

UNIVERSITY OF OSLO
Department of Physics

Suppression of
acoustic feedback by
frequency shifting

An approach for howl
suppression in a public
address system

Master thesis

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Blindern, august 2012
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Nomenclature

<i>AP</i>	Audio Precision
<i>dB</i>	Decibel, logarithmic measurement unit used to describe a physical quantity, e.g. sound pressure, relative to a defined reference
<i>dBA</i>	Denotation for A-weighted decibels, in this thesis for sound pressure level measurements. Also denoted as dB(A) elsewhere
<i>FB</i>	Abbreviation in this thesis for acoustic feedback
<i>Feedback</i>	General term for when a signal of some sort is fed back from the output of a system, to the input. In this thesis as well in the everyday language the term is often used to describe acoustic feedback (Larsen effect)
<i>FPGA</i>	Field-Programmable Gate Array, a programmable digital hardware electronic chip
<i>GS/s</i>	Giga samples per second, 10^9 samples per second
<i>IC</i>	Integrated Circuit
<i>Infrasound</i>	Sound with frequency below the human hearing threshold of 20 Hz
<i>IP</i>	Intellectual Property, a ready-made free or commercial piece of hardware code for FPGA
<i>LED</i>	Light Emitting Diode
<i>op – amp</i>	Operational Amplifier, a type of integrated circuit used in many different variations in electronic circuitry
<i>PA</i>	Public Address. A generic term used for electronic sound systems containing a microphone, amplifier and speaker(s) which purpose is to address the public. This could be used as anything from announcements at a train station to singing in a concert hall
<i>PCB</i>	Printed Circuit Board

<i>PT</i>	Abbreviation in this thesis for Prototype. Four different prototypes were produced, PT1, PT2, PT3 and PT4.1
<i>RCA</i>	Common connector for audio and video equipment. Often also referred to as Phono connector
<i>rms</i>	Root Mean Square. Measure of magnitude for time varying signals, for example sinusoids
<i>SMD</i>	Surface Mount Device, in opposition to for example through-hole components
<i>SPICE</i>	Simulation Program with Integrated Circuit Emphasis. Open source platform to simulate and verify analogue electronic circuits
<i>SPL</i>	Sound pressure level, also denoted in literature as L_p . Measured in dB. In this thesis A-weighted dB, dBA, is used
<i>TFS</i>	Abbreviation in this thesis for The Frequency Shifter, the main circuit described and developed in this thesis
<i>THD</i>	Total Harmonic Distortion. A way of measuring the “perfectness” of a sine wave, as in no harmonics, only a single frequency
<i>THD + N</i>	Total Harmonic Distortion + Noise. A measurement of THD where the noise also are calculated for
<i>UiO</i>	University of Oslo (Norwegian: Universitetet i Oslo)
<i>Ultrasound</i>	Sound with frequency above the human hearing threshold of 20,000 Hz
V_{pp}	Volt, peak-to-peak
<i>VHDL</i>	VHSIC Hardware Description Language, language for programming FPGA and ASIC circuits
<i>XLR</i>	Common 3 pin connector for audio equipment, supporting balanced signaling

Abstract

In this master thesis a way to suppress acoustic feedback in a public address system, has been studied. The setup is typically a microphone, amplifier and loudspeaker(s) used in a confined space, coupled with the solution (TFS) in this thesis as an external circuit between the microphone and the amplifier. The circuit is utilizing a small frequency shift of the sound, to allow the gain to be turned up a few more decibels, before the infamous howling sound takes over. An analogue hardware variant of the circuit described by Martin Hartley Jones [4] was studied, developed and modernized, using current available components. The original circuit by Hartley Jones was greatly altered to implement possibilities for more comprehensive testing scenarios regarding the phase shift frequency. In addition, modifications were made to make the circuit able to handle balanced signals directly from a microphone. This resulted in a completely new quadrature oscillator section, as well as a new preamplifier section.

The testing was performed with three different setups, and the possibility of higher gains before feedback occurs, were confirmed in all three scenarios. Even though the results indicate the possibility of ~ 6 dBA higher sound pressure levels, more testing is required to get explicit accurate numbers.

Chapter 1

Introductions

1.1 Background and motivation

Most people attending a seminar or concert where a public address (PA) system is in use, have experienced the annoying, often ear hurting, phenomenon called acoustic feedback. The sound, also referred to as “howling” or “Larsen effect”, is an unwanted effect of the use of a microphone, amplifier and one or more loudspeakers in a confined space to amplify a sound source. Such an arrangement is presented in figure 1.1. The most common

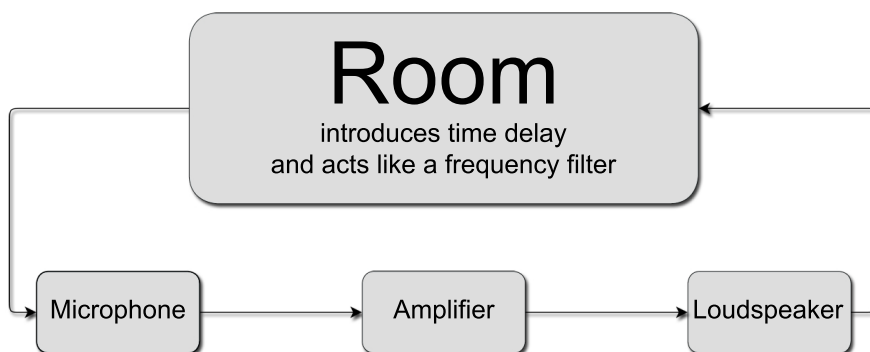


Figure 1.1: The block diagram shows a normal setup of public address systems. The room introduces a small time delay, and acts like a filter for the sound going from the loudspeakers and back into the microphone, forming a positive feedback loop. Because of this, some frequencies will at high gain, result in the phenomenon called acoustic feedback.

example of sources would be a human voice or an instrument. The howling is the result of the sound traveling through the room from the loudspeakers back to the microphone, making a positive feedback loop that in a way amplifies itself. This happens when the gain, or volume in layman’s terms, is turned up high enough. The room acts like a filter in this situation, and all elements within the room will affect at what gain and at which frequencies the acoustic feedback will occur. This includes the placement of the microphone and the loudspeaker(s), the size of the room, the materials

which the floor, walls and ceiling are made of, as well as the objects within the room such as chairs, tables and people. Other factors that may influence when feedback occur, are ambient pressure and temperature in the room. In addition to damage ones hearing temporarily or permanently, the feedback could also ruin the equipment involved, especially the amplifier.

The author's keen interest in noise and sound resulted in studying at a system that can suppress acoustic feedback. There are many ways this could be done, and most known are adaptive filter modeling, automatic notch filtering and frequency shifting. The latter method is studied in this thesis, and a setup in a public address system can be seen in figure 1.2. For the interested reader a numerous articles describes the other two methods, for example Loctwassana et al. [5] and Martinez-Leira et al. [6] on the topic of adaptive filtering and techniques using automatic notch filtering. This thesis takes a starting point in the frequency shifting circuit described by Jones in 1973 [4].

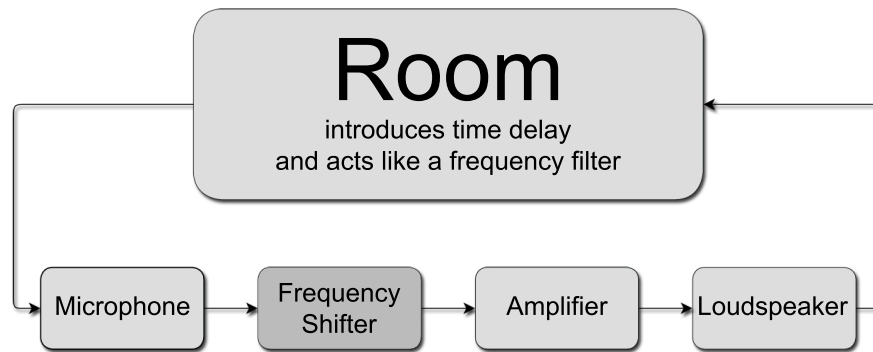


Figure 1.2: In this diagram the frequency shifter can be seen in the signal chain between the microphone and the amplifier. Signals coming from the microphone will be shifted in frequency. The frequency shifter makes sure to even out the loudest frequency peaks, so the gain can be turned up more, before feedback occur.

1.2 Goal of the thesis

There were several goals with this thesis, some major, but a couple of minor as well. The major goals were the main objectives of the thesis, and the focus while studying and developing the circuit and testing scenarios. The major goals of the thesis were defined:

- Build a working version of Jones's frequency shifter from 1973, with components available in 2012
- Modernize the frequency shifter to be able to handle balanced microphone signals
- Verify the anti feedback theory with several different frequency shifts, not only 5 Hz

The minor goals acted more like bonus achievements, and were expected to be fulfilled as the process went on. The minor goals were:

- Get a better understanding of the theory of sound pressure and the practical measurement of it
- Go through all the steps needed when making an electronic product, including drawing schematics, PCB layout, etching and assembly
- Run simulations of a circuit, and later building it to verify and compare the results

1.3 Structure of the thesis

This thesis was structured to look familiar to the reader acquainted with scientific master theses, papers and other writings. Below is a short description of the contents of each chapter.

Chapter 1 presents the motivation and background, as well as enlisting the goals of the thesis.

Chapter 2 goes through the most important theory needed to understand why and how a frequency shifter works.

Chapter 3 describes the frequency shifter in detail, presenting every sub-circuit of the design.

Chapter 4 enlists all the equipment, both soft- and hardware, used in the development and testing processes.

Chapter 5 goes through all methods in this thesis, including the design process and how calibration and testing were performed.

Chapter 6 contains the measurement results from the tests of the frequency shifter in a large lecturing hall.

Chapter 7 discusses the results and draws the final conclusions of the thesis.

Appendix A contains the MATLAB code listing for the MATLAB simulation of the circuit.

Appendix B shows all full scale schematics of prototype 4.1 of the frequency shifter.

Appendix C contains layout of both top and bottom electrical layer of TFS PT4.1, as well as all component positions on the PCB.

Appendix D enlists all parts used in TFS PT4.1.

Chapter 2

Theory

To fully understand the function and purpose of the circuit described in this thesis, a bit of background theory is required. There are mainly three parts of theory that need to be understood: Basic knowledge of sound and sound pressure, the frequency response of a room and the math utilized to deploy the frequency shifter circuit (TFS).

2.1 Sound and sound pressure

Before discussing the nature of sound, some knowledge of measurement units must be acquired. Time is measured in seconds (s). Sound are changes, waves or oscillations of pressure in a medium. The medium considered in this thesis is air. The waves of pressure comes at a rate, called frequency, measured in hertz (Hz). The relation between time and frequency are given by

$$f = \frac{1}{T} \quad (2.1)$$

where f is the frequency in Hz and T is the period of one waveform cycle in seconds. In other words is Hz oscillations per second. The human ear can perceive frequencies from 20 Hz to 20,000 Hz. That being said, frequencies in both ends of this spectrum are not heard by most people. This is especially true for the high end of the spectrum, as the sensibility for higher frequencies is weakened as we grow older.

Pressure is measured in Pascal (Pa). Sound pressure is the difference between the pressure in the air and the pressure in the sound wave. A logarithmic scale is used to denote sound pressure level, SPL (in some books/articles denoted as L_p or A_{SPL}):

$$SPL = 10 \log_{10} \frac{p^2}{p_{ref}^2} \text{ dB}$$

and rewritten, sound pressure level can therefore be defined

$$SPL = 20 \log_{10} \frac{p}{p_{ref}} \text{ dB} \quad (2.2)$$

where p is the rms of the measured pressure and p_{ref} is the reference pressure. The reference pressure used is 20 μPa , approximately the average threshold for human hearing at 1 kHz.

As the calculation of SPL takes into account sound of all frequencies, including both infra- and ultrasound, there is usually a weighting of the different frequency bands involved when measuring sound pressure level. If no weighting is applied, the SPL measured will not reflect what we humans perceive. Many different weighting curves exist, and common for all of them is that frequencies are suppressed outside the hearing range, 20 - 20,000 Hz. Within the hearable spectrum of frequencies some curves try to smoothen out the response, so it is similar to the ear response. The instrument used for measurements in this thesis uses A-weighting, and all measurements are therefore given in dBA. According to Tashev [11] “A-weighting matches the human perception at ambient noise levels around 40 dB SPL”. From figure 2.1 we see that the lower and higher frequencies in the hearable spectrum is attenuated, and the peak of the response lies around 2500 Hz.

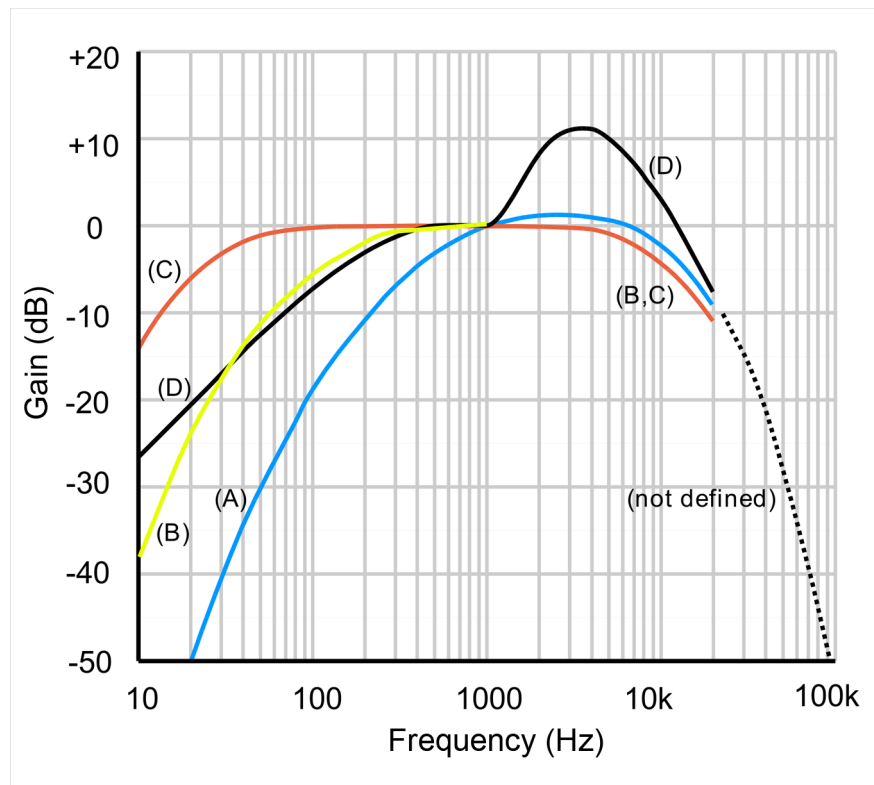


Figure 2.1: The figure shows A-weighting (blue) as well as B-weighting (yellow), C-weighting (red) and D-weightings (black). Source: Wikipedia [12]

2.2 Reverberation time and room frequency response

When sound is transmitted into a room from a source, the room, and all things within it, will have an influence on what we hear. Some frequencies will be absorbed and attenuated, others reflected around the room, depending on placement of objects, and what they are made of. Different materials have different properties regarding absorption and reflection of sound, and these properties are often frequency dependent. Reverberation, or reverb for short, is the phenomenon where the sound bounces around the room, so that the sound can be heard even after the source stops to produce the sound. A classical example is inside a cathedral. Often made of stone, the walls of the cathedral reflect the sound, in opposition of absorbing it, and the sound can be heard echoing around long after the source has become quiet. In other words can reverberation be described as the sum of many echoes. Reverberation time is in short terms the time from a sound source is cut off until the room has become quiet. Different frequencies will have different reverberation time in a given room. The reverberation time is defined

$$RT_{60} = \text{time before the level has decayed 60 dB from initial value} \quad (2.3)$$

In practice RT_{60} can be difficult, or impossible, to measure because of the high background noise. In such cases RT_{30} or RT_{20} are used instead. When a microphone, amplifier and loudspeaker are used in a confined space, the room acts as a filter. For feedback to occur there has to be a sound loop from the loudspeaker and back into the microphone. This is an example of positive feedback. As the frequency response of all the equipment involved, microphone, amplifier, loudspeaker and room, is not flat, some frequencies will make feedback occur earlier than others when the gain is turned up.

In figure 2.2 on the next page we see an example of a frequency response of a room. As we can see, the peaks of different frequencies are several decibels above the average response, shown by the dotted line in the figure. If we therefore shift all frequencies for example 5 Hz up at every loop, the peaks will “travel” upwards the frequency spectrum for every loop through the chain. This will result in all sounds been shifted in frequency, so a perfect pitch A at 440 Hz will no longer sound like that, but rather have the frequency 445 Hz. The feedback of that tone will of course have the frequency 450 Hz the next time around the loop, 455 Hz the time after that and so forth. In general the new frequency will be

$$f_{shifted} = f_{init} + \Delta f \quad (2.4)$$

where f_{init} is the initial frequency sent into the microphone and Δf is the frequency shift. Because the feedback portion of the signal is only a fraction of the total signal sent into the microphone, this loop effectively attenuates feedback at the cost of getting the original sound frequency shifted by a small amount. This is shown by Schröder [9]. In another article by Schröder [10]

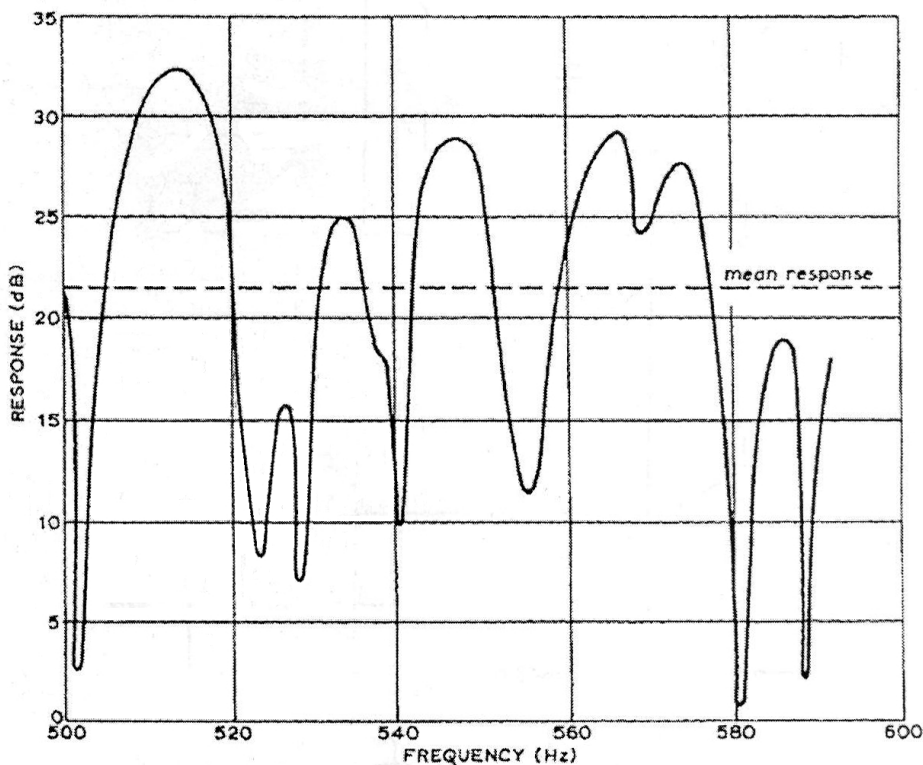


Figure 2.2: A small section, 500 - 600 Hz, of a room's typical frequency response. The distances between the peaks are only a few hertz, and several peaks are above the average room response marked by the dotted line. Source: Jones [4]

it is shown that the average spacing between peaks such as those shown in figure 2.2 is given by

$$\Delta f_{max} \approx \frac{4}{T_r} \text{ Hz} \quad (2.5)$$

where T_r is the reverberation time of the room. This means that if the reverberation time of the room is very short, a large frequency shift must be applied to suppress the feedback, and vice versa. It is therefore interesting to have an idea of the reverberation time of the room that we are going to deploy the frequency shifter. T_r is in this thesis assumed to be equal to RT_{60} .

2.3 Frequency shift and mathematics

The mathematics in this section will be straight forward, and very much reflect the simple formula juggling Hartley Jones [4] did in his article. In subsection 2.2 on the previous page we established that we needed a frequency shift to suppress the feedback. If we send a sinusoidal signal into the frequency shifter, this can be described as a function of time, t , by

$$V_{in}(t) = A \sin(2\pi ft) \quad (2.6)$$

where A and f is the amplitude and frequency of the signal respectively. The circuit should then output a signal with the same amplitude, but shifted in frequency by Δf :

$$\begin{aligned} V_{out}(t) &= A \sin(2\pi(f + \Delta f)t) \\ &= A \sin(2\pi ft + 2\pi\Delta ft) \end{aligned} \quad (2.7)$$

By the trigonometric relation

$$\sin(u + v) = \sin(u)\cos(v) + \cos(u)\sin(v)$$

equation 2.7 can be expanded to

$$V_{out}(t) = A \left(\sin(2\pi ft)\cos(2\pi\Delta ft) + \cos(2\pi ft)\sin(2\pi\Delta ft) \right) \quad (2.8)$$

It is equation 2.8 that will be synthesized by the circuit in this thesis. So far only calculations for simple sinusoidal waves have been shown. Luckily Fourier theory tells us that all signals, no matter how complex, can be represented as a sum of different sinusoidal signals. Therefore equation 2.8 holds for complex signals like speech, as well as simple pure sinusoidal signals.

2.4 A note on frequency shifts

As shown in section 2.2 on page 7, the circuit in this thesis shifts all signals upwards in frequency. This implies that the solution described will not be suitable for use in most music applications. If for example the guitar chord major E is played, it consists of the notes seen in the first row of table 2.1. The second row shows the corresponding frequencies rounded to one decimal, not considering overtones. A tone's pitch is said to be one octave above another when the frequency is doubled. As seen, E3 played on the 3rd string is one octave above the 1st string's note E2, as the frequency of E3, f_{E3} , is bound to the relationship

$$f_{E3} = 2 \cdot f_{E2}$$

where f_{E2} is the frequency of the note E2 (one octave below E3). If we

Table 2.1: The table shows the notes played in a E major chord on a standard tuned guitar (EADGHE - tempered scale) and their corresponding frequencies. In the bottom row is all the notes frequency shifted upwards with 5 Hz.

String	1st	2nd	3rd	4th	5th	6th
Note	E2	B2	E3	G#3	B3	E4
Frequency (Hz)	82.4	123.5	164.8	207.6	246.9	329.6
New frequency (Hz)	87.4	128.5	169.8	212.6	251.9	334.6

introduce a frequency shift of 5 Hz, all the tones will now have the frequencies corresponding to the bottom row. Now the note produced by the 3rd string, formerly E3, is no longer one octave above note the 1st string produced

$$169.8 \text{ Hz} \neq 2 \cdot 87.4 \text{ Hz} = 174.8 \text{ Hz} \quad (2.9)$$

Harmonic relationships would only be preserved if the frequency shift was at a fixed percentage of the signal's frequency. From equation 2.9 on the previous page we see that all harmonic relations are destroyed when introducing a fixed frequency shift, and this implies that the frequency shifter presented in this thesis will not be suitable to music applications where harmonic relations and scales are important, in other words music as most people are used to listening to. This of course does not mean that all kinds of music usage are excluded, a lot of experimental and contemporary music could be played through the shifter, as well as more borderline projects like noise sounds capes and the like. Nevertheless, the main usage for the frequency shifting circuit presented here is speech, not music.

Chapter 3

The frequency shifter circuit

In this chapter all parts of the frequency shifter circuit (TFS) are described in detail. The development and redesigning process of the product is described in chapter 5 on page 25, but in this chapter every bit of the circuit and its sub-circuits are explained.

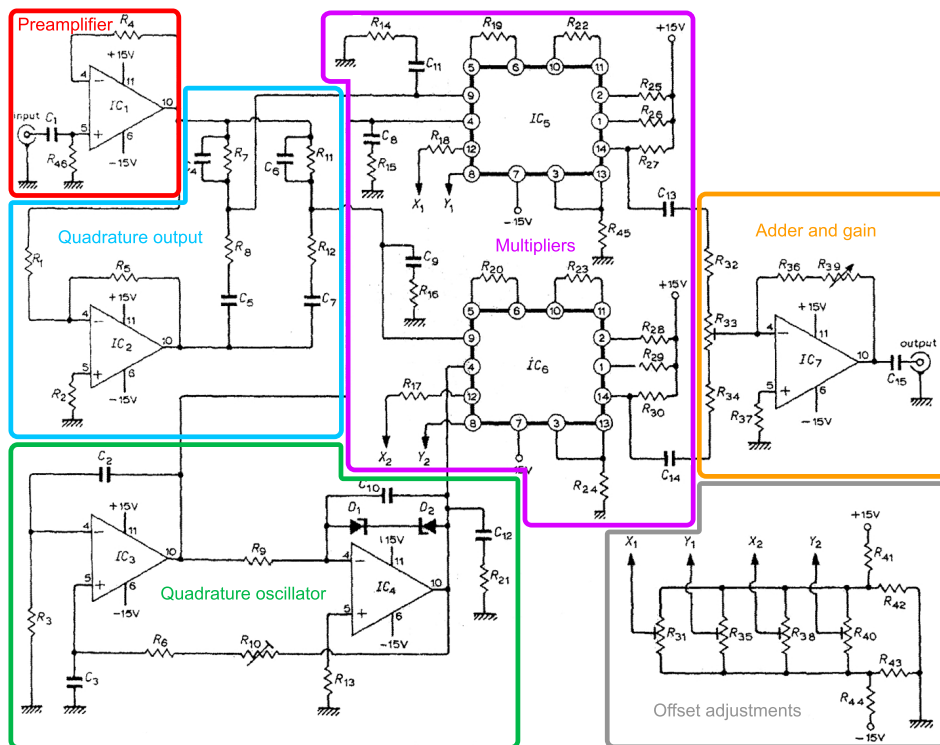


Figure 3.1: Seen here is the original frequency shifting circuit made by M. Hartley Jones as presented in *Wireless World*, July 1973. The power supply part of the circuit is not shown. The sub-circuits have been color coded for easy comparison of the original and new circuit in this thesis.

TFS is based upon a circuit made by Martin Hartley Jones in 1973 [4], as seen in figure 3.1. To clarify the original circuit all sub-circuits are highlighted with different colors. This is to ease the comparison with the one

developed and presented in this thesis. Even though the circuit described in this thesis differ from Hartley Jones’s circuit, the main elements and sub-circuits are mostly the same. For a detailed explanation of Hartley Jones’s circuit, please refer to Hartley Jones’s article [4]. As TFS was designed for testing purposes, a lot of circuitry can be omitted if one would want to build a version for usage. Only one quadrature oscillator, with a fixed frequency, would be needed. The possibility for oscillator selection, *OSCSEL*, and output of the preamplified signal only, *PREAMPED*, would also be eliminated in a circuit not designed for testing purposes.

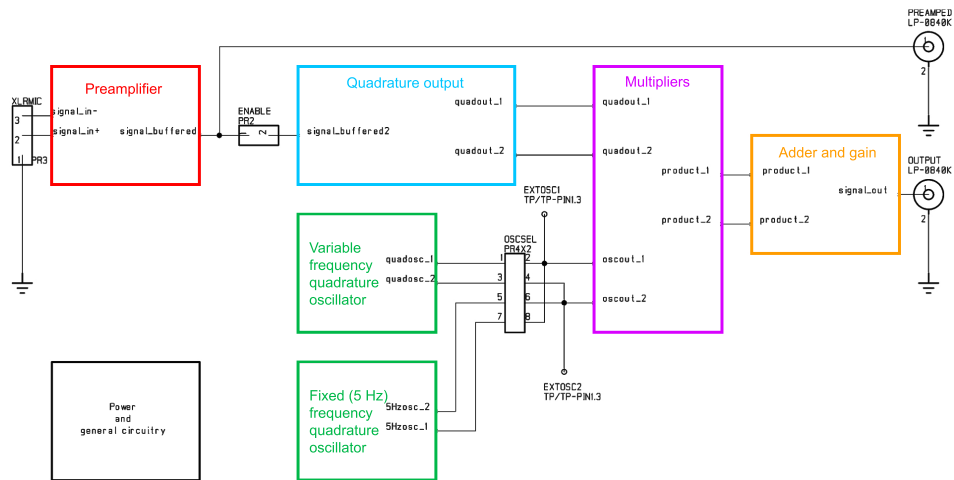
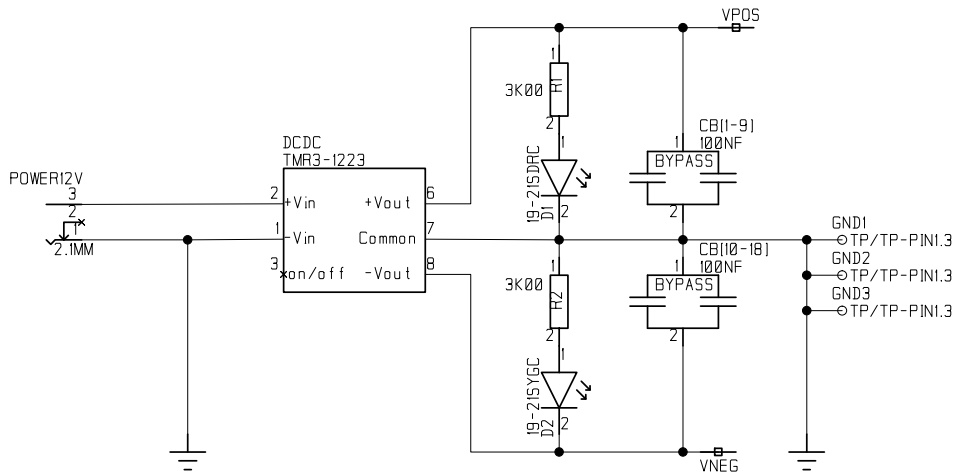


Figure 3.2: The top schematic of the frequency shifter. All sub-circuits are separated in boxes for better readability and overview, and color coded to be comparable to Hartley Jones’s original schematic seen in figure 3.1 on the previous page.

In figure 3.2 the same sub-circuits and elements are seen, save the offset adjustment section, as in Hartley Jones’s original circuit of 1973. Figure 3.2 shows the top view of TFS, and the sub-circuits are color coded in the same way as in figure 3.1 on the preceding page for easier comparison of the two circuits. As described in the theory section, TFS is a realization of equation 2.8 on page 9, and most sections in this chapter will therefore reflect a part of that equation.

3.1 Power and general circuitry

In figure 3.3 the power converting and distribution system can be seen. A regular 12 Volt AC-adaptor is used to power the circuit, and a *TMR – 3 1223* from Traco Power converts the 12 Volt DC to +9 Volt and -9 Volt DC. These voltages are distributed in the circuit as *VPOS* and *VNEG* respectively. The two LEDs, *D1* and *D2*, are for indicating power at the two rails. All active components in the whole circuit have bypass capacitors connected between the power pins and ground. This is to ensure minimal power fluctuations for the components. On the PCB the bypass capacitors are positioned as close to the power pins as possible, but for convenience and readability of the schematic, they are drawn in in the power and general circuitry part of the schematic. Three test points for ground connection, *GND1*, *GND2* and *GND3*, have similar placement due to the same reasons, though they are spread out on the PCB layout for better testability.



*Figure 3.3: The power schematic of TFS PT4.1. To the left a 12 V AC-adaptor is connected, and the DC/DC converter converts the voltage to ± 15 V. These two voltages are distributed to the rest of the circuit as *VPOS* and *VNEG*. The two light emitting diodes serves only as power indication of the two rails, and bypass capacitors is also placed in this sub-circuit for a better readability of the schematics. The same goes for the three ground test points.*

3.2 Preamplifier

This sub-circuit has two main purposes:

- Convert the incoming balanced microphone signal to a more manageable single ended signal
- Amplify the signal to match the signal levels coming from the quadrature oscillator

The first of these two tasks is handled by the integrated circuit *INA134UAE4* (*DIFFREC – B*) made by Texas Instruments, as it converts the signal from balanced to unbalanced. The second task is handled by a two stage amplifier. The first stage is a non-inverting op-amp amplifier, with a set gain, A_v , given by

$$A_v = \frac{R_f}{R_g} + 1 = \frac{R4}{R5} + 1 = \frac{100 \text{ k}\Omega}{5.1 \text{ k}\Omega} + 1 \approx 20.6 \quad (3.1)$$

The second stage have a similar function, but here the gain can be adjusted with *PREAMPGAIN*. The maximum total gain of the preamplifier is 425. This is enough to raise the small signal from a microphone, for example 10-40 mV_{pp}, to the signal levels used elsewhere in this circuit. These signal levels vary, but are in the magnitude of about 10 V_{pp}. *C1* functions as a DC block, removing any DC offset from the signal.

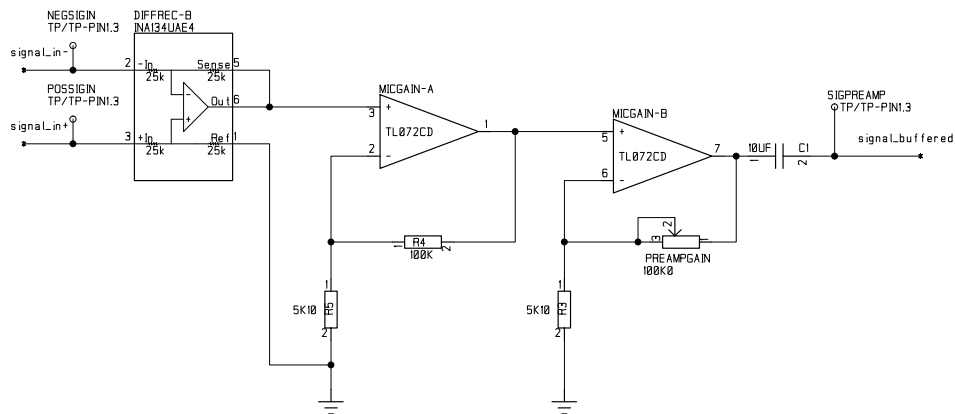


Figure 3.4: Schematic of the preamplifier section of *PT4.1*. The balanced microphone signal is converted to a single ended signal, and amplified in a two stage non-inverting op-amp based amplifier. The gain is adjustable by potentiometer *PREAMPGAIN*. The two test pins *NEGSIGIN* and *POSSIGIN* are mainly for connecting a general signal source to the circuit for testing purposes.

3.3 Quadrature output

As seen in equation 2.8 on page 9, in addition to the sine input signal, we also need a cosine version of the same signal:

$$V_{out}(t) = A \left(\sin(2\pi ft) \cos(2\pi \Delta ft) + \cos(2\pi ft) \sin(2\pi \Delta ft) \right)$$

The relation between sine and cosine is given by

$$\cos(x) = \sin\left(x + \frac{\pi}{2}\right) \quad (3.2)$$

This sub-circuit contains an inverting amplifier and a RC network, working as a broad band phase shifter for the signal from the preamplifier. The phase relationship between the two outputs, *quadout_1* and *quadout_2*, is ideally fixed at 90°. Some variation must be expected. The sub-circuit is an exact copy of Hartley Jones's, except for C4 and C8. These two capacitors were omitted during testing, and were replaced by 0 Ω resistors making the circuit identical to Hartley Jones's. For a detailed description of the circuit see the appendix of Hartley Jones's article [4].

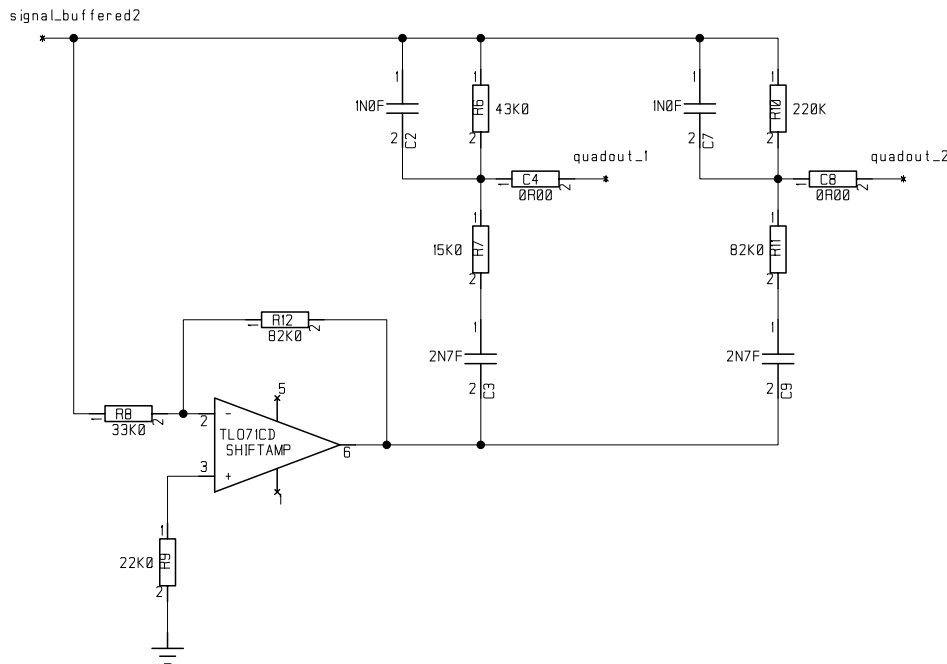


Figure 3.5: Schematic of the quadrature output section of PT4.1. This sub-circuit produces two different versions of the signal coming from the preamplifier section. The two outputs are about similar in amplitude and waveform, but are in quadrature which means there is a 90° phase difference between them.

3.4 The two quadrature oscillators

An oscillator is a circuit that produces a signal without an external source. There are numerous ways to build an oscillator, and choosing one design depends on the usage of the final circuit and any requirements of waveform, frequency and amplitude. In TFS a special type of oscillator called a quadrature oscillator, was needed. It produces not only one, but two waveforms. These are ideally identical in shape, frequency and amplitude, but with a phase relationship of 90° between them. The two outputs are said to be in quadrature. Equation 2.8 on page 9 shows that we need both the sine and cosine version of the shifting frequency, Δf

$$V_{out}(t) = A \left(\sin(2\pi ft) \cos(2\pi \Delta ft) + \cos(2\pi ft) \sin(2\pi \Delta ft) \right)$$

The shifting frequency are therefore produced by the quadrature oscillator. Hartley Jones's version of the frequency shifter produced a frequency shift of approximately 5 Hz. With TFS the possibility to test the circuit within a wide range of frequency shifts, was wanted. A quadrature oscillator with an easily adjustable oscillating frequency was therefore needed. To be able to compare the new circuit to the old one, two oscillators were included, one with an adjustable frequency and one copy of Hartley Jones's 5 Hz oscillator.

3.4.1 Adjustable frequency quadrature oscillator

There exists many well known designs for quadrature oscillators, but common for most of them is that the oscillating frequency cannot easily be adjusted on the fly. That is why the adjustable frequency quadrature oscillator was built. The oscillator consists of a well documented Wien bridge oscillator, with a RC phase shifting network, similar to that in the quadrature output section, attached to it. The frequency of the oscillator is adjustable for generating different frequencies for different frequency shifts. The Wien bridge oscillator produces a sine wave with very low distortion, and is well suited for low frequencies. The oscillating frequency is determined by only two RC circuits. With fixed and equal capacitor values, an external dual-potentiometer is used to adjust the frequency. The oscillating frequency, f , is determined by

$$f = \frac{1}{2\pi RC} \quad (3.3)$$

where R is the resistance set on the two resistors in the dual potentiometer and C is the capacitance of $C5$ and $C6$. The dual potentiometer should be a $2 \times 50 \text{ k}\Omega$ or $2 \times 100 \text{ k}\Omega$ dual potentiometer to get the desired frequency range. PT4.1 used a $2 \times 100 \text{ k}\Omega$ one. As the frequency is changed it may be required to adjust the amplitude of the oscillator with $OSCGAIN$, to maintain oscillation.

“Capacitors” $C13$ and $C14$ were eliminated during the testing process, and replaced by 0Ω resistors. The two signals going out of the phase shifting network, $quadosc_1$ and $quadosc_2$, are amplified to better suit the rest of the circuit.

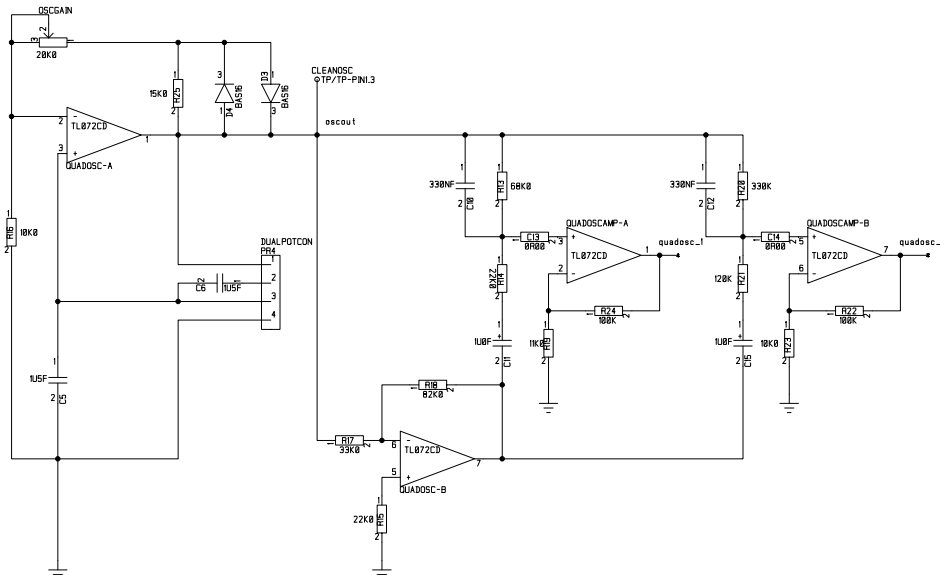


Figure 3.6: The adjustable frequency quadrature oscillator section of PT4.1. A $2 \times 100 \Omega$ dual potentiometer should be connected to DUALPOTCON. $OSCGAIN$ may need to be adjusted as the frequency is increased.

3.4.2 Fixed frequency quadrature oscillator

This quadrature oscillator is a phase shift oscillator, also with 90° phase difference between the two outputs. The oscillating frequency is given by the values of RC pairs $R26-C16$, $R27-C17$ and $R28-C18$. As we can see the RC components values are not equal for each pair, and the oscillating frequency can therefore not easily be determined. The maximum and minimum resonance frequency for the different RC pairs can be found by applying equation 3.3 on the previous page to each of the pairs. The resonance frequency of $R27-C17$ and $R28-C18$ are equal and given by

$$f_1 = \frac{1}{2 \cdot \pi \cdot R27 \cdot C17} = \frac{1}{2 \cdot \pi \cdot 270 \cdot 10^3 \cdot 100 \cdot 10^{-9}} \approx 5.9 \text{ Hz} \quad (3.4)$$

The resonance frequency of $R26-C16$ are given by

$$f_2 = \frac{1}{2 \cdot \pi \cdot R26 \cdot C16} = \frac{1}{2 \cdot \pi \cdot 330 \cdot 10^3 \cdot 100 \cdot 10^{-9}} \approx 4.8 \text{ Hz} \quad (3.5)$$

and the final oscillating frequency was assumed to lie nearby those frequencies. The oscillator in prototype 4.1 of TFS oscillated at 4.5 Hz. The two op-amps used are in one physical package of *TL072*, a JFET input op-amp from Texas Instruments. Other than that the circuit is an exact copy of the one described by Hartley Jones [4], and will not be further discussed here.

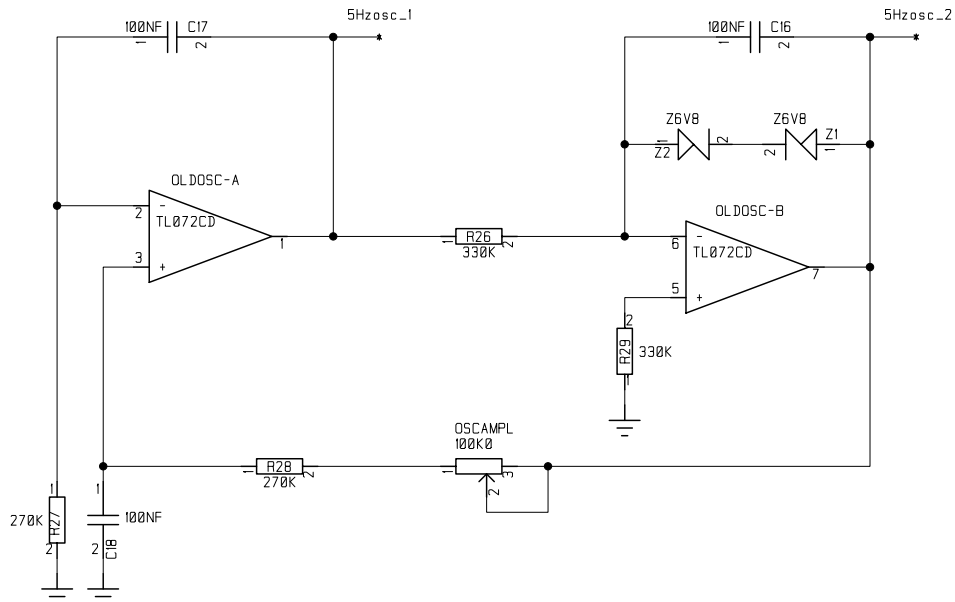


Figure 3.7: Schematic of the fixed frequency quadrature oscillator section of *PT4.1*. This is an exact replica of the circuit proposed by Hartley Jones, and oscillates at about 5 Hz, depending on the variance of the component values of $C16$, $C17$, $C18$, $R26$, $R27$ and $R28$. The oscillator gain can be adjusted with potentiometer *OSCAMPL*.

3.5 Multipliers

Equation 2.8 on page 9 shows that multiplications of the sine and cosine parts of the input signal, with the cosine and sine of the shifting frequency respectively, are needed in the frequency shifter. This is achieved by using two analogue multipliers. Analogue multipliers can be used in many different ways and setups, but the most basic application is used here. It simply takes two signals, multiply them, and give the result. The multipliers Hartley Jones used, Motorola *MC1495L* or Silicon General *SG1495D*, were discontinued, and replaced in TFS PT4.1 by *AD633* from Analog Devices. The specifications for the two are similar when it comes to power supply and amplitude values, and the multiplier could therefore be used without any other alterations to the rest of the original circuit. The *AD633* also feature a high impedance summing input *Z* which is not used here, other than for offset corrections. The offset adjustment circuitry proposed in the data sheet of *AD633* was used, rather than the one described by Hartley Jones. Other than that the multiplier section of TFS and Hartley Jones's design is identical. The four potentiometers, *M1X2OFFSET*, *M1Y2OFFSET*, *M2X2OFFSET* and *M2Y2OFFSET* are used for calibrating purposes only, as described in chapter 5.3.1 on page 38.

The four two-component RC circuits seen in the diagram, are serial Zobel-filters, but acts here as low pass filters. The RC circuit's purpose is to lead any high frequency noise the circuit may have picked up or generated to ground, in stead of to the multipliers inputs.

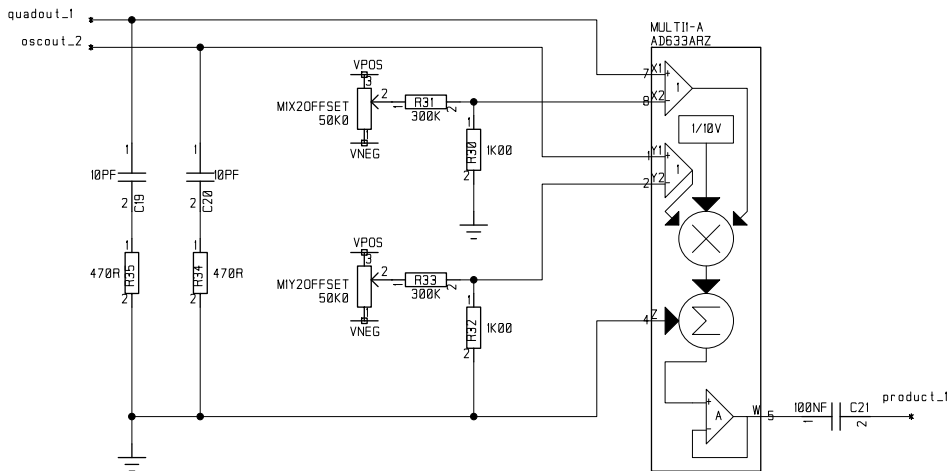


Figure 3.8: Schematic of half the multiplier section of PT4.1. The schematic shows the offset correction at the inputs of the multipliers, as well as the low pass filters leading high frequency noise from the preamplifier and oscillator sections to ground. The other multiplier section that handles the other two signals, is not shown here, but is practically identical as this section.

3.6 Adder and gain

The last sub-circuit of the signal chain in TFS PT4.1 makes sure the outputs from the multipliers are added together, and that the result are amplified to line level. This is achieved by using an inverting summing amplifier setup, using a *TL071* op-amp from Texas Instruments. The gain of the summing amplifier is adjustable by potentiometer *ADDERGAIN*, and is used in the calibrating process. The circuit also features a balance potentiometer, *BALANCE*. This should never need to be adjusted after the calibrating of the circuit. Se chapter 5.3.2 on page 39 for further details.

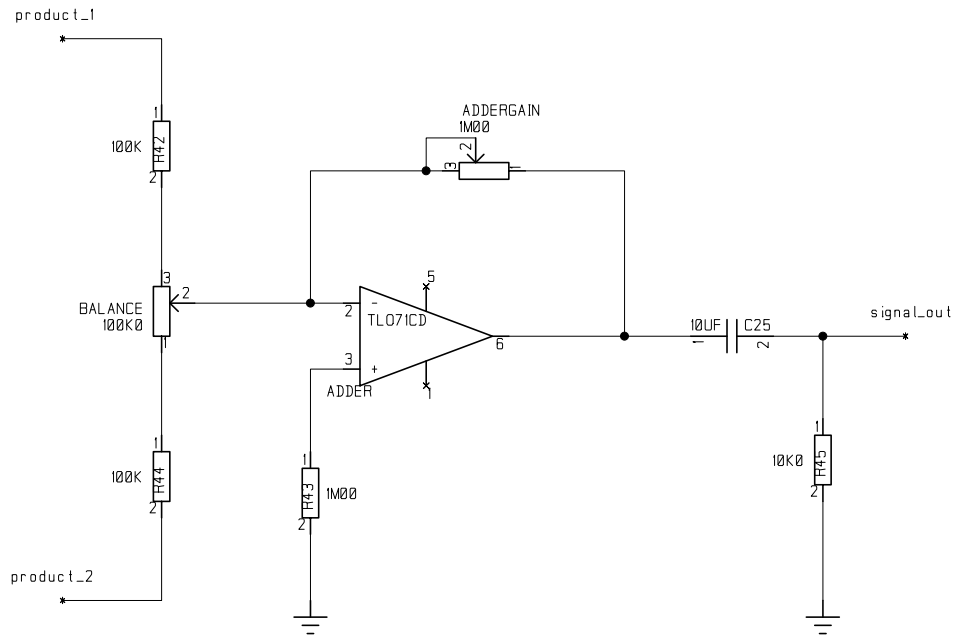


Figure 3.9: Schematic of the summing amplifier and gain section of PT4.1. The weighting of the two signals coming from the multipliers, can be adjusted with potentiometer *BALANCE*. The final gain of the circuit is adjusted with *ADDERGAIN* to make the output match the required line level wanted.

Chapter 4

Equipment

In this chapter all equipment used in the development, testing and the various verifying processes of TFS, are described. The chapter is sectioned in two parts; Hardware and Software.

4.1 Hardware

A lot of different equipment were utilized during the development and testing of TFS. Below is a short description of each and every hardware piece used in this process.

4.1.1 PA amplifier

The amplifier used in the testing of TFS PT4.1 was custom built by the Electronics Workshop at the Department of Physics at UiO. The model is called POW4 and is a MOS-FET amplifier. Not all specifications were available, but the most important ones are seen in table 4.1. The output of TFS was connected to the right RCA line in connector at the back of the unit.

Table 4.1: Specifications, measured in November 2000, of the POW4 custom amplifier used in testing.

<i>Attribute</i>	<i>Testing conditions</i>	<i>Value</i>
Input sensitivity	1 W/8 Ohm	130 mV
Gain	1 W/8 Ohm	+26 dB
Frequency response	1 W/8 Ohm/<10 Hz - 90 kHz	-1.5 dB
Output power	1% THD/1 kHz/8 Ohm	75 W

4.1.2 Audio Interface

A M-Audio Fast Track Ultra USB audio interface, a high quality, low noise sound card, were used for the measurement of reverberation time, RT₃₀.

4.1.3 Digital Multimeter (DMM)

For measuring of DC currents and potentials as well as resistance measurements, two copies, #16 and #20, of model 127 multimeter from Tecpel were used. The model is easy and quick to use, and is able to measure within a large range of values. The testing leads used were normal lab wires with hooks.

4.1.4 Function Generator

For testing and calibrating purposes, TG550 Function Generator from TTI were used for signal generation, and as a comparison reference for the oscillator. It were also used to calibrate the gain of internal circuitry in TFS PT4.1 before the final testing.

4.1.5 Microphone

The microphone used for testing was Shure SM58[®]. It is a cardioid dynamic microphone with a 3 pin balanced XLR output.

4.1.6 Oscilloscope

Through the testing, calibrating and verifying processes of TFS, the oscilloscope TDS1002 from Tektronix was used. This is a two channel digital oscilloscope with 60 MHz bandwidth, capable of sampling rates up to 1 GS/s. Two HZ36 from Hameg were used as test probes. These each have an internal variable capacitance which were calibrated using the oscilloscope's internal 1 kHz oscillator.

4.1.7 Power Supply

Two power supplies were used for the testing of the different prototypes. PT1 used a ± 15 V LAB-supply manufactured by the Electronics Workshop at the Physics Department at UiO. The later prototypes, PT2, PT3 and PT4.1 had DC/DC converters installed, and used a Vanson[®] model no. 230124R, a 12 V regulated AC/DC adapter as power source.

4.1.8 Sound pressure level measurement

Brüel & Kjær Type 2226, Serial Number 1014602, were used for all sound pressure level measurements. This is an integrating impulse sound level meter with measuring range from 20 to 140 dBA at a resolution of 0.5 dBA. Two different settings were used in the measurements, "Fast" and "Slow", within the two ranges 50 - 90 dBA and 80 - 120 dBA.

4.1.9 Speakers

A pair of Sound Master model C5121 PA speakers, 100 W speakers, 8 Ohm impedance, were used when testing TFS PT4.1 in Store Fysiske Auditorium.

4.2 Software

4.2.1 Adobe Photoshop CS3

Graphics tools for various photo and illustration work.

4.2.2 Audio Precision, System One

Audio Precision is the industry standard for testing audio equipment. It was used in this thesis to document the frequency specters and the total harmonic distortion of the waveforms produced by the oscillator prototypes.

4.2.3 LTspice IV and PSpice 9.1, student version

LTspice IV is a free-to-use SPICE software from Linear Technology, well suitable for simulating circuits using integrated circuits from Linear Technology. It was used for testing parts of the circuit design in the development process. PSpice was used in this thesis to simulate small parts of the design through the development process of the quadrature output and oscillator.

4.2.4 MATLAB R2011b

Math simulation tool by Mathworks, used for a mathematic simulation of the circuit and its response to different signals, especially the influence of DC-offsets on the multiplying calculations.

4.2.5 Sonic Visualizer

Graphical audio analysis tool made by Chris Cannam and Mary Queen, University of London. In this thesis used for producing sonograms for the different sound samples used.

4.2.6 TeXnicCenter 1.0

Word processor and \LaTeX tool for writing this report.

4.2.7 Zuken CADSTAR 13.0

Zuken's CADSTAR Design Editor were used for drawing the schematics and layout of the PCB. The routing editor PREditor XR was used for the routing process and some of the PCB layout.

Chapter 5

Methods

The different methods used when constructing, calibrating and testing the frequency shifter will be described in this chapter. It is divided into smaller sections which goes through the setups and methods for each and every procedure used through the development. In addition methods for the production, measurement and calibrations are described.

5.1 Redesigning Hartley Jones's frequency shifter

This section of the chapter is not critical for the understanding and usage of TFS, but could maybe give a better understanding for the choices made for the final PT4.1. When starting the reproduction and redesign process for the circuit in this thesis, there was no doubt that there were going to be some changes from Hartley Jones circuit from 1973. First of all some of the electrical components were no longer available. In addition to that some new features were desirable in the new design. This chapter has a short description of the development up to and including the current version, PT4.1. Earlier versions include PT1, PT2 and PT3. PT1 was nearly a replica of the circuit constructed by Hartley Jones. PT4 only exist as a schematic, and was never produced, as some minor changes were done right before production. The final version of the frequency shifter is therefore PT4.1.

Each subsection will illuminate the process and difficulties encountered while developing a sub-circuit of the grand design. The starting point for design of each function, was the corresponding sub-circuit from Hartley Jones design.

5.1.1 Power supply and distribution

While Hartley Jones built a separate power supply to power his circuit, building and developing a power supply was not one of the goals of this thesis. The first prototype, PT1, used an external LAB-supply as power source, and the circuit had connections for ± 15 Volt and ground directly. While this was time saving for the first prototype, it was not a practical solution for testing situations and never meant to be the final design. All later prototypes

used a regular 12 Volt AC-adaptor, with a suitable DC/DC converter for that specific circuit. As PT2 and PT3 were based on a different analogue multiplier than the first and last prototype, they used *MEA1D1205SC* from muRata, a 12 Volt to ± 5 Volt DC/DC converter. This was to see if a low cost alternative to the more expensive DC/DC converters were usable, without getting too much noise problems. The switching frequency of *MEA1D1205SC* was stated to be 85 kHz, and during testing no switching noise problems were detected.

PT4.1 was based on circuitry running at ± 15 Volts, and a different DC/DC converter was needed. *TMR 3 – 1223* from TRACOPOWER was chosen. Switching frequency was given to be 100 kHz, but no hearable switching noise was detected in practical use here either.

5.1.2 Preamplifier

The preamplifier section of Hartley Jones's design was aimed to handle single ended signals at line level. Even though Hartley Jones presents a possible modification to the circuit to handle microphone signals directly, the circuit could not in it's original form handle the balanced signals produced by most professional microphones used today. In practical use it was found easier if the frequency shifter could be connected between the microphone and the amplifier without any other electronics involved. Because of this the preamplifier section was redesigned.

To convert the balanced signal to a single ended signal, *INA134UAE4* (*DIFFREC*) from Texas Instruments, a specialized circuit just for that purpose, was used. The reason for this was twofold. It was easy to set up and use, and it minimized the number of components, so that the final product could be made small if desired. For the amplifying part a two stage non inverting amplifier structure was chosen, utilizing the low-cost all round op-amp *TL072*, from Texas Instruments. With a unity gain bandwidth of 3 MHz, two amplifiers in cascade were needed to ensure a signal gain of 400 (approximately 52 dB) within 20 - 20,000 Hz to be without phase distortion. As the output signal amplitude varies from microphone to microphone, an adjustable gain were chosen for the last amplifier. This can be adjusted with the potentiometer *PREAMPGAIN*. Only minor problems were encountered during the development of this part of the circuit, one being that the potentiometer had to be flipped, so that the use would be intuitive (more gain as turning clockwise). The only challenge in this sub-circuit was to get the amplification right, so that the range of the potentiometer used could be fully utilized. This was found experimentally.

5.1.3 Quadrature output

The quadrature output section in PT4.1 is identical to Hartley Jones's original circuit. Some experiments were done by adding DC blocking capacitors to the outputs during testing, but these were removed in the final version. More on this in the subsection 5.1.5 on page 33 where the multipliers are described.

5.1.4 Quadrature oscillator

One of the most interesting, and time consuming, parts of the development of TFS was the design of the quadrature oscillator section. In the final version, PT4.1, there are two oscillators. The 5 Hz fixed frequency quadrature oscillator is identical to the one Hartley Jones utilized. It will not be discussed in detail here, but the development of the other, the adjustable frequency quadrature oscillator, starts with a study of the phase shifting oscillator Hartley Jones used.

Hartley Jones's stated in his article that: "...in most applications it turns out that a shift of 5 Hz is sufficient, no further advantage being gained by greater shifts". This was something that the author wanted to confirm in a testing scenario. As mentioned in chapter 3 on page 11 the phase shift oscillator Hartley Jones used, operates at a frequency given by the value of three RC-pairs. To be able to vary the frequency of this oscillator, one needs to accurately change three resistors, or three capacitors, at the same time. Adjusting three potentiometers to change one attribute of the circuit, seemed like a difficult and impractical solution for a simple testing process. One could of course construct many independent oscillators, one for each frequency that was needed. This was not desirable for two reasons. For one it was not considered an elegant solution, and second, there was no guarantee that the oscillators would oscillate at the exact calculated frequency. This lies in the nature of the construction of the phase shifting quadrature oscillator and the tolerance of electrical components. The oscillator used by Hartley Jones was then rejected as a candidate for the final circuit. With the studying of the quadrature output section of the circuit, came the idea that one could build the much simpler Wien bridge oscillator, and then run the signal through a RC network, similar to the quadrature output section, to create a sort of quasi quadrature oscillator.

The Wien bridge oscillator

The Wien bridge oscillator was chosen as it have low distortion, is well suited for low frequency applications and the author had some experience with the oscillator from earlier projects. There exists many different design variations of the Wien bridge oscillator, and two slightly different designs were selected as candidates. The main difference between the two designs were that the first used an incandescent light bulb for stabilizing and amplitude control. The other used a pair of diodes for the same purpose. The light bulb oscillator design was taken from Mancini [3] and can be seen to the left in figure 5.1 on the following page. The diode oscillator was taken from Paynter [8], and can be seen to the right in figure 5.1 on the next page.

Both circuits were built and tested on a breadboard. There was some trouble getting the bulb based oscillator to oscillate. Several op-amps were tested, but with no luck. The design of the light bulb stabilized oscillator was assumed correct, so the specifications of the light bulb was suspected as the fault. The lamp tested was rated 12 V at 200 mA. The specifications of lamps to be used for oscillator stabilizing is not well documented. Moore [7]

claims that the current rating of the lamp has to be equal or less than 50 mA, and two new bulbs were therefor tested. One rated 6.5 V 40 mA, the other 3.5 V 10 mA. Both of these made the oscillator work, and the latter was used in the rest of the experiments.

There were several issues that needed to be considered when choosing one oscillator design from the two candidates. The light bulb stabilized oscillator was thought to produce less distorted waveforms, but to be fragile and sensitive to mechanical vibrations as the filament vibrates and the light bulb is made of glass. The diode stabilized version was assumed more robust, but documentation of distortion to be expected, was not found. Both circuits were therefore tested in Audio Precision (AP) to find total harmonic distortion + noise (THD+N) for the two circuits. As Audio Precision was made for testing at audio frequencies, it was not possible to test at the target frequency band of 3 - 20 Hz. Each circuit was therefore tested at two different frequencies, ~ 1 kHz and the lowest possible to test with AP, ~ 30 Hz. Fast Fourier Transform (FFT) plots were produced to look at the frequency components of the two signals, and THD+N calculated by AP was 0.3 % for the bulb oscillator and 1.1 % for the diode oscillator. Figure 5.2 on the facing page, 5.3 on the next page, 5.4 on page 30 and 5.5 on page 30 show the frequency components of the waveforms produced by the two oscillators at said frequencies. The value of the X-axis is frequency in hertz, and on the Y-axis the dBr value is shown, calibrated to the main frequency at 0 dBr. 1.1 % THD+N was found to be sufficient, and the diode oscillator was selected to be part of the quasi quadrature oscillator.

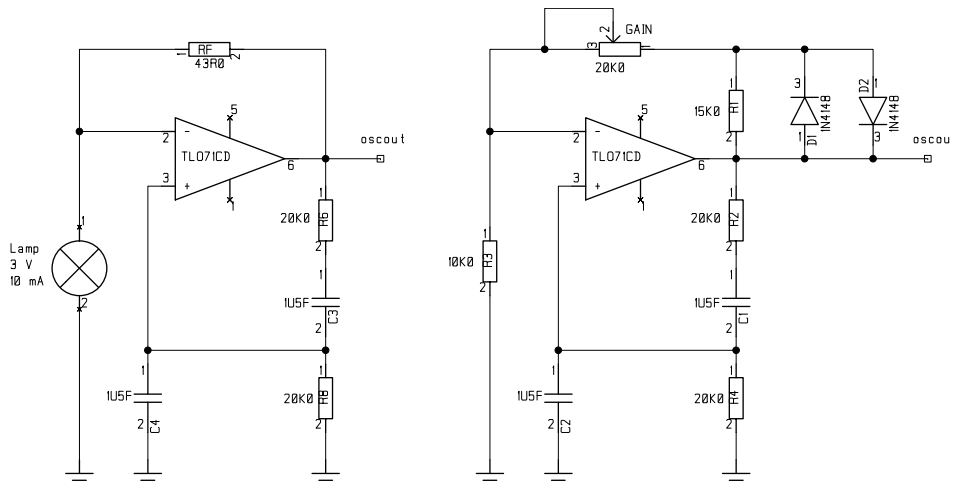


Figure 5.1: The circuit to the left shows a simple Wien bridge oscillator design, where a incandescent light bulb is utilized for stabilizing amplitude control. The oscillator was rejected as candidate for the quasi quadrature oscillator. The Wien bridge oscillator to the right uses two diodes for amplitude control. This variant became part of the final design of the quasi quadrature oscillator. Schematic sources: Source: Mancini [3] (left) and Paynter [8] (right).

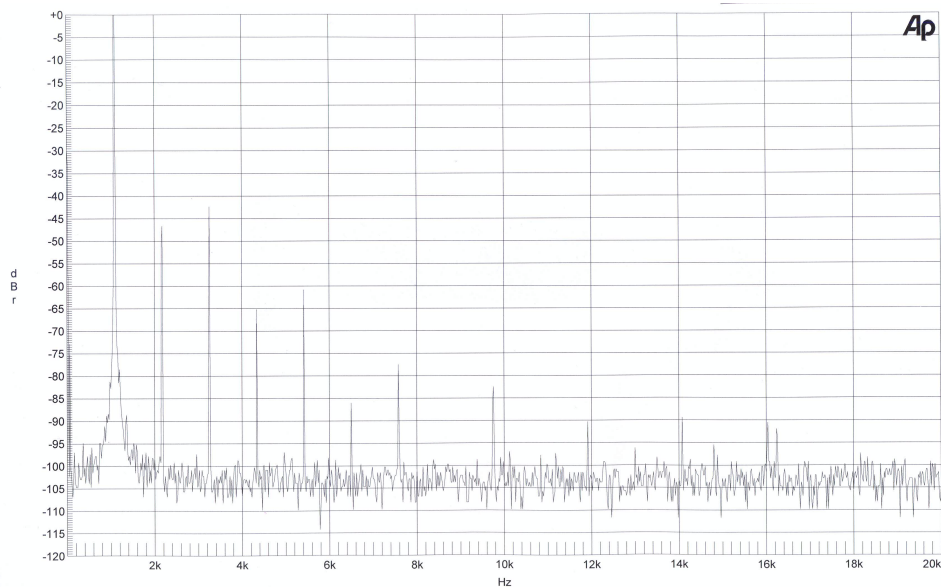


Figure 5.2: Frequency spectrum analysis, 20 - 20,000 Hz, of the light bulb stabilized oscillator running at about 1 kHz. The scale is calibrated to show the main generated frequency, 1.09 kHz, as 0 dBr. The harmonics of the signal can be seen as spikes with attenuated magnitudes with increasing frequency.

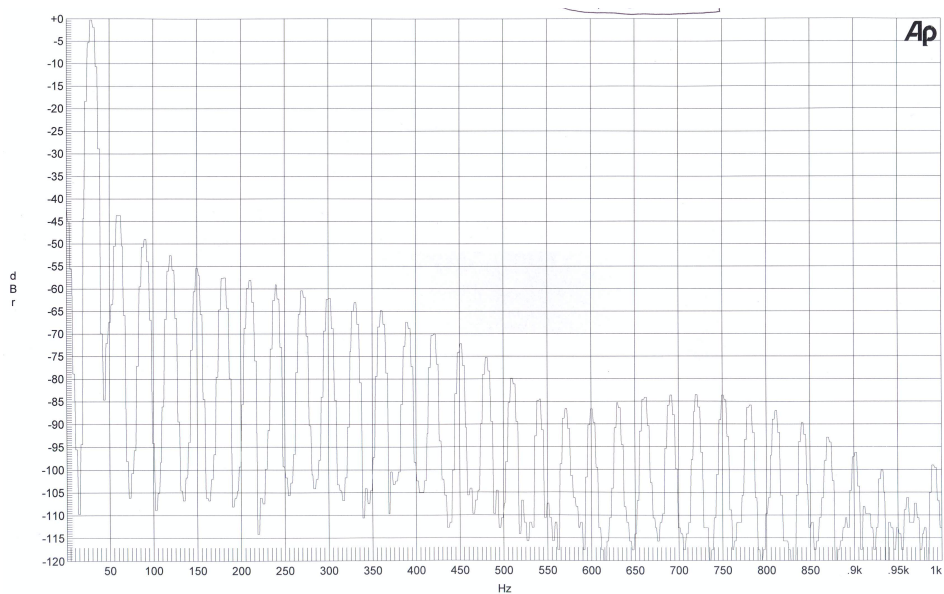


Figure 5.3: Frequency spectrum analysis, 20 - 1000 Hz, of the light bulb stabilized oscillator running at about 30 Hz. The scale is calibrated to show the main generated frequency, 30.04 Hz, as 0 dBr. The harmonics of the signal can be seen as spikes at 60 Hz, 90 Hz and so forth.

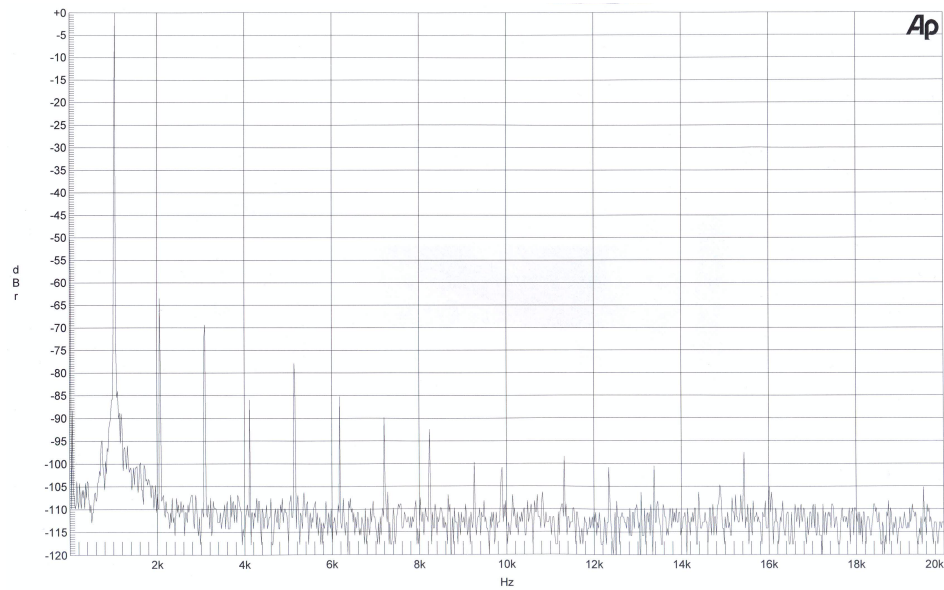


Figure 5.4: Frequency spectrum analysis, 20 - 20,000 Hz, of the diode stabilized oscillator running at about 1 kHz. The scale is calibrated to show the main generated frequency, 1.03 kHz, as 0 dB. The harmonics of the signal can be seen as spikes with attenuated magnitudes with increasing frequency.

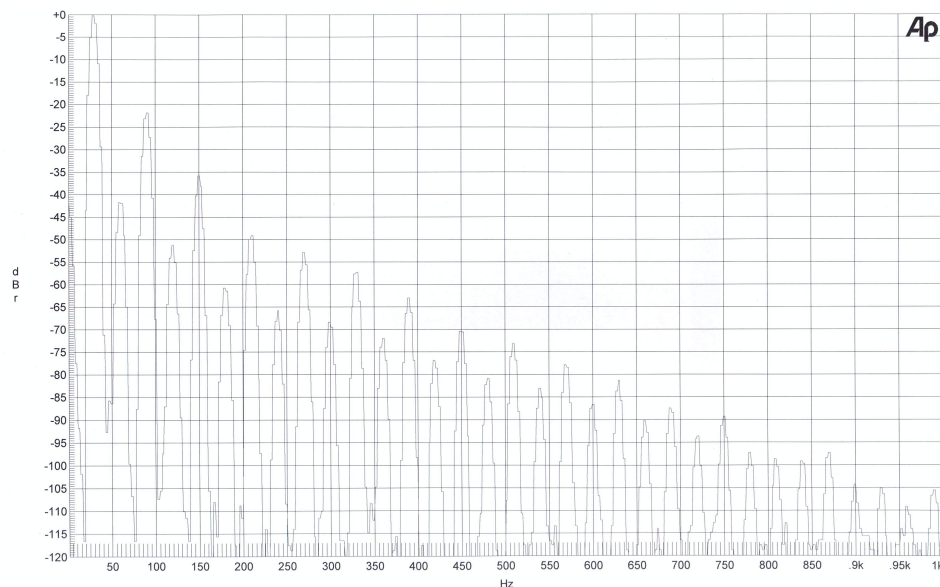


Figure 5.5: Frequency spectrum analysis, 20 - 1000 Hz, of the light bulb stabilized oscillator running at about 30 Hz. The scale is calibrated to show the main generated frequency, 30.10 Hz, as 0 dB. The harmonics of the signal can be seen as spikes at 60 Hz, 90 Hz and so forth.

The phase shifting RC network

When designing a RC network similar to the one used in the quadrature output sub-circuit, it was decided to find the values of the components experimentally, by utilizing SPICE simulations. PSPICE and LTSPICE were used for this. Starting out with the component values from the quadrature output, several simulations and measurements were done with multiples of the starting component values. The simulation frequency range spanned from 3 to 15 Hz. During the testing process an Excel spreadsheet was developed to register all timing data of the curves. The spreadsheet also calculated timing delays into corresponding phase delay based on the current frequency, as well as deviations from the desired 90° phase shift. After a good setup were found, the simulation component values were exchanged with values that were available for physical SMD components. Then the circuit was constructed on a breadboard, and tested to verify the simulated results. As seen in table 5.1, the RC network was well suited for phase shifting of frequencies from about 5 Hz and upwards.

Table 5.1: The table shows measured phase delays, calculated corresponding phase shifts and deviations of the breadboard model of the phase shifting RC network of the quasi quadrature oscillator. All deviations are in relation to the target phase shift of 90°.

<i>Oscillator frequency (Hz)</i>	<i>Measured delay (ms)</i>	<i>Calculated phase shift (°)</i>	<i>Deviation from 90° phase shift (%)</i>
15	18,00	97,20	8,00
14	19,20	96,77	7,52
13	20,40	95,47	6,08
12	21,60	93,31	3,68
11	24,40	96,62	7,36
10	25,60	92,16	2,40
9	28,80	93,31	3,68
8	31,20	89,86	-0,16
7	34,00	85,68	-4,80
6	39,00	84,24	-6,40
5	46,00	82,80	-8,00
4	55,00	79,20	-12,00
3	72,00	77,76	-13,60
2	104,00	74,88	-16,80
1	244,00	87,84	-2,40

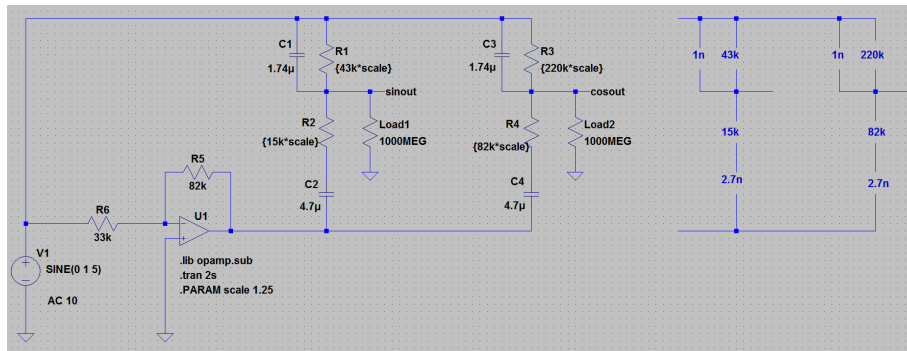


Figure 5.6: An early SPICE schematic for simulating the phase shifting capabilities with different scaling of the component values.

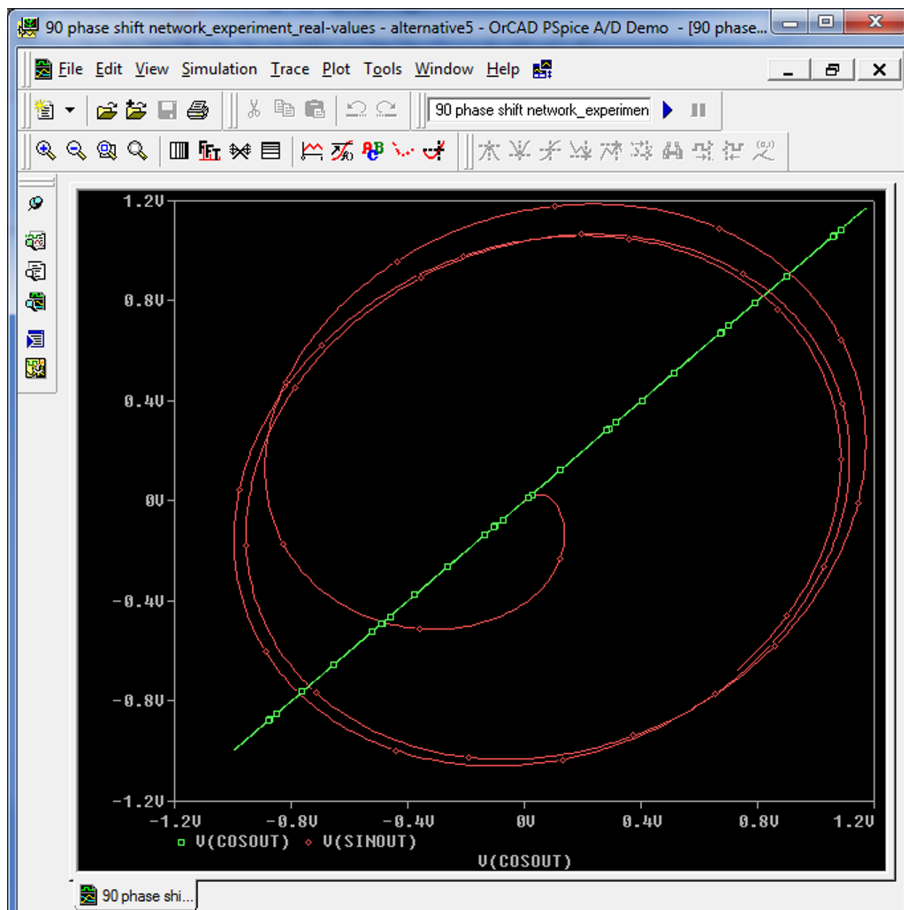


Figure 5.7: During the simulations for finding the correct component values for the phase shifting circuit, it was efficient to look at a XY-plot of the outputs, and check for “roundness”. Exactly 90° phase difference equals a circle if the axis have equal setups. After a promising model was found, accurate measurements of the time plot were done with cursors.

5.1.5 Multiplier section

The development of the multiplier section was thought to be quite straightforward, as the construction was simple. The application notes for the multipliers tested in this thesis, showed the most basic use of the multiplier, namely for multiplying two single ended signals. Nevertheless, the multiplier section was the section that caused the most frustration during the development of TFS.

In PT1 *AD633* from Analog Devices was used. This is a low-cost four quadrant multiplier, and was thought of as a good starting point, as it complied with the same signal amplitudes as the multipliers Hartley Jones used in his circuit. When PT1 was tested, noise problems occurred, and a better more expensive multiplier, *AD835* from the same manufacturer, were chosen instead. The whole design of the circuit had to be redone to comply with power supply at ± 5 Volt instead of ± 15 Volt, and signal levels at ± 1 Volt rather than ± 10 Volt. That new circuit became PT2, but when tested, it performed poorly. Huge amplitude modulation judged the circuit useless for it's purpose. While searching for the error causing the problems, two irregularities/error were found. One op-amp had been flipped during conversion to the new design, meaning the inverting and non-inverting inputs had switched places in the schematic. This resulted of course in an overload and clipping in one of the sub-circuits. The other thing found was huge DC offsets on the multiplier inputs. There was no information in the data sheet of the *AD835* about it setting up its own offsets, so it was believed to come from other sub-circuits of the design. A MATLAB simulation program was made to see what influence a possible offset on the multiplier inputs had on the final result. Manipulation of the offset variables showed that small offsets in the multiplication signals would result in large amplitude modulation on the final result. An example of such a simulation can be seen in figure 5.8.

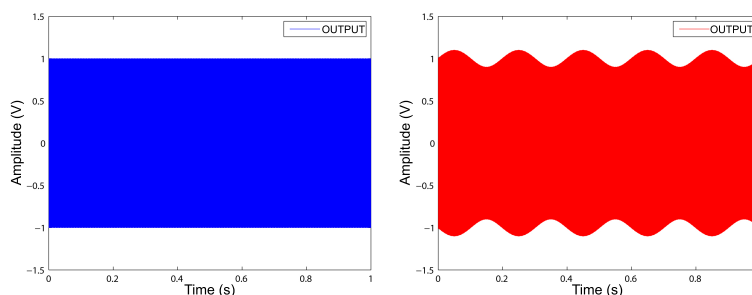


Figure 5.8: The output signal of the simulated TFS. Left (blue): The signal is unmodulated, and have an even amplitude throughout the sample. Right (red): An offset has been introduced at the sine component of the audio signal fed to the multiplier. As the graph shows, the result is huge amplitude modulations of the output signal. With the introduction of offsets at the other multiplier inputs as well, an irregular, but periodic, amplitude modulation of the output is the result.

A new prototype, PT3, was built to solve the challenges encountered in PT2. First of all the op-amp problem was corrected in the schematics. Then DC block capacitors were inserted between the sub-circuits which did not already have this: Both outputs from both the quadrature output and the quadrature oscillator. Now the clipping of the signal was gone, but an offset of ~ 700 mV was still present on all inputs of the multipliers. This was believed coming from the multiplier itself, being a non documented part of the way the multiplier worked internally. A $10\text{ k}\Omega$ resistor was connected between ground and the inputs X1 and Y1 of both multipliers. This was to lead any stray currents to ground, instead of building up a DC offset at the inputs. This lowered the offset to about 70-80 mV, but did not remove it completely. The other inputs of the multipliers, X2, Y2 and Z, previously connected to ground, was connected to ground through a small, $470\ \Omega$, resistor instead. This did not help, the offsets remained unchanged. As a last resort, a DC voltage from an external power supply was injected through $1\text{ M}\Omega$ resistors, at the inputs. This removed the offsets, and the circuit performed as expected. The data sheet of AD835 was closely studied again, but no data was there to suggest any special measures to be taken to avoid offset on the inputs. The design was then discarded, and it was decided to go back to the AD633 multipliers instead.

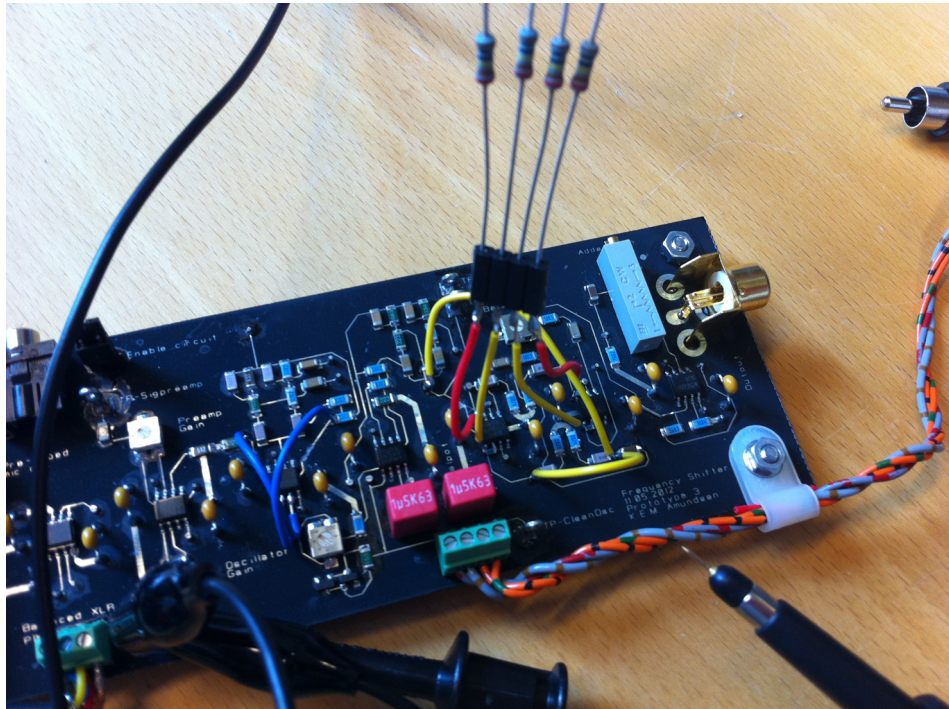


Figure 5.9: Prototype 3 is seen here with all the fixes and workarounds tried for resolving the DC offset problem at the multiplier inputs. In the top we see the four $1\text{ M}\Omega$ resistors used in the DC injection procedure described in subsection 5.1.5 on the previous page.

The following prototype, PT4.1, initially had some offset issues, but the removal of the DC blocking capacitors introduced in the PT3 design, resolved those problems. The circuit, now performing as expected, was declared the final design. The MATLAB code for simulating TFS can be found in appendix A on page 69.

5.1.6 Adder and gain

This sub-circuit was barely changed from the original circuit. Also in this circuit the *TL071* op-amp was used. The structure is an inverting amplifier. From Hartley Jones's circuit the rather large resistor *R43*, 1 M Ω , connected from the non-inverting input of the op-amp to ground, was kept. Originally this was probably present to handle current leakage. Today's op-amps seldom need that structure, and it can most likely be omitted. This is not tested though. To adjust the output impedance, resistor *R45* was added to the circuit .

5.2 Production and assembly of TFS PT4.1

The schematic design and PCB layout were done in Zucken CADSTAR, at first version 12.1 and later 13.0. The PCB was produced from FR4 board at the Electronic Workshop at the Department of Physics at UiO. The board used was a two sided copper board, with photo-resist on both sides. The top and bottom electric layer of the PCB layout was printed on foils, and then transferred to the board using a UV lightning box. The photo-resist was then induced by sodium hydroxide (NaOH, caustic soda). Then the board was placed in the etching chamber for six minutes, to remove the copper not being part of the design. The etching mixture used was sodium persulfate ($\text{Na}_2\text{S}_2\text{O}_8$). When the etching process completed, the remains of photo-resist was removed with Lynol (mixture of toluen, n-butylacetat and ethanol). SUR-TIN was then used to give the copper a thin layer of tin. SUR-TIN consist of sulfuric acid (H_2SO_4), tin(II) sulfate (SnSO_4) and thiourea ($\text{SC}(\text{NH}_2)_2$). This is not the best material in regard of the soldering process later, but at least it is better than oxidized copper, which is the result if the copper is exposed to air for some time.

Following the tining process, holes were drilled, and soldering paste was applied to all pads in the design. All SMD components were placed manually, and then soldered in a vapor phase soldering process. The oven used was Asscon Quicky Vapor-Phase 450, with Galden LS230 vapor phase fluid. Finally the through-hole components, test points, connectors and wires were mounted, and soldered in place by hand. As many different chemicals were used, both goggles and acid resistant apron were used during etching process. The final result PCB can be seen in figure 5.10.

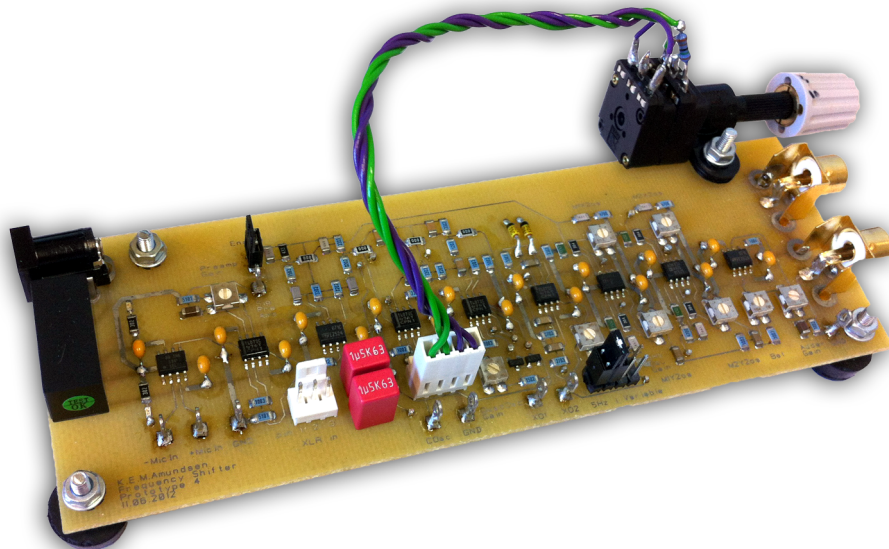


Figure 5.10: Here the final version of the frequency shifter, PT4.1, is assembled and ready for use.

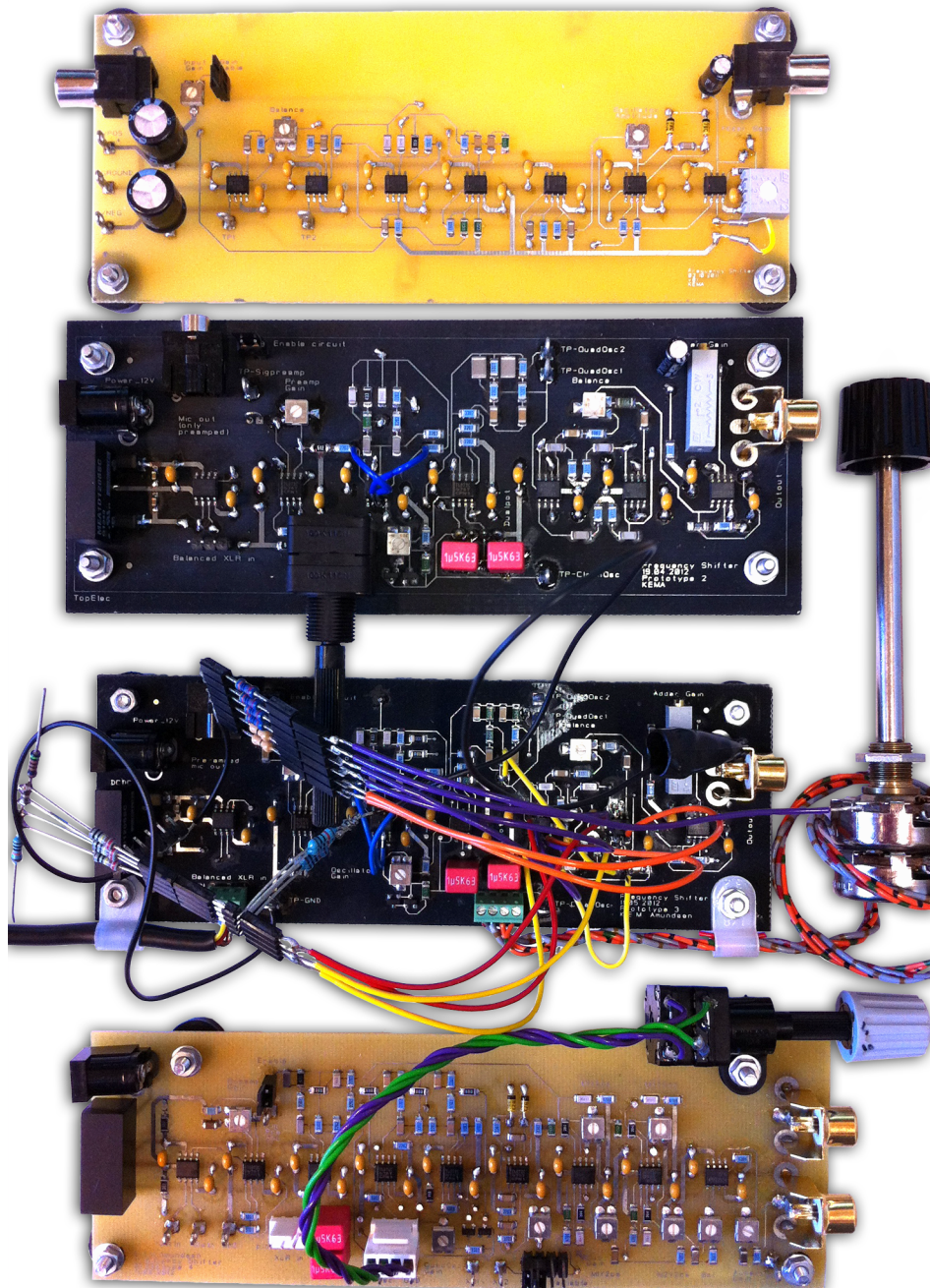


Figure 5.11: Seen here are all the four prototypes developed. From the top down, PT1, PT2, PT3 and PT4.1. The fixes of PT2 can be seen, as well as all the excessive bypasses, fixes and add-ons of PT3.

5.3 Calibrating the frequency shifter

A few simple steps are required to calibrate TFS for optimal performance before usage. The calibration process is divided into three parts: calibrating the multiplier input offsets, adjusting the balance of the summing amplifier and adjusting the preamplifier and summing amplifier gain. To perform the various procedures, some extra equipment are required:

- DC power source. This could be a power supply or simply a battery
- Signal generator capable of producing a sine signal with $1 V_{pp}$ amplitude
- Oscilloscope suited for measuring audio signals

Before any of the adjustments are done, make sure all the potentiometers are at center position.

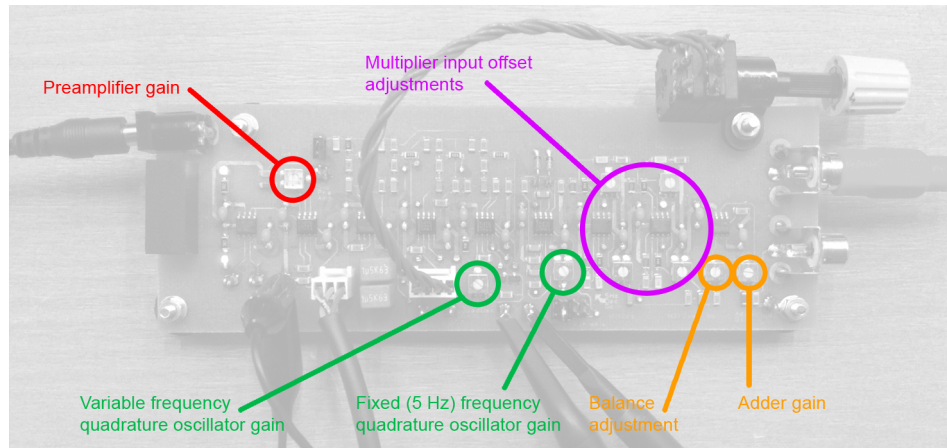


Figure 5.12: TFS PT4.1, with all adjustment potentiometers marked and labeled. The color scheme matches that of figure 3.2 on page 12 for easy navigation and adjustments.

5.3.1 Offset adjustments

To adjust the offsets of the multiplier inputs, the jumper *ENABLE* and the double-jumper *OSCSEL* has to be removed first. This is because pin 2 on *ENABLE* and the test pins *EXTOSC1* and *EXTOSC2* will be used to introduce voltages and signals, while adjusting the potentiometers. The DC power source should be set to anywhere between 5 and 10 V, the exact value is not important. The calibrating done before testing used 5 V DC, and the signal generator was set to 1 kHz sine wave of amplitude $1 V_{pp}$. To calibrate offset for X2 of *MULTI1* and *MULTI2* do the following connections:

- Connect the signal generator to pin 2 on *ENABLE*
- Connect the DC voltage to *EXTOSC2*

Measure on pin 5 of *MULTI1* with the oscilloscope and adjust *M1X2OFFSET* to minimize offset. Then repeat the procedure for *MULTI2* measuring on pin 5 of *MULTI2* and adjusting *M2X2OFFSET* instead.

To calibrate offset for Y2 of *MULTI1* and *MULTI2* a similar setup is used, but now the two connections must be switched:

- Connect the DC voltage to pin 2 on *ENABLE*
- Connect the signal generator to *EXTOSC2*

Again measure on pin 5 of *MULTI1* with the oscilloscope, but now *M1Y2OFFSET* has to be adjusted to minimize offset. Then repeat the procedure for *MULTI2*, measuring on pin 5 of *MULTI2* and adjusting *M2Y2OFFSET*. The offset adjustments are now complete.

5.3.2 Balance

The adjustment of balance is for correcting any differences of the two multipliers, so that each one contributes equally to the final summing. An unbalance would result in an amplitude variations of the output result. The balance adjustment is therefore best done visually, with the help of an oscilloscope, see figure 5.13. Make sure the *ENABLE* jumper is on, and select the 5 Hz fixed frequency quadrature oscillator with the *OSCSEL* double jumper. The balance adjustment are then done by

- Connect the signal generator to test pin *POSSIGIN*
- Look at the *OUTPUT* signal of TFS, and adjust *BALANCE* to get a “steady” sine signal

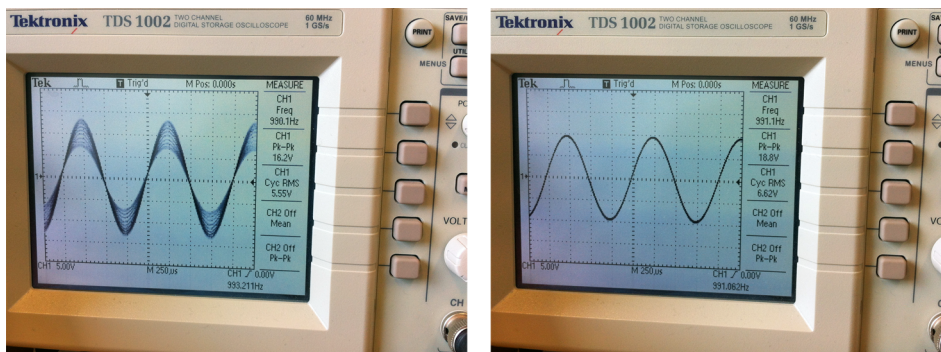


Figure 5.13: To the left an output signal is seen where the two multiplier results are not added equally. In the right picture the balance has been adjusted, and the output is finally “steady”.

5.3.3 Adjusting the quadrature oscillators

TFS PT4.1 contains, as previously mentioned, two quadrature oscillators. One runs at a fixed frequency, about 5 Hz, while the other can be adjusted to produce frequencies between 0.5 and 50 hertz. First the fixed frequency oscillator is adjusted. This is a simple process of three steps:

- Set double jumper *OSCSEL* to the left to select the 5 Hz quadrature oscillator
- Set potentiometer *OSCAMPL* to max (all the way to the right)
- Measure with the oscilloscope on test pin *EXTOSC1* or *EXTOSC2* and turn the *OSCAMPL* down until the signal is about $10 V_{pp}$

The adjustment of the adjustable frequency quadrature oscillator is a bit more complicated, as the process may need to be repeated during testing. This is because frequency and amplitude are tied together, and the oscillator gain need to be adjusted to compensate for change in frequency. The process for adjustment of the adjustable frequency quadrature oscillator follows:

- Set double jumper *OSCSEL* to the right to select the adjustable frequency quadrature oscillator
- Set potentiometer *OSCGAIN* to max (all the way to the right)
- Measure with the oscilloscope on test pin *EXTOSC1* or *EXTOSC2* and adjust the dual potentiometer until the desired frequency is set
- Finally turn the *OSCGAIN* down until the signal is about $10 V_{pp}$

A note on oscillator amplitude

As the amplitude of the oscillator is increased, the total harmonic distortion also increases. This is why one should try to keep the amplitudes as low as possible, if the best possible result in regard of distortion of the final output signal is required. The best result from the multiplying operation is achieved when the audio and oscillator signals have similar amplitudes, but more importantly, the distortion of the sine wave from the oscillator must be as low as possible. In practical use there were found no problems of adjusting the amplitudes low (barely oscillating) to get the best possibly result from the frequency shifter.

5.3.4 Preamplifier and adder gain

The last part of setting up the frequency shifter, is the adjustment of the preamplifier and adder gain. The goal of adjusting the preamplifier gain is to ensure that the audio signal are as strong as possible, and have similar amplitude as the oscillator signal. All this without becoming distorted from the limitations of the preamplifier structure. As different types and brands of microphones deliver different signal amplitudes, there are no blueprint of the perfect setting of the preamplifier gain. The levels of a typical microphone

signal varies a lot, but an assumption of amplitudes between $4 mV_{pk-pk}$ and $40 mV_{pk-pk}$ seems reasonable. To set the preamplifier gain, again connect the signal generator to the test point *POSSIGIN*, and set the amplitude of the sine wave to $20 mV_{pk-pk}$. Measure on test pin *SIGPREAMP* and adjust *PREAMPGAIN* until the signal is approximately $8 V_{pk-pk}$. Amplitudes above this will distort the signal in the quadrature output section.

As for the adder gain adjustment, the goal is to amplify the final signal from the adder enough to reach line level. Definition of line level are expressed in different ways for consumer and professional audio equipment. Where dBV are mostly used in consumer electronics, dBu is used in more professional oriented gear. According to Wikipedia [13] line level are -10 dBV for consumer electronics ($0.894 V_{pp}$) and +4 dBu ($3.474 V_{pp}$) for professional electronics. TFS can easily produce the amplitude requirement for both these standards. A level in between would be to adjust the output signal level to $1 V_{rms}$. In practice this works good for all appliances, but the user is encouraged to set the output level according to the specific needs. As all other AC voltages in this thesis is given in V_{pp} , the peak-to-peak value can be calculated from the rms value:

$$\begin{aligned} V_{pp} &= 2\sqrt{2} \cdot V_{rms} \\ &= 2\sqrt{2} \cdot 1 V_{rms} \\ &\approx 2.83 V_{pp} \end{aligned} \tag{5.1}$$

To adjust the adder gain, simply measure the *OUTPUT* of TFS, and adjust *ADDERGAIN* until $2.83 V_{pp}$ is reached. The setup and calibration of the frequency shifter circuit is now complete.

5.4 Testing the frequency shifter

In this section all necessary information is given to replicate the measurements, as long as one have access to the same testing locale. Store Fysiske Auditorium at the Department of Physics at UiO is a typical large lecture hall, with rows of seats/benches in a slope, with the lecturing area with a desk and blackboards at the bottom. Though it appears box shaped in the illustrations, the shape of the room is irregular.

Three different testing scenarios were created to simulate different situations for the frequency shifter, and different settings of the testing equipment. The setups were quite similar, but with a few key differences in each scenario. Scenario B and C had an identical setup in terms of equipment placement in the room, but the settings of the measurement instrument, Brüel & Kjær Type 2226, were different in the two scenarios. To keep the testing variables to a minimum, the sound source used was not a person talking, but an Apple iPhone 4 playing a pre-recorded sound sample. This was done because it is very hard, if not impossible, to keep an equal volume and pitch for each test.

In the tests where no frequency shift were applied, the *PREAMPED* RCA output was used instead of the *OUTPUT* RCA connector. This was done to make the testing situations as equal as possible by using the same microphone preamplifier. The *ENABLE* jumper was connected at all times. As mentioned in the sections above, the amplitude of the adjustable frequency quadrature oscillator needed minor adjustments at the higher frequency setups. All sound pressure level measurements were done using the following procedure:

- Turn the gain all the way down on the amplifier
- Turn sound source on
- Slowly turn the amplifier gain up until some feedback/coloration is heard
- Register the dBA value as *Start of FB* (start of feedback)
- Watch the sound pressure level measurement instrument closely while turning the gain up slowly
- When total feedback/coloration occurred, register the last stable dBA value seen as *Total feedback*

All test were performed three consecutive times per setting, in each scenario. The microphone was placed precisely at the center of the room, right/left direction when facing the seats.

Two different sound samples were used for testing. The first, called “Human male voice”, was the authors voice saying welcome and goodbye in Norwegian, over and over again in a loop of 60 seconds. This was to simulate a person speaking. The sound samples was recorded in Audacity 2.0.0 using a Shure SM58® microphone through a Sound Blaster USB interface from

Creative. The only sound processing done was normalizing in said software. The second sound sample, “White noise”, was a 60 second clip of white noise generated in Audacity. An attempt to describe the sound sample “Human male voice” is done in a waveform and spectrogram chart presented in figure 5.17 on page 46. There the time-amplitude waveform of the sound is presented, as well as two different sonograms showing different power and frequency properties versus the time of the sound.

5.4.1 Scenario A

Figure 5.14 shows the equipment setup for all measurements in scenario A. The sound pressure level measurement instrument were set to “Fast” for all measurements in Scenario A. Additional information about the setup can be found in table 5.2.

Table 5.2: Distances between equipment used in testing scenario A.

<i>Object and placement</i>	<i>Distance</i>
Microphone to sound source	0.5 cm
Microphone to floor	110 cm
Microphone to speakers closest front corner	110 cm

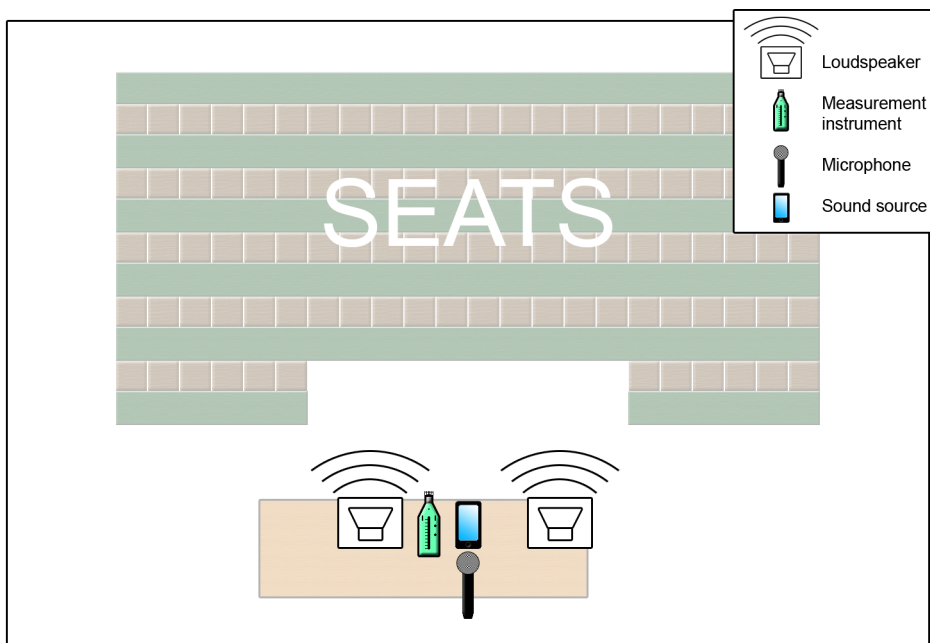


Figure 5.14: Store Fysiske Auditorium is shown with the equipment setup and placement used in scenario A.

5.4.2 Scenario B

This setup was a bit different than scenario A, and can be seen in figure 5.15. Here the setup was “worse”, in terms of sound quality for the audience, as the speakers were partly pointing towards the microphone. This would ensure that feedback occurred easier, and was considered a type of worst case, but nevertheless common, scenario for a lecture or conference setup. The sound pressure level measurement instrument was in this scenario set to “Fast”. Additional information about the setup can be found in table 5.3

Table 5.3: Distances between equipment used in testing scenario B and C.

Object and placement	Distance
Microphone to sound source	0.5 cm
Microphone to floor	110 cm
Microphone to speakers, front center	530 cm

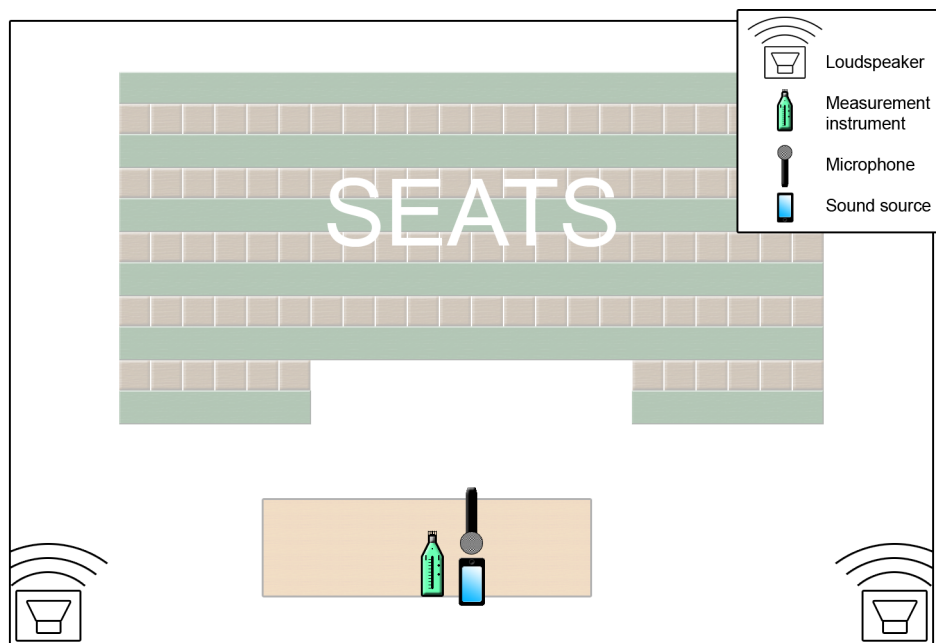


Figure 5.15: Store Fysiske Auditorium is shown with the equipment setup and placement used in scenario B and C.

5.4.3 Scenario C

The measurement setup in scenario C was identical to scenario B, save the setting on the sound pressure level measurement instrument. The Brüel & Kjær Type 2226 instrument was set to “Slow” in scenario C, in opposition to “Fast” in scenario B (and A).

5.5 Measuring of reverberation time

As the frequency shift needed, according to equation 2.5 on page 8, is dependent of the reverberation time of the room, an estimate of RT_{60} for Store Fysiske Auditorium was of certain interest. The method described here is not a way to make an accurate measurement, but it will give a good estimate of the actual value. Because of background noise, RT_{60} is hard to measure. RT_{30} was therefor measured and estimated instead. RT_{30} can then be extrapolated to find RT_{60}

$$RT_{60} = 2 \cdot RT_{30} \quad (5.2)$$

The actual procedure used to estimate RT_{30} consists of both sound capture and a graphical estimation. The procedure is based of the procedure presented by Dr. Dan Boye at Davidson College [1], but revised to be able to give an estimate from a graphical point of view. The sound capture in Store Fysiske Auditorium was made with the microphone connected to the Fast Track Ultra audio interface. The microphone was placed in the middle of the room. Audacity was used for the sound processing on a Macbook. The reverberation time estimate followed the procedure:

- Capture the sound of clapping five times in Audacity
- Extract noise profile
- Remove noise from the sample
- Select strongest and most sharp clap was visually in Audacity, and discard the rest of the sample
- Change the view from amplitude to dB plot
- Estimate RT_{30} from the figure

The last point in the list was done in Adobe Photoshop to ease the process and get a better estimate.



Figure 5.16: Equipment setup for testing the TFS. Seen in the picture is the sound pressure level measurement instrument, sound source (iPhone 4), microphone, oscilloscope and of course the frequency shifter PT4.1.

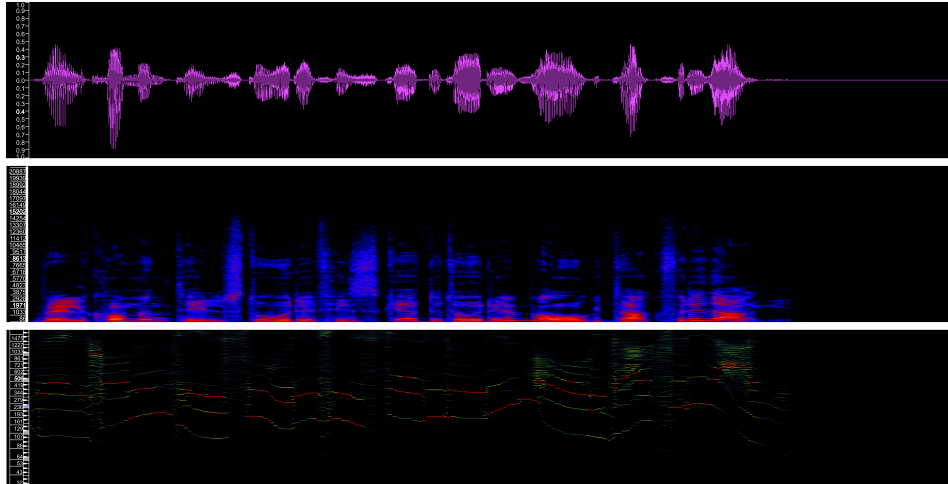


Figure 5.17: Three different views of a small segment of the sound sample “Human male voice”, used in the testing scenarios. The horizontal axis shows time, and the segment shown is about 5 seconds. This segment is repeated in the sound sample 12 times for a total of 60 seconds sound sample. The vertical axis for the top graph, the waveform, is amplitude. In the spectrogram in the middle, as well as the bottom peak-frequency spectrogram, the vertical axis represent frequency. The two lower graphs are in addition color coded. In the center sonogram, warmer colors equals more power. At the bottom, warmer colors equals higher amplitudes of the shown frequency. As we can see from the top waveform, the amplitude is varying throughout the sound clip. This made measurements hard to do, as the feedback often occurred at the exact same position in the sound sample, every time. From the middle graph, the spectrogram, we see as expected that the lower frequencies are over represented, as is normal in speech. The higher frequencies are only present in short bursts. In the bottom graph we see that a select few frequencies holds the most of the power in the clip, except at the end of the sample. There three peaks can be seen, and it was those three peaks that made the feedback occur when the measurements were made.

Settings in Sonic Visualizer for the two sonograms: Middle sample window size: 1024, 50 % overlap, all bins. Bottom sample window size: 4096, 93.75 % overlap, only frequency bins.

Chapter 6

Results

The results in this chapter are the final results of testing the TFS prototype 4.1 in the three different scenarios: A, B and C. Each scenarios results are presented in separate sections, and each section contains the measured sound pressure levels, as well as calculations of mean values of those measurements. The most interesting results are in column 3 and 5 of tables 6.2 on the next page, 6.4 on page 50, 6.6 on page 52, 6.8 on page 54, 6.10 on page 56 and 6.12 on page 58, which shows total dB(A) gained when using the frequency shifter. A graphical representation of these results follows after the tables in each section. The setup and testing procedure for each scenario are described in section 5.4 on page 42.

The last section of this chapter describes the method used for measuring RT_{30} . As all necessary equipment for accurate measurements of the reverberation time were not available, the method used gave a rather rough estimate, not an accurate measurement.

Explanation to the tables

The frequency shift “0 Hz” means the main parts of TFS was bypassed. In these measurements only the preamplifier section of the TFS was used, and no frequency shift was made. The frequency shift denoted as “5 (fixed)” was measured using the fixed 5 Hz quadrature oscillator. All the other tests were using the adjustable frequency quadrature oscillator.

6.1 Scenario A

Table 6.1: Scenario A: “Human male voice”. Measured results. Measuring instrument set to “Fast”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
1	50 - 90	off	0	87,0	87,0
2	50 - 90	off	0	85,0	86,0
3	50 - 90	off	0	85,0	86,0
4	80 - 120	on	3	86,0	91,0
5	80 - 120	on	3	86,0	93,0
6	80 - 120	on	3	87,0	92,5
7	80 - 120	on	5	86,0	91,0
8	80 - 120	on	5	86,5	97,5
9	80 - 120	on	5	87,5	100,5
10	80 - 120	on	5 (fixed)	86,0	88,0
11	80 - 120	on	5 (fixed)	86,0	90,0
12	80 - 120	on	5 (fixed)	85,5	89,5
13	80 - 120	on	10	87,0	92,0
14	80 - 120	on	10	86,0	91,0
15	80 - 120	on	10	86,0	92,0
16	80 - 120	on	20	88,0	91,5
17	80 - 120	on	20	87,0	94,5
18	80 - 120	on	20	87,0	95,0

Table 6.2: Scenario A: “Human male voice”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
1-3	85,7	-	86,3	-
4-6	86,3	0,7	92,2	5,8
7-9	86,7	1,0	96,3	10,0
10-12	85,8	0,2	89,2	2,8
13-15	86,3	0,7	91,7	5,3
16-18	87,3	1,7	93,7	7,3

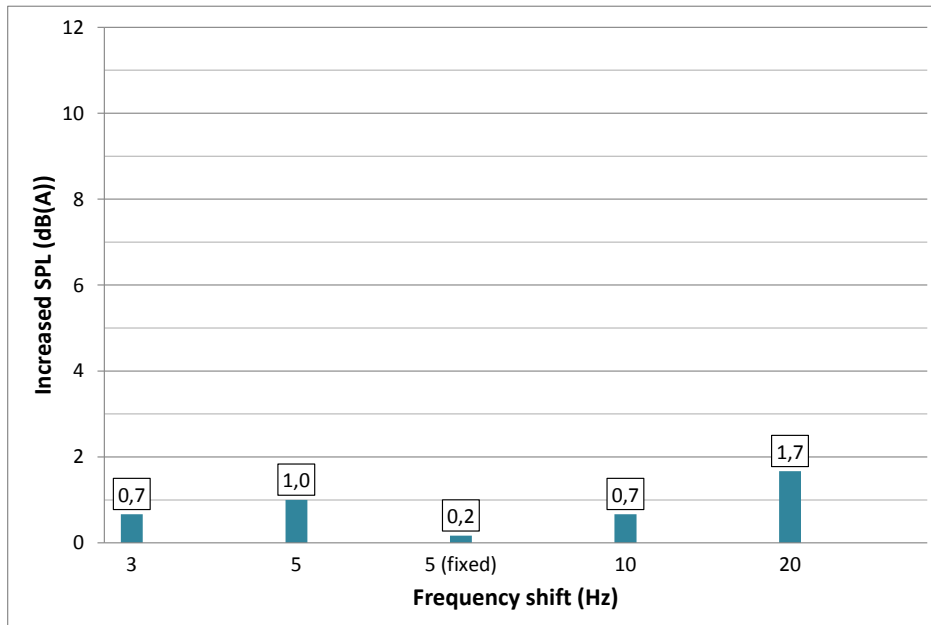


Figure 6.1: Scenario A: “Human male voice”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.2 on the preceding page.

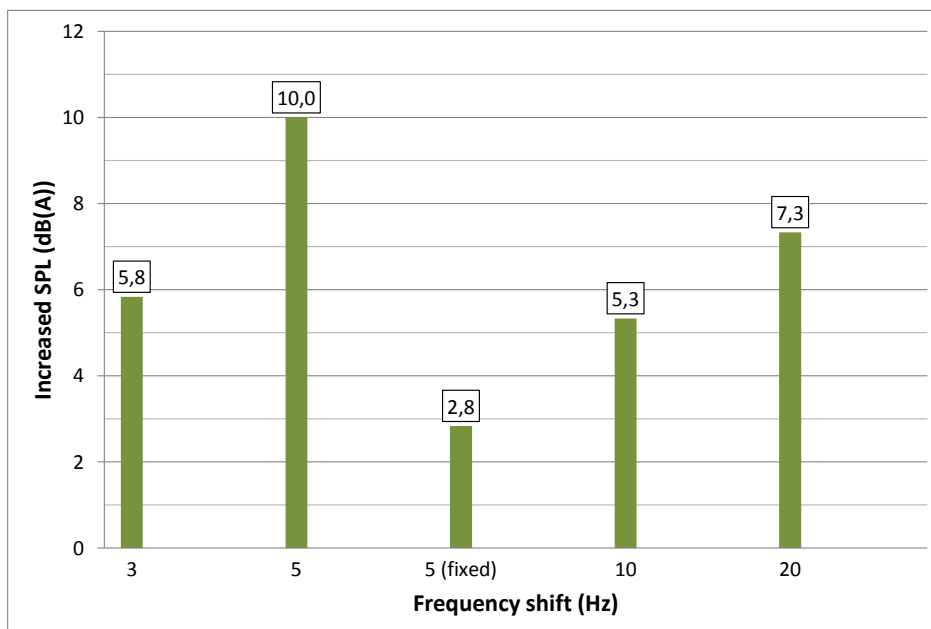


Figure 6.2: Scenario A: “Human male voice”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.2 on the preceding page.

Table 6.3: Scenario A: “White noise”. Measured results. Measuring instrument set to “Fast”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
19	50 - 90	off	0	82,0	82,5
20	50 - 90	off	0	81,5	82,0
21	50 - 90	off	0	81,0	82,0
22	80 - 120	on	3	87,0	87,0
23	80 - 120	on	3	86,5	90,5
24	80 - 120	on	3	86,0	90,0
25	80 - 120	on	5	86,5	91,5
26	80 - 120	on	5	87,0	91,0
27	80 - 120	on	5	87,0	91,0
28	80 - 120	on	5 (fixed)	83,0	88,0
29	80 - 120	on	5 (fixed)	85,0	89,0
30	80 - 120	on	5 (fixed)	85,0	90,0
31	80 - 120	on	10	85,5	90,5
32	80 - 120	on	10	87,0	91,0
33	80 - 120	on	10	87,0	91,0
34	80 - 120	on	20	86,5	94,0
35	80 - 120	on	20	86,0	93,0
36	80 - 120	on	20	86,0	93,5

Table 6.4: Scenario A: “White noise”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
19-21	81,5	-	82,2	-
22-24	86,5	5,0	89,2	7,0
25-27	86,8	5,3	91,2	9,0
28-30	84,3	2,8	89,0	6,8
31-33	86,5	5,0	90,8	8,7
34-36	86,2	4,7	93,5	11,3

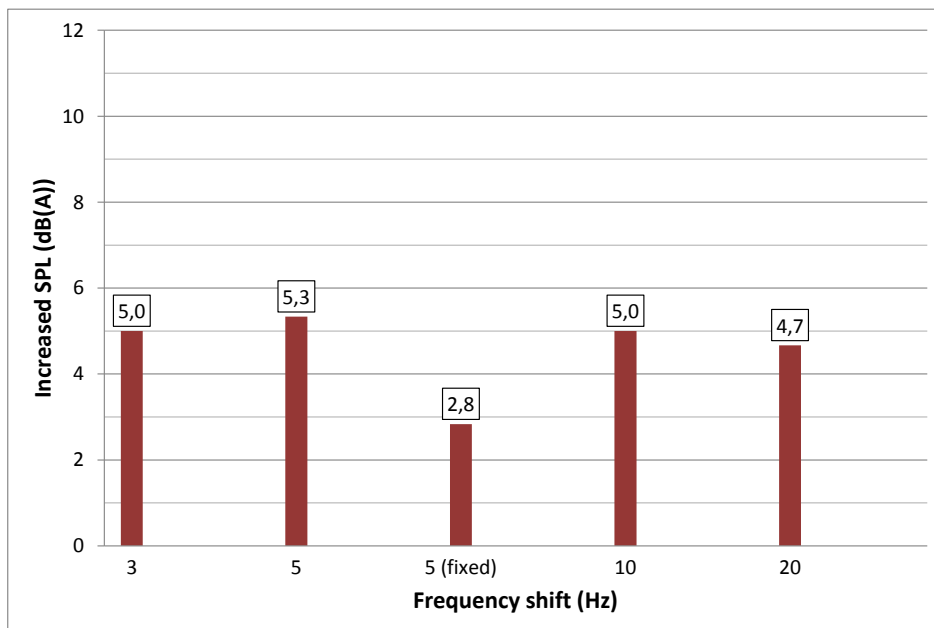


Figure 6.3: Scenario A: “White noise”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.4 on the preceding page.

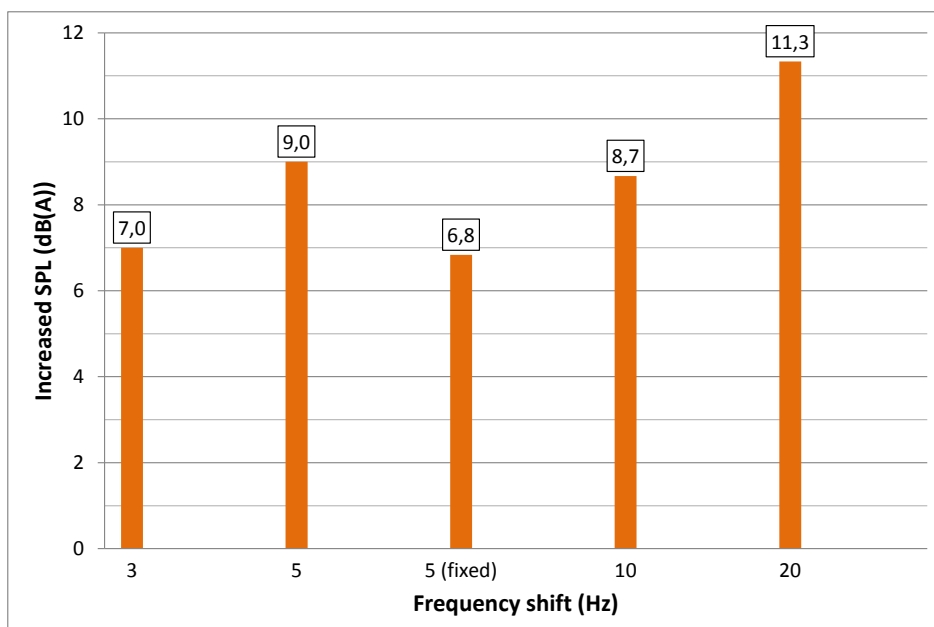


Figure 6.4: Scenario A: “White noise”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.4 on the preceding page.

6.2 Scenario B

Table 6.5: Scenario B: “Human male voice”. Measured results. Measuring instrument set to “Fast”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
1	80 - 120	off	0	85,0	91,0
2	80 - 120	off	0	85,0	90,5
3	80 - 120	off	0	86,0	90,5
4	80 - 120	on	3	87,0	93,0
5	80 - 120	on	3	87,0	92,0
6	80 - 120	on	3	87,0	92,0
7	80 - 120	on	5	89,0	93,5
8	80 - 120	on	5	90,0	93,5
9	80 - 120	on	5	90,5	93,5
10	80 - 120	on	5 (fixed)	86,5	92,0
11	80 - 120	on	5 (fixed)	86,5	92,5
12	80 - 120	on	5 (fixed)	87,0	92,0
13	80 - 120	on	10	85,0	93,0
14	80 - 120	on	10	86,0	92,0
15	80 - 120	on	10	86,0	91,5
16	80 - 120	on	20	86,5	93,0
17	80 - 120	on	20	87,0	93,0
18	80 - 120	on	20	86,0	94,0

Table 6.6: Scenario B: “Human male voice”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
1-3	85,3	-	90,7	-
4-6	87,0	1,7	92,3	1,7
7-9	89,8	4,5	93,5	2,8
10-12	86,7	1,3	92,2	1,5
13-15	85,7	0,3	92,2	1,5
16-18	86,5	1,2	93,3	2,7

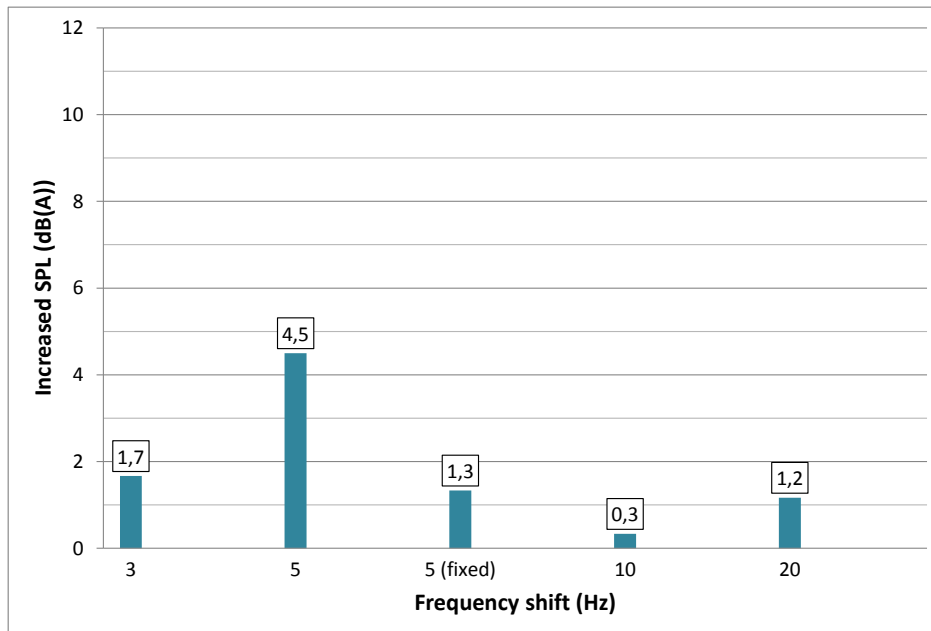


Figure 6.5: Scenario B: “Human male voice”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.6 on the preceding page.

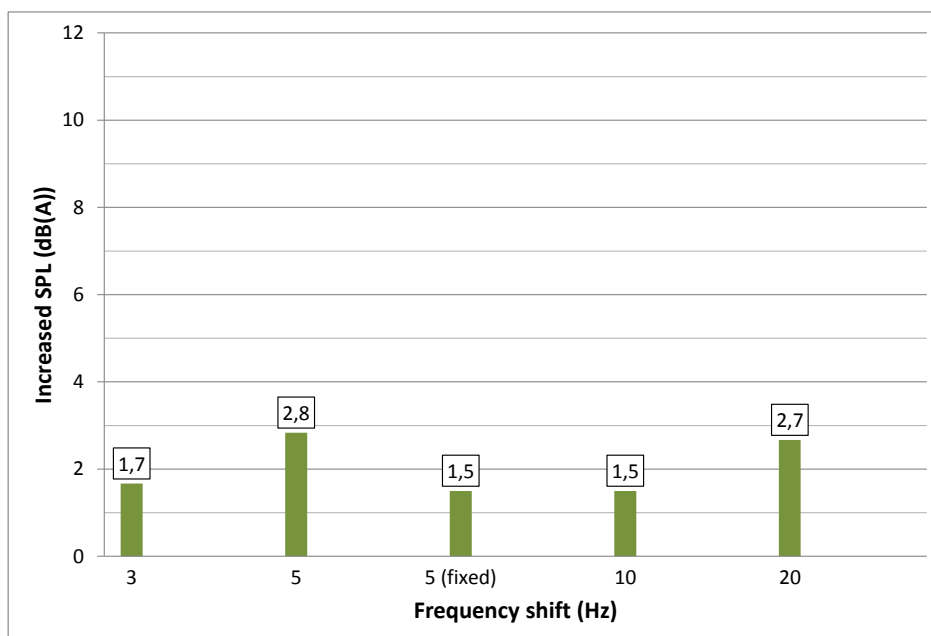


Figure 6.6: Scenario B: “Human male voice”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.6 on the preceding page.

Table 6.7: Scenario B: “White noise”. Measured results. Measuring instrument set to “Fast”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
19	80 - 120	off	0	83,5	84,0
20	80 - 120	off	0	83,5	85,0
21	80 - 120	off	0	84,5	85,0
22	80 - 120	on	3	87,0	91,0
23	80 - 120	on	3	87,0	89,5
24	80 - 120	on	3	85,5	90,0
25	80 - 120	on	5	86,0	92,5
26	80 - 120	on	5	85,0	90,5
27	80 - 120	on	5	86,0	90,0
28	80 - 120	on	5 (fixed)	86,5	90,0
29	80 - 120	on	5 (fixed)	86,0	89,5
30	80 - 120	on	5 (fixed)	86,0	91,5
31	80 - 120	on	10	87,0	92,0
32	80 - 120	on	10	87,0	92,0
33	80 - 120	on	10	86,5	92,0
34	80 - 120	on	20	87,5	93,0
35	80 - 120	on	20	88,0	92,0
36	80 - 120	on	20	87,0	92,0

Table 6.8: Scenario B: “White noise”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
19-21	83,8	-	84,7	-
22-24	86,5	2,7	90,2	5,5
25-27	85,7	1,8	91,0	6,3
28-30	86,2	2,3	90,3	5,7
31-33	86,8	3,0	92,0	7,3
34-36	87,5	3,7	92,3	7,7

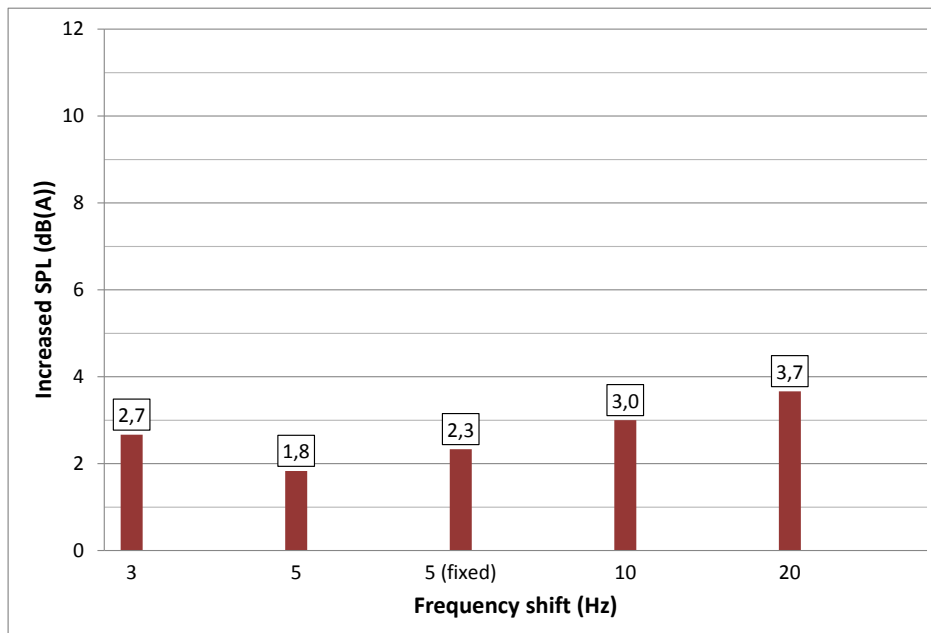


Figure 6.7: Scenario B: “White noise”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.8 on the preceding page.

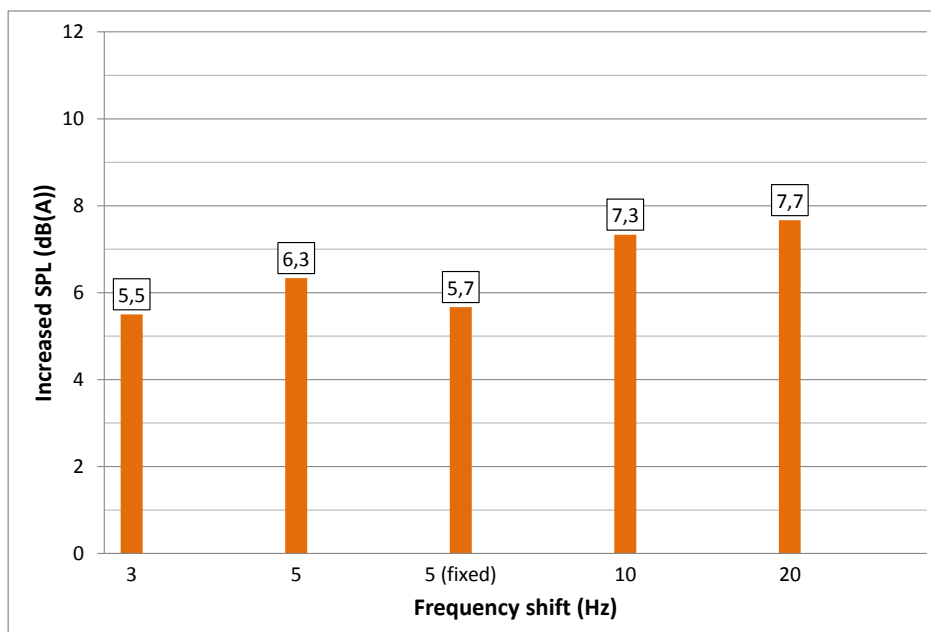


Figure 6.8: Scenario B: “White noise”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.8 on the preceding page.

6.3 Scenario C

Table 6.9: Scenario C: “Human male voice”. Measured results. Measuring instrument set to “Slow”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
1	80 - 120	off	0	83,5	84,5
2	80 - 120	off	0	83,5	85,0
3	80 - 120	off	0	84,0	85,5
4	80 - 120	on	3	86,5	92,5
5	80 - 120	on	3	87,0	92,0
6	80 - 120	on	3	87,0	92,5
7	80 - 120	on	5	88,0	92,5
8	80 - 120	on	5	89,0	92,0
9	80 - 120	on	5	88,5	91,0
10	80 - 120	on	5 (fixed)	85,5	91,5
11	80 - 120	on	5 (fixed)	87,5	94,0
12	80 - 120	on	5 (fixed)	87,0	93,5
13	80 - 120	on	10	86,5	90,0
14	80 - 120	on	10	88,0	93,0
15	80 - 120	on	10	86,5	92,0
16	80 - 120	on	20	86,5	94,0
17	80 - 120	on	20	88,0	94,5
18	80 - 120	on	20	88,0	94,5

Table 6.10: Scenario C: “Human male voice”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
1-3	83,7	-	85,0	-
4-6	86,8	3,2	92,3	7,3
7-9	88,5	4,8	91,8	6,8
10-12	86,7	3,0	93,0	8,0
13-15	87,0	3,3	91,7	6,7
16-18	87,5	3,8	94,3	9,3

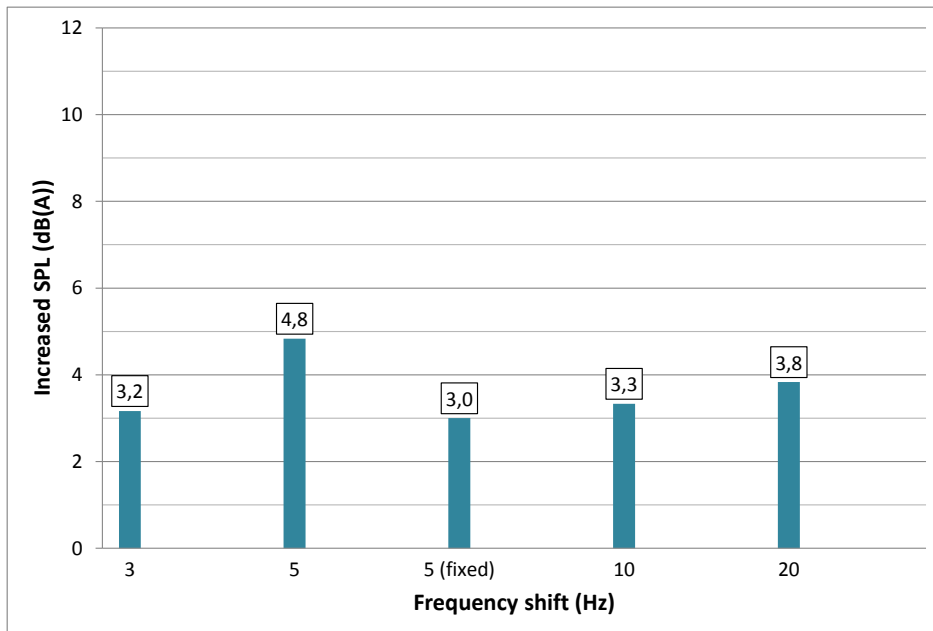


Figure 6.9: Scenario C: “Human male voice”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.10 on the preceding page.

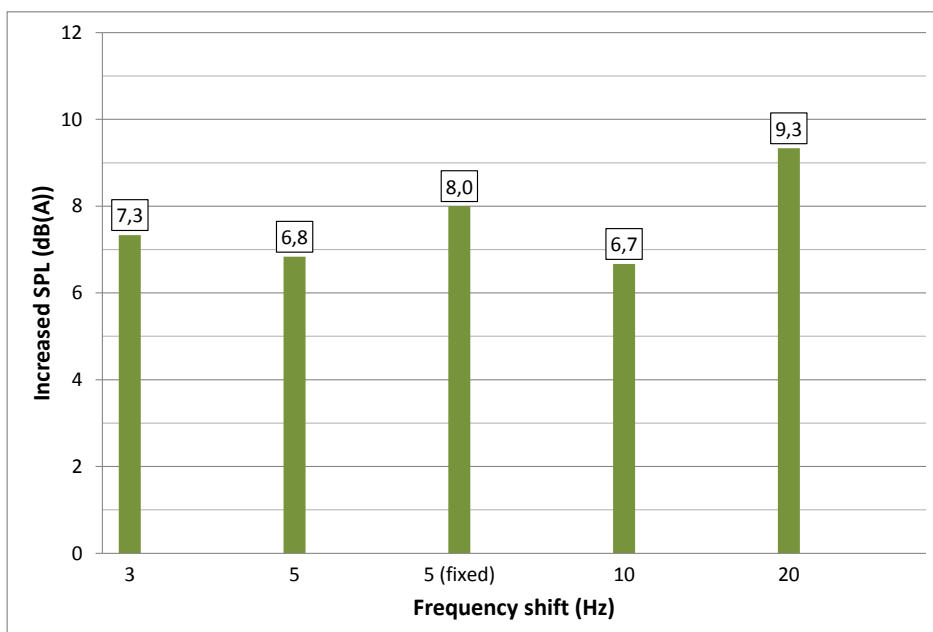


Figure 6.10: Scenario C: “Human male voice”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.10 on the preceding page.

Table 6.11: Scenario C: “White noise”. Measured results. Measuring instrument set to “Slow”.

Test #	Instr. setting (dBA)	TFS (on/off)	Freq. shift (Hz)	Start of FB (dBA)	Total FB (dBA)
19	80 - 120	off	0	83,5	86,0
20	80 - 120	off	0	84,0	85,5
21	80 - 120	off	0	83,5	86,0
22	80 - 120	on	3	89,5	95,0
23	80 - 120	on	3	89,0	95,0
24	80 - 120	on	3	89,5	93,0
25	80 - 120	on	5	90,0	95,5
26	80 - 120	on	5	90,5	98,0
27	80 - 120	on	5	90,0	97,0
28	80 - 120	on	5 (fixed)	88,5	93,0
29	80 - 120	on	5 (fixed)	88,0	93,5
30	80 - 120	on	5 (fixed)	90,0	93,0
31	80 - 120	on	10	87,5	98,0
32	80 - 120	on	10	88,0	97,5
33	80 - 120	on	10	88,5	97,0
34	80 - 120	on	20	89,0	97,5
35	80 - 120	on	20	88,0	95,5
36	80 - 120	on	20	88,0	97,0

Table 6.12: Scenario C: “White noise”. Calculated dBA means for each frequency shift as well as dBA gained when using TFS.

Mean of tests #	Start of FB (dBA)	dBA gained (dBA)	Total FB (dBA)	dBA gained (dBA)
19-21	83,7	-	85,8	-
22-24	89,3	5,7	94,3	8,5
25-27	90,2	6,5	96,8	11,0
28-30	88,8	5,2	93,2	7,3
31-33	88,0	4,3	97,5	11,7
34-36	88,3	4,7	96,7	10,8

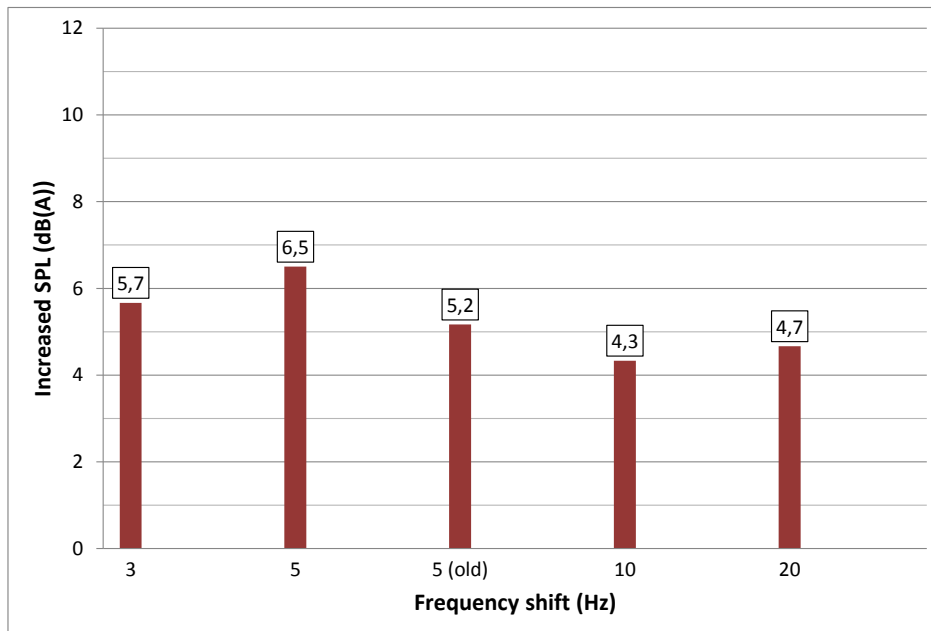


Figure 6.11: Scenario C: “White noise”. The graph shows the dBA gained for each frequency shift, before the feedback started to be heard. The graphs are based on data from the 3rd column in table 6.12 on the preceding page.

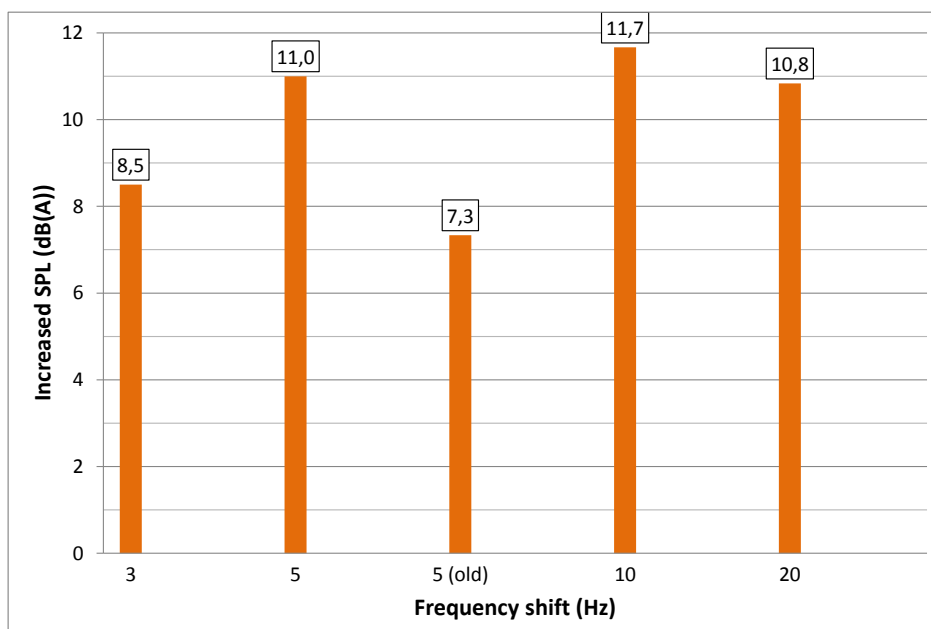


Figure 6.12: Scenario C: “White noise”. The graph shows the dBA gained for each frequency shift, before feedback took over completely. The graphs are based on data from the 5th column in table 6.12 on the preceding page.

6.4 Reverberation time

To estimate RT_{30} , the graph from Audacity was imported into Adobe Photoshop, and manipulated with cursors to be able to get a more accurate reading. As Audacity cannot make a best fit curve of the decaying sound, at least to the authors knowledge, this had to be done visually in Photoshop. The best fit average of the decay can in figure 6.13 on the facing page be seen as a red line. Two reading points, 30 dB apart, has been selected. The first one, -15 dB under the initial value of the clap, is marked with green cursors and occurs after about 0.031 seconds. The second, -45 dB, marked with orange cursors, is at about 0.573 seconds. The reverberation time can then be estimated

$$RT_{30} = 0.573 \text{ s} - 0.031 \text{ s} = 0.542 \text{ s} \quad (6.1)$$

and the result of equation 6.1 can be used with equation 5.2 on page 45 to extrapolate an estimate of RT_{60}

$$RT_{60} = 2 \cdot 0.542 \text{ s} = 1.084 \text{ s} \quad (6.2)$$

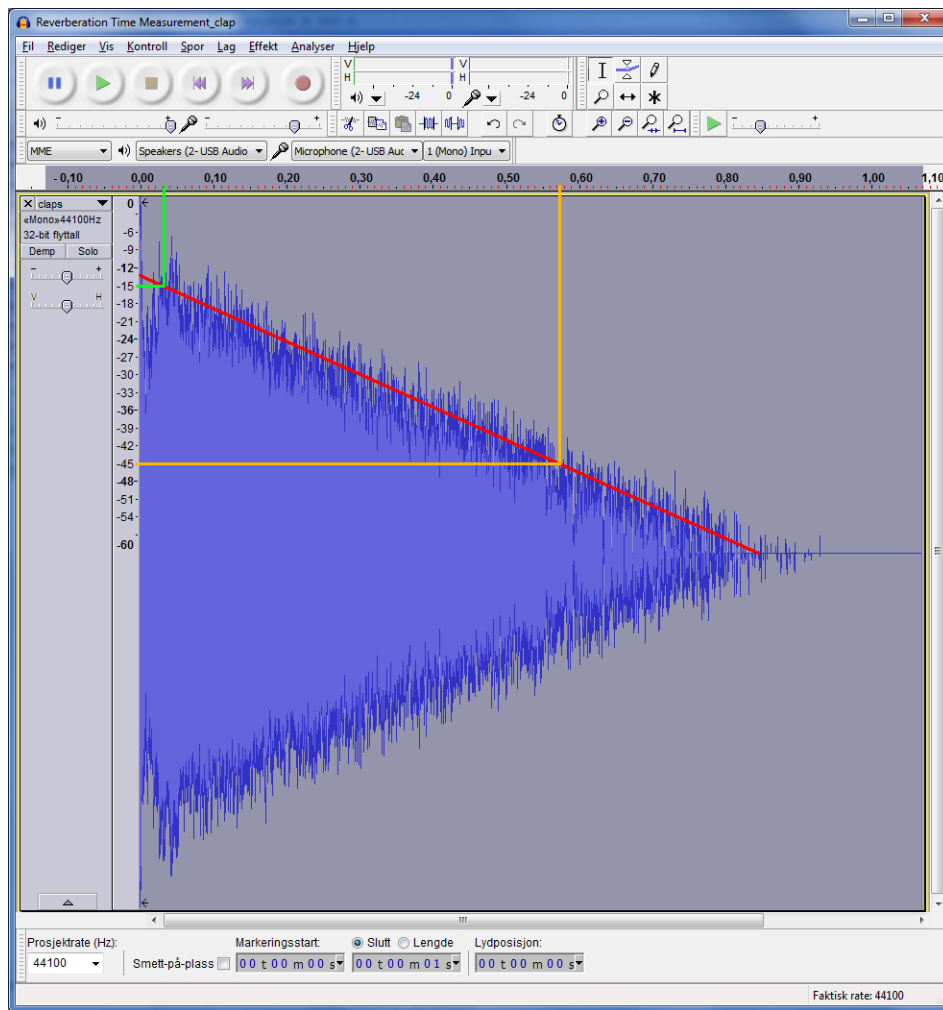


Figure 6.13: The graph seen was imported from Audacity to Photoshop. Here the author has marked the time ruler with red markings to ease the reading process. The average of the decaying sound is marked as a red line. The two measuring points, -15 dB and -45 dB, is seen at about 0.031 seconds and 0.573 seconds. A simple subtraction gives that the sound has decayed 30 dB in the time frame of 0.542 seconds.

Chapter 7

Discussion and conclusion

7.1 Discussion of the measuring results

All measuring results are given in dBA, but the specific values are not of importance. The important numbers are the differences in sound pressure levels between the tests with the frequency shifter and those without. Even though the uncertainty and accuracy of the sound pressure level measurement instrument was unknown, this is of little relevance when all the measurements are seen relative to each other. As seen in the bar diagrams of chapter 6 on page 47, the sound pressure level gained of using the frequency shifter is quite clear. The dBA gained before total feedback occur span from 1.5 to 11.5 dBA, with a mean value for all tests of ~ 6.8 dBA. As for using the frequency shifter in a real world setup, frequency shifts of 10 and 20 Hz are not recommended. The sound artifacts with such high shifts, especially at the 20 Hz test, are distorting the sound quality, and the human voice sounds metalized and unnatural. The mean value of the sound pressure level gained for shifts 3 and 5 Hz (both old and new 5 Hz oscillator) is ~ 6.3 dBA. With a gain of 3 dB, the sound pressure level appear doubled to the human spectator. 6.3 dBA is therefore recognized as a good result.

A comparison of the results of the 5 Hz fixed versus the 5 Hz adjustable frequency quadrature oscillator, shows that shifts with the latter presents larger SPL gains. Why this is, is not clear. The reason why both oscillators were realized, was to see if the quasi quadrature oscillator at least performed as good as the original designed 5 Hz oscillator. This is confirmed by this results.

The measuring setup in the three test scenarios were not found optimal. The fist sound sample, "Human male voice", should have been manipulated more, or recorded completely, to have a better equal/smooth response through the whole clip. It now contains a few peaks, and that made it difficult to make good measurements. This was discovered during testing, too late in the process to be corrected. The white noise sound sample, "White noise", was included as a "catch all", as the possible feedback frequencies of the room were not known. In addition the author had to watch the sound pressure level meter while listening for feedback. This was not optimal, and knowing at what values the feedback "normally occur" may have influenced

the measurement results.

The settings of the Brüel & Kjær Type 2226 sound pressure level measurement instrument was set to “Fast” in scenario A and B. This was changed to “Slow” in scenario C. This meant that the instrument were integrating over a longer time frame, and the measurements became more accurate because the instrument did not “pick up on” every little peak, not even notable by the human ear. Therefore the results for shift frequency 3, 5(fixed) and 5 Hz from scenario C, are, in the authors opinion, the most accurate and reliable measurements. The average sound pressure level gained in those tests is 8.15 dBA. The results found in this thesis complies with the results predicted and measured by Hartley Jones [4].

The measured and extrapolated reverberation times, RT_{30} and RT_{60} , are not accurate measurements, and must be taken as just an indication of the reverberation time of the lecture hall used for the measurements in this thesis. A lot of uncertainty are introduced by a numerous of different sources:

- Lack of good measuring instruments, makes capturing the sound inaccurate. Especially another microphone, preferably omni directional, should have been used.
- The sound source could have been better defined than a clap. Popping a balloon might be better.
- The measurements should not be done visually, but measured directly and calculated from the sampled data.

That being said, $RT_{60} \approx 1\text{second}$, is probably a good estimate of the reverberation time of Store Fysiske Auditorium.

7.2 Discussion of the development process

Most areas of a development process for electronic products, has been part of the work of this thesis. With this comes many different challenges, but also a wider knowledge of the development process of electronic circuits and products. The thesis had in addition to its three main goals, two minor goals as well. Both these were met. During the development process the author has become aware of the importance of “design for testability”. This includes everything from layout of the circuit, so that repairs and fixes easily can be made, having many more test points than just where the final signals are measured, and designing with physical placement of the components in mind.

During the solving of the offset problems with the *AD835* multiplier inputs, we saw that the application notes of a components data sheet can be lacking. Several people were involved in searching for the offsets origin, but sadly it remained unsolved. In this specific case some information about offset adjustments was “kept secret”, or simply not included in the data sheet. One can just speculate for the reason for this, but cut-and-paste based writing of the data sheets seems plausible.

Through the development process, DC blocking capacitors were inserted and then later removed, several times. How a DC block can also introduce an offset, if the current has “nowhere to go”, was valuable “hands on” experience for the author. The importance of seeing a circuit in a wider perspective became clear as the process took on.

Note that the final circuit produced in this thesis, TFS PT4.1, was designed for testing of the concept, and is far more complicated than a commercial product would have to be. Most, if not all, potentiometers in this circuit exists for testing and flexibility purposes, and in a commercial realization those could be replaced by resistors with fixed values. The extra quadrature oscillator is also not needed, and a ready to use product should have just one, running at one shifting frequency. The component/package size used in the circuits in this thesis is 1206. Most components, maybe save the largest capacitors, are available in smaller packages. Would one wish to make a commercial product, all these possible simplifications will result in a much smaller and less expensive circuit.

7.3 Conclusion

A working frequency shifter was developed and tested in three different scenarios. The results shows that a frequency shift of the signal can aid in the problematic area that is acoustic feedback in lecturer halls and the like. The gain in sound pressure level can be up to ~ 8 dBA when the frequency shifter is in use. Frequency shifts above 5 Hz are not recommended as it make the human voice sound unnatural.

The main goals of this thesis has been accomplished:

- A complete and working frequency shifter was designed, built and tested
- The frequency shifter made handles balanced signals directly from a microphone, and uses electronic components widely available today
- The feedback suppression capabilities of the circuit have been tested with frequency shifts of 3, 5, 10 and 20 Hz

7.4 Digital version of the frequency shifter

A brief research was done in terms of the possibility to make a digital version of the frequency shifter based on a FPGA. A Cyclone III developer kit from Altera was studied, as well as the “Data Conversion Card” add on card. Those two development kits appears to contain all hardware needed to test a digital version of TFS, and several instances of the Intellectual Property (IP) called Numerical Controlled Oscillator (NCO) can be utilized to realize most parts of the circuit. The Simulink® software from Mathworks was also considered a possibility for realizing the circuit in VHDL. Only initial research on these topics was made, and further investigation is required before starting a future realization.

7.5 Future work

For the future aspired reader, it is recommended to construct a more automated testing scenario, if more accurate results are wanted. This would also make the testing procedures easier and less ear hurting, as one could wear ear covers all the time. Building a cut down version with just the parts necessary for the frequency shifter to work, could also be the next step in the investigation of feedback suppression by frequency shifting. It would be exciting to see how small and cheap one could be built, and still work. The possibilities are endless.

As the development came to an end, a quadrature oscillator design by Horng [2] was discovered, but sadly there was no time to test this design. The design allows control of the frequency by a dual potentiometer, but if the design is suited for the low frequencies needed in a frequency shifter is unknown to the author.

As a last remark it would of course also be interesting building a digital version of the frequency shifter, and see how it performs side by side to the analogue version.

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Appendix A

MATLAB code

```
% sample length in seconds
time = 1;
% sampling frequency
smpl_freq = 44.1*10^3;
% time vector
vec_time = 0:1/smpl_freq:time;

% amplitude, frequency and offset of the audio signal
ampl_audio = 1;
audio_freq = 1000;
offset_audsin = 0;
offset_audcos = 0;

% amplitudes, frequency and offsets of the oscillator signal
ampl_qosc_sin = 1;
ampl_qosc_cos = 1;
qosc_freq = 5;
offset_qosc_sin = 0;
offset_qosc_cos = 0;

% balance adjustment in the adder sub-circuit
weight_1 = 1;
weight_2 = 1;

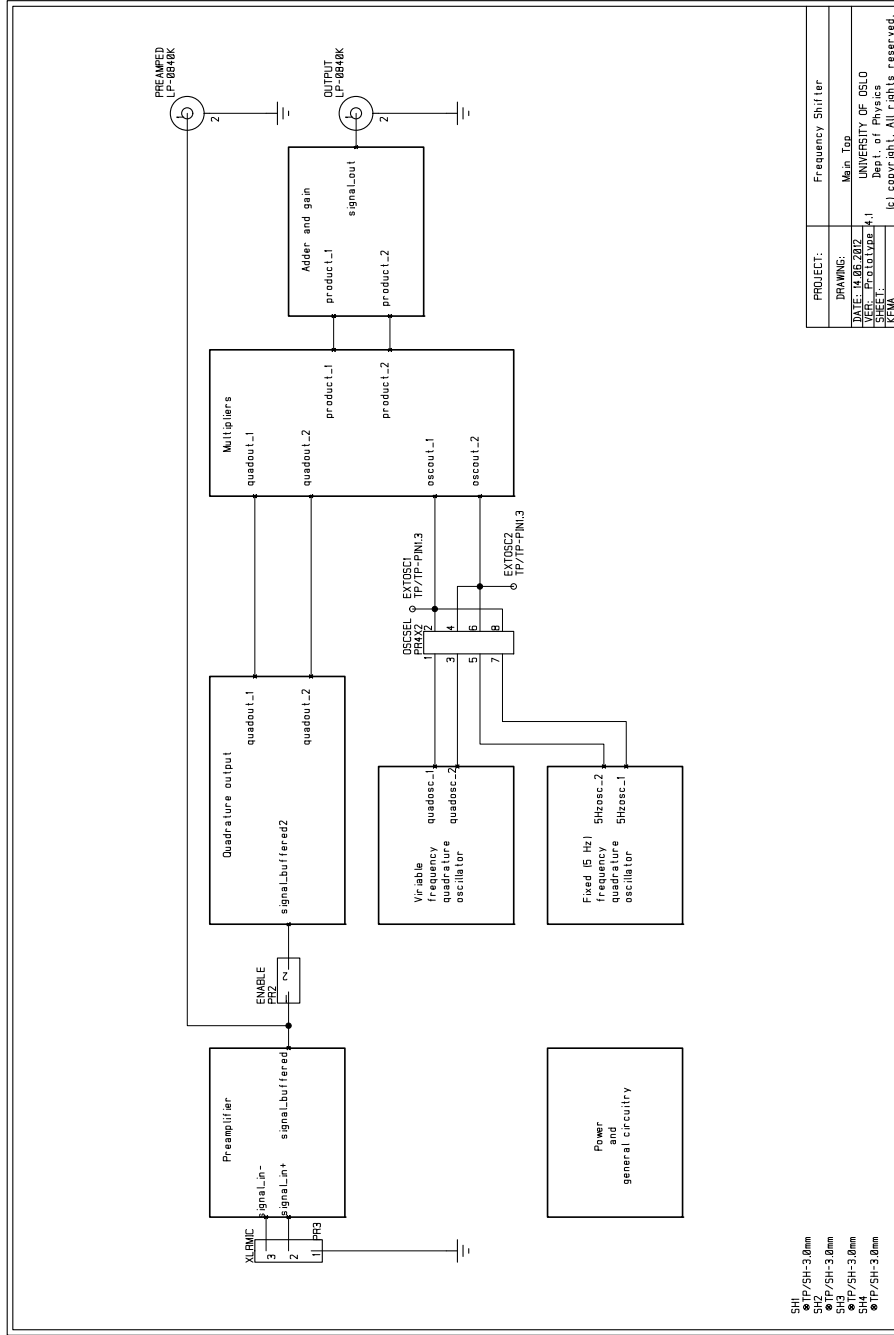
% generate quadrature output of the audio signal
audio_sin = ampl_audio*sin(2*pi*audio_freq*vec_time)+offset_audsin;
audio_cos = ampl_audio*sin(2*pi*audio_freq*vec_time+pi/2)+offset_audcos;
% generate signals from the quadrature oscillator
qosc_sin = ampl_qosc_sin*sin(2*pi*qosc_freq*vec_time)+offset_qosc_sin;
qosc_cos = ampl_qosc_cos*sin(2*pi*qosc_freq*vec_time+pi/2)+offset_qosc_cos;

% calculate the output audio signal
product_1 = audio_sin.*qosc_cos;
product_2 = audio_cos.*qosc_sin;
audio_out = weight_1*product_1 + weight_2*product_2;

% plot the final output audio signal
plot(vec_time(1,1:time*smpl_freq), audio_out(1,1:time*smpl_freq))
legend('OUTPUT')
xlabel('Time (s)')
ylabel('Amplitude (V)')
```

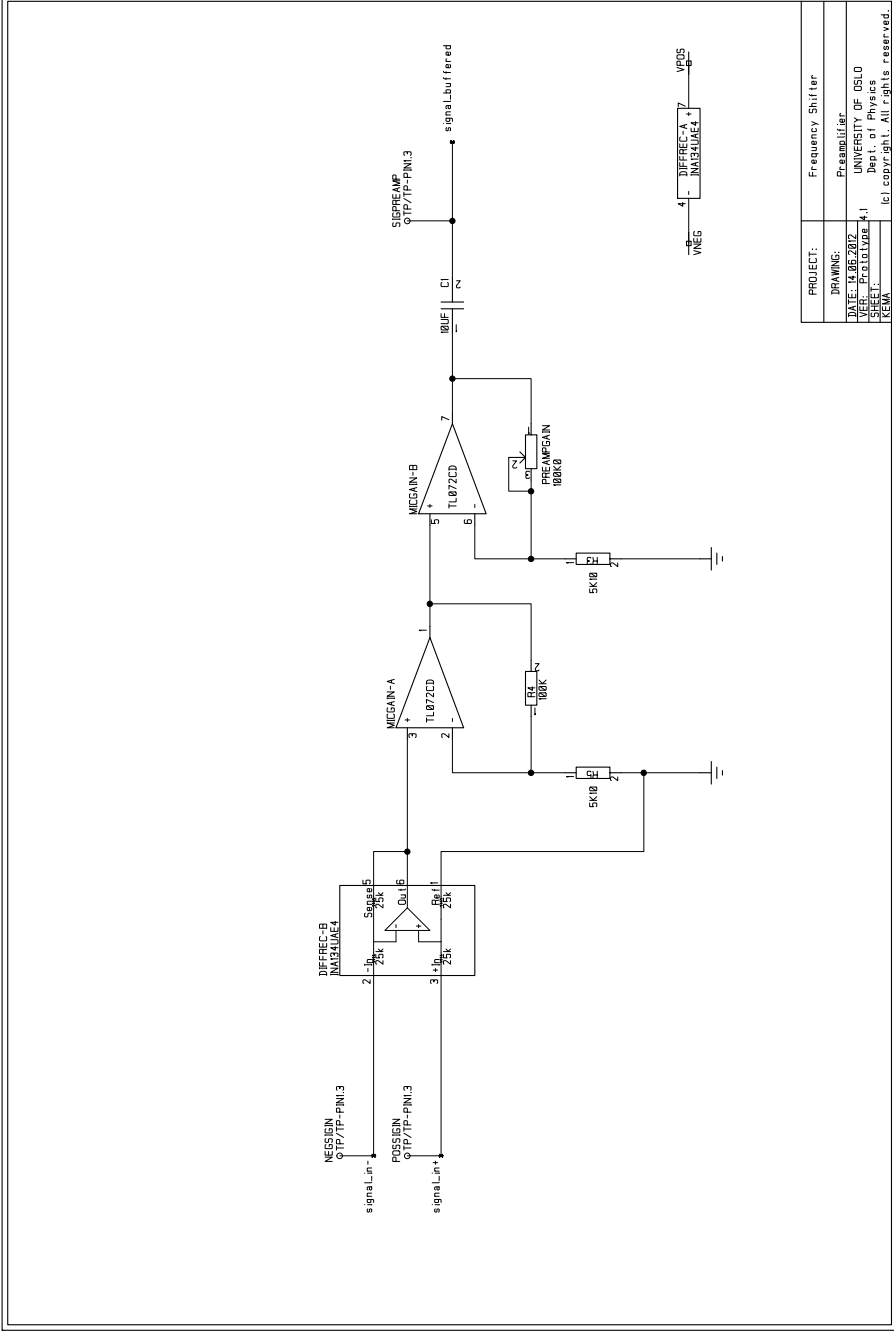

Appendix B

Schematics of TFS PT4.1

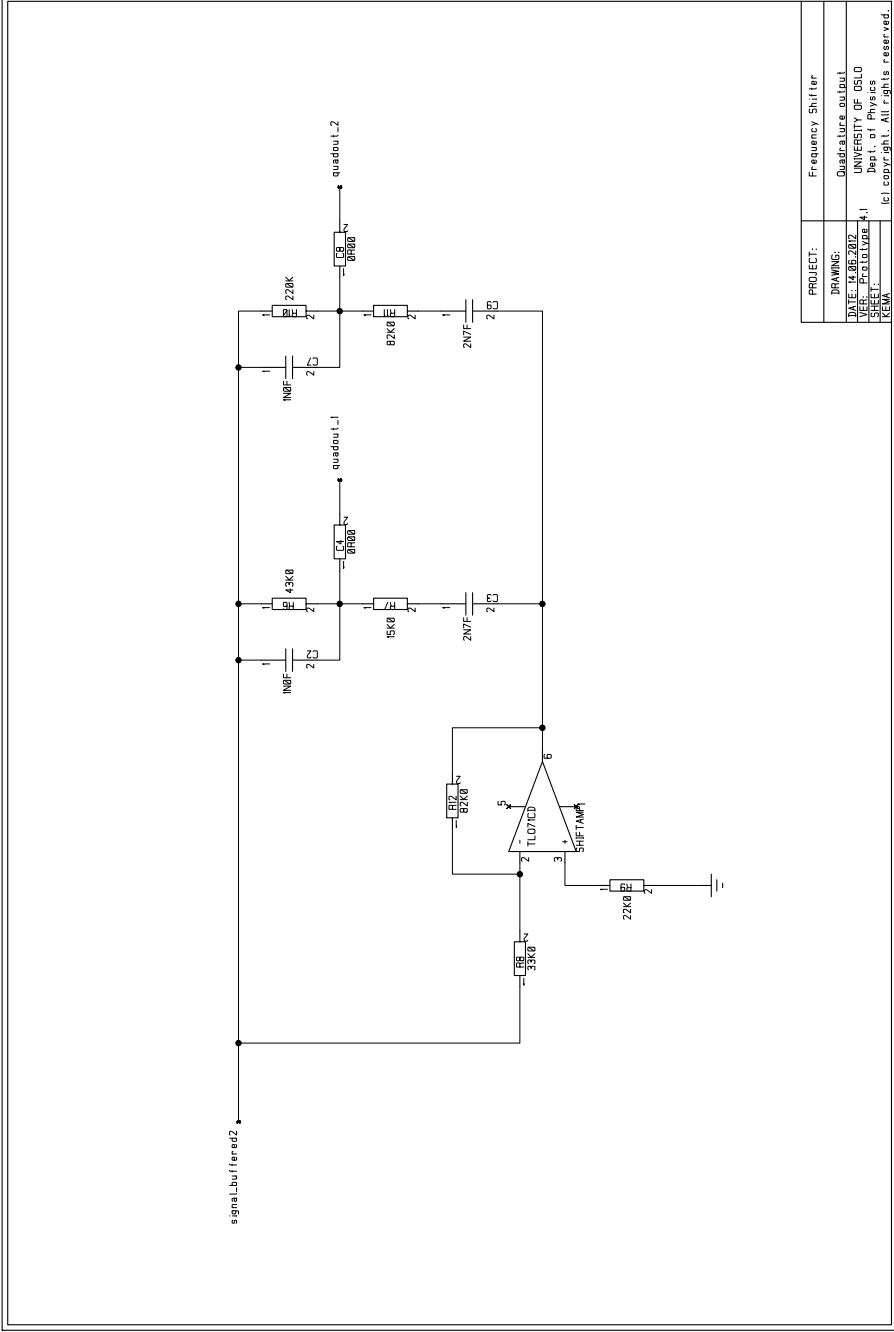


SH
 SHZ/SH-3.0mm
 SHZ
 SHZ/SH-3.0mm
 SHZ
 SHZ/SH-3.0mm
 SHZ
 SHZ/SH-3.0mm

PROJECT:	Frequency Shifter
DRAWING:	Main Top
DATE:	2023/04/11
BY:	PHYSICS
SHEET:	1
REV:	1
FILE:	freq_shifter.kicad_pcb
USER:	ksm

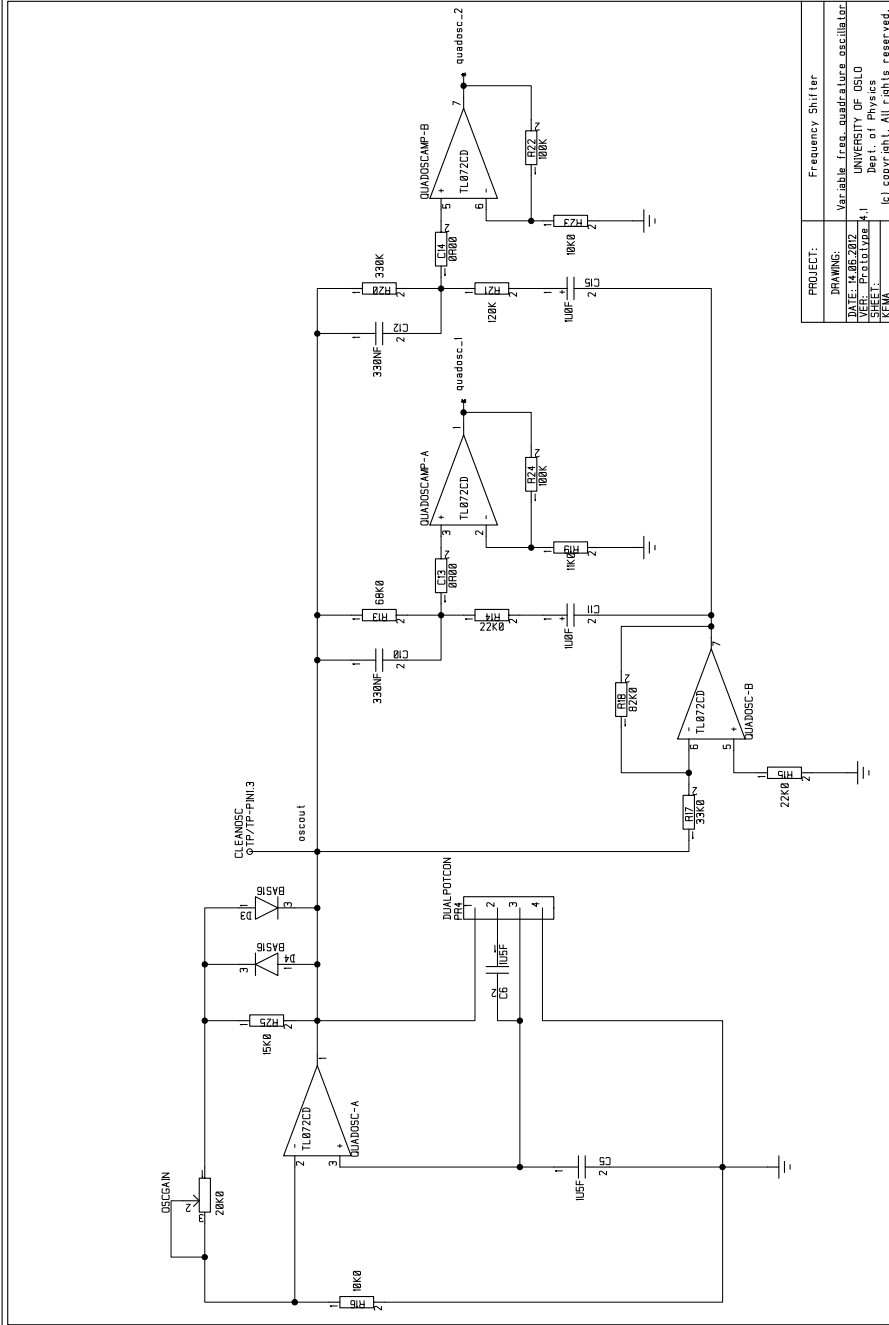


PROJECT:	Frequency Shifter
DRAWING:	PrasannaLilier
DATE:	UNIVERSITY OF OSLO
YEAR:	Dept. of Physics
SHEET:	1
KEY:	tel. copyright. All rights reserved.

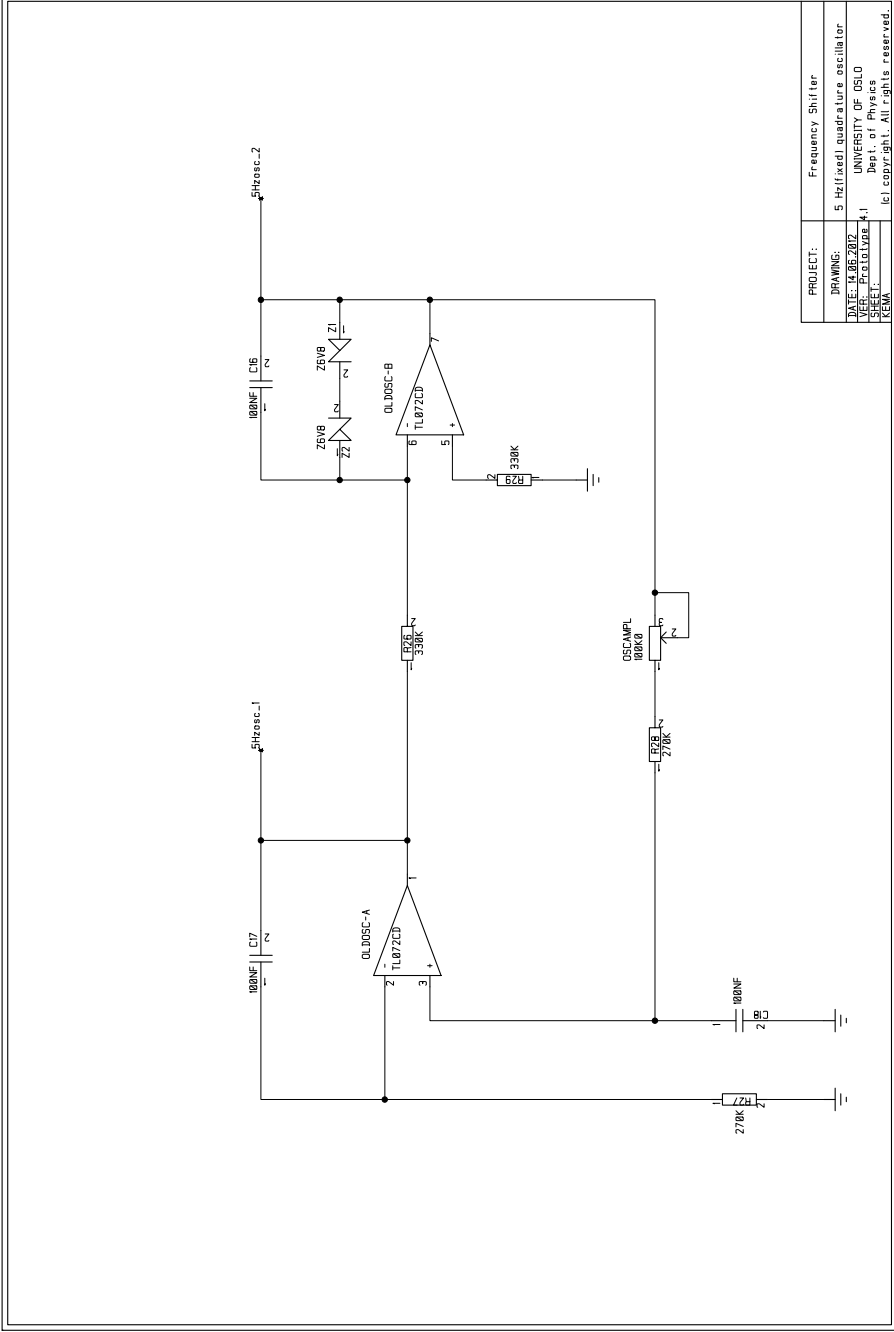


PROJECT:	Frequency Shifter
DRAWING:	Quadrature output
DATE:	2025-07-11
USER:	Physics
SHEET:	1
KEM:	1

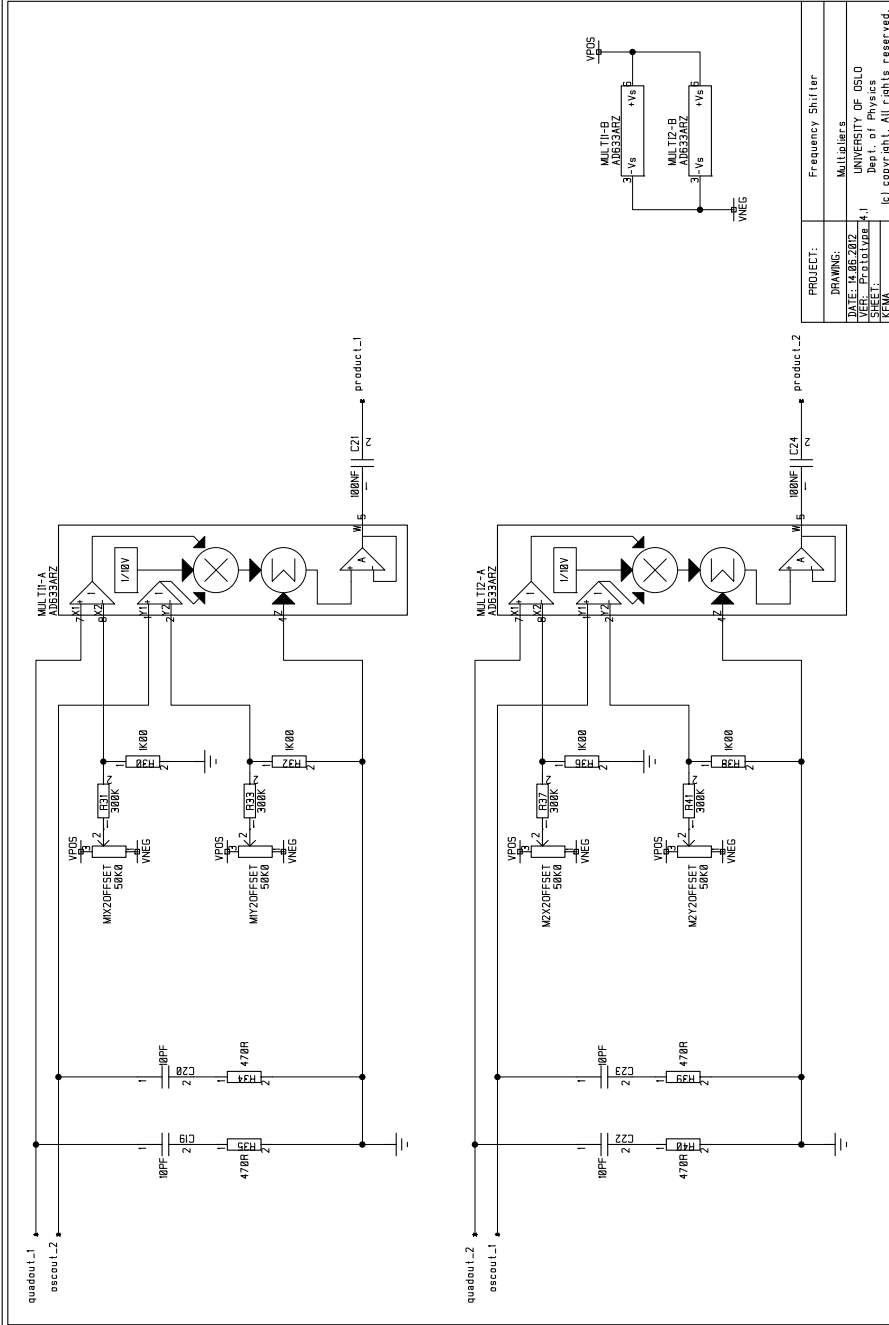
UNIVERSITY OF OSLO
 Dept. of Physics
 All rights reserved.



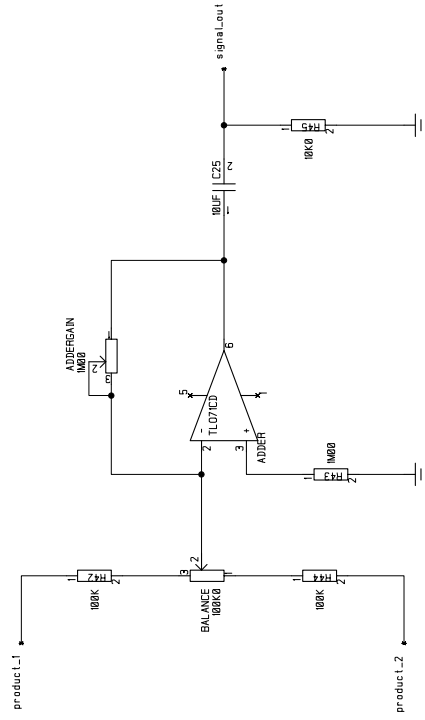
PROJECT:	Frequency Shifter
DRAWING:	Variable-Freq-quadra-lin-a.asc.lib.tor
DATE:	UNIVERSITY OF OSLO
YEAR:	Dept. of Physics
SHEET:	1
KEY:	tel copy right. All rights reserved.



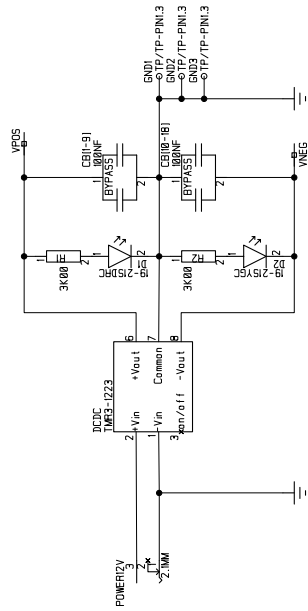
PROJECT:	Frequency Shifter
DRAWING:	5 Hz fixed quadrature oscillator
DATE:	02/20/2011
USER:	UNIVERSITY OF OSLO
DEPT:	Dept. of Physics
SHEET:	1
KEY:	UCL copyright. All rights reserved.



PROJECT:	Frequency Shifter
DRAWING:	MULLIPLIERS
DATE:	05/20/02
USER:	PHYS10004.1
SHEET:	1
KEY:	1cl copyrighl. All rights reserved.



PROJECT:	Frequency Shifter
DRAWING:	AddeR_and_again
DATE:	2023-09-11
TYPE:	UNIVERSITY OF OSLO
YEAR:	Dept. of Physics
SHEET:	1
KEY:	UCL copyright. All rights reserved.



PROJECT:	Frequency Shifter
DRAWING:	Power
DATE:	2025/02/03
USER:	REX@OSLO
SHEET:	1
KEY:	UNIVERSITY OF OSLO Dept. of Physics All rights reserved.

Appendix C

PCB layout and component placement of TFS PT4.1

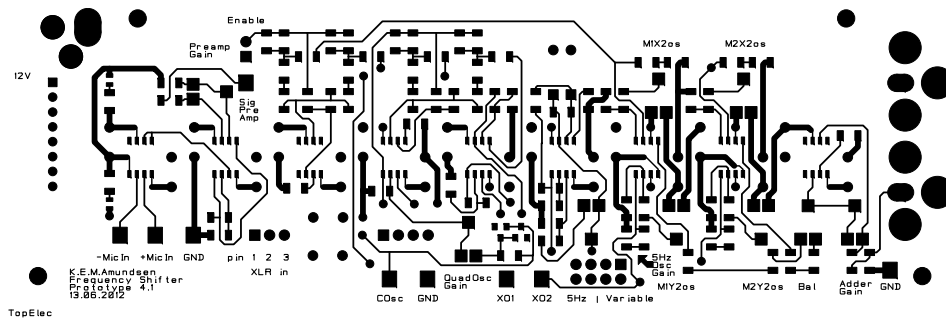


Figure C.1: Top electrical layer of TFS PT4.1 PCB layout. As the production of the PCB were done at the Electronic Workshop at UiO, no silk overlay was possible. Labels for connections and potentiometers were therefore etched in copper in the top electrical layer.

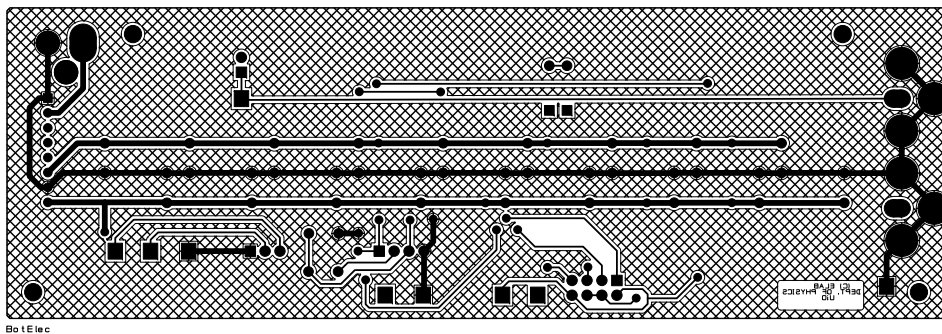


Figure C.2: Bottom electrical layer of TFS PT4.1 PCB layout. The ground grids main purpose was to balance the etching process, as both sides of the board should contain equal amounts of copper to prevent over/under etching of any of the sides.

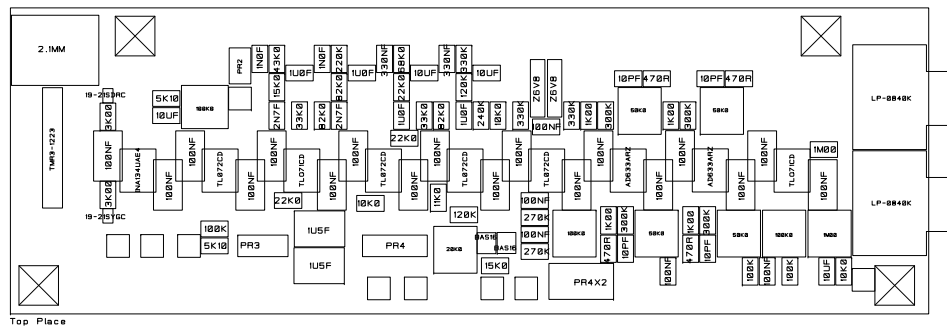


Figure C.3: Component placement on the PCB for TFS PT4.1. The squares without labels are various test points and ground connection for measurement purposes, the four large X-squares are screw holes and “2.1mm” is the DC power connector.

Appendix D

Part list of TFS PT4.1

Table D.1: Part list for TFS PT4.1. The part names correspond to the ElabLib component database, used by the Electronic Workshop at UiO.

Part name	Description	Qty	Component
AD-SSM/AD633ARZ	ANALOG MULTIPLIER, 4-QUADRANT	2	MULTI-2
CAP/100NF/1206R	10% 50V 1206 X7R	5	C16-18; C21; C24
CAP/10PF/1206R	5% 100V 1206 NP0	4	C19-20; C22-23
CAP/10UF/1206R	Kemet 10V 1206 Y5V	4	C1; C13-14; C25
CAP/1N0F/1206R	10% 100V 1206 X7R	2	C2; C7
CAP/1U0F/1206R	MURATA,TYP.GRM, 10% 16V	4	C4; C8; C11; C15
CAP/1U5F/MKS2	WIMA MKS2 10% 63V(5mm)	2	C5-6
CAP/2N7F/1206R	10% 50V 1206 X7R	2	C3; C9
CAP/330NF/1206R	10% 50V 1206 X7R	2	C10; C12
CAP/BYPASS/CER	KEMET GOLDEN MAX X7R/50V 10%	18	CB1-18
CON/ELMCH2/	BAT.ELEM.CONNECTOR 2.1 MM	1	POWER12V
CON/LP-0840K	LIH SHENG PREC.RCA CHASSIS CON.	2	OUTPUT PREAMPED
CON/PR2	LIH SHENG PREC.RCA CHASSIS CON.	1	ENABLE
CON/PR3	3 SCOTT ELEC. PINROW	1	XLRMIC
CON/PR4	4 SCOTT ELEC. PINROW	1	DUALPOTCON
CON/PR4X2	4X2 SCOTT ELEC. PINROW	1	OSCSEL
DIO/BAS16/SMD	SWITCH DIODE 75V/0.1A. SOT23	2	D3-4
DIO/Z6V8	BZX55Cnn ZENER DIODE .5W	2	Z1-2
INA134UAE4	AUDIO DIFF. LINE RECEIVER	1	DIFFREC
LED/19-21SDRC/SMD	SMD LED RED	1	D1
LED/19-21SYGC/SMD	SMD LED GREEN	1	D2
OPAMP/TL071CD/SMD	JFET-OPAMP 3Mhz, 13V/uS.	2	ADDER SHIFTAMP
OPAMP/TL072CD/SMD	DUAL JFET-OPAMP 3Mhz, 13V/uS	4	MICGAIN OLDOSC QUADOSC QUADOSCAMP
POT/100K/23B/SMD	BI-TECH CERMET TR.POT	3	BALANCE OSCAMPL PREAMPGAIN
POT/1M00/23B/SMD	BI-TECH CERMET TR.POT	1	ADDERGAIN
POT/20K0/23B/SMD	BI-TECH CERMET TR.POT	1	OSCGAIN
POT/50K0/23B/SMD	BI-TECH CERMET TR.POT	4	M1X2OFFSET M1Y2OFFSET M2X2OFFSET M2Y2OFFSET
RES/100K/1206R	RESISTOR KOA 1206 1% 0.25W	3	R4; R42; R44
RES/10K0/1206R	RESISTOR KOA 1206 1% 0.25W	3	R16; R23; R45
RES/11K0/1206R	RESISTOR KOA 1206 1% 0.25W	1	R19
RES/120K/1206R	RESISTOR KOA 1206 1% 0.25W	2	R21; R24
RES/15K0/1206R	RESISTOR KOA 1206 1% 0.25W	2	R7; R25
RES/1K00/1206R	RESISTOR KOA 1206 1% 0.25W	4	R30; R32; R36; R38
RES/1M00/1206R	RESISTOR KOA 1206 1% 0.25W	1	R43
RES/220K/1206R	RESISTOR KOA 1206 1% 0.25W	1	R10
RES/22K0/1206R	RESISTOR KOA 1206 1% 0.25W	3	R9; R14-15
RES/240K/1206R	RESISTOR KOA 1206 1% 0.25W	1	R22
RES/270K/1206R	RESISTOR KOA 1206 1% 0.25W	2	R27-28
RES/300K/1206R	RESISTOR KOA 1206 1% 0.25W	4	R31; R33; R37; R41
RES/330K/1206R	RESISTOR KOA 1206 1% 0.25W	3	R20; R26; R29
RES/33K0/1206R	RESISTOR KOA 1206 1% 0.25W	2	R8; R17
RES/3K00/1206R	RESISTOR KOA 1206 1% 0.25W	2	R1-2
RES/43K0/1206R	RESISTOR KOA 1206 1% 0.25W	1	R6
RES/470R/1206R	RESISTOR KOA 1206 1% 0.25W	4	R34-35; R39-40
RES/5K10/1206R	RESISTOR KOA 1206 1% 0.25W	2	R3; R5
RES/68K0/1206R	RESISTOR KOA 1206 1% 0.25W	1	R13
RES/82K0/1206R	RESISTOR KOA 1206 1% 0.25W	3	R11-12; R18
SPES/TMR3-1223	TRACOPOWER DC/DC CONVERTER	1	DCDC