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Further Development of a High Gain Microphone Preamplifier

API 500-Series Compatible Module

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The purpose of this thesis work was to improve an existing microphone preamplifier design for Sandhill Audio Oy in order to accomplish better distortion performance and better suitability for digital control of the unit.

The environment the preamplifier will be used in, as well as its boundary conditions are introduced in the first section of this thesis work. Most recording chain components are briefly explained and signal level ranges are investigated.

The main design focus was put on improving the analog circuit design, although, for the sake of clarity the basic structure of the digital control system is also briefly explained.
The front end gain structure was designed to accommodate a digitally controlled relay switching, the equalizer was changed from active to passive and the drive capability of the output section was drastically improved. Also an improved level indicator circuit was introduced.

To accomplish one of the most important targets, low distortion, characteristics of the input and output transformers were examined in depth. Also operational amplifier nonlinearities were measured and the circuit was fine-tuned for lowest possible distortion.

The result of this thesis work is an optimized design of the analog circuit operating in an API 500-series environment. The circuit is a very high performance preamplifier which is ready to be applied in a production unit with very little or no changes.

Keywords	microphone, preamplifier, analog design, op amp, transformer, ribbon microphone

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Koulutusohjelma	Sähkötekniikka
Suuntautumisvaihtoehto	Elektroniikka
Ohjaaja(t)	Heikki Valmu, yliopettaja

Tämän insinöörityön tarkoituksena oli parannella olemassa olevaa prototyyppiasteelle rakennettua mikrofonietuvahvistinta Sandhill Audio Oy:lle. Tavoitteena oli parantaa laitteen säröarvoja ja mukauttaa analogielektroniikka sopeutumaan digitaaliseen ohjausjärjestelmään.

Työn ensimmäisessä osassa tutkitaan laitteen käyttöympäristöä ja sen analogipiirin suunnittelulle asettamia reunaehtoja. Oleellisimmat äänitysketjun osat selitettään tässä osassa ja samalla tutkitaan sovellettavia signaalitasoja.

Suunnittelussa perehdyttiin pääasiallisesti analogielektroniikan optimointiin, mutta kokonaiskuvan vuoksi myös digitaalisen ohjausjärjestelmän toiminta on selitetty lyhyesti.
Jännitevahvistusasteen rakenne muokattiin toimimaan paremmin mikroprosessoriohjattujen releiden kanssa, taajuuskorjain muutettiin aktiivisesta passiiviseksi ja linjavahvistinasteen suorituskykyä ja luotettavuutta parannettiin huomattavasti. Samalla suunniteltiin paranneltu tasoilmaisin.

Asetettujen tavoitteiden saavuttamiseksi tuli tulo- ja lähtömuuntajien ominaisuuksia tutkia tarkemmin. Myös operaatiovahvistimien epälineaarisuutta tutkittiin ja kytkentä mukautettiin saadun tiedon perusteella tuottamaan mahdollisimman pienen harmonisen särön.

Työn tuloksena saatiin korkealaatuinen API-500 sarjaan yhteensopiva analogikytkentä jota voidaan soveltaa kaupallistettavaan tuotteeseen hyvin pienin muutoksin.

Avainsanat	Mikrofoni, etuvahvistin, analogielektroniikka, operaatiovahvistin, muuntaja, nauhamikrofoni

Contents

Abbreviations

A	ampere (SI-base unit)
AC	alternating current
AD	analog-to-digital
API	Automated Processes Inc.
BBC	British Broadcasting Company
CMRR	common mode rejection ratio
dB	relative level (decibel)
dBu, dBU	signal voltage level (re 0.775 Volt $_{RMS}$)
DC	direct current
DI	direct input, direct interface
E.I.N.	equivalent input noise
emf	electro motive force
FFC	flat flexible cable
H	Henry (SI-base unit)
Hz	Hertz (SI-base unit)
IC	integrated circuit
I/O	input/output
JFET	junction field effect transistor
LAN	local area network
LED	light emitting diode
op amp	operational amplifier
Pa	Pascal (SI-base unit)
PCB	printed circuit board
PWM	pulse width modulation
RCA	Radio Corporation of America
RMS	root mean square
S.I.D.	slewing induced distortion
SPL	sound pressure level
THD	total harmonic distortion
µC	microcontroller
V	Volt (SI-base unit)
FS	full scale

1 Introduction

The work presented in this thesis was done for Sandhill Audio Oy, a Finnish ribbon microphone manufacturer. The goal of this work was to examine and perform improvements to an ongoing microphone preamplifier design. The starting point was a working prototype of an API 500-series compatible preamplifier specially tailored for use with ribbon microphones and especially the Sandhill 6011A active ribbon microphone.

The prototype had been built to test the concept solely from an audio perspective. One part of this thesis was to redesign the analog circuit to accept a fully digital control circuit. In other words all analog switching would be performed with microcontroller driven relays instead of running signals through mechanical front panel switches. This arrangement would guarantee a longer working life for the unit and optimized circuit board layout in the critical circuit stages. Professional audio gear has a fairly long life span expectation compared to consumer audio and faulty switches are the most common source of problems in vintage units. Too often those switch types have become obsolete and the otherwise fully functional gear cannot be repaired. The other motive for designing a fully digital control system is having the option to convert the unit to running remotely controlled in the future with only minor changes to the circuit.

The preamplifier under development has a transformer coupled topology. Audio transformers are known for having a tonal signature of their own, thus changing the signal due to their electric and magnetic limitations. Therefore it was a question of examination to find out the parameters of the input and output transformers picked up for the design, both sourced from Swedish manufacturer Lundahl Transformers AB. However, since the line output stage of the preamplifier has a distortion cancelling low impedance topology, the characteristics of the input transformer were considered more interesting, it being the first component to meet the weak input signal.

The preamplifier utilizes modern JFET-input operational amplifiers for all stages. The first voltage gain stage has a wide adjustable gain range and one part of this thesis was to investigate if an uncompensated operational amplifier with appropriate external compensation could reliably be used for this purpose. The operational amplifier used for the initial prototype was otherwise sufficient, but at the highest gain settings the frequency response was compromised because the operational amplifier was pushed to its limits in terms of open loop bandwidth.

2 Microphone Preamplifier

2.1 Description

A microphone is an electroacoustic transducer which produces an electric signal proportional to variations in air pressure. In more detail microphones can either function as pressure or pressure gradient transducers. Some advanced measurement microphones can even measure the particle velocity and pressure components separately to reveal sound intensity (sound pressure per unit area).

A microphone preamplifier is required to serve two main purposes, voltage amplification and impedance matching. Generally microphone level signals are in the range of a few millivolts per Pascal. In terms of sound pressure level (SPL) 1 Pascal corresponds to 94dB SPL where 0dB refers to 20µPa. 20µPa in turn is the average threshold of human hearing at 1kHz frequency. Normal sensitivity ranges by use are divided into three groups by John Earle in Microphone book as can be seen in Table 1. These guidelines do not take self-noise into account, but they do give a ballpark.

Table 1 Normal sensitivity ranges by use. Reprinted from Eargle (2005) [1,110]

Microphone usage	Normal sensitivity range
Close-in, hand-held	2–8 mV/Pa
Normal studio use	7–20 mV/Pa
Distant pickup	10–50 mV/Pa

Typically in order to adapt to the sensitivity of the following signal chain the microphone's signal level has to be amplified by 0 to 80dB depending on the sensitivity of the microphone and the sound pressure produced by the sound source. The nominal line level for pro audio is +4dBu, 0dBu referring to $0.7746V_{RMS}$. A typical recording signal chain utilizing digital recording media is represented in Figure 1.

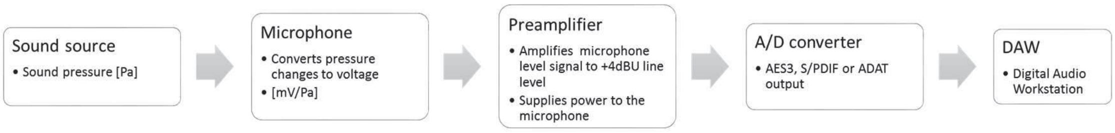


Figure 1 Block diagram of a recording signal chain in modern digital recording environment

Example: required amplification for human voice:

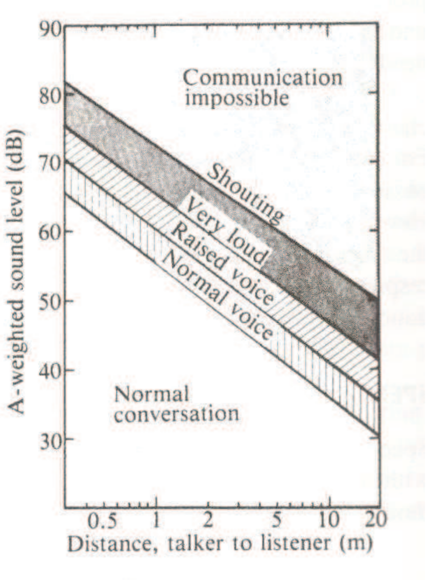


Figure 2 Human voice SPL vs. distance. Reprinted from Rossing, Moore & Wheeler (2002) [2, 728]

A typical A-weighted sound pressure level of human voice outdoors is represented in Figure 2. According to the figure, the SPL for normal voice at 1 meter distance from the talker is approximately 60dB. If recorded with a microphone having sensitivity of 30mV/Pa, the RMS voltage produced is:

$$V_{OUT} = 30mV * 10^{\frac{(60dB-94dB)}{20}} = 0{,}60mV_{RMS} \tag{1}$$

In order to achieve the nominal line level of +4dBu the voltage gain A_v has to be:

$$A_V = \frac{0{,}7746V_{RMS}*10^{\frac{+4dB}{20}}}{0{,}60*10^{-3}V_{RMS}} = 2046 \gg A_{dB} = 20*\log(2046) = 66dB \tag{2}$$

Microphones can be divided into two subgroups: active and passive. Active microphones have a built-in circuitry for impedance matching. This circuit may also include amplification and filtering. The active circuit usually makes the microphone less prone to interaction with the load that it is driving. Active microphones are either powered by their own external power supply or by so called phantom powering provided by the microphone preamplifier. Phantom powering is explained later in this thesis work. Passive microphones on the other hand are unbuffered transducers and they are more demanding in terms of loading. Usually their output impedance is frequency dependent and interactions with demanding loads are more unpredictable than with active microphones. Loads can affect the sensitivity and also degrade frequency and transient response of the microphone.

General requirements for a high quality microphone preamplifier are high amplification, flat frequency response low noise and low distortion. The first three can be considered undisputed but sometimes certain preamplifiers are preferred over others for their sonic signature. This usually corresponds to their overdrive characteristics over other parameters. A soundscape of various sound sources is a complex system highly dependent on psychoacoustic phenomena and nonlinear nature of human hearing. Occasionally a slightly distorted tone with enhanced low order harmonics can be more adequate in a certain context than a pristine reproduction of the original signal.

2.2 Phantom Powering

Phantom power is a method of powering microphones through a differential balanced transmission line that is primarily used for carrying the output signal of a microphone. It's a positive 12-48V voltage applied to XLR connector's pins 2 and 3 with respect to pin 1. This powering method is described in the international standard IEC 61938 and it consists of three variants P12, P24 and P48, the latter being the most common in pro audio and therefore the only one explained here.

The nominal DC voltage for P48-type phantom power is 48 ±4 Volts. Phantom power is used for powering the active circuitry of a microphone and in capacitor microphones the polarization voltage of the microphone capsule is also derived from phantom power.

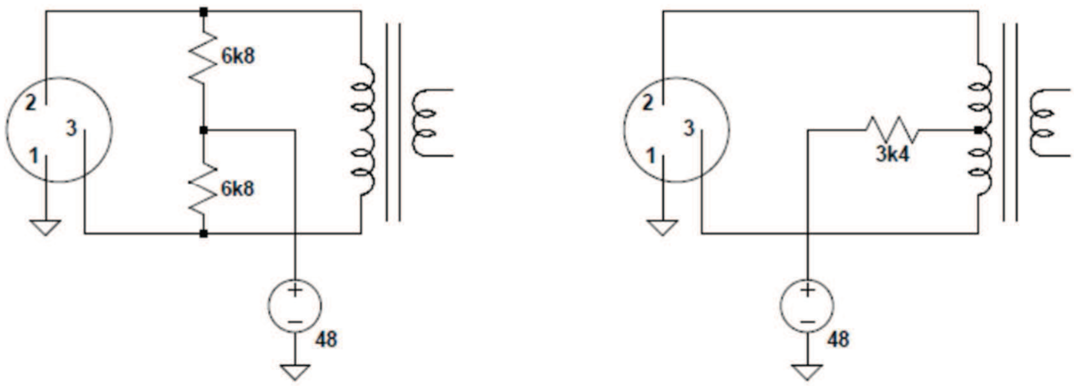


Figure 3 Phantom power feed circuits

Phantom power can be fed into the balanced signal line (pins 2 and 3 in an XLR connector) in two different ways described in Figure 3. Pin 1 is used as a voltage reference and it's tied to the ground lead of the cable, the shield or both. The most common method utilizes two feed resistors, typically having resistance of 6.8kΩ. These resistors have to be closely matched (within 0.4%) to achieve good common mode performance, however the actual value of these resistors is not critical. The drawback with this method is the fact that the maximum input impedance is limited by the phantom feed resistors as they are effectively in parallel with the input resistance of the preamplifier circuit [1, 117-120].

The other method is to feed the phantom through a center tap of an input transformer primary. Naturally this is an option only with preamplifiers that do have an input transformer and in addition one with a suitable primary winding with a symmetrical center tap. The current is limited with one resistor with half the value of the feed resistors, hence 3.4kΩ. In this configuration the common mode performance is determined solely by the input transformers characteristics and the input impedance of the preamplifier is not affected. In both circuits the short condition current is limited to 14mA by the phantom feed resistor(s). According to the IEC standard the maximum current consumption per microphone is 10mA. Power is separated from the balanced signal in a similar way to the phantom feed using either two matched resistors or a center tap of the output transformer secondary.

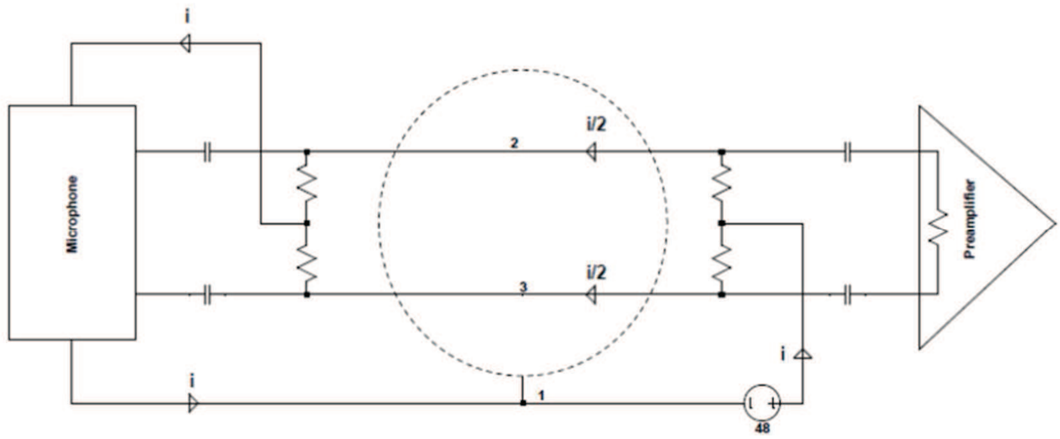


Figure 4 Phantom power current path

In Figure 4 the whole phantom current path is illustrated. It can clearly be seen that should any noise be present in the phantom power, imbalance in the circuit would convert it into a differential signal.

2.3 Connectivity and Level Compatibility

Majority of pro audio devices are connected to each other thorough balanced signal lines typically using 3-pin XLR-connectors. Some devices also have unbalanced connections, more commonly found in consumer level audio. In addition to the previous, microphone preamplifiers sometimes come with an additional unbalanced high impedance input for instrument signals, hence the name direct input, direct interface or D.I.

Microphone preamplifier has to amplify the microphone level signal to a nominal line level of 4dBu.Typically the recorded material is unpredictable and some headroom is required to avoid overloading during peaks. The headroom needed above the nominal level will be different depending on the devices following the preamplifier.

Since the early 1990's digital recording has become increasingly popular and most of audio recording and post processing is now days performed in digital domain. In AD-converters the nominal analog voltage level of +4dBu corresponds to -16dBFS (American standard) and -14dBFS (European standard). Respectively, the output levels required for 0dBFS level are +20dBu and +18dBu. Quality analog gear will handle levels of +22dBu or more and saturation is not considered as hazardous as in digital domain.

On contrary, distortion created by digital devices is not harmonically related to the signal and is considered dissonant and harsh sounding.

To be compatible with any gear the output stage of a microphone preamplifier should have output capability of 22-24dBu.

3 Practical Signal Sources

3.1 Microphone Categorization

A microphone is a transducer whose task is to convert an incident sound (pressure or particle velocity) into an electric signal. The strength of this signal is usually represented as voltage. However, the term microphone is commonly used for both the transducer and the whole device containing the transducer and surrounding components. Microphones can be divided into various categories by the for example the purpose of use or operation principle of the transducer [3, 34].

Recording studio and broadcasting microphones can roughly be divided into two subgroups, capacitor microphones and dynamic microphones. In addition to the principle of operation the most common technical details that determine the practical operation are:

•sensitivity	the output RMS-voltage produced by the microphone at a specified sound pressure level. Typically measured at 94dB and 1 kHz
•frequency response	the output voltage level as a function of frequency, often normalized to the output at 1kHz and shown in dB
•polar pattern	the sensitivity of the microphone as a function of rotation angle. The polar pattern is also frequency dependent (see Figure 7)

3.2 Capacitor Microphone

In a capacitor microphone the transducer consists of a capacitor whose capacitance varies proportionally to air pressure. The change in capacitance occurs due to variation in

distance between the plates of the capacitor, also known as diaphragm and back plate. In order to operate as a transducer the capacitor requires a high impedance series load and a constant voltage called polarization voltage applied across the circuit. The capacitor and series load, which usually is a resistor of very high value form a voltage divider whose output is proportional to pressure changes in air.
The circuit shown in Figure 5 has a very high impedance and therefore requires an external impedance converter in close proximity to the transducer.

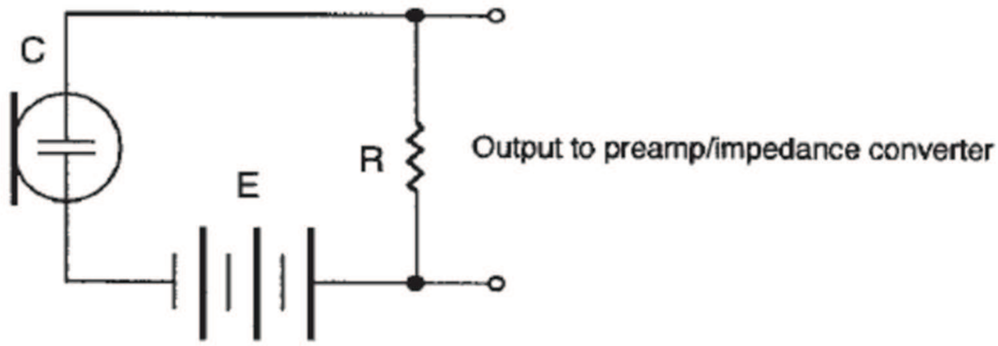


Figure 5 Simplified electrical circuit of a capacitor microphone. Reprinted from Eargle (2005) [1, 23]

Depending on the construction of the microphone capsule the capacitor transducer can be either a pressure microphone, a pressure gradient microphone or a combination of the both.

3.3 Dynamic Microphone

In a dynamic microphone a conductor is placed in a magnetic field. When the conductor moves forced by changes in air pressure, an electro motive force is applied between the ends of the conductor. Dynamic microphones can be further divided into groups of moving coil microphones shown in Figure 6 and ribbon microphones. Ribbon microphones will be discussed in more detail in this thesis work because the designed preamplifier will be tailored to work optimally with a certain ribbon microphone.

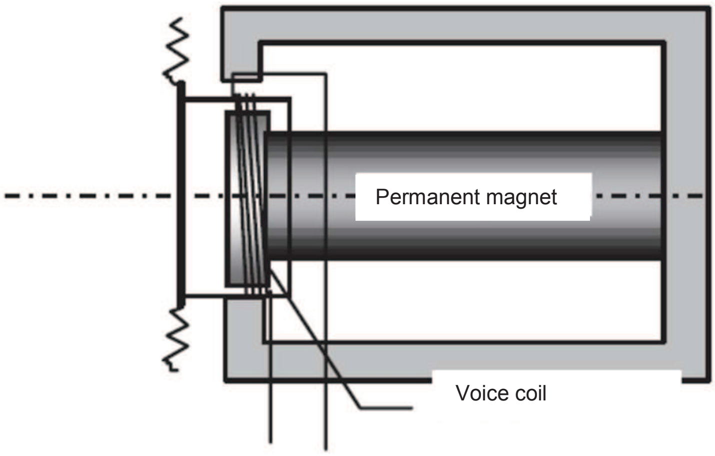


Figure 6 Moving coil dynamic microphone. Modified from Backman (2008) [3,104]

3.4 Ribbon Microphone

3.4.1 Description

A ribbon microphone was invented by German physicist Walter H. Schottky and inventor Dr. Erwin Gerlach while working for Siemens laboratories during the 1920's. The ribbon transducer consists of a very fine ribbon shaped conductor placed in a constant magnetic field. When sound waves displace the ribbon, its velocity induces an electromotive force between the ends of the ribbon according the Lorenz Force Law. Sound waves are transverse vibration and the ribbon is forced to move back and forth by the variations in pressure, thus a continuous time dependent signal is produced. Schottky and Gerlach's invention was based around field coil structure (a.k.a. electromagnet) and this arrangement had some fundamental flaws mostly due to the weak magnetic field and noise. However, it became clear already at this point that the transducer has a reciprocal nature. In other words it could also be used be for transforming electrical energy into sound. After the evolution of permanent magnets the invention was first commercialized by American experimental physicist Harry F. Olson at RCA Viktor in the 1930's. The first commercially meaningful product, RCA 44A, was released in 1934. In the 44A the magnetic field was produced by a single horse shoe shaped AlNiCo (aluminum-nickel-copper) permanent magnet. The 44A was followed by 44B and 44BX, both build around two magnets, and the models stayed unchanged until the whole RCA 44 family of products was finally discontinued in the mid 1950's [21].

3.4.2 Principle of Operation

A ribbon microphone is a pressure gradient transducer. This means that the ribbon is forced to move by the pressure difference between across it. The difference is determined by the wavelength of the sound, the physical dimensions of the transducer and the incidence angle of the sound wave. The frequency response of a pressure gradient microphone increases 6dB/octave and on the other hand far above the mechanical resonance frequency the frequency response of a dynamic microphone increases by 6dB/octave. This would in theory predict a flat response if the ribbon is tuned below the useable frequency range. However, in real life the low frequency range contains some longitudinal harmonics of the mechanic resonant frequency and the upper range is limited due to interference of incident and diffracted waves. [3, 110]

The minimum points in the frequency range are periodic and the minimum in the frequency response is detected every time the so called front-to-back path (a dimension determined by the physical dimensions of the transducer) is a multiple of the wavelength. There is usually only one diffraction minimum within the audio range. At these frequencies the pressure gradient of the sound field (=the difference in pressure) becomes zero due to the deflecting sound waves having the same phase. This dependency also affects the transducers polar pattern making it bi-directional. A polar pattern of this kind is also called figure-8 for resemblance of the digit 8 in 2D- plane as can be seen from Figure 7.

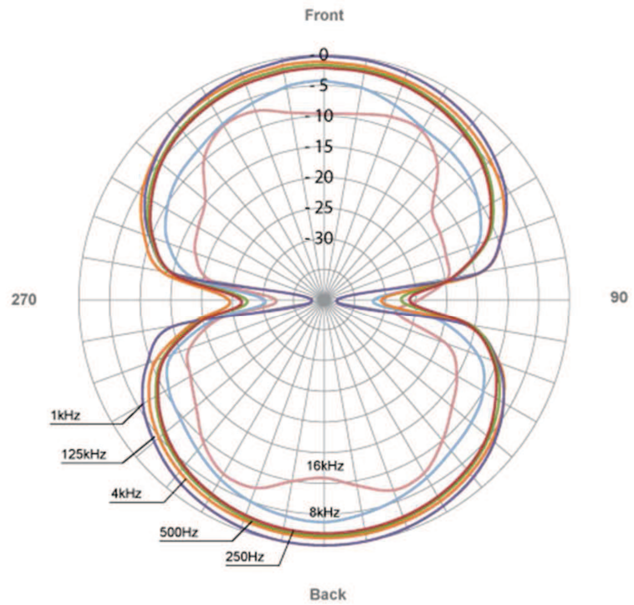


Figure 7 Ribbon microphone polar pattern showing the bi-directivity of a pressure gradient microphone. Reprinted from Sandhill Audio [4]

3.4.3 Electric Properties

The resistance of the ribbon is in range of few hundred milliohms depending on its physical dimensions. Traditional material for the ribbon element is aluminum with thickness ranging from 1.5 to 3µm. Ribbon dimensions vary between different models and manufacturers but usually the width is a few millimetres and length is less or equal to 60mm. The corrugated ribbon is placed inside a magnetic loop consisting of a permanent magnet and a return path made of iron or steel.

The movement of the ribbon produces a very small emf and making use of this tiny signal requires impedance matching with a step-up transformer. Ribbon microphone transformers come with a turns-ratio of at least 1:25. Some ribbon microphones utilize a transformer with ratio as high as 1:100 but without active buffering the output impedance of these devices is too high for most microphone preamplifiers. The transformer, however, provides virtually noise free voltage gain before active impedance conversion. Loading causes more current to flow in the primary circuit forced by the emf. This current creates a force opposite in direction to the one caused by the pressure gradient, thus the ribbon movement will be damped. With passive ribbon microphones a large preamplifier input impedance is usually preferred due to less damping and smaller voltage division between the source and the load. However some users prefer the effects caused by loading.

3.4.4 Requirements for Preamplifier

Ribbon microphones generally have a low output compared to other microphone types. Their sensitivity usually ranges from 0.2mV to 10mV with some exceptions. Moving coil microphones are in the same sensitivity range, but they benefit from the option of being placed closer to the sound source without significant amount of tone coloration. Ribbon microphones have a strong proximity effect. This means that the sensitivity of a pressure gradient microphone, such as a ribbon microphone, increases by 6dB/octave in a close proximity of a sphere wave centre compared to the sensitivity defined on a planar wave. As mentioned in chapter 4.4.2 there is a minimum in the frequency range in the upper half of the audio range. Having the opportunity to boost these frequencies slightly and cut the low frequencies is a useful feature. [3, 42]

Due to the low sensitivity ribbon microphones usually require more amplification from the preamplifier than other microphone types. The resistive nature of the ribbon material also creates thermal noise, which is the dominant form of noise in ribbon microphones limiting the low end of the dynamic range. Passive ribbon microphones have a frequency dependent impedance and therefore they greatly benefit from the use of a bridging input load.

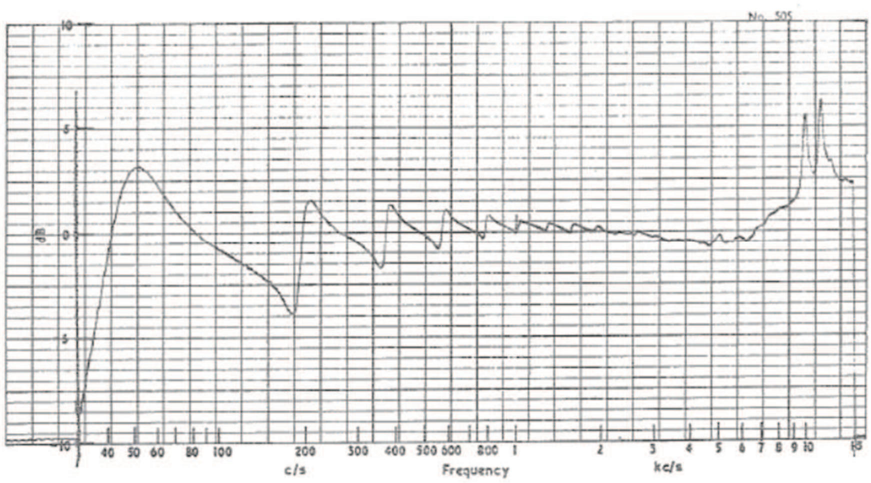


Figure 8 Example impedance curve of a passive ribbon microphone. The impedance variation is shown in decibels relative to a dummy microphone placed in the adjacent arm of a motional impedance bridge, frequency of x-axis. Reprinted from Shorter & Harwood [5, 11]

An impedance plot of a BBC experimental microphone is shown in Figure 8. Using a very high impedance load compared to the impedance of the microphone minimizes the frequency response effects from variation in microphone impedance and therefore flattens the frequency response of the system.

4 API 500-Series

4.1 Description

Automated Processes Inc. is an American company established in 1968. They pioneered the modular recording console design in the late 1960's and 1970's. API recording consoles utilized a motherboard design that later become a standard in personal computer design. The consoles provided signal routing, inputs, outputs and powering for modular preamplifiers and equalizers and dynamics processors. The modular units were connected to the motherboard via 15-pin card edge connectors and could be removed by opening two screws in the front panel. API called this design the 500-series. More recently API launched a modular racking system called Lunchbox. Since its release in 1985 it has become increasingly popular among recording engineers and many manufacturers have started producing their own modules for it. The major advantages of this design are low cost due to a shared power supply and easy customizability. In the recent 20 years digital recording has become an industry standard and mobile recording gear, consisting of only a computer, preamplifiers and AD-converters is often all that is needed for high quality recording. API racks come in 6- 8- and 10-slot versions and multiple other manufacturers have started producing their own similar racks

4.2 I/O

The 500-series modules are connected to the motherboard via a 15-pin EDAC card edge connector. The original pinout is specified for various API modules as described in Table 2 below. A preamplifier follows a connection scheme of a 512c preamplifier module. The first Lunchbox modules had low level (-2dB) inputs and outputs but this option is no longer available.

Table 2 API 500-Series pin configuration

Pin	Connection	Module
1	CHASSIS	
2	OUTPUT ı4dB (ı)	512c 525 550b 560b
3	OUTPUT -2dB (ı)	512c 550b 560b
4	OUTPUT LOW	512c 525 550b 560b
5	AUDIO GND	
6	STEREO LINK	525
	N/C	512c 550b 560b
7	INPUT LOW -2dB	560b
	INPUT LOW +4dB	512c 525
	N/C	550b
8	INPUT LOW ı4dB	512b 525 560b
	INPUT LOW (GND)	550b
9	INPUT HI 2dB	550b 560b
	INPUT HI ı4dB	512c 525
10	INPUT +4dB	512c 525 550b 560b
11	GAIN ADJ.	550b 560b
	N/C	512c 525
12	ı16VDC	
13	POWER GND	
14	-16VDC	
15	ı48V PHANTOM	512c

5 Design Architecture

5.1 Analog Circuit Architecture

5.1.1 ISA Heritage

The preamplifier gain stage is loosely based on a topology first introduced by British company Focusrite in their ISA110 preamp module. The ISA amplifier was designed by Focusrite founder Rupert Neve in 1985 for a series of recording consoles built for George Martins AIR Montserrat studios. After more than three decades variations of the same circuit are still being manufactured by Focusrite. The gain stage was built around a then new NE5534 op amp which is a bipolar input low-noise high-speed audio operational amplifier. The ISA 110 had a maximum gain of 70dB and the output transformer was driven with a discrete NPN-PNP complimentary pair.

Because of the bipolar input op amps, the ISA circuit had multiple large DC-blocking capacitors in the signal chain. Electrolytic capacitors are known to be the least preferred option for coupling audio signals. They have high dissipation factor and a high dielectric absorption, both causing unwanted effects in the audio range. High dissipation factor results in a phase displacement, thus distortion and high dielectric absorption causes lack of sound clarity due to residual recharging effect within the capacitor. In addition to the previous, electrolytic capacitors are notorious for having a short operating life [6]

The preamp circuit was designed to have a minimal amount of capacitors and those that are crucial are preferably polypropylene capacitors. This is accomplished by using a JFET input op amps that have very low bias currents and low offset voltages. Polypropylene is widely accepted as the best all-around dielectric for capacitors used in audio with a wide selection of capacitances available. The only drawback is their relatively large size which will have to be taken into consideration in PCB layout design. Polypropylene capacitors are not available in surface mounted versions so they will add to the number of hand-soldered components.

5.1.2 Design Priorities

In the very beginning of the preamplifier design a decision was made to design a transparent sounding amplifier. Because the preamplifier is tailored for a specific ribbon microphone, it is highly likely that it will be used for amplifying other ribbon microphones as well also passive ones with a very low sensitivity and high self-noise. Therefore the preamplifier must have a high gain reserve and very low noise.

Usually preamplifiers that are designed to provide “transparent” or “clean” gain use transformerless input and output topologies. This is due to the fact that transformers have limitations that if not designed right will add substantial amount of tonal signature. Transformer imperfections will be discussed later in chapter 6.1.2. However, transformers have great benefits over transformerless solutions in improved signal balance and with careful design a transparent preamplifier can be built. An input transformer also helps to reduce the self-noise of the amplifier by providing virtually noise free step-up voltage gain when the signal enters the device.

In other stages of the circuitry low noise and low distortion were the primary targets. It was necessary to substitute the 19MHz ADA4627-1 in the gain stage with an uncompensated 79MHz ADA4637-1 for more bandwidth. This was essential in order to keep the frequency response perfectly flat at maximum gain.

5.1.3 PCB Layout

The initial preamplifier prototype was designed and built using through-hole components. This is not a cost effective way of building a product that will be sold in multiples of more than few tens so one goal of this thesis work was to convert the design to contain as many surface mounted components as possible. Only transformers and a handful of larger components such as capacitors and connectors will be hand-assembled. The parts count was to be kept as low as possible to optimize the pick-and-place machine setup. The actual PCB layout design will be performed with a software called Sprint Layout by ABACOM. It is a low level design program for layout design only, hence it has no schematic editor. Therefore some extra attention must be paid in order to avoid layout errors. However, for example creating custom components is a lot less time consuming than it would be with a more complex design tools such as Mentor Graphics PADS.

5.1.4 Design Architecture

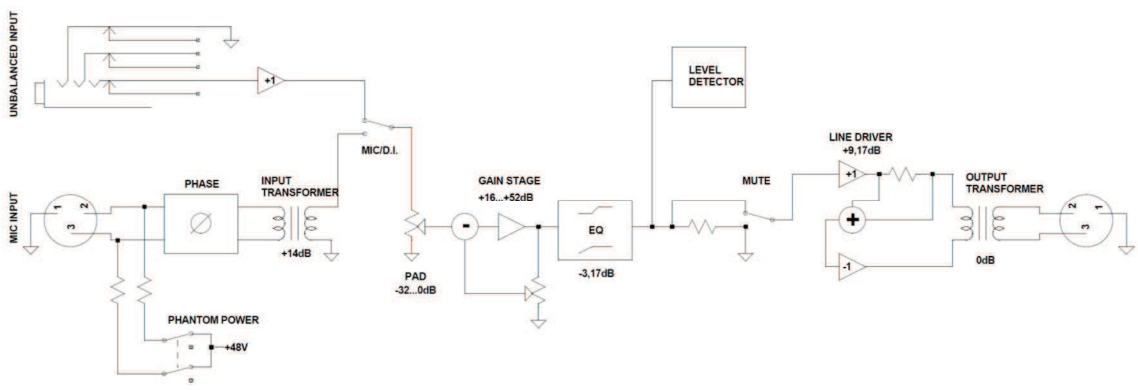


Figure 9 Analog circuit block diagram

A block diagram of the analog circuit is represented in Figure 9. The biggest topology-level changes from the prototype are the addition of an unbalanced DI and an improved mute circuit

5.2 Digital Control Circuit

5.2.1 Design Goals

Although the main emphasis of this thesis work is put on improving the analog circuit of the preamplifier, the digital control circuit is also briefly described.

The goal was to convert all switching in the device from mechanical switches to microprocessor controlled relays. Relays provide a hermetic means of switching signals and enable the design with optimized PCB layout.

Often in a recording situation it is necessary to have the microphone preamplifier placed close to the signal source if the distance between the source and the recording media is very long. Running very long microphone level lines can lead to loading and noise problems which may, if not get recognized in situ, ruin the whole recording. Therefore having the opportunity to control the preamplifier remotely from a distance would be ideal. There are a couple of LAN-based solutions in the market, such as the AMS Neve 4081 in Figure 10, but the possibility of having a reliable wireless system is worth investigating in the future. To allow easy upgradeability, the microcontroller was placed on a separate PCB which can easily be changed without any changes to the analog motherboard in the future should a remote controllability be required. This would be the case if the manufacturer would design a rack with remote controlled master µC.

Figure 10 AMS-Neve 4081 Quad Mic Amp. Reprinted from AMS Neve Ltd (2011) [22, 1]

5.2.2 ELMA X4 Hall Effect Switch

X4 is a rather recent addition to the ELMA range of switching components. The operation of the rotary switch is contactless and relies on a magnetic Hall-effect sensing system. The switch has a microcontroller built in which operates over a voltage range from $2.85V_{DC}$ to $5.25V_{DC}$. The X4 switch also has an optional momentary push button and the output is either absolute (Gray) code or encoder. The switch also has a PWM and analog outputs but they are not of interest in this application. The only possible drawback is the sensitivity to external magnetic fields which could potentially cause problems. The switch is available with two different connectors, FFC as seen on left and Micro MaTch on right in Figure 11.

Figure 11 ELMA X4 Hall-effect rotary switch. Reprinted from Elma Electronics Inc. (2017) [23, 1]

In this application a rotary encoder could be used but then a display would be required to indicate the gain. Recording engineers are used to mechanical switches so having a digitally controlled circuit with functionality that users are used to is preferred over the rotary encoder. Another great feature is that X4 has exactly the same indexing feel as the very common 04-series of mechanical switches. A 23-position version with a stop pin was chosen to provide gain range from 6dB to 72dB in 3dB increments. Thanks to the upgraded input op amp the analog circuit has enough bandwidth to allow shifting the gain range to even higher by changing the program.

5.2.3 Functions

In addition to gain switching the preamplifier has five switching functions:

- Mute
- Phantom power
- High pass filter
- Shelving filter
- Phase reversal

The mute relay will short the signal to ground between the equalizer and the line output stage muting the output of the device. The microcontroller switches the mute on as default at start-up and also silences the output while other switching functions is performed. This will prevent any unwanted switching noises that could occur. The mute is switched on slightly before the switching action and released automatically after. The Mute function can also be manually switched on from the built-in push button on the ELMA X4 rotary switch.

The remaining functions will have separate LED illuminated tactile switches placed in the front panel.

6 Circuit Design

6.1 Transformers

6.1.1 Transformer Coupling

In the vacuum tube era all preamplifiers used transformers for input and output coupling. Transformers offered a nearly perfect high broadband impedance balancing (i.e., 60 dB or better) [7, 489], hence a very good CMRR and virtually noise-free voltage gain that was necessary in order to achieve good signal-to-noise ratio with vacuum tube circuits. Transformers also provide galvanic isolation which is very useful especially with high voltage vacuum tube circuits. The alternative way of DC-decoupling at very low source impedances would require large capacitors and electrolytic capacitors are known for their relatively short life span and less than perfect impedance characteristics. Although vast majority of microphone preamplifiers of the modern day use transformerless topologies in both input and output there are undisputable benefits from using transformer coupling. Transformerless inputs and outputs can offer good balance and noise rejection in the audio range but they often fall down at radio frequencies. However, a good transformer is expensive to manufacture and takes a lot of room in the amplifier chassis adding cost, size and weight to a product. In the 1970’s and 1980’s manufacturers came up with advanced solid state based circuits utilizing feedback to compensate for the distortion from using smaller size transformers.

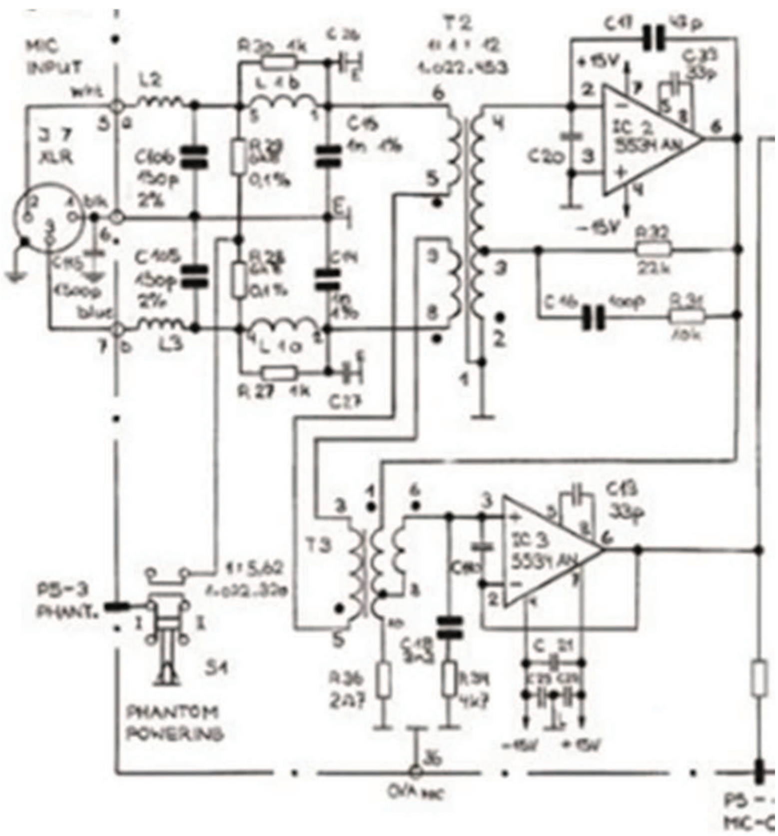


Figure 12 Studer 961/962 mic input schematic. Reprinted from Studer (1990) [8]

A good example is Studer 962 preamplifier in Figure 12 introduced in the late-eighties. Another topology called mixed feedback output driver used in the Sandhill preamplifier origins from a German patent DE 29 01 5671979 filed in 1979 will be discussed in more depth later in this thesis work.

6.1.2 Transformer Imperfections

An audio transformer does have some practical limitations due to the physical construction and design. The operation of a transformer is dependent upon the source and load connected to it. The primary impedance, which is mainly comprised of the primary winding inductance should be significantly larger than the reflected load impedance in order to achieve a flat frequency response. [9] Transformers have a limited bandwidth and they do induce some losses and distortion. In the following some mechanisms of transformer issues are explained in more detail.

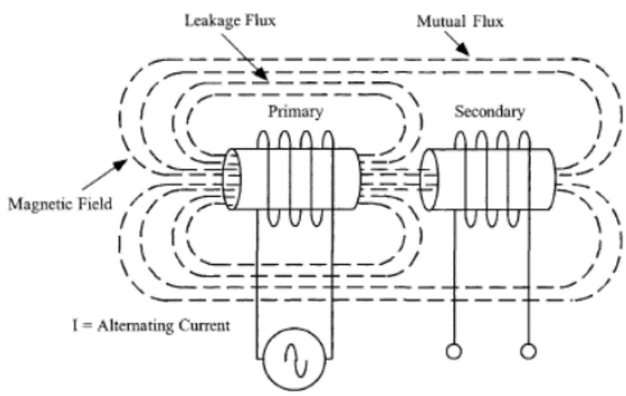


Figure 13 Simplified transformer. Reprinted from McLyman (2004) [10, 4]

Leakage inductance is the part of the primary induced flux that does not link to the secondary. This is illustrated in Figure 13. For simplicity's sake it is often shown as a lumped constant although it is actually distributed throughout the windings. In the same manner winding capacitances are shown as simplified lumped elements. An equivalent circuit for a transformer is shown in Figure 14.

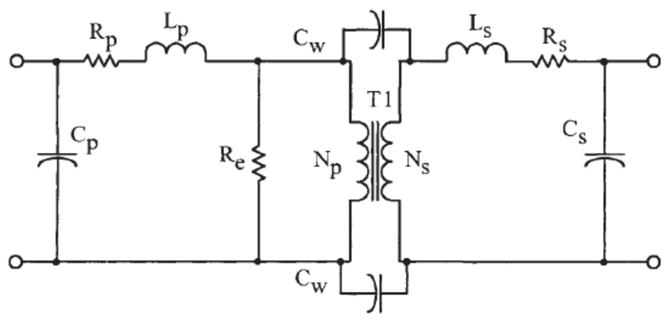


Figure 14 Transformer equivalent circuit. Reprinted from McLyman (2004) [10, 427]

In this circuit leakage inductances for primary and secondary are represented by Lp and Ls and the corresponding winding capacitances by Cp and Cs. Rp and Rs are the primary and secondary resistances and Re represents the core loss. Cw is winding-to-winding capacitance. Leakage inductance can be minimized by interleaving primary and secondary windings. However leakage inductance and winding capacitance are inversely proportional, thus if one is decreased, the other will increase [10].

In a recording situation it's often a matter of context and taste whether transformer coloration is considered desirable. Although most quality audio transformers do have a flat

frequency response in the audio range and beyond (20Hz-20kHz) at a moderate signal level, the level handling at low frequencies is limited resulting in harmonic distortion with higher level signals. Transformer saturation is often described as a mechanism providing “warmth” to the sound and can in certain contexts be used as an advantage. It is not uncommon that transformers, especially in the output stage, are driven into saturation on purpose. Transformers also add some distortion at all frequencies due to the hysteresis phenomena but that is rarely a problem.
Transformers also have a tendency to ringing in the high frequency range due to a self-resonant condition formed by leakage inductance and winding capacitances. A common way to reduce the ringing is connecting a damping circuit, sometimes called Zobel-network, across the transformer secondary winding. In a transient response measurement this phenomenon can be seen as an overshoot or decaying ringing on a leading edge of a square wave test signal.

6.1.3 Lundahl Transformers

Both transformers used in the Sandhill preamplifier design are manufactured by Swedish manufacturer Lundahl Transformers AB. Lundahl transformers differ from most audio transformers in the way the cores are constructed. All Lundahls are wound bobbinless on in-house manufactured laminated C-cores made of either mu-metal or amorphous steel. They have static shields between the interleaved primary and secondary windings and most of their microphone transformers are also externally mu-metal shielded. Each winding layer is separately insulated yielding lower winding capacitances and better winding repeatability

6.2 Transformer Measurements

In order to verify the manufacturer given specifications and get more in-depth information some basic measurements were carried out.

6.2.1 Lundahl LL1538 Microphone Input Transformer

Lundahl 1538 is a mu-metal core 1+1:5 microphone transformer with three-section coils. The key design values according to the manufacturers datasheet (appendix 1):

- Static resistance of each primary 44Ω
- Static resistance of each secondary 880Ω

- Distortion (primaries connected in parallel, source impedance 200Ω) — 0.2%@0dBu primary level, 50Hz; 1%@10dBu primary level,50Hz

- Self-resonance point — >120kHz

The inductances and level handling of the input transformer were examined. The transformer was measured in 1:5 configuration hence primaries wired in parallel. Coil inductances in audio transformers are tricky to measure due to the distributed capacitance in them. It was not possible to measure the primary inductance with an LCR-meter as the results were highly frequency dependent and clearly the measurement was disturbed by the winding capacitances. Transformers are also known to be level dependent what comes to inductance. For the primary inductance measurement a known resistance was connected in series with the coil and the corner frequency of the formed RL-filter was used to calculate the inductance. The resistance value was chosen so that the corner frequency was below 100Hz. The inductance was measured at three different voltage levels in 10dB increments. Leakage inductance was measured with a handheld LCR-meter and the values were not frequency dependent.

The self-resonance frequency was measured both with a swept sine wave and a 10kHz square wave. The self-resonance appeared as a decayed ringing on the leading edge of the test signal and is shown in Figure 15. The self-resonance is at such a high frequency that it’s highly unlikely to have any sonic effect. Therefore a Zobel-network is not considered necessary.

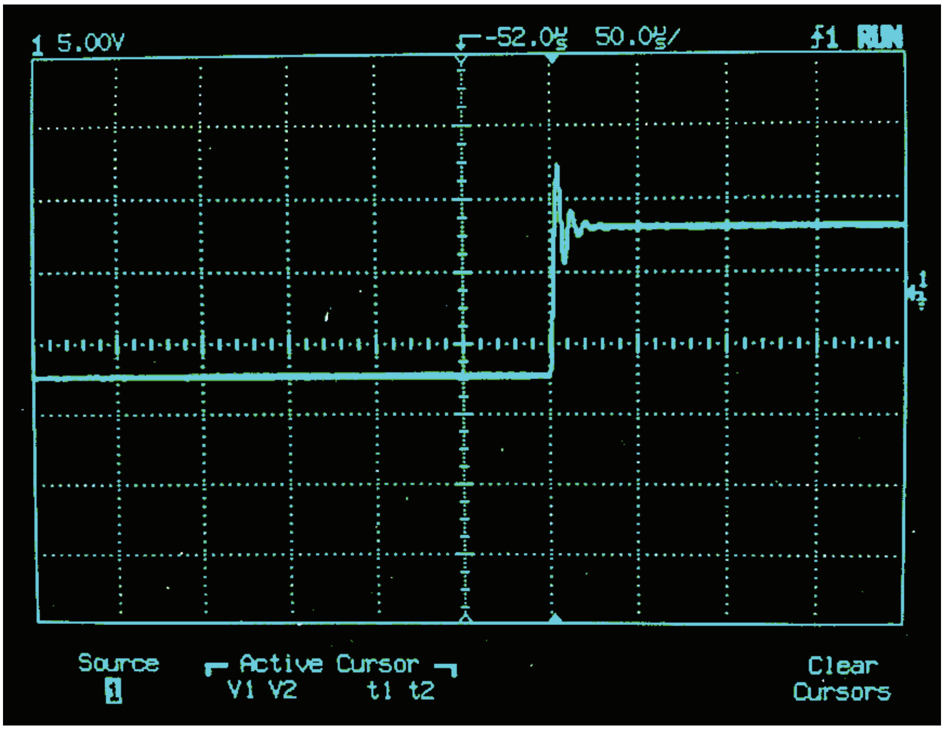


Figure 15 Transformer overshoot and ringing with a 10 Volt 10kHz square wave output. Voltage on y-axis (5V/div.) and time on x-axis (50µs/div.).

Measured values:

• Primary resistance	25.2Ω
• Primary inductance	20H(±20%)
• Secondary resistance	805.4Ω
• Leakage inductance (primary to secondary)	320uH
• Distortion (primaries connected in parallel source impedance 50Ω)	0.2%@0dBu primary level, 50Hz 1%@+13dBu primary level,50Hz
• Self-resonance frequency	121kHz

The transformer was measured for low frequency headroom and distortion. The input level resulting in 0.1% and 1.0% THD+noise was measured from 10Hz to 300Hz.The 0.1% THD+noise at 10-20Hz was immeasurable due to measurement system noise and distortion. The results are shown in Figure 16. The generator residual THD+noise was 0.004% at 1Vrms level.

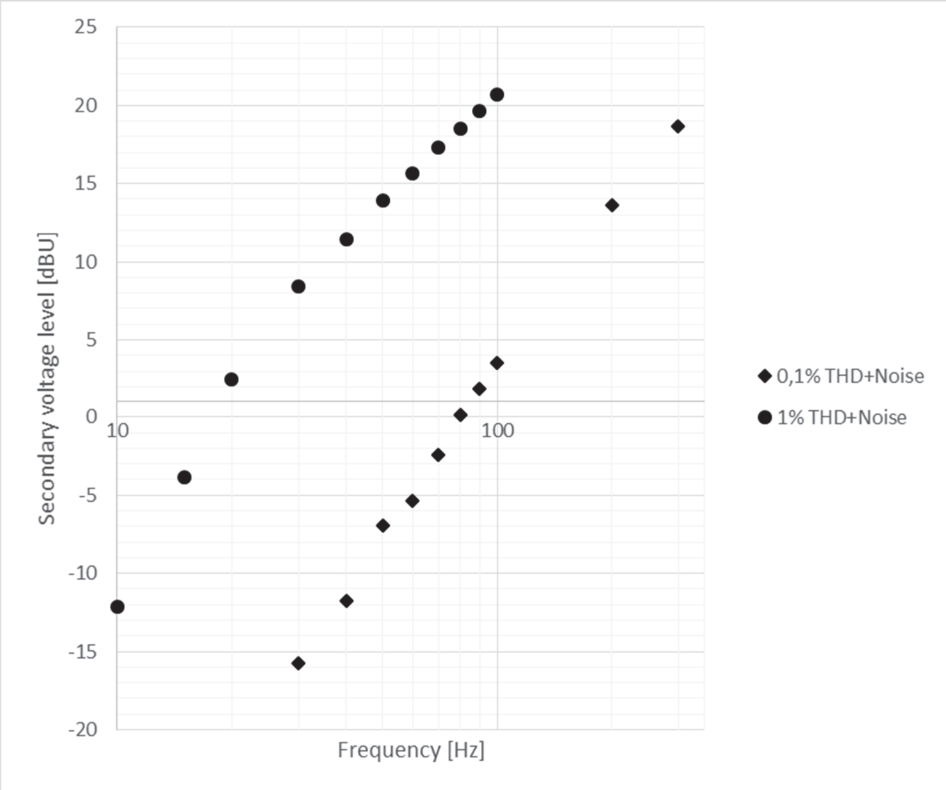


Figure 16 LL1538 low frequency headroom for 0.1% and 1% THD+Noise, frequency on x-axis and secondary voltage level in dBu on y-axis

The manufacturer guarantees an input level of 10dBu at 50Hz into an infinite load from a 200 ohm source impedance at 1.0% THD. Considering the slight differences in measurement system specifications the results agreed well with the published figures. However an unexpected amount of second harmonic was present before saturation level. The output (upper) and distortion (lower) waveforms at 50Hz and 1% THD+noise are shown in Figure 17.

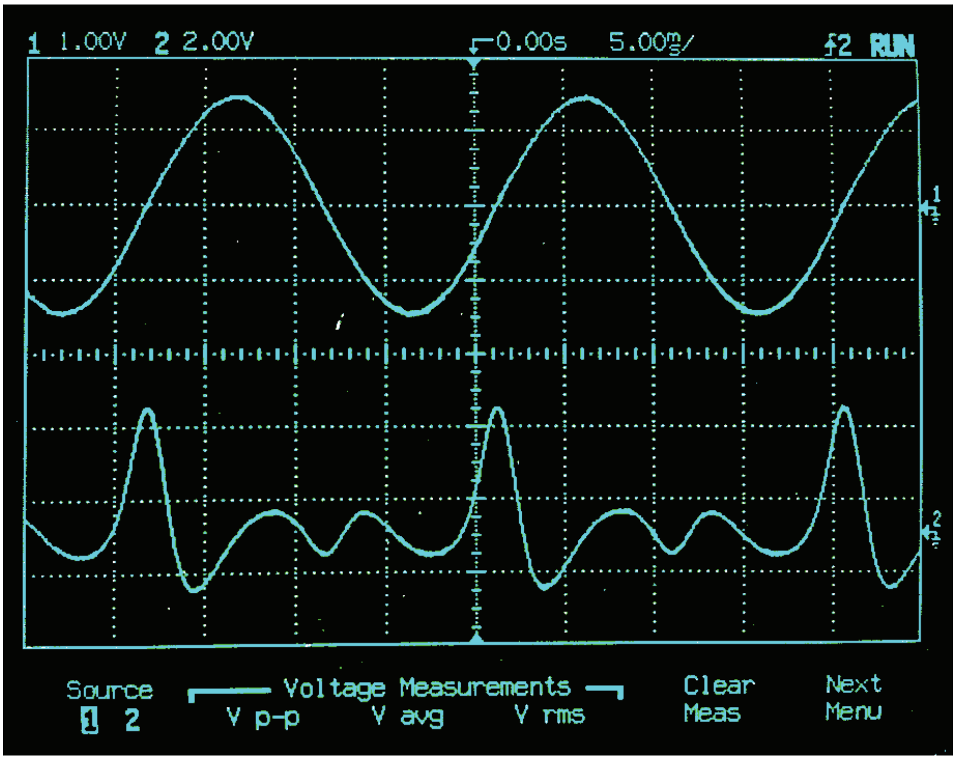


Figure 17 Output and distortion waveforms at 1% THD+noise, 50Hz. The signals are sourced from distortion analyser monitor outputs and shown voltages are scaled. Time on x-axis (5ms/div.)

The THD+noise was also measured at a constant input level of 1Vrms and it is shown in Figure 18.

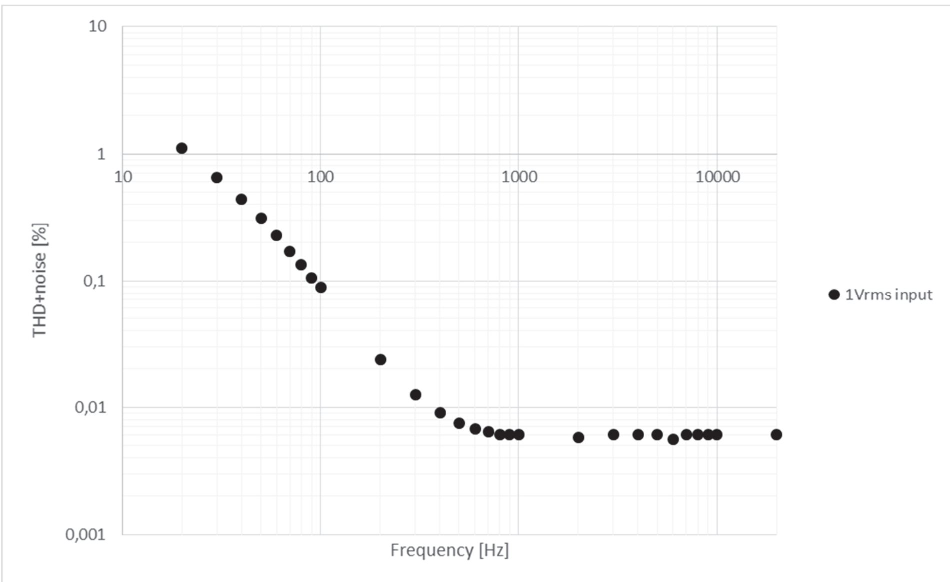


Figure 18 THD+noise at $1V_{RMS}$ input, frequency on x-axis and THD+Noise on y-axis

Both measurements show a limited headroom at lower frequencies. With Sandhill 6011A ribbon microphone the input level for 0.1%THD at 30Hz corresponds to approximately 123dB SPL and 1.0%THD is reached at around 148dB. Lundahl offers an XL-version of the input transformer with a 45% larger core and 3dB higher headroom which is pin compatible with the standard version.

6.2.2 Lundahl LL1517 Audio Output Transformer

Lundahl 1517 is a mu-metal core audio output transformer with three-section coils. According to the manufacturers datasheet (appendix 2) the transformer is ideally used with mixed feedback drive circuits. In this design the output transformer will indeed be driven by a distortion cancelling mixed feedback circuitry so datasheet will provide all necessary level handling figures. For circuit simulation purposes and op amp choice however knowing the primary inductance is useful. The transformer was measured in the suggested 1:1 configuration with primaries and secondaries wired in series. The primary inductance was measured in the same manner as the LL1538 but it turned out the inductance level dependency of the LL1517 was much more obvious showing a clear proportional relation between the drive level and primary inductance as shown in Figure 19. The transformer was measured at four different signal levels ranging from 100mV to 10Vrms in 10 dB increments with two different resistor values. Under these conditions the primary inductance varied between roughly 0.7 and 2 Henries. This figures correspond to 80-250Ω load impedance at 20Hz which is challenging to drive and will have to be taken into consideration when choosing the op amp for the line driver.

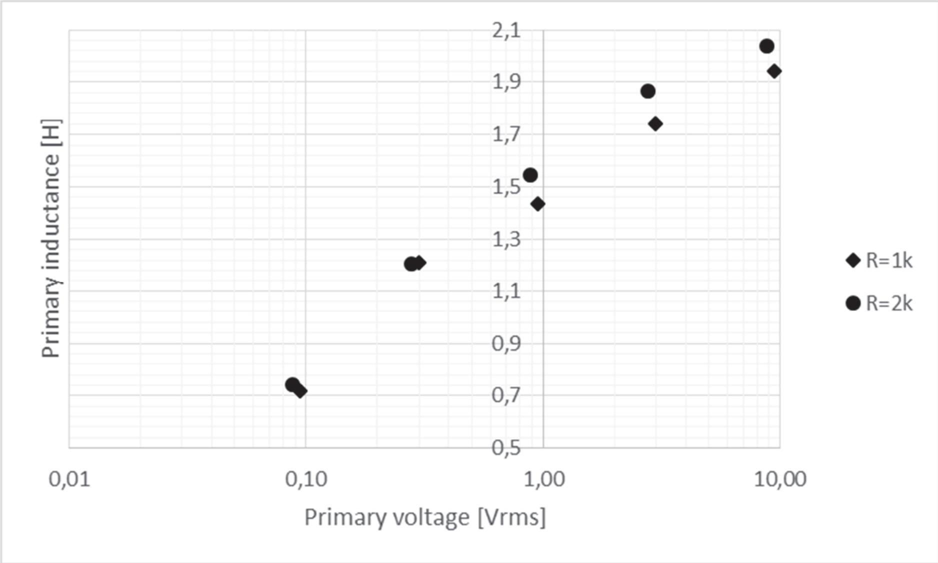


Figure 19 LL1517 primary inductance vs. level, primary voltage on x-axis and primary inductance on y-axis

The key design values according to the manufacturers datasheet:

• Static resistance of each primary	9.2Ω
• Static resistance of each secondary	9.5Ω
• Leakage inductance of secondaries (sec. in series)	0.3mH
• No-load impedance	Typically >600Ω@ 50Hz,+20dBu
• Distortion (achieved with mixed feedback drive circuit, load 600Ω)	<0.03%@24dBu, 30Hz
• Optimum source impedance	-18Ω

Measured values:

• Primary inductance	0.7-2H
• Primary resistance	17.8Ω

- Secondary resistance 18.9Ω

- Leakage inductance immeasurable

6.3 Gain Stage

6.3.1 Initial Design

The initial prototype amplifier used an Analog Devices ADA4627-1 operational amplifier IC for voltage amplification. ADA4627-1 is a high speed, low noise, low bias current JFET op amp with 19 MHz gain bandwidth product. It is a very high performance unity gain stable op amp with an exceptionally low offset voltage. A simplified schematic of the design is presented in Figure 20.

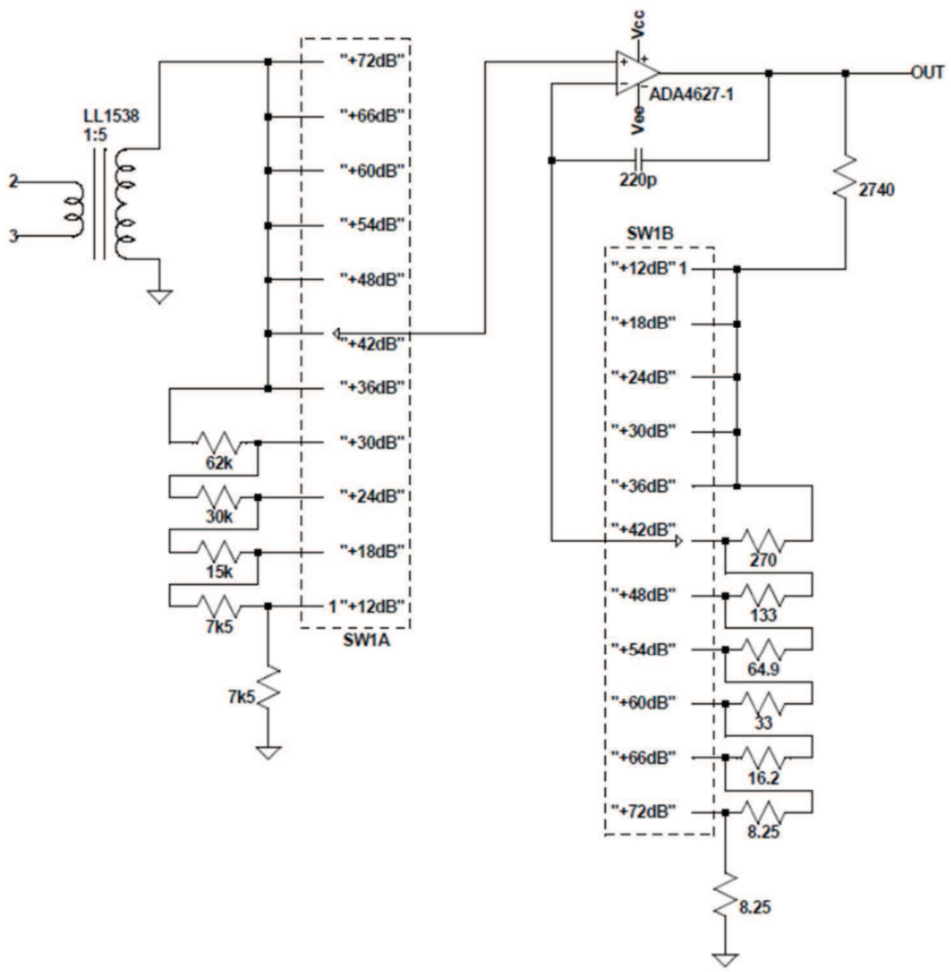


Figure 20 Prototype Gain Stage Schematic showing the gain switching of the mechanical rotary switch

Prior to the active gain stage the signal is boosted 14dB by the input transformer wired in 1:5 configuration. Between the transformer and the op amp lies a resistive attenuator circuit, also known as PAD, which protects the amplifier from overloading at the four lowest positions of the 11-position gain control. This PAD cuts the signal by 6 to 24dB in 6 decibel increments depending on the gain control position. The op amp is set up in a non-inverting configuration with a switchable feedback divider network. The gain of the op amp is set between 16 and 52 decibels. 52dB from a single op amp is not common in the world of audio devices and it pushes the ADA4627-1 to its limits in terms of bandwidth. The corner frequency at maximum gain is at roughly 48kHz and there is a 1dB attenuation showing at 20kHz. The limited bandwidth can lead to excess distortion at higher frequencies which should be avoided. One other thing of great importance that came up with this design is the premature distortion due to input capacitance nonlinearity of the IC. That will be also discussed later in this chapter.

6.3.2 Decompensated Op Amps

For the new design the IC was substituted with a decompensated version of the ADA4627-1, the ADA4637-1. Compensated op amps are traditionally designed to be stable for gains down to and including unity gain. Decompensated or less compensated op amps have less internal compensation which allows them to work only at noise gains greater than a specified minimum. The advantages compared to a compensated unity gain op amp:

- An open-loop gain which extends to a higher frequency
- A higher frequency closed-loop bandwidth
- A better slew rate

The ADA4637-1 is stable at noise gains (NG) over five and has a gain bandwidth product of 79MHz. In a non-inverting amplifier noise gain equals signal gain [7, 12]:

$$A_{V\,CL} = 1 + \frac{R_F}{R_G} = NG \qquad (3)$$

The lowest gain setting from the op amp is configured at 16dB or six. It is above the stability criteria but ever so slightly that extra care has to be taken in the circuit board design.

External compensation methods for decompensated amplifiers are discussed in Texas Instruments Application Report AN-1604 and although they are primarily intended for stabilizing decompensated op amps to run below the minimum required gain they obviously apply to bullet proofing potentially stable circuits as well.

Feedback stability theory states that the closed-loop gain must intersect the open-loop gain at a slope no greater than 6 dB/octave (single pole response) for the system to be unconditionally stable. If the response is 12 dB/octave (two pole response), the op amp will oscillate. The easiest way to think of this is that each pole adds 90° of phase shift. Two poles yields 180° phase shift, and 180° of phase shift turns negative feedback into positive feedback which means oscillations [11, 6].

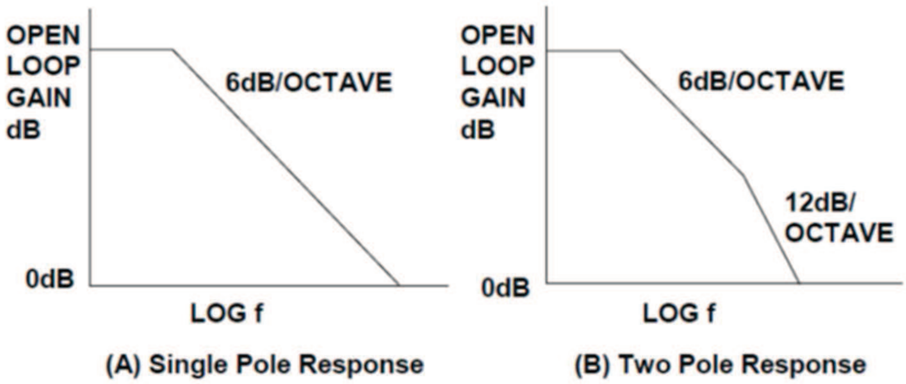


Figure 21 Two different op amp responses, the one left is a single pole response and the one on right has two poles. Reprinted from Analog devices Inc. (2009) [11, 1]

Two different responses are pictured in Figure 21. The main reasons for amplifier oscillation are discussed in the following.

6.3.3 Feedback Network Phase Lag

Parasitic capacitance between PCB traces cannot in real life be completely avoided. This capacitance together with the resistance from the feedback network will create a pole. According to Linear Technologies Application Note 148 it is recommended to size the feedback network resistor values so that the pole they create with circuit board stray capacitance is well above the loops unity gain frequency. [12, 3] The stray capacitance pictured in Figure 22 can be assumed to be in the range of 5pF. In addition the input common mode capacitance of the ADA4627-1 and ADA4637-1 op amps yields another 7pF in parallel with the stray capacitance.

The feedback network has a pole at:

$$f_{pole} = \frac{1}{2\pi * R_F || R_G * C_{PAR}} \tag{4}$$

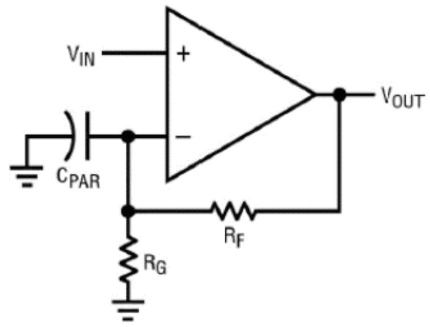
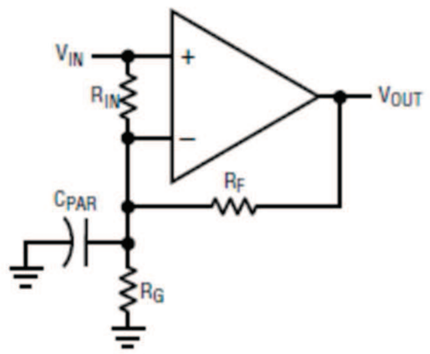


Figure 22 Parasitic capacitance. Reprinted from Harvey (2014) [12, 3]

Figure 23 shows two possible methods to make the feedback divider more tolerant to parasitic capacitance. The one on the left has a decreased input impedance due to R_{IN} and cannot be used in this design. The one on the right adds a compensation capacitor across R_F which increases noise gain at high frequencies (Figure 24). The capacitor should be sized so that the time constants of the RC-pairs are equal [12,4]:

$$C_{PAR} * R_G = C_F * R_F \tag{5}$$



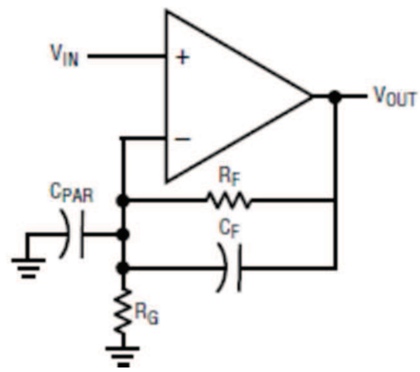


Figure 23 External compensation methods for non-inverting amplifier. Reprinted from Harvey (2014) [12, 4]

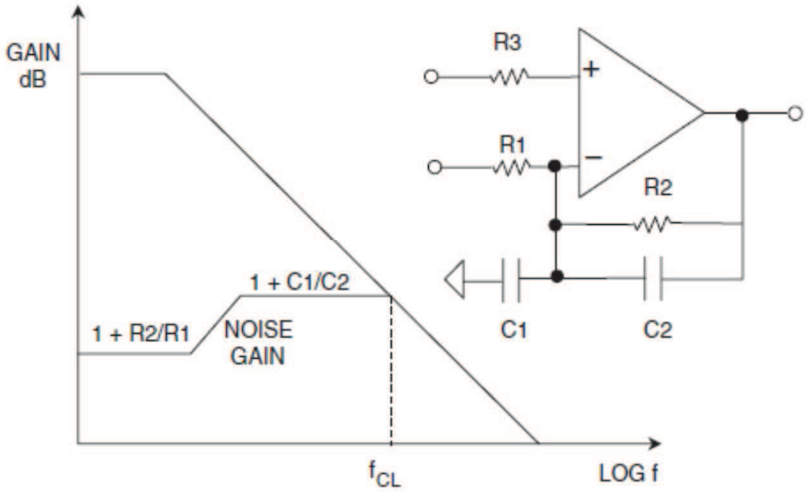


Figure 24 The effect of C_F on noise gain resulting in increased stability. Reprinted from Jung (2005) [7, 70]

6.3.4 Capacitive Loading

The output impedance of the op amp varies with frequency and closed loop gain. The flat region in Figure 25 shows the open-loop output impedance of the ADA4627-1 which is approximately 40Ω. Full feedback at gain of -1 allows the op amp to reduce the output impedance at lower gains while higher gains have a higher output impedance.

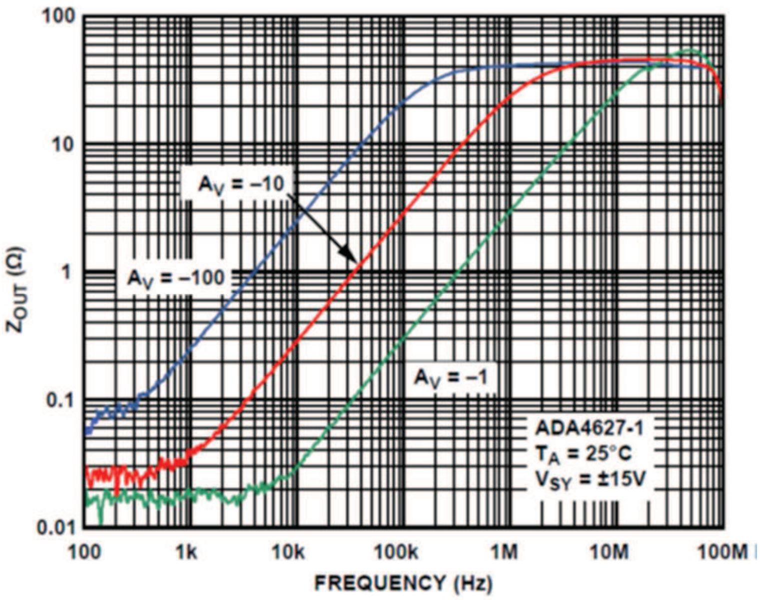


Figure 25 ADA4627-1 closed-loop Z_{out} vs. frequency at different closed loop voltage gains. Reprinted from Analog Devices Inc. (2015) [13, 6]

6.3.5 Revisited Design

The prototype had an 11-position 2-pole rotary switch for gain control that was controlling the PAD and feedback networks simultaneously. Instead of a mechanical switch the new design consists of two relay controlled four-bit 3dB/step ladder attenuators, shown in Figure 26, controlled by a microcontroller.

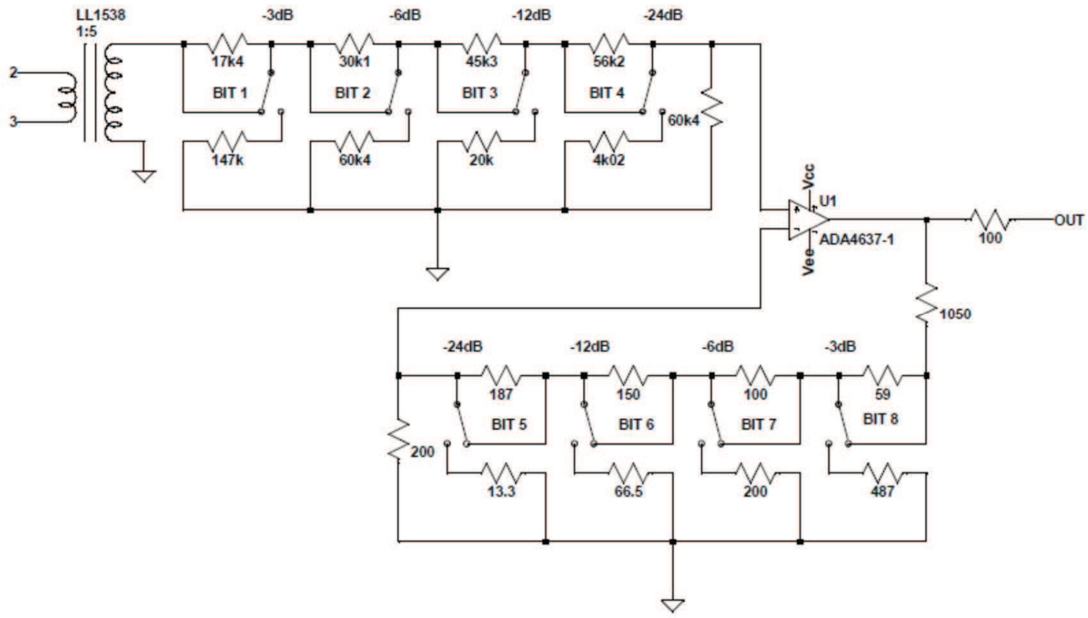


Figure 26 Ladder attenuators of the gain set circuit

The gain switch chosen for this preamp is a 23-position digital ELMA X4 hall-effect rotary switch with a 6-bit absolute code output. Both ladders are of a constant load type with E96 series resistor values. This configuration has the advantage of having hard wired feedback to the inverting input independent of relay positions, thus there will never be a situation where op amp would suddenly go to open loop again should a something unexpected happen with the relays. In addition this arrangement makes make-before-break switching unnecessary. Thanks to the extra bandwidth the gain structure can be reconfigured programmatically to different range if needed. The relay control also enables the use shortest possible physical signal routing on the PCB reducing stray capacitances and unwanted inductive signal contamination to minimum.

The load resistance of the pad is reflected to the primary side of the input transformer and is seen as the input impedance in parallel with the reactance from the primary winding inductance. The 20H primary winding has a reactance of approximately 2.5kΩ at 20Hz which limits the size of the load resistance in order to keep the input impedance of the device relatively constant throughout the audio range. Modern microphones having output impedance between 100Ω and 300Ω a load of 1.5kΩ, or five times the microphone impedance is sufficient. However, a larger input impedance rarely has negative effects. The ladder resistance will also have effect on the distortion behaviour of the amplifier, which will be discussed in chapter 6.3.5.2. The PAD ladder resistance was chosen to be 60kΩ, which results in a nominal impedance of 2,4kΩ decreasing to 1.2kΩ at 20Hz.With a 150Ω microphone the signal will be attenuated 0.5dB at 20Hz which is tolerable. The negligible effect of phantom feed resistors is omitted in these calculations.

6.3.5.1 External Compensation

The circuit was built on a breadboard and tested with a square wave input into a demanding load of 470Ω. The gain was set at 6 or 16dB. Testing circuits on a breadboard as shown in Figure 27 effectively mimics the worst case layout scenario revealing any tendencies towards instability.

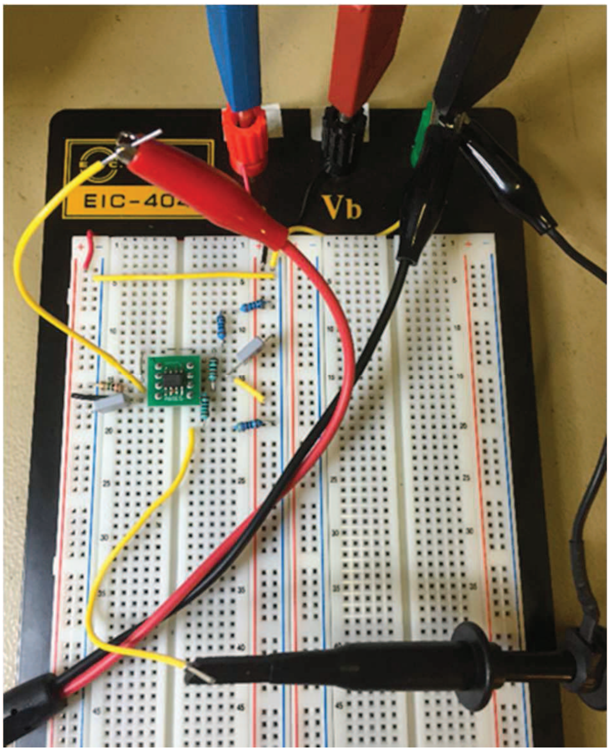


Figure 27 Test circuit on Breadboard

If we assume the total feedback network capacitance to be 12pF, equation 4 states that the network will have a pole at 79MHz.
According to equation 3 the feedback compensation capacitor should be:

$$C_F = \frac{C_{PAR}R_G}{R_F} = \frac{12pF*200\Omega}{1050\Omega} = 2.3pF \qquad (6)$$

This is less than the parasitic capacitance of the PCB traces so placing an actual component would be pointless. The circuit was showing some slight overshoot with 1X oscilloscope probe connected to the output, but placing a 150Ω resistor between the output and the load made the amplifier stable with a perfect square wave output at 100kHz. Adding a 47pF feedback capacitor actually made the op amp oscillate at a very high frequency. Both waveforms are shown in Figure 28.

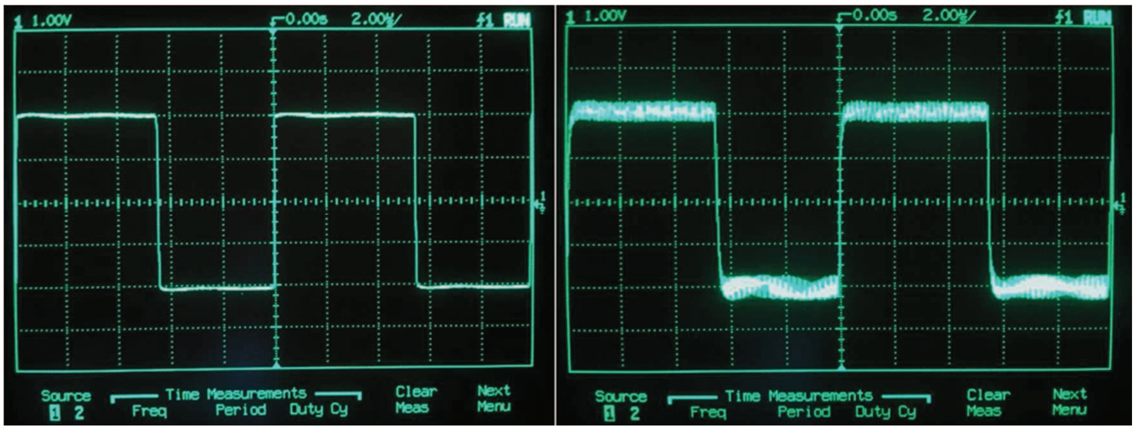


Figure 28 100kHz square wave, with and without output isolation

The gain stage in this design will see a relatively large impedance load. Therefore it is not likely it will ever oscillate. Nevertheless, a 100Ω resistor will be used to isolate the op amp from the following circuit anyway just for extra safety.

6.3.5.2 Source Impedance

Generally speaking a microphone preamplifier should have a bridging input. This means that the impedance reflected to its input, seen as a load by the microphone, should be much greater than the output impedance of the microphone itself. Active microphones are reasonably tolerant to loading due to their low output impedance but passive ribbon and moving coil microphones usually benefit from large loads. The input impedance in this case is defined by the PAD circuit load resistance which is reflected to the primary side as:

$$R_{IN} = R_{PAD} * \left(\frac{N_P}{N_S}\right)^2 = \frac{R_{PAD}}{{N_S}^2} = \frac{120k\Omega}{5^2} = 4.8k\Omega \tag{7}$$

Op amps have input capacitance to the substrate that is not linear with voltage [14, 1]. This is particularly significant in FET input op amps and can cause measurable distortion if the source impedance is too high. If the capacitance changes by 1pF, its reactance at 10kHz will change by:

$$\Delta X_C = \frac{1}{2\pi f \Delta C} = \frac{1}{2\pi * 10kHz * 1pF} \approx 15.9M\Omega \tag{8}$$

With 10kΩ series source impedance this will result in attenuation of:

$$A_V = 1 - \frac{15.9M\Omega}{15.9M\Omega + 10k\Omega} = 0.00063 \; or \; 0.063\% \tag{9}$$

This is enough to cause a measurable amount distortion. According to Walt Jung’s Op Amp Applications Handbook the source impedance should ideally equal to the parallel equivalent of resistors R_F and R_G for smallest possible distortion [7, 469] The achieved impedance symmetry effectively cancels out even order harmonic components.

The linearity of input capacitance in ADA4637-1 was measured with different source impedances ranging from 1kΩ to 1MΩ. Gain was set to approximately 6 or 16dB and output level was $5V_{RMS}$ with ±15V supply voltage (Figure 29). Figure 30 shows the growth of harmonic distortion at test frequencies 1kHz and 10kHz. In order to avoid noise related error THD was measured with Keithley 2015 THD multimeter in THD mode.

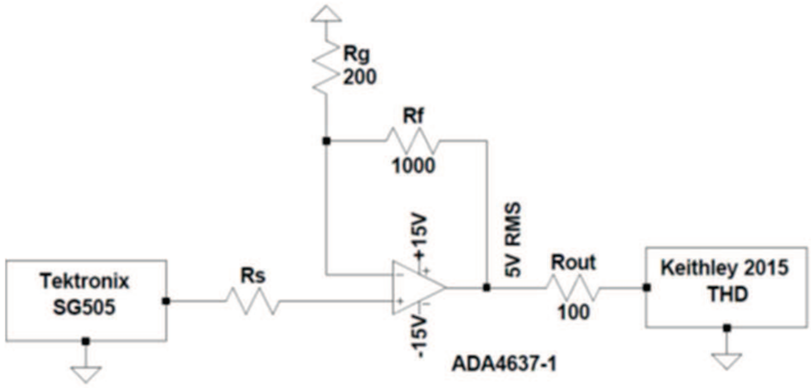


Figure 29 Test setup for input nonlinearity, R_s 1kΩ…1MΩ

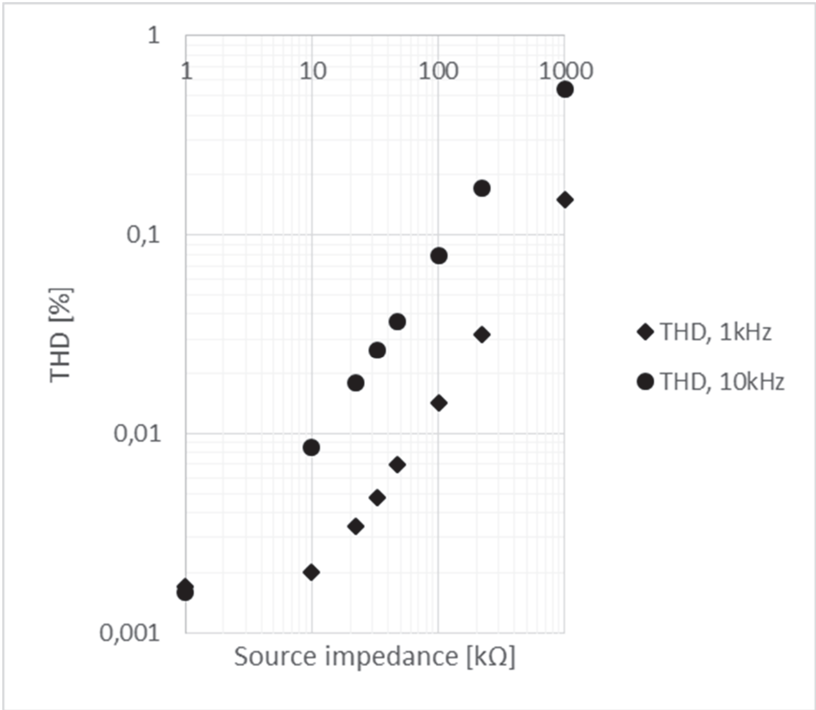


Figure 30 THD vs source impedance at 1kHz and 10kHz

The design of the PAD is a give and take between input impedance and distortion. With a 60kΩ four step ladder attenuator the source impedance ranges between 3.5kΩ and18kΩ if the microphone has an input impedance of 150Ω. Looking at the distortion figures in Figure 30 source impedances above 10kΩ are risky.

Direct Input

The prototype preamplifier had no DI-input for instruments and the only microphone input connection was through the card edge connector. The new design features a Neutrik Combo series 6.3mm phone jack/XLR-connector placed in the front panel for easier access.

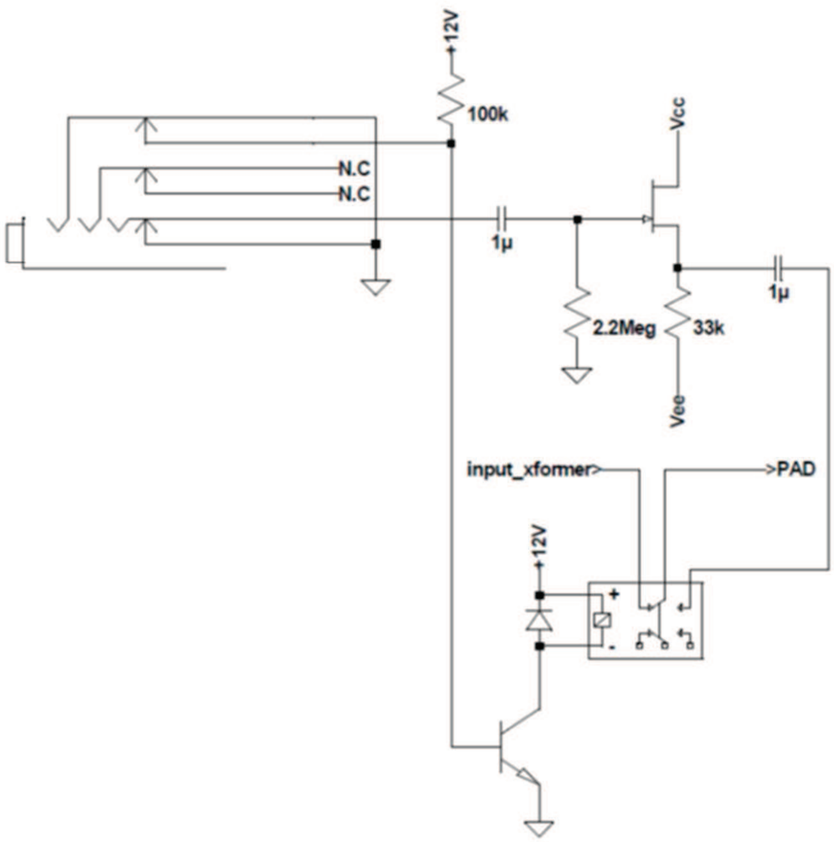


Figure 31 DI schematic

Plugging a 6.3mm mono plug to this connector controls a relay which selects between mic signal and a buffered instrument signal. The minimum recommended load for a DI input is 1MΩ and in this case 2.2MΩ was used. The buffer in Figure 31 consists of a single resistor loaded common drain N-channel JFET which is AC-coupled with capacitors.

6.3.5.3 Level Handling

The bare op amp circuit was measured for THD+noise at 1kHz with 16dB gain. The amplifier was able to put out 8.8V_{RMS} or 21.1dBu at 0.1% THD+noise level into a 200kΩ load. Clipping occurred rapidly above this level as one would expect from an amplifier with a large amount of feedback.

The output was also loaded with a pull-down constant current source to the negative rail to see if forcing one of the output devices into class-A had any effect to the performance. With a 4mA pull-down current the measured distortion figures didn't change but the clipping waveform became slightly asymmetrical with one side having rounded edges.

At maximum gain the upper corner frequency was measured at 137kHz and the output DC offset voltage was 8mV. 10kHz square wave in Figure 32 showed some slight ringing but apart from than that it was surprisingly clean. The output voltage for 0.1% THD +noise was 8.52V_{RMS} or 20.8dBu at 1kHz and 7.99V_{RMS} or 20.3dBu at 20kHz

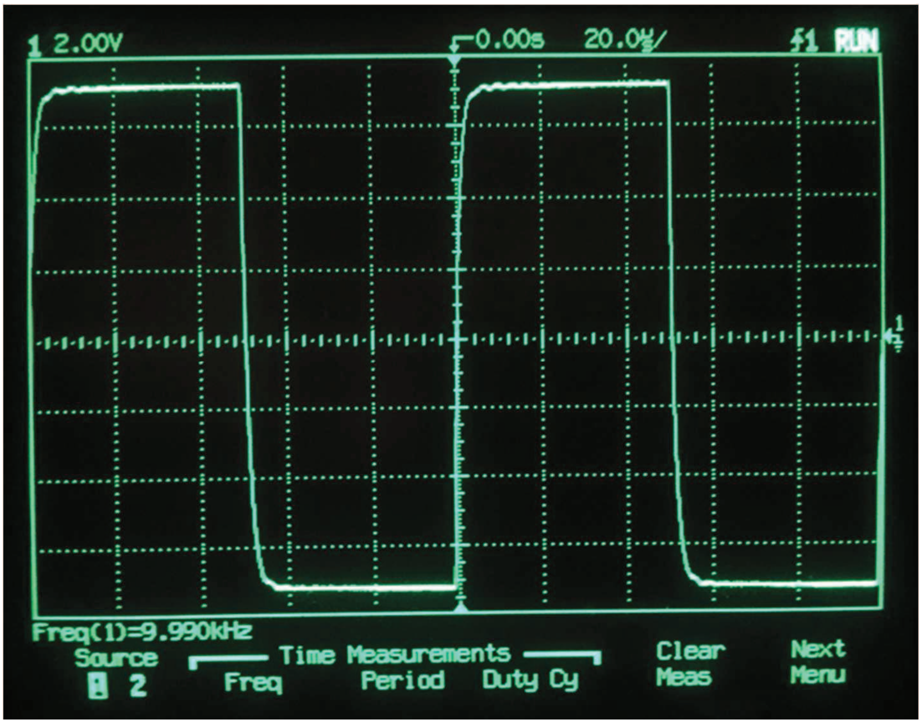


Figure 32 10kHz square wave at full gain, time on x-axis (20µS/div.) and voltage on y-axis (2V/div.)

Slewing induced distortion (SID) measurements for this circuit are too large a topic to cover in this thesis work but it would be a very interesting area for further study [15].

6.3.6 Noise

The total noise level of an op amp circuit is calculated piece by piece using superposition. Each source is isolated assuming everything else to be noiseless. The calculated noises are then added together as independent sources. The equivalent noise sources for inverting and non-inverting amplifiers are represented in Figure 33.

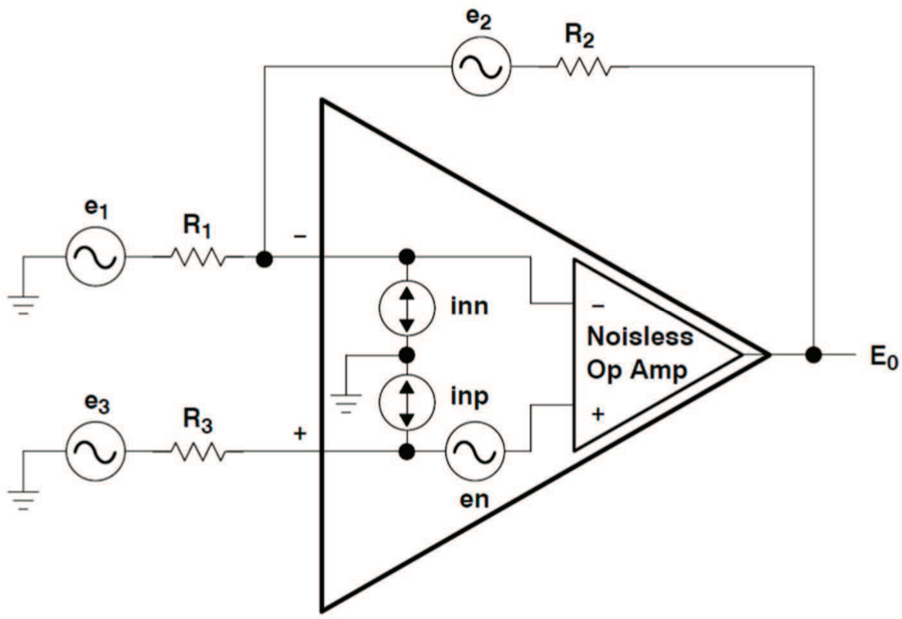


Figure 33 Equivalent noise sources in an op amp. Reprinted from Texas instruments Inc. (2007) [19, 11]

Thermal noise of a resistor is given by equation:

$$e^2 = \int 4kTR\,df\,,\ where\ k = 1.38 * 10^{-23}\frac{J}{K} \quad (10)$$

For a bandwidth BW between frequencies f_L to f_H this simplifies to

$$e = \sqrt{4kTR(f_H - f_L)} \quad (11)$$

Thermal noise in the op amp output from resistors in $R_1 \ldots R_3$ Figure 33:

$$E_1 = e_1\frac{R_2}{R_1} \quad (12)$$

$$E_2 = e_2 \quad (13)$$

$$E_3 = e_3\frac{R_1 R_2}{R_1} \quad (14)$$

The noise voltages of the gain stage at maximum gain:

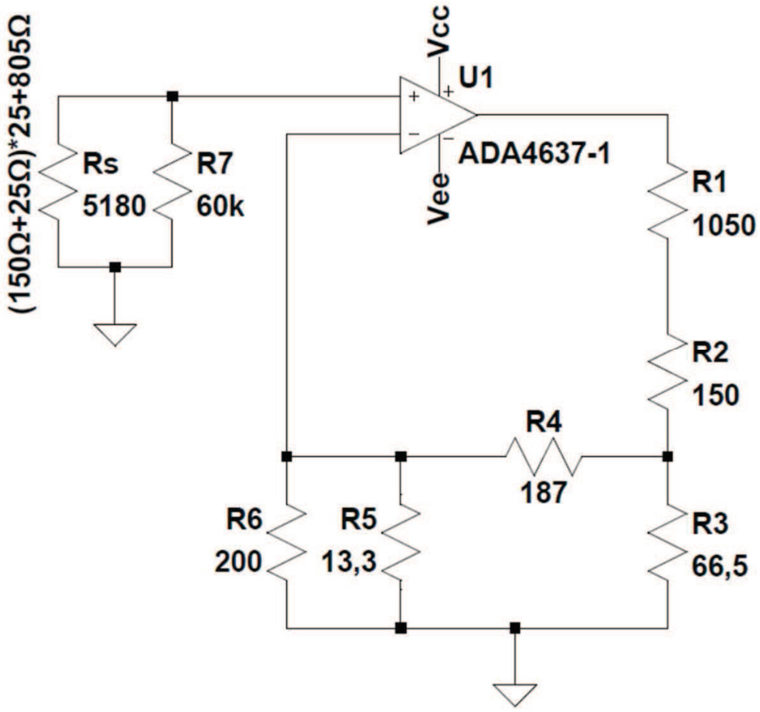


Figure 34 Gain stage resistances at full gain

Due to the ladder attenuation in the feedback loop, the circuit differs slightly from the basic amplifier presented in Figure 33. The resistor network at full gain is shown in Figure 34. Source resistance R_s consists of the reflected 150Ω microphone impedance used for calculating the equivalent input noise or E.I.N. in series with the primary resistance and the secondary resistance of the transformer. In order to calculate the resistor induced noise the ladder network resistors are substituted with tree resistors R_1, R_2 and R_3 in Figure 35 creating equal amount of noise. The equivalent resistors are denoted with * in the following equations. When calculating E_1 and E_3 the generator noise is multiplied by the noise gain of the circuit which for both non-inverting and inverting inputs is 52dB or 398.

$R_1^* = R_5||R_6 = 12.5\Omega$

$R_2^* = ((R_1 + R_2)||R_3) + R_4 = 250\Omega$

$R_3^* = R_S||R_7 = 4768\Omega$

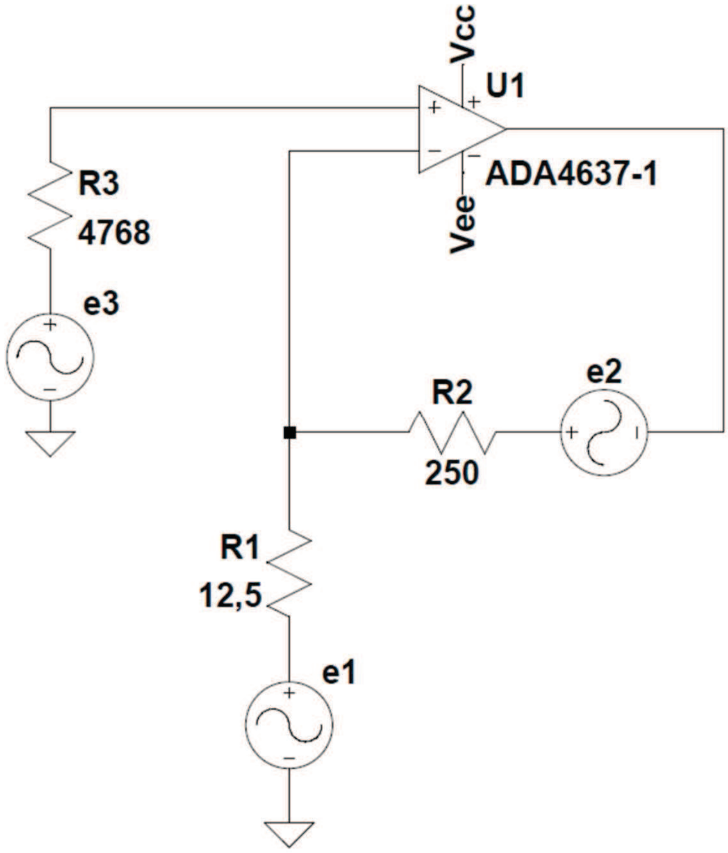


Figure 35 Equivalent resistances and noise sourced used for thermal noise calculations

$e_1 = 64.5nV$

$E_1 = 64.5nV * 398 = 25.7uV$

$e_2 = 289nV$

$E_2 = 289nV$

$e_3 = 1.26\mu V$

$E_3 = 1.26\mu V * 398 = 501.5\mu V$

Op amp voltage noise:

The op amp RMS voltage noise is calculated with equation [19,11]:

$$E_n = e_n * \sqrt{f_{nc} * \ln\left(\frac{f_H}{f_L}\right) + (f_H - f_L)} \tag{15}$$

where f_{nc} is the 1/f voltage noise corner frequency and f_H and f_L are the upper and lower limits of the chosen bandwidth. The voltage noise corner frequency can be sourced from the datasheet graph as shown in Figure 36. It is the point where the asymptotes of 1/f-noise and white noise regions cross. According to the graph for ADA4637-1 this is at about 110Hz, which is typical for a JFET input op amp.

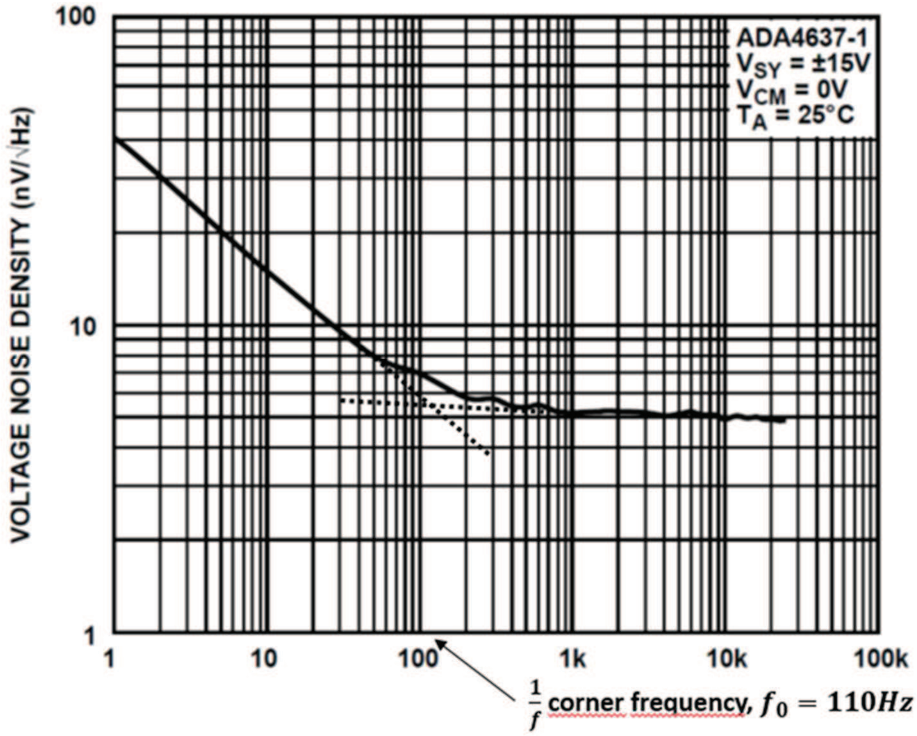


Figure 36 ADA4637-1 Voltage noise density vs frequency. Asymtotes drawn in dashed lines to indicate 1/f corner frequency. Modified from Analog Devices Inc. (2015) [13, 13]

An alternative way of estimating the noise corner frequency f_{nc} from datasheet figures is described in Texas Instruments Application report Noise Analysis in Operational Amplifier circuits [19, 7]. The corner frequency is derived from a device constant K

$$K^2 = \left[\left(\frac{e_{\frac{1}{f}}}{\sqrt{Hz}}\right)^2 - \left(\frac{e_{white}}{\sqrt{Hz}}\right)^2\right] * f_{LOW} \quad (16)$$

$$K^2 = \left[\left(\frac{16{,}5}{\sqrt{Hz}}\right)^2 - \left(\frac{6{,}1nV}{\sqrt{Hz}}\right)^2\right] * 10Hz = 2350(nV)^2$$

In this equation $e_{1/f}$ is the equivalent noise voltage density at the lowest possible frequency in the 1/f noise range given by the manufacturer, f_{LOW} is the corresponding frequency and e_{white} is the white noise voltage density. The corner frequency f_{nc} is derived from this value by dividing with white noise density squared.

$$f_{nc} = \frac{K^2}{(e_{white})^2} \tag{17}$$

$$f_{nc} = \frac{2350(nV)^2}{\left(\frac{6.1nV}{\sqrt{Hz}}\right)^2} = 63Hz$$

The op amp RMS voltage noise can now be calculated [20,2]:

$$e_n = e_n * \sqrt{f_{nc} * \ln\left(\frac{f_H}{f_L}\right) + (f_H - f_L)} \tag{18}$$

$$e_n = 6.1\frac{nV}{\sqrt{Hz}} * \sqrt{110Hz * \ln\left(\frac{20kHz}{20Hz}\right) + (20kHz - 20Hz)} = 0.883\mu V$$

The voltage at the output is the voltage noise multiplied by the noise gain:

$$E_{n\ output} = 0.883\mu V * 398 = 351\mu V$$

Op amp current noise:

The ADA4637-1 current noise density is not specified thoroughly in the data sheet. Only two values are given:

- Current noise density $1.6\frac{fA}{\sqrt{Hz}}, f = 100Hz$
- Current noise 30fA p-p, f=0.1Hz to 10Hz

These values do not provide enough information for audio range current noise analysis, but due to the very small current noise density value the current noise can quite safely be omitted without a significant error.

Total noise:

The total RMS noise in the gain stage output can be calculated with equation [19, 5]:

$$E_{TOT\ RMS} = \sqrt{{E_1}^2 + {E_2}^2 + {E_3}^2 + {E_n}^2} \tag{19}$$

$$E_{TOT\ RMS} = \sqrt{(25.7\mu V)^2 + (0.289\mu V)^2 + (501.5\mu V)^2 + (0.351\mu V)^2} = 613\mu V$$

Equivalent input noise or E.I.N. can be calculated by dividing the output noise with the voltage gain. In this case it is the sum of the transformer and op amp gains, hence 66dB or 1990.

$$E.I.N. = \frac{502\mu V}{1990} = 307nV \qquad (20)$$

In dBu this refers to:

$$E.I.N = 20 * \log\left(\frac{307nV}{0.775V}\right) = -128dBU \qquad (21)$$

The theoretical limit is set by the thermal noise from 150Ω resistor which results in:

$$E.I.N._{theor.} = 20 * log\left(\frac{\sqrt{4*kT*150\Omega*19980Hz}}{0.775}\right) = -130.8dBU \qquad (22)$$

This means that the noise figure of the gain stage is 2.8dB. The following stages will contribute a few microvolts to the output noise but its relevance to the overall noise performance of the preamplifier is negligible.

6.4 Equalizer

The preamplifier has an equalizer with two functions, high pass filter and a shelving filter. The prototype had active filters but the design was simplified in order to minimize the amount of active components on the signal path. The task of the low pass filter is to attenuate unwanted low frequency content from the audio and the corner frequency of the first order RC-filter is set to 85Hz. The shelving filter will boost the signal by 0.7dB at 10kHz and 1.8dB at 20kHz compensating the attenuated high frequency response typical to all bidirectional ribbon microphones. The circuit of the filters is shown in Figure 37. The equalizer has a nominal attenuation of 3.17dB which will be compensated in the line driver stage. The responses of both filters were fine-tuned on LTSpice simulator software to emulate the target curves sourced from listening tests. The frequency and phase responses with both filters switched on are shown in Figure 38. Listening tests were performed using digital software equalizers in a recording studio environment on audio material recorded with a Sandhill 6011A ribbon microphone.

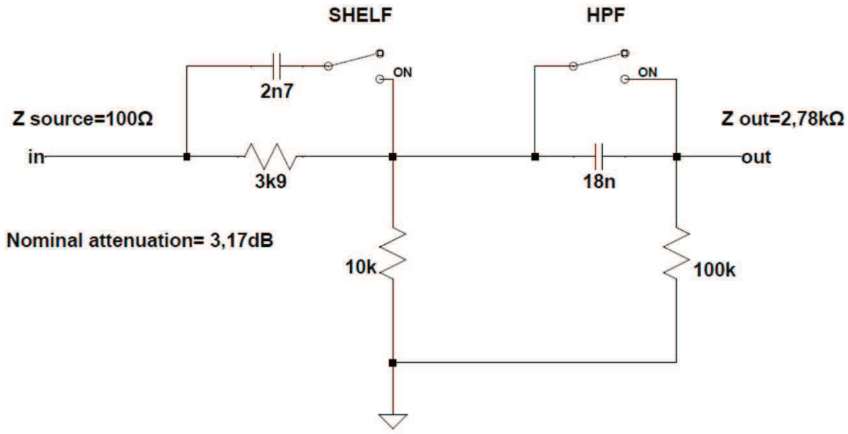


Figure 37 Passive equalizer circuit

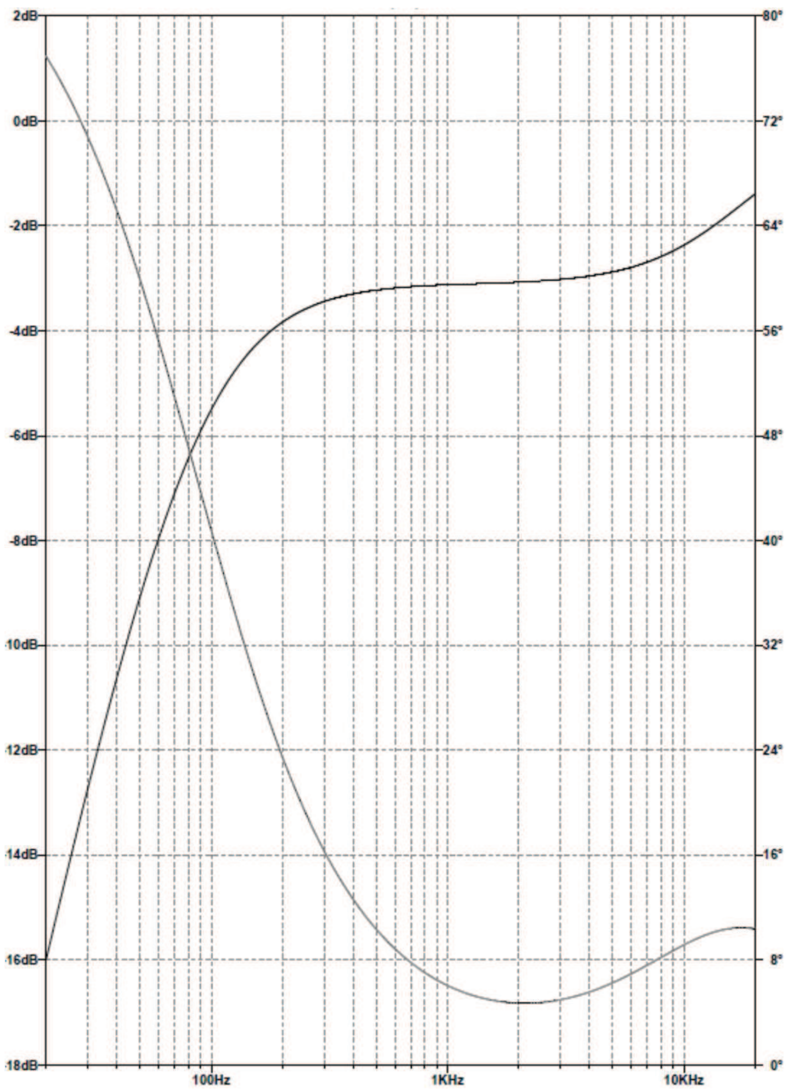


Figure 38 Simulated equalizer frequency and phase responses. Both filters are active in this simulation

6.5 Line Driver

The output driver of a microphone preamplifier will have to be able to deliver a large voltage swing into a load that can in some cases have a large reactive component. The load often consists of a cable with an unknown length and capacitance. A balanced audio cable typically has a capacitance in range of 10-20 pF/meter from one conductor to the other and five times or more than that from both conductors to the shield. This means that a 100 meter cable run will add 5-10nF capacitance in parallel with the load. Therefore the output driver has to have a reasonable current capability and its stability cannot be compromised by capacitive loads.

Lundahl Transformers AB recommends that for optimal distortion performance their output transformers should be driven by a circuit using a mixed feedback topology. Therefore a balanced mixed feedback driver was chosen for the preamplifier prototype. The circuit worked reasonably well but some improvements were still required for reliable operation.

6.5.1 Mixed Feedback Topology

The line output driver utilizes a circuit presented by Walt Jung in his book Op Amp Applications Handbook. This is a variation of a mixed feedback design first introduced by Werner Baudisch in a German patent DE2901567, issued July 24, 1980. This circuit uses positive feedback to compensate for low frequency distortion created by transformer core nonlinearities. The primary inductance, L_h in Figure 39 and copper resistance R_{cu} of the output transformer form a series RL circuit and if either one, in this case the inductance is nonlinear, distortion will occur. R' is a current sampling resistor which will create positive feedback voltage equal to the voltage loss created by the copper loss the primary winding, hence the circuit will have a negative output impedance.
In this arrangement the following should hold true:

$$R_n = R_e \ and \ R' = R_{cu} \qquad (23)$$

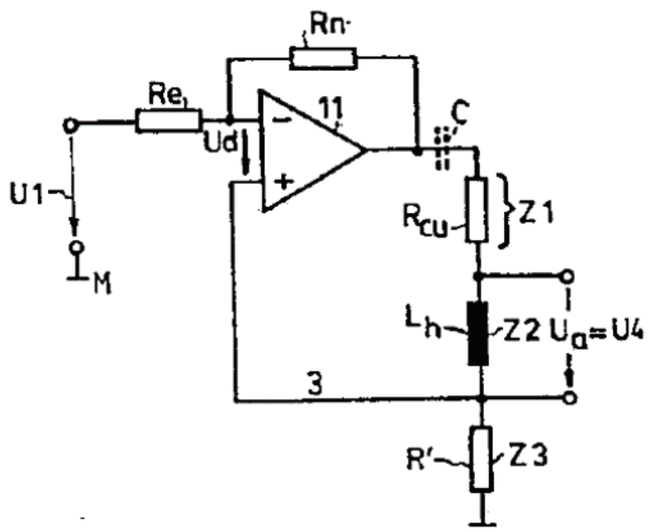


Figure 39 Mixed feedback circuit from Werner Baudisch's patent. Reprinted from Baudisch (1980) [16]

The improved circuit presented by Walt Jung has some major advantages compared to the original mixed feedback driver design. The circuit, which is shown in Figure 40 has a balanced drive consisting of two inverting op amps. The gain of the driver can be set by the voltage divider R_1 and R_2 and the ratio does not affect the distortion compensation. Due to the balanced drive, 6dB more headroom is available from the same supply voltages than would be possible with a single-ended circuit and the arrangement also has a gain of 6dB if resistor values R_1 and R_2 are equal. Similar to the circuit in Figure 39 the following ratios should apply for maximum distortion cancelling operation:

$$R_3 = R_{primary} \ and \ R_4 = R_6 \tag{24}$$

The distortion null can be fine-tuned with a trimmer R_5, however some error sources are obvious. The thermal coefficients of the resistor R_3 and the primary resistance will differ, thus the ambient temperature will change the ratio to some degree causing less than perfect distortion performance. The trimmer R_5 position will also affect the gain of the driver stage which potentially can cause variation between units. This could be compensated by placing another 500Ω trimmer between R_1 and R_2 for gain fine-tuning.
The capacitor C_3 is placed in the loop to prevent DC-latch up should the positive feedback override negative. According to Jung the capacitor should be of a polymer type.

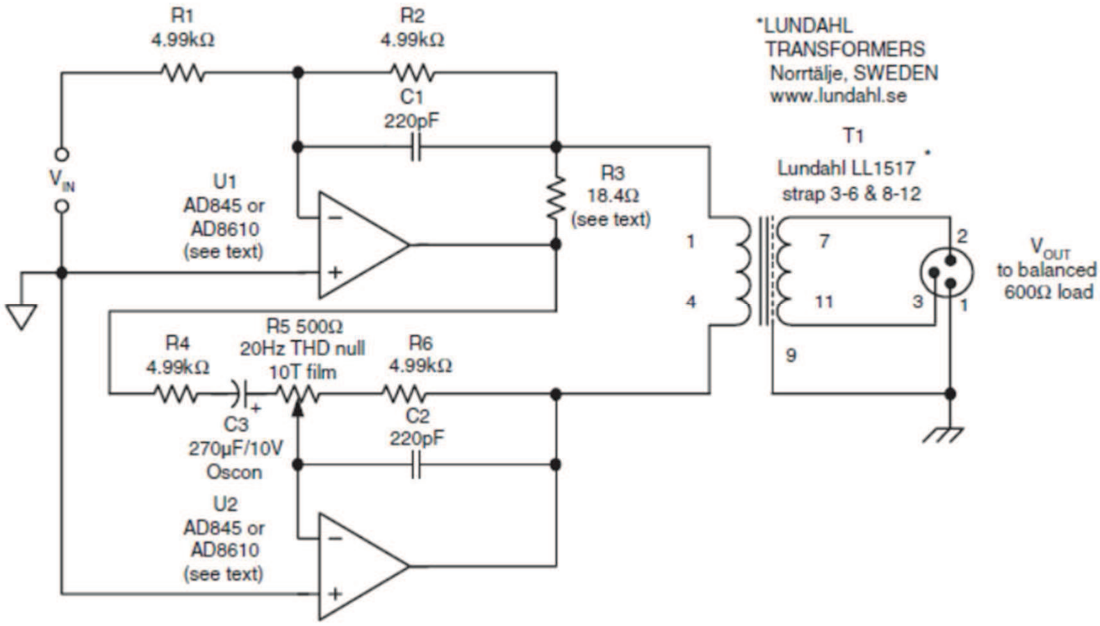


Figure 40 Balanced mixed feedback circuit by Walt Jung. Reprinted from Jung (2005) [7, 489]

The distortion figures reported by Jung are as such excellent but in actual tests driving low frequency signals into a 600Ω load resulted in severe temperature rise within the op amps. A $5V_{RMS}$ output at 20Hz caused enough heat to evaporate isopropanol (boiling point 82.3 °C) instantly from the surface of a SOIC-8 package of the op amp. In normal use such continuous signals are not likely to ever be present, but in a test situation this condition could result in a failure. Therefore an appropriate means to increase the current drive capacity should be used.

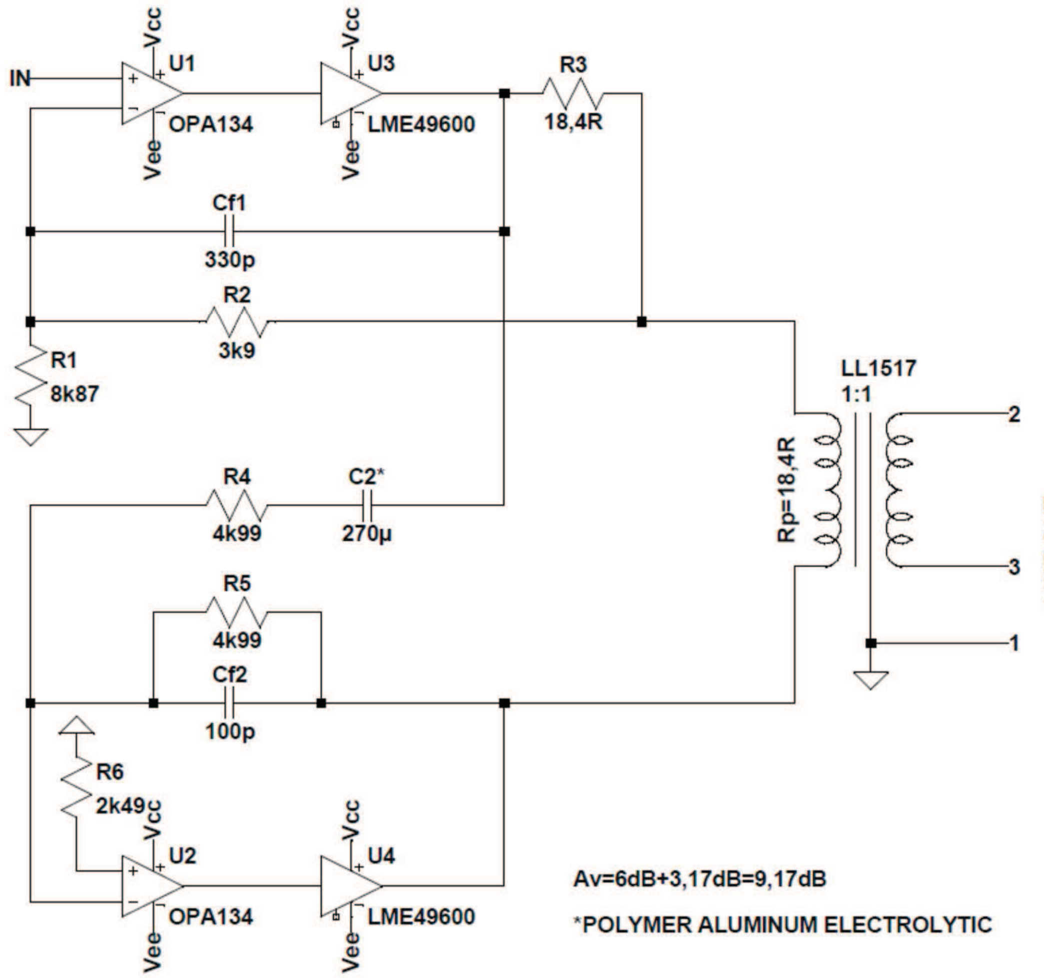


Figure 41 Line driver circuit with current buffers

Since the revisited equalizer circuit requires a large load impedance, it is necessary to convert the first op amp, U_1 in Figure 41, into a non-inverting configuration as suggested by A. Offenberg [18].The gain of the stage is set to 3.17dB to compensate for the equalizer attenuation. The feedback resistors are sized in such way that the impedances seen by the op amp inputs are equal. It is noted in OPA134 datasheet that should one of those impedances exceed 2kΩ the impedances should be matched for lowest distortion. The nominal impedance of the equalizer is 2.78kΩ and resistors R_1 and R_2 in parallel result in 2.72kΩ. Also the inverting input of the second op amp was tied to ground with a resistor half the value of the feedback resistors.

The distortion null can be adjusted by substituting the current sampling resistor R_3 with a trimmer potentiometer. However in this design it was omitted because the manufacturing tolerances of Lundahl transformers are known to be tight. It should be noted that despite having a DC-blocking capacitor C_2 in the circuit, a resistor of larger value than the primary resistance resulted in pulsing low frequency DC-latch up with $\tau = C_2 * R_4 = 1.3s$. This happened due to excessive positive feedback.

6.5.2 Amplifier Choice

The circuit presented by Jung has a major drawback which by looking at the primary inductance measurements of the LL1517 in chapter 6.2.2 becomes quite obvious. At low frequencies the primary impedance of the output transformer is determined by the (level dependent) primary inductance and the total impedance of the primary paralleled with a 600Ω reflected secondary load varies roughly between 75 and 180 ohms at 20Hz. This means that for a full 24dBu output the corresponding primary current range is 69 to 160mA. No regular IC op amp is able to source or sink such current and as mentioned in the previous chapter the thermal dissipation limits of the op amps may be unintentionally exceeded resulting in a failure of the devices. Therefore a pair of IC buffers were added to the circuit as suggested by Jung. However Jung's arguments are solely based on distortion performance rather than thermal issues.

The op amps recommended by Jung are both Analog Devices JFET input units. The ±15V supply voltage is too high for AD8610, but AD845 could be used in this circuit. However, since the op amp only has to drive the input of the buffer IC instead of the transformer, the op amp choice is not as crucial as in Jung's circuit and a more generic and cost effective OPA134 can be used. The current drive IC chosen for the task is a Texas Instruments LME49600 250mA headphone buffer. It is cased in a TO-263 package which has a cooling tab that when soldered to a copper plane on a PCB will decrease the thermal impedance of the system thus decreasing the junction temperature. On the contrary to op amp ICs the LME49600 also has a built in thermal shutdown mechanism which further protects the output from misuse.

6.5.3 Performance

The line driver circuit was tested for distortion on a breadboard. The measurements were carried on an Audio Presicion Portable One audio analyser. The circuit was loaded with 600Ω, which in a modern recording environment is very uncommon as most devices come with much higher input impedances.

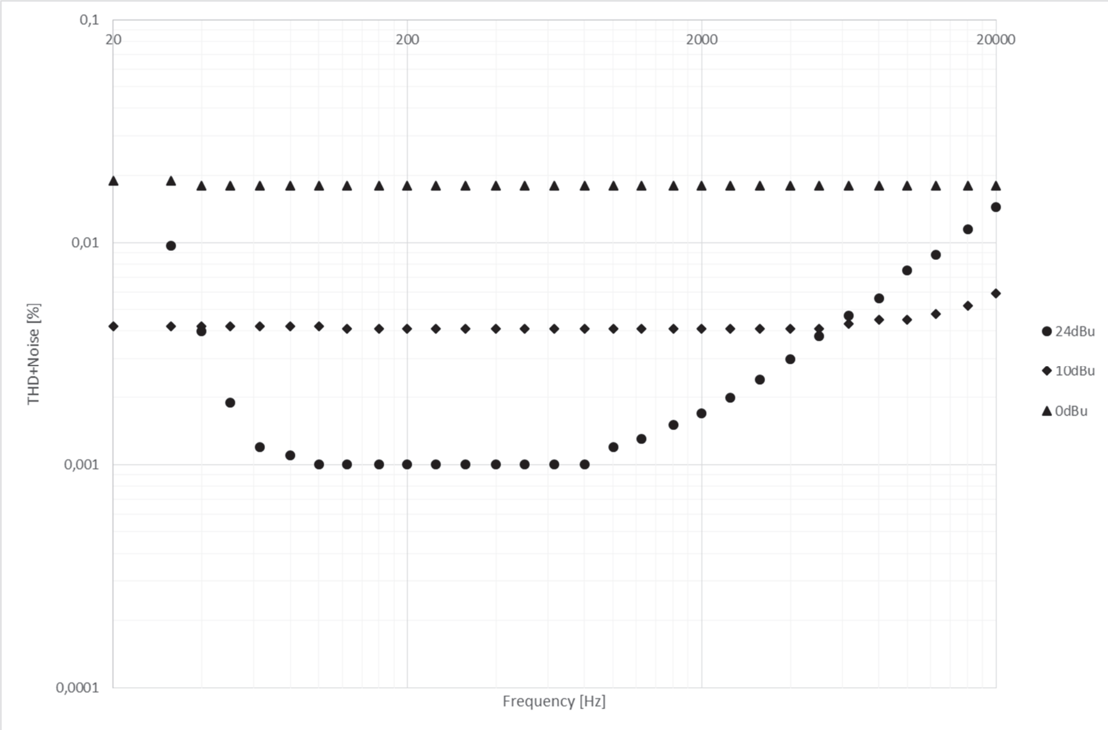


Figure 42 Line driver THD+noise into a 600Ω load at three output levels.

At 20Hz heavy crossover distortion occurred at 22.5dBu (10.3VRMS). At frequencies higher than that the driver was able to deliver 24dBu signal level into a 600Ω load as shown in Figure 42.The 20Hz distortion value at 24dBu is omitted in Figure 42. The rising distortion in the upper range at 24dBu level is most likely due to JFET input capacitance nonlinearity discussed earlier in this thesis work. In this test it became clear that cooling of the output buffers has to be well thought in the PCB layout design as the devices ran really hot on breadboard. At 10dBu output level the frequency response was flat (±0.1dB) from 20 Hz to 20kHz. The upper -3dB corner frequency was at 80kHz.

6.6 Level Indicator

The design goal was to replace an existing two led signal and peak detector with one that would use only one bi-color led to indicate the signal level. This simple indicator serves two purposes. It gives an indication that a signal is present and warns the user of signals that can overdrive the circuit.

Since the signal may have either positive or negative transients a two-sided indicator is useful for at least the peak detector. A two-sided detector was designed for signal indication as well because the quad OPA4134 IC has an even number of gates. The detectors shown in Figure 43 are op amp comparators with resistor divider set threshold voltages. The dividers are ground referenced to minimize offset error from supply voltage variation. The choice of OPA4134 is overkill for the application, but it is chosen to keep the parts count as low as possible for pick-and-place machine setup. It is essential to use a JFET input op amp in this position to keep bias currents from affecting the threshold voltage divider operation. In the signal detector the op amp has to have a low input offset voltage compared to the threshold voltage which is about 38mV. The OPAx134 family of amps has a maximum input offset voltage of ±2.0mV [17] which is sufficient resulting in an error of less than 0.5dB.

The comparators are followed by inverting Schmitt triggers which adjust the release time of the indicator. The release time, hence the time the trigger output will stay in a low state after the stimulus disappears is set for 227mS with resistors R_5, R_6, R_{12} and R_{23}. Negative output voltages from the Schmitt triggers (U_3 and U_6 in Figure 43) to the single supply logic ICs are disabled with output diodes (D_3 D_6).

The logic operations are performed with a CD4011B CMOS quad NAND-gate logic IC followed by op amp voltage followers. The logic circuit can only source about 3.6mA and sink even less current, so using voltage followers (U_{11} and U_{12}) is necessary to drive the LEDs. R_{16} limits the LED current to about 12mA.

Logic operations:

output level	**detector output**	**signal**
< -20dBu	peak low, signal low	led off
≥ -20dBu	peak low, signal low	led green
≥ 24dBu	peak high, signal high	led red

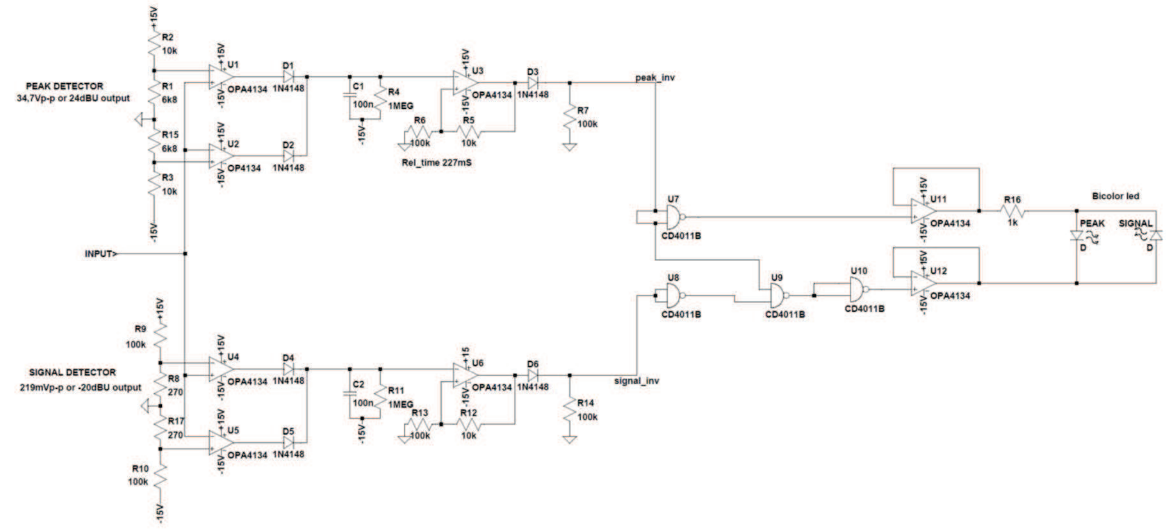


Figure 43 Level indicator schematic

The 24dBu peak indication offers approximately 2dB headroom before the output waveform clips. However, the peak indicator is not able to fully predict transformer saturation at lowest frequencies. Especially the input transformer the transformer will saturate earlier than other parts of the circuit when driven hard with a low frequency signal. However, distortion caused by transformer saturation does not have a hard clipped waveform and consists of low order harmonics.

7 Conclusions

The result of this thesis work is a versatile high performance analog circuit for a microphone preamplifier, which can be implemented in both an API 500-series module as well as in a standalone device with a built-in power supply. Implementing the amplifier in a standalone unit would make it possible to use slightly higher supply voltages, hence the performance would be further improved.

The design was studied in great detail and most of the pitfalls were avoided. Finding somewhat nontrivial distortion sources such as the nonlinearity of the JFET op amp input capacitance and the level dependent inductance of the output transformer brought a great satisfaction. It can also be said that most of the design goals were fully met, although the heat issue in the line driver will require great care later in the PCB design phase.

The design process was performed almost solely from a technical perspective with a money no object philosophy. However in a design of a commercial product the manufacturing costs are always a matter of great importance. Despite the fact that the performance of the amplifier is pristine, a high retail price can make it impossible to commercialize the product. It is left for further study to find means of saving costs by simplifying the design. One possible action would include replacing the output transformer with a capacitor coupled circuit as transformers are by far the most expensive single components in the circuit.

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Lundahl LL1538 Technical Datasheet

LUNDAHL
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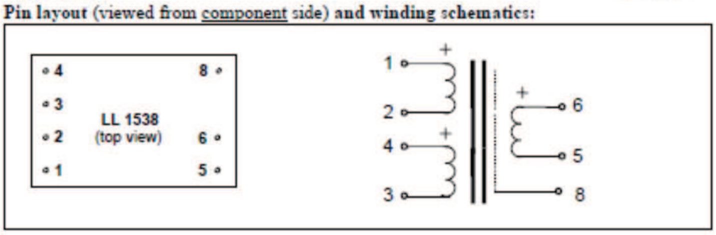
Microphone Input Transformers
LL1538 and LL1538XL

The LL1538 and the LL1538XL are high performance microphone input transformers, each with a high permeability mu-metal core and two three-section coils.
In the LL1538XL the core is about 45% larger than in the LL1538, resulting in a larger level capability. In both types, primary and secondary windings are separated by electrostatic shields. The three-section winding structure of the transformers results in a very low leakage inductance and thus an excellent frequency response.
The transformers are encapsulated in mu-metal cases for magnetic shielding.

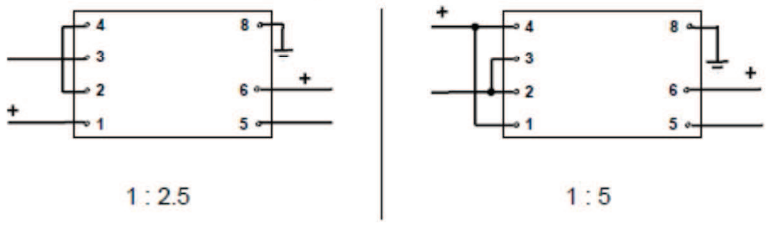
Turns ratio: 1 + 1 : 5

Pin layout (viewed from component side) and winding schematics:



	LL1538	LL1538XL
Dimensions (Max. Length x Width x Height above PCB (mm))	38 x 24 x 17	38 x 24 x 20.5
Spacing between pins	5.08 mm (0.2")	5.08 mm (0.2")
Spacing between rows of pins	27.94 mm (1.1")	27.94 mm (1.1")
Weight	46 g	65 g
Rec. PCB hole diameter	1.5 mm	1.5 mm
Static resistance of each primary	44Ω	61Ω
Static resistance of each secondary	880 Ω	975 Ω
Distortion (primaries connected in parallel, source impedance 200Ω)	0.2 % @ 0 dBU (0.775V rms) primary level, 50 Hz	0.2 % @ + 3 dBU (1.1V rms) primary level, 50 Hz
	1 % @ + 10 dBU (2.5 V rms) primary level, 50 Hz	1 % @ + 13 dBU (3.5V rms) primary level, 50 Hz
Self resonance point	> 120 kHz	> 120 kHz
Optimum termination for best square-wave response (Connection 1:5, source imp. 200Ω)	No termination necessary	No termination necessary
Frequency response (source 200 Ω, no termination)	10 Hz - 100 kHz +/- 0.3 dB	10 Hz - 80 kHz +/- 0.3 dB
Isolation between windings/ between windings and shield	4 kV / 2 kV	4 kV / 2 kV

Connection alternatives



R980616

Lundahl LL1517 Technical Datasheet

LUNDAHL
– TRANSFORMERS –

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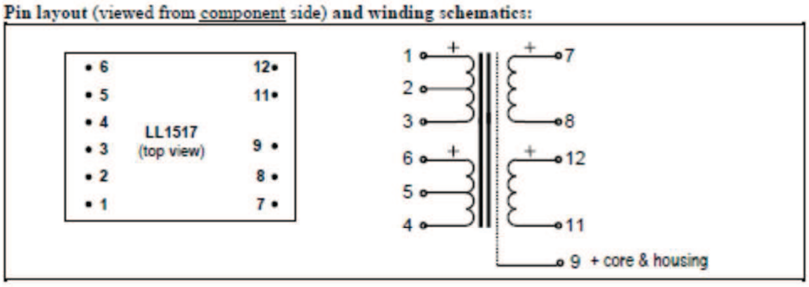
Audio Output Transformer
LL1517

LL1517 is an audio output transformer for balanced or unbalanced drive. The transformer is built from two three-section coils, with primaries and secondaries separated by electrostatic shields, and a audio C-core of our own production. The transformer is housed in a mu-metal housing.
The LL1517 has sufficient low copper resistance to meet broadcast specifications in a conventional drive configuration, but is (as all output transformers) ideally used with mixed feedback drive circuits. (See separate paper for mixed feedback design principles).

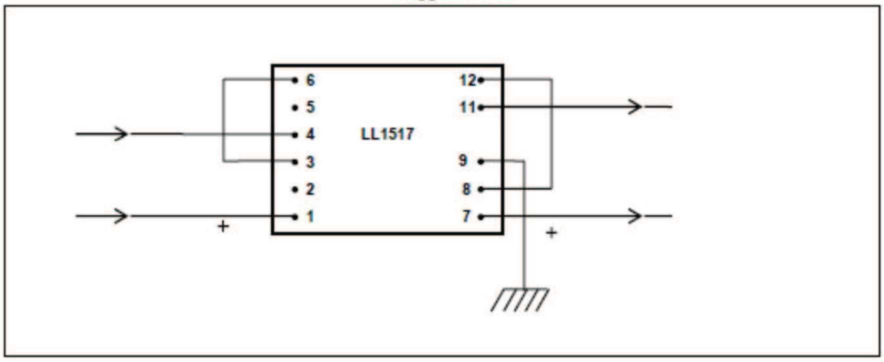
Turns ratio:	1 + 1 : 1 + 1
Dims (Length x Width x Height above PCB (mm)):	47 x 34 x 18

Pin layout (viewed from component side) **and winding schematics:**



Spacing between pins:	5.08 mm (0.2")
Spacing between rows of pins:	35.56 mm (1.4")
Weight:	105 g
Core:	Audio C-core
Housing:	Mu-metal
Rec. PCB hole diameter:	1.5 mm
Static resistance of each primary:	9.2 Ω
Static resistance of each secondary:	9.5 Ω
Leakage inductance of secondaries (sec. in series):	0.3 mH
No-load impedance:	Typically > 600Ω @ 50 Hz, +20 dBU
Optimum source impedance:	Minus 18 Ω (See above)
Balance of output (according to IRT, source < 10 Ω , Load 600 Ω):	> 60 dB
Maximum output level before saturation (sec. in series, load 600 Ω)	+ 24 dBU @ 30 Hz
Distortion (achieved with mixed feedback drive circuit, load 600 Ω)	< 0.03 % @ 20 dBU, 30Hz
Frequency response (source 10 Ω, load 600 Ω):	10 Hz -- 80 kHz +/- 0.3 dB
Loss across transformer (at midband with 600 Ω load):	0.3 dB
Isolation between primary and secondary windings / between windings and core:	4 kV / 2 kV

Suggested use



R160916 PL

Measurement Gear

Agilent 1732C impedance meter

HP 54600B oscilloscope, 2ch, 100MHz

General Radio Decade Resistor Type No. 1432-M

Keithley 2015 THD Multimeter

Wavetek Model 184 5MHz Sweep Generator

Tektronix SG505 low distortion generator, modified for 50 ohm source impedance and +22dBu output level

Tektronix AA501 distortion analyser, differential input, 200kΩ input impedance

Audio Presicion Portable One Dual Domain Audio Test System