alpha_{ramp}\right)dV_{ref} + \frac{V_{ref}}{R_{out}^2}\left(\alpha_{bias} + \frac{R_i}{R+R_i}\alpha_{ramp}\right)dR_{out} - \frac{V_{ref}}{R_{out}}d\alpha_{bias}$$

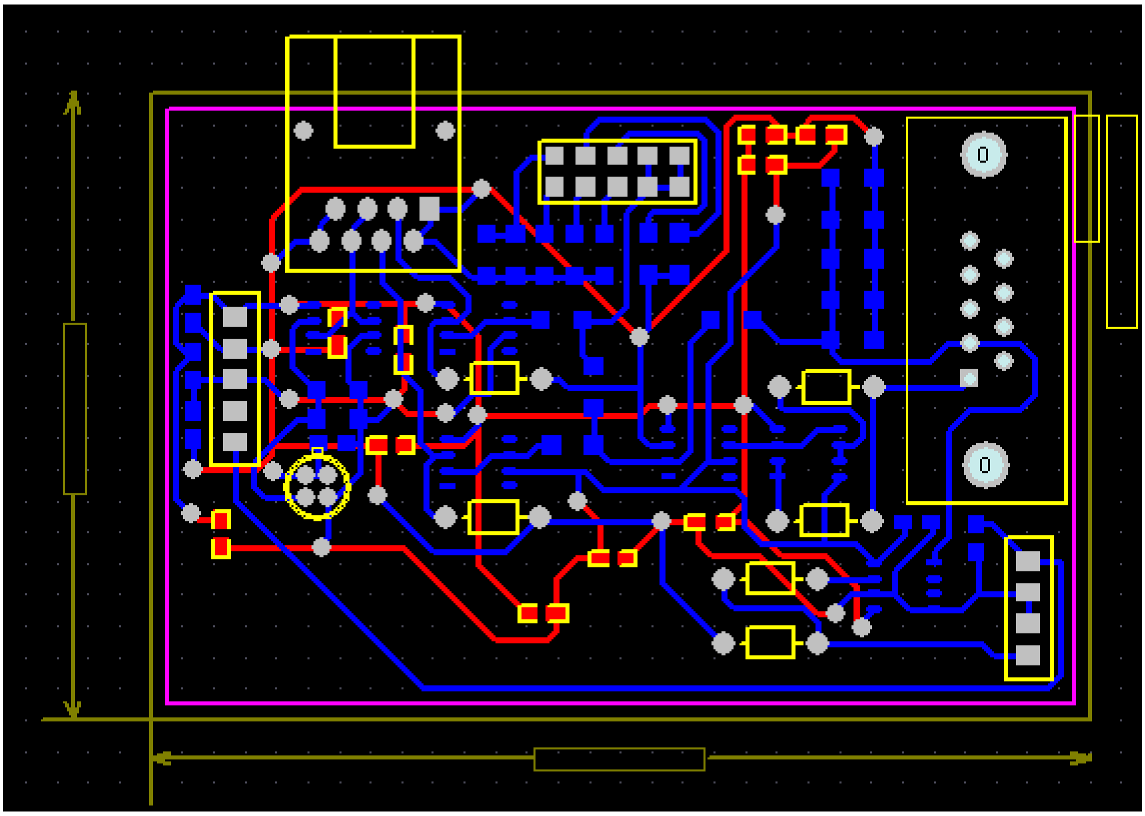
$$-\frac{V_{ref}}{R_{out}}\left(\frac{R_i}{R+R_i}\right)d\alpha_{ramp} - \frac{V_{ref}}{R_{out}}.\frac{\alpha_{ramp}}{\left(R+R_i\right)^2}\left(RdR - R_i dR_i\right)$$

h) Construction and test of the circuit

In the following figure, we are the final schematics we used to design the PCB:



And here one version of a functional PCB:



IV : Design of the temperature controller

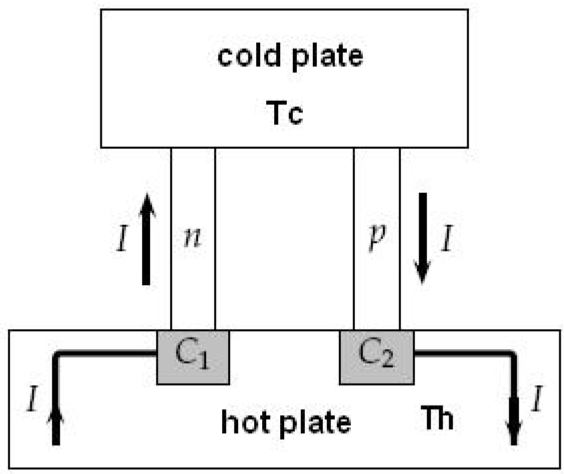
1) Specification and constraints

The laser diode is sensitive to the variation of the temperature. Its wavelength depends on the temperature of the device. In the case of the Hitachi HL7851G, the emitted wavelength changes from 781 to 793 nm when the temperature increases from 0 to 60°C, that means a sensitivity over temperature of about 0.2 nm/°C. When the laser diode is working, it dissipates heat and thus the temperature of the device increases, shifting the working wavelength. For stable and accurate operation, like in spectroscopy, an uncontrolled shift of the wavelength is not acceptable. That is the reason why, a temperature controller is necessary to make the temperature of the device and the ECDL stable. For accurate operations, temperature changes of $\pm 0.3\,°C$ are acceptable, so we will see that the electronic circuit needs to have fast response to bridge temperature changes as quick as possible and accurate performance.

To cool the laser diode and the ECDL structure, it is common and convenient to use a Peltier element. So we are going first to explain the theory of the Peltier effect before to study the circuit of the temperature controller which drive the TEC.

2) Theory of the Peltier effect

The Peltier effect is a physical phenomenon of heat moving in presence of a current between two different conductor materials. A Peltier element consists of two ceramics plates linked together by N-types and P-types semiconductor materials. The two semiconductor materials are connected electrically in series and thermically in parallel as shown in the following figure:



When a DC current is applied, (see figure), the carries moves from P-types material to N-type material, the holes follow the opposite way. By passing through the heat plate, the carries absorb heat so their thermical move is increase and number of impacts with material molecules also increase creating a conduction thermal current from the hot to the cold side.

Now we will describ the behaviour of the a Peltier element by its physical equations to understand how it works in practice.

But first, we need to speak about the different ways of heat transfer:

a) The different heat transfer modes

α) Radiation

All materials at a temperature different than the absolute zeroth, emitt electromagnetic radiations. The power of the emitted radiations depends on the temperature of the material. The more the material is hot the more the radiation is intense.

When two materials, at a different temperatures, are in a small distance each other, the both receive the electromagnetic wave emitted by the other, but as the hot material emitt more radiations than the cold one, heat is exchanged between the hot and cold object. The hot material globally loss heat (because it emitt more than it receive) and the cold one gains heat until both reach the same temperature.

The radiation heat trasnfert is expressed by the following equation:

$$Q_{rad} = F.e.s.A.\left(T_{hot}^4 - T_{cold}^4\right)$$

Where

-F is the shape factor of the material (The electromagnetic transfer depends on the shape of the material).

-e stand for the emissivity of the material

-s is Stefan-Boltzmann constant ($5.667 \times 10^{-9} W / m^2 K^4$)

-A is the area which receives radiations of the cold material

-T_{hot} is the temperature of the hot material

-T_{cold} is the temperature of the cold material

In the case of a Peltier element, the radiation heat transfer is not desired and is often negligeable in comparison with the other heat transfer modes. It is the case when the Peltier element operates in a gazeous environment with small difference of temperature.

β) Convection

The convection heat transfer occurs when a fluid (gas) flows between two materials at different temperatures. The amount of gas in the vicinity of the hot material is heated and finally is at about the same temperature than the object and in cold side the fluid in the vicinity of the object reaches about the cold temperature. The difference of temperature of the gas between the two sides generates a difference of pressure. The hot gas is more dense than the cold one. Thus, a gas current flowing from the hot side to the cold one is created producing heat transfer between hot material and cold one until the both reach the same temperature (the hot material loss heat whereas cold one gains it).

The convective heat load is proportional to the surface of exchange and the difference of temperature as discribed the analytical equation:

$$Q_{conv} = h.A.(T_{hot} - T_{cold})$$

where
-h is the convective heat transfer coefficient ($W / m^2 K$)
typicaly, in the normal pressure $h = 21.7 W / m^2 K$
-A is the common surface between the two materials
-T_{hot} is the temperature of the hot material
-T_{cold} is the temperature of the cold material

For small temperature's differences, the convective heat transfer mode is negligeable compared to the conduction one.

χ) conduction

Conduction heat transfer occurs with molecular thermical moving. When the temperature of one material increase in one of its point, the molecules composing the material moves more and more fast and heat is transfered by impact to others molecules in the colder area. Thus heat is transfered by impact from hot side to cold side of one material.

The equation describing the conduction heat transfer mode is the following:

$$Q_{cond} = \frac{\lambda.A}{L} \Delta T$$

where:
-λ is thermal conductivity of the material ($W / m.K$)
-A is the cross section area of the material
-L is the length of the heat path
-ΔT is the difference of temperature across the heat path
In the Peltier element, ΔT is the difference of temperature between the hot and the cold plate and L is the length of the semiconductor materials.

The conduction transfer mode is the most significant in the TEC because, on one hand the effects are not negligeable compared to the other modes and, on the second hand the conduction mode is directly linked to the electric current flowing in the Peltier element as we will see more closely later (The reason is that the electrons flowing in the materials heat in the vicinity of the hot plate and spread heat by impact to the molecules of the material).

In the theory of the Peltier element we will only consider the case of the conduction because this is the dominating way of heat transfer.

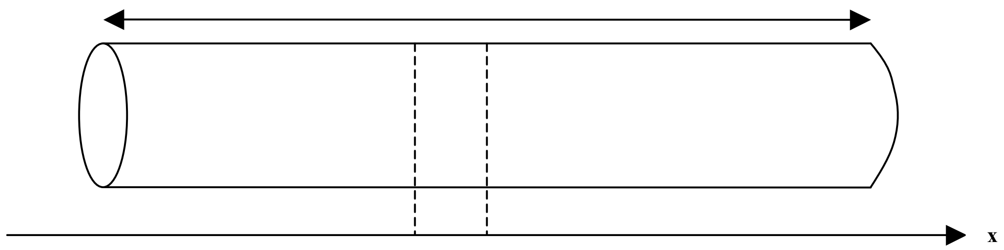
b) The physical equations of the Peltier effect

As we saw before, the Peltier is composed by two semiconductors materials (N-type and P-type) between two plates (One of them is the cold one and the other the hot one). In the Peltier element, the heat transfer occurs between the two plates via the two semiconductor materials.

So we need to explain the thermoelectric effect that appears in one semiconductor (or conductor) material before studying the complet structure. For this we will study the case of a cylindric material.

α) case of one conductor (thermo electrical coupling)

We consider a cylindric semiconductor whose length is L and surface is Σ, represented in the following figure:



In the cylindrical material, we will express the expression of the entropy created per volume unit and per time unit. In a thermodynamical system, the entropy stands for the degree of desorder in this system. When the entropy is null, the system is in its initial and stable state and its order is perfectly respected (In our application, the order would represent the difference of temperature between the cold and the hot plate). In all system, the natural variation of the system (the spontaneous variation) tends to increase the entropy, in other words, the desorder (in our case the increasing of the temperature's difference rises the entropy). To invert the spontaneous processus, it is nessecary to supply energy to the system but, the more entropy is high, the more the required energy to invert the spontaneous processus is high. The goal is to have a global created entropy null.

To express the entropy created by volume unit and by time unit, we will first express the one created by a system where only a conduction heat transfer occurs, and then the one where only an electrical conduction occurs before seeing the case of a thermoelectrical coupling system, and discussing about it.

In the first case, the entropy created per volume unit and time unit by a system with only conduction phenomenon is expressed as follows, if we consider the problem in one dimension (x):

$$\sigma_q = j_q \frac{d}{dx}\left(\frac{1}{T}\right)$$

where j_q is the thermal current density (W/m^2)

This formula implies that the creation of entropy depends on the variation of the temperature following the x axis. If between x and x+dx, inside the material, the temperature T(x) of the x position is superior of the one T(x+dx) of the x+dx position, exchanged entropy between x and x+dx is negative. So energy (thermical current) is exchanged from x to x+dx because the spontaneous reaction occurs where creation of entropy is negative. We have the opposit case if T(x+dx) is superior to T(x).

If we generalize this expression in three dimentions, the previous formula is written as follows:

$$\sigma_q = \vec{j}_q.\vec{\nabla}\left(\frac{1}{T}\right) = \vec{j}_q.\vec{f}_q$$

Where f_q is the thermodynamic thermal force conduction: $\vec{f}_q = \vec{\nabla}\left(\frac{1}{T}\right)$

In the second case, the entropy created per volume unit and time unit by a system where only electrical current is flowing inside the component following the x axis is expressed as follows:

$$\sigma_e = -j_e.\frac{1}{T}\frac{\partial(V)}{\partial x}$$

Where V is the electrical potential in one point (position x) of the material.

This formula implies that the created entropy is the opposit of the variation of the potential following the x axis. If, between x and x+dx, V(x) is superior to V(x+dx), the created entropy is negative.

If we generalize this expression in three dimentions, the previous formula is written as follows:

$$\sigma_e = -\vec{j}_e.\frac{\vec{\nabla}(V)}{T} = -\vec{j}_e.\vec{f}_e$$

where $\vec{f}_e$ is the electric force.

Now, we are considering the entropy created in a case of a thermoelectrical coupling. It is in fact the sum of the entropy created by the thermal force and the one created by the electric force.

$$\sigma = \vec{j}_q.\vec{\nabla}\left(\frac{1}{T}\right) - \vec{j}_e.\frac{\vec{\nabla}(V)}{T}$$

We can see with this formula that in order to cancel the creation of entropy, if the thermal exchange occurs in one direction, the opposit electrical force must be applied. I mean that if T(x)>T(x+dx) and $\vec{j}_q$>0 so σ_e is negative. To have a creation of entropy null, if $\vec{j}_q$>0, the appropriate value of the difference of potential has to be applied and has to be negative.

As we can see, thermal transfer and electric current are linked and the thermoelectrical coupling is expressed by the following expression

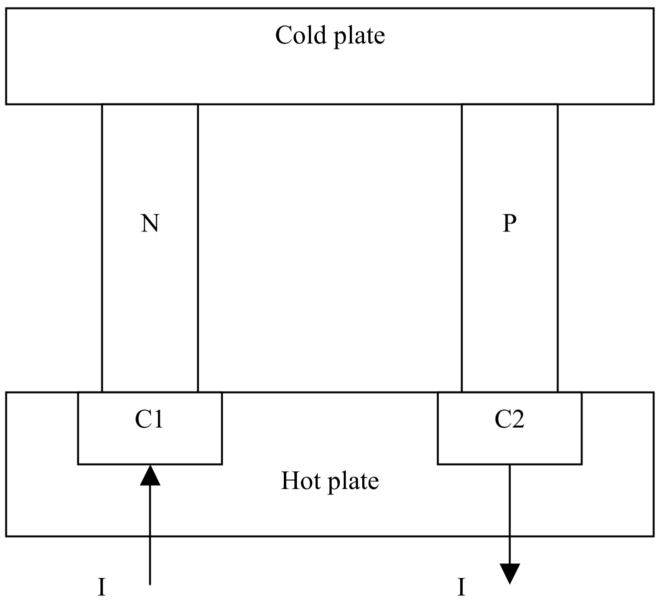
$$\vec{j}_q = -\lambda.\vec{\nabla}(T) + \varepsilon T \vec{j}_e$$

where ε is the thermoelectrical coupling coefficient and λ is the thermal conductivity of the material.

From this equation, which characterizes the thermoelectrical coupling in one material, the expression of the exchanged power follows from it.

β) case of the TEC structure

We consider now one peltier composed by two plates, one N-type semiconductor material and one P-type semiconductor material. The significant carriers in the N-type semiconductor are electrons and holes for the P-type semiconductor.



The expression of the exchanged power between the two plates arises directly from the relation of the conduction current and the electric current (see the previous formula).

The heat power absorbed by the cold plate is expressed as follow:

$$Q_{cold} = (\varepsilon_p - \varepsilon_n).I.T_{cold} - \frac{1}{2}RI^2 - K(T_{hot} - T_{cold})$$

And the heat power lost by the hot plate is given by:

$$Q_{hot} = (\varepsilon_p - \varepsilon_n).I.T_{hot} + \frac{1}{2}RI^2 - K(T_{hot} - T_{cold})$$

Where

- ε_p is the thermoelectrical coupling coefficient of the P-type semiconductor
- ε_n is the thermoelectrical coupling coefficient of the N-type semiconductor
- R is the equivalent resistance of the Peltier element (we suppose that the resistance is only due to the semiconductor materials and it is the same for the both)

$$R = \frac{L}{\Sigma}\left(\frac{1}{\gamma_n} + \frac{1}{\gamma_p}\right)$$

and γ_n and γ_p are respectively the electrical conductivity of the N-type and the P-type semiconductors.

-K is the thermal conductance of the Peltier element

$$K = \frac{\Sigma}{L}(\lambda_n + \lambda_p)$$

and λ_n and λ_p are respectively the thermal conductivity of the N-type and the P-type semiconductors.

-I is the current flowing through the Peltier element.
- T_{hot} is the temperature of the hot plate
- T_{cold} is the temperature of the cold plate

Thus the power exchanged (corresponds to the require electric power) between the hot and the cold plate is expressed as follow:

$$P_e = Q_{hot} - Q_{cold} = (\varepsilon_p - \varepsilon_n).I.(T_{hot} - T_{cold}) + R.I^2$$

In an industrial Peltier element, there are more than one pair of semiconductors. The previous formula are valid but we have to multiply them by Nb, the number of pairs of semiconductors (N-type and P-type).

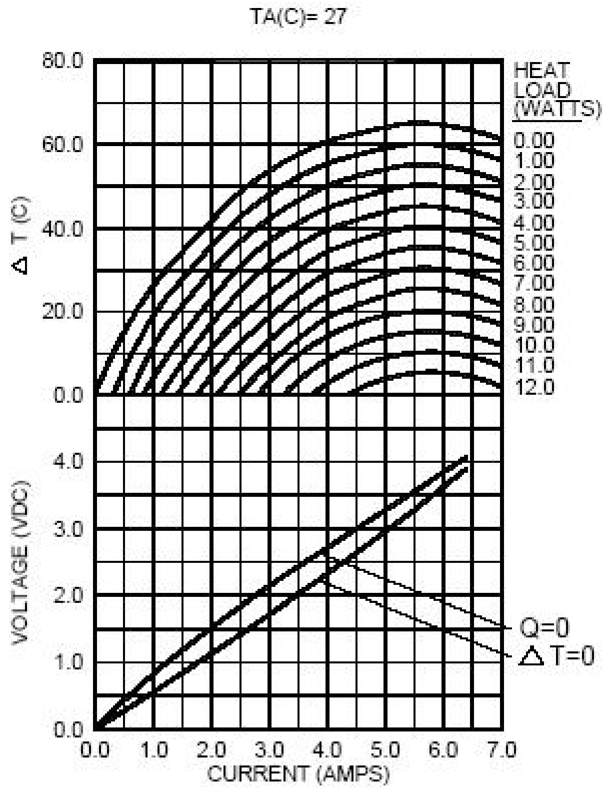
$$P_e = Nb.(\varepsilon_p - \varepsilon_n).I.(T_{hot} - T_{cold}) + Nb.R.I^2$$

and the corresponding voltage is optained by deviding the expression of the power by the current I:

$$U_{DC} = Nb\left[\left(\varepsilon_p - \varepsilon_n\right).\left(T_{hot} - T_{cold}\right) + R.I\right]$$

Our goal is to keep the difference of temperature between the cold and the hot plate as constant as possible. We expect to have temperature's variation smaller than at least 0.5 mK on the hot plate (laser diode).

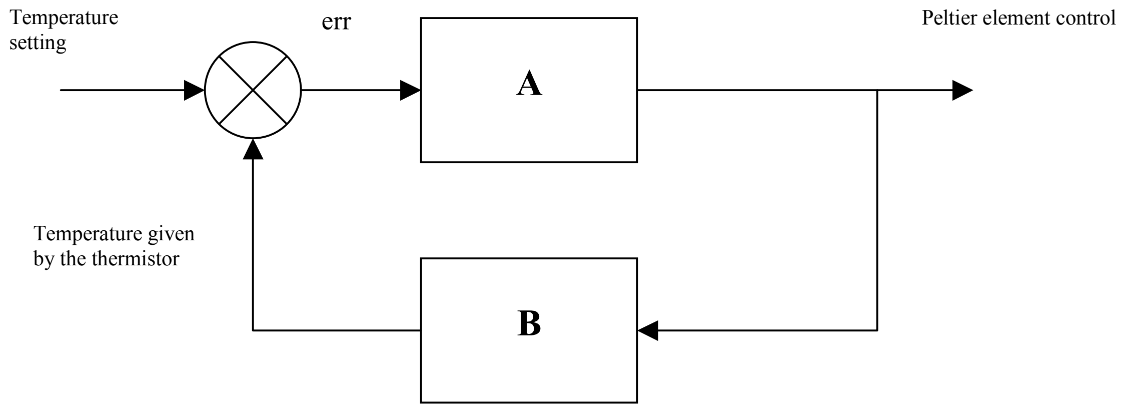
Initially, the entropy of the system is null when the electrical power P_e is null. If the temperature change, the current increase according to the zeroth-power curve. To bridge the change of temperature, electrical power is increasing by injecting proper driven current or driven voltage. The speed of the temperature change depends on the response of the temperature controller.



These previous curves show the characteristics of the TEC 3-6, used to regulate the temperature of the laser diode. The device can support a maximal voltage of 3.2V and a maximal current of 5.4A. So our temperature controller must includ a protection part in order to avoid being out of the specification ranges and destroying the device.

3) The different electric architecture to control the Peltier element

The power dissipation of the peltier element is controlled by a buckled system, where the direct function transfer is aimed to process the error signal, i. e. the difference between the input voltage, proportional to the temperature setting and the voltage given by the feedback (proportional to the real temperature of the laser diode). The feedback depends on the response of the Peltier element, the nature of the thermical contact between the Peltier element and the laser diode mount and the response of the thermistance.



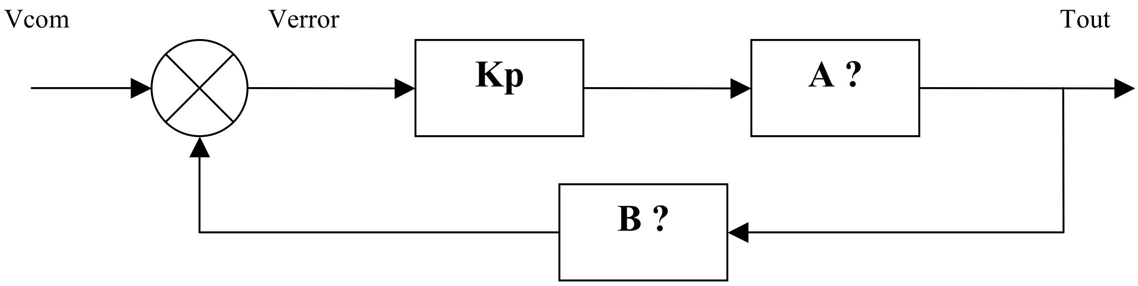
The transfer function in the openned loop is determined by the architecture of the electronic circuit whereas the feedback depends on different unknown parameters like the analytic transfer function of the Peltier element, the way how is the thermical contact between the TEC and the laser mount and the response of the thermistance. So, although the open-loop transfer function is well known, it is impossible to simulate the stability of the system due to the unknown feedback. What we can do is to build a circuit whose transfer function is adjustable to be adapted to the feedback.

We have different possible structures which are proposed by the TEC manufacturers or described in scientific reviews to drive a Peltier Element. We will describ the different architectures and discuss about which of them performs better for our application.

a) First structure : the proportional control

The proportional control architecture is the simpliest structure to drive the peltier element. In this architecture, the output is proportional to the signal error, that is to say, the difference beteween the temperature setting and the temperature given by the feedback.

The following diagram in the Laplace field describs the relationship between the temperature setting and the output:



Kp is the static gain of the controller circuit.
A is the unknown peltier element response
B is the response of the thermistor

The transfer function of the close-loop is expressed by the following expression:

$$T_{out} = \frac{K_P A}{1 + K_P AB} V_{com}$$

We don't know the analytical transfer function of the peltier element and those of the thermistor. Consequently it is impossible to study the stability of the system with precision. We can just approximate the transfer function by simple known models. For example, concerning the thermistance, we know that the time constant is about 15s. So we will express the response of the thermistance as a first order filter (just to simulate a model and see which controller performs better):

$$B = \frac{B_o}{1 + \tau S}$$

where $\tau = 15s$ and B_o is the static gain.

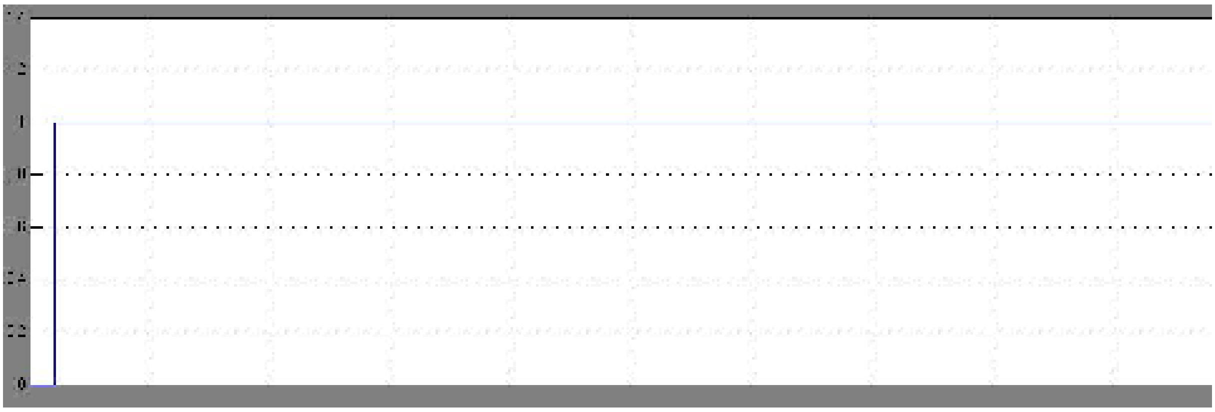
Conserning the TEC, we could read on several manufacturer websites than the typical time response to an unit step is between 15s and 40s and often arround 30s. So if we consider that our Peltier element has a first order transfer function with a time constant $\tau_p = 30s$, we have

$$A = \frac{A_o}{(1 + \tau_p S)}$$

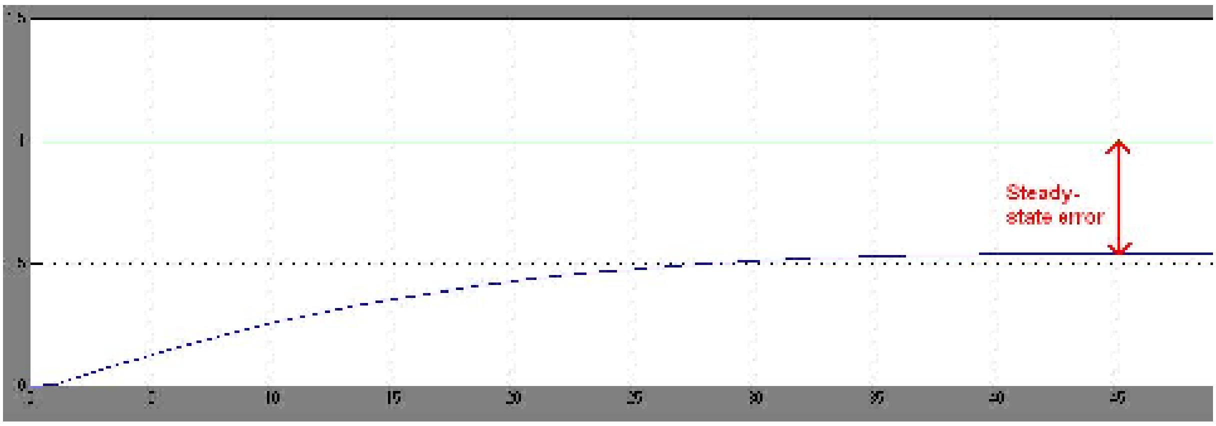
we can simulate close-loop transfer function with this model that give us the following expression:

$$H = \frac{K_P A_o (1+\tau S)}{(1+K_P A_o B_o)+(\tau+\tau_p)S+(\tau\tau_p)S^2}$$

With modelsim (Matlab7), we simulate the unit step response of this system. The step signal is shown in the following figure:

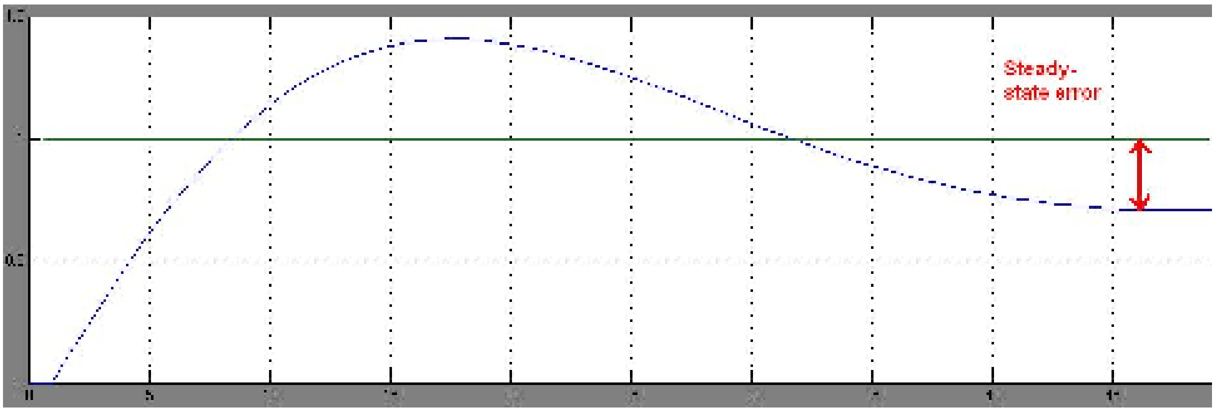


First, we simulate with a proportional gain of 1 and the response of the system is shown on the following figure:

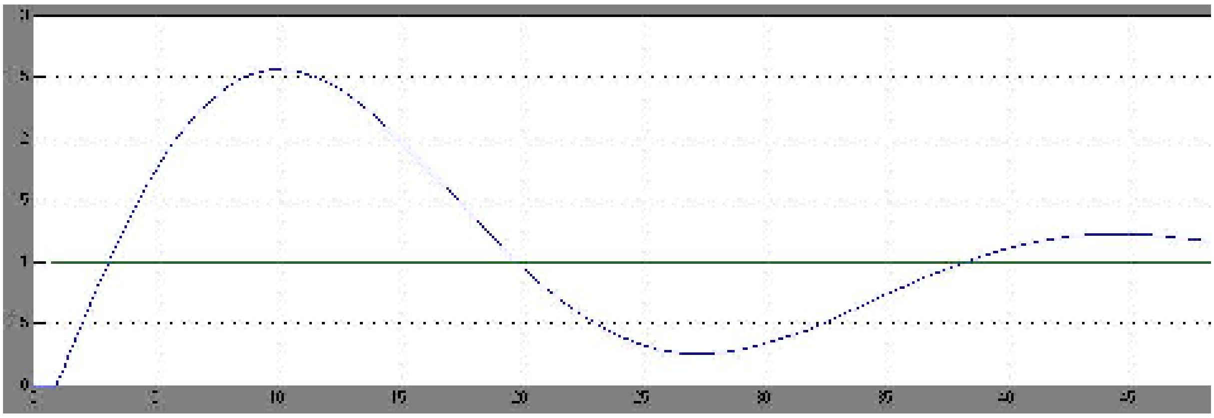


As we can see, on one hand, the time rise of the response is very big and correspond to a time constant of 30s, in other words the time constant of the transfer function **A** (the supposed Peltier element's transfer function). On the other hand, the steady-state error is not negligeable. In this case it is about 0.55.

Now, if we increase the value of the proportional gain at 5, the optained response is the following:



In comparison with the previous case, the time rise and the steady-state error have been decreased (we will see why after the introduction of the matlab simulations). However, we can observe an undesirable overshoot in the transient response of the system. If we increase more the proportional gain (at 15 in the following figure), absorbed oscillations appear in the transient response arround the final value and the first overshoot magnitude is bigger than in the previous simulation. But the time-rise is still decreasing and the steady-state error is becoming negligeable (but is not canceled, we are going to see why).



All the previous simulation illustrate the advantages and disadvantages of a proportional gain controller. The advantages are that it is a very simple structure easy to implement and if the plant transfer function is stable, it also give a stable response. Moreover, it reduces the time rise of non-controlled system (i.e., the system defined only by the transfer function of the Peltier Element A) and decreases the steady-state-error, with a big proportional gain. Indeed, the expression of the steady-state-error is optained by the following formula:

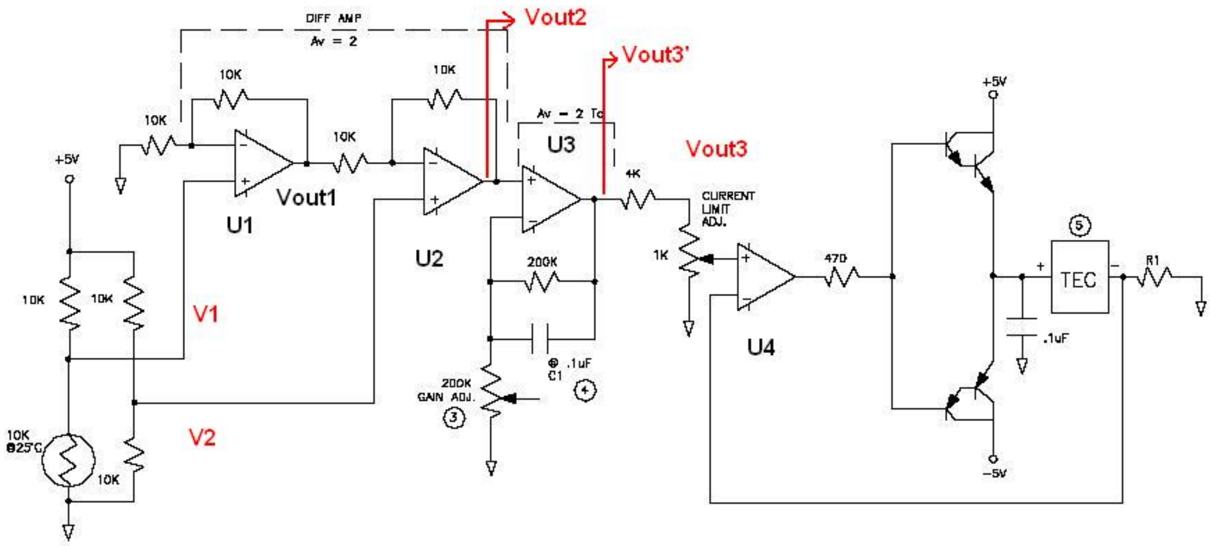
$$err = \lim_{S \to 0} S \times \frac{Vcom}{1 + K_p AB}$$

For our application, since the laser system is aimed to work at a constant temperature, the command temperature is a step. In the Laplace field, $Vcom = \frac{Vo}{S}$

$$\text{So: } err = \lim_{S \to 0} \frac{Vo}{1 + K_p AB} = \frac{Vo}{1 + K_p A_o B_o} \approx \frac{1}{2} \text{ in our case.}$$

The expression of the steady-state-error shows us that the more the proportional coefficient K_p is high, the more the steady-state-error is low. However, a big value of K_p degrades the transient response. As a consequence, with this kind of controller, it is impossible to have at the same time a fast response and a correct transient without overshoot. Moreover, if we find a compromise between low rise time and good transcient response, although the steady-state error is reduced by the proportional gain, it is not canceled. Consequently for our application, since we need a good precision, this basic controller is not convenient.

For this basic kind of temperature controller, Marlow industries (the manufacturer of the TEC we use) propose a schematic as shown in the following figure:



This is not the architecture we used for our application because of the reasons we have just explained, but we will nevertheless detail it quickly. Indeed the general architecture is the same for all kind of controllers. It can be divided into 3 parts:

-(1) the error detection
-(2) the control transfer function
-(3) the output current setting

(1) The error detection circuit:

The detection is realized by a wheaston bridge where the two output potentials are expressed as follow:

$$V_1 = \frac{R_T}{R_o + R_T} V_{cc} \qquad \text{and} \qquad V_2 = \frac{V_{cc}}{2}$$

Two operational amplifiers downstream are aimed to give a the difference between the reference signal (given by the reference resistor of the bridge) and the feedback (temperature detected by the thermistor). For that, the two amplifiers (U1 and U2) are configured to multiply the differential voltage ($V_2 - V_1$) of the detection bridge by a gain of 2. Indeed, if we consider the first operational amplifier working in the ideal conditions, we have:

$$V_{U1}^{+} = V_1 = V_{U1}^{-} = \frac{V_{out1}}{2} \text{ consequently } V_{out1} = 2V_1 = \frac{2R_T}{R_o + R_T} V_{cc} .$$

And with the second operational amplifier, we get:

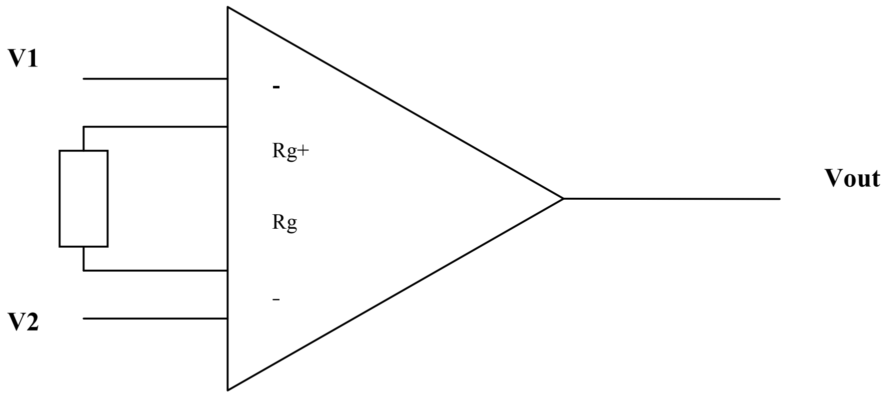
$$V_{U2}^{+} = V_2 = V_{U2}^{-} = \frac{V_{out1} + V_{out2}}{2} = \frac{V_{cc}}{2}$$

$$\text{so} \qquad V_{out2} = 2.(V_2 - V_1)$$

The output given by the two successive first operational amplifiers is the differential voltage of the detector bridge multiplied by two. The reason for that is to optain the following expression:

$$V_{out2} = \left(1 - \frac{2R_T}{R_o + R_T}\right).V_{cc} = \frac{R_o - R_T}{R_o + R_T}.V_{cc}$$

We will see further (in the description of the selected architecture: the PID controller) that this function allows to have a linear variation of the voltage V_{out2} according to the temperature. The architecture of the error detector can also be improved by using a instrumentation amplifier instead of two operational amplifiers. It allows to cancel better the common mode parasites and to set the input gain of the controller loop via a resistance connected to the two gain-setting inputs of the instrumentation amplifier (see the following figure):



The gain of the instrumentation amplifier is given by $G = 1 + \frac{R_{intern}}{R_G}$ where R_{intern} is the specific resistance of the instrumentation amplifier (given in the datasheet of the device) and R_G is the resistance settled to select the appropriate gain.

<u>(2) The amplification stage</u>

The amplification stage in this architecture is composed by one operational amplifier (U3) followed by a voltage divider.

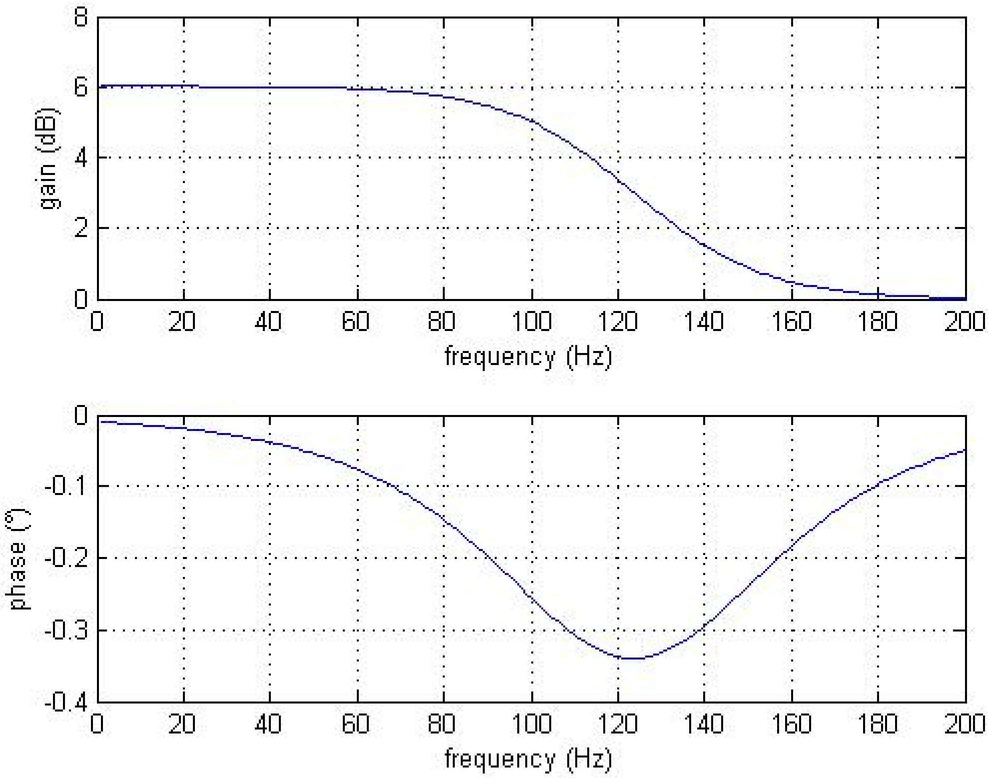
If we suppose the operational amplifier ideal, we can write that:

$$V_3^+ = V_{out2} = V_3^- = \frac{R_{gain}}{R_{gain} + \left(\frac{R}{1+PCR}\right)} V'_{out3} = \frac{R_{gain}}{R_{gain}+R} \times \frac{1+PCR}{1+PC\frac{R_{gain}R}{R_{gain}+R}} V'_{out3}$$

Consequently $$V'_{out3} = \left(1 + \frac{R}{R_{gain}}\right) \times \frac{1 + j\frac{\omega}{\omega_1}}{1 + j\frac{\omega}{\omega_2}} V_{out2}$$

Where $$\omega_1 = \frac{R_{gain}+R}{R_{gain}RC}$$ and $$\omega_2 = \frac{1}{RC}$$

The bode diagram corresponding to a static gain of 2 ($R_{gain} = R$) is given in the following figure:



The low frequencies (lower than 100 Hz) are amplified by the static gain ajusted by the potentiometer R_{gain} wherease the frequencies above are twice less amplified (shift of -6 dB between low and "high frequencies). Hence, the filter privileges the low variations of the useful error signal in comparison with the high frequencies that correspond to parasits.

After the operational amplifier, the voltage divider finally gives the following voltage:

$$V_{out3} = \left(\frac{\alpha R_2}{R_1 + R_2}\right)\left(1 + \frac{R}{R_{gain}}\right) \times \frac{1 + j\dfrac{\omega}{\omega_1}}{1 + j\dfrac{\omega}{\omega_2}} \times \frac{R_o - R_T}{R_o + R_T} V_{cc}$$

In low frequencies, that is to say, in the useful working, the proportional gain is adjusted by the potentiometers R_{gain} and R_2 in order to avoid having a too big maximal current flowing through the Peltier element and to adjust the stability. Indeed, since the feedback depends on the Peltier element and the thermal isolation, we need to adjust the gain of the action transfer function before using the temperature controller. The gain have to be selected in order to have an as quick and stable response as possible following the used Peltier element.

Now we have to explain how the peltier element is controlled by the circuit. It is the role of the third part which turn the output voltage of the proportional gain into a proportional currant that drive the peltier element.

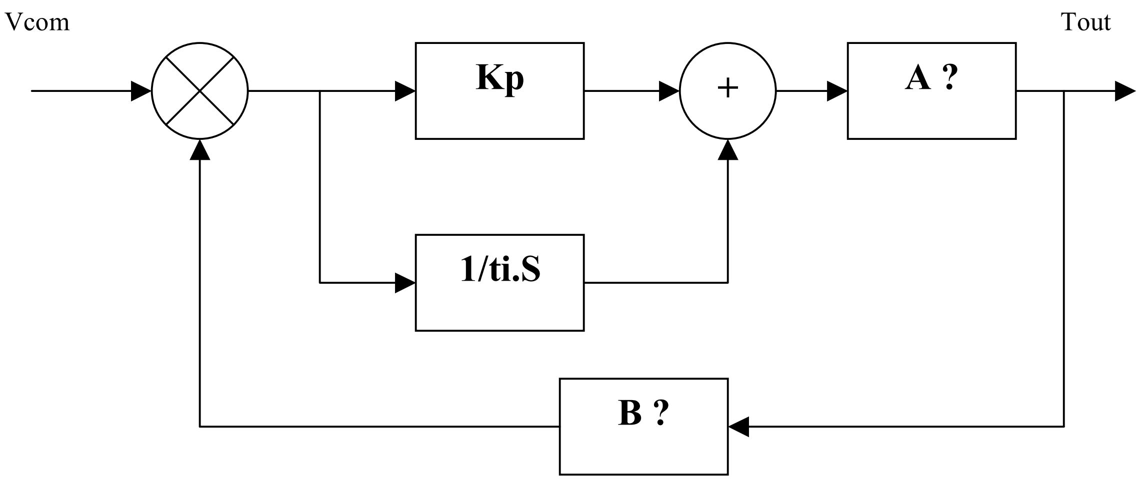
(3) The convetion voltage/current

The third part is similar to the end of the current source where the voltage is converted into a proportional current, and it is exactly the same architecture that we will use in the schematic of the temperature controller we designed. So we will describe this part more closely later.

b) Second structure : the PI control

The main problem with the proportional gain temperature controller was that there was always a residual error, even after the controller has settled to the final state. The add of an integrator in parallele with the proportional gain allows to eliminate the steady-state error and to decrease the rise time as we are going to see it.

The general architecture is summarized in the following bloc schematics:



τ_i is the integrator time constant.

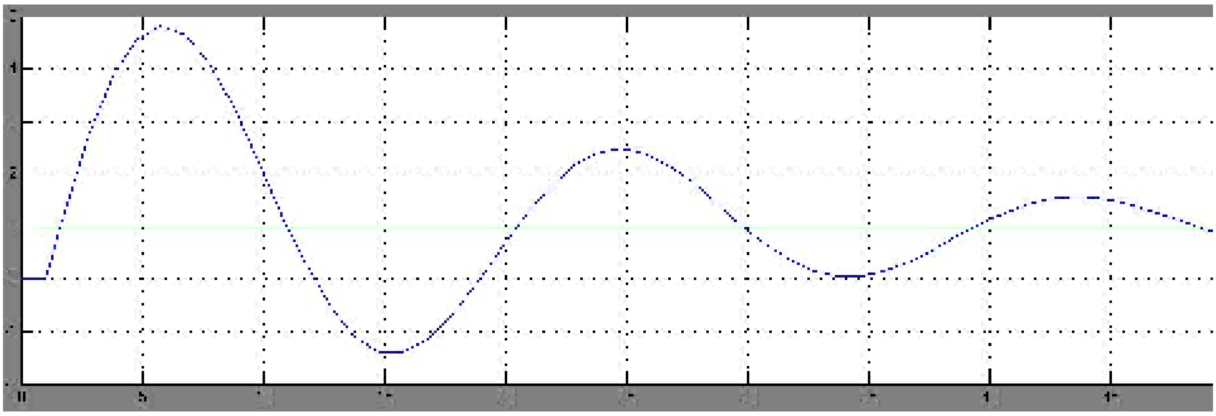
The transfer function of the PI controller is expressed as follows:

$$H_{cont\_PI} = \frac{1 + \tau_i S}{\tau_i S}$$

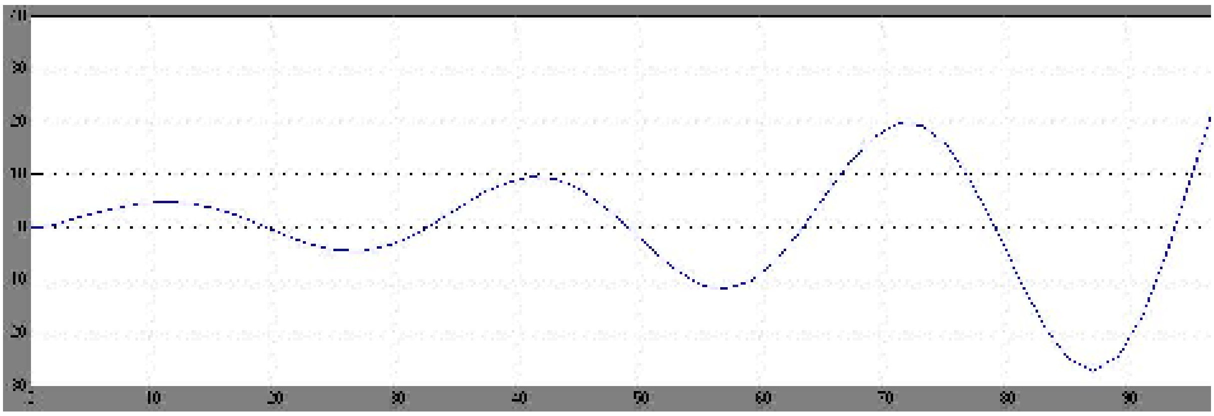
If we approximate the transfer function of the peltier element and the thermistance as the same way as for the proportional gain, we get the close-loop PI transfer function expressed as follows:

$$H_{PI} = \frac{A_o\left(1 + \left(\tau + K_p \tau_i\right)S + K_p \tau_i \tau\, S^2\right)}{A_o B_o + \left(1 + K_P A_o B_o\right)\tau_i\, S + \left(\tau + \tau_p\right)\tau_i S^2 + \left(\tau_i \tau_p \tau\right)S^3}$$

As we can see, in comparison with the proportional gain controller, the addition of one integrator increments the degree of the denominator, that is to say it adds a pole to the transfer function increasing the instability of the system. But at the same time, it improve the time rise. This fact is illustrated in the following simulation where the proportional gain K_P is equal to 15 (like the last proportional gain simulation) and the integrator time constant τ_i is 10s.



When we decrease the integrator time constant, the time rise decrease (for the same value of the proportional gain as previously), but at the same time the transient response get worse and the overshoot become larger. If we decrease more the integrator time constant, it is possible that the system become unstable. It is illustrated with following simulation where K_P is 15 and τ_i is 0.33s (the response is unstable):



So, like in the case of the proportional gain, this type of controller push us to do a compromise between the speedness of the system and the quality of the response. But if we express the steady-state error as follows, we can observe that it is canceled.

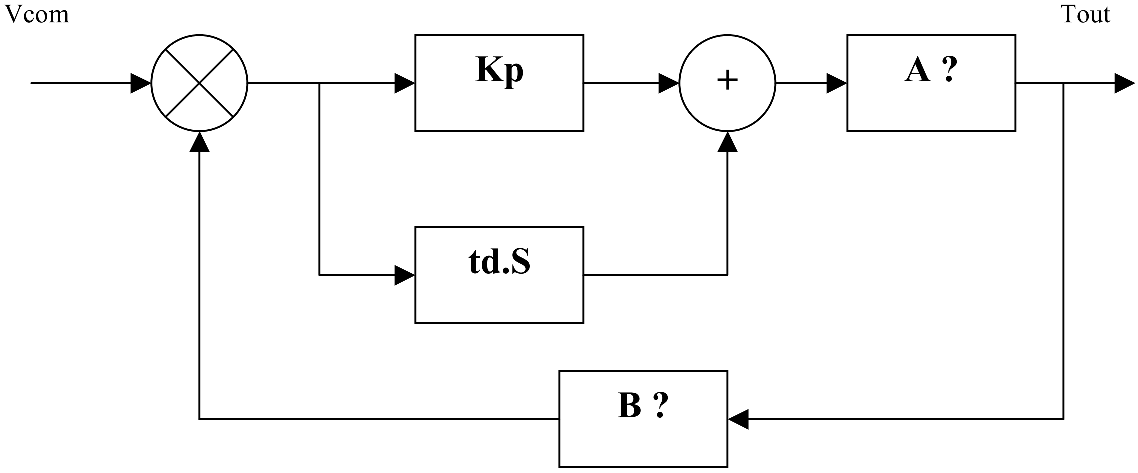
$$err_{PI} = \lim_{S \to 0} \frac{Vo}{1 + \frac{1 + \tau_i S}{\tau_i S} AB} = 0$$

Consequently, even if we are not able to have, at the same time, a good transient response and a low rise time, in comparison with the proportional gain, the PI controller allows to cancel the static error. But for our application, we need to have a fast response, so such a system is not enough convenient. We need to improve it.

Now we are going to study the effect with the addition of a derivative in parallel with the proportional gain.

c) Third structure : the PD control

We add a derivative in parallele with the proportional gain as shown as follows:



where τ_d is time constant of the derivative.

The transfer function of the PD controller is expressed as follow:

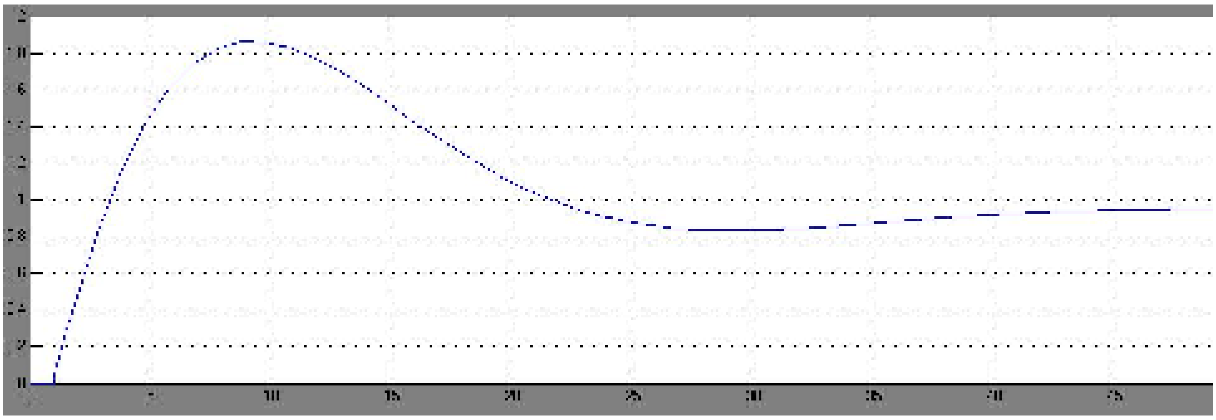
$$H_{cont\_PD} = K_p + \tau_d S$$

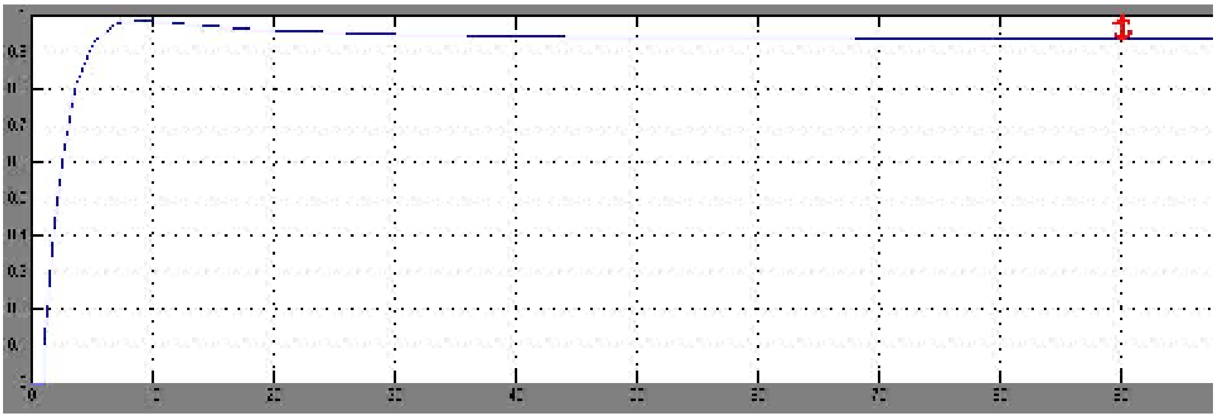
If we consider that the same transfer function A and B than the others case the close-loop transfer function of the system is expressed as follow:

$$H_{PD} = \frac{A_o\left(K_P + \left(K_P\tau + \tau_d\right)S + \tau_d\tau S^2\right)}{\left(1 + K_P A_o B_o\right) + \left(\tau + \tau_p + A_o B_o \tau_d\right)S + \left(\tau\tau_p\right)S^2}$$

As we can notice when we look the expression of the PD-transfer function is that in comparison with the proportional gain, the derivative does not add poles (The degree of the denominator of the PD controller is the same than that of the proportional gain). But the derivative adds one degree to the numerator that increase the stability of the system giving the transient response better.

The two next figures show two step simulations of the PD structure with two different derivative time constants: The first one is lunched with K_P=15 and τ_d=50s and the second one with K_P=15 and τ_d=200s





As we can see, the more the derivative time constant is high, the more the transient response is better. Moreover, in comparison with the others systems, this structure does not impose us any compromise. But if we express the steady-state error (with a step input command) we notice that it is the same error than with only one the proportional gain (in our example we mean a proportional gain equal to 15), so the derivative structure does not improve the static-error:

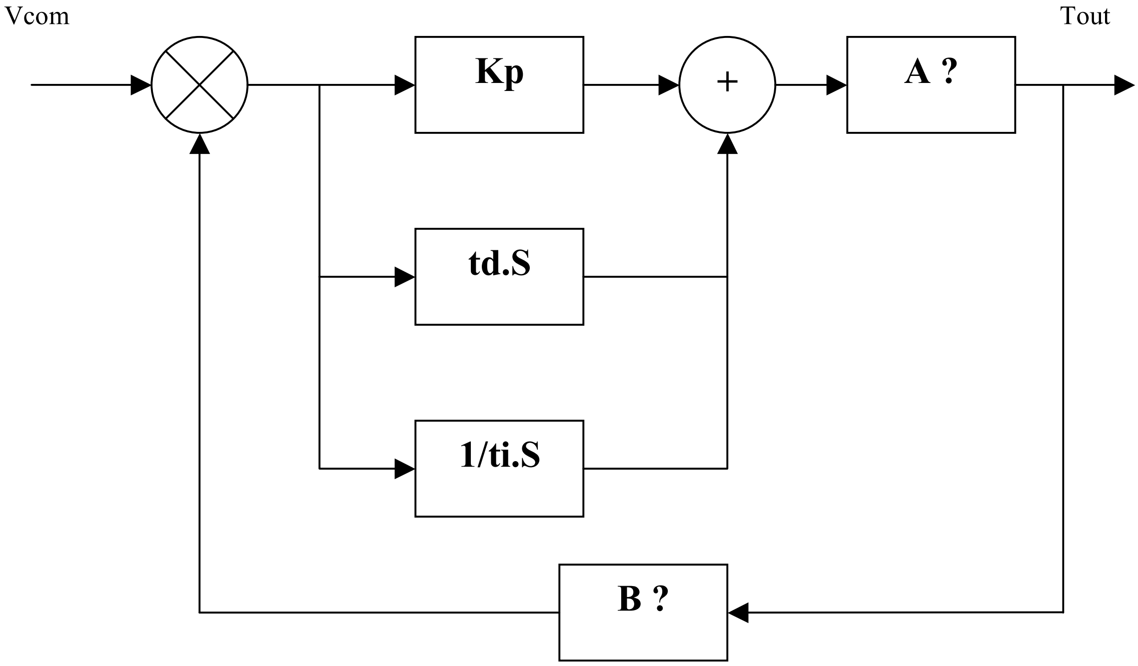
$$err_{PD} = \lim_{S \to 0} \frac{Vo}{1 + (K_P + \tau_d S)AB} = \frac{Vo}{1 + K_p A_o B_o}$$

As a consequence, a simple PD controller is not enough for our application given that it does not allow a cancellation of the steady-state error.

But the characteristics of a PD controller can bridge the default a PI controller and in the opposite side, the qualities of the PI controller bridge the poor rise time and precision of the PD controller. This structure is called PID (Proportional Integrator Derivative) controller.

d) Fourth structure : the PID control

In the PID controller, the proportional gain, the integrator and the derivative are connected in parallele as show the following bloc schematics:



Following this configuration, the transfer function of the PID controller is expressed as follows:

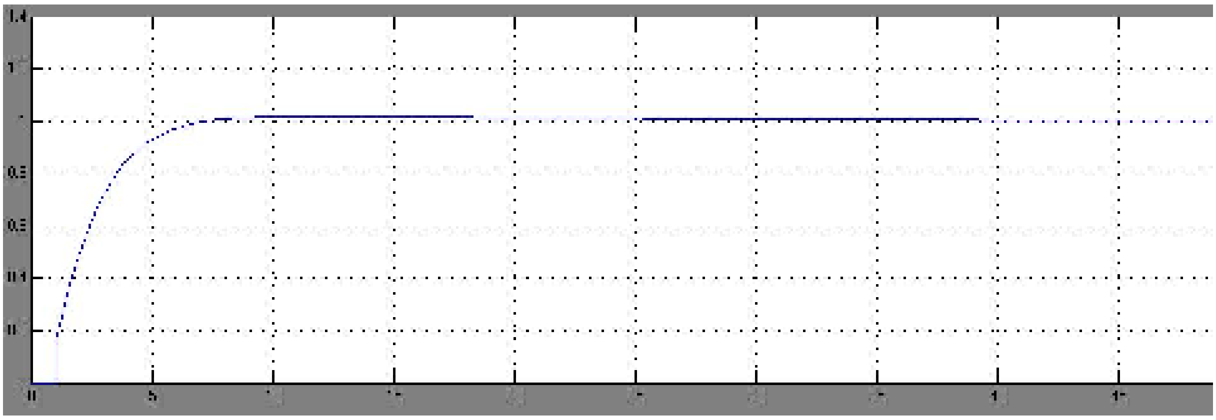
$$H_{PID} = \frac{1 + K_P \tau_i\, S + \tau_i \tau_d\, S^2}{\tau_i\, S}$$

If we consider the same transfer function A and B than in the previous cases, we express the close-loop transfer function as follows:

$$H_{PI} = \frac{A_o\left(1 + \left(\tau + K_p \tau_i\right)S + \tau_i\left(\tau_d + K_P \tau\right)S^2 + \tau_i \tau_d \tau\, S^3\right)}{A_o B_o + \left(1 + K_P A_o B_o\right)\tau_i\, S + \left(\tau + \tau_p + A_o B_o \tau_d\right)\tau_i S^2 + \left(\tau_i \tau_p \tau\right)S^3}$$

If we look at the transfer function, we can see that the integrator adds a pole, increasing the instability of the system, but the action of the derivative is to bridge that by adding a zero in the numerator. So to have a good stability and at the same time good performances, we understand that we have to balance the value of τ_i, τ_d and also K_P.

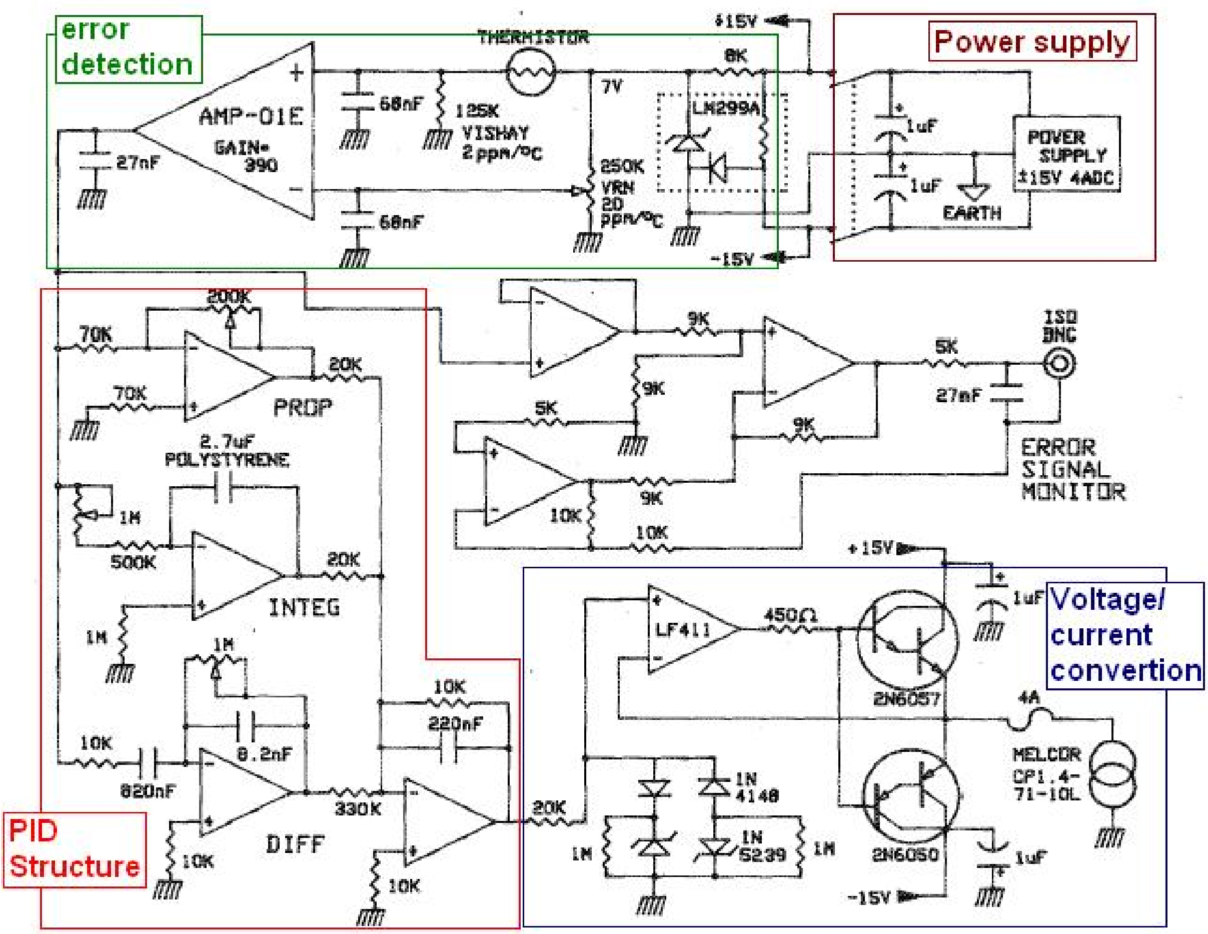
On the following figure, we can see an adapted step response with the proper PID coefficients: The rise time is short and the steady-state error is cancelled:



An electronic circuit based on the PID architecture has been found in the litterature (***ref 18***) allowing to balance the PID parameters to find at the same time the best stability and the as fast response as possible (***ref 20***).

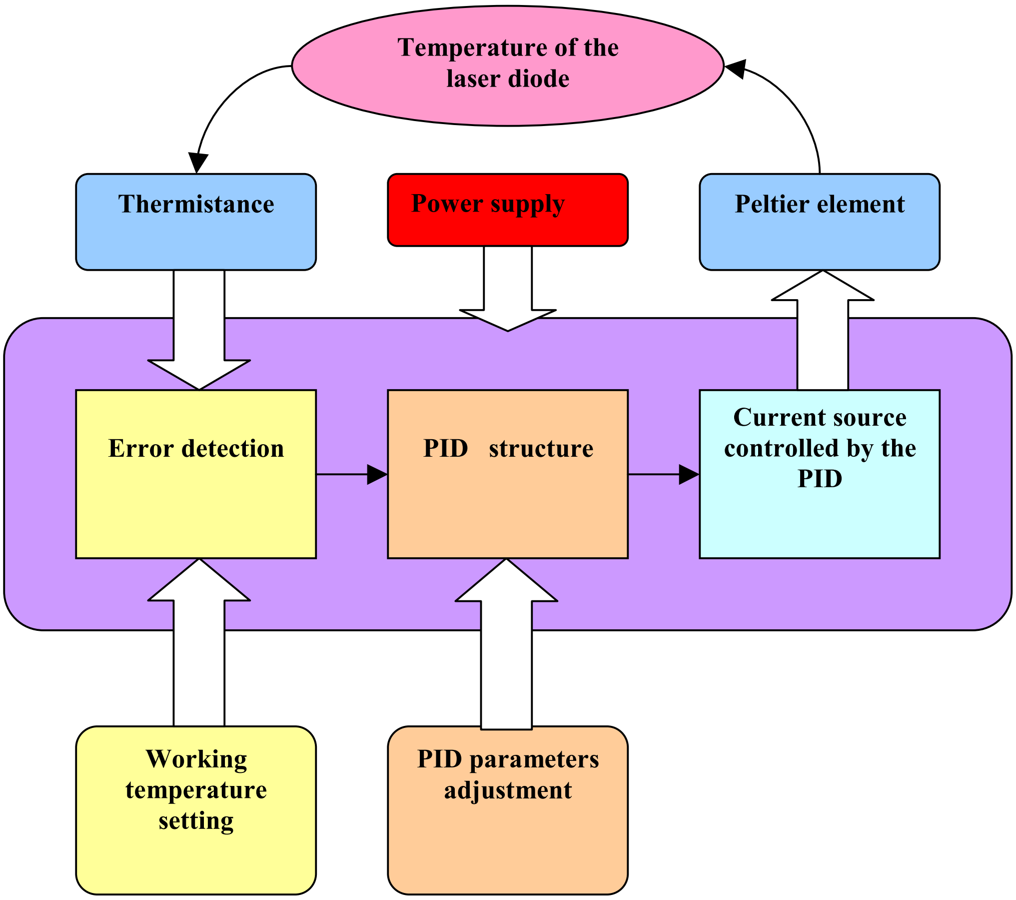
e) Description of the PID electronic circuit

This PID controller architecture was described in 1990 by C. C. Bradley, J. Chen and Randall G. Hulet (***ref 18***) and it is shown in the following figure:



In this part we are going to study how works the circuit by locating the critical points and by expressing the analitical transfer function of the controller bloc.

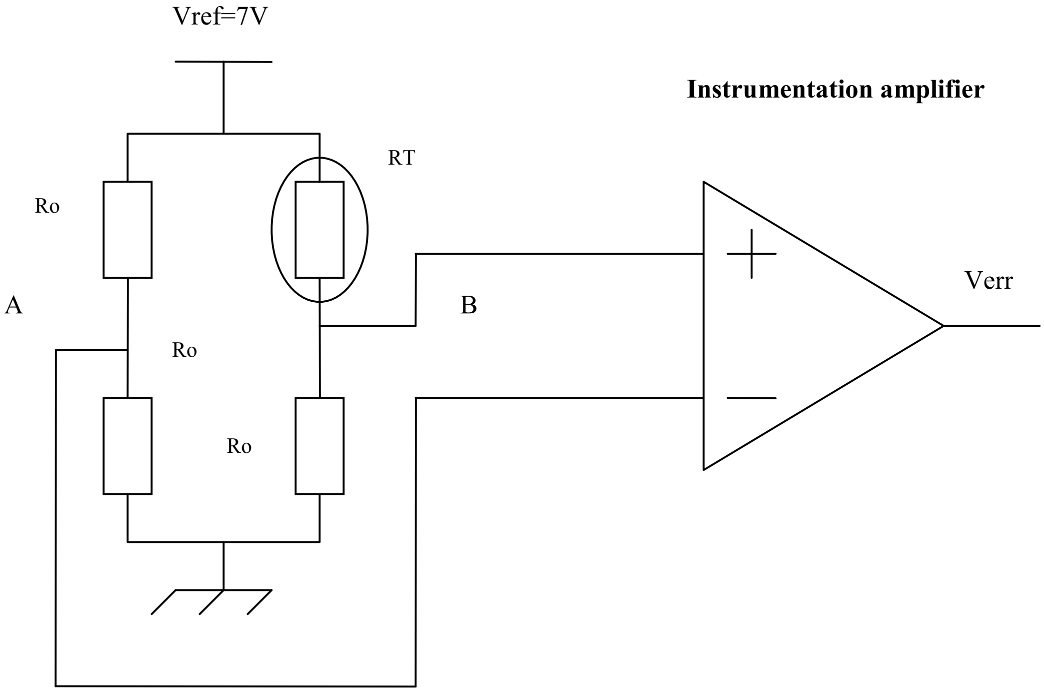
The architecture can be sumerize by the following bloc schematics:



α) Error detection circuit

This part is aimed to detect the difference between the wished temperature and the temperature of the laser diode. For this work, a resistor bridge whose differential outputs are connected to an instrumentation amplifier is particularly good adapted because it cancels the common mode error and at the same time increase the linearity of the value of the thermistance according to the temperature as we are going to see it.

The basic bridge detector is shown in the following figure:



Ro is the reference resistor whose value is the value of the thermistance at 25°C.

R_T is the thermistance. The chosen model is the TH10K what means that the value of the resistance at 25°C (the typical temperature working) is $10\,K\Omega$. The temperature coefficient of the TH10K is -4.4 %/°C that allows us to express the resistance value of the thermistance according to the temperature:

$$R_T = R(25°C) \times (1 - 0.044)^{T-25} \qquad \text{with } R(25°C)=10\,K\Omega$$

As the typical working temperature is 25°C, we choose Ro=R(25°C)=10 $K\Omega$.

So, we have finally this following expression:

$$R_T = R_o \times 0.956^{(T-25)}$$

The expression of the potential A connected to input (-) of the instrumentation amplifier is the following:

$$V_A = \frac{V_{ref}}{2}$$

And the potential B of the bridge connected to the input (+) of the instrumentation amplifier is expressed as follows:

$$V_B = \frac{R_o}{R_o + R_T} V_{ref}$$

The instrumentation amplifier amplifies the difference between the two potentials of the bridge:

$$V_{err} = G(V_B - V_A)$$

This differential configuration allows to eliminate or at least significaly decrease the common mode parasites, if the two tracks A and B are enough near (In fact, to have the same common mode distortions on the two wires, these have to be in the same electromagnetic configuration, in other words the tracks A and B have to be as near as possible). More over, we don't have to care a lot to the coupling effects between the two lines because the signals does not variate very quickly.

By replace V_A and V_B by there respective expression in the previous relation, we optain:

$$V_{err} = \frac{G}{2}\left(\frac{R_o - R_T}{R_o + R_T}\right)V_{ref}$$

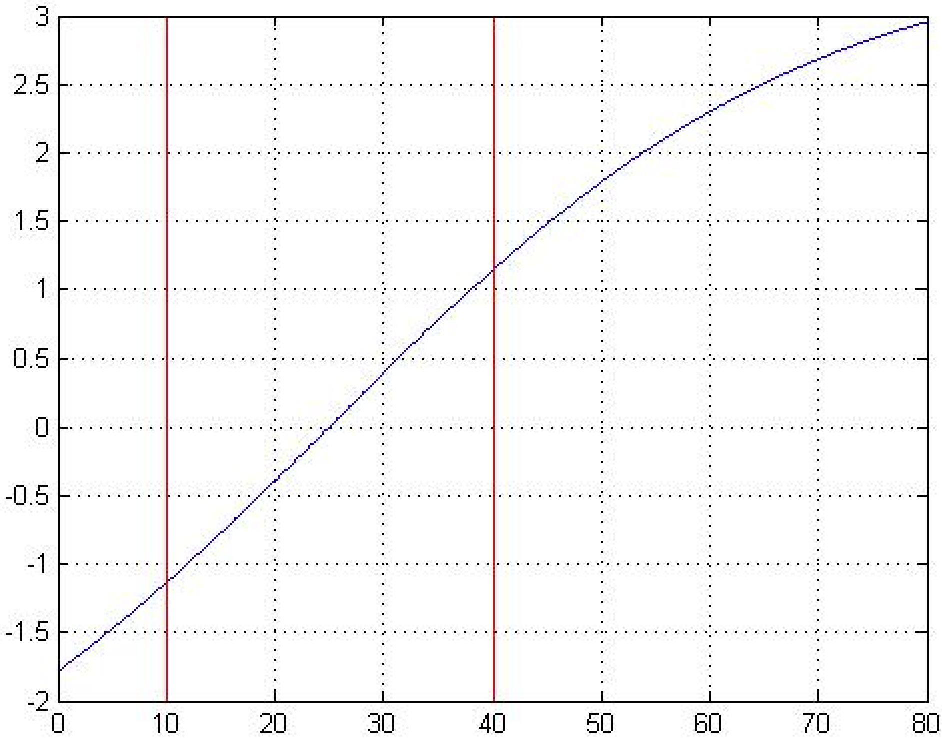
If we look at this expression, we notice that at 25°C, the value of V_{err} is null since the value of the thermistor at 25°C is Ro, in other words when there are no differences between the working temperature (25°C) and the temperature of the diode.

However, the expression is not linear even by replacing R_T by its expression according to the temperature:

$$V_{err} = \frac{G}{2}\left(\frac{1 - 0.956^{(T-25)}}{1 + 0.956^{(T-25)}}\right)V_{ref}$$

Consequently, we have to check on which temperature scale the expression behaves linearly to determin if it is convenient for our application.

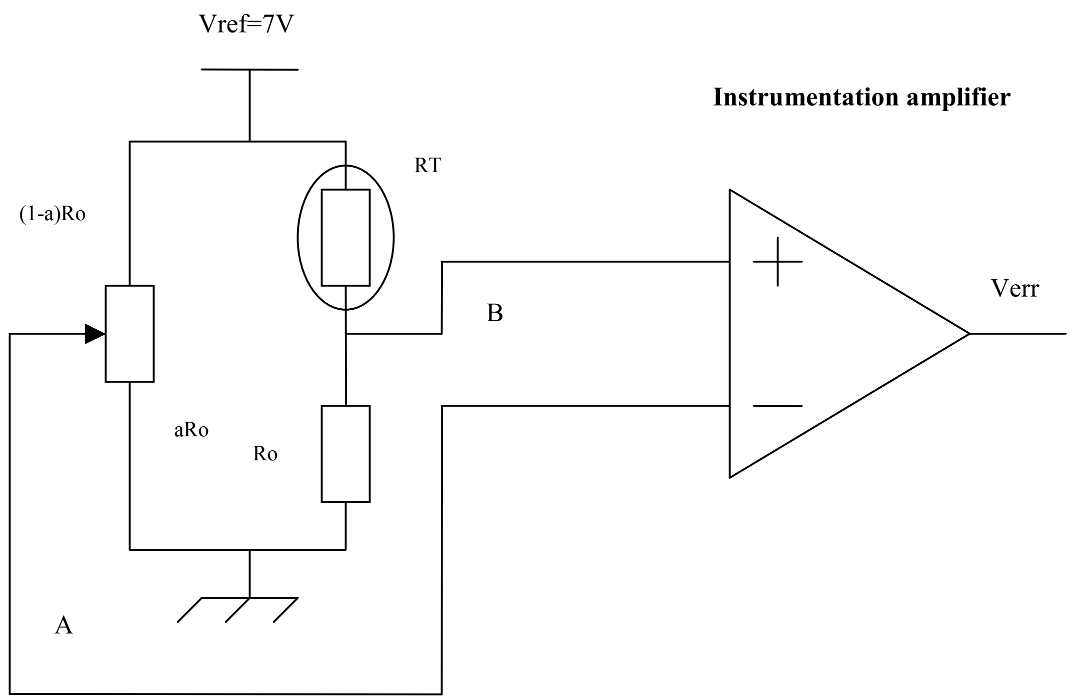
On Matlab we draw the curve of V_{err} according to the variation of the temperature (For V_{ref} =7V):



According to the curve, we can concidere that the expression of V_{err} is linear between 10°C and 40°C, including the working temperature (25°C). This domain of linearity is enough wide for our application, because we need to have an as stable temperature as possible excluding large excursion in the temperature variation. Our circuit is designed to response as quick as possible to temperature variation less than 3mK. Consequently, we can consider the error detection circuit perfectly linear in our use.

This detection structure built like this only allows to work at 25°C, but in our use it would be convenient to change the working temperature around 25°C.

This function is realized by removing the two $10\,K\Omega$ resistances composing the side A of the detector bridge and putting instead one $20\,K\Omega$ potentiometer whose pins 1 and 2 are respectively connected to the voltage reference and the ground and the floating pin w is connected to the input (-) of the instrumentation amplifier as shown in the following figure:



In this configuration, we have between the voltage reference and the point B a resistance of $(1-\alpha)R_{pot} = 2(1-\alpha)R_o$ and between the point B and the ground a resistance of $2\alpha R_o$ (because we choose a potentiometer whose maximal resistance is twice the reference resistance value of 25°C).

Consequently, with the same calculation, the V_{err} voltage is defined by:

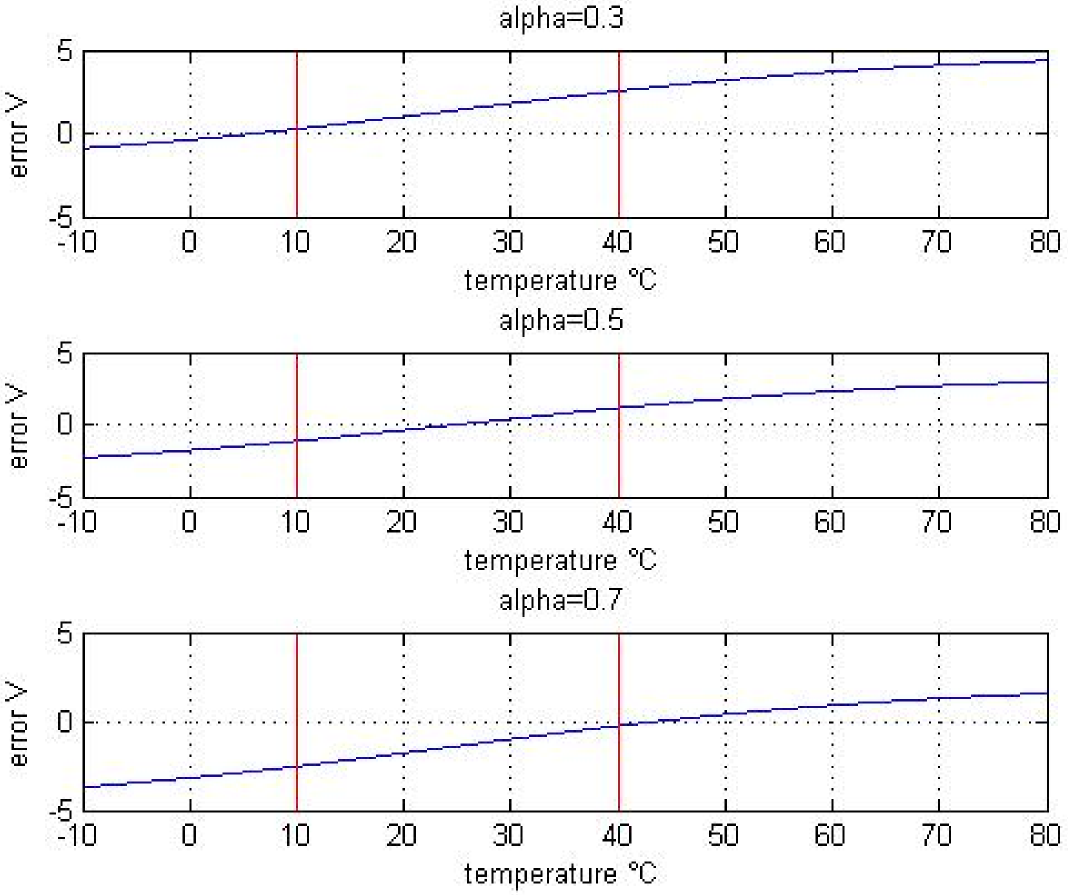
$$V_{err} = G\left(\frac{R_o}{R_o + R_T} - \alpha\right).V_{ref} = G\frac{(1-\alpha).R_o - \alpha.R_T}{R_o + R_T}V_{ref}$$

When the position of the potentiometer is in the middle, i. e. α=0.5, we are in the previous situation where the working temperature is adjusted at 25°C. But when we change α, the previous curve taken a 25°C setting is verticaly shifted:

-positively if α is decreasing

-negativly if α is increasing

(Seeing the following figure)

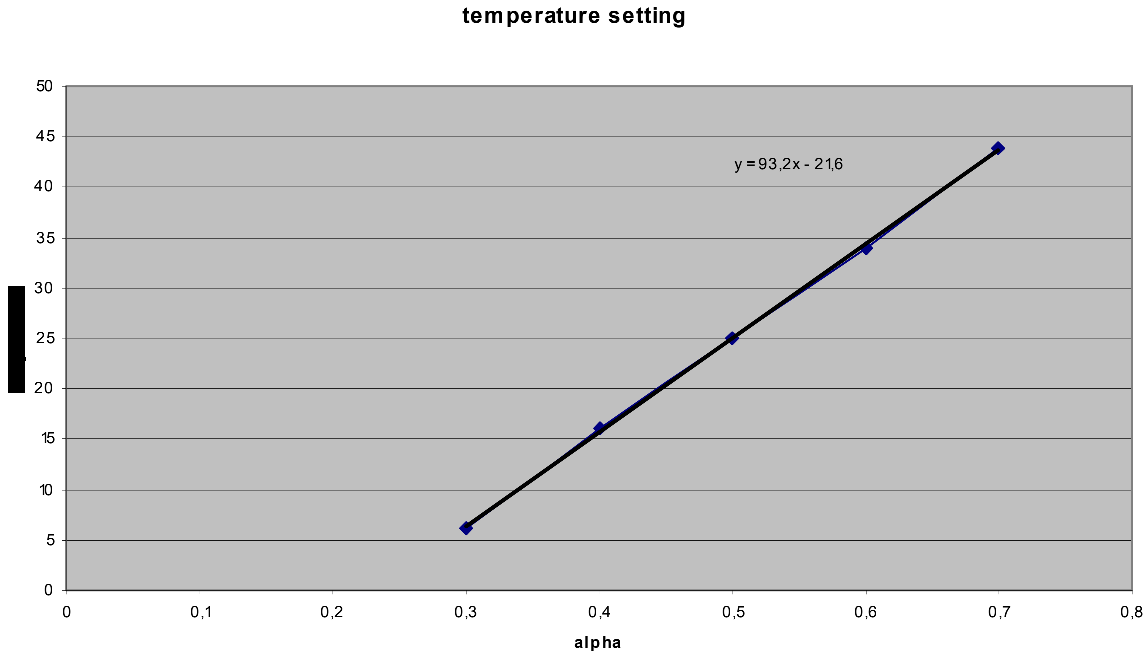


That means that when the curve is shifted down, the value of the temperature for which we have a voltage error null increase, in other words the working temperature and inversely when the curve is shifted up.

So with this architecture, we increase the working temperature by increasing α and we decrease it by decreasing α .

But the problem is that shifting verticaly the curve does not shift the linearity domain. Consequently, the range of temperatures we can select is limited by the linearity domain between 10°C and 40°C but it is not a problem concerning our application because the ambiant laboratory temperature is generally never out this range.

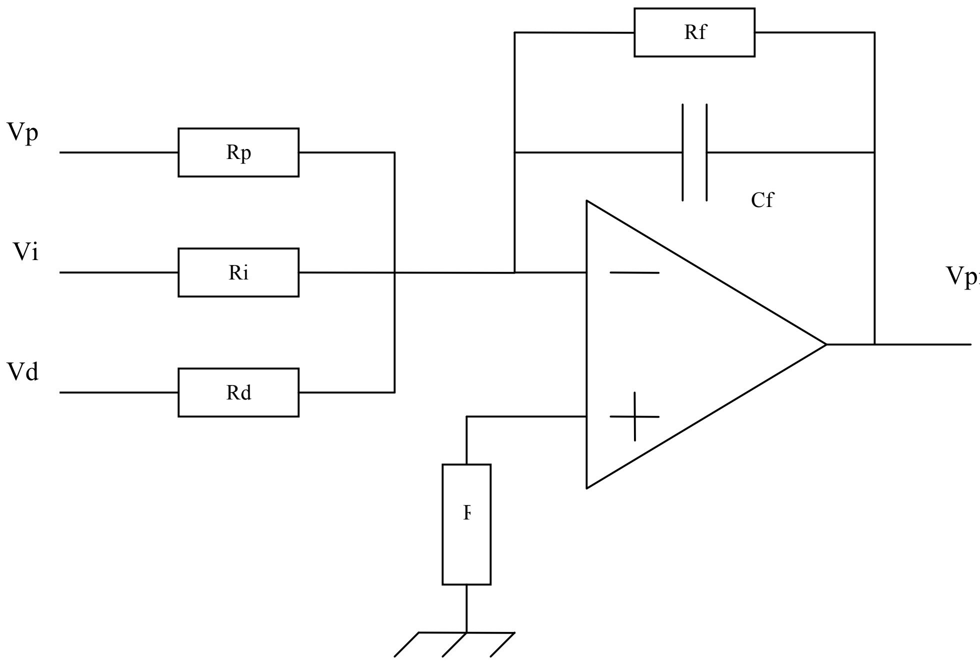
The following curve shows the value of the theoretical working temperature according to the position of the potentiometer (on the domain of linearity):



β) PID structure

The electronic architecture of the PID part is a copy of the theorical one. It is composed by three operational amplifiers realizing respectively the function proportional gain, integrator and derivative, followed by a fourth which add and ponderate the three signal (proportional, integrator and derivative signals).

We are going to express the analytical expression of the parameters (K_P, τ_i and τ_d) of the PID controller according to the value of the components. For that, we begin by expressing the transfer function of the adder (which acts as a low-pass filter).



If we concider the operational amplifier ideal and working in the linear part (because the output is looped in the minus input) we can write that:

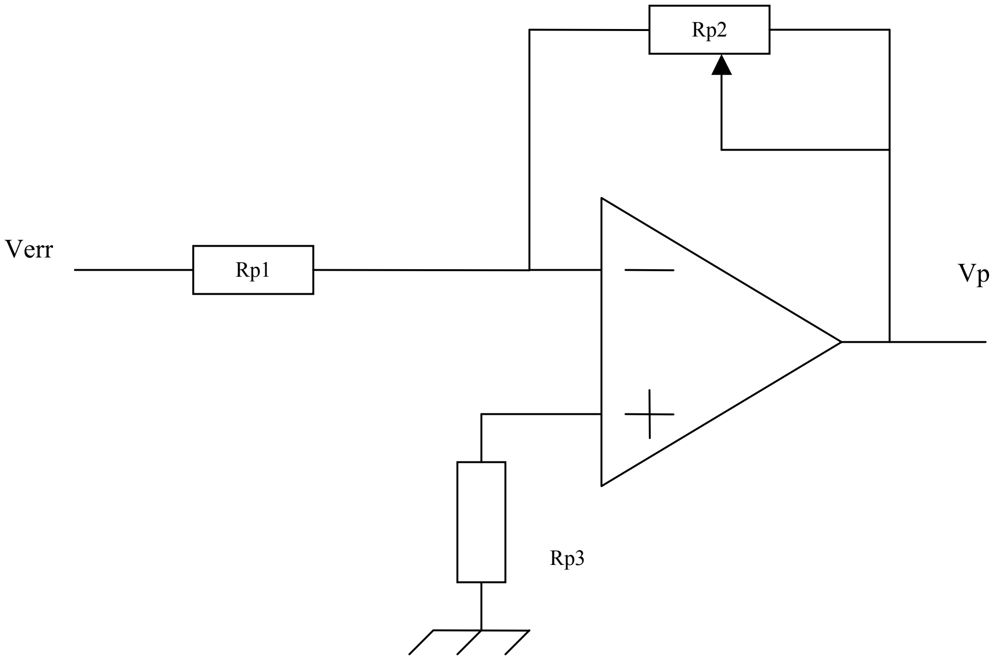
$$V^- = V^+ = 0 = \frac{V_P}{R_P} + \frac{V_i}{R_i} + \frac{V_d}{R_d} + \left(\frac{1}{R_f} + j\omega C_f\right)V_{pid}$$

so $$V_{pid} = -\frac{1}{1 + j\omega R_f C_f}\left(\frac{R_f}{R_P}V_P + \frac{R_f}{R_i}V_i + \frac{R_f}{R_d}V_d\right)$$

So this part of the PID controller adds and inverts the three signals coming from the proportional gain, the integrator and the derivative with ponderation coefficients $\frac{R_f}{R_P}$, $\frac{R_f}{R_i}$ and $\frac{R_f}{R_d}$ but also acts like a low-pass filter to eliminate high frequencies parasites.

Proportional stage

The function of the proportional gain is realized by the following schematics:



The proportional gain is given by an inverter amplifier whose transfer function is the following:

$$V_P = -\frac{R_{P2}}{R_{P1}} V_{err}$$

where R_{P1} is a fixed resistance and R_{P2} a variable one.

By multiplying the last expression by the ponderation coefficient associated to the proportional gain and the gain of the instrumentation amplifier, we get the expression of the proportional parameter K_P:

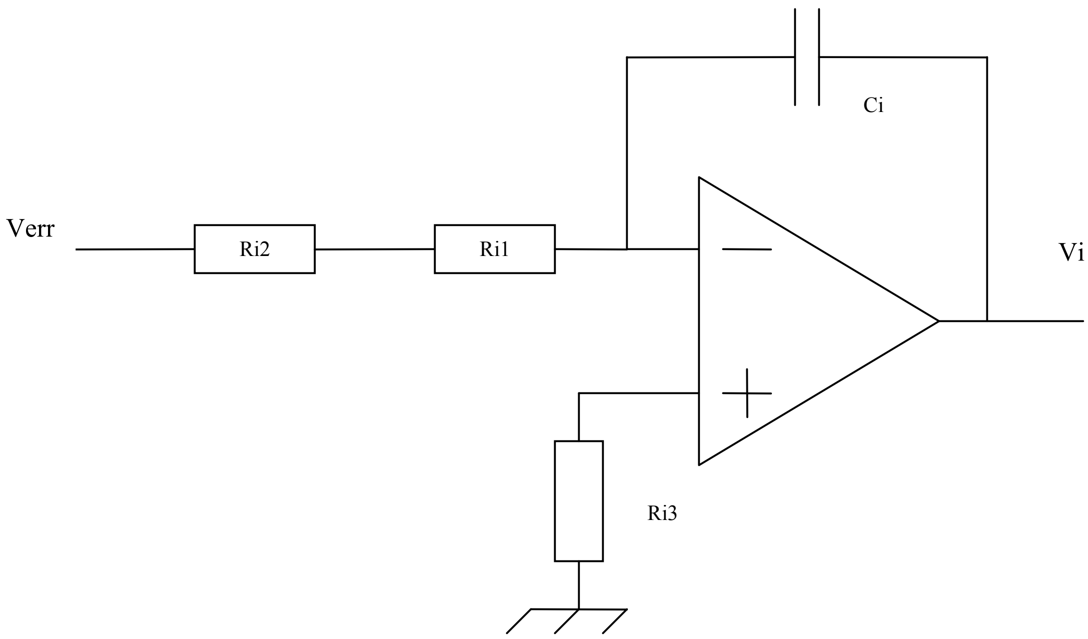
$$K_P = R_{P2} \times \frac{R_f}{R_{P1} R_P}$$

So in this architecture, the value of the proportional gain is adjustable by the variable resistance R_{P2}. With the following value of the componants:

$$G = 400,\ R_{P1} = 70K\Omega,\ R_{P2} = 200K\Omega,\ R_P = 20K\Omega \text{ and } R_f = 10K\Omega$$

The proportional gain is ranged from 0 to 571.

Integrator stage



The transfer function of the integrator is given by the following expression:

$$V_i = -\frac{1}{j\omega(R_{i1}+R_{i2})C_i}$$

The time constant of this integrator is given by $(R_{i1}+R_{i2})C_i$, but to have the complete integrator parameter, we have to divide this previous expression by the gain of the instrumentation amplifier and the ponderation coefficient of the integrator determined by the adder/filter upstream.

So, the complete integrator time constant is expressed as follows:

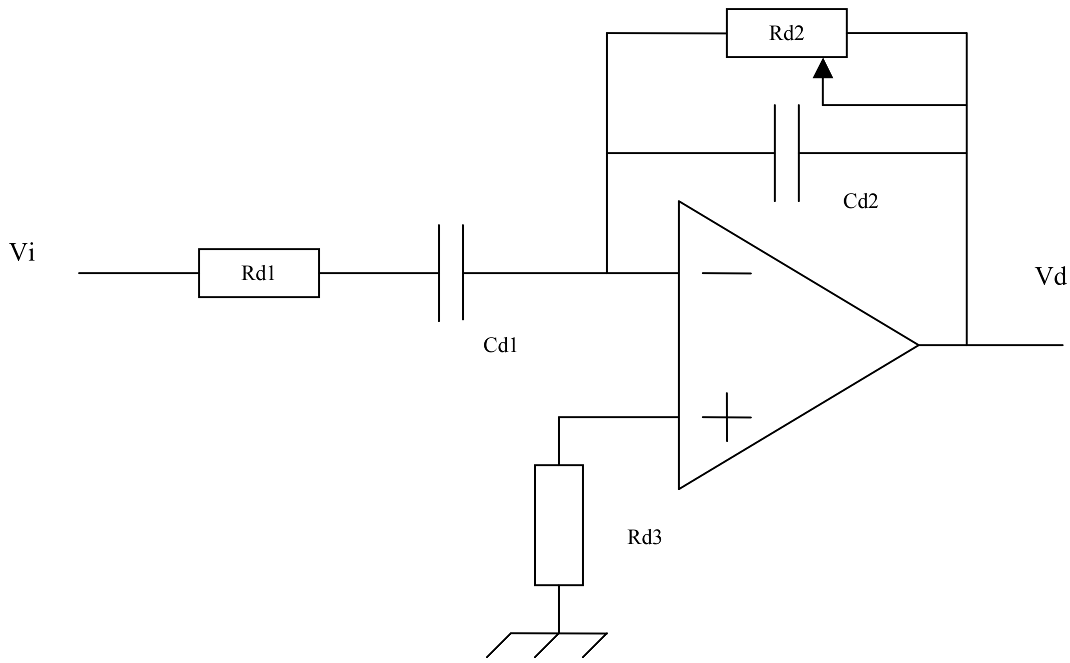
$$\tau_i = \frac{R_i}{G\,R_f}(R_{i1}+R_{i2})C_i$$

The integrator time constant is adjustable by the variable resistance R_{i2}. with the following value of the componants:

$$G = 400,\ R_{i2} = 1M\Omega,\ R_i = 20K\Omega,\ R_f = 10K\Omega \text{ and } C_i = 2.7\mu F$$

the integrator time constant is ranged from 6.75ms to 20.25ms.

Derivative stage



The transfer function of the derivative structure is given by the following expression:

$$V_d = -\frac{j\omega R_{d2} C_{d1}}{1 + \left(\frac{R_{d1}}{R_{d2}} + \frac{C_{d2}}{C_{d1}}\right) j\omega R_{d2} C_{d1} + (j\omega)^2 R_{d1} R_{d2} C_{d1} C_{d2}}$$

In high-frequency, this function acts like a filter, but in low variations of the signal error, it behaves like a derivative structure whose expression is: $j\omega R_{d2} C_{d1}$.

The value of the derivative time constant is obtained by multiplying the time constant of this previous bloc by the gain of the instrumentation amplifier G and the and the coefficient $\frac{R_f}{R_d}$ associated to the derivation.

$$\tau_d = G \frac{R_f}{R_d} R_{d2} C_{d1}$$

The integrator time constant is adjustable by the variable resistance R_{d2} and with the following value of the components:

$$G = 400,\ R_{d1} = 10K\Omega,\ R_{d2} = 1M\Omega,\ R_d = 330K\Omega,\ R_f = 10K\Omega,\ C_{d1} = 820nF \text{ and } C_{d2} = 8.2nF$$

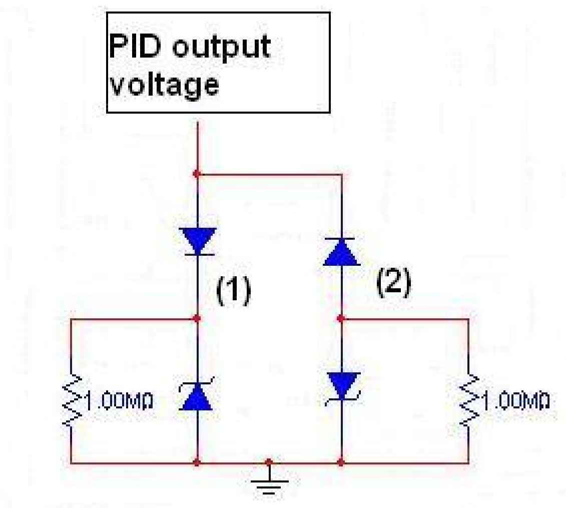
the derivative time constant is ranged from 0 to 9.94ms.

χ) The conversion voltage/current

The last part of the schematics to provide a driving current to the peltier element is exactly based on the same principle than in the end of the structure of the current source built previously. Indeed, the circuit produces a current that is proportional to the output voltage of the proportional gain by fixing this potential at an output resistor thanks to a loop. Inside the loop, we have one follower whose inverting input is used to close the unit gain loop and non inverting input connected to the output of the PID voltage control, and just after a current amplifier composed by an operational amplifier or a darlington structure (like in this case) in order to allow having a big output current without dissipating a big quantity of heat. A protection resistor can be positionned between the follower and the current source (or just after the current source because the voltage gain of the darlington structure is 1) to limit the voltage at the edges of the peltier element.

Like in the case of the current source, the gain of the current amplifier bridge the defaults of all components enclosed in the loop. Especially in the case of a darlington structure where the transistor work in class B: When the input voltage is superior than V_{BE} (the voltage between the base and the emetor of the transistor), the PNP is locked whereas the NPN works and it is the opposit case when the input current is less bigger than -V_{BE}. So because of the threshold voltage V_{BE} of the two transistor the output signal is distorded if it often go up and down between positive and negative. But with the use of the follower, the effect of V_{BE} is divided by the high gain of the amplifier. This architecture is not sensitive to the default of the components.

To avoid current overflow in the Peltier element, a protection circuit is added in parallele at the input of the voltage/current converter. This circuit avoid to have a driven current out of the specification range of the Peltier element. It is composed of one diode and one zener diode in parallele with two reversed other one as shown in the following figure:



The maximal current supported by the Peltier element is 5.4A, but we don't need to reach the limit current. It is sufficient to fixe the current range flowing throw the Peltier element from -4 to 4A in order to avoid to stress a lot the device. In the protection circuit, we use the zener diode BZX55-C12 whose reverse voltage is 12V. Hense, if in the input of the voltage/current converter, the voltage is positive, the diode in the branch (1) (on the previous figure), is driving wherease the diode in the branch (2) is locked and the zener diode avoid having a voltage superior to 12 V. In the opposit case, the diode of the branch (2) is driving and the one in the branch (1) is locked and the zener diode prevent against a voltage inferior to -12V. Consequently, the protection circuit limit the output PID voltage range to $\pm 12V$. If we want a maximal output current flowing in the TEC of 4A, the output resistance should be equal to 3Ω .

δ) The displaying circuits

In our circuits, two values are important to be displayed:

-(1) the signal error which indicates by a differential voltage how much is the difference between the setled temperature and the real one.

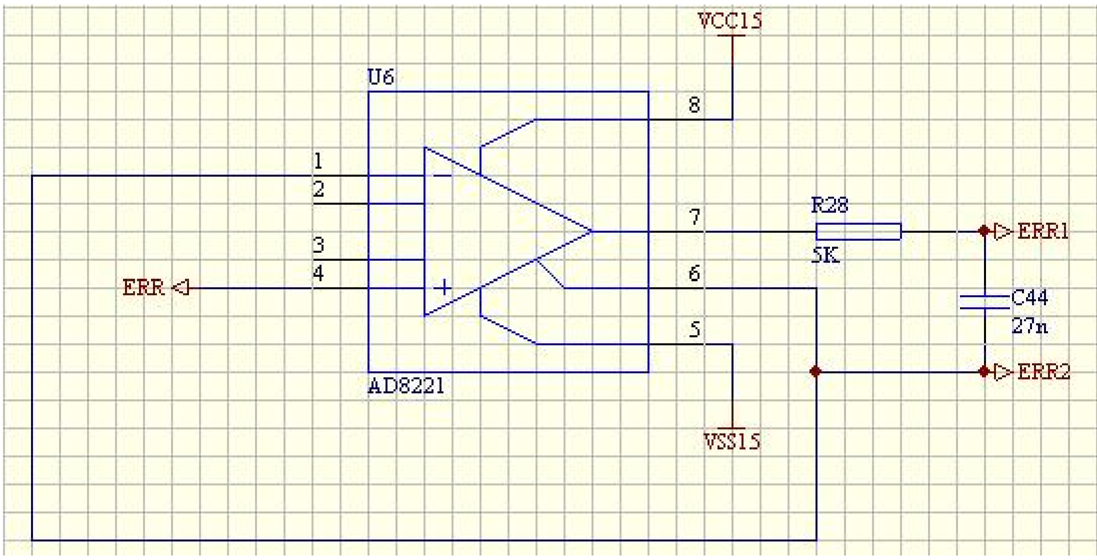
-(2) the settled working temperature adjustable by the potentiometer of the bridge detector.

For the first point, the instrumentation amplifier AD8221, used for the detection circuit, is convenient for this work, because one of its pin can be use as a reference. With this additional pin, the function of the instrumentation amplifier is expressed as follows:

$$V_{out} = G \times \left(V^{+} - V^{-} + V_{ref}\right)$$

Thanks to this reference pin, we can create a floating ground by connecting the V^{-} to the reference pin and the V^{+} to the error voltage at the output of the detector circuit. In this way the voltage difference between the output of the instrumentation amplifier and the reference pin corresponds to the error voltage, and this signal is perfectly isolated from the ground of the circuit avoiding to have distortions in the work of the PID controller. To attenuate eventual high-frequency parasites, a first order low-pass filter is added at the out put of the instrumentation amplifer and the error signal is gotten back at the filter capacitor to be displayed.

The following figure shows the circuit displaying the error signal:



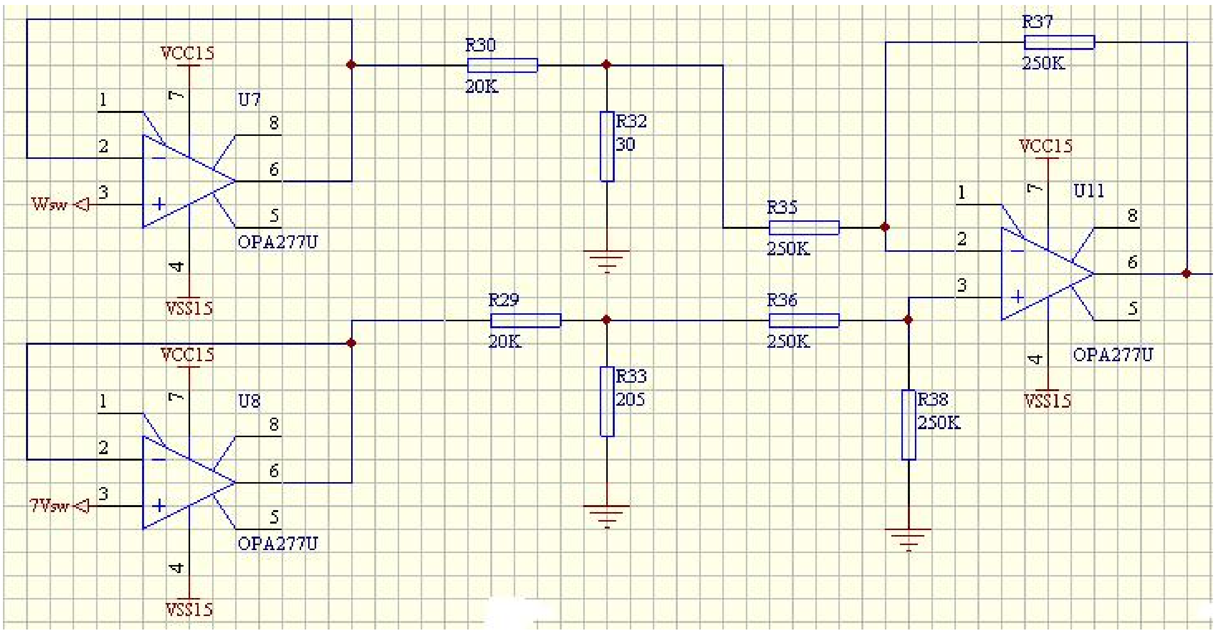
For the second point, we want to display the value of the working temperature selected by the potentiometer. We saw previously that the settled temperature is proportional to the position of the floating pin of the potentiometer only from 30% to 70%. It corresponds to a temperature setting ranged from 5 to 45°C. In the linear area, the convertion between the voltage at the floating pin of the potentiometer and the temperature is given by:

$$T = -13.3 \times V_{set} + 71 = -13.3 \times V_{set} + 10.2 \times V_{ref}$$

To display the temperature on a voltage displayer, we need to work in milivoltage ranged otherwise we are out of the linearity work of the used operational amplifiers. That means that we have to divide by 1000 the previous equation. To realize this function, we use two opperational amplifiers as two follower, one connected to the voltage reference and the other connected to the potentiometer floating pin to islolate this displayer circuit to the controller circuit. For a good isolation, the operational amplifiers used as follower must have very low bias and offset current. One voltage divider follows each of these two operational amplifiers to fixe the coefficients of the previous equation: 0.0133 for the setting voltage and 0.0102 for the reference voltage. The a substractor realize the difference between the two branches to finally have the output voltage to display in milivolts:

$$T = -0.0133 \times V_{set} + 0.0102 \times 6.95$$

The schematics of the displaying circuit is shown as follow:



The risk with this circuit is that it could add drift in the inputs of the instrumentation amplifier of the detector part, even if the bias current of the operational amplifier used as follower is low. So we added a switch between the controller and the displayer circuit allowing to disconnect the displaying setting temperature from the detector in order to avoid to decrease the performence of the error detector.

ε) The critical points of the temperature controller

The critical point does not come from the PID structure itself, because, as we saw, it is enclosed in a loop by about 400-gain instrumentation amplifier. Consequently, all undesired drift effects or any offset voltages are devided by 400 and thus components inside the loop do not need to be really accurate.

The true critical point is the detection part of the circuit because the response of the controller depends on the quality of the detection. Any offset or drift in the detection inputs disturbs the error signal and thus the work of the circuit. We have to discuss about the characteristics that should have detector's components to get acceptable detection quality.

As we saw ealier, the voltage difference between the two inputs of the instrumentation amplifier is given by:

$$\Delta V_{inst} = \left(\frac{R_o - R_T}{R_o + R_T} \right) V_{ref}$$

and as $R_T = R_o \times C_T^{(T-25)}$ where C_T is the temperature coefficient of the thermistance (C_T =0.044 for the TH10K)

$$\Delta V_{inst} = \left(\frac{1-(1-C_T)^{(T-25)}}{1+(1-C_T)^{(T-25)}} \right) V_{ref}$$

If we consider a small variation of the temperature, we can write that:

$$\frac{\delta}{\delta T}(\Delta V_{inst}) = \left(\frac{-2 \times \ln(1-C_T) \times (1-C_T)^{(T-25)}}{\left(1+(1-C_T)^{(T-25)}\right)^2} \right) V_{ref}$$

Since C_T <1, we can write that, on one hand $\ln(1-C_T) \approx C_T$ and on the other hand, $(1-C_T)^{(T-25)} \approx 1$ for very small variations of the tempetature. This is justified by the fact that we expect that our circuit detects temperature variations as small as 0.1mK.

$$\delta(\Delta V_{inst}) = -\left(\frac{C_T}{2} \right) V_{ref} \, \delta T$$

Consequently, a variation of the temperature 0.1 mK correspond to a variation of the voltage between the two input of the instrumentation amplifier of:

$$\delta(\Delta V_{inst}) = \left(\frac{0.044 \times (0.956)^{(0.0001)}}{2} \right) \times 6.95 \times 0.0001 = 15.3 \mu V = \delta(\Delta V_{inst})_{\min}$$

Hense, if we want the detector be able to detect temperature variations as small as 0.1mK, the addition of all sources of voltage drift have to be less than 15.3μV.

The voltage drifts are due to the bridge resistances sensitivity over temperature, to the voltage reference drift and to the instrumentation amplifier input characteristics (offset and bias input current, offset voltage, drift over temperature).

characteristics of the resistances of the detection bridge

The temperature coefficient of the resistances must be as small as possible. The resistance variation over temperature is expressed as the same way than that of the thermistor (but the temperature coefficient of the thermistor is of course quite bigger):

$$R_{bridge} = R_o \times (1 - C_R)^{\Delta T}$$

Consequently, the voltage variation between the two inputs of the instrumentation amplifier due to the drift of resistors of the bridge has a similar expression than previously:

$$\delta(\Delta V_{inst})_R = \left(\frac{C_R}{2}\right) V_{ref}\ \delta T_R$$

As we said before, we need to have $\delta(\Delta V_{inst})_R << \delta(\Delta V_{inst})_{\min}$, in other words:

$$C_R \delta T_R << C_T \delta T_{\min}$$

If the variation of the temperature at the resistances is in the same range than the temperature at the thermistance, we must have in any case $C_R << C_T$. The temperature coefficient of the thermistor TH10K is -4.4%/°C, i. e. -440 $\Omega / °C$. In the scientific review, it is used resistances of 2 ppm/°C, what is enough to perform good but the article is 17 years old and now it is possible to find resistance whose temperature coefficient is 0.2 ppm/°C in order to have a better detection signal.

choice of the voltage reference

As we saw ealier, in the formula of the detectors' sensitivity over temperature, the output voltage is proportional to the voltage reference. So any variation in the voltage reference over temperature disturbs the work of the detector. The variation of the voltage reference must be smaller than the minimal voltage difference between the two inputs of the instrumentation amplifier.

If we use the same voltage reference than in the case of the current source (the LM399H), the typical temperature coefficient is 0.00003%/°C and the typical value of the voltage reference is 6.95V. So for a temperature variation at the voltage reference of 10°C, the reference drift is 20.85 μV and it produce a drift of:

$$\delta(\Delta V_{inst}) = -\left(\frac{0.044}{2}\right) \times 20.85 \times 10^{-6} \times 10 = 4.6 \mu V < \delta(\Delta V_{inst})_{\min}$$

Thus, the LM399H, has acceptable performances for our application.

considerations concerning the instrumentation amplifier

As small differences must be detected in the inputs of the instrumentation amplifier, the common mode parasites must be negligeable compared to the useful differential mode. In other words, the common mode rejection must be as big as possible or at least bigger than:

$$-20.\log\left(\frac{\delta(\Delta V_{inst})_{\min}}{V_{com}}\right) = -20.\log\left(\frac{\delta(\Delta V_{inst})_{\min}}{\frac{V^{+}}{2}}\right) = -20.\log\left(\frac{C_T \delta T_{\min}}{2}\right) = 113.3dB$$

For minimal variation of the temperature, the common mode rejection of the instrumentation amplifier must be bigger than 113.2 dB.

The bias and the offset input currents are also sources of voltage drift. So these input characteristics must have negligeable effects on the offset voltage over temperature in order to be able to detect very small temperature variations.

The offset voltage generated by the bias current of the IC is expressed as follow:

$$\Delta V_{bias} = I_B\left(R_o + R_T - R_P\right)\frac{\max\left(R_o, R_T\right)}{R_o + R_T}$$

Where R_P is the maximal resistance of the potentiometer (20 $K\Omega$)

As we can see with this formula, given that at 25°C, the value of the thermistance is $10\,K\Omega$, the resistance R_o is 10 $K\Omega$ and the potentiometer as a maximal resistance of 20 $K\Omega$, the offset due to the bias current is null. But if we have a difference of temperature, the value of the thermistance changes and the offset increases. What is the reason why the bias current must be as small as possible.

The offset voltage due to the offset input current is expressed as follow:

$$\Delta V_{Ioff} = \frac{\partial I_{off}}{\partial T}\delta T \max\left(R, R_T\right)$$

This formula does not only show that offset current have to be small but the sensitivity of it must also be tiny.

Finally, the input voltage drift must be negligeable compared to the minimal voltage difference between the two inputs equal to $15.3\mu V$. So the temperature coefficient must be smaller than this previous value divided by ten, in other words: $1.5\mu V/°C$.

For our detection circuit, we have chosen the AD8221BR whose critical characteristics are the followings:

-Common mode rejection: between $130\,dB$ and $140\,dB$ for a gain of 400.
-Max input voltage drift: $0.3\mu V/°C$
-Typical input offset current drift: $1pA/°C$
-Max input bias current: $0.4\,nA$

As we can see, the common mode rejection and the input voltage drift are excellent for our application.

Concerning the input offset current drift, if we have a temperature's variation of about 1°C, the value of the thermistance decreases of $440\,\Omega$ and by applying the formula of the voltage drift generated by the bias current, we have:

$$\Delta V_{bias} = -0.4 \times 10^{-9} \times 440 \frac{10000}{10000 + 9560} = 0.09\mu V$$

So the drift produced by the bias current can be considered negligeable

Concerning the effect of the offset current, if we considere a temperature's variation of 1°C, we have:

$$\Delta V_{Ioff} = 1 \times 10^{-12} \times 1 \times 10000 = 0.01\mu V$$

Consequently, we can consider that the voltage offset produced by the bias current and the sensitivity of the offset current over temperature is negligeable in comparison with the minimal voltage difference between the two inputs.

To increase the quality of the error-signal, two capacitor of 68nF are placed between the two inputs of the instrumentation amplifier and the ground to eliminate high-frequency parasites.

For the same reason, another capacitor is added at the output of the instrumentation amplifier to be combinated with the internal resistance of the device in order to have a low-pass filter.

Finally, if the previous condition are respected, the sensitivity of our error-detector (in the output of the instrumentation amplifier) is:

$$\frac{\Delta V_{out}}{\Delta T} = \left(\frac{0.044}{2}\right) \times 6.95 \times 400 = 61.2\,mV/mK$$

<u>considerations of the others points of the circuit</u>

The others parts of the circuit are not critical because there are enclosed in a loop started by the instrumentation amplifier (whose gain is 400). So, all bad effects are devided by

400. However, for the PID structure, we will use precision OPA277 operational amplifier for its low noise.

f) Adjustment method of the PID parameters

When the temperature controller is built, it is necessary to set the parameters of the PID structure in order to have an as fast and as stable response as possible. For this, a simple methode, known as *Ziegler Nichols closed loop method* exists. This method consists in adjusting at first the proportional gain by cancelling the integrator and derivative ones. We start to adjust the proportional gain from the begining (the lowest value). Normaly, the response of the system is slow and there are now oscillation in the transient. Step by step the rule consists in increasing by factor two the value of the proportional gain until optening continuous oscillation. The value K_u of the proportional gain for which we have continuous oscillation and the period P_u of the optained oscillations are the values which determin the others parameters τ_i and τ_d :

$$K_p = \frac{K_u}{1.7}$$
$$\tau_i = \frac{P_u}{2}$$
$$\tau_d = \frac{P_u}{8}$$

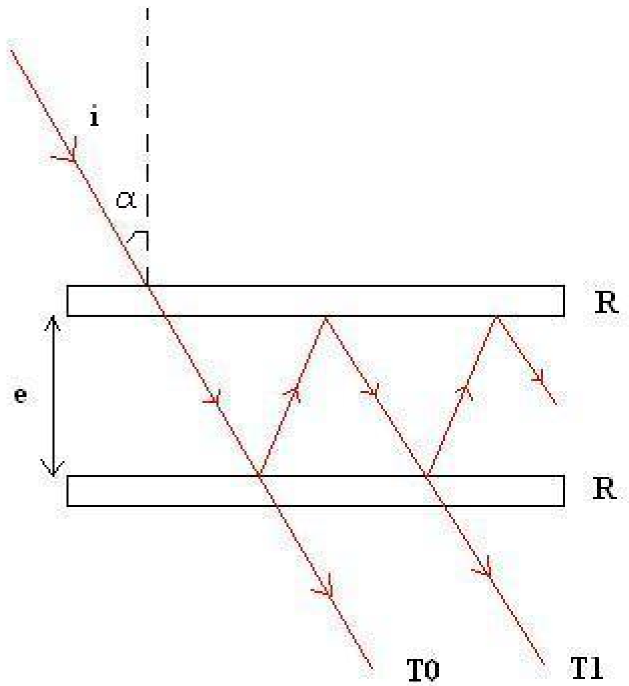
In practice, it is hasardous to reach the continuous oscillation because it is the limit between the stable state and the unstable one. It is better to increase the proportional gain only until a point where some oscillations has been optained. Then it is possible to approach the value of K_u and P_u or at least be near this setpoint.

V : The Fabry Perot interferometer

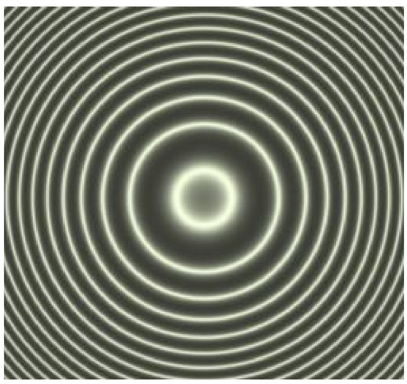
1) Principe

The aim of the Fabry Perot interferometer is to increase the visibility of a spectrum (it’s a little bit like a zoom onto the spectrum) to make it more easy to analyse on an oscilloscope. It’s a way to separate closed wavelengths.

It consists in a middle of index n and width e between two plane or concave mirrors with a high refractive index R.



The light goes from a surface to the other a lot of time and a transmitted light is produced at each time. All the transmitted lights are coherent because coming from the same input light but the phase is different because the optical paths have different values. It results interferences at the output side, if we place a lens at the output and look with our eyes we see rings system.



T

he output amplitude is about $A_T = t^2 [1 + r^2 e^{-j\varphi} + r^4 e^{-2j\varphi} \ldots\ldots\ldots\ldots]$

The output transmittance as a function of wavelength is :

$$T = \frac{1}{1 + \frac{4R}{(1-R)^2} \times \sin^2(\frac{\delta}{2})}$$

With $\delta = \frac{2 \times \pi}{\lambda} \times n \times e$ the optical path difference between two consecutive output beams.

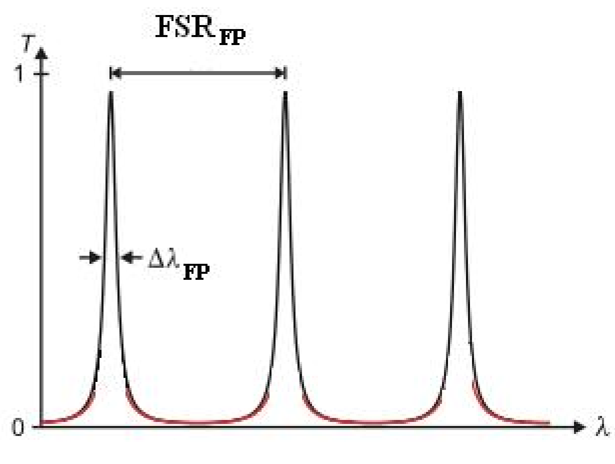
Because the F P Interferometer is a cavity only some frequencies will be transferred, those whose the value is :

$\nu_m = \frac{m \times c}{2 \times n \times e}$ with m an integer, n the index of the cavity (1 for our device) and e the space between the two mirrors.

So the interferometer free spectral range is : $\mathrm{FSR}_{\mathrm{FP}} = \frac{c}{2 \times n \times e}$

If we smoothly change the distance e we can scan wavelength and some peaks will appear at the wavelength that the input light contain.

Here is the theoretical curve we can obtain :



The intensity of the output beam can be express by :

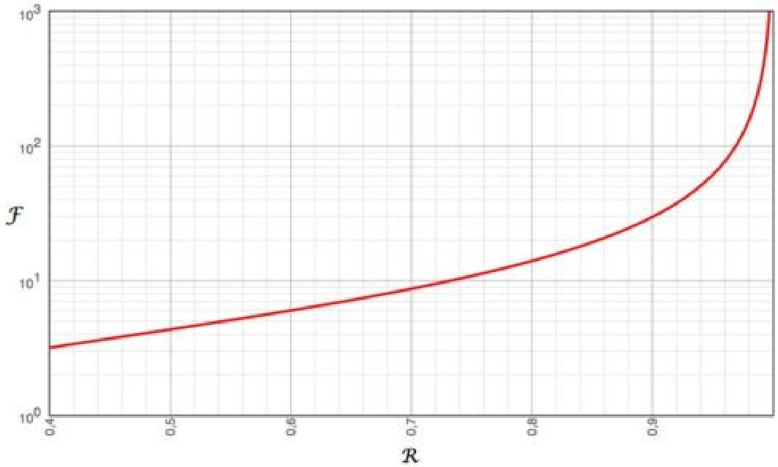
$$I = I_{max} / (1 + F \sin^2\varphi/2) \text{ with } \varphi = 2\pi\, \delta / \lambda$$

For $\varphi = 2\,k\,\pi$ which correspond to $\lambda = \delta / k = (2\,e\,\sin\alpha) / k$ we have $I = I_{max}$.

Dl_{FP} represents the line width of the interferometer at the half power. If the linewidth of the analysed beam is superior to Dl_{FP} it represents the linewidth of the analysed system (Dl_{ECDL} in our case) otherwise it represents the linewidth of the Fabry Perot interferometer.

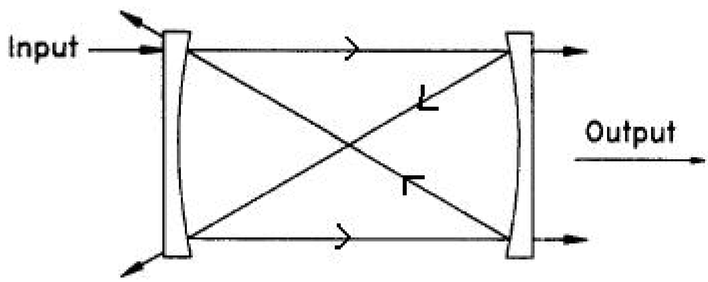
$$\Delta\lambda_{FP} = \frac{c \times (1-R)}{2 \times n \times e \times \pi\sqrt{R}}$$

The finesse of the interferometer is $F = FSR_{FP} / \mathrm{Dl}_{FP}$ and $F = \dfrac{\pi \times \sqrt{R}}{(1-R)}$



The Finesse must be the most high than possible to have the best signal quality. So the aim is to have a very high refractive index R (near 1) but less than one to have enough output power.

Another structure exists, it consists in two spherical mirrors which allow an input angle a = 0°. The optical path is different as we can see below :



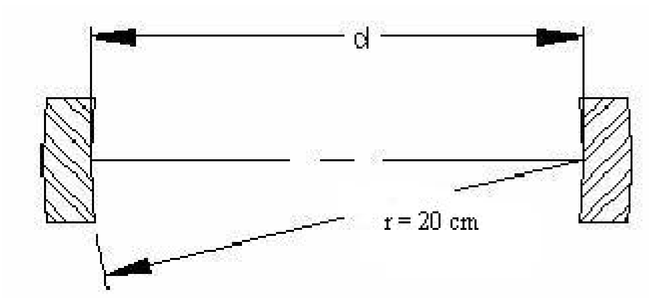
In this configuration we have :

$$\text{FSR}_{\text{FP}} = \frac{c}{4 \times n \times e}$$

$$\Delta\lambda_{FP} = \frac{c \times (1 - R)}{4 \times n \times e \times \pi\sqrt{R}}$$

2) Design

Our Fabry Perot will be composed by two concave mirrors. The FSR in this configuration is c/ 4.d with n = 1 (air). The distance between the two mirrors e is equal to their curvature radius. We choose a long distance of 20 cm to have good performance (narrow line width and wide FSR which imply a good resolution).



e = 20 cm so we can deduce FSR_{FP} = 750 MHz.

Dl_{FP} limits the linewidth we can observe, with ECDL application we can obtain a linewidth less than 1Mhz. So the ideal case would be a Fabry Perot with $Dl_{FP} < 1$ MHz to be sure to well determine the ECDL linewidth. But such a narrow linewidth imply to have high reflectance mirror whose price is very expensive. So we will work with a line width of about 1Mhz.

Here is a table which depict the line width for different reflectance R :

line width (MHz)	Reflectance
22.2	0.8
10.5	0.9
5	0.95
1	0.99

The refractive index is chosen to be about 0.99, we can deduce F = 312.6 and Dl_{FP} = 1MHz.

So the distance between the two mirrors must be around $e = \dfrac{c}{2 \times FSR_{FP}}$ = 0.2 m.

To scan the wavelength we fix a PZT element on the back of one of the two mirrors which is connected to an alimentation providing a tooth saw signal. A photodiode is placed in the path of the output light. The intensity of the diode (in Ampere) is directly proportional to

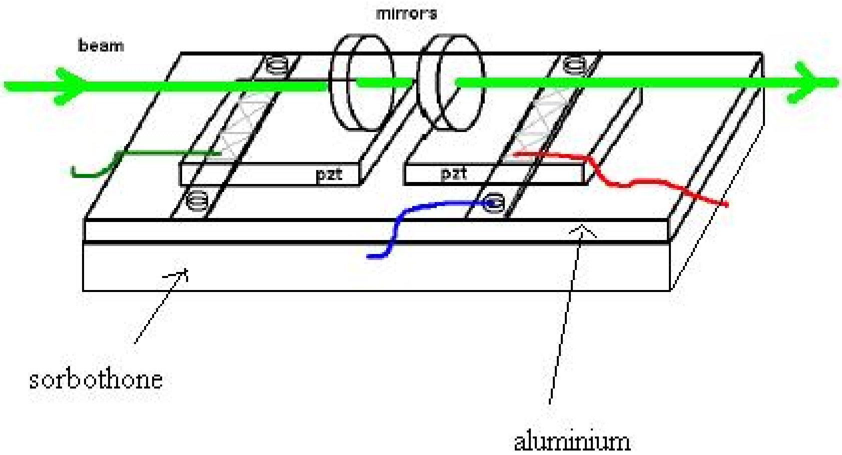
the intensity of the light (in Candela). So it is possible to observe the spectrum of the beam with an oscilloscope. The needed shift for the length is $\Delta e = \lambda_{min}\,\lambda_{max} / (2\,(\lambda_{min}-\lambda_{max}))$
We will scan from 760 nm to 810 nm so De = 6,156 mm

The PZT equation is :

$$\Delta e = \frac{d \times L \times \Delta V}{e}$$ L : length , d : strain coefficient, V : voltage, e : thickness

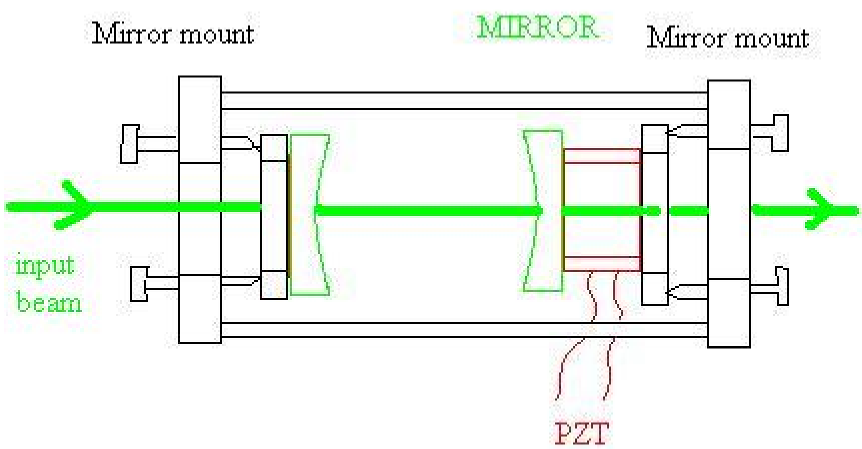
We need a voltage of 100 V to scan properly the output beam.

The first idea was to fix the mirrors onto two flat rectangular PZT :



But this assembly is not enough stable and this kind of PZT doesn't exist.

The second idea is to use two mirror mounts like that in the ECDL system and a cylindrical PZT element :

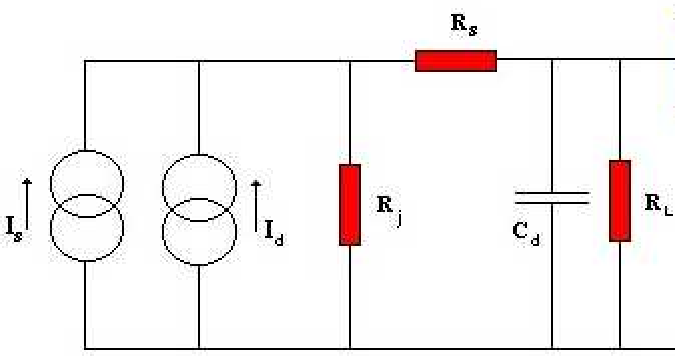


This system is more stable and easy to make without any aluminium plate needed. The PZT is stacked on the mirror in one side and on the mirror mount in the other side. The mirrors are just fixed by a M4 screw inside the mirror mount. The problem is the high price of this kind of PZT.

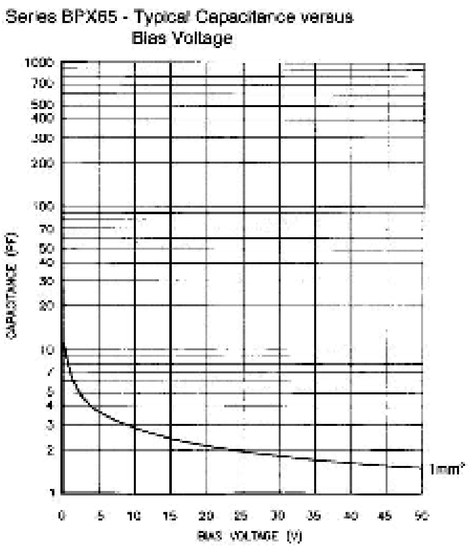
3) The photodiode circuit

To analyse the output beam spectrum we use a BPX65 photodiode link to an oscilloscope by a current amplifier level. The BPX65 is placed near the output beam but not on his path to not destroy the photodiode by a too high energy level.

Here is a schematic of the equivalent circuit of a photodiode :

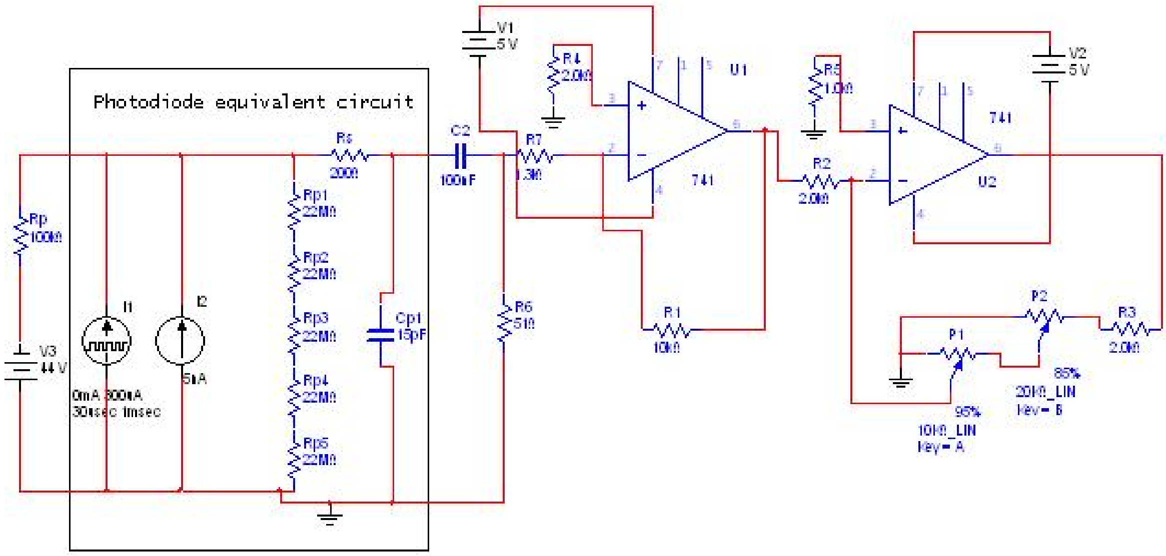


The photoconductive current has been modelled as a current source, Is, whose magnitude depends on the incident optical power. The constant current source Id models the dark current due to background radiations (5 nA max for our diode). The shunt resistance Rj has a value between 1 and 100 GΩ. The series resistance Rs corresponds to the bulk semiconductor and the contact resistance, its value is around 200Ω. Cd is the capacity of the PN junction due to the depletion region, its value depends of the polarisation.



The load resistor, R_L, shunts the total diode capacitance Cd and this time constant usually limits the speed of response. So we have to limit this value to have good features that imply a high bias voltage.

Here is the schematic of the amplifier circuit on Multisim :



The aim of the first level is to convert the photocurrent into voltage without changing the response time. So we use a transimpedance amplifier assembly with an operational amplifier. The advantage is that the noise of the backlash resistor is divided by the amplifier gain. The AOP must have low current entry (less than 10mA) and a wide bandwidth (>200MHz).We took a OPA355 with a 200MHz GBW , high slew rate of 360V/ms and low input bias current of 3 pA. It’s a CMS component.

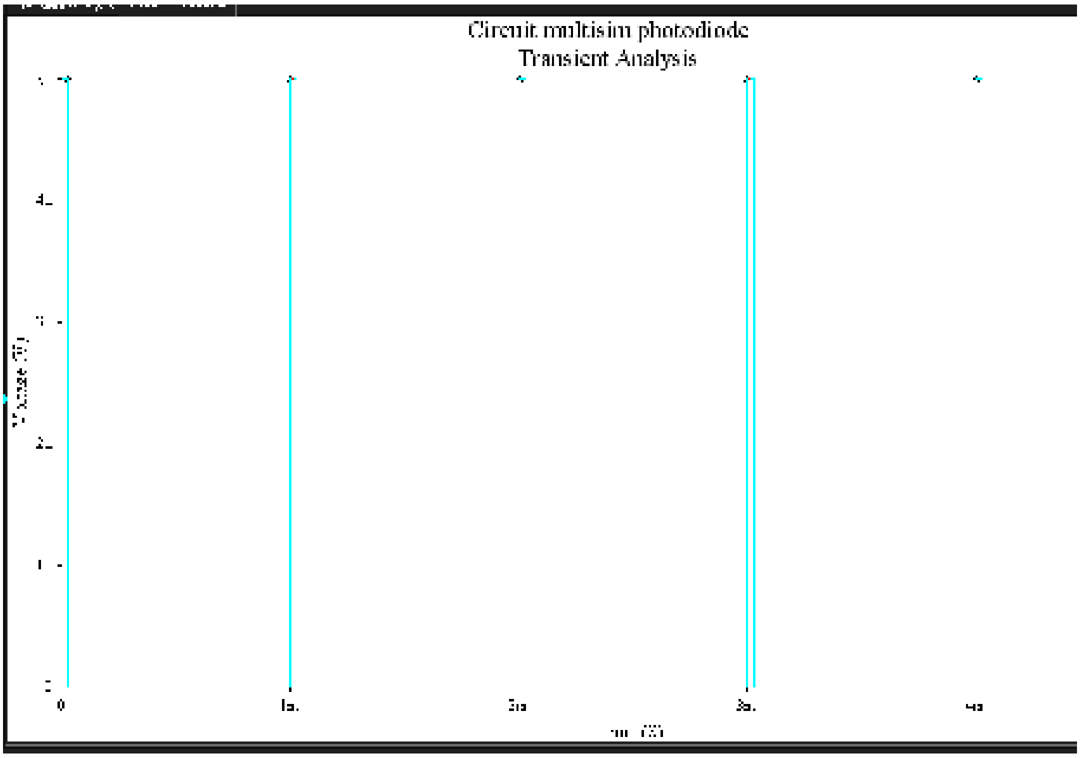
The second level allows the user to choose the amplification level by using the two potentiometers (one for a fine adjustment and another for a coarse adjustment). The AOP can be a simple 741. The relation between the output and the input of the second level is :

$$Vout = -\frac{P1 + P2 + R3}{R2} \times Vin$$

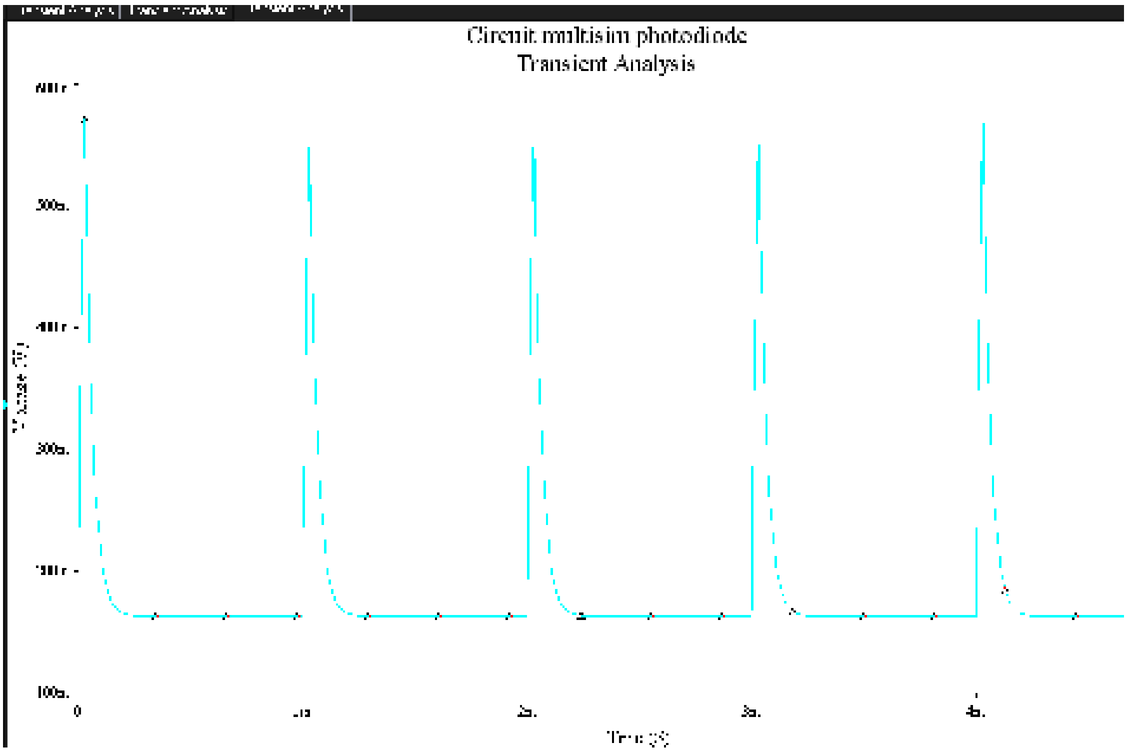
R4 and R5 protect the AOP of a saturation in the case of a cut on the circuit.

Below is an example of simulation, the photocurrent is settle to be a periodical pulse signal with these features : rise time : 1ns, fall time : 1ns, period : 1ms, max : 1uA, min : 0 A, pulse width : 30 us.

Here is the voltage which correspond to the input current passing into a 1Ω resistor :



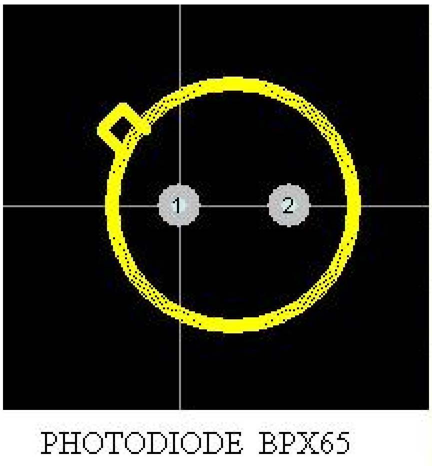
And below is the corresponding output with the potentiometers at 90 % :

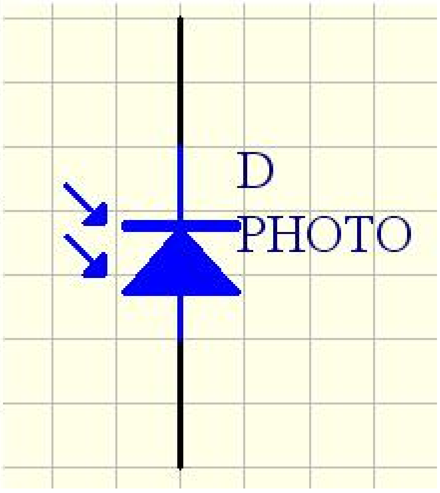


The capacitor has an effect on the shape of the output voltage but this impact does not disturb the analyse of the different peaks. To decrease this effect we have to apply the maximum reverse voltage that is possible. The limit for a BPX65 diode is 50 V , we will use a voltage of 2 x 22V = 44 V. The output maximum is above 500 mV for an input current from 300 uA to 10 mA and reach 5V between 1mA and 10mA that allow to analyse properly the spectrum of the beam.

After that we made the PCB :

We didn't find the footprint of the photodiode, so we have created it :



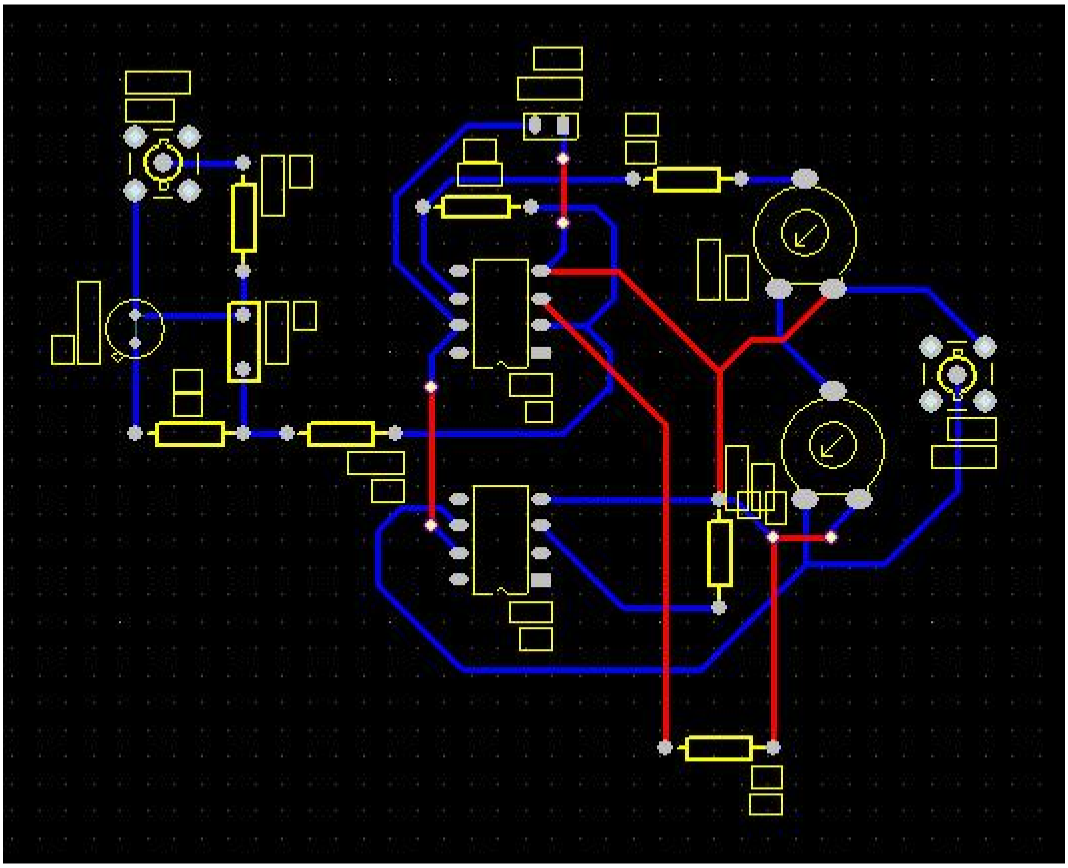


It is important to give the same names for the pins into the footprint and the schematic. Otherwise an error is generated when we download the netlist for the PCB.

Here is the list of the footprints for the other elements :

Component	Footprint
resistance	AXIAL0.4
capacitor	RAD0.2
AOP	DIP8
Potentiometer	POT_H
BNC connector	BU_BNC5X5V
2 pins connector	HDR1X2

Below is the PCB realised with Protel :



Unfortunately we did not have enough time to make and test this card.

Conclusion

For the moment we have done the temperature and current driver PCB and the boards will be make during this week.

The ECDL is almost finish, we just need to make holes for the photoresistance, the baseplate TEC and the output beam.

For the Fabry Perot we need to find another structure because the price of the PZT is too high.

The photodiode circuit will be probably make during the week.

The test will be make after the boards will be finish.

Bibliography :

Physic and optic bases theory :

- *(ref 1)* Grant R. Fowles. **Introduction to modern optics (second edition)**, Dover edition 1989

- *(ref 2)* N.W. Ashcroft, **Solid State Physics**, HRW International Editions

- *(ref 3)* Kittle C.. **Introduction to solid physics**. 5^{th} edition Wiley 1982

Theory of diode laser :

- *(ref 4)* Ye C. (Editor : Tjan Kwang Wei 2004). **Tunable External Cavity Diodes Lasers**. pp 7-27

- *(ref 5)* Silvast,W.T.. **Laser Fundamentals**. Cambridge University Press, Cambridge, 1996

- *(ref 6)* Svelto O.. **Principles of Lasers**. Plenum Press, New York, 1977

Theory of ECDL system :

- *(ref 7)* Ye C. (Editor : Tjan Kwang Wei 2004). **Tunable External Cavity Diodes Lasers**. pp 63-169

- *(ref 8)* G.-Y. Yan, Arthur L. Schawlow. **Measurement of diode laser characteristics affecting tunability with an external grating**. JOSA B, Vol. 9, Issue 11, pp. 2122- (November 1992)

Different design of ECDL system :

- *(ref 9)* Hong W. , Painter O.J.. **Design and characterisation of a Littrow configuration External Cavity Diode Laser** pp 1-8

- *(ref 10)* Andalkar A., Lamoreaux S.K., Warrington R.B.. **External Cavity Diode Laser System for Cesium D1.** Review of Scientific instruments

- *(ref 11)* O.I. Permyakova, A.V. Yakovlev, P.L. Chapovsky. **Simple external cavity diode laser**. 2003

- *(ref 12)* Toni Laurila, Timo Joutsenoja, Rolf Hernberg, Markku Kuittinen. **Tunable External-Cavity Diode Laser at 650 nm Based on a Transmission Diffraction Grating**

Applied Optics, Vol. 41, Issue 27, pp. 5632-5637 (September 2002)

- *(ref 13)* Lars Hildebrandt, Joachim Sacher, Richard Knispel. **Rubust external cavity diode laser (ECDL) with implemented antireflection coated blue laserdiodes and their performance in atom absorption spectroscopy.** Laser Applications to Chemical and Environmental Analysis (LACEA) 2002, paper SuB1

- *(ref 14)* Jianwei Zhang, Kaikai Huang, Donghai Yang. **Compact extended cavity diode laser system for small optically pumped cesium beam frequency standards** Chinese Optics Letters, Vol. 4, Issue 9, pp. 525-528 (September 2006)

Disign of a precision current source for laser diode application :

- *(ref 15)* P. Saunders and D. M. Kane, **A driver for stable-frequency application of laser diodes**, Rev. Sci. Instrum**, 63**, 2141 (April 1992).

- *(ref 16)* C. C. Bradley, J.Chen and Randall G. Hulet,**Instrumentation for a stable operation of laser diodes,** Rev. Sci. Instrum.,**61**, 2097 (May 1990)

- *(ref 17)* Michael S. Cafferty and Eric D. Thompson, **Stable current current with protection circuits for a lead-salt laser diode**, Rev. Sci. Instrum.,**60**, 2896 **(**May 1989)

Disign of a precision temperature controller for laser diode application :

- *(ref 18)* C. C. Bradley, J.Chen and Randall G. Hulet,**Instrumentation for a stable operation of laser diodes,** Rev. Sci. Instrum.,**61**, 2097 (May 1990)

- *(ref 19)* D.D.L. Wijngaards and R. F. Wolffenbuttel, **Characterisation and modelling of planar on-chip integrated Peltier elements for highly localised thermal stabilisation and cooling**

- *(ref 20)* Dr M. J. Willis, **Proportional-Integral-Derivative Control**, Dept. Of Chemical and Process Engineering, **(**October 1999)

- *(ref 21)* Carl E. Wieman and Leo Hollberg, **Using diode laser for atomic physics**, Rev. Sci. Instrum**, 62**, 1 (August 1990).

Fabry Perot :

- *(ref 22)* D. M. Kane, **The use of the nearly confocal interferometer for the study of transverse modes of a cavity**. Am J. Phys. 59, 235-239

- *(ref 23)* J. C. Amato, R. E. Williams and H. Helm. **An inexpensive, easy to build Fabry-Perot interferometer and its use in the introductory laboratory.** American Journal of Physics -- December 2005 -- Volume 73, Issue 12, pp. 1135-1138

Annexes

AutoCad :

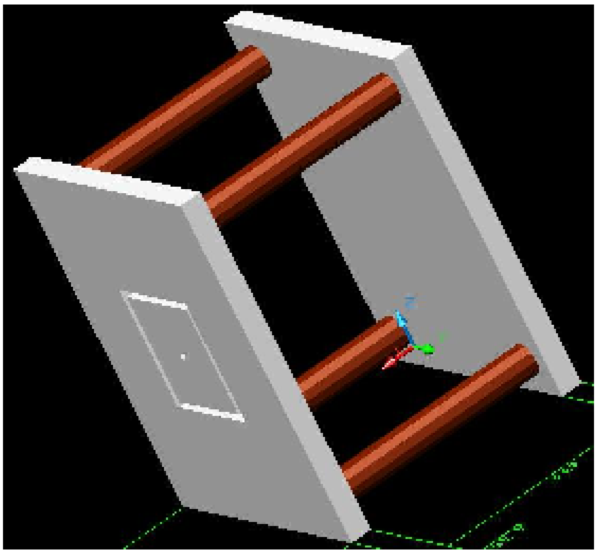
The advantages of the software :

To make the external cavity properly it is essential to well choose the software which will provide the design. AutoCad is probably one of the most convenient to make it, because it includes intelligent dimensioning.

AutoCAD can be useful for a wide variety of disciplines :

- Architectural, Engineering, and Construction
- Mechanical
- Geographic Information Systems
- Surveying and Civil Engineering
- Facilities management
- Electrical/electronic

It is possible to make 2D or 3D drawings. Here is an example of 3D drawing :

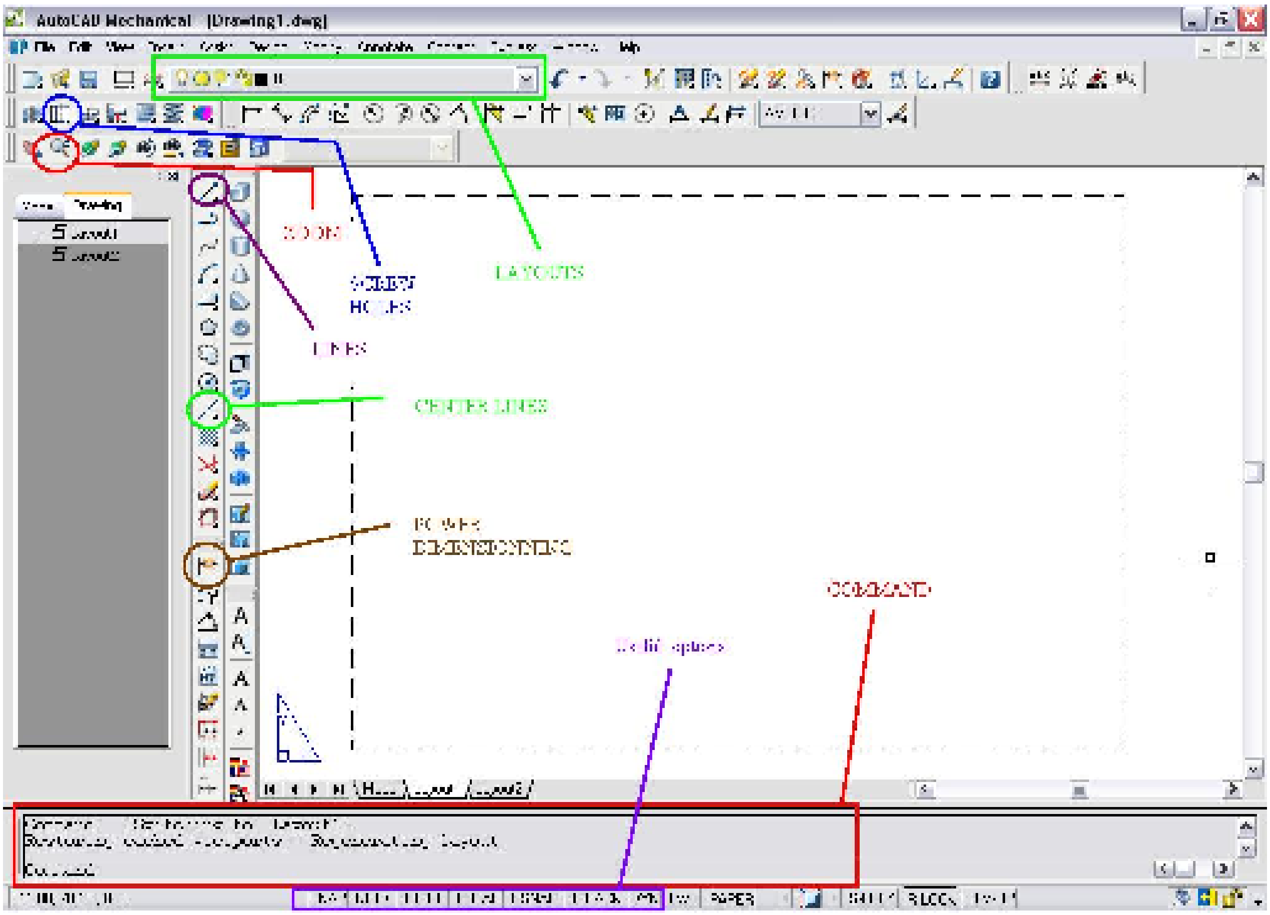


It is possible to add titles and descriptions on the drawing.

The software can easily change the scale of the drawing.

After have been drawn the 2D designs can be easily print to be mill.

The environment :

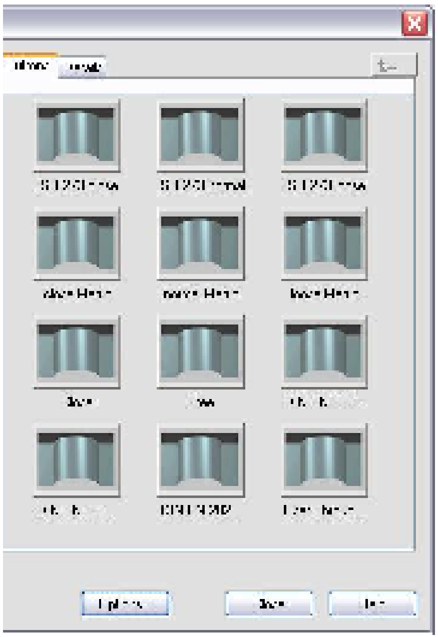


The main part of AutoCad is a drawing board. The limits of the drawing are defined by the dashed lines.

To display only some elements but not all it is possible to use the different Layouts. To display or not a Layout light on or light off the corresponding bulb. For example if we choose a Layout and after draw a circle, this circle will be include in the corresponding Layout. If we turn of the bulb of the Layout the circle disappear.

In the command we can indicate the features of each part, for example the length and the angle of a line, the height and the width of a rectangle. And an error signal is indicated if there are any problems.

The holes for the screws can be accurately define choosing the good grade among a large choice :



The useful options :

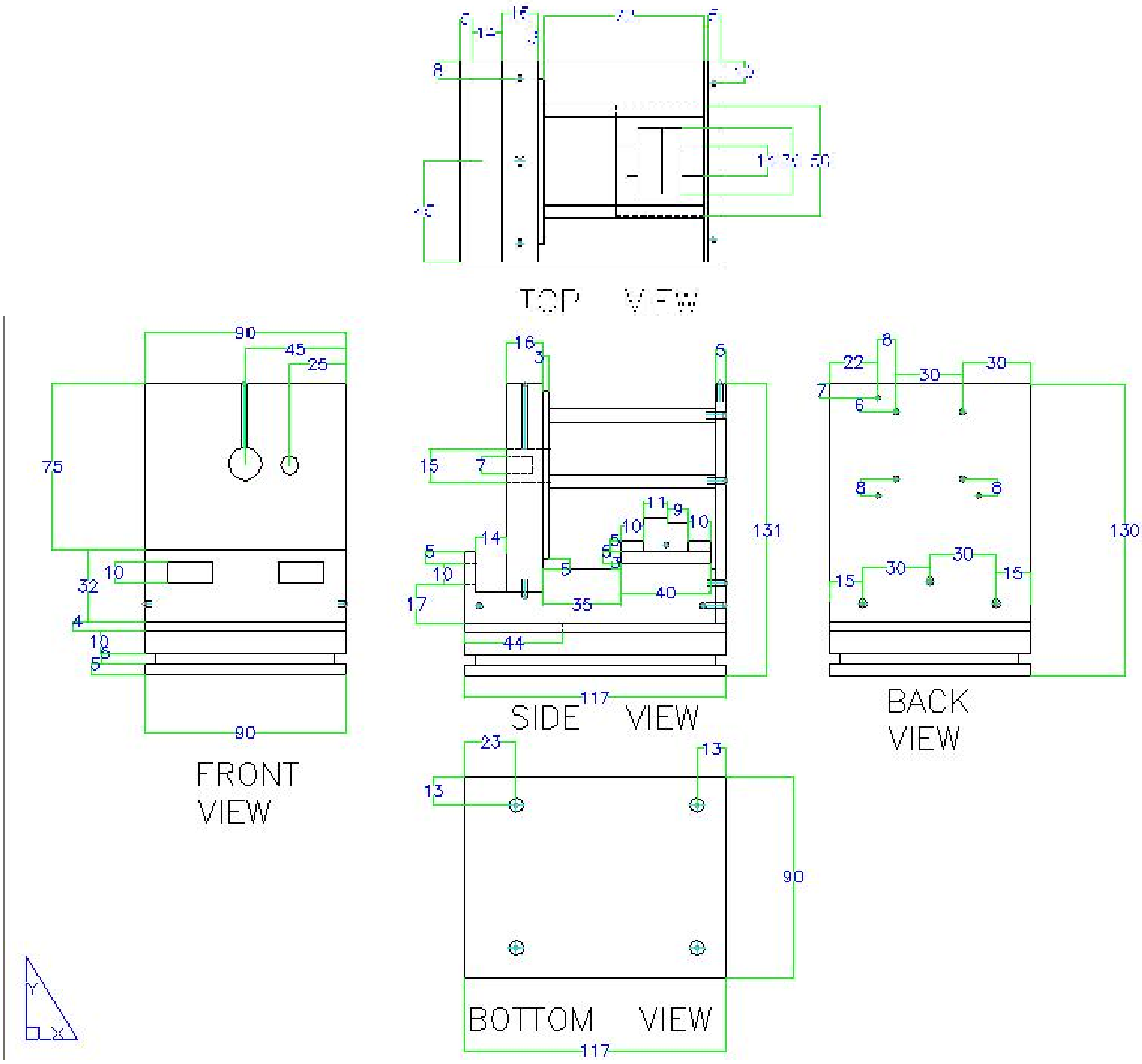
DYN : to put directly the dimensions of the lines and circles without write anything in the command frame.
OSNAP : Allow to directly click on a tip or a middle of a line.
GRID : to display a reference grid whose step can be defined in the options.
ORTHO : to make 90° angles lines.
POLAR : to make for example 45° lines (it depends how is settle this option by the user).

Some advices to well use AutoCad :

Don’t forget to change the unit : Assist/Format/Units.
To zoom in the all schematic : View/Zoom/All.
To define the angle and displacement step : Assist/Drafting settings/Drafting settings.
The centrelines are very useful to align for example side view and front view.
To put the dimensions click on Power Dimensioning then click on the two points where you want measure the distance.
To change the format of the sheet : Assist/Format/Drawing limits (click on GRID button to see the result).

ECDL Dimensions :

Here are all the dimensions of our ECDL design, the values are all in millimetres.



The total volume of this ECDL device is 117 x 90 x 130 = 1368900 mm^3

Below are the dimensions of the cover :