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CONTROL ELECTRONICS FOR AN IODINE STABILIZED HE-NE LASER

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#### Abstract

This work describes the design of an electronic controller unit for an iodine stabilized $\mathrm{He}-\mathrm{Ne}$ laser. This unit contains frequency stabilization electronics utilizing the third harmonic locking technique. Also provided in this unit are temperature regulation electronics for iodine cell cooling and high--voltage electronics for laser tube operating current generation.

The theory of frequency stabilization is very briefly explained. Electronic design considerations and sources of errors are described more thoroughly, however. Also, test results are given and analyzed. An operational test was performed on the electronic unit by comparing this unit equipped with a laser unit to another similar system by beat frequency measurement. The observed frequency difference was between 5 kHz and 8 kHz , which agrees with results of previous comparisons between the laser systems.


## Keywords:

Iodine stabilized laser, wavelength standard, length metrology.

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Tässä työssä selostetaan jodistabiloidun $\mathrm{He}-\mathrm{Ne}$ laserin elektroniikan suunnittelu. Suunniteltu laite sisältää taajuuden stabilointielektroniikkaa, joka käyttää hyväkseen kolmannen harmonisen lukitusmenetelmää. Laitteessa on myös lämpötilasäätäjä jodikennon jäähdyttämiseen ja suurjännite teholähde laserputken käyttövirran generoimiseen.

Taajuuden stabiloinnin teoria on selitetty erittäin lyhyesti. Elektroniikan suunnittelu ja eri virhelähteiden vaikutukset on esitetty yksityiskohtaisemmin. Testituloksia on myös esitetty ja analysoitu. Testejä suoritettiin pelkälle elektroniikkayksikölle ja myös vertaamalla tätä yksikköä, laserputkella varustettuna, toiseen vastaavaan järjestelmään. Vertailu tehtiin mittaamalla lasersäteiden sekoitustaajuutta. Taajuusero oli $5 \mathrm{kHz}: \mathrm{n}$ ja $8 \mathrm{kHz}: \mathrm{n}$ välillä, mikä vastaa aiemmin tehtyjen vertailujen tuloksia.

## Avainsanat:

Jodistabiloitu laser, aallonpituuden normaali, pituuden metrologia.

## Preface

The work described in this thesis took place in Helsinki University of Technology, faculty of electrical engineering at the laboratory of metrology and the metrology research institute. The subject was given to me by Assoc. prof. Pekka Wallin. The electronics was built for University of Turku. I thank Prof. Jyrki Kauppinen of the applied physics laboratory at University of Turku for the support that made this work possible. I also like to thank Dr. Birger Ståhlberg from University of Helsinki, Dr. Kari Riski from Centre for Metrology and Accreditation and Lic. Techn. Jari Koskela for giving me valuable advice regarding the electronics. Dr. Erkki Ikonen and the rest of the laboratory staff I thank for giving me practical advices and support. Special thanks go to Lic. Techn. Jianpei Mu, who patiently helped me to understand the physics side of the laser and who let me use the iodine stabilized laser he built, for testing of the electronics.

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# List of symbols and abbreviations 

Component reference designators:<br>C Capacitor<br>D Diode<br>J Connector or jumper wire<br>L Inductor<br>R Resistor<br>U Integrated circuit<br>X Crystal

| Symbols: |  |
| :--- | :--- |
| $A$ | Scale factor (value unknown) |
| $c$ | Speed of light in vacuum |
| $G$ | Gain |
| $I$ | Intensity |
| $k$ | Iodine absorption contrast |
| $l_{0}$ | Laser cavity length |
| $l_{1}$ | Relative position of mirror attached to modulation PZT |
| $l_{2}$ | Relative position of mirror attached to tuning PZT |
| $n$ | Number of standing wave periods in laser cavity or mode number |
| $T$ | Known temperature |
| $V_{p e a k}$ | Peak voltage (at the 3f detector output) |
| $V_{o f f s e t ~}$ | Offset voltage (at the 3f detector output) |
| $X$ | Unknown temperature |
| $\nu_{L}$ | Laser frequency |
| $\lambda$ | Laser wavelength |
| $\nu_{g}$ | Center frequency of laser gain profile |
| $\Delta \nu_{D}$ | Laser gain profile width |
| $\beta$ | Frequency spacing between positive and negative peaks |
| $\gamma$ | of the 3rd derivative of the iodine absorption profile |
| $\tau$ | Iodine absorption natural linewidth |
| $\tau$ | Time constant |
| $\tau_{g}$ | Group delay |
| $\Delta V_{o u t}$ | Integrator output voltage change |
| $\Delta t$ | Time difference |
| $\Delta l$ | Position or length change |
| $\Delta f$ | Laser frequency offset |
| $\theta$ | Phase angle |
| $\omega$ | Angular frequency |


| Abbreviations: |  |
| :--- | :--- |
| AC | Alternating current |
| BIPM | Bureau International des Poids et Mesures |
| BPF | Bandpass filter |
| CIPM | Comité International des Poids et Mesures |
| CMOS | Complementary symmetry metal oxide semiconductor (logic type) |
| DC | Direct current |
| ECO | Engineering change order |
| FET | Field effect transistor |
| HCMOS | High speed CMOS |
| He-Ne | Helium-neon (gas mixture) |
| HUT | Helsinki University of Technology |
| IC | Integrated circuit |
| LED | Light emitting diode |
| LS-TTL | Low power schottky TTL |
| NEC | Nippon electric company |
| NPL | National Physical Laboratory |
| PI | Proportional-Integrating (Regulation method) |
| PZT | Piezoelectric transducer |
| RMS | Root mean square |
| THD | Total harmonic distortion |
| TIM | Transient intermodulation distortion |
| TTL | Transistor transistor logic |
| XOR | Exclusive OR (logic gate) |

## Chapter 1

## Introduction

Iodine stabilized He -Ne lasers are not generally used in everyday laser measurements. This is because they are sensitive to interference (mechanical vibrations and optical feedback), they have low output power and they need a lot of attention from the operator. Some low noise laser tubes used in iodine stabilized lasers have a relatively short lifetime, which results in a high operation cost aswell. However, the high frequency stability of the iodine stabilized lasers make them an excellent choise for realization of the lenght standard. Many laboratories have built portable iodine stabilized laser systems which can be brought together and compared. In practice it has been noticed that lasers built at different locations show very small frequency differences at these comparisons [Chartier]. The relative uncertainty can be better than $10^{-11}$.

Commercial lasers (e.g. Zeeman stabilized lasers) that are more practical to use but are less stable can be calibrated regularly by comparing them with the iodine stabilized laser. The comparison is made by combining the beams from both lasers and detecting the resulting beat signal with a fast photodetector. The frequency of the photodetector output signal is measured with a frequency counter and is the frequency difference of the lasers. Some measurements that require a stable reference are performed with the iodine stabilized laser directly. One example of this kind of measurement is the geophysical investigation with earth strainmeters. Because the drift of an iodine stabilized laser is very small, it is suitable for measurement of small and slow seismic events. The main problem is that the laser might jump out of lock during a long measurement. The requirements of the electronics are
relaxed by the fact that modern electronics outperform the laser tube and iodine cell as far as stability is concerned. The goal for instability due to the control electronics built in this work was set to $\pm 1 \mathrm{kHz}$ relative to the laser frequency ( $\approx 473612370 \mathrm{MHz}$ ).

In chapter 2 of this thesis the operating theory of the frequency stabilization scheme is briefly explained from the electronics point of view. Computer simulation of the physics of the iodine absorption is presented in this chapter aswell. Also described in this chapter is the theory behind the temperature control electronics and the laser power supply. Chapter 3 reviews the electronic implementation of the different electronic modules. Some design formulas are given and alternative solutions are discussed. In chapter 4 the performance of the built unit is evaluated. Measurement results are presented and analyzed. Chapter 5 explains how to use the equipment to get a good laser stability. In chapter 6 advanced instructions for calibration of the electronics are given. Also explained here are different methods of verifying and measuring the accuracy of the electronics unit.

## Chapter 2

## Operating theory

### 2.1 Frequency stabilization

The frequency stabilization of the $\mathrm{He}-\mathrm{Ne}$ laser in this work utilizes the third harmonic locking technique to one of the iodine absorption lines at 633 nm . The iodine is situated in an iodine cell in the light path inside the laser cavity. This iodine cell is temperature controlled to stabilize the iodine vapour pressure. The iodine absorption cell makes small intensity peaks to appear in the intensity of the laser beam. The topmost graph in figure 2.1 shows the intensity versus frequency function of the laser. The absorption peaks are marked $\mathrm{d} \ldots \mathrm{j}$. Note that the heights of the absorption peaks are small compared to the intensity background. The slope that the absorption peaks are riding on is from the laser tube gain profile. For some laser tubes the gain maximum is at a higher frequency than shown in the figure. Beneath the intensity curve in figure 2.1 there are curves that show the derivatives of the intensity function. The third harmonic locking technique uses the third derivative for locking as there are steep zero crossing points at the center of the absorption lines. This makes locking easy and accurate.

The locking is performed by modulating the laser frequency slightly with a sinusoidal modulating signal and by detecting the laser output intensity variations. The third derivative of the intensity curve can be measured by detecting the third harmonic component of the modulation frequency in the laser intensity output with a phase sensitive detector. The output of the phase sensitive detector is integrated twice and then fed to a tuning piezo that tunes the laser frequency. When there is a voltage at the output of


Figure 2.1: Calculated laser output intensity. The vertical scales are normalized to the peak values.


Figure 2.2: Frequency locking block diagram
the phase sensitive detector, the integrators will integrate this voltage and adjust the laser frequency until the detector output voltage is zero. This means that the laser frequency will be locked to the zero crossing point of the third derivative curve of the intensity function, or in other words, the center of the iodine absorption peak. An integrator is needed in the control loop to remove any static offset errors. A second integrator is needed to remove the error caused by linear drift of the laser cavity length caused by temperature change [Konttila]. If the temperature change is not linear with time, however, errors in the laser frequency will occur and therefore the laser cavity is designed to be insensitive to temperature changes and other environmental effects [Latvala] [Hu].

The relative strength and the form of the harmonic components were studied by computer simulation of the laser tube equipped with an iodine cell. The formulas for the laser intensity profile and for the iodine absorption profile given in reference [Shotton/Rowley] was used. The signal flow block diagram of the third harmonic locking scheme is shown in figure 2.2. Note that the blocks in this diagram do not necessary exist as separate physical
modules in the actual laser system. The laser block is considered as a length to frequency converter. The laser frequency $\nu_{L}$ is dependent on the relative positions of the mirrors $l_{1}$ and $l_{2}$ as:

$$
\begin{equation*}
\nu_{L}=\frac{c n}{2\left(l_{0}+l_{1}+l_{2}\right)}, \tag{2.1}
\end{equation*}
$$

where

$$
\begin{equation*}
n \approx \frac{2 l_{0}}{\lambda} \tag{2.2}
\end{equation*}
$$

is the number of periods of the standing wave in the cavity. The refractive index of air is considered to be 1 in this investigation. Constant $c=$ $299792458 \mathrm{~m} / \mathrm{s}$ is the speed of light (in vacuum), $\lambda=632.991 \mathrm{~nm}$ is the wavelength of the laser and $l_{0}$ is the length of the laser cavity. $l_{0}=371.0 \mathrm{~mm}$ in this simulation. $n$ is an integer number in theory but is approximated here in the simulation by a real number.

The absorption cell block implements a frequency to intensity conversion function:

$$
\begin{equation*}
I=A e^{-\left(\frac{\nu_{L}-\nu_{g}}{\Delta \nu_{D}}\right)^{2}} \cdot\left(1+k \frac{1}{1+\left(\frac{\nu_{L}-\nu_{p}}{\gamma}\right)^{2}}\right) . \tag{2.3}
\end{equation*}
$$

The exponential factor is the approximation of the gain profile of the laser tube. $A$ is a scale factor. The last factor is the absorption peak of the absorption cell. In the actual laser unit the absorption cell is situated inside the laser cavity. This equation is valid only for frequencies close to one of the absorption lines. To get a wider range for the equation, several absorption peak factors would have to be multiplied together. Only one absorption peak is investigated here though. The other parameters in the intensity equation for the absorption cell block are listed in the table below.

| parameter | value |  |
| ---: | :---: | :---: |
| laser resonance frequency | $\nu_{g}$ | $\approx 473612370 \mathrm{MHz}$ |
| laser gain profile width | $\Delta \nu_{D}$ | $\approx 1400 \mathrm{MHz}$ |
| iodine absorption contrast | $k$ | $\approx 5 \cdot 10^{-3}$ |
| iodine absorption width | $\gamma$ | $\approx 2.4 \mathrm{MHz}$ |

Additionally there is the iodine absorption line frequency $\nu_{p}$. The absorption lines typically used with 633 nm He-Ne lasers [BIPM] [Metrologia] are listed in the following table:

| absorption line | frequency $\left(\nu_{p}\right)$ |
| :---: | :---: |
| j | 473612193.140 MHz |
| i | 473612214.705 MHz |
| h | 473612236.644 MHz |
| g | 473612340.399 MHz |
| f | 473612353.597 MHz |
| e | 473612366.960 MHz |
| d | 473612379.821 MHz |

The PZT blocks are piezoelectric transducers (PZT) that convert a voltage to a displacement. The topmost PZT in figure 2.2 is called the modulation piezo as it is controlled by the sinusoidal modulation voltage $f$ from the modulation frequency generator. One of the laser cavity mirrors is attached to this piezo. The position $l_{1}$ of this mirror is zero when no voltage is connected to the piezo. If the modulation of the laser frequency is 6 MHz peak-to-peak the position of the mirror moves between $l_{1}=-2.35 \mathrm{~nm}$ and $l_{1}=+2.35 \mathrm{~nm}$. The modulation voltage amplitude needed for this movement is 1.70 V for the piezo used (PZT 5A material) [ Hu ]. The sensitivity of the modulation piezo is then $\Delta l / \Delta V \approx 1.38 \mathrm{~nm} / \mathrm{V}$.

The other PZT is the high voltage piezo or the tuning piezo. This piezo moves the mirror on the other side of the laser cavity and tunes the average laser frequency. The position $l_{2}$ of the tuning piezo is zero when no voltage is applied. The sensitivity of this piezo is $\Delta l / \Delta V \approx 3 \mathrm{~nm} / \mathrm{V} \ldots 4 \mathrm{~nm} / \mathrm{V}$. In the simulation this sensitivity was mistaken to be slightly smaller and thus the tuning voltage shown in the simulation results is a little high.

The photodetector block in figure 2.2 converts the intensity to a voltage. The PZT, laser, absorption cell and photodetector blocks were simulated with a computer. The Matlab ${ }^{1}$ script program used can be found in the appendix. Figure 2.3 shows the results of the simulation. The DC component is the intensity, the $f$ component is the intensity first derivative, the $2 f$ component is the intensity second derivative and so on. The graphs show the magnitude (absolute value) of the harmonic components, except for the $3 f$ component that shows the amplitude as recovered by phase sensitive detection. The magnitudes of the harmonic signals in the graphs in figure 2.3 can be compared to each other whereas in the case of figure 2.1 they could not.

[^0]The effects of modulation amplitude changes ${ }^{2}$ and laser background tilting can be easily studied with this program but this has not been done within this work.

The $3 f$ bandpass filter after the photodetector filters out the $3 f$ component from the noisy detector signal. This makes detection easier and allows a bigger amplification of the detector signal without saturating the amplifiers. The components of the frequency stabilization are situated in four separate modules in the actual electronic implementation. The modulation frequency generator is one. The amplifier marked $G_{2}$ in the block diagram is the PZT amplifier and is also one separate module. The photodetector and the 3 f bandpass filter are situated on a small printed circuit board inside the laser unit with the laser tube, absorption cell, cavity mirrors and the piezos. The photodetector amplifier and filters are situated on the same circuit board with the photodiode to minimize noise and electrical interference. The phase sensitive detector, integrator I, amplifier G and integrator II are situated in a module called the lock-in amplifier. Part of the $3 f$ bandpass filter can also be considered to be located inside the lock-in amplifier.

### 2.2 Temperature control

The temperature control of the iodine cell is performed by cooling the cell's cold finger down to $15.0^{\circ} \mathrm{C}$ and keeping it stable withing $\pm 0.2^{\circ} \mathrm{C}[\mathrm{Hu}]$. The cooling is achieved with a peltier element. The peltier element is a semiconductor component with several specially arranged junctions in series. When a current is driven through the element, it will cool down a cooling well with the iodine cell's cold finger inside. By reversing the peltier element current the cooling well could be heated, but in this application only the cooling feature is used.

Inside the cooling well there is a temperature probe that measures the temperature of the well. The temperature control electronics compares the well temperature sensed by the probe to a reference setting set by the operator (usually $15.0^{\circ} \mathrm{C}$ ). The electronics will adjust the cooling current so that the temperature in the cooling well is kept the same as the reference setting.

[^1]


Figure 2.3: Simulation results of the absorption peaks

The regulation is linear unlike the thermostats in commercial refrigerators that switch the cooling abruptly on and off. The regulation is performed by a typical PI-controller circuit, which ideally removes any static temperature offset errors.

### 2.3 Laser power supply

The NEC He-Ne laser tubes used in 633 nm iodine stabilized lasers operate at a current of approximately 3.0 mA . A current adjustment range of $2 \mathrm{~mA} . . .4 \mathrm{~mA}$ is usually desired. The operating voltage requirement depends on the size of the ballast resistor and is usually $800 \mathrm{~V} \ldots 1700 \mathrm{~V}$. This voltage will rise when the tube gets older and eventually the voltage output by the power supply will not be high enough to light the tube. The ballast resistor is situated in series with the tube and close to it. The function of the ballast resistor is to limit the current and to protect the laser tube. The initial voltage requirement to start the discharge in the laser tube is higher than the operating voltage. This voltage is in the range 1600 V... 3200 V . Many laser power supplies have a special button that can be pressed to start the discharge. This button will induce a high voltage over the tube until it hopefully discharges. These buttons should only be pressed if the laser tube won't start discharging by themselves.

When operating high voltage devices one should be careful not to touch, or even come close to, unshielded high voltage conductors. The voltages may stay charged in cables (even unconnected) and power supplies for several hours after the power switch has been turned off.

## Chapter 3

## Electronic implementation

### 3.1 General design

The laser electronics unit built by Jari Koskela [Koskela] at Helsinki University of Technology was used as a model when designing this new unit. Most parts were changed, however. The lock-in amplifier and the PZT high voltage amplifier were redesigned completely. Some changes were made to fix some known problems with this older unit, others were made just as a technical experiment. In this chapter references to parts are frequently beeing made. References are given as part reference designators and part type numbers. These parts can be found and studied in the schematic drawings and part lists in the appendix.

### 3.2 Modulation frequency generator

### 3.2.1 Digital part

In the design of the digital circuitry, care has been taken to avoid interference by digital noise in the rest of the unit. The 5 V power supply needed for digital logic circuits is separated from the power supplies for analog circuits. The digital circuits ground and the analog circuits ground connects only at one point close to the digital to analog conversion. HCMOS logic chips have been used whenever possible as these don't generate interfering spikes in the operating current and ground lines. Because some HCMOS logic chips
weren't available at this time some TTL and LS-TTL chips had to be used aswell.

The modulation frequency generator uses a 1 MHz Crystal (X1) as a frequency reference. This results in a very stable modulating frequency compared to LC or RC oscillators. Even though the iodine cell and the phasesensitive detection are insensitive to modulation frequency drift, the bandpass filters before the phase-sensitive detector may cause unwanted changes in loop-gain if the frequency drifts. The crystal oscillator is built with HCMOS inverters (U1, 74HC04). The crystal's resonating frequency is dependent by a small amount on the capacitive loading of the crystal. However, as the absolute frequency of the oscillator is not important, this has not been accounted for.

The 1 MHz clock signal generated by the crystal oscillator is divided by 60 to get a 16666.67 Hz clock signal. The division is performed with logic counter and flip-flop chips in three steps. First counter chip (U2, 74LS92) divides the clock rate by 3 , then another counter chip (U3, 74LS90) divides the clock rate by 10 . Finally a D-flip-flop (U4, 74HC74) divides the rate by 2 and outputs a symmetrical 16666.67 Hz square wave. This rate is 6 times the modulating frequency (6f). Another D-flip-flop (U4, 74HC74) divides this rate further by 2 to get 8333.33 Hz . This is 3 times the modulating frequency ( 3 f ) and the frequency used as reference in the phase-sensitive detector for 3rd harmonic locking.

The reference output is taken through a 4-position rotary switch so that one can select between phase shifted versions of the reference frequency. $0^{\circ}$ phase shift reference is taken directly from the noninverting output of flipflop U4. $180^{\circ}$ phase shifted reference is available at the inverting output of the same flip-flop. $90^{\circ}$ and $270^{\circ}$ phase shifted references are generated at noninverting and inverting outputs respectively of a D-flip-flop (U5, 74HC74) that delays the $0^{\circ}$ reference by half a period of the 6 f clock signal (that is one quarter period of the 3 f clock). After the rotary switch the reference signal is buffered by two inverters ( $\mathrm{U} 1,74 \mathrm{HCO4}$ ) and finally routed to the BNC connector on the front panel.

The 2777.78 Hz modulating frequency (f) is generated from the reference frequency (3f) by a divide by 3 circuit consisting of three D-flip-flops (U5, U7, 74HC74) and some logic gates (U8, U9, 74HC00, 74HC20). This divider circuit is clocked by the 6 f clock signal explained earlier. This 6 f clock signal is fed through a monostable flip-flop (U6, 74121) that performs a delay on the signal's rising edge. This delay is variable by a potentiometer situated


Figure 3.1: Frequency generator clock signals
at the front panel. This control allows fine adjustment of the relative phase between the reference clock signal and the modulating frequency signal within a range of $0^{\circ}$ to $140^{\circ}$. This value has to be added to the $0^{\circ}, 90^{\circ}, 180^{\circ}$ or $270^{\circ}$ reference phase that is selected by the rotary switch. It has to be noted that $360^{\circ}$ phase shift corresponds to one period of the 3 f reference clock here. From the divider circuit two 2777.78 Hz clock signals are outputted $60^{\circ}$ out of phase. These two clock signals are used to generate a smooth 2777.78 Hz sinewave by the analog circuitry explained next. Figure 3.1 shows the relationship between the different clock signals. $\operatorname{Mod}_{1}$ is the output from U7's pin 6 and Mod ${ }_{2}$ is the output from U7's pin 8.

### 3.2.2 Analog part

The digital modulation frequency signals $\operatorname{Mod}_{1}$ and $\operatorname{Mod}_{2}$ are summed thru R4 and R5 into U103. U103 works as a differential amplifier, thus eliminating possible common mode interference from the digital domain. When Mod ${ }_{1}$ and $\mathrm{Mod}_{2}$ are summed, a waveform shown in figure 3.2 is generated. If we examine the fourier series of $\operatorname{Mod}_{1}$ and $\operatorname{Mod}_{2}$, which are:

$$
\begin{align*}
\operatorname{Mod}_{1} & =\sum_{i=0}^{\infty} \frac{2 E}{\pi(2 i+1)} \sin (2 \pi f(2 i+1) t) \quad \text { and }  \tag{3.1}\\
\operatorname{Mod}_{2} & =\sum_{i=0}^{\infty} \frac{2 E}{\pi(2 i+1)} \sin \left(2 \pi f(2 i+1) t-\frac{(2 i+1) \pi}{3}\right) \tag{3.2}
\end{align*}
$$



Figure 3.2: Generated waveform by summing two square waves $60^{\circ}$ apart


Figure 3.3: Frequency components of generated waveform
we note that they contain 3rd harmonics. $E$ is the peak-to-peak voltage of the squarewave. If we sum these squarewaves we get:

$$
\begin{align*}
& \operatorname{Mod}_{1}+\operatorname{Mod}_{2}= \\
& \quad \frac{4 E}{\pi} \cos \left(\frac{\pi}{6}\right) \sin \left(2 \pi f t-\frac{\pi}{6}\right)+ \\
& \quad+\sum_{i=2}^{\infty} \frac{2 E}{\pi(2 i+1)} \cos \left(\frac{(2 i+1) \pi}{6}\right) \sin \left(2 \pi f(2 i+1) t-\frac{(2 i+1) \pi}{6}\right) . \tag{3.3}
\end{align*}
$$

As we can see the 3 rd harmonic component is cancelled when the two $60^{\circ}$ out of phase square waves are summed. Figure 3.3 shows the frequency spectrum of the combined square waves.

The choice of a HCMOS chip ( $\mathrm{U} 7,74 \mathrm{HC74}$ ) to feed the summing resistors R4 and R5 is motivated by the symmetry and stability of it's pull-up and pulldown characteristics. There might be some differences, however, and also the resistors might not match. This could cause some 3rd harmonic leakage to the output waveform. To remove this residual 3rd harmonic frequency and
to smooth the waveform to a sinewave, a sharp bandpass filter is placed after the differential amplifier U103. The filter has an unadjusted center frequency of $2760 \mathrm{~Hz} \pm 100 \mathrm{~Hz}$ and a Q -value of $100 \pm 2$. The circuit that consists of three operational amplifiers (U101, U102, NE5534, NE5532) with surrounding resistors and capacitors implements a transfer function:

$$
\begin{equation*}
H(s)=-\frac{1}{R_{101} C_{102}} \frac{s}{s^{2}+\frac{1}{R_{103} C_{102}} s+\frac{1}{R_{107} C_{101} R_{104} C_{102}} \frac{R_{11}}{R_{109}}} \tag{3.4}
\end{equation*}
$$

This filter structure is called a Tow and Thomas biquad [Ghausi/Laker] and is relatively insensitive to component drift as investigated by Thomas [Thomas]. After the bandpass filter comes two identical cascaded second order lowpass filters to filter out any residual harmonics. The transfer function for each of these filters is:

$$
\begin{equation*}
H(s)=\frac{1}{s^{2} R_{115} R_{116} C_{105} C_{106}+s R_{116} C_{106}+s R_{115} C_{106}+1} \tag{3.5}
\end{equation*}
$$

The attenuation of harmonic components by the bandpass filter should be as follows:

| component | attenuation |
| :---: | :---: |
| f | 0 dB |
| 2 f | 43.5 dB |
| 3 f | 48.5 dB |
| 4 f | 51.5 dB |
| 5 f | 53.6 dB |

Figure 3.4 shows the combined transfer characteristics of all the filters, from the output of the digital logic to the modulation frequency output connector. This curve was plotted by simulating the circuits with PSPICE ${ }^{1}$ simulation software.

[^2]

Figure 3.4: Transfer function magnitude for digital wave to output

### 3.3 Lock-in amplifier

### 3.3.1 Block diagram

Figure 3.5 shows the main functional blocks of the lock-in amplifier. The signal from the photodetector located in the laser unit is connected to the input of the lock-in amplifier. The input gain is selectable between 2 dB and 7 dB . A 3 f bandpass filter may be enabled in the input signal path if desired. This filter provides some additional gain for the 3 f signal at the same time it attenuates undesireable noise and other harmonic frequency components.

Next the input signal reaches the input of the 3 f detector. To be safe the input of the 3 f detector is monitored by an overflow LED that lights up if the signal level is too high. The detector input signal is also available at the signal monitor output connector at the front panel of the unit. The 3 f detector gets it's two phase clock signals from a digital circuit that provides toggling between $0^{\circ}$ and $180^{\circ}$ phase shifts by a switch at the front panel. The digital reference input is isolated from this unit by an optocoupler to


Figure 3.5: Block diagram of the lock-in amplifier
minimize digital noise interference.
The output of the detector is integrated first by integrator I, then by integrator II. Integrator I has selectable gain and time constants. Integrator II can be turned on or off. The output of integrator II is the control output available at the front panel. This output voltage is also shown at a panel meter in the unit. The output of integrator I is monitored by a drift sensor that turns on the tracking LED that lights if the integrator output changes.

The output of the detector is buffered and monitored with a $3 f$-detect indicator LED. The buffer amplifier has a gain of 11 . This buffered detector signal is also fed thru a passive RC lowpass filter and connected to a BNC connector at the front panel. This detector monitor output is used for $\mathrm{Y}-$ deflection in the XY-oscilloscope that is used when searching for absorption lines.

The 3 f filter, detector and integrator I are each situated on separate printed circuit boards that are attached to the main lock-in amplifier board with pinheader connectors. This arrangement allows experimentation with different types of modules. The idea was to try several solutions for the $3 f$ detector circuit to find the best one. Unfortunately there was time to build and test only one detector. See the appendix for the dimensions and electrical connections for the plug-in units.

### 3.3.2 Input amplifier and 3 f bandpass filter

The input amplifier (U1, OP27) has a frequency range from 800 Hz to 11 kHz . The gain can be switched between 2 dB and 7 dB at 8333.33 Hz . The gain peak is at approximately 3.3 kHz . The gain peak is chosen a little lower than the $3 f$ frequency to give more attenuation for higher frequencies. This is because the detector is more sensitive to error signals high in frequency.

The 3 f bandpass filter is situated on a separate circuit board. This filter consists of a 8333 Hz biquad bandpass filter and a 5555 Hz notch filter. The bandpass filter is of the Tow and Thomas biquad [Ghausi/Laker] type and is relatively insensitive to component drift as investigated by Thomas [Thomas]. The unadjusted center frequency of the filter is $8420 \mathrm{~Hz} \pm 400 \mathrm{~Hz}$ and the Q -value is $40.3 \pm 0.5$. The transfer function of this filter is:

$$
\begin{equation*}
H(s)=-\frac{1}{R_{1} C_{1}} \frac{s}{s^{2}+\frac{1}{R_{3} C_{1}} s+\frac{1}{R_{5} C_{2} R_{2} C_{1} \frac{R_{9}}{R_{7}}}} \tag{3.6}
\end{equation*}
$$



Figure 3.6: The transfer function of the input amp and the 3 f filter

The attenuation of harmonic components relative to the 3 f component by the bandpass filter should be as follows:

| component | attenuation |
| :---: | :---: |
| 2 f | 30.5 dB |
| 3 f | 0 dB |
| 4 f | 27.4 dB |
| 5 f | 32.6 dB |

The notch filter is tunable and is tuned to have a zero in the transfer function at 2 f frequency ( 5555 Hz ). The transfer function is biquadratic:

$$
H(s)=\frac{\left(1+\frac{R_{11}}{R_{10}}\right)\left(C_{7} C_{8} s^{2}+\frac{1}{\left(R_{12}+R_{13}\right)\left(R_{14}+R_{15}\right)}\right)}{C_{7} C_{8} s^{2}+\left[\frac{C_{7}+C_{8}}{R_{14}+R_{15}}-\left(\frac{1}{R_{12}+R_{13}}+\frac{1}{R_{14}+R_{15}}\right) C_{8} \frac{R_{11}}{R_{10}} s+\frac{1}{\left(R_{12}+R_{13}\right)\left(R_{14}+R_{15}\right)}\right.} .
$$

It is assumed that:

$$
\begin{align*}
C_{9} & =C_{7}+C_{8} \quad, \quad \text { and }  \tag{3.8}\\
R_{16}+R_{17} & =\left[\frac{1}{R_{12}+R_{13}}+\frac{1}{R_{14}+R_{15}}\right]^{-1} . \tag{3.9}
\end{align*}
$$

The tuning of the notch filter is done with three trimmers (R12, R14, R17) so that the feedthrough of an applied 5555 Hz signal is minimized. The total transfer function from the lock-in amplifier input to the detector including


Figure 3.7: The group delay of the input amp and the 3 f filter
the input amplifier at low gain and the 3 f filter is shown in figure 3.6. With the input gain at high position 5.5 dB should be added to the gain curve. Figure 3.7 shows the group delay $\left(\tau_{g}\right)$ for the same transfer function. It shows that $3 f(8333.33 \mathrm{~Hz})$ signal components are delayed 1.5 ms in the input circuits before entering the detector. These graphs were plotted with the circuit simulation software APLAC ${ }^{2}$.

There are three reasons not to make the $3 f$ bandpass filter too selective. One restriction comes from the used operational amplifier. The gain $\left(A_{0}\right)$ and dominant pole ( $\omega_{0}$ ) of the op-amp limit the maximum reliable Q -value to:

$$
\begin{equation*}
Q_{\max } \approx 0.025 \cdot A_{0} \cdot\left(\frac{\omega_{p}}{\omega_{0}}-0.5\right)^{-1}, \tag{3.10}
\end{equation*}
$$

where $\omega_{p}$ is the center frequency of the filter [Thomas]. Another restriction comes from the fact that the group delay will rise if the Q-value of the filter is increased. The group delay will account to a linear phase shift $\left(\theta=-\tau_{g} \omega\right)$ added to the control loop phase curve of the lock-in system. This will reduce the phase margin, or worse, cause the system to loose stability. The third restriction is simply the fact that in a very high $Q$ filter the drifting of the components will cause the center frequency to wander off, and thus cause attenuation for the $3 f$ signal.

[^3]
### 3.3.3 Digital 3f reference input

Digital signals in a close proximity to sensitive analog circuits always cause trouble. Because the digital signals are square waves, they contain powerful harmonics. These harmonics high in frequency act as a carrier and help the fundamental frequency to connect and interfere with other circuits, even if separated galvanically. In the lock-in amplifier it is important that no $3 f$-components leak from the reference signal to the input of the detector. This would give rise to an offset in the system. The reference signal could of course be a sine wave to reduce the chance of interference. However, if a sinewave reference is used it has to be regenerated to digital anyway if a switching-type detector is used.

The digital reference input of this unit is isolated from the rest of the electronics by an optocoupler (U2, TIL117). This removes the possibility of digital interference spikes through ground loops but does not remove the overhearing noise. R28 is connected to the base of the phototransistor in the optocoupler and allows trimming of the symmetry of the square wave output. Asymmetric clock signal to the detector will cause some loss of sensitivity and increase the number of harmonics to cause a false detector output. Q1 ( BC 239 B ) transforms the optocoupler output current to a voltage that is fed to a CMOS logic chip (U4, 4070) carrying four XOR gates. The switch K5 that is used to invert (shift phase $180^{\circ}$ ) the clock signal is designed in such a way that it doesn't switch the fast changing logic signal directly, but operates with DC-current only. This avoids long wires carrying interfering digital clock signals to the switch at the front panel. The CMOS chip operates at 8 V regulated from the +15 V used by the analog circuitry. Two clock signals are outputted by the logic circuit to the detector. One is directly the $3 f$ reference, the other is an inverted version.

### 3.3.4 3 f detector

The detector functions as a synchronously switching $\pm 1$ gain amplifier. The N-FET transistors (Q1, Q2, 2N3819) operate as switches, toggling at a rate of the 3 f reference clock. The most critical feature of the detector is the output DC offset. One cause for offset is the leakage of reference clock signal into the detector input signal path. This leakage most easily happens through the FET switching transistors. The designed detector avoids this by using a balanced detector scheme. The charge leakage induced by one of the switching transistors is compensated by a charge leakage of the opposite
sign by the other transistor. The other cause for offset is the operational amplifier. The op-amp used (U5, TLE2027CP) has excellent DC-offset stability. This op-amp also has a very large bandwidth, unlike other amplifiers with good DC characteristics. This is important because a slow amplifier might distort ${ }^{3}$ the signal that is very rich on high frequencies because of switching. As the distortion of the transients might very well be unsymmetric, this would cause unwanted offsets at the output.

Capacitors C1 and C11 are chosen so that they compensate for the different parasitic gate capacitances of the FET transistors and for the different impedances of the balanced amplifier circuit seen from the drain of the transistors. Impedance matching would have been easy by adding two more op-amps, but the number of offset causing active components was tried to keep at a minimum. The difference of the FET transistors' channel resistance might cause asymmetry of the detection. This effect is reduced by the series resistances R7 and R9 with the cost of loosing some detector gain. Likewise the difference in leakage resistance of the FETs is made insignificant with resistors R8 and R10. Trimmers R24 and R25 are used to trim the offsets for both halves of the clock period separately. Trimmer R13 is used to adjust the symmetry balance of the detector. See the service chapter for details of trimming.

The use of a four quadrant analog multiplier as the detector was also considered. The benefit of a detector of this type is the reduced need for wide bandwidth filtering after the detector as no transients are generated. Also, the detector would not be sensitive to $9 \mathrm{f}, 15 \mathrm{f}, 21 \mathrm{f} \ldots$ harmonics in the input signal and a sine wave reference with low interference could be used. However, the disadvantages seemed to be worse, so the idea was abandoned. The disadvantages of an analog multiplier are the unstability of the DC-offset and the harmonic distortion that causes additional offset.

### 3.3.5 Integrator I

Integrator I is the integrator situated right after the $3 f$ detector. The purpose of this integrator is to integrate the output of the detector and thus greatly amplify the DC component in the detector output signal while removing the AC ripple. Theoretically the DC gain of an integrator should be infinite but the used operational amplifier (U1, OP27) has a limited DC gain of approximately $1.5 \cdot 10^{6}$. This is high enough to reduce any externally induced

[^4]

Figure 3.8: The simplified integrator I circuit
static errors in the control loop to a virtually nonexistent level. K5 is a two pole switch that is situated at the front panel. In lock position (open connection) the integrator functions normally keeping the system locked to an absorption line. If the switch is switched into search position (closed connection) the integrator is nulled and the output is zero. This allows manual search for absorption lines with the search wheel on the PZT amplifier unit. The small resistors R27 and R28 reduce the capacitor discharge current and thus reduce the strain on the switch, op-amp and capacitors.

The simplified circuit of the integrator is shown in figure 3.8. The actual integrator is symmetric to reduce the offset drift. $R_{A}$ adjusts the overall gain of the integrator. This control is a 6 position rotary switch. The resistance $R_{B}$ is changed by the time constant rotary switch. There are five different time constants and a test position to choose from. The test position changes the integrator to a simple first order lowpass filter with 3.4 Hz cutoff frequency. This position is not used normally and is provided only for experimental reasons.

The gain is given as the gain of the integrator at 0.1 Hz frequency when the time constant knob is at 0 s position ( $R_{B}=0 \Omega$ ). At this frequency the time constant knob can maximally change the gain by approx. 0.06 dB . Even at 1 Hz the gain change by the time constant knob is 0.5 dB at maximum. The transfer function of the integrator is:

$$
\begin{equation*}
H(s)=-\frac{s R_{B} C_{2}+1}{s R_{A} C_{2}} \tag{3.11}
\end{equation*}
$$

At frequencies much lower than the corner frequency set by the time
constant the gain of the integrator rolls off $-20 \mathrm{~dB} /$ decade and the phase shift is $-90^{\circ}$. The front panel shows approximate gain values only. The following table gives the more exact gain values for the gain knob:

| gain setting | real gain | $R_{A}$ value |
| :---: | :---: | :---: |
| 3 dB | 7.08 dB | $320 \mathrm{k} \Omega$ |
| 10 dB | 14.9 dB | $130 \mathrm{k} \Omega$ |
| 20 dB | 24.1 dB | $44.9 \mathrm{k} \Omega$ |
| 30 dB | 33.8 dB | $14.8 \mathrm{k} \Omega$ |
| 40 dB | 43.8 dB | $4.68 \mathrm{k} \Omega$ |
| 50 dB | 53.7 dB | $1.50 \mathrm{k} \Omega$ |

The time constant determines the frequency corner point above which the integrator starts working as a normal amplifier. The time constant is calculated ${ }^{4}$ as $\tau=R_{B} C_{2}$. The frequency corner point is calculated as $f_{c}=1 /\left(2 \pi R_{B} C_{2}\right)$. The following table lists the approximate time constants noted at the front panel together with the more exact time constants. Also listed are the corner frequency and gain at 8333.33 Hz .

| $\tau$ setting | real $\tau$ | $R_{B}$ | $f_{c}$ | $\mathrm{G}(8333.33 \mathrm{~Hz})$ |
| :---: | :---: | :---: | :---: | :---: |
| 0 ms | 0 ms | $0 \Omega$ | $\infty$ | -98 dB |
| 0.2 ms | 0.22 ms | $99.0 \Omega$ | 730 Hz | -77 dB |
| 1 ms | 0.99 ms | $449 \Omega$ | 160 Hz | -64 dB |
| 2 ms | 2.0 ms | $909 \Omega$ | 80 Hz | -58 dB |
| 10 ms | 11 ms | $5.00 \mathrm{k} \Omega$ | 14 Hz | -43 dB |

To the 3 f gain component mentioned in the table above the gain knob gain value has to be added to get the total gain of the integrator. For instance if we use the maximum gain $(53.7 \mathrm{~dB})$ and biggest time constant ( 11 ms ) the gain at 8333.33 Hz is $53.7 \mathrm{~dB}-43 \mathrm{~dB}=10.7 \mathrm{~dB}$. This means that the annoying 3f ripple present at the detector output will actually be amplified by the integrator instead of being smoothed out.

### 3.3.6 Integrator II

The second integrator is actually not an integrator but a special kind of a lowpass filter that has characteristics that remind an integrator. In a limited frequency range the gain of this filter rolls off by $-10 \mathrm{~dB} /$ decade and the

[^5]phase shift is approximately $-45^{\circ}$. This filter is sometimes called a $1 / 2-$ integrator. It is implemented in this unit by a 4 -pole and 4 -zero filter built around one operational amplifier (U3, OP27). The poles are situated in the frequency range $0.1 \mathrm{~Hz} \ldots 1 \mathrm{kHz}$ one decade apart. The zeros are placed one half decade above the poles in frequency. This cancels the phase shift by the poles while maintaining 10 dB attenuation per pole-zero pair. The DC-gain of the filter is 0.35 dB and the attenuation at 8333.33 Hz is 40 dB . This integrator smooths out the rest of the ripple from the phase sensitive detector that passes thru the first integrator.

### 3.3.7 Indicator circuits

The three indicators in the lock-in amplifier unit are a great help to avoid and handle problem situations. The red overload LED (D107) lights when the detector input signal peak value exceeds about 7.5 V . The positive or negative peak of the detector input signal, whichever is larger, is charged into capacitor C101. This voltage representing the signal peak value is compared to approx. 7.5 V by comparator U103 (LM311N). If the voltage exceeds 7.5 V the comparator drives a current through the LED which then lights. R114 achieves a small hysteresis of about 0.2 V in the comparator. The LED currents are passed thru a resistor (R144) and capacitor (C122) network to reduce interference from the fast on and off switching LEDs to the +15 V operating power.

The $3 f$-detect indicator monitors the buffered detector output signal. If this DC voltage is greater than 0.7 V or less than -0.7 V the comparator U102 will drive the yellow 3f-detect LED. Before the comparator the detector output signal is lowpass filtered with a 16 Hz corner frequency first order filter to remove noise that would blink the LED unnecessarily.

The tracking indicator detects changes in the integrator output. This is done by derivating. The output of the derivator is proportional to the drift speed of the integrator output. This derivator output is lowpass filtered to remove noise and finally connected to a similar comparator as the $3 \mathrm{f}-$ detect indicator had, that turns on the green tracking LED when the absolute voltage is greater than 0.7 V . The LED indicator turns on when the integrator drift speed exeeds approx. $0.6 \mathrm{~V} / \mathrm{s}$ to either direction.

### 3.4 PZT amplifier

The PZT amplifier is a high voltage amplifier that drives the compensation piezo in the laser mechanics. For the end stage a high voltage hybrib amplifier component (U1, U2, PA88) manufactured by Apex Microtechnology was chosen. The choise was motivated by high reliability, safety reasons, low quiescent power consumption, design time savings and unavailability of proper discrete high voltage components. These amplifiers function as any normal operational amplifier with the exception of permissible high operating and output voltages. The input impedance of this hybrid is very high $\left(10^{11} \Omega\right)$, and thus simplifies the designing.

Two of these hybrid amplifiers are used. They each feed a separate output connector. This arrangement allows a balanced configuration to be used where both ends of the piezo are connected to separate amplifiers. The amplifiers are connected in such a way that they output the same voltage but with different polarity. This doubles the control voltage range and the voltage gain compared to using only one amplifier. It is of course also possible to use only one of the amplifier outputs as an unbalanced configuration to control the piezo. The hybrids use a separate power supply that generates a $\pm 200 \mathrm{~V}$ operating voltage for them. This means that the output of the amplifier can swing between -190 V and +190 V approximately. In the unbalanced configuration this is directly the control voltage range over the piezo. If the balanced configuration is used the control voltage range over the piezo is from -380 V to +380 V approximately.

The high voltage components are situated on a separate circuit board from the rest of the PZT amplifier components to increase safety. The gain of the end stage amplifiers is designed to be $-20(26 \mathrm{~dB})$ and they have a pole at 3.4 kHz . The output current is limited to 18 mA . Resistors R7 and R17 limit the current of possible interference spikes that might damage the hybrid units. These resistors also reduce the capacitive loading that might cause the hybrid amplifiers to oscillate. The third function of these resistors is to act as a lowpass filter in conjunction with the cable and piezo capacitance to filter out some noise. The pole of this filter is estimated to be in the range 50 kHz to 1 MHz . Diodes D1...D8 are protection diodes that prevent high voltages to enter the low voltage section of the unit during startup conditions or error situations. The high voltage output is probed by a resistor voltage divider network and fed to the low voltage board in the unit to be monitored. These monitor outputs are also protected against high voltages by zener diodes D9...D12.

The low voltage section of the PZT amplifier unit buffers the input and feeds to the high voltage end stage amplifiers the voltages opposite in polarity. Diodes D1 and D2 provide soft clipping when the voltage approaches it's limits. This reduces strain on the hybrid high voltage amplifiers. This also means that the gain of this unit will decrease at higher voltages and thus the output voltage is not linearly dependent on the input voltage. Linearity is not important for this unit, as it functions inside the control loop of the laser stabilization scheme. Also provided is a comfortable offset adjustment knob that is called the search wheel. This knob is used to search for absorption peaks by altering the laser's frequency. The search knob's adjustment range is set by zener diodes and is a little less than the full adjustment range of the unit. The offset might drift a little but this is not important as the control electronics will compensate for this.

The output monitor and meter are linear, however, and show the correct output voltage. A rotary switch is provided to select monitoring mode. One can monitor the unbalanced voltage at outputl or output2 or the balanced voltage as the difference of these two. The resolution of the voltage meter is 0.1 V in the unbalanced mode and 1 V in the balanced mode. The monitor output connector outputs a voltage that is $1 / 201$ times the actual output voltage.

### 3.5 Temperature regulator

The temperature regulator compares the setting by the set temperature knob to the temperature measured by the temperature sensor in the cooling well and adjusts the peltier element current accordingly. The temperature knob setting is kept stable by a precision voltage reference (U7, LM399H). The temperature setting range is adjusted to $-4^{\circ} \mathrm{C} \ldots+20^{\circ} \mathrm{C}$ by two trimmers (R24, R26). The temperature sensor is a semiconductor based integrated chip (U1, AD590KH) that outputs a current proportional to the temperature over a wide range. The sensitivity of the sensor is $1 \mu \mathrm{~A} / \mathrm{K}$ and the current at $0^{\circ} \mathrm{C}$ is $273 \mu \mathrm{~A}$. Amplifier U 2 changes this current to a voltage that is compared to the voltage set by the set temperature knob. Temperature calibration is performed by offset and gain trimmers ( $\mathrm{R} 2, \mathrm{R} 5$ ) around this amplifier.

Operational amplifier U3 is an integrator that integrates the error between the actual temperature and the temperature setting. Diode D2 prevents the integrator to integrate further down when the peltier element current is zero and the cooling well is warming up. This makes the controller faster.

Transistor Q2 controls the peltier current. The transistor is controlled by the output of the integrator that is buffered by Q3. Trimmer R12 adjusts the maximum current fed thru the peltier element during cooling down. A appropriate cooling current is $4 \mathrm{~A} \ldots 5 \mathrm{~A}$. A meter is provided to monitor the peltier current. Trimmer R16 calibrates this meter. The following table lists the peltier element specifications:

## PELTIER ELEMENT SPECS

| type | PKE 36 A 001 |
| :---: | :---: |
| manufacturer | PELTRON GMBH |
| ohmic resistance | $R_{i} \approx 0.38 \Omega$ |
| thermoelectric force | $\alpha=14 \mathrm{mV} / \mathrm{K}$ |
| maximum voltage | $V_{\max } \approx 4 \mathrm{~V}$ |
| maximum current | $I_{\max } \approx 9 \mathrm{~A}$ |
| max. voltage ripple | $\Delta \mathrm{V} \leq 15 \%$ |

A digital panel meter is provided to show the temperature. A toggle switch toggles between temperature setting value and actual temperature reading from the sensor. The meter's resolution is $0.01^{\circ} \mathrm{C}$ but the accuracy of the control system is $0.1^{\circ} \mathrm{C}$ or better. The most obvious source of the error would be the sensor. As the sensor output is slightly dependent on the operating voltage and the operating voltage is regulated with zener diodes, changes in the unit's ambient temperature or operating current may cause temperature errors. This feature was not improved as the design of this unit was almost directly copied from Jari Koskela [Koskela].

The power supply for the peltier cooler has to be able to deliver the 5 A current needed for cooling. It is also important to keep the ripple in the voltage under $15 \%$ as the peltier element functions badly with unsmooth voltages. A voltage of 5 V to 6 V would be high enough to supply the needed current thru the peltier element and regulation electronics. This would set the power requirement for the transformer to about $30 \mathrm{~W} \ldots 35 \mathrm{~W}$. As the design of this unit is also copied directly from Jari Koskela the voltage of the transformer is unnecessary high ( 12 V ). This gives a lot of extra power to get rid of and results in extreme heating of the regulation semiconductors. In addition, the transformer has to be more powerful and heavier.

There are seven regulator ICs in parallel that perform the regulation. This reduces the loading to max 0.7 A per regulator. Still as the voltage drop has to be high $(4 \mathrm{~V} \ldots 7 \mathrm{~V})$ the power dissipation per regulator is several watts. The voltage output of the supply is adjustable by R10. This voltage is adjusted so that the regulation transistor in the temperature regulation
unit and the seven regulators at the power supply heat up the same amount during the cooling down period of the cooling well. After the well has cooled down the current needed will reduce and the regulation semiconductors will not heat up to extreme temperatures anymore.

### 3.6 Photodetector

The photodetector senses the small intensity changes in the laser beam caused by the iodine absorption cell when the laser frequency is being modulated. The photodetector unit circuit board is situated at the laser unit just behind one of the laser cavity mirrors. For the photodiode (D1) Hamamatsu's S2386 was chosen. The sensitivity for red light, linearity and low noise are the important features of the photodiode. The speed is not important as the bandwidth needed is only approx. 10 kHz . A low noise op-amp (U2, OPA111) functions as a current to voltage converter for the photodiode current. The gain of this stage can be selected by jumper wires (J1, J2, J3). The output of U 2 is lowpass filtered and connected to the DC MONITOR connector at the laser unit. This voltage can be used to monitor the laser beam intensity and should be checked once in a while to make sure the photodetector amplifier (U2) does not saturate.

Next the output of U2 is amplified by a broadband 8333.33 Hz bandpass filter (U3A). The gain is 15 . The output of this filter is monitored at the AC MONITOR connector at the laser unit. The $\mathrm{f}, 2 \mathrm{f}, 3 \mathrm{f}$ and 4 f harmonic components of the modulation frequency can be monitored here. The $f$ component is, however, a little attenuated due to the bandpass filtering. Finally a more narrow ( $\mathrm{Q}=29$ ) bandpass filter with adjustable gain filters out the 8333.33 Hz signal of interest from the noisy detector signal. The gain can be trimmed from 0.5 to 25 . U1 is an op-amp with good output drive capabilities and it functions as a buffer for the output connector.

This unit has not been tested because the new laser unit to be constructed is not ready at this time.

### 3.7 Laser power supply

The laser power supply has to provide a 2 mA to 4 mA current to the laser tube operating at approx. $0.8 \mathrm{kV} \ldots 3 \mathrm{kV}$. It is very hard to find components
that operate safely at these high voltages and they are very expensive. The power supply consists of a line transformer that outputs a 1400 V AC voltage. This voltage is rectified and doubled with high voltage diodes D1 and D2. The DC voltage is charged into capacitors C1...C10 and amounts to about 3940 V . The capacitors are connected in series to reduce the voltage per capacitor to a safe value. The resistors R1...R10 that are connected in parallel with the capacitors divide the voltage evenly over all the capacitors. Next this DC voltage is lowpass filtered to remove ripple by resistors R21...R24 and capacitors C11...C20. The corner frequency of this lowpass filter is 0.06 Hz . Again there are resistors (R11...R20) parrallel with the capacitors to reduce the voltage over the capacitors to a safe level. After the lowpass filter the voltage is between 3 kV and 4 kV depending on the loading current.

The current regulator comes next. The current passes through four regulating transistors (Q1, Q2, Q3, Q4, BUV47A) in series. The transistors are connected in series to reduce the voltage drop over each of them. The resistors (R26...R33) at the base of the transistors divide the voltage evenly over the transistors. Actually this is not exactly true as the base currents of the transistors add up and cause a higher voltage drop over the transistors earlier in the chain. Q5 compares the voltage over R37 and R38 to the voltage over zener diode D5 and adjusts the current thru the resistors at the regulator transistor bases accordingly. The voltage over R37 and R38 is proportional to the laser current and thus the laser current is regulated. By adjusting these resistors the current can be adjusted. One of these resistors is a trimmer and the other is a potentiometer. The trimmer sets the maximum limit for the laser current adjustable by the potentiometer.

After the current regulator there are two RC lowpass stages to filter away some noise and interference. The corner frequency is several hundred Hertz. Just before the output connector there is a voltage doubler that doubles the output voltage to approx. 8 kV if the discharge button is pressed. This feature should only be used in extreme cases when the laser tube won't start discharging because of old age. It works by charging the 4 kV peak to peak voltage of the transformer into C22 and adding this voltage to the output voltage with the help of the high voltage diodes D3 and D4. Capacitor C21 is actually not necessary for this but is included anyway to perhaps give some protection for the high voltage diodes.

The laser current is measured with a panel meter. This meter is connected at the ground end of the power supply. If it were connected to the 4 kV end, problems would occur with isolation and with erroneous readings due to high static voltage fields. The meter is calibrated with trimmer R44.

### 3.8 Equipment case

The equipment case is a standard modular 19 " sub-rack with 6 U high slots. The front end consists of five sections, four modular units and a front panel with power switches and the laser operating current section. The modular units are the modulation frequency generator, lock-in amplifier, PZTamplifier and the temperature controller. These units are all shielded modules except for the frequency generator that is a plain stick-in circuit board with a front panel attached. All the units have the same H 11 backplane connector that provides them with operating power. All connectors deliver $\pm 15 \mathrm{~V}$ operating voltage. The frequency generator card gets additionally +5 V operating voltage for digital circuits, the high voltage amplifier unit gets the $\pm 200 \mathrm{~V}$ operating voltage also needed and the temperature controller is also provided with 5 A peltier current thru these backplane connectors. See the appendix for details of the connections of the voltages to the backplane connectors.

The laser power supply and the power supply for the high voltage amplifier are each situated in separate diecast boxes mounted in the back of the case. The circuit board with the regulators for the $\pm 15 \mathrm{~V}$ and the +5 V operating voltages is secured with screws to the backplane structures also in the back of the case. The same goes for the peltier current regulator board. The toroidal transformers for the $\pm 15 \mathrm{~V},+5 \mathrm{~V}$ and peltier operating voltages are screwed to the wall inside the back of the case. The mains cord socket and the mains fuses are placed at the rear panel of the electronics unit. The red LED indicators above the PZT, LASER and PELTIER power switches are powered by the 6 V transformer of the digital +5 V power supply.

## Chapter 4

## Performance

### 4.1 Modulation frequency generator

The first thing to check with the modulation frequency generator is the purity of the sinusoidal modulation signal. If the modulation signal contains a too high third harmonic component, it might offset the laser's frequency from the center of the absorption line. This is because the third harmonic distortion component present in the modulation frequency will be multiplied by the first derivative of the laser's intensity function and connected to the input of the phase sensitive detector, which then gives an offset at the output. The first derivative of the iodine absorption profile should be zero at the line center, but the slope of the background (laser gain profile) will give a nonzero first derivative. This means that the resulting offset will be opposite in sign on different sides of the laser tube center frequency and can thus be identified.

The frequency spectrum of the modulation signal output was measured first without load. The level of the third harmonic component relative to the fundamental frequency changed with the output level set by the amplitude potentiometer. With small output levels the relative level of the distortion increased. This is explained by the fact that the digital $3 f$ reference signal connects as an interference directly to the modulation output connector. At small output levels the third harmonic component was measured to be more than 8.5 dB down from the fundamental frequency component. At typical output levels the harmonic component was more than 95 dB down. Figure 4.1 shows the measured frequency spectrum of the signal after the biquad bandpass filter at jumper J101. The measured spectrum of the unloaded output


Figure 4.1: Spectrum of the filtered digital modulation waveform
signal with the amplitude set to 1 V is presented in figure 4.2 .
When the modulation piezo was connected to the output of the generator the third harmonic component rised a little. The ratio of the fundamental component to the third component was now measured to be 88 dB with a modulation amplitude corresponding to a 6 MHz modulation of the laser. This increase of the harmonic component could be caused by the capacitive loading of the output amplifier or by some distortion in the modulation piezo. The worst case THD ${ }^{1}$ was measured to be less than $0.02 \%$.

The frequency of the modulation output was measured to be 2777.9 Hz . The range of the amplitude knob was measured to be $0.00 \mathrm{~V} \ldots 3.74 \mathrm{~V}$ which corresponded to a range of $0.00 \mathrm{MHz} \ldots 12.2 \mathrm{MHz}$ modulation (peak-topeak) of the laser frequency. The range of the phase potentiometer was verified to be $+0^{\circ} \ldots+140^{\circ}$. The 3f reference output "high" and "low" logic levels were measured while connected to the reference input in the lock-in amplifier:

[^6]

Figure 4.2: Spectrum of the modulation output

| logic level | voltage |
| :---: | :---: |
| "low" | 0.180 V |
| "high" | 4.40 V |

### 4.2 Lock-in amplifier

The lock-in amplifier is the most complicated module in the electronics unit. It is also the most difficult to measure standalone. In this section only standalone measurements are reported. In a later section the overall perfomance of this unit is discussed also as a part of the whole frequency stabilization electronics.

The gain of the input amplifier was measured at 8333.33 Hz . The gains with the two positions of the input gain switch are listed in the following table:

| input gain | gain |
| :---: | :---: |
| low | -3.8 dB |
| high | +4.2 dB |

These input gains are different from the ones reported in chapter 3 as


Figure 4.3: Frequency response of input amplifier (gain=low)
the attenuation of the input network was not accounted for in the design. The frequency response of the input amplifier was measured also. Figures 4.3 and 4.4 show the frequency response with low and high gain, respectively. The frequency response maximum is below the $3 f(8333.33 \mathrm{~Hz})$ frequency as designed. The measurement was done by averaging noise spectra and therefore the curves look a little noisy.

The $3 f$ filter was measured next. The bandpass center frequency was at $8375 \mathrm{~Hz} \pm 16 \mathrm{~Hz}$. The gain of the filter at 3 f was 14.6 dB . The combined transfer function of the input amplifier at high gain and the $3 f$ filter was measured and is shown in figure 4.5. This curve corresponds very well to the simulated curve in figure 3.6 if one takes into account that the simulation was done with low input gain. The harmonic distortion of the combined input amplifier and 3 f filter was also investigated. The distortion caused by the amplifier and filter themselves was very low but when the $3 f$ digital reference was connected to the unit some interference spikes appeared in the signal to be detected by the phase sensitive detector. The 3 f harmonic component was 61 dB down from the fundamental f frequency at a worst case situation. Because of the arrangement of the input gain switch, the interference from the digital reference is worse with a low input gain setting.

The input overload indicator LED was also tested. It turned on when the signal at the input of the phase sensitive detector was $15 V_{p p}$ or higher.


Figure 4.4: Frequency response of input amplifier (gain=high)


Figure 4.5: Frequency response of input amplifier and 3 -filter


Figure 4.6: Frequency response of integrator I
This was exactly as designed ( $7.5 V_{p}$ ).
The detection gain of the phase sensitive detector is defined here as the ratio between the detector voltage output and the amplitude of the input signal, when the input signal is a sinewave in phase with the reference. When the detector is an amplifier with a gain that alters between -1 and +1 synchronously with the reference. the theoretical detection gain is $2 / \pi \approx 0.637$. The detection gain was measured by applying a sinewave with a frequency slightly different from the reference to the input of the detector. By lowpass filtering the output of the detector a sinewave with a frequency that is the difference of the reference frequency and the input frequency can be discovered. Assuming that the lowpass filter doesn't attenuate this signal, the detector gain can be measured as the ratio of this signal's amplitude to the input sinewave's amplitude. The result of this measurement was a detector gain of $0.62(-4.2 \mathrm{~dB})$. This gain is slightly less than the theoretical gain mentioned above because of the resistor arrangement that reduces sensitivity to variations in the channel resistances of the switching FET transistors (see chapter 3.3.4, page 21 ).

The DC gain of integrator I should be infinite, but in practice it is limited by the DC gain of the op-amp and the parasitic parallel resistance of the integrator capacitor. The DC' gain was not measured, but it is expected to
be better than 100 dB as the op-amp used (OP27) has a gain of 120 dB and the capacitor used is of a good quality. The AC gain of integrator I was measured with a spectrum analyzer. The gain knob was set to 20 dB and the time constant knob was set to 10 ms . The result is shown in figure 4.6. The DC gain shown in the figure is not reliable as the test signal was AC coupled to the integrator and also the frequency span used on the spectrum analyzer was quite high. The absolute value of the AC gain seems to be a little lower than estimated in chapter 3.3.5. This is probably because of a measurement error as there were practical problems with performing the measurement. The important result of this measurement is, however, the verification of the straightness of the gain curve. The gain curve should be straight at frequencies higher than the corner frequency. The corner frequency is 14 Hz in this case with a 10 ms time constant.

### 4.3 PZT amplifier

The PZT amplifier module contains two circuit boards. One board for the high voltage components and another for the low voltage parts. The high voltage board was tested separately before final assembly of the module. The high voltage power supply was designed so that it could deliver a lower voltage for initial test by changing the connection of the cables to the transformer. At this lower operating voltage the basic function of the hybrid amplifiers was checked. This step is important for high voltage electronics because, if there are shorts in the circuit or if the amplifier has a tendency to oscillate, expensive components might be damaged by overheating if full tension is connected.

The full operating voltage was connected to the board next, after the basic functions were checked. This voltage was measured to be $424 \mathrm{~V}( \pm 212 \mathrm{~V})$ with normal operating currents. The operating current of the amplifier was 3.25 mA . The gain of the PZT1 end stage was measured to be -19.88 ( 25.97 dB ). The gain of the PZT2 end stage was measured to be -19.93 $(25.99 \mathrm{~dB})$. The gain in differential mode is the sum of these and is 39.81 $(32.00 \mathrm{~dB})$. The offset voltages of the PZT1 and PZT2 stages were +1.6 mV and -0.5 mV respectively. The total noise voltage over a frequency range from approx. 45 Hz to 20 kHz was measured to be around 2 mV (RMS) for both of the amplifiers.

More tests were performed on the high voltage amplifier module fully assembled. Figure 4.7 shows the measured transfer function of the amplifier


Figure 4.7: Transfer function of the PZT amplifier


Figure 4.8: Noise spectrum of the PZT amplifier
from the input to output 1. The same function for output 2 was measured and it seemed to have an approximately identical transfer function. The noise spectra at the outputs were also measured. Figure 4.8 shows the noise for output 1. Output 2 had an similar graph. The noise of the high voltage amplifier is connected to the tuning piezo of the laser and causes the laser frequency to change randomly. From these noise measurements one could expect the rms noise in the laser frequency to be somewhere in between 1 kHz and 10 kHz if the high voltage piezo sensitivity is about $4 \mathrm{~nm} / \mathrm{V}$. One way to reduce the noise would be to reduce the bandwidth of the output amplifier. This would slow down the stabilization, however, which in turn would make the system more sensitive to outside noise.

### 4.4 Temperature regulator

The maximum current output for the peltier element was designed to be 5 A . This goal was not achieved because the current gain of the output transistor was less than expected. The approx. 4 A now available is high enough, however. The cooling down of the cooling well from room temperature $\left(23^{\circ} \mathrm{C}\right)$ to $15^{\circ} \mathrm{C}$ within a $0.1^{\circ} \mathrm{C}$ stability took 2 minutes and 15 seconds. The temperature of the cooling well will oscillate around the reference value a while just after cooling down until it stabilizes. The time for the temperature to stabilize within $\pm 0.02^{\circ} \mathrm{C}$ from the start of the cooling was about 3 minutes. The cooling current at stable operating conditions was 0.9 A .

The temperature reading at the unit display was calibrated by measuring the cooling well temperature with a calibrated temperature meter. The accuracy should be better than $\pm 0.2^{\circ} \mathrm{C}$ in the range $10^{\circ} \mathrm{C} \ldots 20^{\circ} \mathrm{C}$. For an iodine cell at the recommended $15^{\circ} \mathrm{C}$ temperature this corresponds to a laser frequency uncertainty of $2 \mathrm{kHz}[\mathrm{Hu}]$.

### 4.5 Laser power supply

The laser power supply was not tested with a real laser tube. Tests with artificial loads were performed, however. The maximum current adjustable by the current adjustment potentiometer was set to 5.2 mA . A $400 \mathrm{k} \Omega$ resistor was used as a load and the unit was kept running at 3.5 mA output current for a longer time. The heating of the unit was considered to be within
a safe range. A switching power supply would be more efficient than the configuration used in this power supply, but the risk of switching spikes getting into to laser voltage was considered too big. A linear power supply with the regulation electronics at the primary side of the transformer could also be considered as an alternative solution.

### 4.6 Equipment case

The accuracy of the $+15 \mathrm{~V},-15 \mathrm{~V}$ and the +5 V operating voltages from the backplane connectors is better than $5 \%$. The operating current drain from the backplane connectors of the different modules was measured and is listed in the following table:

| module | +15 V drain | -15 V drain | +5 V drain |
| :---: | :---: | :---: | :---: |
| mod. freq. gen. | 30 mA | 30 mA | 52 mA |
| lock-in amp. | 71 mA | 59 mA | 0 mA |
| PZT amplifier | 78 mA | 28 mA | 0 mA |
| temp. regulator | 85 mA | 36 mA | 0 mA |

The total power consumption of the modules is then 6.5 W . The power loss in the voltage regulators is almost 3 W . The maximum power consumption of the high voltage power supply is about 6 W . The laser power supply consumes about 14 W . Lastly the peltier power supply consumes approx. 16 W when the temperature is stable. Thus, the total power consumption of the electronics unit is 46 W .

### 4.7 Overall performance of frequency stabilization

For a short time the portable $\mathrm{He}-\mathrm{Ne}$ laser built by Jianpei $\mathrm{Hu}[\mathrm{Hu}]$ at HUT was borrowed so that the overall performance of the electronics unit connected to the laser could be tested. All connectors of this electronics unit were designed to be compatible with the connectors of the electronics unit normally used with the laser. Only output 1 of the high voltage amplifier was used as the laser unit didn't have connectors for balanced operation of the tuning piezo.

The first problem occurred with the digital memory xy-oscilloscope. As
the output level of the monitor output at the high voltage amplifier that is connected to the x -channel of the xy -oscilloscope was designed to be very low ( $-1 \mathrm{~V} \ldots+1 \mathrm{~V}$ ), the oscilloscope was very sensitive to high frequency noise in this monitor signal. Part of this noise came from the digital panel meters. This noise was eliminated by connecting a capacitor across the operating power pins of the meter and by adding a RC lowpass filter to the input of the meter (see the engineering change order in the appendix). However, there was still noise present in the oscilloscope graph. This noise was independent of the electronics unit and was induced into the oscilloscope cables from the environment. This noise was eliminated by putting a passive RC lowpass filter with a corner frequency of about 100 Hz at the input of the oscilloscope. The oscilloscope was also sensitive to ground loop problems as the inputs of the oscilloscope were not floating. This resulted in a severe 50 Hz disturbance if the power cords of the oscilloscope and the electronics unit were not plugged into the same mains outlet.

The strength of the absorption lines was investigated next. A modulation of 6 MHz was used and the laser output power was $52 \mu \mathrm{~W}$. The input gain of the lock-in amplifier was set to high and the 3 f filter was turned on. While sweeping the laser frequency over the absorption lines the peak value of the third harmonic component detected by the photodetector was measured. The peak amplitude of the third harmonic at the input of the phase sensitive detector was measured at the signal monitor output to be $\approx 0.45 \mathrm{~V}$. The peak voltage at detector monitor output was found to be $\pm 2.8 \mathrm{~V}$. As the voltage at the detector monitor output is the voltage at the detector output amplified 11 times, the peak voltage at the detector output can be calculated to be $\approx \pm 0.25 \mathrm{~V}$. The detection gain is then $0.25 \mathrm{~V} / 0.45 \mathrm{~V} \approx 0.56$. This corresponds very well to the detection gain measured earlier. The small difference is explained by the inaccuracy of the measurement of the noisy signal at the signal monitor output. The strength of the absorption peaks is highly dependent on the laser intra-cavity power and thus the peak voltage may change if the power drifts. The distance between the positive and negative peaks of the third harmonic at the opposite sides of an absorption line corresponded to a 0.6 V change of the tuning piezo voltage.

From the values measured above the combined gain of the tuning piezo, the laser, the absorption cell, the photodetector, the $3 f$ bandpass filter and the phase sensitive detector (see figure 2.2 on page 5 ) was calculated to be $0.85(-1.4 \mathrm{~dB})$. If the gain of the lock-in amplifier input circuits and the phase sensitive detector is reduced from this value, the gain from the tuning piezo input connector to the photodetector signal output connector at the laser unit is found. The value obtained for this gain from the measurements


Figure 4.9: Simulated bode diagram of the frequency control
is -16 dB . Note that this gain is valid only when the laser is in a locked condition. This gain is useful when calculating the gain and phase margins of the stability analysis of the control system.

Figure 4.9 shows the open loop gain and phase for the frequency stabilization electronics. This figure was simulated before the electronics was built. As the gain of the laser unit was not properly measured before, the gain was assumed to be -25 dB in the simulation. Further the simulation assumed a high input gain in the lock-in amplifier and that the 3 f filter was turned on. Integrator I had 50 dB gain and 0 ms time constant in the simulation. Integrator II was turned on aswell. The differential mode of the high voltage amplifier was included in the simulation. The simulation showed a phase margin of $45^{\circ}$ and a gain margin of 53 dB . As the simulation was done with worst case parameters the result was very satisfactory. The simulation showed that the gain could be increased by as much as 53 dB before the control would be instable. With the new measurements of the gain of the laser unit the gain marigin reduces to 44 dB while the phase margin stays approximately the same. These results are still very good.

As the signal level at the input of the detector was quite low and as
there is a big gain marigin, I would recommend that the gain of the photodetector should be made higher in the laser unit used with this electronic unit compared to the one borrowed from HUT. By increasing the gain before the phase sensitive detector and reducing the gain after the detector (e.g. the gain of integrator I) the laser frequency offset error caused by the detector can be reduced while keeping the total gain of the control loop constant.

The following table shows a summary of the measured gains of the different units in the control loop:

| input amplifier | high gain | +4.2 dB |
| :--- | :---: | ---: |
|  | low gain | -3.8 dB |
| 3f filter | on | +14.6 dB |
|  | off | 0 dB |
| 3f detector |  | -4.2 dB |
| PZT amplifier | single | +26.0 dB |
|  | differential | +32.0 dB |

The spacing between the iodine absorption lines d and g was measured to correspond to a voltage change of 8.4 V over the tuning piezo. This voltage was slightly different for different modes of the laser. This indicates that some nonlinearities exist in the high voltage piezo. The frequency spacing between d and g absorption lines is 39.422 MHz . From this the piezo sensitivity can be calculated to be $3.68 \mathrm{~nm} / \mathrm{V}$. The piezo voltage corresponding to a change from one laser mode to the next was measured to be 60 V in average. The mirror position change corresponding to one mode change of the laser is $\lambda / 2$. This gives a piezo sensitivity of $5.27 \mathrm{~nm} / \mathrm{V}$. The sensitivity mentioned earlier is probably more accurate as the temperature drift of the laser made the measurement of the mode spacing inaccurate. This sensitivity might change as much as $20 \%$ depending on the absolute compression or expansion of the piezo material.

The offset of the phase sensitive detector and integrator I can be trimmed to zero. The offset might change, however, with time and temperature. The offset of this unit was observed and it appeared to stay within $\pm 130 \mu \mathrm{~V}$ over a longer observation time. This corresponds to a accuracy of $\pm 800 \mathrm{~Hz}$ or better of the laser frequency. See the service chapter for details about measuring the offset.

This electronics unit together with the laser L3 built by Hu was compared to another iodine stabilized laser L2 at HUT by beat frequency measurement. See reference $[\mathrm{Hu}]$ for details about the measurement methods.

Laser frequency matrix measurements were made to find out the frequency differences of the lasers. The measurement was repeated with different gain and time constant settings. The results are listed in the following table:

|  | $\tau=0 \mathrm{~ms}$ | $\tau=0.2 \mathrm{~ms}$ | $\tau=2 \mathrm{~ms}$ |
| :---: | :---: | :---: | :---: |
| $\mathrm{G}=3 \mathrm{~dB}$ | $1.4 \mathrm{kHz} \pm 6 \mathrm{kHz}$ | - | - |
| $\mathrm{G}=10 \mathrm{~dB}$ | - | $7.5 \mathrm{kHz} \pm 4 \mathrm{kHz}$ | - |
| $\mathrm{G}=20 \mathrm{~dB}$ | - | $7.0 \mathrm{kHz} \pm 4 \mathrm{kHz}$ | $5.8 \mathrm{kHz} \pm 4 \mathrm{kHz}$ |
| $\mathrm{G}=30 \mathrm{~dB}$ | $5.1 \mathrm{kHz} \pm 3 \mathrm{kHz}$ | $5.9 \mathrm{kHz} \pm 3 \mathrm{kHz}$ | - |
| $\mathrm{G}=40 \mathrm{~dB}$ | - | $7.1 \mathrm{kHz} \pm 5 \mathrm{kHz}$ | - |

The detailed results of the matrix measurement with 30 dB gain and 0.2 ms time constant can be found in the appendix. The frequency difference between the two lasers was between 5 kHz and 8 kHz . This result agrees with results of previous comparisons between the two lasers. The result of the measurement with 3 dB gain is different from the others, probably because of temperature drift of the laser cavity. When low gain is used the compensation for temperature changes is reduced. Measurements with high gain and a big time constant was not performed as the attenuation of the $3 f$ ripple output from the phase sensitive detector was not big enough and the xy-oscilloscope used to monitor the absorption lines showed a sloping background under the third derivative curves. A passive LC $3 f$ notch filter was added to the output of the phase sensitive detector after the measurements (see the engineering change order in the appendix). This should allow higher gains and time constants to be used. Unfortunately there was not time to make measurements with the new filter for this report.

The short time stability of the laser system was measured next. This measurement was made again by beat frequency measurement between lasers L2 and L3 at HUT. The problem with this kind of measurement is that there is no simple way to tell which one of the lasers is unstable. The Allan variance of 20 frequency measurements with 10 s gatetime was calculated with different gain and time constant settings. The Allan variances are listed in the following table:

|  | $\tau=0 \mathrm{~ms}$ | $\tau=0.2 \mathrm{~ms}$ | $\tau=1 \mathrm{~ms}$ | $\tau=2 \mathrm{~ms}$ | $\tau=10 \mathrm{~ms}$ |
| :---: | :---: | :---: | :---: | :---: | :---: |
| $\mathrm{G}=3 \mathrm{~dB}$ | - | 6.9 kHz | - | - | - |
| $\mathrm{G}=10 \mathrm{~dB}$ | - | 4.2 kHz | - | - | - |
| $\mathrm{G}=20 \mathrm{~dB}$ | 6.7 kHz | 4.7 kHz | 4.3 kHz | 6.9 kHz | 5.1 kHz |
| $\mathrm{G}=30 \mathrm{~dB}$ | - | 5.4 kHz | - | - | - |
| $\mathrm{G}=40 \mathrm{~dB}$ | - | 4.3 kHz | - | - | - |
| $\mathrm{G}=50 \mathrm{~dB}$ | - | 5.4 kHz | - | - | - |

No big differences in the Allan variances with different settings are spotted. This was not expected either for different time constants as the time constant knob affects only frequencies above 10 Hz , and these frequencies are effectively averaged out by the 10 s gatetime of the frequency counter. By changing the gain, the speed of the compensation for errors induced in the laser frequency is varied and this should generally give different Allan variance measurement results. Maybe the other laser system (L2) is the major cause of instability in the Allan variance measurement above when no significant change of the Allan variance with different gain settings is seen.

As a last measurement a long overnight measurement of the beat frequency stability was measured. The Allan variance of the measurement was calculated. The stability optimum is obtained with a gate time of 2000 s . The Allan variance of 20 frequency samples with this gate time was 332 Hz . The relative Allan variance is then $7 \cdot 10^{-13}$. Figure 4.10 shows the graph of the Allan variance measurement. The same measurement data can also be found in numeric format in the appendix.


Figure 4.10: Allan variance measurement result

## Chapter 5

## Using the equipment

### 5.1 Cable connections

Most of the cables used with this unit are RG-58 coaxial cables with BNC connectors at the ends. The laser current cable and the high voltage piezo cables are RG-59 cables because of the slightly higher voltage strength ( 1.7 kV ). The high voltage piezo cable has MHV connectors (Suhner) at the ends while the laser current cable has SHV connectors (Suhner). The outer conductors (the shield) of the coaxial cables are all connected to ground. The electronics units internal ground is connected to the chassis at the high voltage amplifiers output connectors. The case is grounded via the power line cord. The SIGNAL INPUT connector at the lock-in amplifier is a 6 -pole Lemo connector. The TEMPERATURE SENSOR connector at the temperature regulation unit is a 4 -pole Lemo connector. The PELTIER CURRENT OUTPUT connector is a 4 -pole current connector. The pin connection of these multipole connectors can be found in the appendix. The unit is designed so that any cables except for the high voltage cables can be connected or disconnected with the power switched on.

THE HIGH VOLTAGE CABLES HAVE TO BE HANDLED WITH CAUTION. THERE MIGHT BE HIGH VOLTAGE CHARGES PRESENT IN THE EQUIPMENT AND THE CABLES A LONG TIME AFTER THE POWER HAS BEEN SHUT DOWN AND THE CABLES BEEN DISCONNECTED.


Figure 5.1: The front panel of the electronics unit


Figure 5.2: The cable connections
Figure 5.1 shows the front panel of the electronics unit and figure 5.2 shows the cable connections between the electronics unit, the laser unit and the oscilloscope. The first two things to connect are the $3 f$ REFERENCE output from the modulation frequency generator to the 3 R REFERENCE INPUT at the lock-in amplifier and the CONTROL OUTPUT from the lock-in amplifier to the INPUT at the PZT amplifier. These connections should be made with as short cables as possible because the outputs and inputs are not designed to drive long transmission lines. A cable length of less than 30 cm is recommended.

The $f$ MODULATION output of the modulation frequency generator is connected to the modulation piezo of the laser. The SIGNAL INPUT at the lock-in amplifier is connected to the photodetector connector at the laser. The operating power for the photodetector is provided thru this connector and the detector output signal is returned to the lock-in amplifier via the
same cable. If differential operation of the high voltage (tuning) piezo is used, the outputs at the PZT amplifier are each connected to one end of the high voltage piezo so that the voltage over the piezo is the difference of these output voltages. If single output operation is used, the connection is made in such a way that the output voltage is directly connected over the piezo. The connectors at the temperature regulation unit are both connected to the corresponding connectors at the laser with appropriate cables. Likewise the LASER POWER output from the electronics unit is connected to the laser with a special cable.

An xy-oscilloscope is used as an absorption line monitor. A digital or analog memory oscilloscope is recommended but a normal oscilloscope can also be used. The bandwidth required is very low ( $\leq 100 \mathrm{~Hz}$ ). If the oscilloscope has a large bandwidth some high frequency noise might disturb the reading especially with memory oscilloscopes. If this is the case, it is recommended to reduce the bandwidth by a (passive) lowpass filter near the input connector at the oscilloscope. The oscilloscope's x-channel is conneted to the MONITOR OUTPUT at the PZT amplifier. This sweeps the trace in horizontal direction when the frequency of the laser is tuned. The DETECTOR MONITOR OUTPUT at the lock-in amplifier is connected to the $y$-channel of the oscilloscope. Now a figure of the third derivate of the laser intensity is drawn on the oscilloscope screen when the laser frequency is tuned with the search wheel on the PZT amplifier.

### 5.2 Temperature regulation

The MAINS power switch must be switched on, of course, for any operation with the unit. The iodine cooling well temperature desired (typically $15.0^{\circ} \mathrm{C}$ ) is set by the SET TEMPERATURE knob. The setting can be viewed from the digital display when the toggle switch beneath the display is set to REFERENCE position. The actual temperature of the cooling well can be viewed from the same display when the switch is set to IODINE CELL position. The peltier element cooling current has to be swithed on with the PELTIER switch underneath the MAINS power switch. The highest temperature displayable by the temperature display is $19.99^{\circ} \mathrm{C}$. If the temperature is higher than this, the display will show an overflow indication typical for 7 -segment displays.

### 5.3 Laser power supply

The laser power supply is turned on with the LASER switch under the MAINS power switch. It takes some time before the voltage rises high enough to start the laser tube discharge. If the tube hasn't started discharging in about five minutes the connections should be checked again. If the connections are correct the DISCHARGE START button can be pressed. Don't hold the button pressed down for a too long time as it can damage the cables or laser tube. The discharge button can deliver as high voltage as 8 kV to the laser power output connector. Note that the cable used (RG-59) is rated for only 1.7 kV operation.

### 5.4 Locking to an absorption line

Assuming that all the connections and operations explained earlier are made, the search for absorption lines can begin. First make sure that the AMPLITUDE potentiometer of the modulation frequency generator is not at zero position. Next set the toggle switches on the lock-in amplifier in the following positions:

| INPUT GAIN | $H$ HIGH |
| :---: | :---: |
| 3f FILTER | $O N$ |
| REF. PHASE | $0^{\circ}$ |
| INTEGR. I | SEARCH |
| INTEGR. II | $O N$ |

If the input OVERLOAD indicator lights set the INPUT GAIN switch to LOII: Turn on the high voltage power supply needed for the tuning piezo next with the PZT switch under the MAINS power switch. Now the frequency of the laser can be tuned by turning the SEARCH wheel on the PZT amplifier unit. Adjust the position and volts/div knobs on the oscilloscope so that the trace becomes visible. $1 \mathrm{~V} / \mathrm{div}$ on the y -channel and $10 \mathrm{mV} / \mathrm{div}$ on the $x$-channel are good values of the oscilloscope sensitivity to start with. Now a trace like the third derivate curve in figure 2.1 on page 4 should be obtained. If there are many overlapping peaks in the trace the laser is probably operating with multiple modes of oscillations. The absorption peaks should then be searched at another position of the search wheel. An appropriate search range is -60 V to +60 V of the PZT amplifier output voltage. Select
the PZT amplifier meter operating mode with the rotary switch according to the connection of the tuning piezo. If the multimode operation of the laser continues over the whole search range the power of the laser could be reduced to improve the situation. If the absorption peaks are weak the phase adjustment knobs of the modulation frequency generator should be turned so that maximum peak heights are obtained. If the absorption lines are still weak the amplification of the photodetector or the power of the laser could be increased.

If the absorption peaks are found we can try to lock the laser to one of them. This is done by tuning the laser frequency close to the center of the absorption line and switching the INTEGR. I toggle switch at the lock-in amplifier to the LOCK position. The oscilloscope beam should be on the slope between the negative and positiove peaks of the trace before switching into lock position to ensure locking. If the beam on the oscilloscope jumps away from the slope when the INTEGR. I toggle switch is switched into LOCK position, the gain of the control loop should be inverted. The inversion of the gain can be done by changing the phase $180^{\circ}$ with the rotary switch on the modulation frequency generator or with the REF. PHASE switch on the lock-in amplifier. The gain can also be inverted by swapping the cables between the outputs of the PZT amplifier.

The GAIN and TIME CONSTANT knobs of the lock-in amplifier can be adjusted at any time. The gain should be high enough to make the compensation of cavity temperature drift fast. A too high gain, however, could make the stabilization unstable and oscillation might occur. By increasing the time constant the stabilization of high frequencies can be improved. The stability with high gains can also be improved with a bigger time constant. The disadvantage of a high time constant, however, is the reduced attenuation of the ripple from the phase sensitive detector.

When the laser is successfully locked, none of the indicators should light. If the input OVERLOAD indicator lights, the input signal level is too high and should be reduced. The 3 f DETECT indicator lights if a nonzero third harmonic signal is detected. As the locking is performed to the zero of the third derivative of the intensity function, this indicator should not light when the laser is locked to an absorption line. The TRACKING indicator lights if the output of integrator I changes. It lights even for fast vibrations that can't be viewed from the output meter. In a stable lock condition this indicator won't light. With high gain and time constant settings the ripple from the phase sensitive detector can pass thru the integrator and light the TRACKING indicator. Integrator II smooths out the ripple after the first integrator so
the laser could be operated even with the TRACKING indicator lighting, but be warned.

## Chapter 6

## Servicing the equipment

### 6.1 General service

The most probable parts to cause failure in the electronics unit are the connectors. The next probable are the power components. If some of the plug-in modules cease to function, the first thing to check is the operating voltage from the backplane connectors. The pin configuration of the backplane connectors can be found in the appendix. Note that the high voltage at the PZT amplifier backplane connector will remain dangerously high a long time after the power has been shut down.

There are three fuses on the back panel of the unit. One fuse is for the backplane power supplies, one is for the peltier element power supply and the last one is for both the laser power supply and the high voltage (PZT) power supply. There are also fuses inside the unit. Two fuses are inside the high voltage power supply and three fuses can be found on the backplane power supply card. Only fuses of the correct rating should be used to ensure fire safety.

### 6.2 Modulation frequency generator

This unit contains no trimmers. The components in this unit are very reliable so no frequent testing should be neccesary. One thing to test for is the frequency output at the $f$ and 3 outputs, as the aging of the crystal could
cause a frequency drift. The aging of the crystal is slow, however. The frequency at the f output should be $2777 \mathrm{~Hz} \pm 5 \mathrm{~Hz}$ and $8333 \mathrm{~Hz} \pm 15 \mathrm{~Hz}$ at the 3 f output. Another thing to check for more frequently is the distortion of the modulation output signal. It should be checked both with and without load. The third harmonic component should be more than 70 dB down from the fundamental frequency component to ensure reliable operation.

### 6.3 Lock-in amplifier

There are quite a few trimmers in the lock-in amplifier to adjust. Not all of them are very critical, however. On the main board there are two trimmers. R28 trims the symmetry of the 3 f clock signal. This trimming is not very critical. It can be trimmed by connecting the $3 f$ reference from the modulation frequency generator to the lock-in amplifier and by looking at the signal at pins 10 or 11 of U4 (4070). The trimmer should be adjusted so that one square wave period has an equally long "high" and "low" period. The other trimmer (R22) calibrates the output meter. This adjustment is not very important either. It can be calibrated by locking the laser to an absorption line first. Then the search wheel is turned until the output voltage of the lock-in amplifier is 10 V measured by an external voltmeter. Finally the trimmer is trimmed until the lock-in amplifier meter shows the same as the external meter.

To get to the trimmers on the plug-in units, the panel meter has to be removed. The $3 f$ bandpass filter (lock-in amp. filter bank) plug-in unit has three trimmers. These trim the 5555 Hz notch filter center frequency. To trim these, connect a pure 5555.6 Hz sinewave to the input of the lockin amplifier. Now trim the trimmers until the output of the filter (at the SIGNAL MONITOR OUTPUT connector) is minimized. The notch filter adjustment is not very critical as the phase sensitive detector is not sensitive to $2 f$ frequencies, nor to the harmonics of it.

The integrator has one trimmer. This is the offset trimmer. To adjust the integrator offset remove the detector (with the power shut down) and short the output pins (pins 7 and 8 ) of the detector connector at the main board. Turn off integrator II with the toggle switch on the front panel. Now trim the offset trimmer while monitoring the voltage at the CONTROL OUTPUT connector with an oscilloscope or a sensitive voltmeter. The adjustment should be done with the INTEGR. I switch in LOCK position. If the output
voltage drifts far from zero, reset the integrator with the INTEGR. I switch. Adjust the trimmer so that the output voltage stays constant or changes randomly around a constant average voltage. The offset is slightly different with different gain settings of the integrator. Adjusting the offset with the gain at maximum ( 50 dB ) gives the best overall result, however. The offset of the integrator is generally only needed to bee adjusted once. Replace the detector plug-in unit after the adjustment has been done.

The detector plug-in unit has three trimmers. First make a rough adjustment by trimming the offset without the $3 f$ reference connected to the lock-in amplifier. Keep the SIGNAL INPUT of the lock-in amplifier connected to the photodetector in the laser unit but disconnect the cable to the modulation piezo. Have the INPUT GAIN and 3 FILTER switches in the normal operating position (e.g. HIGH and ON respectively). The INTEGR. II switch should be in OFF position. Monitor the CONTROL OUTPUT voltage the same way as when adjusting the integrator offset. Again the INTEGR. I switch can be used to reset the integrator if the output voltage drifts too far. First adjust trimmer R25 so that the output voltage stays constant or changes randomly around a constant average voltage when the REF. PHASE switch is in $0^{\circ}$ position. Then switch REF. PHASE into $180^{\circ}$ position and do the same adjustment with trimmer R24. The adjustments should be done with a high gain setting ( 40 dB or 50 dB ) and a zero time constant of the integrator.

Now reconnect the 3 f reference to the lock-in amplifier. Adjust R25 until the output voltage stays constant again and then adjust it approximately halfway back. Then adjust R24 until the output voltage stays constant once again. Finally adjust the balance of the detector by applying a voltage (e.g. $+15 \mathrm{~V}^{\circ}$ ) to the leg of C2 that is connected to trimmer R13 and trim R13 until the output voltage stays constant.

Small corrections to the offset of the whole system can be made by simply trimming R24 or R25 without the elaborate procedure explained above. The locations of the offset trimmers are shown in the drawing in the appendix.

The offset of the lock-in amplifier can be checked at any time by removing the modulation from the laser and turning off integrator II. If there is an offset the output meter will start moving when the INTEGR. I switch is put in the LOCK position. The integrator gain should be set high to see this. Just aiter the unit has been turned on there will be a bigger offset noticeable in this way. This will go away after a couple of minutes when the unit has "warmed up". The offset voltage at the detector output can be calculated from the time it takes for the output voltage to drift a certain voltage. If the
output voltage drifts $\Delta V_{\text {out }}$ volts in a time of $\Delta t$ seconds, the offset voltage $V_{o f f \text { set }}$ is:

$$
\begin{equation*}
V_{o f f \text { set }}=R_{A} C_{2} \cdot \frac{\Delta V_{\text {out }}}{\Delta t} \tag{6.1}
\end{equation*}
$$

where $R_{A}$ and $C_{2}$ are the integrator components shown in figure 3.8 on page 23. With an integrator gain of $50 \mathrm{~dB}, R_{A}$ is $1.50 \mathrm{k} \Omega$ and $C_{2}$ is $2.2 \mu \mathrm{~F}$. From the offset voltage the laser frequency offset $\Delta f$ can be estimated as follows:

$$
\begin{equation*}
\Delta f \approx \frac{V_{\text {off set }}}{V_{\text {peak }}} \cdot \beta / 2, \tag{6.2}
\end{equation*}
$$

where $V_{\text {peak }}$ is the peak voltage of the detector output when sweeping the laser over an absorption peak and $\beta$ is the frequency spacing between the positive and negative peaks of the detector output. The value of $\beta$ was measured to be $\beta \approx 2.2 \mathrm{MHz}$. This is quite close to the iodine absorption natural linewidth ${ }^{1} \gamma \approx 2.4 \mathrm{MHz}$. $V_{\text {peak }}$ can be calculated by dividing the peak voltage at the detector monitor output connector at the lock-in amplifier by 11. To reduce the effect of the offset voltage on the laser frequency, the peak voltage at the detector output should be maximized by increasing the gain of the photodetector. Note that if there is an offset and the strength of the absorption peak changes, the laser's frequency will shift. This effect could explain some of the frequency shifts caused by power shift, modulation amlitude shift and iodine cell temperature shift.

The offset checking method described in reference [Shotton/Rowley] on pages 35-37 can be performed with this electronics unit. If the REF. PHASE switch on the lock-in amplifier is toggled and the cables at the outputs of the PZT amplifier are swapped, the operation of the electronics unit is the same with the exception that the sign of the offset is reversed. If the laser frequency is compared to another laser by beat frequency measurement both ways, the difference in the measurement result will be twice the offset of the electronics. The problem with this measurement is that the measurement results are usually quite noisy and the iodine cell and the laser itself might drift. These uncertainties are probably greater than the offset of the electronics and thus the offset of the electronics cannot be reliably determined by this method.

The possible offset caused by the third harmonic in the modulation frequency or nonlinearities in the modulation piezo or photodetector can be reduced by a simple adjustment. Instead of removing the modulation from

[^7]the laser when calibrating the detector offset as explained earlier, keep it connected. Tune the laser frequency in between the absorption lines (e.g. lines $h$ and $g$ ) with the search wheel. The cable from the control output of the lock-in amplifier to the PZT amplifier has to be removed. If the phase sensitive detector offset is adjusted to zero now, the total offset will be eliminated. This adjustment is not valid for all frequencies of the laser as the offset caused by modulation frequency distortion will change with the slope of the laser gain curve.

### 6.4 Photodetector

The photodetector has three jumper wires which select the gain of the current to voltage converter. The gain should be set as high as possible without causing the amplifier output to saturate. An appropriate DC output voltage is approx. 10 V or less. The beam intensity at the photodiode can be reduced by optical methods if the amplifier saturates even with the lowest gain. Trimmer R4 adjusts the DC offset of the current to voltage converter. If the DC monitor output is used as an absolute intensity monitor the offset can be calibrated with this trimmer.

Trimmer R11 adjusts the AC gain of the photodetector amplifier. The gain is adjusted so that the peak voltage at the detector monitor output connector of the lock-in amplifier is approx. $8 \mathrm{~V} \ldots 10 \mathrm{~V}$. The input gain of the lock-in amplifier should be high and the $3 f$ filter should be on. If the overload indicator on the lock-in amplifier lights, however, the gain must be reduced.

### 6.5 PZT amplifier

The trimmers on the PZT amplifier board calibrate the voltage monitoring function of the unit. The calibration is performed by measuring the outputs with an accurate voltmeter and adjusting the trimmers so that the built in digital display shows the same value. Leave the input connector unconnected when doing the calibration. First adjust output 1 to zero with the search wheel. Trim R27 so that the display shows zero when the display rotary switch is set to output 1. Ajust the output 2 to zero next and trim R29 so that the display shows zero when the display rotary switch is set to output 2 . Next
adjust the search wheel so that the voltage between the outputs is zero. Now trim R33 until the display also shows zero when it is set to differential mode. Now adjust a relatively high voltage to the output 1 and trim the display to the correct value with trimmer R36 when the display swith is in output 1 position. Do the same calibration in differential mode with trimmer R35. In differential mode the display should show the voltage difference between the two ouṭputs.

### 6.6 Temperature regulator

The temperature regulator has many trimmers. R24 and R26 adjust the range settable by the temperature setting potentiometer. The simplest way to adjust them is to adjust the voltages at the ends of the potentiometer as shown in the schematic diagram. R16 calibrates the current reading on the panel meter. This is done by feeding a known current thru the resistors R14 and R15 and adjusting the meter to show the correct value. R12 can be trimmed during the cooling down of the cooling well to adjust the maximum cooling current. The output voltage of the peltier power supply should be adjusted so that the current regulating transistor heats up approximately as much as the power supply regulators themself.

The temperature sensor reading is calibrated with trimmers R2 and R5. Place the temperature sensor in a known temperature $T_{1}$ (e.g. $10^{\circ} \mathrm{C}$ ) and write down the temperature reading $X_{1}$ from the temperature display of the temperature controller unit. Now place the temperature sensor in another known temperature $T_{2}\left(\right.$ e.g. $\left.18^{\circ} \mathrm{C}\right)$ and write down the new temperature reading $X_{2}$. Now calculate the following coefficients:

$$
\begin{align*}
G & =\frac{X_{1}-X_{2}}{T_{1}-T_{2}}  \tag{6.3}\\
S & =X_{1}-G \cdot T_{1} \tag{6.4}
\end{align*}
$$

Assuming that the sensor is still at the known temperature $T_{2}$, trim R5 so that the display reads $T_{2}+S$. Finally trim R 2 so that the display shows the correct temperature $T_{2}$. Check the reading with a few known temperatures of the sensor and repeat the calibration procedure if necessary.

### 6.7 Laser power supply

The laser power supply is not a very reliable device because of it's high voltage. With trimmer R44 the current meter can be calibrated. The meter was initially calibrated by measuring the current thru an artificial load in place of the laser tube and adjusting the reading of the meter to show the same value. The maximum current was adjusted with R37 with an operating voltage that was a quarter of the full voltage (for safety reasons).

## Chapter 7

## Conclusions

Even if the electronics unit built in this work could not be tested with a laser specifically adjusted to meet it's needs, the measurement results were very good. The accuracy goal of $\pm 1 \mathrm{kHz}$ was reached. The comfort of using the equipment was also very satisfying. Things that could be improved still are the peltier cooling current regulation and the laser power supply. The power losses of the peltier current regulator could be easily reduced by lowering the operating voltage. The laser power supply reliability could be increased by reducing the amount of components on the high voltage side of the transformer.

The frequency stabilization electronics could also be improved. Digital signal processing techniques could be used to detect and integrate the third harmonic locking signal. This would practically eliminate offsets caused by the electronics and improve stabilization speed. The faster stabilization control would make the laser system more insensitive to temperature or other drift factors. Also, automatization of the locking would be easy to implement. Automatic locking would enable long time frequency matrix measurements without human assistance. As digital hardware has become very cheap, the cost of implementing a digital laser control unit is comparable to an analog one. The analog electronics is, however, already more accurate than the laser and iodine absorption cell. This situation might change in the future.

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## Appendix

A Backplane connector voltages
B Multipole connector pin configurations
C Lock-in amplifier plug in units
D Drawing of offset trimmers in the lock-in amplifier
E Laser frequency matrix measurement report
F Allan variance measurement report
G Matlab script for absorption line simulation
H List of schematic diagrams
I Schematic diagrams
J Engineering change order (ECO)
K Part Lists
L Printed circuit board diagrams

## BACKPLANE CONNECTOR VOLTAGES



H11 backplane connector (almost DIN 41612)

## MULTIPOLE CONNECTOR PIN CONFIGURATIONS

## TEMPERATURE SENSOR CONNECTOR



$$
1 \text { SENSOR - }
$$

2 SENSOR +

## PHOTODETECTOR CONNECTOR



1 GND
2-15 V
3 DETECTOR SIGNAL 4 SIGNAL GND 5 NO CONNECTION $6+15 \mathrm{~V}$

## PELTIER CURRENT CONNECTOR



1I_PELT 21 PPELT + 3 I_PELT + 41 _PELT.

JACKS, FRONT VIEW

## LOCK-IN AMPLIFIER PLUG IN UNITS



| conn. | pin | integrator | detector | 3f filter |
| :--- | :--- | :--- | :--- | :--- |
| J 1 | 1 | INPUT + | INPUT | INPUT |
| J 1 | 2 | INPUT + | INPUT | INPUT |
| J 1 | 3 | INPUT - | GND (IN) | GND |
| J 1 | 4 | INPUT - | GND (IN) | GND |
| J 1 | 5 | GND (OUT) | GND (OUT) | GND |
| J 1 | 6 | GND (OUT) | GND (OUT) | GND |
| J 1 | 7 | OUTPUT | OUTPUT | OUTPUT |
| J 1 | 8 | OUTPUT | GND (OUT) | OUTPUT |
| J2 | 1 | VCC | +8V | VCC |
| J2 | 2 | VCC | VCC | VCC |
| J2 | 3 | GND | GND | GND |
| J2 | 4 | GND | GND | GND |
| J2 | 5 | VEE | VEE | VEE |
| J2 | 6 | VEE | VEE | VEE |
| J2 | 7 |  | CLK |  |
| J2 | 8 |  | CLK |  |

## LOCK-IN AMPLIFIER OFFSET TRIMMERS



Laser frequency difference measurement
26.10.1993 Tue at 15:05:43

KAU test 6 AIN $=30 \mathrm{~dB}, \mathrm{t}=0.2 \mathrm{us}$

| LASER1 L2 |  |
| ---: | :--- |
| Pover output | $=35 \mathrm{uW}$ |
| Cell temperature | $=15.0 \mathrm{C}$ |
| Modulation ampl. | $=6 \mathrm{MHz}$ |
| LASER2 L3 (tkau) |  |
| Pover output | $=52 \mathrm{uW}$ |
| Cell temperature | $=15.0 \mathrm{C}$ |
| Modulation ampl. | $=6 \mathrm{MHz}$ |

LASER FREQUENCY DIFFERENCE MATRIX:


Frequency differences (Laser2-Laser1) :
$(g, f)=12.28557166 \mathrm{kHz}$
$(\mathrm{g}, \mathrm{e})=4.494953323 \mathrm{kHz}$
$(\mathrm{g}, \mathrm{d})=-3.478573334 \mathrm{kHz}$
$(f, e)=14.76043500 \mathrm{kHz}$
$(f, d)=11.03516333 \mathrm{kHz}$
$(\mathrm{e}, \mathrm{d})=-3.547395836 \mathrm{kHz}$
Average frequency difference $=5.9250257 \mathrm{kHz}+/-3 \mathrm{kHz}$
Measurevent standard deviation $=4.97 \mathrm{kHz}$
Component difference std. dev. $=8.06 \mathrm{kHz}$

Differences from recomended values

|  | Measured | Standard | Difference |
| :---: | :---: | :---: | :---: |
| e1-g2 $=$ | 26.552 MHz | 26.561 MHz | -8.8979 kHz |
| e1-f2 $=$ | 13.354 MHz | 13.363 MHz | -9.2157 kHz |
| e1-d2 $=$ | 12.862 MHz | 12.861 MHz | 783.38 Hz |
| e2-g1 $=$ | 26.561 MHz | 26.561 MHz | 92.053 Hz |
| e2-f1 $=$ | 13.383 MHz | 13.363 MHz | 20.305 kHz |
| e2-d1= | 12.869 MHz | 12.861 MHz | 7.8782 kHz |

ALLAN VARIIANCE MEASUREMENT
L3 ( $+K$ KAU) ( g ) and L2 (d)
27.10.1993 Wen at 08:1E:15

| Timeperiod | Allan Variance | Std. Deviation | Samples |
| ---: | :---: | :---: | :---: | :---: |
| 0.15 | 48.25916 kHz | 47.42876 kHz | 1000 |
| 0.25 | 33.47167 kHz | 32.79228 kHz | 500 |
| 0.45 | 21.89555 kHz | 22.32659 kHz | 250 |
| 15 | 15.67252 kHz | 15.77221 kHz | 1000 |
| 25 | 11.02770 kHz | 10.96048 kHz | 500 |
| 45 | 7.611860 kHz | 7.674849 kHz | 250 |
| 105 | 5.094670 kHz | 5.217011 kHz | 4000 |
| 205 | 3.627268 kHz | 3.780401 kHz | 2000 |
| 405 | 2.544117 kHz | 2.748953 kHz | 1000 |
| 1005 | 1.611832 kHz | 1.925175 kHz | 400 |
| 2005 | 1.104688 kHz | 1.563882 kHz | 200 |
| 4005 | 847.0491 Hz | 1.345337 kHz | 100 |
| 10005 | 445.2476 Hz | 1.160364 kHz | 40 |
| 20005 | 331.9622 Hz | 1.127608 kHz | 20 |
| 40005 | 373.5608 Hz | 1.135494 kHz | 10 |
| 100005 | 649.9835 Hz | 1.168728 kHz | 4 |

Mean frequency $=39.423377212 \mathrm{MHz}$
Max frequency $=39.581219000 \mathrm{MHz}$
Min frequency $=39.271252000 \mathrm{MHz}$

ALLAN VARIANCE MEASUREMENT
L3 (+KAU) (G) and L2 (d)
27.10.1593 Wen at 08:17:50

| Timeperiod | Allan Variance | Std. Deviatign |  | Samples |
| :---: | :---: | :---: | :---: | :---: |
| 0.15 | 1.02028E-0010 | 47.42876 | kHz | 1000 |
| 0.25 | $7.07646 \mathrm{E}-0011$ | 32.75228 | kHz | 500 |
| 0.45 | 4.62908E-0011 | 22.32659 | kHz | 250 |
| 15 | 3.31343E-0011 | 15.77221 | kHz | 1000 |
| 25 | $2.33144 \mathrm{E}-0011$ | 10.96048 | kHz | 500 |
| 45 | 1. E0927E-0011 | 7.674849 | kHz | 250 |
| 10 s | $1.07710 \mathrm{E}-0011$ | 5. 217011 | kHz | 4000 |
| 205 | 7.EE8E4E-0012 | 3.780401 | kHz | 2000 |
| 405 | 5.37868E-0012 | 2.748953 | kHz | 1000 |
| 1005 | 3.40768E-0012 | 1.925175 | kHz | 400 |
| 2005 | $2.33549 E-0012$ | 1.563882 | kHz | 200 |
| 4005 | 1.79080E-0012 | 1.345337 | kHz | 100 |
| 10005 | Э.41327E-0013 | 1. 160364 | kHz | 40 |
| -000s | $7.01823 E-0013$ | 1.127608 | kHz | 20 |
| .0005 | $7.89769 E-0013$ | 1. 135494 | kHz | 10 |
| 100005 | 1.37417E-0012 | 1.168728 | kHz | 4 |

[^8]\% Matlab script:
\% Study of the intensity profile of an iodine stabilized He-Ne Laser
\% The outspectra array contains the harmonics generated by the iodine
\% absorption peak. Change the variable tunevolt to sweep the laser \% frequency. Variable fmodpp outputs the modulation amplitude in Hz . Tom Ahola 7.5.1993

```
%tunevolt=0; % tuning voltage (high voltage piezo)
modvolt=1.7; % modulation voltage peak-to-peak
pztsensl=2.765e-9; % modulation piezo sensitivity [m/V]
pztsens2=3.68e-9; % tuning piezo sensitivity [m/V] (high voltage piezo)
k=5e-3; % absorption contrast
gamma=2.4e6; % absorption width [Hz]
vp=473612236.7e6; % absorption resonance (h-line) [Hz]
dvd=1400e6; % laser gain profile width [Hz]
vg=473612214e6; % laser resonance [Hz]
10=371e-3; % laser cavity length [m]
c0=2.997924562e8; % speed of light in vacuum [m/s]
ny=vp; % laser frequency (same as absorption resonance)
lambda=c0/ny; % laser wavelenght [m] (632.8nm)
n=10/lambda;
Vmod=[-16:16]/32*modvolt; % modulation voltage
ll=pztsensl*Vmod; % modulation piezo movement
l2=pztsens2*tunevolt; % tuning piezo movement
nyout=c0*n*(10+11+12).^(-1); % laser momentary frequency output
fmodpp=max(nyout) -min(nyout);
I=exp(-((nyout-vg)/dvd).^2).*(1+k*gamma^2*ones(nyout)./(gamma^2+(nyout-vp).^2));
Ipol=polyfit(Vmod,I,6); % intensity profile approximation
\begin{tabular}{ll} 
modsig=modvolt \(/ 2 * \sin ([0: 255] / 256 * 2 * \mathrm{pi}) ;\) & \(\%\) sine modulation signal \\
detectsig=polyval(Ipol,modsig) ; & \% simulate laser with absorption cell \\
outspectra=fft(detectsig); & \% output spectra
\end{tabular}
```


## LIST OF SCHEMATIC DIAGRAMS

| number | page | filename | date | title |
| :---: | :---: | :---: | :---: | :---: |
| KAU-932900A | 1/1 | KAUOO | 20.07.1993 | SUBRACK WIRING DIAGRAM |
| KAU-932011A | 1/1 | KAUll | 18.05.1993 | PZT-AMPLIFIER END STAGE |
| KAU-932012A | 1/2 | KAU12-1 | 18.05.1993 | PZT-AMPLIFIER |
| KAU-932012A | 2/2 | KAU12-2 | 18.05.1993 | PZT-AMPLIFIER |
| KAU-932013A | 1/1 | KAU13 | 19.05.1993 | PZT POWER SUPPLY |
| KAU-932121A | 1/2 | KAU21-1 | 25.05 .1993 | LOCK-IN AMPLIFIER |
| KAU-932121A | 2/2 | KAU21-2 | 27.05.1993 | LOCK-IN AMPLIFIER |
| KAU-932122A | 1/1 | KAU22 | 25.05.1993 | LOCK-IN AMP DETECTOR |
| KAU-932123A | 1/1 | KAU23 | 25.05.1993 | LOCK-IN AMP INTEGRATOR |
| KAU-932124A | 1/1 | KAU24 | 31.05 .1993 | LOCK-IN AMP FILTER BANK |
| KAU-932131A | 1/1 | KAU31 | 28.05.1993 | LASER POWER SUPPLY |
| KAU-932241A | 1/2 | KAU41-1 | 31.05 .1993 | MOD FREQUENCY GENERATOR |
| KAU-932241A | 2/2 | KAU41-2 | 04.06.1993 | MOD FREQUENCY GENERATOR |
| KAU-932251A | 1/1 | KAU51 | 02.06.1993 | MAIN POWER SUPPLY |
| KAU-932461A | 1/1 | KAU61 | 16.06.1993 | PELTIER POWER SUPPLY |
| KAU-932462B | 1/2 | KAU62-1 | 18.08.1993 | TEMPERATURE CONTROLLER |
| KAU-932462B | 2/2 | KAU62-2 | 18.08.1993 | TEMPERATURE CONTROLLER |
| KAU-934671B | 1/1 | KAU71 | 18.11.1993 | PHOTODETECTOR |


















## ENGINEERING CHANGE ORDER

ECO-1
PHASE SENSITIVE DETECTOR (KAU-932122A)


ECO-2
PZT AMPLIFIER (KAU-932012) AND TEMPERATURE CONTROLLER (KAU-932462)


| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 1 |  |  | MAINS SWITCH 2-POLE ON/OFF |
| 2 | 3 |  |  | SECONDARY CIRCUIT SWITCHES |
| 3 | 3 |  |  | FUSEHOLDER |
| 4 | 2 |  |  | FUSE 500mA-T |
| 5 | 1 |  |  | FUSE 2A-T |
| 6 | 1 |  |  | MAINS SOCKET |
| 7 | 1 |  |  | 9 POLE TERMINAL BLOCK |

KAU11 MATERIALS (PZT-AMPLIFIER END STAGE)

| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 2 | PA88 | U1 U2 | APEX POWER OP-AMP (8-PIN TO-3) |
| 2 | 4 | ZD | D9 D10 D11 D12 | ZPD2.7 ZENER DIODE |
| 3 | 8 | D | D1 D2 D3 D4 | RECTIFIER DIODE |
|  |  |  | D5 D6 D7 D8 |  |
| 4 | 2 | CAPC | C6 C2 | 15pF CER. CAP. 200M |
| 5 | 6 | CAPC |  | 100pF CER. CAP. 200M |
| 6 | 4. | CAPP | C11 C12 C14 C13 | 100nF POL. CAP. 250V 400M |
| 7 | 2 | CAPER | C10 C9 | $330 \mathrm{uF} / 200 \mathrm{~V}$ AL EL. CAP. 400M |
| 8 | 2 | R | R14 R4 | 33 |
| 9 | 2 | R | R13 R3 | 100 |
| 10 | 2 | R | R7 R17 | 1K |
| 11 | 2 | R | R10 R20 | 10K |
| 12 | 2 | R | R1 R11 | 44.2K |
| 13 | 2 | R | R2 R12 | 47K |
| 14 | 4 | R | R6 R5 R16 R15 | 470 K |
| 15 | 4 | R | R8 R9 R19 R18 | 1M |

## KAU12 MATERIALS (PZT-AMPLIFIER)

| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | REG7800 | U3 | 7805 (=LM340T5) |
| 2 | 2 | TLO74 | U1 U2 | QUAD LOWNOISE OP-AMP |
| 3 | 2 | ZD | D1 D2 | ZPD3.9 ZENER |
| 4 | 2 | ZD | D3 D4 | ZPD5.1 ZENER |
| 5 | 4 | R | R101 R100 R103 R102 | 10 |
| 6 | 1 | R | R104 | 22 |
| 7 | 2 | R | R30 R31 | 100 |
| 8 | 3 | R | R7 R14 R13 | 1 K |
| 9 | 1 | R | R11 | 5.10K |
| 10 | 1 | R | R12 | 6.8 K |
| 11 | 18 | R | R8 R4 R2 R1 R3 R5 | 10K |
|  |  |  | R6 R10 R9 R17 R15 |  |
|  |  |  | R18 R19 R20 R22 |  |
|  |  |  | R23 R21 R37 |  |
| 12 | 1 | R | R16 | 12K |
| 13 | 1 | R | R24 | 21 K |
| 14 | 4 | R | R40 R41 R42 R43 | 33K |
| 15 | 1 | R | R25 | 39K |
| 16 | 1 | R | R32 | 100K |
| 17 | 1 | R | R34 | 470K |
| 18 | 2 | R | R26 R28 | 3.9 M |
| 19 | 3 | TRMPOTH | R27 R29 R33 | IOK 1-TURN CERMET TRIM. POT. HOR. |
| 20 | 2 | TRMPOTH | R36 R35 | 1 K 1-TURN CERMET TRIM. POT. HOR. |
| 21 | 3 | CAPC | C1 C9 C10 | 220 pF CER. CAP. 200M |
| 22 | 6 | CAPC | $\begin{aligned} & \text { C11 C12 C105 C106 } \\ & \text { C107 C108 } \end{aligned}$ | 100 nF CER. CAP. 200M |
| 23 | 1 | CAPER | C111 | 100uF/16V AL. EL. CAP. 150M |
| 24 | 5 | CAPER | C101 C109 C102 C103 C104 | $470 \mathrm{uF} / 16 \mathrm{~V}$ AL. EL. CAP. 200M |
| 25 | 1 | CAPP | C110 | 100 nF POL. CAP. 400 M |
| 26 | 1 | CAPP | C2 | luF POL. CAP. 600M |
| 27 | 1 | SW3 | K1 K2 K3 K4 | 4XON-ON-ON ROTARY SWITCH |
| 28 | 2 | BNC |  | BNC FEMALE SOCKETS 500HM |
| 29 | 2 | MHV |  | MHV FEMALE SOCKETS 500HM |
| 30 | 1 | POT |  | 10-TURN POTENTIOMETER 20K (BOURNS |
| 31 | 1 | KNOB |  | KNOB TO THE POTENTIOMETER |
| 32 | 1 | KNOB |  | KNOB TO THE ROTARY SWITCH |
| 33 | 1 | DP-416-01 |  | KUWANO DIGITAL PANEL METER |
| 34 | 1 | H11 |  | VERO IlPIN TYPE HIl CARD Edge plu |

KAU13 MATERIALS (PZT POWER SUPPLY)

| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |  |
| :---: | :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | TRAFO | X1 | 230/2X250-15VA TRANSFORMER |  |
| 2 | 2 | L | L1 L2 | CHOKES (FERRITE BEADS ON WIRE) |  |
| 3 | 2 | FUSE | F2 F3 | 100 mA -T FUSE $5 \times 20 \mathrm{~mm}$ |  |
| 4 | 2 |  |  | PCB FUSEHOLDERS $5 \times 20 \mathrm{~mm} 900 \mathrm{M}$ |  |
| 5 | 1 |  |  | PANEL FUSEH | OLDER $5 \times 20 \mathrm{~mm}$ |
| 6 | 4 | CAPER | C1 C2 C3 C4 | 150uF/400V | AL-EL. CAP. 400 M |
| 7 | 4 | ZD | D7 D8 D5 D6 | ZPY100 | ZENER DIODE |
| 8 | 4 | D | D3 D4 D2 D1 | 1N4007 | RECT. DIODE |
| 9 | 2 | R | R5 R6 | $330 \quad 0.5 \mathrm{~W}$ | 900M |
| 10 | 4 | R | R1 R2 R3 R4 | 8.2K 1W | 900M |




| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 2 | NFETSGD | Q2 Q1 | 2N3819 NFET TRANS. |
| 2 | 2 | PNPCBE | Q3 Q4 | BC307B PNP TRANS. |
| 3 | 1 | OPAMP1 | U1 | TLE2027CP PRECISION OPAMP |
| 4 | 4 | R | R7 R9 R20 R21 | 100 |
| 5 | 2 | R | R22 R23 | 1 K |
| 6 | 2 | R | R14 R15 | 2.2K |
| 7 | 9 | R | R1 R2 R3 R4 R5 R6 R11 R12 R26 | 10K |
| 8 | 2 | R | R18 R19 | 22K |
| 9 | 2 | R | R8 R10 | 1 M |
| 10 | 1 | TRMPOTV | R13 | 500 10-TURN CERMET TRIM. POT. |
| 11 | 2 | TRMPOTV | R24 R25 | IOK 10-TURN CERMET TRIM. POT. |
| 12 | 2 | CAPC | C1 Cll | 4.7 nF CER. CAP. 200M |
| 13 | 2 | CAPC | C9 Cl0 | 470 pF CER. CAP. 200M |
| 14 | 3 | CAPC | C6 C8 C7 | 100 nF CER. CAP. 200M |
| 15 | 1 | CAPP | C2 | 0.22 UF POL. CAP. |
| 16 | 3 | CAPER | C3 C4 C5 | 47uF/16V AL EL. CAP. 150M |
| 17 | 2 | PH2X4M | J1 J2 | PINHEADER 2 X 4 MALE (ANGLE) |



| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | TLO74 | U1 | TLO74N QUAD LOWNOISE OP-AMP |
| 2 | 1 | R | R16 | 3.01 K |
| 3 | 1 | R | R4 | 3.74 K |
| 4 | 3 | R | R2 R5 R6 | 4.02 K |
| 5 | 2 | R | R13 R15 | 6.19 K |
| 6 | 1 | R | R11 | 9.09 K |
| 7 | 2 | R | R8 R10 | 10.0K |
| 8 | 2 | R | R7 R9 | 20.0K |
| 9 | 1 | R | R1 | 80.6K |
| 10 | 1 | R | R3 | 162 K |
| 11 | 3 | TRMPOTH | R12 R14 R17 | 500 10-TURN CERMET TRIM. POT |
| 12 | 4 | CAPP | C1 C2 C7 C8 | 4.7 nF POL. CAP. 200M |
| 13 | 1 | CAPP | C9 | 10 nF POL. CAP. 200M |
| 14 | 2 | CAPC | C3 C4 | 100 nF CER. CAP. 200M |
| 15 | 2 | CAPER | C5 C6 | 100uF/16V AL EL. CAP. 150M |
| 16 | 1 | JUMPER | J1 | JUMPER WIRE 100M |
| 17 | 2 | PH2X12M | J2 J3 | PINHEADER 2 X 4 MALE |


| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |  |
| :---: | :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | TRAFO | X1 | 230V/1400V | 40VA TRANSFORMER |
| 2 | 4 | D | D1 D2 D3 D4 | HSKE3500 | HIGH VOLT. DIODE |
| 3 | 4 | NPNCBE | Q1 Q2 Q3 Q4 | BUV47A | HIGH VOLT. TRANSISTOR |
| 4 | 1 | PNPCBE | Q5 | 2N2905 | PNP TRANSISTOR |
| 5 | 1 | ZD | D5 | 6.2 V | ZENER DIODE |
| 6 | 2 | R | R35 R42 | 220 0.4W |  |
| 7 | 1 | R | R36 | $1 \mathrm{~K} \quad 0.4 \mathrm{~W}$ |  |
| 8 | 1 | R | R43 | $8.2 \mathrm{~K} \quad 0.4 \mathrm{~W}$ |  |
| 9 | 1 | R | R41 | 10 K 2W |  |
| 10 | 1 | R | R40 | 15K 2W |  |
| 11 | 5 | R | $\begin{array}{llll}\text { R21 } & \text { R22 } & \text { R23 } & \text { R24 }\end{array}$ | 47K 6W |  |
| 12 | 1 | R | R34 | 44.2K 0.4W |  |
| 13 | 1 | R | R25 | 220K 1W |  |
| 14 | 8 | R | R26 R27 R28 R29 | 220 K 0.4 W |  |
|  |  |  | R30 R31 R32 R33 |  |  |
| 15 | 20 | R | R1 R2 R3 R4 R5 R6 | 1.5M 0.4W |  |
|  |  |  | R7 R8 R9 R10 R20 |  |  |
|  |  |  | R19 R18 R17 R16 |  |  |
|  |  |  | R15 R14 R13 R12 |  |  |
|  |  |  | R11 |  |  |
| 16 | 2 | TRMPOTV | R44 |  |  |
| 17 | 1 | POT | R38 | 5K 1-TURN CERMET TRIM. POT. VERT 5K POTENTIOMETER PLASTIC SHAFT |  |
| 18 | 20 | CAPER | $\begin{array}{cl}\text { C1 } & \mathrm{C} 2 \\ \mathrm{C} 3 & \mathrm{C} 4 \\ & \mathrm{C} 5\end{array} \mathrm{C} 6$ | 150uF 400y | AL EL. CAP. 400M |
|  |  |  | C7 C8 C9 C10 C20 |  |  |
|  |  |  | C19 C18 C17 C16 |  |  |
|  |  |  | C15 C14 C13 C12 |  |  |
|  |  |  | C11 |  |  |
| 19 | 2 | CAPC | C21 C22 | 2.2 nF 4 KV CER. CAP. 400 M |  |
| 20 | 1 | CAPP | C23 | 150 nF POL | CAP. 200 M |
| 21 | 2 | CAPC | C24 C25 | $4.7 \mathrm{nF} \quad 5 \mathrm{kV}$ | CER. CAP. 400 M |
| 22 | 1 | SW1 | K1 | HIGH VOLTAGE PUSH BUTTON |  |
| 23 | 1 | TEM-65E |  | KUWANO PANEL METER 100uA 8A |  |
| 24 | 2 | SHV |  | SHV FEMALE SOCKET |  |
| 25 | 1 | SHV |  | SHV MALE PLUG <br> RG59 CABLE |  |
| 26 | 1m | RG59 |  |  |  |  |
| 27 | 2m |  |  | ```RG59 CABLE 10 kV HIGH VOLTAGE CABLE``` |  |
| 28 | 1 | HV-CABLE TERMINAL BLOCK |  | 10 kV HIGH VOLTAGE CABLE <br> 3-WAY TERMINAL BLOCK 250VAC |  |

KAU41 MATERIALS (MOD FREQUENCY GENERATOR)

| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | 7404 | U1 | 74HCO4 HEX INVERTTER |
| 2 | 1 | 7492 | U2 | $74 \mathrm{LS92}$ COUNTER |
| 3 | 1 | 7490 | U3 | $74 \mathrm{LS90}$ COUNTER |
| 4 | 3 | 7474 | U4 U5 U7 | 74HC74 D-FLIP FLOP |
| 5 | 1 | 7400 | U8 | 74 HCOO QUAD NAND |
| 6 | 1 | 7420 | U9 | $74 \mathrm{HC20}$ DUAL NAND |
| 7 | 1 | 74121 | U6 | 74121 MONOSTABLE F.F. |
| 8 | 2 | OPAMP2 | U102 U104 | NE5532 LOW NOISE DUAL OPAMP |
| 9 | 2 | OPAMP1 | U101 U105 | NE5534 LOW NOISE SINGLE OPAMP |
| 10 | 1 | OPAMP1 | U103 | TLO71 LOW NOISE SINGLE OPAMP |
| 11 | 2 | R | R130 R131 | 220.4 W |
| 12 | 7 | R | R1 R2 R3 R7 R8 R40 R106 | 1 K |
| 13 | 1 | R | R121 | 1.8 K |
| 14 | 1 | R | R110 | 2.87 K |
| 15 | 1 | R | R119 | 4.7 K |
| 16 | 5 | R | R107 R108 R109 | 5.76 K |
|  |  |  |  |  |
| 17 | 1 | R | R120 | 6.8 K |
| 18 | 1 | R | R27 | 7.5K |
| 19 | 6 | R | $\begin{array}{lllll}\text { R112 } & \text { R114 } & \text { R115 } & \text { R116 } \\ \text { R117 } & \text { R118 } & & \end{array}$ | 10K |
| 20 | 4 | R | R4 R5 R9 R10 | 22K |
| 21 | 1 | R | R102 | 191K |
| 22 | 2 | R | R101 R103 | 576K |
| 23 | 1 | R | R20 | 10M |
| 24 | 1 | xtal | x 1 | $1 \mathrm{MHz} \mathrm{CRYSTAL} \mathrm{500M}$ |
| 25 | 3 | CAPC | C43 C128 C129 | 22p CER. CAP. 200M |
| 26 | 1 | CAPC | C42 | 82p CER. CAP. 200M |
| 27 | 11 | CAPC | C114 C115 C117 C118 | $100 \mathrm{nF}, \mathrm{CER}$. CAP. 200M |
|  |  |  | C119 C120 C124 C125 |  |
|  |  |  | C121 C122 C123 |  |
| 28 | 2 | CAPP | C104 Clll | 4.7nF POL. CAP. 200M |
| 29 | 1 | CAPP | C1 | 3.3 nF POL. CAP. 200M |
| 30 | 2 | CAPP | C106 C108 | 2.2 nF POL. CAP. 300 M |
| 31 | 2 | CAPP | C101 C102 C105 C107 | 10 nF POL. CAP. 300M |
| 3.2 | 1 | CAPP | C103. | 100 nF POL. CAP. 200M |
| 33 | 1 | CAPP | C109 | luF POL. CAP. 600M |
| 34 | 5 | CAPER | C112 C113 C126 C127 | $470 \mathrm{uF} / 16 \mathrm{~V}$ AL EL. CAP. 200M |
|  |  |  | C116 |  |
| 35 | 1 | CAPT | C130 | 1uF/10V TANT. EL. CAP. 200M |
| 36 | 1 | JUMPER | J101 | JUMPER |
| 37 | 1 | SW4 | K1 | 1 POLE 4 POSITON ROTARY SWITCH |
| 38 | 1 |  |  | BOURNS 10-TURN POTENTIOMETER 20 K |
| 39 | 1 |  | . | BOURNS 10-TURN POTENTIOMETER 5K |
| 40 | 2 |  |  | BOURNS H490 SERIES 10-TURN 1" DIA |
| 41 | 2 | BNC |  | BNC FEMALE SOCKET 500HM |
| 42 | 1 | KNOB |  | KNOB FOR ROTARY SWITCH |


| ITEM | QTY | COMP-NAME | REFERENCE-DESIGNATOR | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | TRAFO | X1 | 230V/2x15V 40VA TRANSFORMER |
| 2 | 1 | TRAFO | X2 | 230V/2x6V 25VA TRANSFORMER |
| 3 | 2 | REG7800 | U1 U2 | LM7815CT REGULATOR |
| 4 | 2 | REG7900 | U3 U4 | LM7915CT REGULATOR |
| 5 | 1 | REG7800 | U5 | LM7805CT REGULATOR (LM340T5) |
| 6 | 3 | D | D1 D2 | 1N4005 RECT. DIODE |
| 7 | 3 | FUSE | F1 F2 F3 | 2A-T FUSES |
| 8 | 3 |  |  | PCB FUSE CLIPS |
| 9 | 1 | RECTBDGE | X3 | RECTIFIER BRIDGE 3A/60V |
| 10 | 2 | CAPER | C1 C2 | 4700 OF 25V AL EL. CAP. 300M |
| 11 | 1 | CAPER | C11 | 4700 UF 16V AL EL. CAP. 300M |
| 12 | 5 | CAPER | C7 C8 C9 C10 Cl3 | 470 uF 16V AL EL. CAP. 200M |
| 13 | 5 | CAPP | C3 C4 C5 C6 C12 | 100 nF CER. CAP. 200M |
| 14 | 1 | TERMINAL | BLOCK | 2-WAY TERMINAL BLOCK 250VAC |
| 15 | 5 | TERMINAL | BLOCK | 4-WAY TERMINAL BLOCK PCB MOUNT |
| 16 | 3 | LED |  | 5 mm RED LED |
| 17 | 3 | R |  | 470 |
| 18 | 3 |  |  | 5 mm LED CLIP |




## KAU71 MATERIALS (PHOTODETECTOR)

| ITEM | QTY | PART NAME | VALUE | REFERENCE | DESCRIPTION |
| :---: | :---: | :---: | :---: | :---: | :---: |
| 1 | 1 | PD | S2386-5K | D1 | Photodiode |
| 2 | 1 | OPA111 |  | U2 | Low Noise Single Op-Amp |
| 3 | 1 | OP27 |  | U6 | High Speed Precision Op-Amp |
| 4 | 2 | TL072 |  | U3-4 | Low Noise Dual Op Amp |
| 5 | 1 | NE5534 |  | U1 | Low Noise Single Op-Amp |
| 6 | 1 | LM301 |  | U5 | Single General Purpose Op-Amp |
| 7 | 2 | D | 1N4005 | D2-3 | Silicon Diode |
| 8 | 1 | CAPC | 15 pF | C1 | CERAMIC CAPACITOR, 200M |
| 9 | 1 | CAPC | 22 pF | C2 | CERAMIC CAPACITOR, 200M |
| 10 | 1 | CAPC | 47 pF | C3 | CERAMIC CAPACITOR, 200M |
| 11 | 2 | CAPC | 33 pF | C9 C11 | CERAMIC CAPACITOR, 200M |
| 12 | 6 | CAPC | 0.14 F | C16-21 | CERAMIC CAPACITOR, 200M |
| 13 | 1 | CAPP1 | 10 nF | C4 | POLYESTER CAPACITOR, 200M |
| 14 | 1 | CAPP1 | 1 nF | C10 | POLYESTER CAPACITOR, 200M |
| 15 | 4 | CAPP2 | 10 nF | C5-8 | POLYESTER CAPACITOR, 200M |
| 16 | 1 | CAPT | 22uF | C22 | CAP TANTALUM RADIAL, 200M |
| 17 | 4 | CAPER7 | 470 uF | C12-15 | ELECTROLYTIC RADIAL CAP, 200M |
| 18 | 1 | R | 1.00 M | R1 | RES $1 / 4 \mathrm{~W}$ |
| 19 | 3 | R | 470 k | R2 R5 R23 | RES 1/4W |
| 20 | 1 | R | 365 | R6 | RES 1/4W |
| 21 | 4 | R | 10.0k | R7-8 R18 R20 | RES $1 / 4 \mathrm{~W}$ |
| 22 | 1 | R | 2.20 k | R12 | RES $1 / 4 \mathrm{~W}$ |
| 23 | 1 | R | 1.80 k | R13 | RES $1 / 4 \mathrm{~W}$ |
| 24 | 1 | R | 56.0 k | R14 | RES $1 / 4 \mathrm{~W}$ |
| 25 | 3 | R | 1.91 k | R15-17 | RES $1 / 4 \mathrm{~W}$ |
| 26 | 4 | R | 100k | R9-10 R21-22 | RES 1/4W |
| 27 | 1 | R | 4.70 k | R19 | RES 1/4W |
| 28 | 1 | R | 220 k | R3 | RES 1/4W |
| 29 | 4 | R | 22 | R24-27 | RES 1/4W |
| 30 | 1 | R | 100 | R28 | RES 1/4W |
| 31 | 2 | TRIMPOT | 100k | R4 R11 | 10 TURN CERMET TRIMPOT |
| 32 | 3 | JUMPER |  | J1-3 | JUMPER BLOCK, 2 PINS, 100M |



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$\square$


KAU-932121




## KAU-932123








KAU-932462




[^0]:    ${ }^{1}$ Trademark of The Math Works, Inc.

[^1]:    ${ }^{2}$ The dip at the center of the absorption line in the DC component output graph (figure 2.3) and the nonlinear slope between the negative and positive peaks in the 3 f component output graph are probably caused by the modulation. These effects can not be observed in the intensity graphs in figure 2.1 because the derivation of the intensity corresponds to an infinitely small modulation amplitude.

[^2]:    ${ }^{1}$ by MicroSim Corporation

[^3]:    ${ }^{2}$ by Helsinki University of Technology and Nokia Corporation, Research Center

[^4]:    ${ }^{3}$ TIM distortion or transient intermodulation distortion.

[^5]:    ${ }^{4}$ Sometimes the time constant of an integrator is reported to be $\tau=R_{A} C_{2}$. This is especially true if $R_{B}=0 \Omega$. When comparing the time constants between different equipment it should be checked which definition is used.

[^6]:    ${ }^{1}$ THD $=$ total harmonic distortion. It was measured with a digital spectrum analyzer from the first five harmonic components.

[^7]:    ${ }^{1}$ The iodine absorption linewidth will be larger than the natural linewidth in the laser cavity because of power broadening. However, as the separation of the negative and positive peaks of the third derivative of the intensity profile is less than the real linewidth, it appears that the separation is the same as the natural linewidth.

[^8]:    Nean frequeney $=39.423377212 \mathrm{MHz}$
    Max frequen $=35.581219000 \mathrm{MHz}$
    Nin frequeney $=39.271252000 \mathrm{MHz}$

