Development of a Hybrid Over Ear Active Noise Cancellation Headphone based on a DSP Chip

Master Thesis completed by

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STATUTORY DECLARATION

I declare that I have authored this thesis independently, that I have not used other than the declared sources / resources and that I have expressly marked all material which has been quoted either literally or content-wise from the used sources.

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Martin Denda

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Abstract

This thesis describes the development of a hybrid active noise cancellation (ANC) headphone based on a digital signal processor.
At the moment, the majority of active noise cancellation headphones are using analog techniques to provide active noise reduction, especially for the consumer electronics market. In this thesis, a digital signal processor has been chosen to build a hybrid ANC headphone which can be compared to an existing equivalent analog ANC headphone. The main question was about the feasibility of designing a digital ANC headphone with the same performance as the analog reference ANC headphone solution.
Both systems have been built in order to make it possible to compare them in terms of ANC performance, power consumption and audio quality.
At the end of this thesis, a listening test has been set up, where test persons where able to subjectively compare the two systems to evaluate whether there is any audible difference between the analog and digital solution.

Kurzfassung

Der Großteil der zurzeit am Markt erhältlichen Active Noise Cancellation (ANC) Kopfhörer verwendet analoge Schaltungstechnik um den Umgebungslärm zu unterdrücken.
Die vorliegende Arbeit beschäftigt sich mit der Entwicklung eines Hybrid ANC Kopfhörer, welcher mit Hilfe eines digitalen Signalprozessors realisiert worden ist.
Ziel dieser Arbeit war es herauszufinden, ob mit Hilfe digitaler Technik ein ANC Kopfhörer aufgebaut werden kann, welcher ähnliche Ergebnisse liefert wie eine analoge Referenzlösung. Zu diesem Zweck wurde neben einem geeigneten digitalen Signalprozessor ein analoger Referenzkopfhörer ausgewählt. Beide Systeme wurden so aufgebaut, dass ein Vergleich hinsichtlich der ANC Performance, des Stromverbrauches und der Klangqualität möglich ist. Am Ende der Arbeit wurde ein Hörtest abgehalten, bei dem die Probanden die beiden Systeme direkt miteinander vergleichen konnten um herauszufinden, ob es einen hörbaren Unterschied gibt oder nicht.

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Chapter 1

Introduction

Nowadays one is confronted with unwanted noise almost 24 hours a day. Travelling by train, subway or airplane, for instance, saves time and is comfortable, but travelers are often negatively affected by high noise levels. Noise may especially cause concentration and communication problems in people.

Also, long-term noise exposure poses a threat to the health of most beings.

This thesis deals with the development of a hybrid active noise cancellation headphone based on a digital signal processor (DSP). Active noise cancellation headphones reduce unwanted ambient noise by means of active noise control (ANC).

With ANC headphones, it is also possible to listen to music at low volumes although the ambient noise level is very high, thereby protecting the ears against long-term damage.

1.1 Motivation

Currently, the majority of ANC headphones are based on analog technology to render active noise reduction. Analog solutions sport specific advantages over digital solutions. For example, analog systems are able to process a signal in real time while power consumption of analog systems is limited and generally not as high as in digital systems.

With the development of modern digital signal processing, digital microchips are performing a lot faster and more efficient. Thus, the possibility of building a digital ANC headphone performing much like the analog solution is getting close to reality.

1.2 Goal

The goal of this thesis is to build a hybrid over-ear active noise cancellation system using a digital signal processor (DSP) and compare this system to an analog reference ANC solution.

In order to achieve this, a DSP with the technical requirements to build a digital ANC system has to be identified. Also, an analog reference ANC headphone has to be identified for comparison. The development of the ANC system includes building the hardware, programming the DSP and comparing the digital to the analog system.

The main questions are:

- What are the technical requirements for the DSP needed to build a good working ANC system?
- Which settings of the DSP are most efficient and are leading to the best performance?
- Is it possible to achieve the same or better ANC performance with the digital system in comparison to the analog solution?
- How much power would the digital solution need, is it possible to use it as a portable system?
- Which benefits would a digital solution have when compared to the analog solution?

At the end of this thesis, a listening test will also be set up to compare the digital ANC system to the analog ANC system, in order to find which of the solutions – subjectively – can be considered better and feels more comfortable and may, therefore be more acceptable.

Chapter 2

Principles of Active Noise Cancellation

2.1 Definition of Noise

In physics, noise is a non-periodic random signal with frequencies randomly spread over a continuous frequency range. In a more common definition, noise is defined as unwanted sound, which is disturbing and makes people feel uncomfortable.
The frequency spectrum of noise can be highly broadband, like daily traffic noise for example, or can be more narrow band, like the low-frequency noise of an aircraft turbine.

Traditional passive hearing protectors, like earmuffs or earplugs, are widely known and protect the ears from harmful or unwanted noise. Both of them attenuate the noise because of their mechanical structure, blocking the air transmission pathway to the ear. Therefore, they are called **passive** hearing protectors.

Figure 2.1 shows two popular traditional passive hearing protectors:

Figure 2.1: Traditional passive hearing protectors

Although passive hearing protectors perform well, they are not able to attenuate all frequencies in the same way. In general, they are able to block out high frequencies very well, but they do have a bad attenuation performance for frequencies below 500 Hz.
As mentioned before, an aircraft-turbine generates low-frequency noise below 500 Hz, so damping these frequencies with passive hearing protectors would not be successful.

2.2 Active Noise Cancellation

Indicated by the name, active noise cancellation is an active process in contrast to the passive hearing protection. The basic principle, which was invented by Dr. Lueg in the year 1937, is about producing an "anti-noise" signal which mimics the unwanted noise in every respect, but the phase is shifted by 180°.
Therefore, the "anti-noise" signal can be seen as the exact opposite of the unwanted noise. If these signals are mixed, they will cancel each other out because of destructive interference, as shown in figure 2.2:

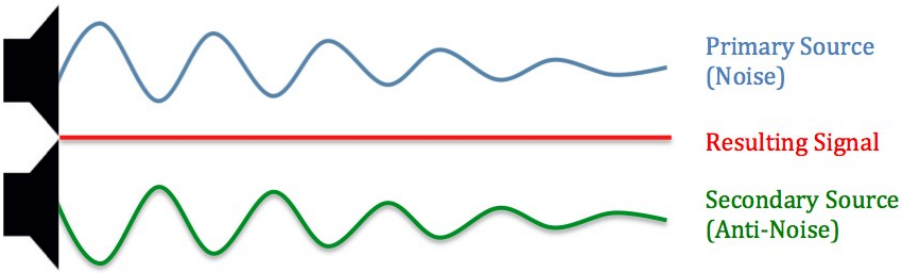

Figure 2.2: Destructive Interference

An exact phase-inverted copy of the noise signal will cause a resulting signal equaling zero, so it would be the perfect cancellation.

Up to now perfect cancellation could not be achieved because it has, so far, not been possible to produce such an ideal "anti-noise" signal. This is because the components of the ANC system, like the microphone, the loudspeaker and the headphone cup, themselves have their own phase characteristics. If the ANC system is a digital system, the DSP will also bring in its own phase characteristics due to propagation delay.
The ANC controller can, in part, compensate for the phase characteristics, but it is not possible to erase them completely.
Therefore, a real "anti-noise" signal will always be an approximation to an ideal "anti- noise" signal. The better the approximation to the ideal "anti-noise" signal, the better the ANC system will perform.

The components required for building an ANC system are shown in figure 2.3:

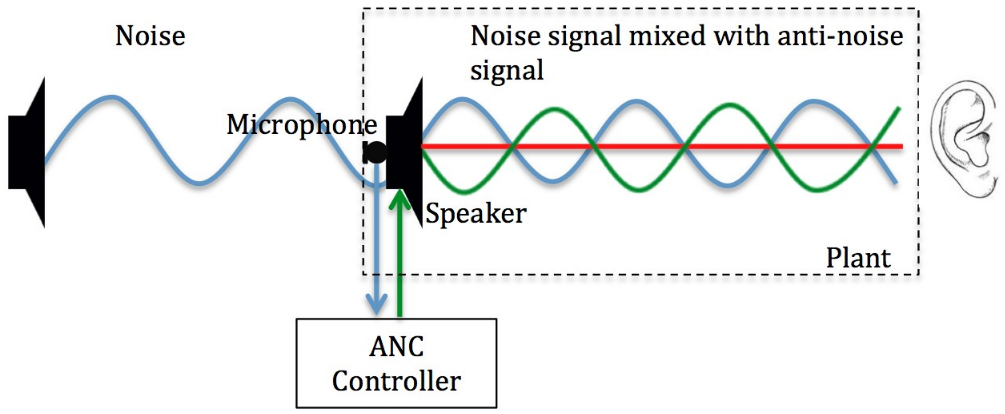

Figure 2.3: Components of an ANC system

The microphone senses unwanted noise and sends it to the ANC controller. The ANC controller produces the anti-noise signal by inverting the phase of the signal received by 180˚. In addition to this phase shift, the ANC controller tries to compensate for the non-ideal transfer function of the other components, which is accomplished by several filters inside the controller.
The loudspeaker renders the "anti-noise" signal back to the physical system (plant), which may, for instance, be the headphone cavity.
With figure 2.3, it is easy to understand that processing of the signal by the ANC controller has to be very fast. Otherwise, the "anti-noise" signal would always be "too late" for the noise signal, causing phase mismatches, which are reducing the ANC performance.

2.3 ANC in a Headphone

Basically, there are three approaches to building an ANC system in a headphone:

- Feed-forward ANC system
- Feed-back ANC system
- Hybrid ANC system

The main difference between these systems is in the position of the microphone that senses unwanted noise. In the feed-forward ANC system, the microphone is positioned outside the ear cup, whereas the position of the microphone in a feed-back ANC system is inside the ear cup, as shown in figure 2.4.

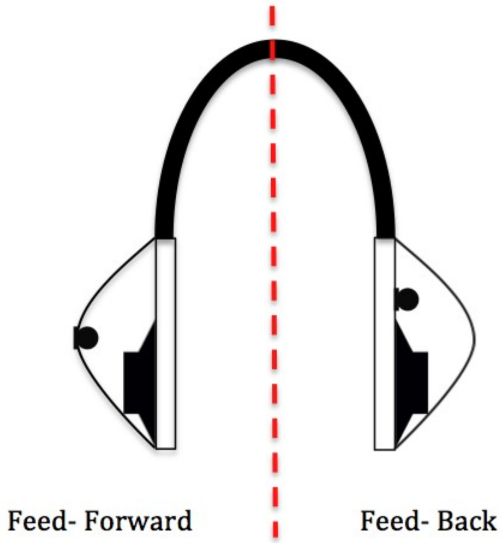

Figure 2.4: Different ANC types in a headphone

2.3.1 Feed-forward ANC system

The microphone in a feed-forward system is positioned outside the ear-cup. Therefore, the unwanted noise can be sensed directly by the microphone and there is no feedback between the loudspeaker and the microphone. This is a big advantage of feed-forward systems:
If they are built correctly, they are always stable because they represent an open-loop system.
Due to this reason, it is important to – acoustically - decouple the microphone from the loudspeaker, so that there is no crosstalk between them.

Figure 2.5 shows a block diagram of a feed forward ANC system, where X(s) represents the command input, Z(s) the external disturbance (noise) and Y(s) the system output. The command input X(s) dictates the desired system output Y(s), so it can be set to zero. The headphone transfer function $HH(s)$ includes the passive damping of the head- phone and the reflections inside the headphone. The loudspeaker transfer function $HL(s)$ includes the transfer function of the loudspeaker, as well the reflections inside the headphone.

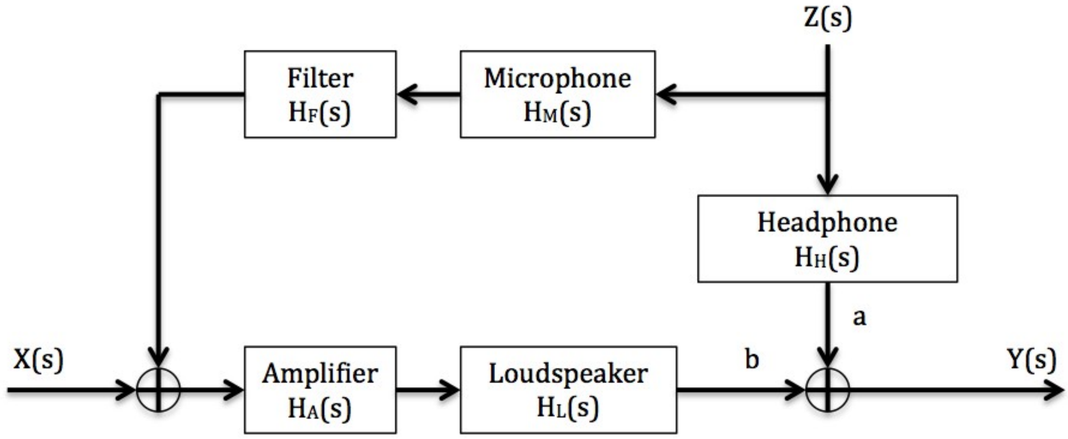

Figure 2.5: Feed forward system

The goal is to keep the system output Y(s) equal to zero:

$$Y(s) = a + b \overset{!}{=} 0 \tag{2.1}$$

with

$$a = Z(s) \cdot H_H(s) \tag{2.2}$$

and

$$b = Z(s) \cdot H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s) \tag{2.3}$$

Equation 2.2 and 2.3 put into 2.1 result in:

$$Y(s) = Z(s) \cdot (H_H(s) + H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s)) \overset{!}{=} 0 \tag{2.4}$$

so the transfer function from the disturbing input to the output is

$$\frac{Y(s)}{Z(s)} = H_H(s) + H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s) \overset{!}{=} 0 \tag{2.5}$$

If the filter is designed so that

$$\boxed{H_F(s) = -H_H(s) \cdot H_M(s)^{-1} \cdot H_A(s)^{-1} \cdot H_L(s)^{-1}} \tag{2.6}$$

the transfer function from the disturbing input to the output equals zero.

The **filter** has to compensate the **headphone** transfer function $H_H(s)$, the **microphone** transfer function $H_M(s)$, the transfer function of the **amplifier** $H_A(s)$ and the transfer function $H_L(s)$ of the **loudspeaker**. The minus in equation 3.6 stands for the phase shift of 180˚.

2.3.2 Feed-back ANC system

In a feedback system, the microphone is positioned inside the ear cup, as shown in figure 2.3.6. The microphone is sensing the noise coming from outside, already damped by the passive attenuation of the headphone cup. Also, the microphone senses the output of the loudspeaker, building a feedback loop, therefore the system is called a **closed loop** system.
Figure 2.6 shows again the block diagram of the feedback system:

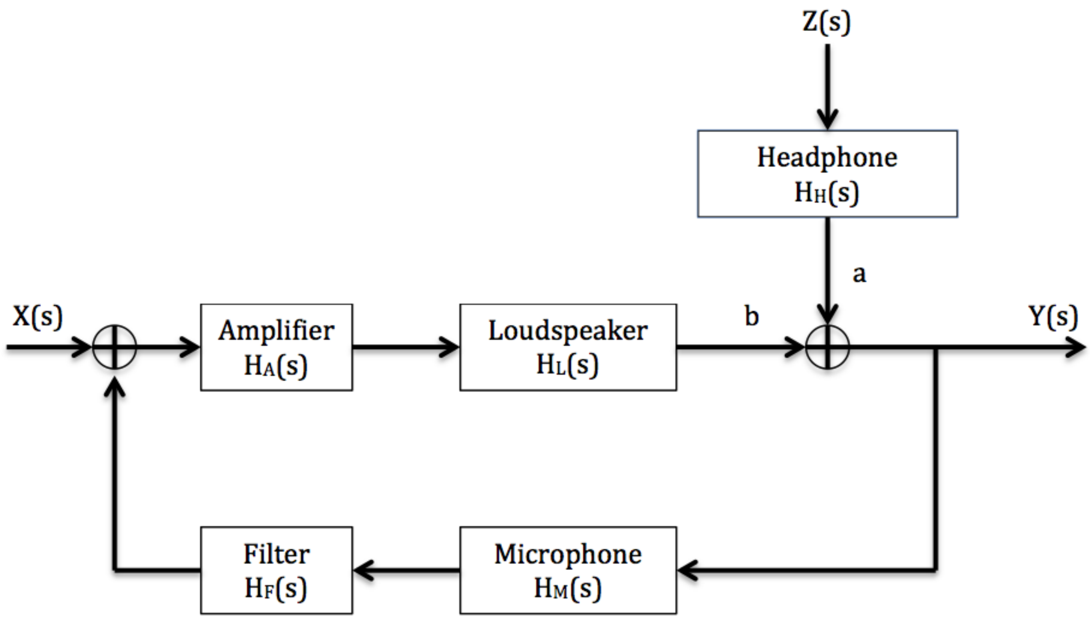

Figure 2.6: Feedback system

The goal is to keep the system output Y(s) equal to zero:

$$Y(s) = a + b \overset{!}{=} 0 \tag{2.7}$$

with

$$a = Z(s) \cdot H_H(s) \tag{2.8}$$

and

$$b = a \cdot H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s) \tag{2.9}$$

Equation 2.8 into 2.9 results in

$$b = Z(s) \cdot H_H(s) \cdot H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s) \tag{2.10}$$

Equation 2.10 put into 2.7 results in

$$Y(s) = Z(s) \cdot (H_H(s) + H_H(s) \cdot H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s)) \tag{2.11}$$

so the transfer function from the disturbing input to the output is

$$\frac{Y(s)}{Z(s)} = H_H(s) + H_H(s) \cdot H_M(s) \cdot H_F(s) \cdot H_A(s) \cdot H_L(s) \tag{2.12}$$

If the filter is designed so that

$$\boxed{H_F(s) = -H_M(s)^{-1} \cdot H_A(s)^{-1} \cdot H_L(s)^{-1}} \tag{2.13}$$

the transfer function from the disturbing input to the output equals zero.

The **filter** has to compensate for the **microphone** transfer function $H_M(s)$, the transfer function of the **amplifier** $H_A(s)$ and the transfer function $H_L(s)$ of the **loudspeaker**. The minus in equation 2.13 stands for the phase shift of 180˚.

Note that the filter transfer function $H_F(s)$ of a feedback system does not compensate the headphone transfer function $H_H(s)$ like in a feed-forward system, since the microphone is being positioned inside the ear cup.

Suppression of the music signal

ANC headphones are normally be used for listening to music, giving the consumer the opportunity to listen to their music at low volume levels despite the ambient noise level being very high. In a feed-forward system, there is no closed-loop; the music signal can simply be fed into the headphone amplifier.
For a feedback system, this is not trivial: The ANC system would also try to compensate for the music signal, together with the noise signal. The result is usually a decreased low frequency level in the music signal.

A solution for this problem is shown in figure 2.7:

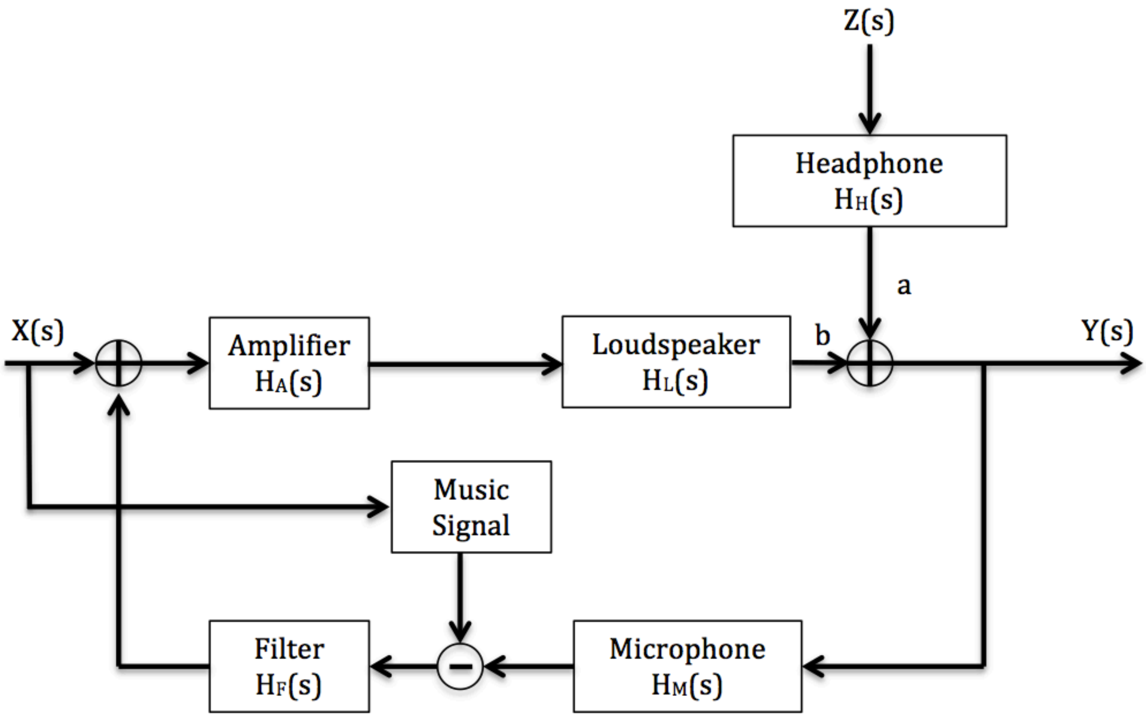

Figure 2.7: Suppression of the music signal

To distinguish the noise signal from the music signal, the music signal is subtracted from the sensed microphone signal via another filter.

2.3.3 Hybrid ANC system

If a headphone combines both feed-forward and feed-back ANC systems, it is called a **hybrid** ANC headphone.
The advantage of implementing both systems in one headphone is in better ANC performance. The feed-forward microphone is sensing the noise and damping it by playing back the "anti-noise" signal. At the same time, the feed-back microphone senses the – already - attenuated noise, playing back its own "anti-noise" signal to damp it even more.

2.3.4 Filter determination feed-forward

As shown in chapter 2.3.1 and 2.3.2, the filter tries to compensate for the transfer functions of the headphone cup (path 1, only feed-forward), the microphone (path 2), the amplifier and the loudspeaker (path 3). The path for these transfer functions is shown in figure 2.8:

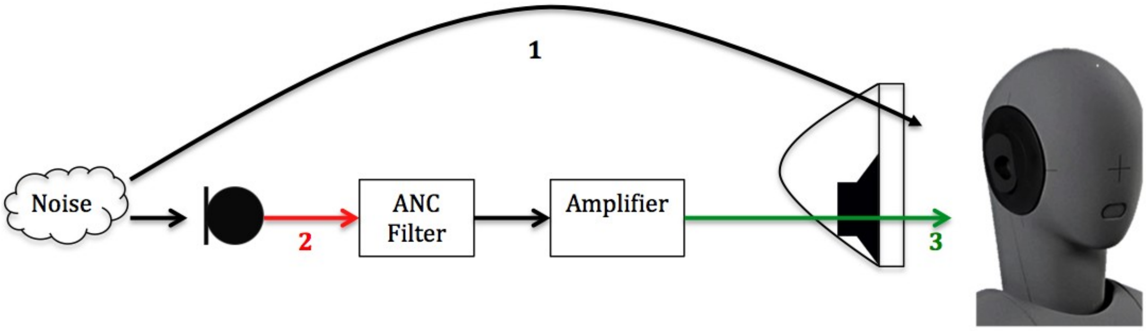

Figure 2.8: Filter determination (Feed-Forward)

The key to a properly working ANC system is in determining the accurate filter gain and phase characteristics, corresponding with the existing system.
To obtain these values, the measurement of the acoustic and electric paths, which are shown in figure 2.8, is necessary. These 3 measurements are fundamental to determining the filter coefficients for the ANC controller.

Path 1: Passive attenuation of the headphone

For this measurement, a loudspeaker is put in front of a dummy head, which has measurement microphones inside the artificial ear canal. The headphones which have to be characterized are put onto the dummy head. The loudspeaker plays back an exponential sine signal sweep and the measurement microphones inside the dummy head measure the amplitude and phase response.
As explained above, the transfer function of the headphone is only needed for the feed- forward ANC system.

Path 2: Microphone

It is recommended to use an omnidirectional microphone to sense the noise coming from all directions, since the position of noise source and head position are changing all the time in a real-world situation. Omni directional microphones tend to act like cardioid microphones for high frequencies. This does not have a big impact because the higher frequency range is already covered by the passive attenuation effect and is not captured by the filter anymore. [9]
The headphone with the ANC microphones stays on the dummy head.
Again, an exponential sine sweep is played back by the loudspeaker. The amplitude and phase response of the ANC microphones is now measured at the out- put of their corresponding microphone amplifiers.

Path 3: Amplifier and Loudspeaker

To measure the last path the exponential sine sweep is fed directly into the amplifier of the loudspeaker inside the headphone. The measurement microphones inside the dummy head are now recording the amplitude and phase response. This measurement represents the exact transfer function of the headphone amplifier, the loudspeaker and the acoustic path between headphone loudspeaker and the ear.
The next step is about the calculation of the filter curve for the active filter.

Filter calculation feed-forward

After the measurement of the individual transfer functions, the right amplitude and phase values of the ANC filter have to be calculated.
The amplitude function A_{Filter} of the desired ANC filter can be calculated as shown in the next equation, with A_1, A_2 and A_3 as amplitudes of the single measurements:

$$A_{Filter} = A_1 - (A_2 + A_3) \tag{2.14}$$

The required phase is

$$\phi_{Filter} = 180° + \phi_1 - (\phi_2 + \phi_3) \tag{2.15}$$

This resulting filter now includes all three parts of the system.

Figure 2.9 shows a typical ANC filter curve for the left and right side of a feed-forward system:

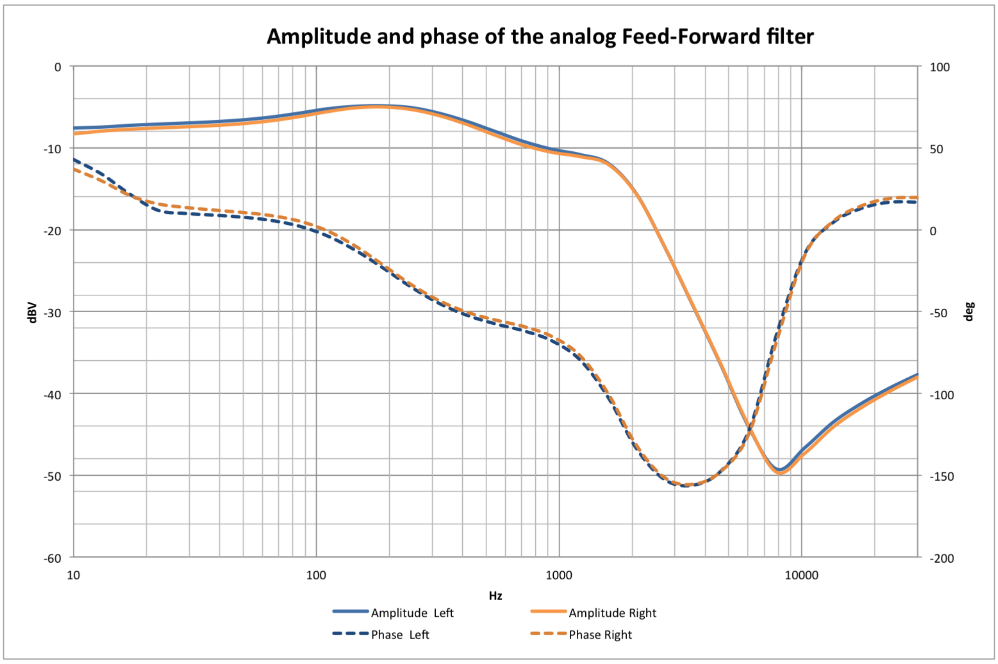

Figure 2.9: Bose Feed-Forward Filter

2.3.5 Design challenges for feed-back

The biggest challenge when designing a feed-back ANC system for a headphone is stability.

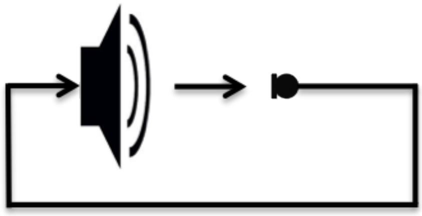

Figure 2.10: Acoustic feedback loop

As explained earlier, the loudspeaker is inverting the phase of the noise picked up by the mic by 180° to cancel the noise in an ANC system. In the higher frequency range above 1000 Hz large phase shifts can occur due to several reasons:

- The physical distance between microphone and loudspeaker produces a phase-shift which can lead to unwanted feedback noise. It is necessary to keep the microphone as close to the loudspeaker as possible, in order to shift the feedback frequencies into high frequencies.
- The reflections inside the ear cup are introducing a lot of different phase-shifts and resonances. Because of the small distances inside the ear cup, resonances will mainly occur in the high frequency region.
- None of the components inside an ANC system have a linear transfer function with linear amplitude and phase. The microphone and the loudspeakers have their own specific transfer function, and the analog filters and controllers introduce their own phase-shifts. If digital filters are used, the DSP itself causes phase shifts because of its propagation delay.

Another problem occurs in the very low frequency range, typically below 100 Hz. Every loudspeaker in a headphone has a high-pass filter characteristic. This high-pass filter introduces a phase shift, which also has to be compensated for. The problem is that the high-pass filter is not time invariant; the cut-off frequency depends on how tight the headphone fits the listener.

2.3.6 Choosing the right headphone

There are different types of headphones on the market, and not all of them perform in the same way for ANC.
As we have learned in chapter 2.3.5, ANC in headphones only works for the low frequency range. The attenuation of the high frequency region can be accomplished by the passive damping of the headphone shell, sealed to the head using an appropriate cushion to block the sound from the outside. The cushion of the headphone has to be soft and flexible for a good fit, but also has to be stiff and heavy enough not to vibrate and radiate sound into the shell cavity (as shown in figure 2.11).

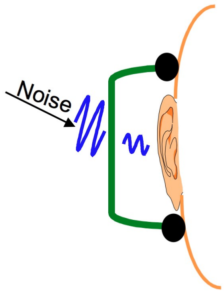

Figure 2.11: Passive Attenuation of a headset (Figure;[2])

In 1962, Shaw and Thiessen demonstrated [10] that the passive attenuation of an headset behaves like a second order mechanical system, here M is the shell mass, R is the cushion damping and K_v and K_c the stiffness of the air in the shell volume and the cushion, respectively:

$$ATT = \frac{K_v}{K_v + K_c + j\omega R - \omega^2 M} \tag{2.18}$$

Since the mechanical impedance increases with frequency, due to the shell mass, the attenuation is bigger for high frequencies.
The combination of damping the low frequencies with active noise cancellation and damping the high frequencies with the acoustic isolation of the headphones results in broadband-noise cancellation:

"A good active headset will effectively combine low frequency-active attenuation with high frequency-passive attenuation to provide high attenuation of the external noise at a wide frequency range."[2]

Circum aural versus supra- aural

Circum aural headphones have ear pads, completely surrounding the ear. Because they can be designed to fully seal the head, the passive attenuation of circum aural headphones is generally very high. A negative aspect is that they are very big and heavy, so this may not be the most appropriate headphones type for mobile applications.

Supra aural headphones have ear pads which are pressing against the ear, rather than surrounding the ear completely. These headphones are smaller and lighter than circum aural headphones, but do attenuate less noise from outside. Because of their size and weight, they are very popular for mobile applications.

Open-back versus closed-back

Both circum aural and supra aural can be open-back or closed-back headphones.
Open-back headphones have the back of the ear cups open. Therefore, the passive damping of noise from outside is not very good; also, more sound from the speaker inside the headphone radiates into the environment. This is especially critical for feed-forward ANC systems because of acoustic feedback with the microphone.
Closed-back headphones block more ambient noise from outside. Most ANC headphones are closed-back, in order to provide the best possible passive damping.

In ear headphones

There is another basic group of headphones, which are very small and directly fit into the outer ear or are inserted into the ear canal. Because these headphones are very small, it is complicated to build them as ANC headphones, especially hybrid ANC systems. Because this work does not deal with ear-fitting headphones, they will not be discussed any further.

Chapter 3

Digital Signal Processing

3.1 Introduction

To understand the content below, some basic principles of digital signal processing have to be considered first.

Figure 4.31 shows the structure of a typical digital signal system:

Figure 3.1: Typical digital system

The analog signal $x_s(t)$ is converted into a digital signal x[n] by the A/D converter. To avoid anti-aliasing effects due to the analog-to-digital conversion, the analog signal $u_e(t)$ has to be band-limited with an analog lowpass filter (see chapter 3.2).
The next part in the signal flow is the digital signal processor (DSP), which is basically a microprocessor that has been especially designed for audio-processing tasks (for example filtering). The digital output signal y[n] of the DSP is converted into the analog output signal $u_a(t)$ using a D/A converter and an analog reconstruction filter.

3.2 Analog - Digital Conversion

To convert an analog signal into a digital signal (A/D conversion), two basic steps are necessary:

- Periodic sampling
- Amplitude quantization

With periodic sampling the time-continuous analog signal is converted into a time- discrete analog signal (pulse-amplitude modulated signal = PAM signal). With amplitude quantization, each analog voltage value of the sampled signal is converted into a corresponding binary digital value.
To convert the digital signal into an analog signal (D/A conversion), the binary values are converted into their corresponding analog values. After that, a low pass filter (reconstruction filter, see 3.2.2) is needed to reconstruct the original analog signal.

3.2.1 A/D conversion

Periodic sampling of an analog signal

The most typical method of obtaining a discrete-time representation of a continuous-time signal is through periodic sampling where the analog signal $x_c(t)$ is modulated with the periodic pulse train

$$s(t) = \sum_{n=-\infty}^{\infty} \delta(t - nT) \tag{3.1}$$

where T is the sampling period, and its reciprocal $f_s = 1/T$ is the sampling frequency in samples per second.
$\delta(t)$ is the unit impulse function, or Dirac delta function:

$$\delta(t) = \infty \quad \text{for } t = 0$$
$$\delta(t) = 0 \text{ for } t \neq 0$$

The resulting signal $x_s(t)$ is

$$x_s(t) = x_c(t)s(t) = x_c(t)\sum_{n=-\infty}^{\infty} \delta(t - nT) = \sum_{n=-\infty}^{\infty} x_c(nT)\delta(t - nT) \tag{3.2}$$

So, the time-continuous signal $x_s(t)$ is a signal which is zero except at integer values of T. The final output sequence

$$x[n] = x_c(nT) \tag{3.3}$$

is now indexed only on the integer values of T (nT), therefore introducing a time normalization.

Frequency domain representation of sampling

The multiplication of $x_c(t)$ and s(t) in the time domain corresponds to the convolution of $x_c(t)$ and s(t) in the frequency domain:

$$X_s(j\omega) = X_c(j\omega) * S(j\omega) \tag{3.4}$$

Figure 3.2 shows the spectrum of the original signal $x_c(t)$:

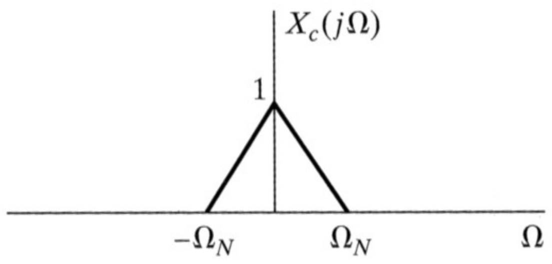

Figure 3.2: Spectrum of the original signal (Figure:[1])

The Fourier-transformation of the periodic impulse train s(t) again is a periodic pulse train, with the fundamental frequency of $\Omega_s = f_s$

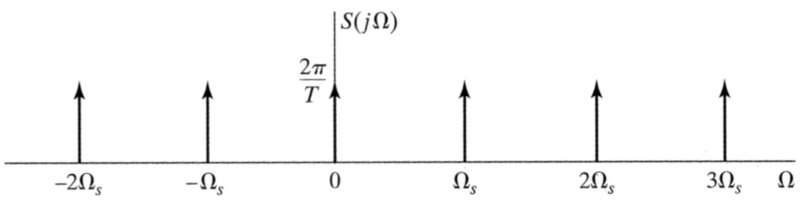

Figure 3.3: Spectrum of the sampling function (Figure:[1])

The convolution of these signals is the spectrum of the sampled signal:

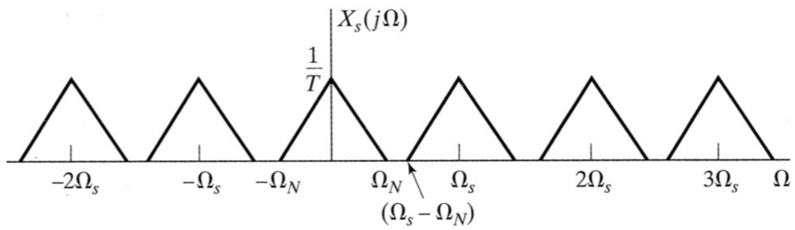

Figure 3.4: Spectrum of the sampled signal with $\Omega_s > 2\Omega_N$ (Figure:[1])

The original spectrum of $x_c(t)$ is copied periodically by multiple integers of the sampling frequency f_s. To reconstruct the original signal $x_c(t)$ out of $x_s(t)$, an ideal lowpass filter is needed to eliminate the replicas.
From figure 3.4, it is easy to see that the original spectrum will not be falsified, as long as

$$\Omega_s - \Omega_N > \Omega_N, \text{ or } \Omega_s > 2\Omega_N \tag{3.5}$$

Equation 3.5 describes the **Nyquist-Shannon sampling theorem** which says that a band-limited function can be perfectly reconstructed from an infinite sequence of samples if the band-limit of the function is not greater than 1/2 the sampling rate f_S (samples per second).
If the Nyquist-Shannon theorem is violated, the replicas of $X_c(j\Omega)$ overlap and the original signal is no longer recoverable. This type of distortion is called **aliasing distortion**, or simple **aliasing.**

Figure 3.5 shows the effect of aliasing:

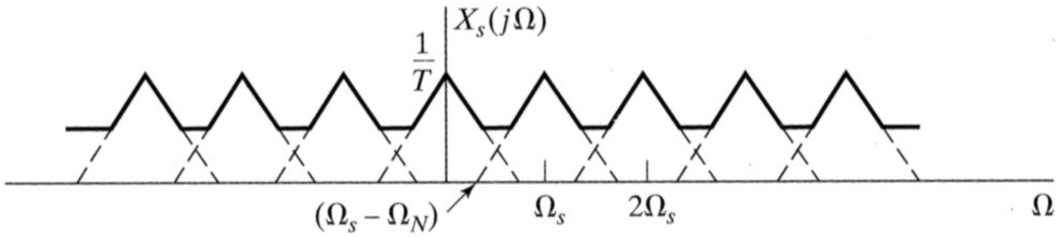

Figure 3.5: Spectrum of the sampled signal with $\Omega_s < 2\Omega_N$ (Aliasing) (Figure; [1])

To avoid these aliasing effects, an analog anti-aliasing filter before sampling has to be inserted. The anti-aliasing filter is a low-pass filter that suppresses all frequencies above $\frac{f_S}{2}$ not to violate the assumptions of the Nyquist-Shannon theorem.

Oversampling

In a real-life system, it is not possible to build an ideal anti-aliasing filter (or reconstruction filter, see 3.2.2) as shown in figure 3.8 because such a filter would have an infinite impulse-response, which is impossible to perform in real-time.
Therefore, the sampling frequency f_S in a real system has to be chosen being greater than the highest frequency of interest (for audio, typically 20 kHz). This means that, in the frequency domain, the replicas of the signal are shifted towards higher frequencies. Because of that, the transition between the pass-band (frequency band where the filter passes frequencies) and the stop band (frequency band where the filter is suppressing any frequencies) does not have to be infinitely steep, as shown in figure 3.6:

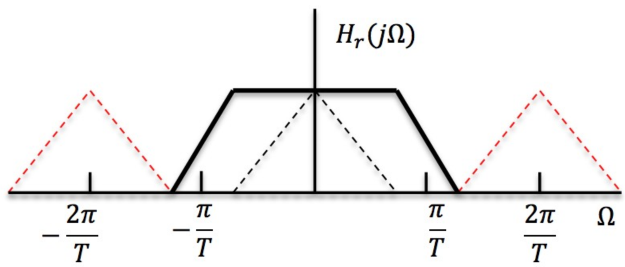

Figure 3.6: Real anti-aliasing / reconstruction filter

To choose a sampling rate higher than theoretically needed (double the highest frequency of interest in the signal = nyquist sampling) is called **oversampling**. The higher the oversampling ratio (OSR) is, the more relaxed are the requirements for the anti-aliasing filter and the reconstruction filter.

Amplitude Quantization

After periodic sampling, the band-limited signal is still an analog signal, but it is now discrete in time. Amplitude quantization turns the analog system into a digital signal, which is discrete in time and amplitude. Therefore, the continuous voltage values of the analog signal are converted to corresponding digital binary numbers.
The most common method of amplitude quantization is the multi bit-quantization, also known as pulsecodemodulation (PCM).
The possible analog voltage range is divided into quantization steps, the amount of quantization steps is given by $n = 2^k$ where k is the number of bits available. The analog voltage is now compared to the quantization curve and gets approximated to the nearest quantization step.
The next figure 3.7 shows an example of the characteristic quantization curve of a four-bit amplitude quantizer, q is called the step size of the quantizer:

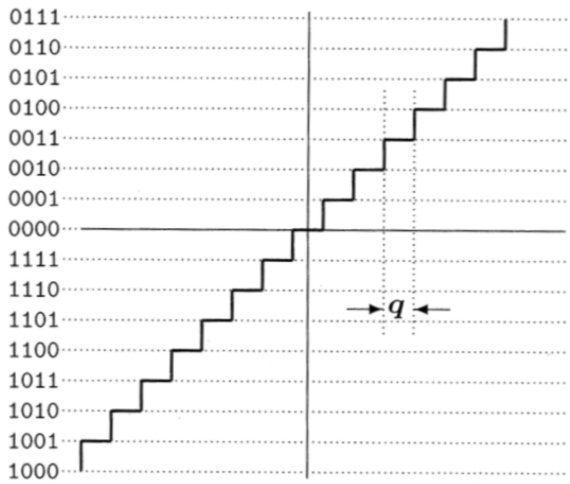

Figure 3.7: 4-bit amplitude quantization (figure; [5])

The error, which occurs due to the approximation of analog voltage, is called **quantization noise**. In general, this quantization noise is white noise, which does not correlate with the sampled signal ([5] [1]).

3.2.2 D/A conversion

Reconstruction of the analog signal

For D/A conversion, the digital binary values are first converted back into their corresponding analog values.
If the conditions of the Nyquist-Shannon theorem are met, it is possible to reconstruct the exact original signal $x_c(t)$ out of $x_s(t)$ using an ideal low-pass filter (compare figure 3.4 and 3.2).
The ideal low pass filter is filtering out the original signal and is eliminating the replicas, so a good choice for the cut-off frequency Ω_c is half the sample rate:

$$\Omega_c = \frac{\Omega_s}{2} = \frac{2\pi f_s}{2} = \pi f_s = \frac{\pi}{T} \tag{3.6}$$

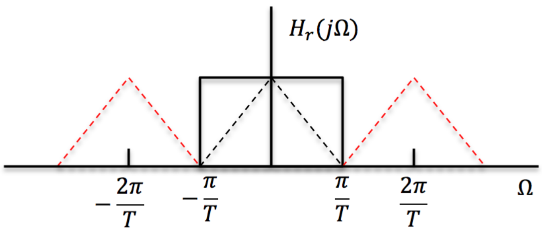

Figure 3.8: Frequency response of an ideal reconstruction filter (Figure:[1])

In the time domain, the corresponding impulse response $h_r(t)$ of this ideal filter with the cut-off frequency of $\Omega_s = \pi/T$ is given by the sinc-function:

$$h_r(t) = \frac{sin(\pi t/T)}{\pi t/T} \tag{3.7}$$

Filtering in the frequency domain is a multiplication which is a convolution in the time domain:

$$x_r(t) = \sum_{n=-\infty}^{\infty} x[n] \frac{\sin[\pi(t-nT)/T]}{\pi(t-nT)/T} \tag{3.8}$$

If we look at the properties of the sinc-function 3.7 more closely, we note that $h_r(0) = 1$ and $h_r(nT) = 0$ with $n = \pm 1, \pm 2, \ldots$.
Putting these values into equation 3.8, then $x_r(nT) = x[n] = x_c(nT)$, so the reconstructed signal $x_r(t)$ is the same as the original signal $x_c(t)$ at multiple integers n of the sample period T.

Between this sampling period, the signal $x_r(t)$ gets interpolated by the sinc-function, as shown in figure 3.9:

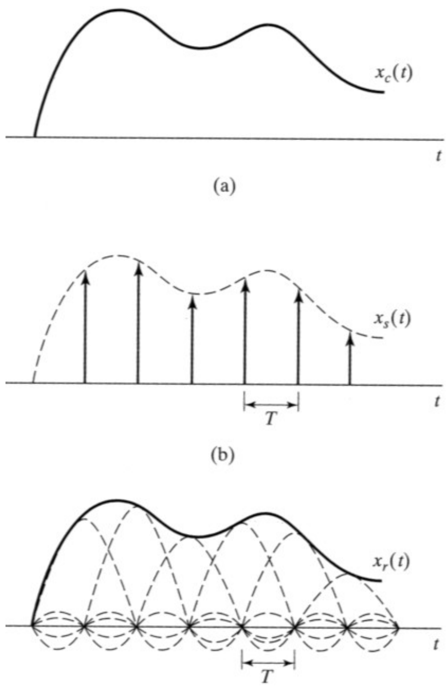

Figure 3.9: Ideal Interpolation Filter (Figure;[1])

3.2.3 Delta Sigma Converter

The Delta Sigma converter (ΔΣ Converter) is the most popular converter in modern digital systems. The basic principle is not to quantify the absolute value of the analog signal (like PCM), but the difference to the previous converted value. Therefore, this type of converter is also called **Differential Converter** [5].

The ΔΣ Converter is using a very high oversampling ratio (up to several MHz) but a very low bitrate, typically 1 bit (2 states are enough: Value is higher / lower than previous sample). There are two main benefits when using an ΔΣ converter:

- Because of the high sampling rate, the requirements for the analog anti-aliasing and the reconstruction filter are relaxed (see 3.2.1). The filter curve does not have to be very steep, therefore, the analog anti-aliasing filter introduces less phase distortion in the band of interest. This is very important, in particular for a good working ANC system.
- By changing the sampling rate, the signal power and total quantization noise power is not affected. Therefore, the signal-to-quantization noise ratio is not changed. However, the quantization noise is spread over a larger frequency range, reducing the spectral density of the quantization noise. When only considering the original Nyquist band, the quantization noise power is reduced by 3 dB for every doubling of the oversampling ratio and the signal-to-quantization noise ratio is improved accordingly ([4]). In addition to that, most ΔΣ Converters are using a technique called **Noiseshaping** that shapes the quantization noise so that most of it is shifted outside the band of interest, as shown in figure 3.10:

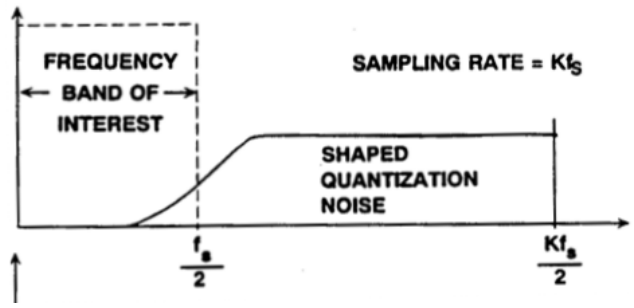

Figure 3.10: Noiseshaping (Figure:[3])

This is achieved by putting the quantization error into a noise-shaping feedback loop. [1] [4]

AD Conversion

The ΔΣ Converter includes a Delta Sigma modulator and a Decimator. The Delta Sigma Modulator converts the analog input signal into a digital signal with a very high sampling rate f_S (up to several MHz) and a very low bitrate of 1 bit. The decimator contains a digital filter and a decimator for converting the sampled data into a signal with a low sampling rate f_D, but high resolution (PCM Signal).

Figure 3.11 shows a simplified block diagram of an ΔΣ AD converter:

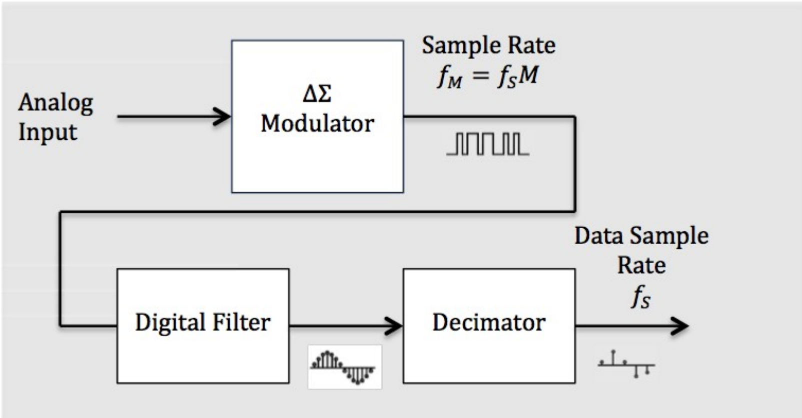

Figure 3.11: Block diagram of an ΔΣ AD converter

The main reason for converting the current signal into a signal with a low sampling rate is that subsequent digital signal processing is easier to realize with low sampling rates.

First, a digital filter is used for converting the output of the modulator into a signal with the same sample rate, but high resolution.
After this, the signal is sent through a decimator which reduces the sample rate by the factor f_S/f_D which is called the decimation ratio or downsampling factor M. Reducing the sampling rate is also called **downsampling**.
Downsampling includes two basic steps, as shown in figure 3.12:

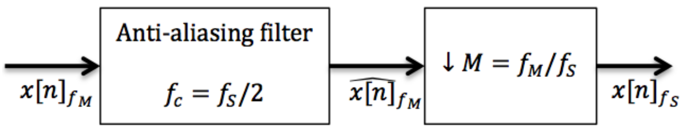

Figure 3.12: Decimator: Downsampling by a factor of M

As shown in chapter 3.2.1, aliasing distortion occurs if the signal contains frequencies higher than half the sampling rate. If reducing the sampling rate by downsampling, the first step is about reducing the bandwidth of the signal to half of the new sampling rate f_D. This is achieved by using a digital low-pass filter.
After that, the sampling rate is reduced by discarding every sample between the new sampling period $T = 1/f_D$.

DA Conversion

Figure 3.13 shows a simplified block diagram of an ΔΣ DA converter which converts a digital PCM signal into an analog output signal:

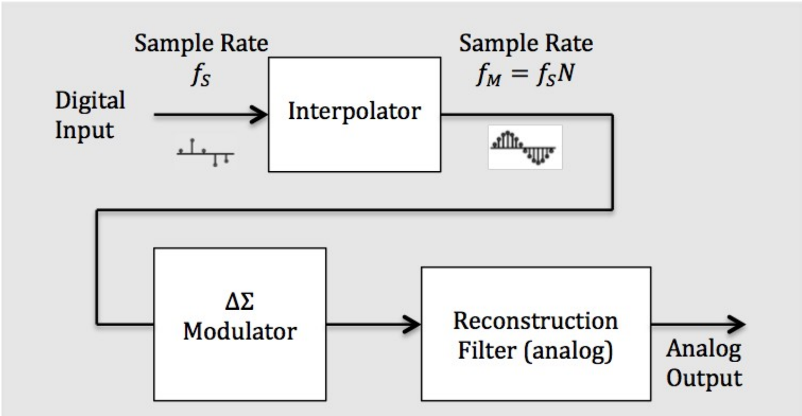

Figure 3.13: Block diagram of an ΔΣ DA converter

The digital input is sent through an interpolator to increase the low sample rate f_S to f_M.
This process is called **upsampling** which includes two basic steps, as shown in figure 3.14:

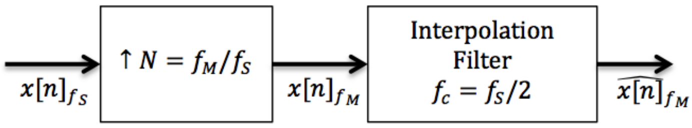

Figure 3.14: Interpolator: upsampling by a factor of M

The first step is about increasing the sample rate by inserting additional zero-valued samples between the samples of the signal $x[n]_{fS}$, also known as "zero-padding" :

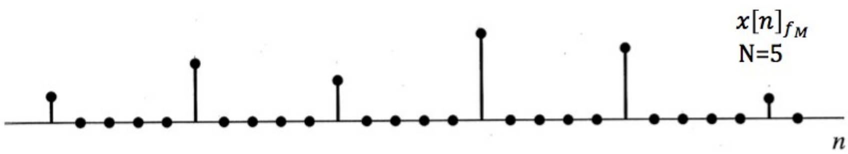

Figure 3.15: Zero padding with N=5 (Figure;[1])

With increasing the sample rate, new replicas of the baseband signal are generated. These replicas are filtered out using a low-pass interpolation filter.

In the time domain, the low-pass filtering fills in the missing values of the inserted zero-samples, therefore "interpolating" between the original samples of $x[n]_{fs}$[1]:

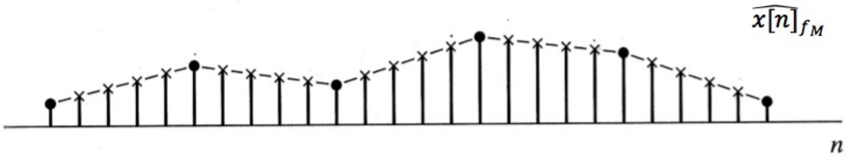

Figure 3.16: Interpolation (Figure;[1])

After increasing the sample rate, the ΔΣ modulator noise-shapes the signal and reduces the sample width to one bit.
The analog reconstruction filter (usually a switched-capacitor filter [7]) now averages the one bit DAC output, removes the shaped quantization noise and rejects any images, which result in the output Nyquist frequency. [3]

3.3 Digital Filters

A digital filter is a system that performs mathematical operations on a sampled, discrete- time signal to reduce or enhance certain aspects of that signal.
This is in contrast to the analog filter, which is an electronic circuit operating on continuous time analog signals. Analog filters are using components like resistors, capacitors and inductors, and in the case of active analog filters, additional operation amplifiers are used.
Digital filters rely on basic mathematical operations, like summing up, delaying and multiplication. Digital filters do have some advantages when compared to analog filters:

- Digital filters are not subject to the component non- linearity that complicates the design of analog filters. Analog filters consist of imperfect electronic components, the values of which are specified to a limit tolerance (e.g., capacitor values often have a tolerance of ±10%) and which may also change with temperature and drift with time. In digital filters, the coefficient values are stored in computer memory, making them far more stable and predictable. [http://en.wikipedia.org/wiki/Digital_filter]
  Therefore, it is possible to reproduce exact copies of any digital filter if the transfer function of the filter is known.

- Changing the transfer function of the digital filter is easier than for analog filters because there is no need to change any hardware, like capacitors or resistors. Changing the behavior of a digital filter means to compute new coefficients and loading new instructions into the DSP. It is also easy to test different filters in "real-life" applications. There is no need for simulating the filters like it is common for analog filter design.

- It is possible to design digital filters which do not exist in the "analog world", for example linear-phase filters or adaptive filters.

There is also one big disadvantage when using digital filters: They introduce a certain latency to the system. In an analog filter, latency is mostly negligible; strictly speaking, it is the time for an electrical signal to propagate through the filter circuit. [http://en.wikipedia.org/wiki/Digital_filter]
In digital filters, latency is a function of the number of delay elements in the system and the calculation speed of the DSP. Also, the power consumption is, usually higher when building a system with digital filters than with analog filters.

Digital Filters can be defined in the time domain by their impulse response $h[n]$.
The input $x[n]$ and the corresponding output $y[n]$ are linked through a simple convolution operation:

$$y[n] = h[n] * x[n] \tag{3.9}$$

There are two main types of digital filters: **finite duration impulse response filters (FIR)** for which the impulse response $h(n)$ is non-zero for only a finite number of samples, and **infinite-duration impulse response filters (IIR)** for which $h(n)$ has an infinite number of non-zero samples.
In the FIR case, the samples of the sequence $h(n)$ are commonly referred to as filter coefficients; for the IIR case, the filter coefficients include feedback terms in a difference equation. [8]

3.3.2 IIR filter

IIR Filters are also known as feedback systems because they do have a recursive path in their structure.
Figure 3.17 shows a direct-form realization of an IIR Filter, using separate summing units:

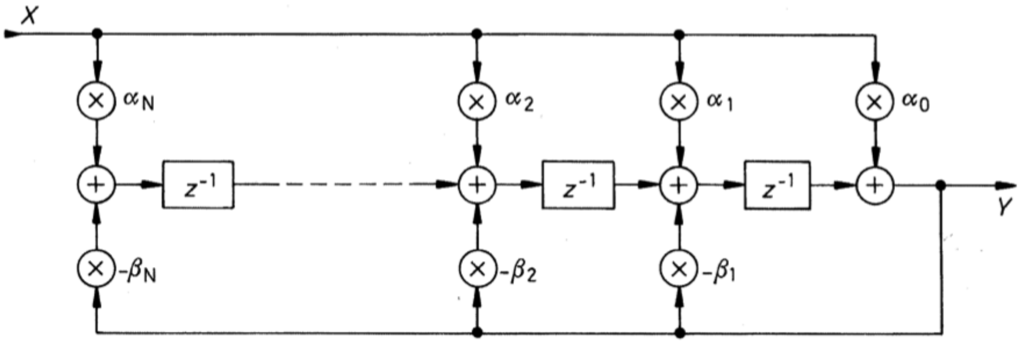

Figure 3.17: Direct Form IIR Filter (Figure: [14])

The number of filter stages is called the order N of the filter. For each filter stage, one delay, two multiplier and one summing units are needed. The difference equation is

$$y[n] = \sum_{k=0}^{N} \alpha_k x[n-k] - \sum_{k=1}^{N} \beta_k y[n-k] \tag{3.10}$$

With the z-transformation [1], the transfer function results in

$$H(z) = \frac{Y(z)}{X(z)} = \frac{\sum_{k=0}^{N} \alpha_k z^{-k}}{1 + \sum_{k=0}^{N} \beta_k z^{-k}} \tag{3.11}$$

The filter is stable if the poles of the transfer function are inside the unit circle in the z-plane. This corresponds to the stability criteria of the analog continuous time filter, where the poles are only in the left half of the s-plane in case the filter is stable. [1]

IIR filters can be designed to provide a frequency response that is a discrete version of the frequency response of an analog filter. A common method to transfer analog filters to their corresponding digital versions is the **Bilinear transformation**. [1]

IIR biquads

Because IIR Filters do have a recursive path, the stability of the filter is always a critical issue, especially if the order of the filter increases. A common method to build filters with high order is to cascade multiple filters with a low order, for example an order of two (the same principle is also often used for analog filter systems).
These filters are also called **biquad filter**, because of their quadratic terms in the nominator and denominator of their transfer function:

$$H(z) = \frac{\alpha_0 + \alpha_1 z^{-1} + \alpha_2 z^{-2}}{1 + \beta_1 z^{-1} + \beta_2 z^{-2}} \tag{3.12}$$

An example for a biquad filter structure is shown in figure 3.18:

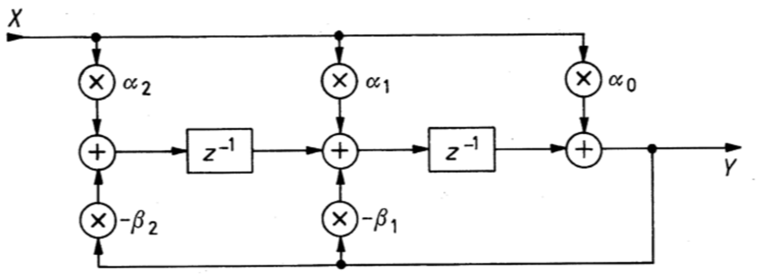

Figure 3.18: Biquad IIR Filter (Figure;[1])

3.3.3 FIR filter

An FIR Filter does not have any feedback structures; the output is a weighted sum of the current and a finite number of previous values of the input. This is the reason why FIR filters are always stable, there are no poles except at z=0 or ∞.
The difference equation is given by

$$y[n] = \alpha_0 x[n] + \alpha_1 x[n-1] + ...\alpha_\alpha x[n-N] = \sum_{k=0}^{N} \alpha_k x[n-k] \tag{3.13}$$

The transfer function results in:

$$H(z) = \frac{Y(z)}{X(z)} = \sum_{k=0}^{N} \alpha_k z^{-k} \tag{3.14}$$

α_k are the filter coefficients, also known as tap weights, that make up the impulse response. The Σ denotes summation from $i = 0$ to $i = N$ where N is the number of feed-forward taps in the FIR filter. Note that the FIR filter output only depends on previous M inputs. This feature explains why the impulse response for a FIR filter is finite. [6]

Figure 3.19 shows a direct form implementation of an FIR Filter:

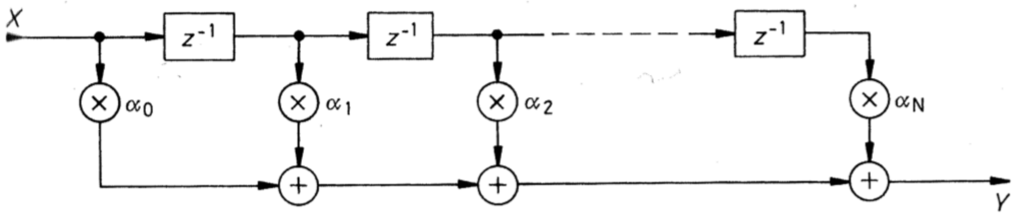

Figure 3.19: Direct Form FIR Filter (Figure; [1])

If the coefficients are chosen so that the impulse response of the FIR filter is symmetric, the filter does have a linear phase characteristic. This means that the propagation delay of the filter is the same for all frequencies, so there is a constant group delay. This is important for several audio applications, for example digital anti-aliasing filtering. [1]

3.3.4 IIR vs. FIR

In general, there is no "better" when comparing an IIR filter to a FIR filter. It depends on the application when choosing between an IIR or an FIR filter because they both have their advantages and disadvantages.
In general, FIR filters do need more than twice the order that IIR filters need for a similar filter curve.

In figure 3.20, there is a comparison between an FIR and an IIR filter for a low-pass filter with comparable filter characteristics:

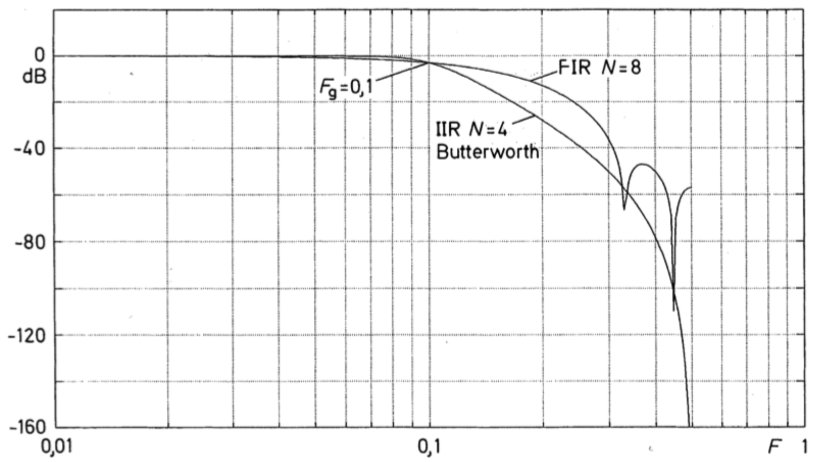

Figure 3.20: Comparison of IIR and FIR filter (Figure; [1])

FIR filters need more computation power than IIR filters, which results in more power consumption of the DSP and/or a longer propagation delay of the overall system. The main advantage of an FIR filter over an IIR filter is that it is always stable and can have linear phase, which is also not possible with analog filters. A comparison between the two filter types can be found in in the next table [1]:

Attribute	FIR Filter	IIR Filter
Required Order	high	less
Required DSP power	high	less
Propagation Delay	high	low
Linear Phase	possible	not possible
Constant Group-delay	possible	not possible
Stability	always	not always
Adaptiv Filter	possible	very dificult

In an ANC system, one of the most critical issues is the propagation delay. Also, power consumption is important when building a portable ANC system, like ANC headphones. Additionally, the filter curves needed are generally not linear phase, as seen from chapter 2.8. So an IIR filter (especially a biquad) may be the better choice for an ANC filter.

3.4 Requirements of the DSP

For building a good working hybrid ANC headphone, the DSP should meet the following requirements:

- As seen in chapter 2.2, it is important that the DSP is able to process the signals very quickly, in order to minimize the propagation delay introduced by the DSP itself. As we will see in chapter 4.2.1, the propagation delay is tightly coupled to the sample rate of the DSP: the higher the sample rate, the faster the DSP is able to process a signal.
- There should be enough microphone inputs and a stereo headphone output (with the corresponding A/D and D/A converter).
- As we have seen from chapter 3.2.3, there is a risk of phase distortion in the band of interest if the A/D converter is not oversampling, because the anti-aliasing filters have to be very steep. This phase distortion will also make it very hard to build a properly-working ANC headphone. Therefore, the A/D converter of the DSP should be able to provide high oversample ratios ($\Delta\Sigma$ converter).
- The DSP should provide the possibility of implementing several filters (FIR or IIR) for the feed-forward and feed-back ANC filters.
- Because the power supply for portable headphones is limited, the DSP should be a low-power device.

Digital low power devices with integrated inputs and outputs and on-board DSPs are often called **audio codecs**. They are often used for portable audio applications like cellular phones or mp3 players.
Currently, there are several audio codecs on the market, giving the engineer the possibility to pick the codec which fits best for the particular application.
For this thesis, the Texas Instruments TLV320AIC3254 audio codec has been chosen, which will be introduced in the next chapter.

Chapter 4

The TI TLV320AIC3254EVM-K Evaluation Board

The TLV320AIC3254EVM-K offers a lot of features; only the key-feature which are important to build up the ANC system can be discussed in this thesis. To learn more about the evaluation module, the TLV320AIC3254EVM-K user's guide [12] is recommended. More information about the audio codec itself can be found in the reference handbook ([13]) by Texas Instruments. Both documents were used to write the following chapter.

4.1 Overview and Features

The TI TLV320AIC3254EVM-K is a complete evaluation kit for the TI TLV320AIC3254 audio codec. The kit includes the TLV320AIC3254EVM board and a USB-based motherboard called the USB-MODEVM interface board.

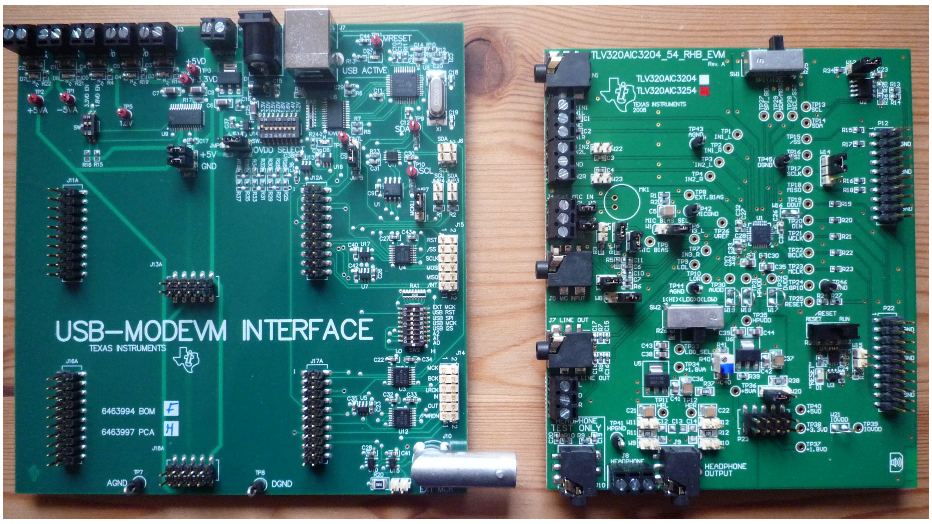

Figure 4.1: USB-MODEVM and TLV320AIC3254EVM

The USB-MODEVM interface board is build around a TAS1020B streaming audio USB controller with an 8051-based core. It can be connected to a PC via the USB port to provide power and control to the TLV320AIC3254EVM. In addition to that, the USB-MODEVM can be used as a sound card to stream audio between the TLV320AIC3254EVM board and the PC.
The TLV320AIC3254EVM includes several analog and digital I/Os and the TLV32AIC3254 audio codec, as seen from figure 4.2.

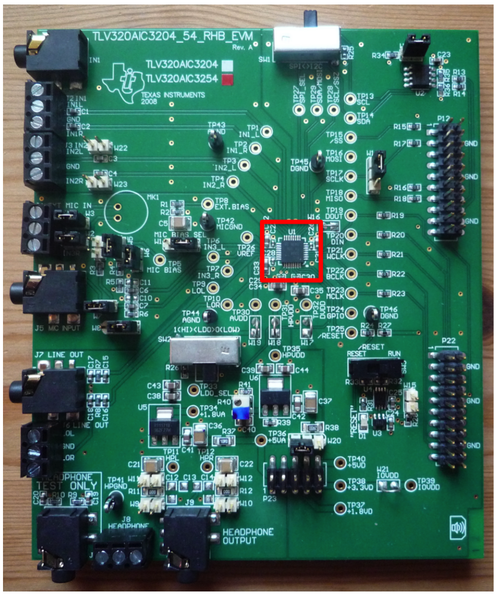

Figure 4.2: TLV320AIC3254 Audio codec on the TLV320AIC3254 Evaluation Board

This codec is a low-power stereo audio codec with integrated stereo AD and DA converters, a miniDSP and programmable inputs and outputs, making it ideal for portable battery-powered audio applications.
The main features of the TLV320AIC3254 audio codec are:

- Stereo audio ADC
- Stereo audio DAC
- Embedded miniDSP
- Six single-ended or 3 fully-differential analog inputs with programmable microphone bias
- Stereo headphone outputs

4.1.1 Analog audio I/O

Analog inputs

There are 6 analog mono inputs or 3 analog stereo inputs in single-ended and/or differential configuration. The gain of the inputs is controlled with two programmable gain amplifiers (PGA) with a range of 0 to +47.5 dB for single-ended inputs and 6 dB to 53.5 dB for differential inputs.
The input impedance can be selected, its either 10 kΩ, 20 kΩ or 40 kΩ.
There is a built-in microphone bias for electret-condenser microphones. The bias-amplifier provides programmable bias voltages and supports up to 3 mA load current to supply multiple microphones.

Headphone output

The stereo headphone driver can drive loads with impedances down to 16 Ω in single ended AC-coupled headphone configuration. Each headphone amp (left and right) can drive up to 15 mW power while operating from 1.8 V analog supplies. In case more power is needed, the headphone amps can also be powered from higher supply voltage, using the LDO_{in}.
The voltage output of the headphone amps can be controlled digitally, providing a range from -6 dB to +29 dB.

4.1.2 A/D and D/A converters

A/D converter

The analog-to-digital converter inside the TLV320AIC3254 is a stereo delta-sigma modulator with a programmable oversampling ratio (OSR). The maximum sampling rate f_M of the modulator is 6.758 MHz. Because of the high sampling rates, the requirement for anti-aliasing filtering is very relaxed. There is an analog anti-aliasing filter before the modulator, with 28 dB attenuation at 6 MHz.

After the modulator, three types of digital decimation filters are available, depending on which sampling rate is chosen. The decimation filter removes high frequencies to avoid aliasing when downsampling from the initial sampling rate of $f_M = OSR * f_s$ to the output data sampling rate of f_s. The output of the analog-to-digital converter supports data sampling rates f_s from 8 kHz up to 192 kHz.

As mentioned above, there are three different filters: A, B and C. They cannot be chosen separately, because they are implicitly set through the chosen data sampling rate f_s and the oversampling ratio.
Filter A is used for sample rates up to 48 kHz, Filter B is intended to be used with sample rates up to 96 kHz and Filter C is designed for a sampling rate of 192 kHz.

The following figure 4.3 shows the frequency responses of filter A:

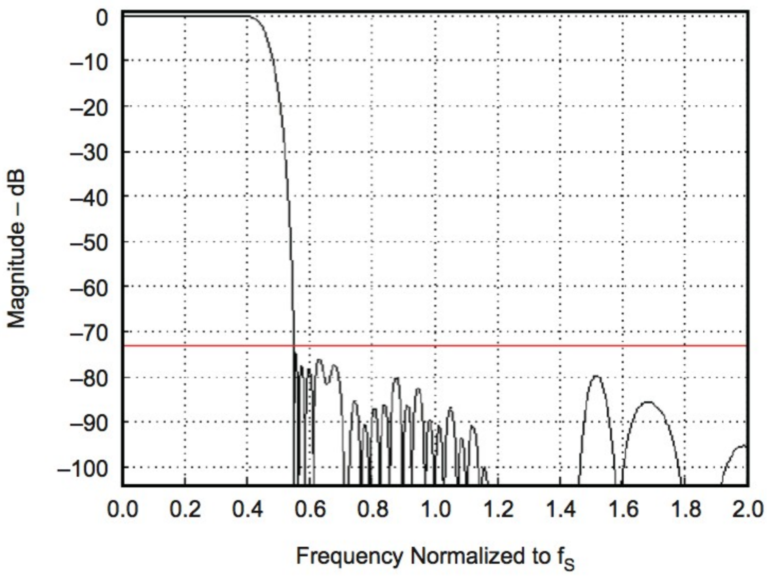

Figure 4.3: ADC frequency response for decimation filter A (Figure; [13])

The common mode of the differential input amplifier can be switched between 0.9 V or 0.75 V. When the gain of the ADC channel is kept at 0 dB and the common mode is set to 0.75 V, a single-ended input of 0.375 V_{RMS} results in a full-scale digital signal at the output of the ADC channel. With the common mode set to 0.9 V, a single-ended input of 0.5 V_{RMS} will result in a fully-scaled signal at the output of the ADC.

D/A converter

The DAC is a delta-sigma modulator with a programmable oversampling ratio and supports input data sampling rates f_S from 8 kHz to 192 kHz. The maximum sampling rate of the modulator is again 6.758 Mhz. There are three different types of digital interpolation filters (A, B and C) before the delta sigma modulator, and an analog reconstruction filter after the delta-sigma modulator.

4.1.3 Clock generation

To build a high performance system, it is important to generate the optimal clock frequencies for each digital processing block, like ADCs, DACs and miniDSPs. Starting from the codec clock-in, the TLV320AIC audio codec provides several programmable clock dividers for generating the required clock frequencies.
Figure 4.4 shows a block diagram of the clock generation:

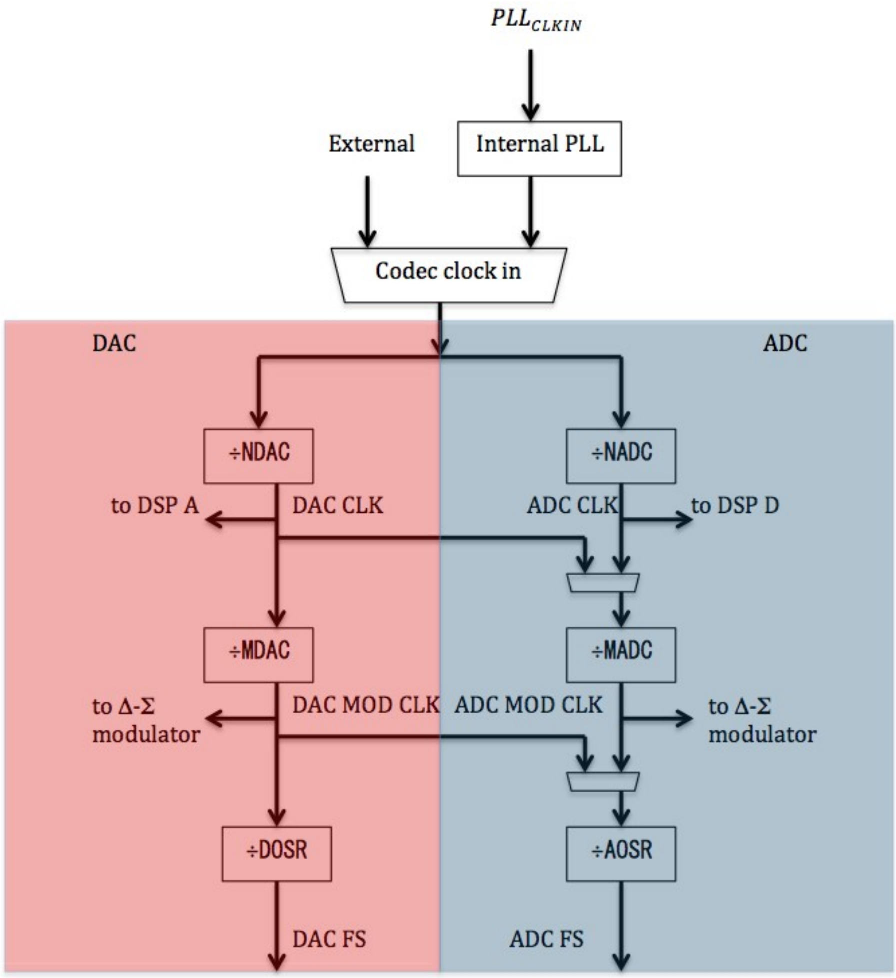

Figure 4.4: Clock generation

In an ideal system, the codec clock-in is connected to an external master clock which is a suitable integer multiple of the desired frequencies. If the system cannot provide the right master clock frequency, the internal PLL can be used to generate the required codec clock-in frequency. The disadvantage of using the internal PLL is in the higher power consumption of the codec.

On the evaluation board, the master clock is generated by a 6 MHz crystal oscillator and a PLL, which are located on the USB-MODEVM.

The Output of the PLL on the USB-MODEVM is connected to the MCKL pin of the TLV320AIC audio codec and sends a master clock signal with a frequency of 11.29 MHz or 12.29 MHz. If a higher master clock frequency is needed, this signal can be routed to the input of the internal PLL of the audio codec, generating the needed codec clock-in frequency.
The PLL output frequency is generated as shown in figure 4.5:

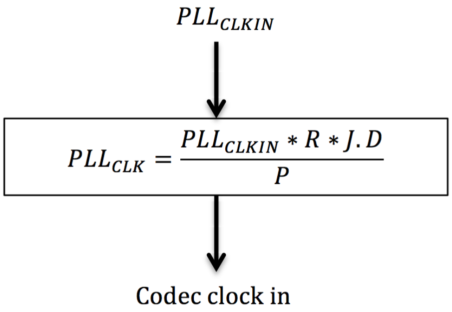

Figure 4.5: Internal PLL

The maximum output frequency of the PLL is 137 MHz.

Starting from the codec clock-in, the first dividers after the PLL are the NDAC, respectively the NADC. These dividers can be configured to generate DAC_{CLK} and ADC_{CLK} which are the clocks for the mini DSPs. The required clock frequency depends on the processing power needed for a certain application.
Very complex applications need more computation steps for one sample; this value can be expressed through the number of **cycles** needed. Thus, if an application requires 96 cycles for both ADC and DAC channels at 44.1kHz, the DAC_{CLK} and ADC_{CLK} has to be set to minimum

$$ADC_{CLK} = DAC_{CLK} = f_S * cycles = 44100 * 96 = 4.233600\ MHz \tag{4.1}$$

With lower sampling rates, more computation power is available. The maximum frequency of the DAC_{CLK} and ADC_{CLK} is 55.296 MHz.

The next blocks are the DAC_{MODCLK} and the ADC_{MODCLK}. These are supplying the delta-sigma modulators. The clock frequency can be adjusted using the MDAC and the MADC dividers. The maximum frequency of the delta-sigma modulators is 6.758 MHz.
With the DOSR and with the AOSR divider, it is possible to adjust the needed audio sampling frequency f_S, which can be in the range between 6 kHz and 192 kHz.
The higher the DAC_{MODCLK} and the ADC_{MODCLK}, the DOSR and the AOSR are, the better the audio quality (see chapter 4.2).

4.1.4 Power

The basic power scheme of the TLV320AIC3254 Evaluation Module can be seen from figure 4.6:

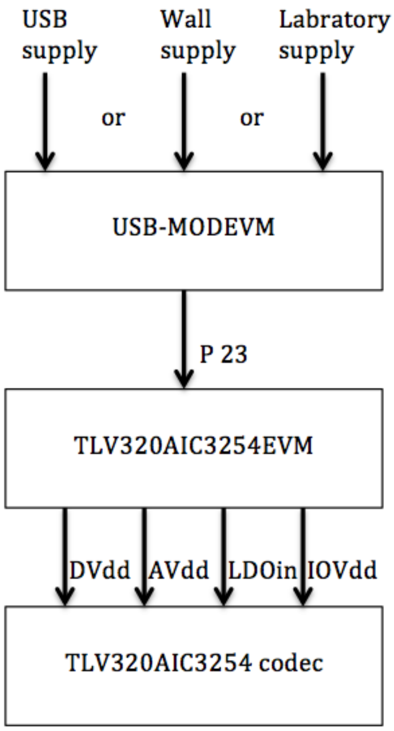

Figure 4.6: Basic power schematic

USB-MODEVM

The USB-MODEVM can be powered from several different sources:

- USB
- 6-Vdc to 1-Vdc AC/DC external wall supply
- Laboratory power supply

To change the power supply source for the USB-MODEVM JMP6 has to be configured: When powered from the USB connection, JMP6 must have a shunt from pins 1-2, if power is provided via the external wall supply, JMP6 must have a shunt from pins 2-3.There are several on-board regulators, which generate the required supply voltages.

If the laboratory power supply is used, JMP6 must not have shunts installed and the different voltages are applied to J2 (+5 VA), J3 (+5 VD), J4 (+1.8 VD) and J5 (3.3 VD). The +1.8 VD and +3.3 VD can also be generated by the onboard regulators from the +5 VD supply; to enable this configuration, the switches on SW1 need to be set to enable the regulators by placing them in the ON position.

If +1.8 VD and +3.3 VD are supplied externally, disable the onboard regulators by placing SW1 switches in the OFF position.

TLV320AIC3254EVM

The TLV320AIC3254EVM is supplied from the USB-MODEVM via P23.
There are four supply voltages: +5 VD, +5 VA, +3.3 VD and +1.8 VD. These voltages do supply the EVM onboard regulators, which generate the required voltages for the TLVAIC3254 audio codec.

TLV320AIC3254 Audio codec

The TLV320AIC3254 audio codec itself has four different power supply pins:

- **Dv**$_{dd}$**:** This pin supplies the digital core of the device. To support the full clock range, Dv$_{dd}$ has to be in the range of 1.65 V to 1.95 V.
- **AV**$_{dd}$**:** This pin supplies the analog core of the device, for example microphone inputs. The AV$_{dd}$ supply can also be generated by the internal LDO from LDO$_{in}$. The range of AV$_{dd}$ should be in the range of 1.5 V to 1.95 V, depending of the selected internal common-mode: If the common mode is 0.9V, AV_{dd} should be above 1.8 V, if a common mode of 0.75 V, AV_{dd} could be below 1,8 V.
- **LDO**$_{in}$ **/ HPV**$_{dd}$: The LDO$_{in}$ pin supplies to the internal LDOs as well as the analog-output amplifiers of the device. The LDO$_{in}$ voltage can range from 1.9 V to 3.6 V.
- **IOV**$_{dd}$**:** The IOVDD pin supplies the digital IO cells of the device like the digital microphone input or the GPIO pin. The voltage of IOVDD can range from 1.1 V to 3.6 V and is determined by the digital IO voltage of the rest of the system.

4.1.5 Control interfaces

The TLV320AIC3254 is programmed by writing to registers that can be accessed by using the I^2C or the SPI protocol. The protocol to use can be selected by switching the hardware pin on top of the evaluation board. Because in this work only I^2C will be used, the SPI protocol will not be discussed.

I^2C is a serial data bus supporting multiple devices on a single bus, using two bidirectional open-drain lines, Serial Data Line (SDA) and trial Clock (SCL). Communication on the I^2C always takes two devices, a master and a slave. Both master and slave can read and write, but the slave can only do so under the direction of the master.
In the evaluation module, the TAS1020B streaming audio USB controller on the USB-MODEVM motherboard acts as a master, the TLV320AIC3254 audio codec as a slave.

For evaluation purpose, TI is offering the AIC3254 Control Software to control the communication between these two devices. The AIC3254 Control Software is running on a PC and provides easy access to the functions of the audio codec.
The AIC3254 Control Software also offers a Command-Line Interface panel, providing to program the registers of the TLV320AIC audio codec using a simple scripting language, as seen in figure 4.7:

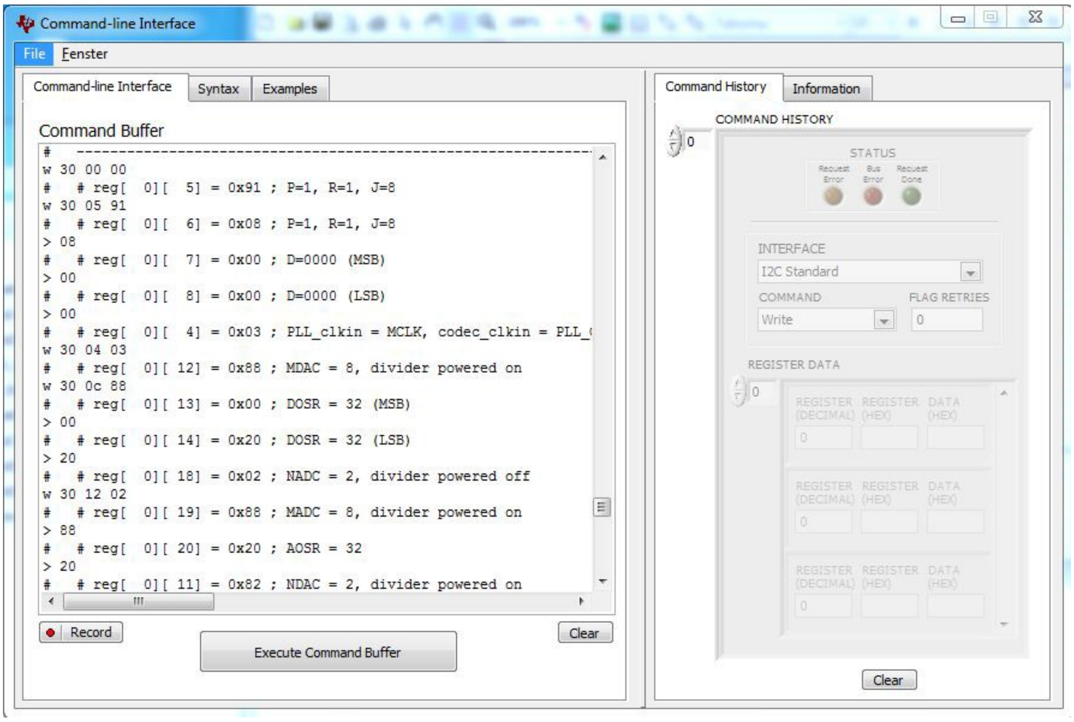

Figure 4.7: Command Line Interface

The first character of each line in a script file is the command:

i Set interface bus to use

r Read from the serial control bus

w Write to the serial control bus

> Extend repeated write commands to lines below a **w**

\# Comment

b Break

d Delay

f Wait for flag

The first command in every script sets the interface to use for the commands to follow:

i2cstd Standard mode I^2C bus

i2cfast Fast mode I^2C bus

spi8 SPI bus with 8-bit register addressing **spi16**

SPI bus with 16-bit register addressing **gpio** Use

the USB-MODEVM GPIO capability

So for example, if a standard mode I^2C bus is to be used, the script starts with:

i i2cstd

The syntax for a function **f** is the following:

f [i2c adress] [register] [D7] [D6] [D5] [D4] [D3] [D2] [D1] [D0]

Where "i2c address" and "register" are in hexadecimal format and "D7" through "D0" are in binary format.
A detailed register map can be found in the TLV320AIC3254 reference guide [13]. There are hundreds of registers, which seem to be overwhelming for an untrained user. Actually, most of them are not needed for typical audio applications.
For example, to power up the NDAC divider and set it to the value "2", the following instructions are needed:

5.2.11 Page 0 / Register 11: Clock Setting Register 6, NDAC Values - 0x00 / 0x0B (P0_R11)

BIT	READ/ WRITE	RESET VALUE	DESCRIPTION
D7	R/W	0	NDAC Divider Power Control 0: NDAC divider powered down 1: NDAC divider powered up
D6-D0	R/W	000 0001	NDAC Value 000 0000: NDAC = 128 000 0001: NDAC = 1 000 0010: NDAC = 2 ... 111 1110: NDAC = 126 111 1111: NDAC = 127 Note: Please check the clock frequency requirements in the Overview section

Figure 4.8: NDAC register map (Figure; [13])

First, Page 0 has to be selected; this is done by the following command:

w 30 00 00

(The first 255 of D7-D0 are selecting the page for the next instructions. To select Page 1, the command would be "w 30 00 01")

To power up the NDAC, "D7" hast to be set to "1" and "D6" to "D0" to 000 0010; this results in a binary digit of "1 0 0 0 0 0 1 0", which is "00x130" in hexadecimal format. The register number is "11" which is "0B" in hexadecimal format. Lets say the I^2C Address is "0x30", then the command to set the NDAC value to "2" is:

w 30 0B 82

Script files are generated by using the command panel window or a simple text editor.

4.1.6 Signal processing

On the TLV320AIC3254, there are four signal processing cores. Two cores are coupled to the ADC channel (recording path) and two of them are coupled to the DAC channel (playback path). Each channel can be programmed individually.
The TLV320AIC3254 offers a range of processing blocks, which implement different possibilities of signal processing, along with decimation and interpolation filtering.
The different processing blocks give users the choice of how much and what type of signal processing they may use and which decimation filter will be used. The programming of the processing blocks is done via register writing.

The processing blocks offer the following signal processing features:

- First-order IIR filter
- Biquad filters
- Variable-tap FIR filter
- AGC (automatic gain control)

The IIR, Biquad and FIR filters have user-programmable coefficients.
The next figure 4.9 shows a table of the different process blocks for the ADC channel. The resource class column (RC) gives an approximate indication of power consumption.

Processing Blocks	Channel	Decimation Filter	1st Order IIR Available	Number BiQuads	FIR	Required AOSR Value	Resource Class
PRB_R1 (1)	Stereo	A	Yes	0	No	128,64	6
PRB_R2	Stereo	A	Yes	5	No	128,64	8
PRB_R3	Stereo	A	Yes	0	25-Tap	128,64	8
PRB_R4	Right	A	Yes	0	No	128,64	3
PRB_R5	Right	A	Yes	5	No	128,64	4
PRB_R6	Right	A	Yes	0	25-Tap	128,64	4
PRB_R7	Stereo	B	Yes	0	No	64	3
PRB_R8	Stereo	B	Yes	3	No	64	4
PRB_R9	Stereo	B	Yes	0	20-Tap	64	4
PRB_R10	Right	B	Yes	0	No	64	2
PRB_R11	Right	B	Yes	3	No	64	2
PRB_R12	Right	B	Yes	0	20-Tap	64	2
PRB_R13	Stereo	C	Yes	0	No	32	3
PRB_R14	Stereo	C	Yes	5	No	32	4
PRB_R15	Stereo	C	Yes	0	25-Tap	32	4
PRB_R16	Right	C	Yes	0	No	32	2
PRB_R17	Right	C	Yes	5	No	32	2
PRB_R18	Right	C	Yes	0	25-Tap	32	2

(1) Default

Figure 4.9: Processing blocks of the ADC channel (Figure:[13])

The user has to pick the processing blocks fitting a certain application best. The Signal chain for each processing block is documented in the Reference Guide ([13]). For example, processing block number three (PRB_R3) has the following signal chain:

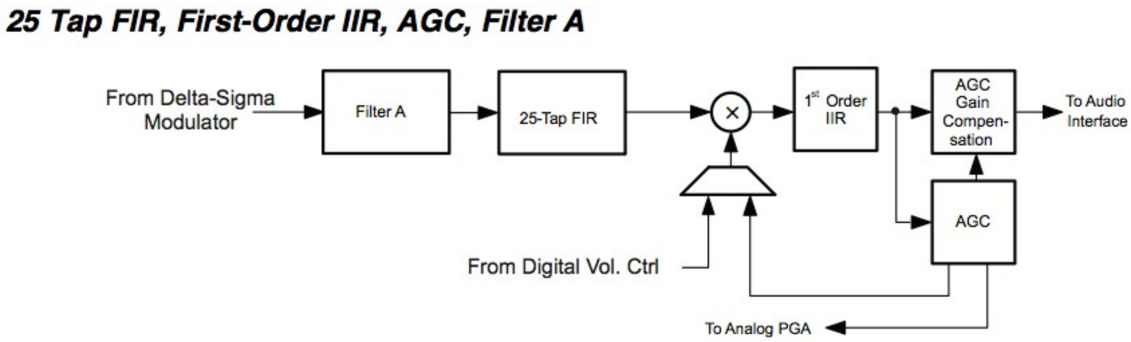

Figure 4.10: Signal chain of PRB_3 (Figure;[13])

The signal chain and the elements inside a processing block cannot be changed or re- placed. Therefore, they are not highly flexible if the application is getting more complex. Another disadvantage of using the processing blocks is that there is no possibility of deactivating the decimation and interpolation filter because they are always part of each processing block. As it will be shown in chapter 4.2 it is necessary to deactivate them for building up a properly-working ANC headphone.

Another possibility for signal processing is in the use of the miniDSPs inside the audio codec. The first core (miniDSP_A) is tightly coupled to the ADC channel; the second core (miniDSP_D) is tightly coupled to the DAC (see 4.11).

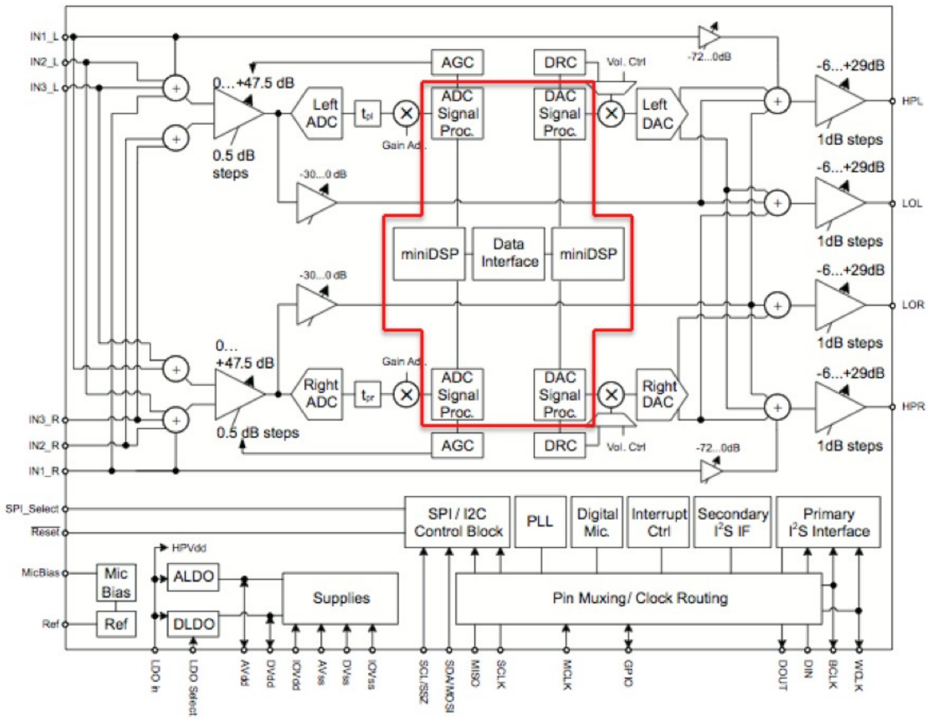

Figure 4.11: Block diagram of AIC3254 (Figure; [13])

When the audio codec is running in "miniDSP mode", the user can create his/her own signal chain, so there are no preassembled signal chains like in the "processing block mode". Also, it is possible to deactivate the decimation and interpolation filters when running in "miniDSP-mode".

To program the miniDSPs, Texas Instruments provides the comprehensive Software program "Pure Path Studio" which provides the means for creating and editing process flows. Pure Path Studio creates a script file out of the build-process flow which can be downloaded to the audio codec via the USB connection (see chapter 4.1.5).

There are three main windows in Pure Path for creating the signal flow, as shown in figure 4.12:

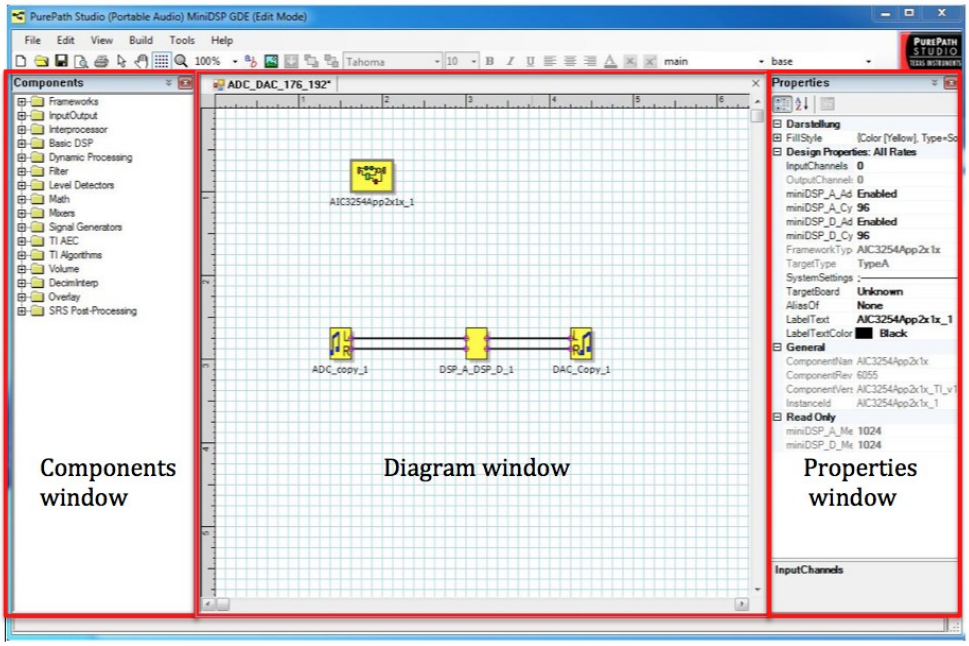

Figure 4.12: Purepath Studio

The components for building the desired signal flow are listed in the **Components window**. Components can be as small as a basic single operation, like multiplying or summing up, or more complex, like an algorithm for a bi-quad filter with a separate GUI.
The components can be placed in the **Diagram window** using the drag-and-drop functionality.
The properties of each component can be adjusted in the **Properties window**.
Each process flow needs an application framework; this framework is the basis for an audio application. Parameters like clock settings, I/O routing or power management can be adjusted in the properties window if the application framework is selected.

4.2 Performance Tests

The goal of the performance tests is to find out about the most efficient settings for the TLV320AIC3254 for building an ANC system. The most important factors are:

- Amplitude response
- Phase response
- Propagation Delay
- Power consumption
- Audio quality

4.2.1 Amplitude response / Phase response

In order to build a properly-working ANC headphone, it is important that the audio codec itself renders a linear amplitude response. If the amplitude-over-frequency is not linear, additional filtering would be needed for compensating for the amplitude response. This filtering would result in additional phase shifts which would lead to unwanted phase mismatches between the noise signal and the generated anti-noise signal.
For the phase response, it would be ideal for the audio codec not to create any phase shift itself, so to get as much phase margin as possible. Because the audio codec needs time to process the analog signal, which results in a phase shift, this cannot be achieved in a real-life system.
The faster the analog audio signal can be passed through the audio codec (including analog- to-digital conversion, signal processing, digital-to-analog conversion), the better the phase response will be in terms of building a properly-working ANC headphone. It is also important that the phase response of the system is linear phase in the audible frequency range. A linear phase means that the propagation delay is constant over the frequency, so all frequencies in the audible spectrum experience the same propagation delay.

As mentioned above, the phase response is getting "better" the faster the audio can be sent and processed through the DSP. The speed on how fast this can be achieved is heavily dependent on the used sampling frequency f_S. The higher the sampling frequency f_S, the more samples can be processed in the same amount of time.
For example, using a sampling rate of 44,1 kHz, there is a time lag of

$$\Delta t = \frac{1}{44100} = 22,6 * 10^{-6}s \tag{4.2}$$

between each sample. This means that the DSP has to wait 2.6 microseconds until getting the next sample to process. If using a 192 kHz sample rate for example, the DSP only has to wait

$$\Delta t = \frac{1}{192000} = 5,2 * 10^{-6}s \tag{4.3}$$

for the next sample to process.

To measure the amplitude and the phase response, the DSP is programmed to send the input IN3 through the DSP to the headphone output, directly, without any additional signal processing inbetween. This is the "basic" signal flow set-up to only measure the performance of the audio codec itself.
Figure 4.19 shows the Pure Path setup to achieve this routing:

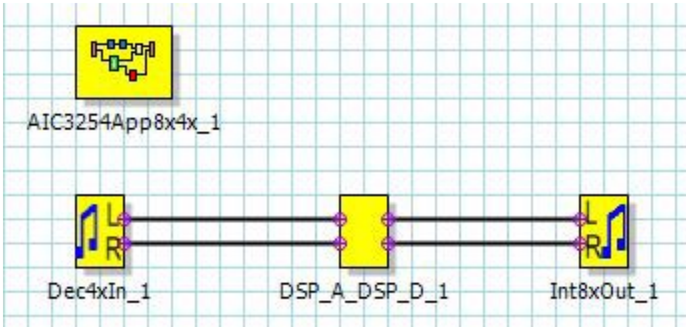

Figure 4.13: Pure Path setup

”*AIC*3254*App*8*x*4*x*_1” is the application framework needed for the process flow. ”*Dec*4*xIn*_1” is a decimation input, which uses a digital anti-aliasing filter before downsampling the delta-sigma modulator‘s output to the sampling frequency f_S.
The ”*DSP _A_DSP _D*” component simply copies the data from the DSP A to the DSP D to sent it to the component ”*Int*8*xOut*_1”, which is an interpolation output with a digital anti aliasing filter before up sampling is done by the delta-sigma modulator.

The clock setting to generate the various sample frequencies can be seen from the following table, it is the same for the ADC and the DAC channels:

Fs [kHz]	44,1	48	88,2	96	176,4	192
MCLK [MHz]	11,29	12,29	11,29	12,29	11,29	12,29
PLL out [MHz]	45,16	49,15	45,16	49,15	45,16	49,15
N	1	1	1	1	1	1
M	8	8	8	8	8	8
OSR	128	128	64	64	32	32

The measurement was done with the Audio Precision 2700 Series measurement system. The generator of the Audio Precision played back a sine sweep signal at -15 dBV while the analyzer was connected to the headphone output of the evaluation board.

In figure 4.14, the effect of different sample rates on the amplitude and phase responses can be seen:

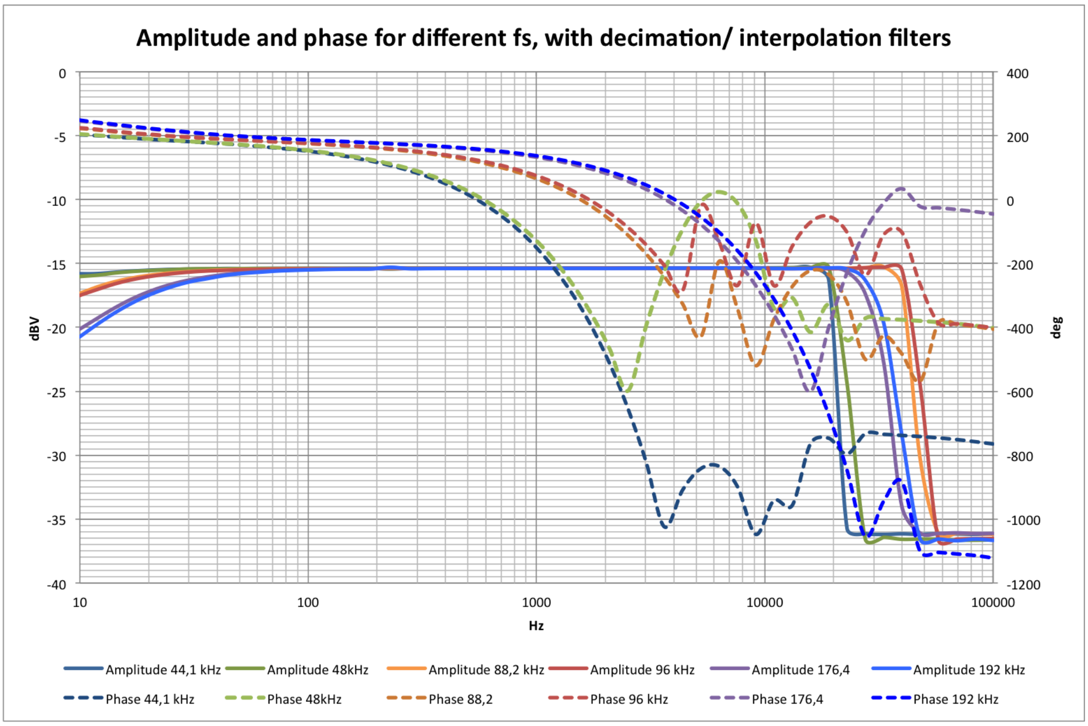

Figure 4.14: Different sample rates using interpolation inputs and decimation outputs

The amplitude responses are linear within the audible frequency range from 20 Hz to 20 kHz. At higher frequencies, the digital anti-aliasing filters are damping the audio signal above their cut-off frequency. We can observe that the higher the sample rate f_s is, the higher is the cut-off frequency of the filters (see chapter 4.1.2).

The phase response also depends on the sampling rate f_s. Because the microphone input operates as an inverting amplifier, the phase is shifted by 180 degrees.
The higher the sampling frequency, f_s the fewer phase shifts occur at a certain frequency, because the DSP is able to calculate faster.
Such a phase response is not very useful when it comes to building up a properly-working ANC headphone. Even for the best sampling rate of $f_s = 192\,kHz$, the phase has shifted 180° at a frequency of 4 kHz. This means that there will be a phase mismatch of 100%, resulting in positive feedback if this anti-noise signal is mixed together with the original noise signal.
Keeping in mind that this measurement is done without any ANC filters in the signal chain (which is mandatory for an ANC system), this is not a good starting point

Because the digital anti-aliasing filters are very steep, it can be assumed that the phase distortion is caused by the digital anti-aliasing filters. Without the filters the resulting phase response should be much better suited for use in ANC headphones.

It is possible to deactivate the decimation and interpolation filters by using the components ”*ADCcopy*” and ”*DACcopy*” instead of the decimation input and the interpolation output in Pure Path studio:

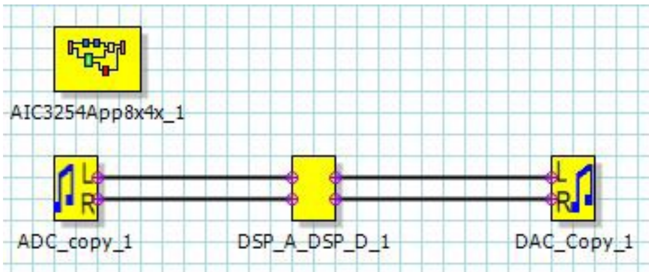

Figure 4.15: Pure Path setup

Again, the amplitude and phase response is measured with the Audio Precision 2700 measurement system, the result is shows in figure 4.16:

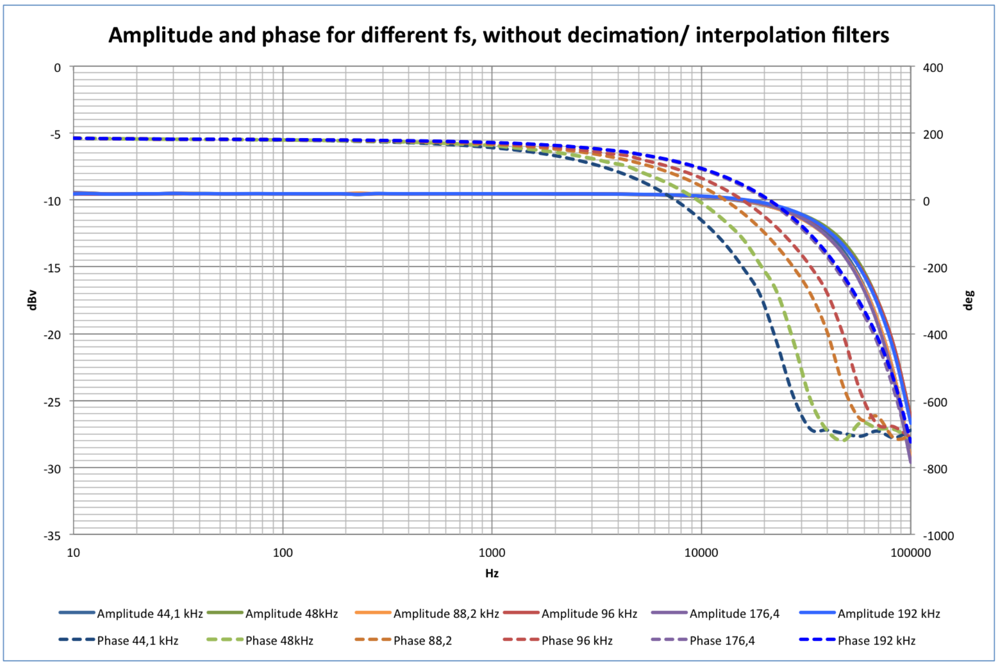

Figure 4.16: Different sample rates using the interpolation inputs and decimation outputs

As expected, the phase response without the anti-aliasing filter is now much better than the phase response when using the anti-aliasing filter. For a sampling rate of f_s = 192 *kHz*, we get a phase shift of about 30˚ at 4000 Hz. Compared to the measurement with the anti- aliasing filters, this represents a difference of 150˚.

***NOTE**: It is not possible to see any aliasing-effects from figure 4.16 although no anti- aliasing filters have been used. This is because the Audio Precision measurement system only analyses the frequency which is actually generated during the swept sine analysis.*

In order to see the aliasing effects, all frequencies have to be analyzed at the same time (for example by using an FFT), which is not the case with this measurement method.

There is still a phase shift because of the time needed to calculate the signal, and because there is an analog reconstruction filter at the output of the DA converter.
The behavior of this filter can be detected if we look at the amplitude response. Because it is an analog filter, the response is the same for all sampling rates.
As mentioned above, it is also important for the phase response to be linear within the audible frequency range so to achieve a constant propagation delay over all frequencies. In figure 4.14, the x-axis is logarithmically scaled, in the next figure 4.17, the phase response is plotted using a linear scale:

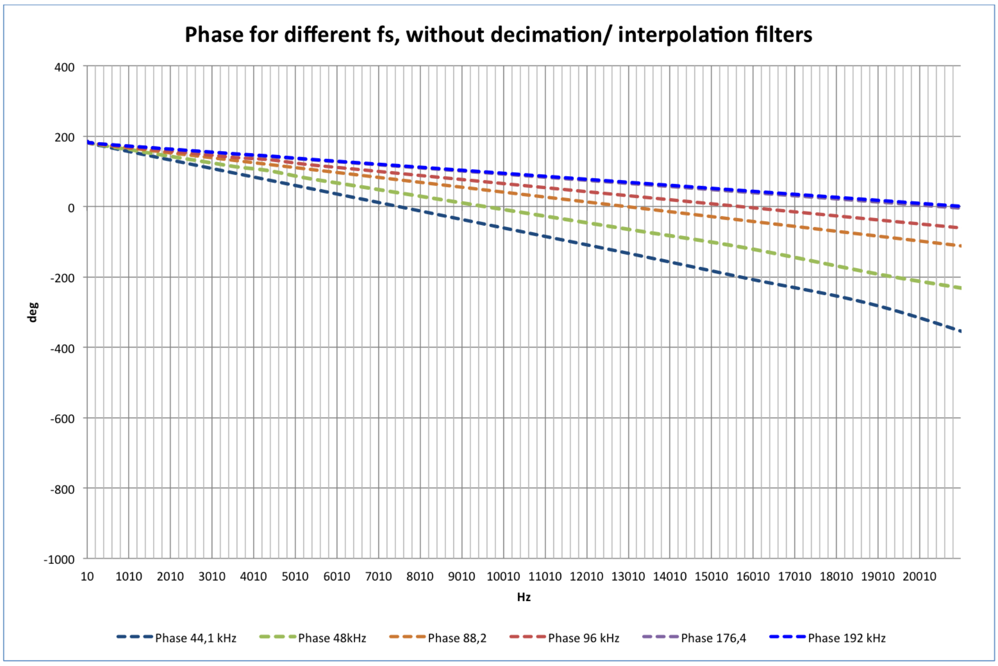

Figure 4.17: Phase of different f_s using a linear scale

As we can see, the phase response is linear within the audible frequency range from 20 Hz to 20000 Hz. This means the audio codec causes a propagation delay which is frequency independent but sampling frequency f_s dependent.

4.2.2 Propagation delay

The propagation delay describes the amount of time for an analog audio signal to travel from the input to the output of the codec. This round-trip includes analog-to-digital conversion, digital signal processing and digital-to-analog conversion at the output of the codec. To build an efficient ANC system, it is important for the propagation delay to be as small as possible.
We have seen that the phase response is linear within the audible frequency range and, so the propagation delay is the same for all frequencies; thus we can measure the propagation delay at one frequency.
To achieve this, a square signal with a frequency of 8000 Hz was sent through the TLV320AIC3254 audio codec. The propagation delay between the input and the output is then measured using an Oscilloscope.
The measured delay between input (orange) and output (green) is $25 * 10^{-6}$s without the anti-aliasing filters, as shown in figure 4.18:

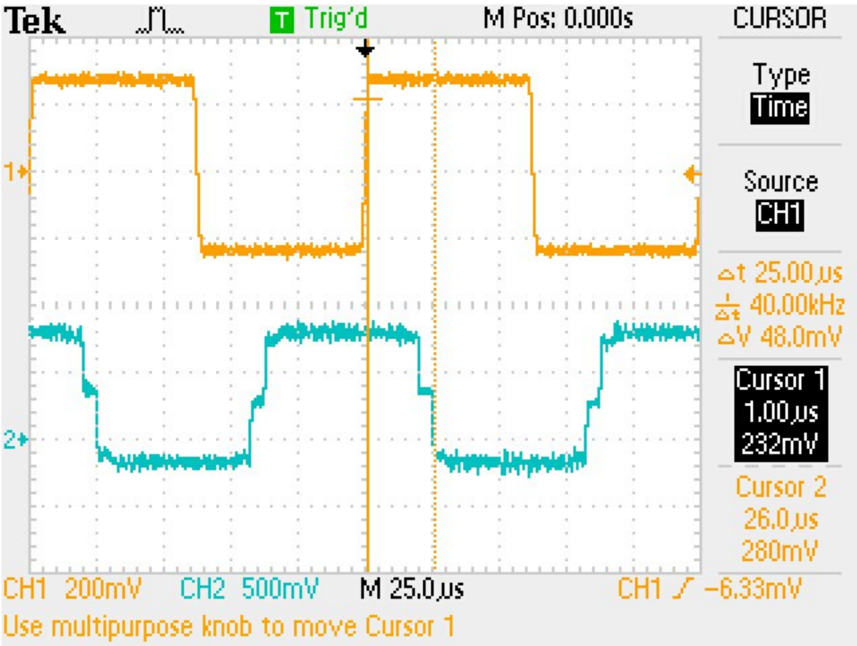

Figure 4.18: Propagation delay

4.2.3 Power consumption

As mentioned in chapter 4.1.4, the codec is supplied by the Dv_{dd}, AV_{dd}, IOV_{dd} and $LDO_{in}/HP\ V_{ddt}$ pins. While the AV_{dd} and $LDO_{in}/HP\ V_{ddt}$ pins are supplying the analog parts of the codec, like microphone input amplifier or headphone output amplifier, the Dv_{dd} and IOV_{dd} pins supply the digital parts of the codec.

To measure power consumption, there are pins and jumpers on the evaluation board to have access to measure voltages and currents. For the power measurements, the evaluation board is powered via laboratory supply (see chapter 4.1.4).
The full-chip common mode is set to 0.9 V, the voltages of the four supply pins of the TLV320AIC3254 audio codec can be measured through separate pins on the evaluation board. The onboard LDO which supplies AV_{dd} is switched off, so every pin is supplied separately.

The following voltages can be measured at the pins:

Dv_{dd}: 1.8V

AV_{dd}: 1.79V

IOV_{dd}: 3.3 V

$LDO_{in}/HP\ V_{ddt}$: 3.3V

The headphone output is supplied via the $LDO_{in}/HP\ V_{ddt}$ with the headphone amplifier common-mode set to 3.6 V. All gains of the input and output amplifiers are set to 0 dB for the measurements.

Different sample rates

The power consumption depends on the sampling frequency f_s used. To get a basic idea of the power consumption in dependence of the sampling rate, the basic signal flow with decimation and interpolation filters was constructed in pure path (same as in chapter 4.2.1):

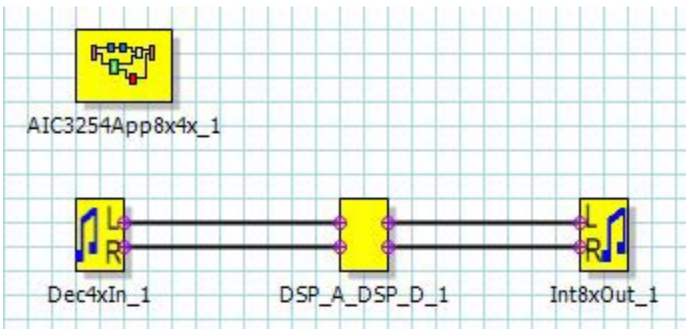

Figure 4.19: Pure Path setup

The clock settings below were used for achieving the different sample rates:

Fs [kHz]	44,1	48	88,2	96	176,4	192
MCLK [MHz]	11,29	12,29	11,29	12,29	11,29	12,29
PLL out [MHz]	45,16	49,15	45,16	49,15	45,16	49,15
N	1	1	1	1	1	1
M	8	8	8	8	8	8
OSR	128	128	64	64	32	32

The cycles (see chapter 4.1.3) are the same (256 cycles for both ADC and DAC) for all sampling frequencies f_S, to compare the power consumption independently of the computing power of the codec.

Figure 4.20 shows the power consumption of all four supplies with different sample rates f_s :

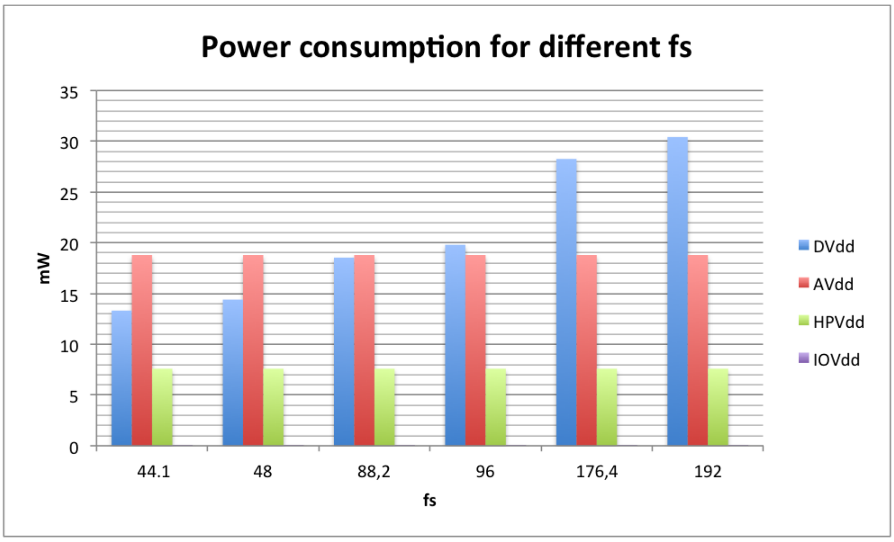

Figure 4.20: Power consumption with different sample rates (with interpolation/decimation filters)

Only the power consumption of the Dv_{dd} pin is changing with different sampling rates. The higher the sampling rate f_S, the more power is needed, because the DSP has to compute faster for high sampling frequencies f_S. At the same time the propagation delay decreases.

The next figure shows the total power consumption of the codec for different sample rates:

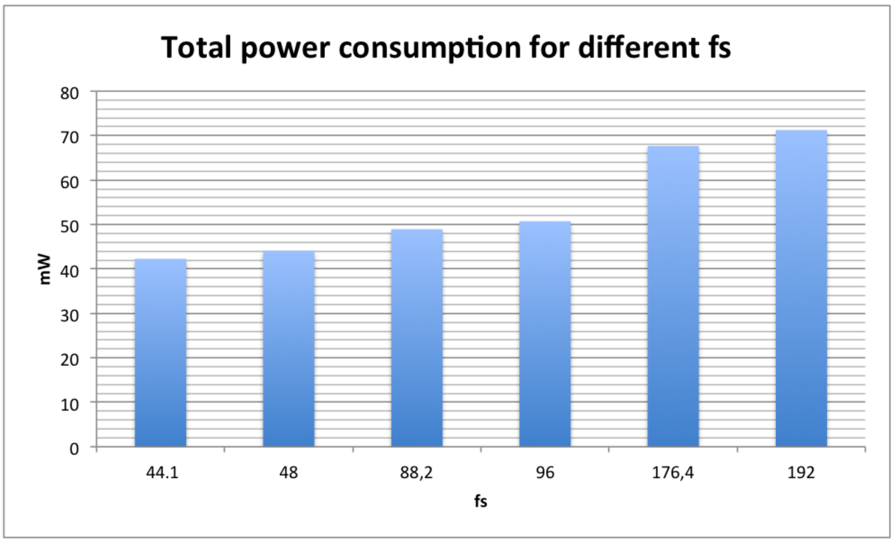

Figure 4.21: Total power consumption with different sample rates (with decimation/ interpolation filters)

As we have seen from chapter 4.2.1, the phase margin increases for higher sampling rates f_S. On the other hand also the power consumption increases with higher sampling frequencies f_S. So, there has to be a trade-off between phase response and power consumption when choosing the sampling rate f_S for building an efficient ANC headphone.

As seen from chapter 4.2.1, it is possible to deactivate the decimation and interpolation filters to get a better phase response for an ANC system. Therefore, the power consumption was also measured without the decimation and interpolation filters. All other settings, like clock settings or gain setting were the same.

Figure 4.22 shows the power consumption of the codec without decimation and interpolation filter:

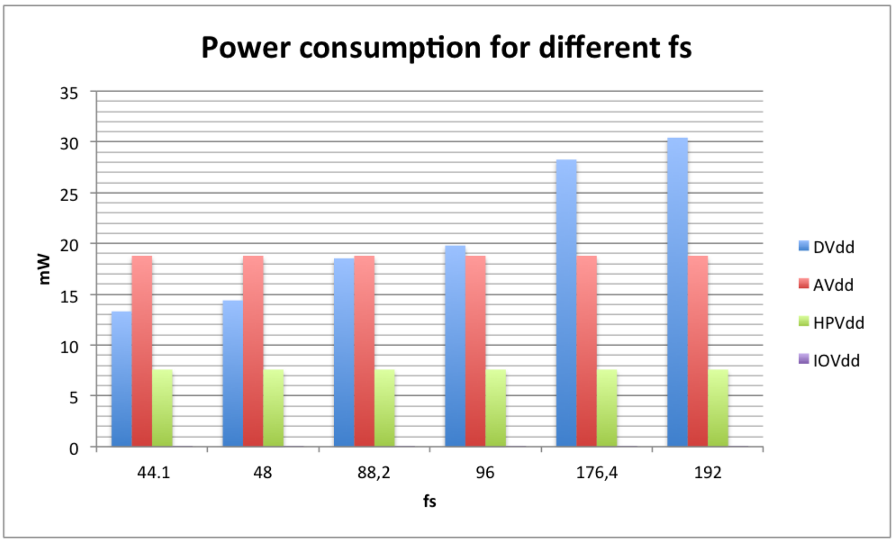

Figure 4.22: Power consumption with different sample rates (without interpolation/decimation filters)

The comparison (see figure 4.23) between the total power consumption with and without decimation/ interpolation filters shows that the power consumption is lower when we are not using the decimation/ interpolation filters. This is because the decimation/ interpolation filters are done in the digital domain and so the codec needs more processing power.

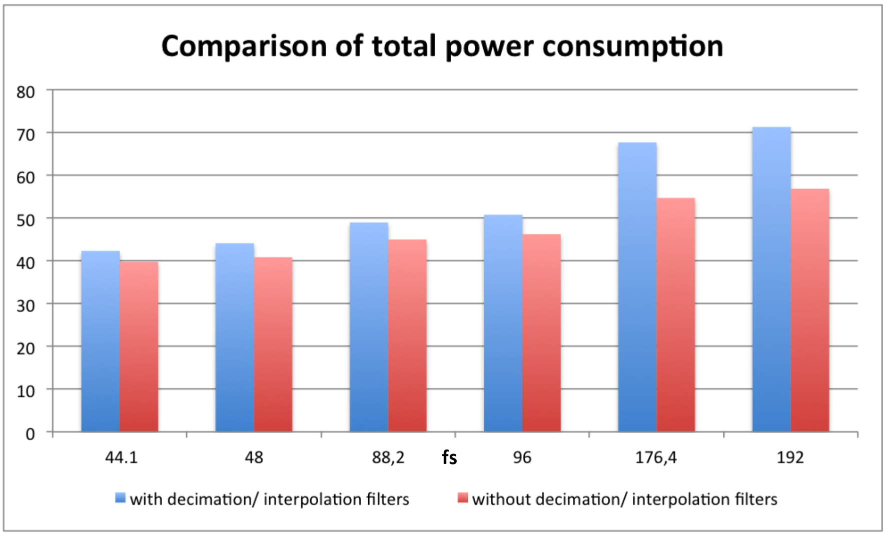

Figure 4.23: Comparison of power consumption

Different cycles

As described in chapter 4.1.3, the processing power of the miniDSP depends on the allocated cycles. The more processing power is needed for a certain task, the more cycles have to be allocated to the miniDSP.

Figure 4.24 shows the power consumption of Dv_{dd} for different cycles at a sample rate of 192 kHz. The measurement is done without using the decimation/ interpolation filters; the following clock settings were used:

MCLK [MHz]	12,29	12,29	12,29	12,29	12,29	12,29
PLL out [MHz]	49,15	49,15	49,15	49,15	49,15	49,15
N	1	1	1	1	1	1
ADC DAC CLK	49,1536	49,1536	49,1536	49,1536	49,1536	49,1536
M	8	8	8	8	8	8
ADC DAC MOD CLK	6144,2	6144,2	6144,2	6144,2	6144,2	6144,2
OSR	32	32	32	32	32	32
Fs [kHz]	192	192	192	192	192	192
cycles	96	128	160	192	224	256

The power consumption of Dv_{dd} rises with the number of cycles that are allocated.

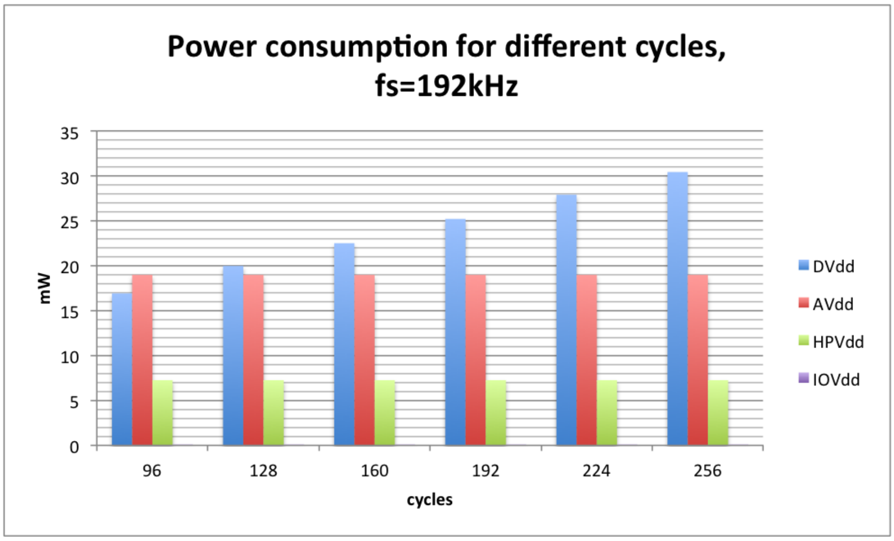

Figure 4.24: Power consumption for Dvdd for different cycles, fs=192kHz

The next tradeoff, which has to be made, is about cycles versus power consumption. The more cycles are needed for a certain task, the higher the power consumption. In Pure Path, the maximum number of cycles can be set in the properties window, if the framework is selected. If the application needs more cycles, Pure Path will not build the code.

The ADC_{CLK}/DAC_{CLK} clock frequencies were the same for all cycles in the measurement of figure 4.24.
As explained in chapter 4.1.3, the minimum ADC_{CLK}/DAC_{CLK} clock frequency must be

$$(ADC_{CLK}/DAC_{CLK})_{minimum} = f_s * \; cycles \tag{4.4}$$

The following table shows the minimum ADC_{CLK}/DAC_{CLK} that were calculated:

Fs [kHz]	192	192	192	192	192	192
cycles	96	128	160	192	224	256
ADC DAC min CLK [MHz]	18,4326	24,576	30,721	36,8652	43,0094	49,1536

In order to generate the desired ADC_{CLK}/DAC_{CLK} clock frequency, the PLL can be configured to generating the required frequency, while setting NDAC/ NADC to 1:

MCLK [MHz]	12,29	12,29	12,29	12,29	12,29	12,29
PLL out [MHz]	18,432	24,576	30,721	36,8652	43,0094	49,1536
N	1	1	1	1	1	1
ADC DAC CLK [MHz]	18,432	24,576	30,721	36,8652	43,0094	49,1536
M	8	8	8	8	8	8
ADC DAC MOD CLK [MHz]	6144,2	6144,2	6144,2	6144,2	6144,2	6144,2
OSR	32	32	32	32	32	32
Fs [kHz]	192	192	192	192	192	192
cycles	96	128	160	192	224	256

In the following figure, the power consumption of Dv_{dd} was measured again, but now using the minimum ADC_{CLK}/DAC_{CLK} clock frequency:

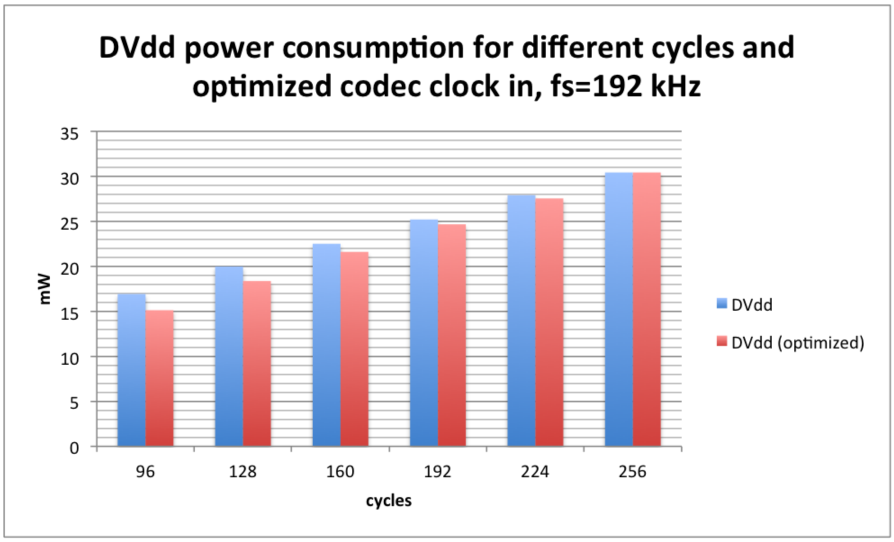

Figure 4.25: Power consumption for Dv_{dd} for different cycles and optimized clock, f_s=192kHz

If power consumption should be at a minimum, it is important to run the codec at the lowest clock settings possible. If running the clocks on a higher rate than needed results in a waste of power.

Different oversampling rates (OSR)

The last clock setting is the AD/DA_{MODCLK}, which is the clock for the Delta-Sigma modulator. Depending on the oversampling-ratio (OSR) needed, the clock can be set with the help of the MDAC/ MADC divider.
So far, we used the AD/DA_{MODCLK} at 6144.2 MHz, which is the maximum frequency of the Δ-Σ modulator. At 192 kHz, this results in an oversampling factor of

$$OSR = \frac{MOD_{CLK}}{f_s} = \frac{6144.2}{192} = 32 \tag{4.5}$$

The higher the OSR should be at a certain sample frequency f_s, the higher the clock of AD/DA_{MODCLK} has to be. It can be assumed that with a low OSR value (which means a possible AD/DA_{MODCLK}), the power consumption will decrease.
Also it's very likely that with a lower OSR more noise will be shifted into the audible frequency range, as described in chapter 3.2.3. Therefore, the audio quality will decrease with decreasing OSR (see next chapter 4.2.4).

The following settings were used for generating the different *ADC/DAC$_{MODCLK}$* clock frequencies:

MCLK [MHz]	12,29	12,29	12,29	12,29	12,29	12,29
PLL out [MHz]	18,432	18,432	18,432	18,432	18,432	18,432
N	1	1	1	1	1	1
ADC DAC CLK [MHz]	18,432	18,432	18,432	18,432	18,432	18,432
M	3	6	12	24	48	96
ADC DAC MOD CLK [MHz]	6144,2	3072,1	1536,05	768,025	384,012	192
OSR	32	16	8	4	2	1
Fs [kHz]	192	192	192	192	192	192
cycles	96	96	96	96	96	96

The effect on the power consumption on Dv_{dd} shows in figure 4.26:

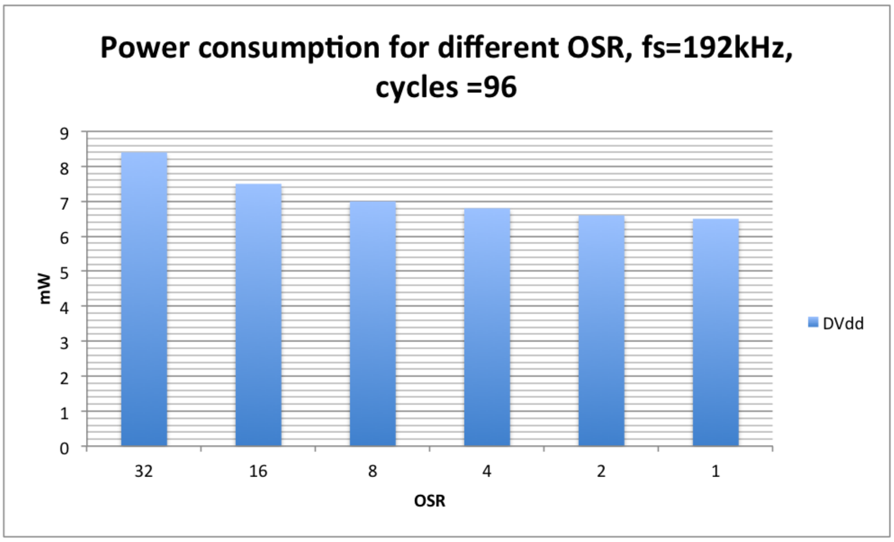

Figure 4.26: Power consumption for Dv_{dd} for different OSR, f_s =192kHz

The power consumption decreases with a lower OSR because the frequency of the *ADC/DAC$_{MODCLK}$* is also decreasing with lower oversample ratios. Therefore, power can be saved at the cost of decreased audio quality as shown in the next chapter 4.2.4.

4.2.4 Audio quality

In order to get an idea of the audio quality in dependence from the oversampling ratio, the noise floor and the total harmonic distortion and noise (THD+N) of the audio codec is being measured. The clock settings are the same as those used in the previous chapter, so the sample rate is 192 kHz for all measurements.
The noise floor has been measured by connecting the headphone output to the input of the Audio Precision 2700 measurement system. All inputs of the audio codec have been shortened to avoid unwanted incoming noise; the gain of the headphone amplifier is 0 dB.

Figure 4.27 shows the result:

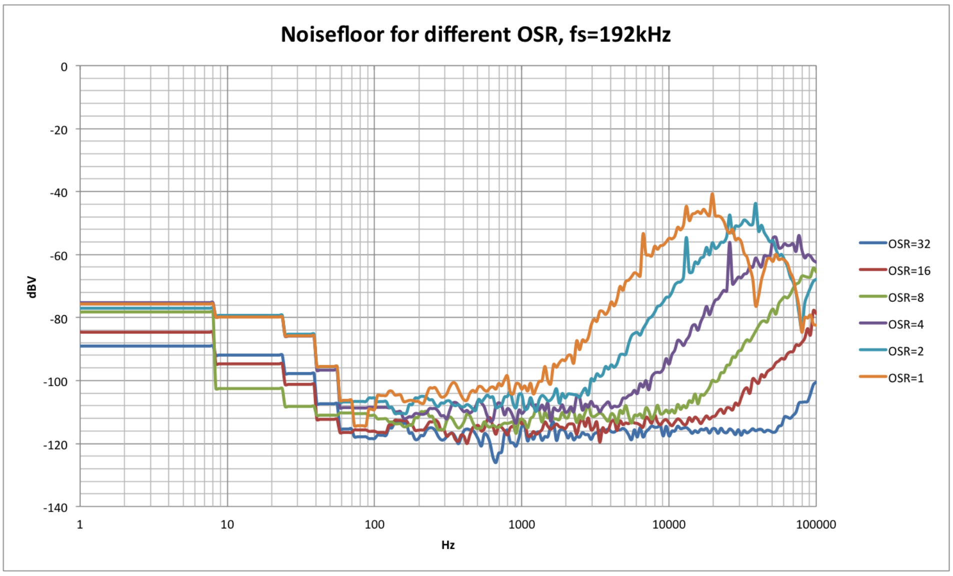

Figure 4.27: Noise floor for different OSR, f_s =192kHz

The noise is increasing with decreasing oversample ratio, as mentioned earlier in chapter 3.2.3. To measure the total harmonic distortion and noise (THD+N), the microphone input (0 dB gain) and the headphone output (0 dB gain) were connected to the Audio Precision measurement system.

The harmonic distortion is increasing when lowering the oversampling ratio too as shown in figure 4.28:

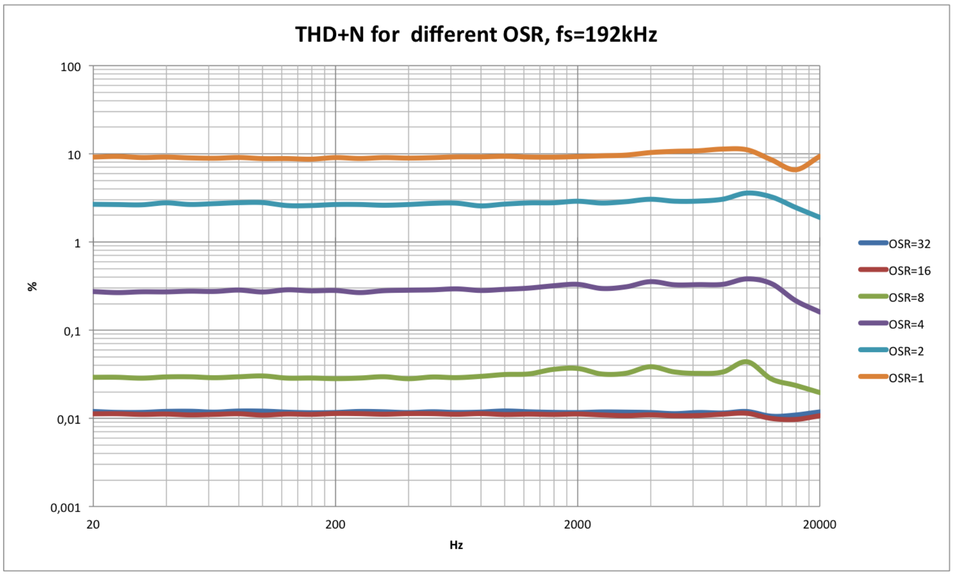

Figure 4.28: Total harmonic distortion for different OSR, $f_s = 192kHz$

It is important to pick high oversample ratios to guarantee the best possible audio quality for the ANC headphone.

4.2.5 Conclusion

When building an ANC system, a balance between the key parameters of the DSP has to be found.

- In terms of phase response and propagation delay, higher sample rates will lead to better results. The higher the sample rate, the higher the power consumption of the audio codec is. So the better the ANC performance should be, the more power is needed.
- The next trade-off is computing power: The more cycles are allocated to the DSP, the more computing power is available. Thus, for complex computations like filtering, more cycles are needed. High cycle values again result in higher power consumption, so it is important to keep the complexity of computation as low as possible.
- Another compromise has to be found in terms of audio quality: The higher the oversampling ratio, the higher the audio quality is. This also results in a higher power consumption of the device.

The trade-offs between the factors to build up a properly-working system can be seen in the following table:

	Power consumption	Propagation delay	Processing power	Audio quality
Fs ↑	↑	↓	↓	–
Cycles ↑	↑	–	↑	–
OSR↑	↑	–	–	↑

4.3 Digital Setup

4.3.1 Music equalization

To be able to tune the sound of the headphone when listening to music, music equalization is necessary. Typically, an equalizer with 5 bands is enough to achieve a satisfactory result. For music equalization, high sample rates are not necessary, because there is no need for real-time processing. Therefore, a sample rate of 44.1 kHz will be sufficient.

For inputs and outputs, there are no interpolation / decimation filters used to lower the power consumption. Because music content is typically band-limited, there is no need for using these filters to prevent aliasing.

There are two IIR biquads with 5 fully-parametric bands for the left and the right side to tune the sound of the headphone. The biquads are inserted into the ADC channel, so their computation is coupled to the ADC processing blocks.

The Oversampling ratio should be as high as possible to provide the highest audio quality possible. The maximum OSR for 44.1kHz is 128, so both the ADC_Mod_{CLK} and the DAC_Mod_{CLK} should have a clock frequency of $44.1 \cdot 128 = 5644.8$ kHz.

For the ADC channel, the minimum value of cycles, which is working to compute both of the Biquads is 256 for 44.1 kHz sampling rate. Therefore, the required minimum ADC_{CLK} is $44.1 \cdot 256 = 11289.6$ kHz. For the DAC, there is no further processing necessary, the minimum value of cycles is 8.

The next figure 4.29 shows the signal flow in the Pure Path project for the music equalization setup:

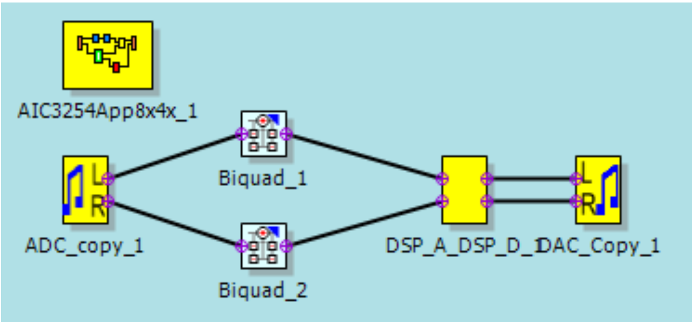

Figure 4.29: Pure Path signal flow for music equalization setup

The divider settings to generated the required clock frequencies for the audio codec can be seen from the next table:

MCKL	11.2896 MHz		
PLL	OFF		
Codec_Clkin	11.2896 MHz		
NDAC	2	NADC	1
DAC_CLK	5644.8 kHz	ADC_CLK	11289.6 kHz
ADC Cycles	256	DAC_Cycles	8
MDAC	1	MADC	2
DAC_Mod_Clk	5644.8 kHz	ADC_Mod_Clk	5644.8 kHz
DOSR	128	AOSR	128
DAC_Fs	44.1 kHz	ADC_Fs	44.1 kHz

The power consumption when using this setup can be seen from the next figure 4.30:

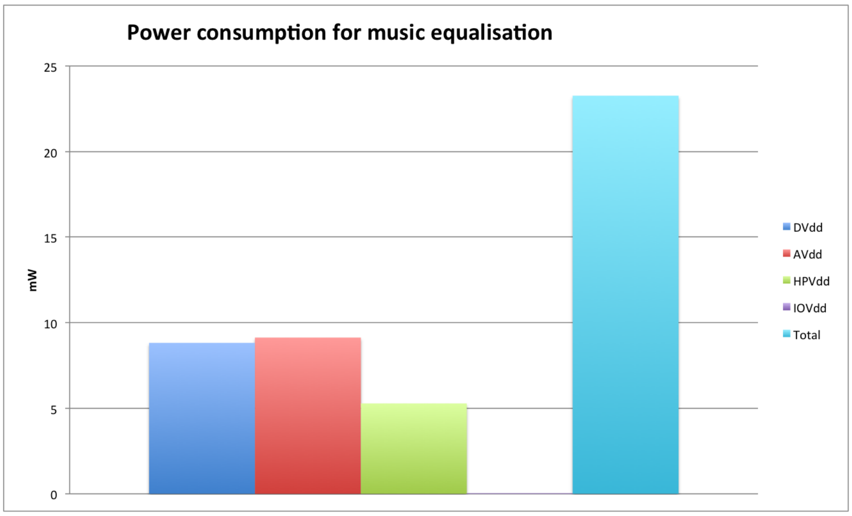

Figure 4.30: Power consumption for music equalization Setup

The total power consumption for music equalization is about 23 mW.

4.3.2 ANC system 1 (Master)

In contrast to music equalization, high sample rates are necessary for building a properly-working ANC system to achieve as much phase margin as possible. When looking at figure 4.16, apparently there is no significant difference in the phase response between 176.5 kHz and 192 kHz, but there is a difference in power consumption (see figure 4.22).

So, the choice for the sample rate is 176.4 kHz because it is the best trade-off between propagation delay and power consumption. Because the TLC320AIC audio codec has only one stereo ADC, two audio codecs are necessary to build a hybrid ANC system.

There are four mono microphone inputs:

- Feed-forward left
- Feed-forward right
- Feed-back left
- Feed-back right

For each side of the headphone, there is one TLC320AIC3254 audio codec with two microphone inputs, one for the feed-forward microphone (left in) and one for the feed-back microphone (right in). The loudspeaker of the headphone is connected to the left side of the headphone amplifier output of each audio codec.

The required digital signal flow can be seen from the following figure 4.31:

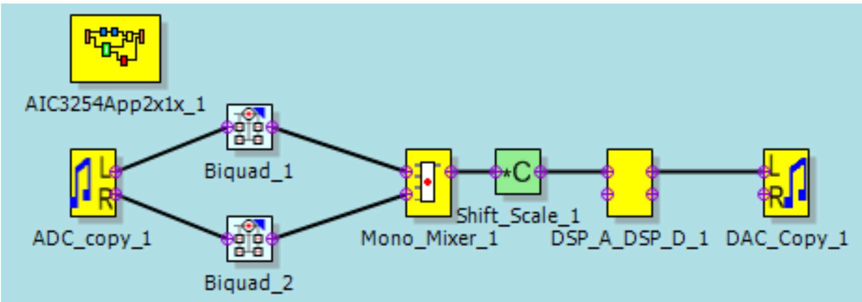

Figure 4.31: Digital signal flow

Again, the OSR should be as high as possible to provide the best audio quality. The maximum OSR for 176.4 kHz is 32, so both the ADC_Mod_{CLK} and the DAC_Mod_{CLK} should have a clock frequency of $176.4 \cdot 32 = 5644.8$ kHz

Each microphone input has its own IIR biquad filter to generate the ANC filter curves. "Biquad_1" is the feed-forward filter; "Biquad_2" is the feed-back filter.
The two biquads are placed on the ADC side, so the computation of the filtering is coupled to the ADC processing blocks. For the ANC filter, a biquad instance with 7 bands for each the feed-forward and the feed-back path is usually enough.
The minimum ADC cycle value, which is needed for computation, is 96. Therefore, the minimum ADC_{CLK} clock frequency needed is $176,4 \cdot 96 = 16934,4$ kHz.
On the DAC side, there is no additional computation required. The lowest cycle value, which works, is 8 for the DAC channel.

After the biquads, there is a "Mono_Mixer_1" instance, which mixes the two inputs to one mono output. The "Shift_Scale_1" instance is used to scale the amplitude of the digital signal, which is then sent to the left side of headphone output.

The divider settings to implement the needed clock frequencies can be seen from the next table:

MCKL	11.2896 MHz		
PLL	1x9.0000/6		
Codec_Clkin	16.9344 MHz		
NDAC	3	NADC	1
DAC_CLK	5644.8 kHz	ADC_CLK	16934.4 kHz
ADC Cycles	96	DAC_Cycles	8
MDAC	1	MADC	3
DAC_Mod_Clk	5644.8 kHz	ADC_Mod_Clk	5644.8 kHz
DOSR	32	AOSR	32
DAC_Fs	176.4 kHz	ADC_Fs	176.4 kHz

The power consumption when using this setup can be seen from the next figure 4.32:

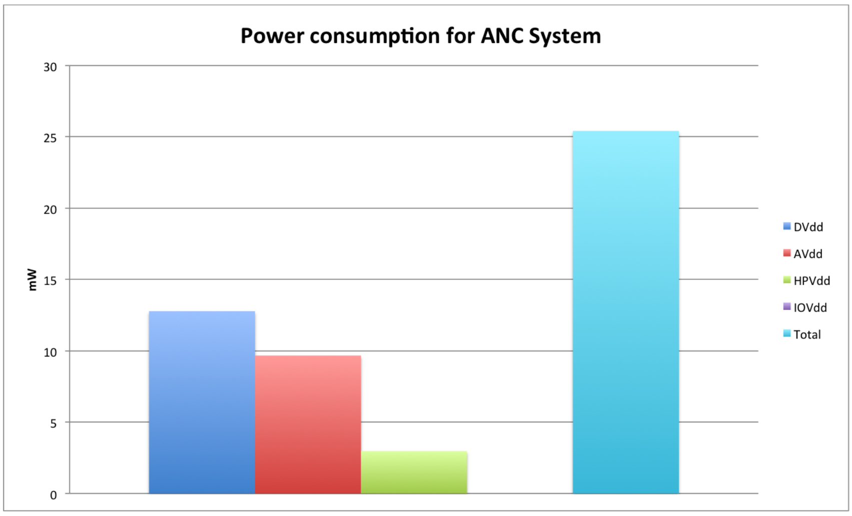

Figure 4.32: Power consumption for ANC Standalone Setup

The total power consumption for the ANC system for one headphone side is about 25 mW. This is a little more than for the music equalization setup, although the sample rate has been increased. It has to be kept in mind that one side of the headphone amp has been powered down for the ANC setup, saving about 3 mW of total power.

4.3.3 ANC system 2 (Master-Slave)

Another way to go, which has been tested, is about generating the required clock frequency only using the PLL of one audio codec (master). This clock can then be sent to the other audio codec (slave) using the GPIO pin and a common coaxial cable. Because of this, the PLL of the slave audio codec can be powered down to reduce power consumption.

The required signal flow is the same as before, the clock setting for the master audio codec, too. The only difference for the slave audio codec is that the PLL is powered down because the audio codec directly receives the maximum required clock frequency of 16.9344 MHz.

The power consumption of both audio codecs can be seen from the next two figures 4.33 and 4.34:

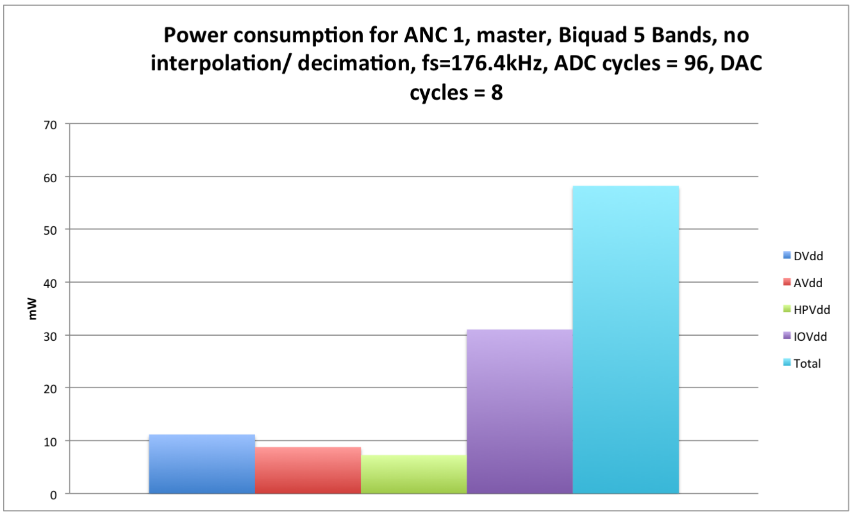

Figure 4.33: Power consumption for ANC Master Setup

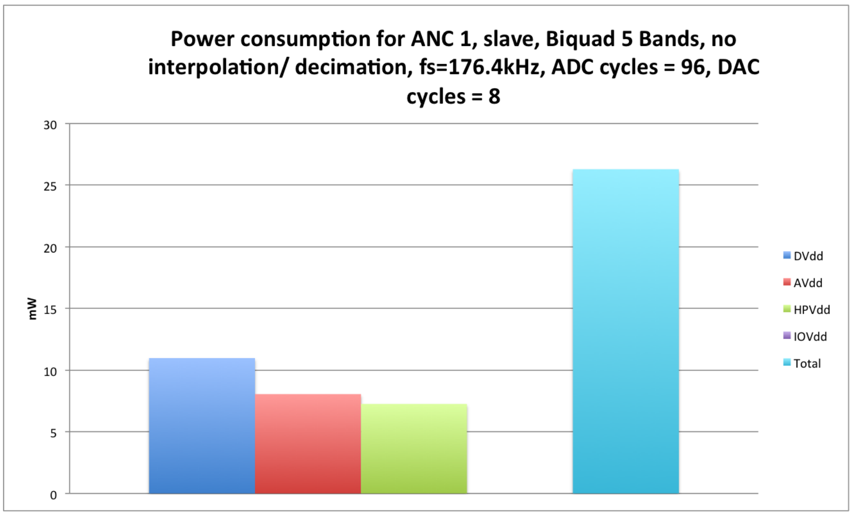

Figure 4.34: Power consumption for ANC Slave Setup

The slave does not need as much power as the ANC master set-up before, but there is no significant difference. Because the master audio codec is sending out the clock frequency, the IOV_{dd} pin power consumption is very high. Therefore, we continue to use the standalone ANC master version.

Chapter 5

Building the ANC headphone

An analog reference ANC headphone had to be identified, to compare the performance of the digital ANC system to the analog reference ANC system. The analog reference headphone should show a good ANC performance and it should be possible to disassemble the analog ANC controller without changing the acoustic behavior of the headphone. The microphones and the loudspeaker of the headphone should be connected to the TLV320AIC3253 audio codec and the original analog ANC controller by using a switch to toggle between the two systems. The goal was to rebuilding the analog ANC filters as close as possible using the digital biquad filters of the TLV320AIC3254 audio codec to match the ANC Performance of the analog reference system.

5.1 Analog Reference

As an analog reference, the Bose Quiet Comfort 15 has been chosen, which can be seen as the benchmark analog hybrid over-ear ANC headphone currently on the market (see figure 5.1).

Figure 5.1: Bose Quiet Comfort 15

The ANC controller is implemented on two circuit boards on each side of the headphone cup, the feed-forward microphones are mounted directly onto the circuit boards, as shown in figure 5.2.

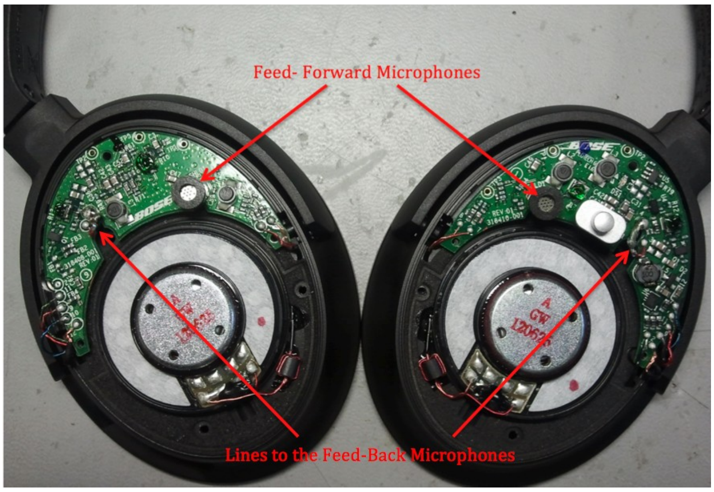

Figure 5.2: Circuit boards

The feed-back microphones are mounted inside the cup near the loudspeaker, the cables to these microphones are connected on the front of the circuit boards, as seen in figure 5.2.
The circuit boards can be removed without changing the acoustic behavior of the headphones, because the circuit boards are located in a separate cavity which is acoustically isolated from the loudspeaker. The headphones are powered by a single cell 1.5 V battery (AAA) and the overall power consumption is about 24 mW when the ANC system is turned on (no music playback).

5.1.1 Bose QC 15 ANC performance

Because the Bose QC15 headphone was the reference for the digital system, it was mandatory to know the ANC performance for comparing the digital to the analog system.
For measuring ANC performance of the Bose QC 15 headphone, a loudspeaker was placed in front of the dummy head. An exponential sine sweep signal was generated and the measurement microphones inside the artificial ear canal measured the transfer function. This measurement was used as a reference transfer function to eliminate the influence of the room and the loudspeaker itself.
To measure the passive attenuation, the headphone was placed on the artificial head without turning the ANC system on. The difference between this transfer function and the reference transfer function represents the passive attenuation of the headphone. For measuring the active attenuation, the ANC system was now switched on. The difference between this transfer function and the transfer function of the passive attenuation results in the active noise cancellation performance.

The active noise cancellation performance of the Bose Quiet Comfort 15 is shows in the following two figures 5.3 and 5.4:

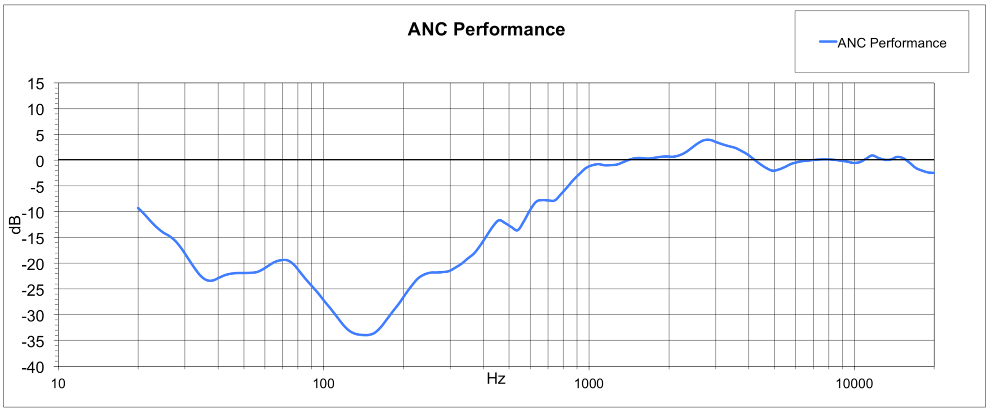

Figure 5.3: Bose QC 15 Performance left

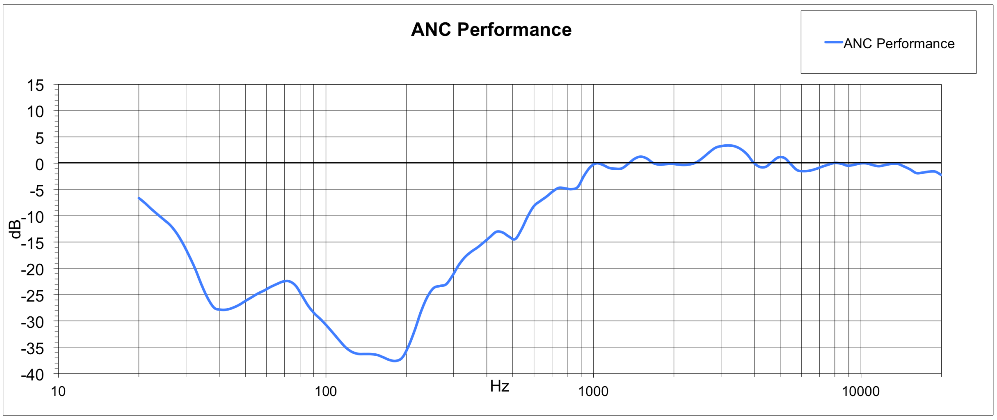

Figure 5.4: Bose QC 15 Performance right

The active attenuation is only working for the lower frequency range from about 20 Hz to 1000 Hz. The maximum damping is 37 dB at 180 Hz. As mentioned before, this is only the active without the passive damping of the headphone. The passive damping should be able to attenuate the higher frequency range to provide overall noise cancellation over a broad frequency range.

The passive attenuation of the headphone can be seen from the next figures 5.5 and 5.6, as well as the combination of the active and the passive damping.

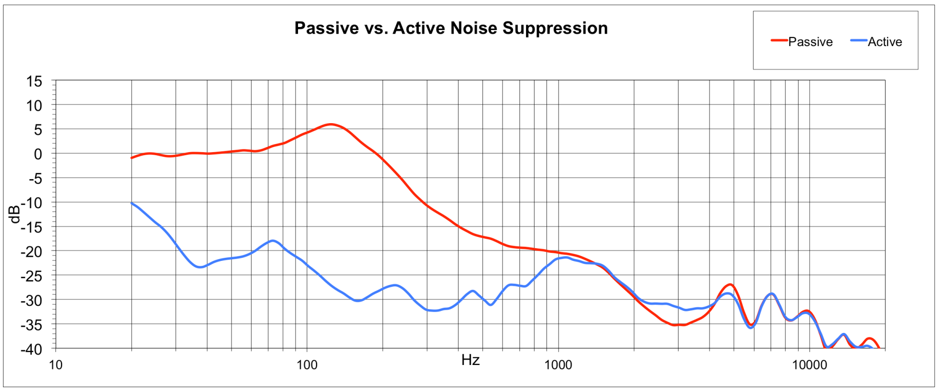

Figure 5.5: Active vs. Passive Performance left

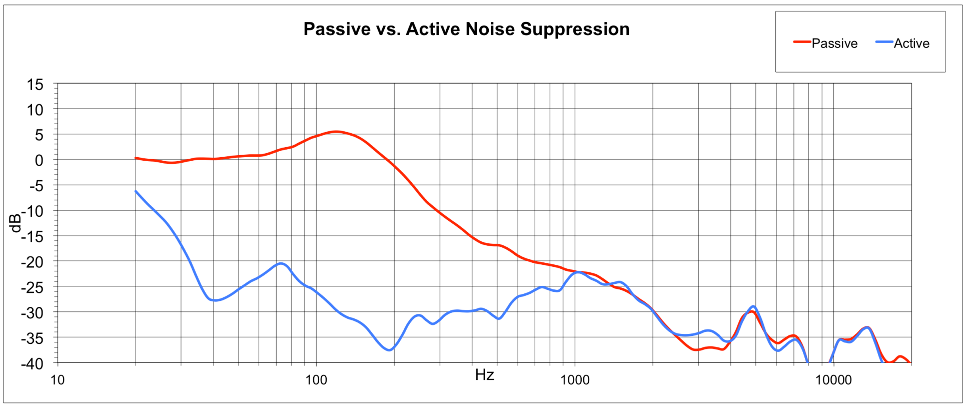

Figure 5.6: Active vs. Passive Performance right

Together, the noise cancellation covers a very broad frequency range with an average overall damping of about 30 db.

5.1.2 Analog filter curves

The discrete analog filters for the feed-back and the feed-forward ANC are implemented on the backside of the two circuit boards (one on the left, one on the right) (see figure 5.7).

Figure 5.7: Analog filter on the circuit board

To design a comparable digital ANC system, the filter transfer functions of both, feed-back and feed-forward path, have to be well-known in order to rebuild them using the digital biquad filters in the TLV320AIC3254 audio codec. The filter transfer functions were measured using the Audio Precision 2700 measurement system with an exponential sine sweep fed into the microphone inputs of the feed-forward and the feed-back microphone amplifier. The transfer functions of the filters were measured at the output of the headphone amplifier.

The measured transfer function of the feed-forward filter for the left and the right side can be found in figure 5.8:

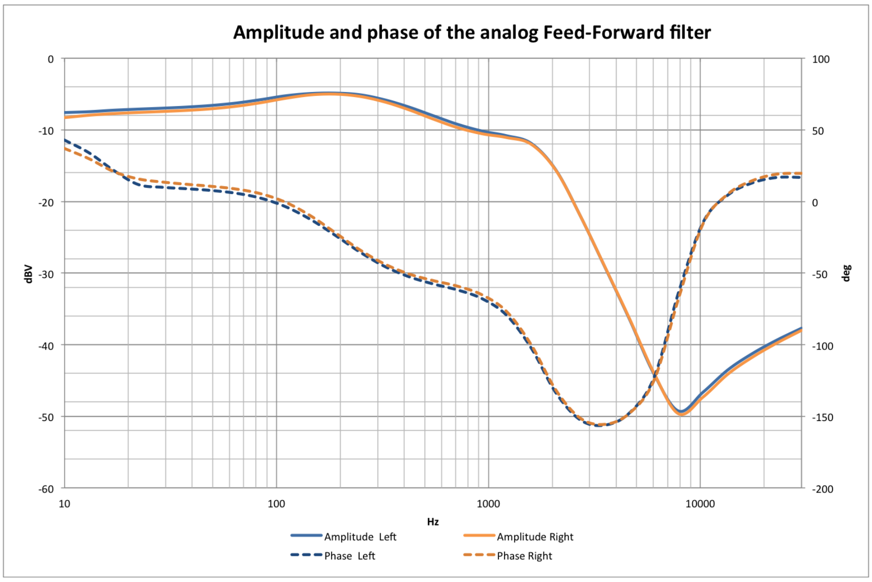

Figure 5.8: Transfer function of the Bose feed-forward filter

The transfer function of the feed-forward filter is the same for both the left and right side. The transfer function of the feed-back filter for the left and right side can be found in figure 5.9:

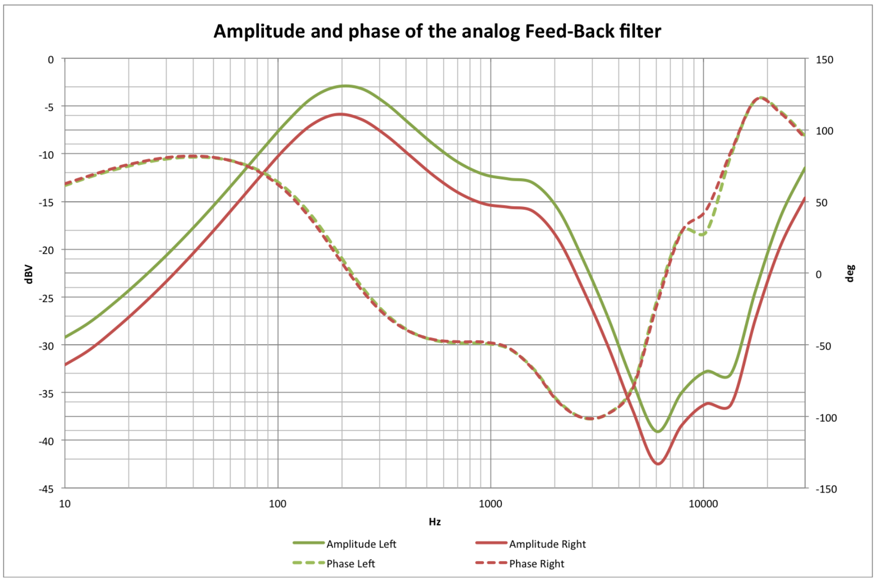

Figure 5.9: Transfer function of the Bose feed-back filter

The transfer function of the feed-back filter is the same for both the left and right side, too, but the gain of the left microphone is about 3 dB higher than the right microphone, obviously to compensate a gain mismatch between left and right microphone.

5.2 Hardware

For comparing the digital to the analog system it is important that the acoustic environment is the same for both systems. In order to achieve this, all input lines (loudspeaker) and output lines (microphones) of the headphone had to be moved to the outside of the headphone, to connect them either to the analog or the digital ANC system, as shown in figure 5.10.

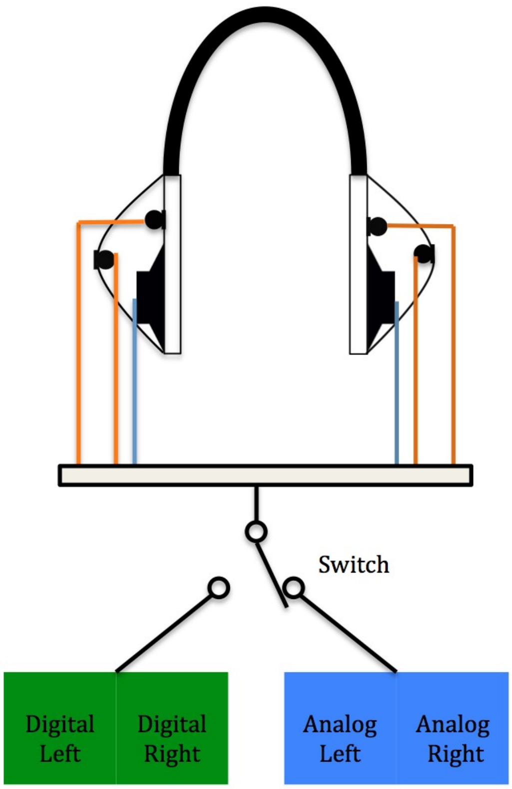

Figure 5.10: Hardware schematic

The analog Bose circuit boards were removed from the headphone to connect them externally to a switch for easily comparing them to the digital system.
Because the microphones for the feed-forward ANC system are mounted directly onto the Bose circuit board, new microphones of the same type have been installed at the same spot inside the headphone cup, as seen in figure 5.2.

For each side of the headphone, five wires have been pulled outside:

- Feed-forward microphone (+)
- Feed-back microphone (+)
- Microphone ground (-)
- Loudspeaker (+)
- Loudspeaker ground (-)

The wires and the new microphones can be seen from figure 5.11.

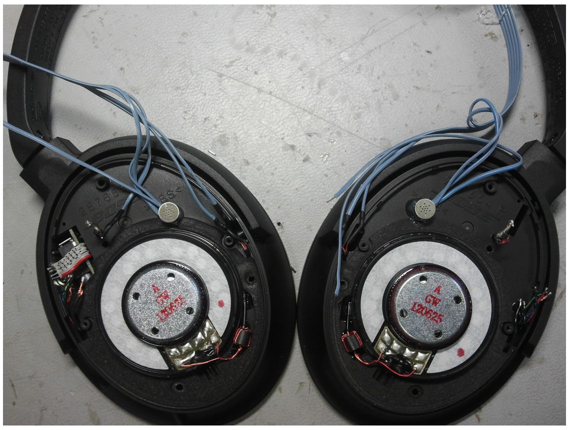

Figure 5.11: Headphone with lines pulled-out

The wires are connected to five stereo relays (Omron G5V-2 type), which can be toggled using a simple switch. The outputs of the relays are connected to the corresponding microphone inputs, headphone amplifier outputs of the analog circuit board and the digital circuit board. There is also a mute button, which mutes the headphone loudspeaker to completely turn off the ANC systems.

The complete hardware setup is shown in figure 5.12:

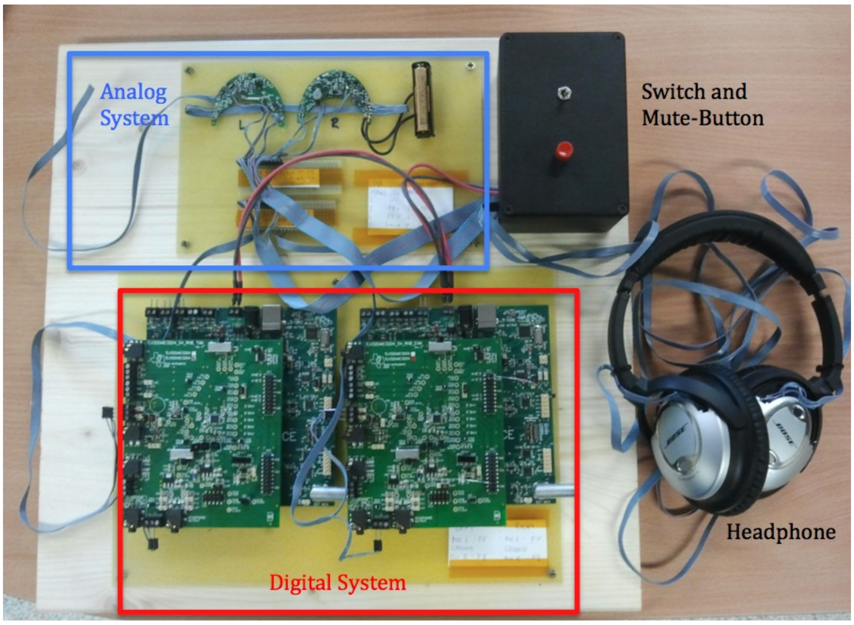

Figure 5.12: Hardware

5.3 Programming of the digital Filters

To check the performance of the digital filters, the microphone inputs and the headphone outputs of the TLV320AIC3254 were connected to the Audio Precision 2700 measurement system. The amplitude and phase response was measured using a sine sweep similar to how the analog filter curves have been measured.
The goal was to match the amplitude and the phase response of the analog reference as good as possible. Because the TLV320AIC3254 audio codec does introduce a propagation delay which causes a phase shift (see chapter 4.2.1 and 4.2.2), it was not possible to match both amplitude and phase completely by using one filter. So two filters (Filter A and Filter B) were built for the feed-forward and feed-back system, respectively: Filter A tries to match the amplitude response of the Bose filter as close as possible, Filter B tries to match the phase response as close as possible. After programming the filters, the performance of the two filters was analyzed and compared to the analog system.

5.3.1 Feed-forward filter

For the feed-forward filter, five bands of the biquad filter instance were used. There is a treble shelf (4000 Hz), a notch equalizer (7500 Hz) and three more equalizers (50 Hz, 240 Hz and 1700 Hz) for reshaping the analog filter curve.

Filter A: Matching the amplitude response

Figure 5.13 shows the amplitude response and phase response of filter A, compared to the analog filter (left side):

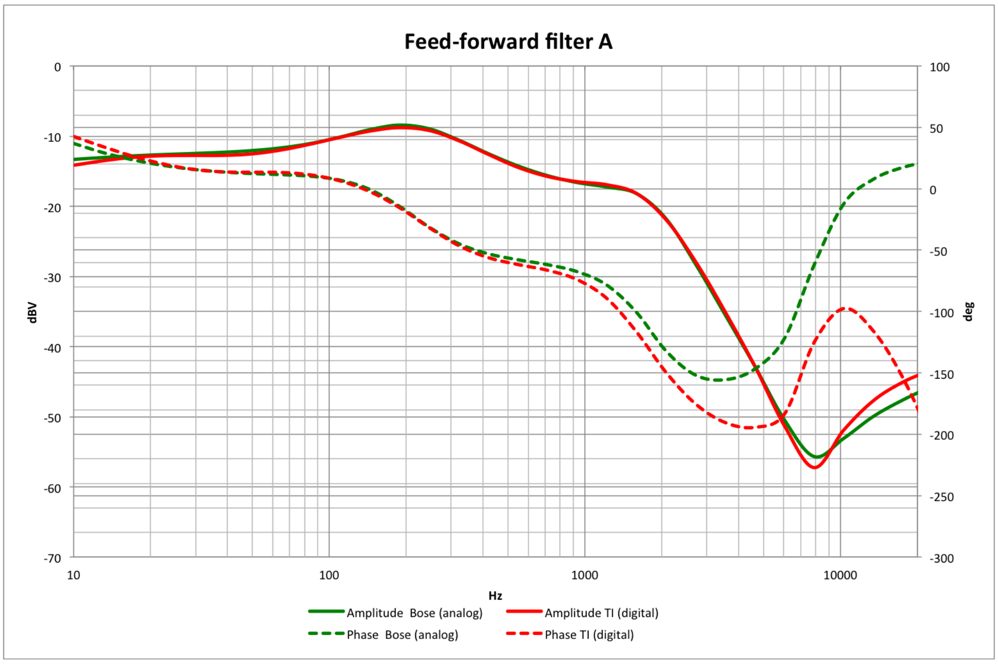

Figure 5.13: Comparison of Filter A to the analog filter

The amplitude of the analog and the digital filters is very similar. Only for the very low frequencies (below 20 Hz) and for the very high frequencies (above 8000 Hz) they are not matching completely. Because the human ear is hardly able to sense frequencies below 20 Hz, the amplitude and phase mismatch is negligible for this frequency area.
Also, the mismatch at the very high frequencies is not important for the ANC performance because the passive damping of the headphone itself is working very well in this frequency region.
When looking at the phase response, the phase of the digital filter starts to drift away at about 400 Hz compared to the analog filter phase response. This is because the DSP itself introduces a phase shift due to the propagation delay (see chapter 4.2.1 and 4.2.2)).

Filter B: Matching the phase response

Given that filter A is supposed to match the amplitude of the analog filter as close as possible, Filter B is supposed to match the phase response of the analog filter. The phase response can be changed by using additional filtering, which means changing the amplitude response in a way that the phase response is matching the analog filter curve more closely.
Figure 5.14 shows the amplitude response and phase response of filter B, compared to the analog filter:

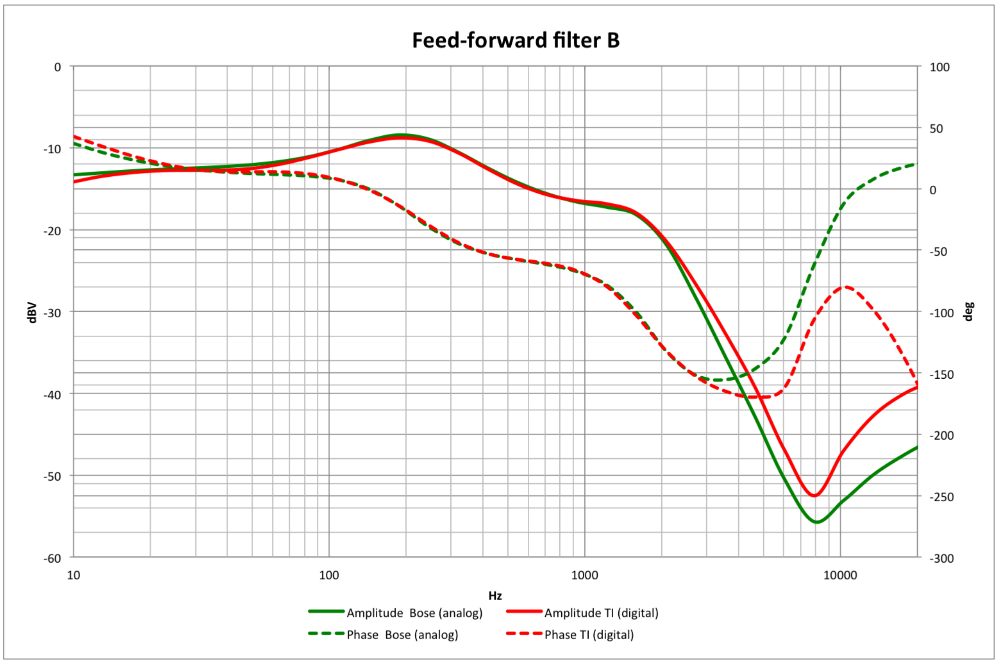

Figure 5.14: Filter B

It is possible to match the phase response of the analog filter up to 3000 Hz by adding more gain in the high frequency region (compare figure 5.13 and figure 5.14).
The problem is that with adding more gain in the high frequency region also unwanted noise or overshoot (negative ANC performance at higher frequencies) increases.

5.3.2 Feed-back filter

For the feed-back filter, also five bands of the biquad filter have been used. There is one high shelf (12000 Hz), two notches (6000 Hz and 13000 Hz) and two more parametric equalizer bands (17hHz and 1800Hz) for reshaping the analog filter curve. The high-pass characteristic of the analog feed-back filter was reproduced by adapting the capacitor of the analog high-pass (before the A/D converter) on the digital evaluation board.

Filter A: Matching the amplitude response

Like on the feed-forward filter, Filter A tries to match the analog amplitude as close as possible, which can be seen from the next figure 5.15:

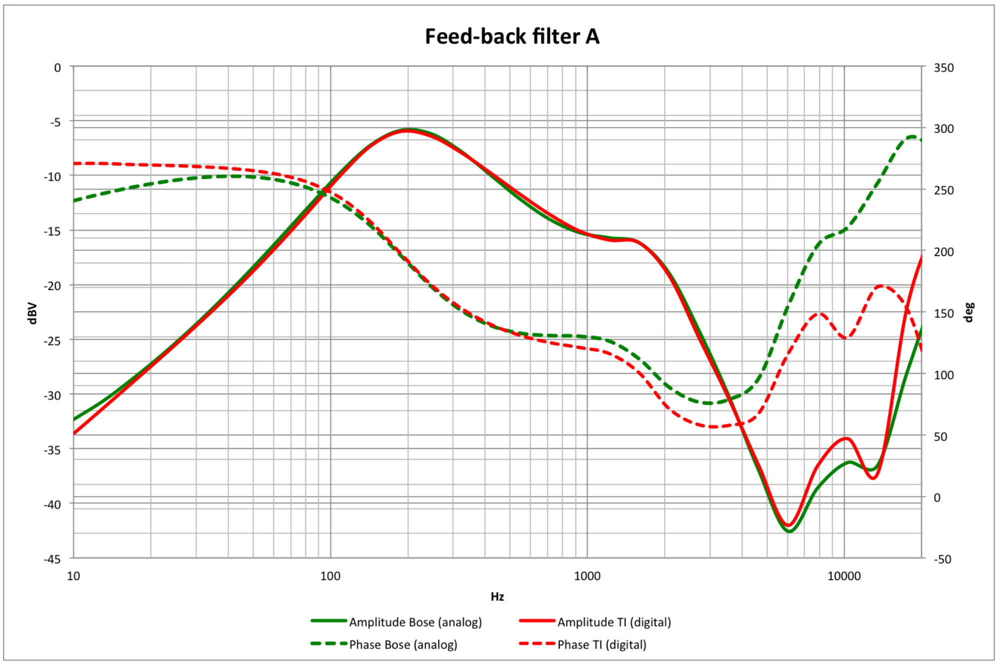

Figure 5.15: Analog vs. digital filter

Again, the amplitude of the digital filter matches the amplitude of the analog filter very well, except for the very low frequencies and the very high frequencies.
Considering the phase response in the low frequency region (below 100 Hz), it is noticeable that the analog phase is drifting off the digital phase. We can assume that the analog filter produces a lot of gain in the very low frequency region (below 10 Hz), which would explain the behavior of the analog phase. Because the Audio Precision measurement system is not able to measure frequencies below 10 Hz, it is not clear how the analog filter curve looks like in this frequency area.

Additionally, the phase of the digital filter starts to drift up again at 400 Hz because of the propagation delay of the DSP.

Matching the phase response

Filter B tries to match the phase response of the analog feed-back filter. To compensate for the phase mismatch because of the propagation delay, some additional gain in the high frequency area is needed.
The comparison between filter B and the analog filter can be found in figure 5.16:

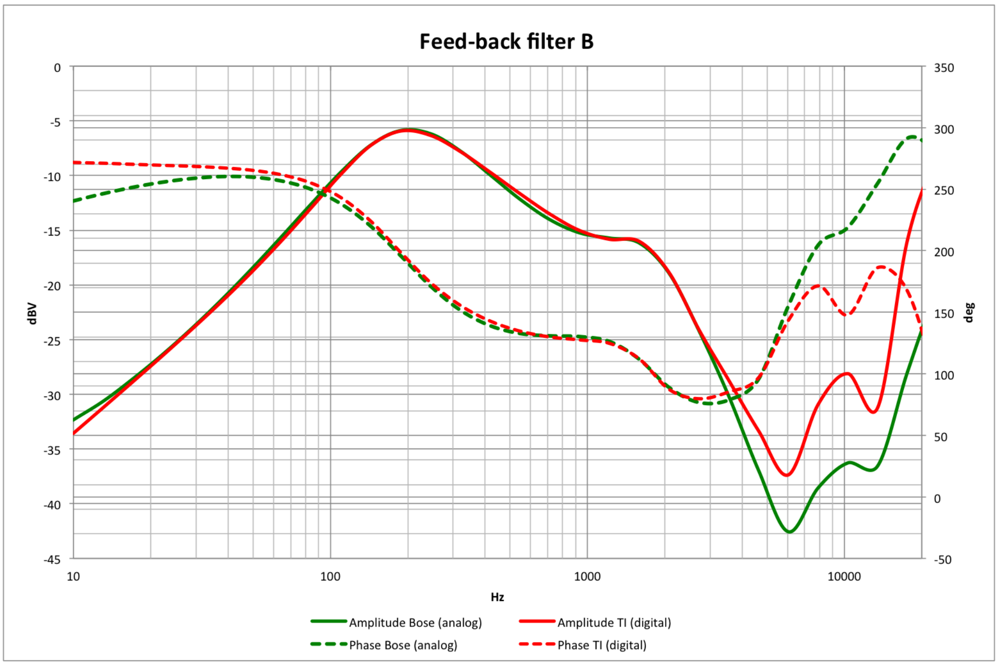

Figure 5.16: Analog vs. digital filter

Because the feed-back system is a closed-loop system, the risk of unwanted acoustic feedback and instabilities increases when adding more gain in the high frequency region.

5.4 ANC Performance of the digital system

Before measuring the ANC performance with both filter A and filter B, the right gain setting of the feed-forward and feed-back microphones had to be identified. It is important for the headphone to be acoustically stable so that no unwanted feedback occurs, which means the headphone starts to oscillate and "whizzles" especially at higher frequencies. Special use cases like pressing the headphone cup to the ear can change the open loop characteristics of the headphone and lead to such instabilities if there's too little gain or phase margin in the feedback ANC filter.

Because filter A is damping the high frequencies better than filter B, it is possible to add more microphone gain before the headphone gets unstable.
The comparison of the ANC performance for both filters can be seen in the next two figures 5.17 and 5.18:

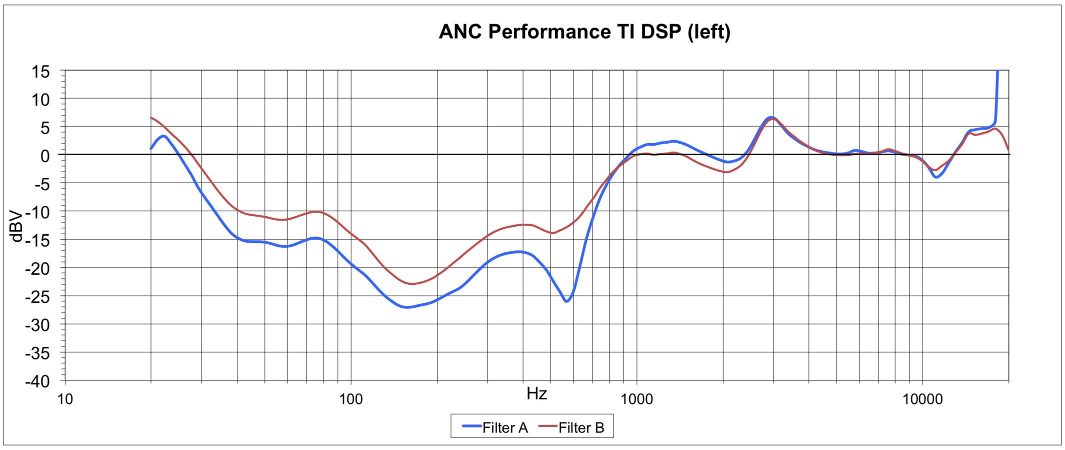

Figure 5.17: ANC Performance TI left

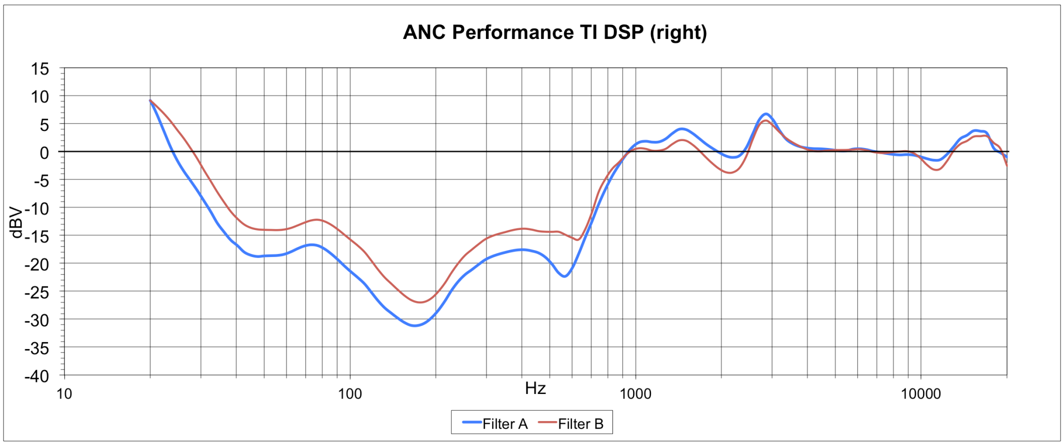

Figure 5.18: ANC Performance TI right

Because the gain of the feed-back microphone is higher for filter A, the ANC performance of filter A is better than filter B for frequencies below 800 Hz. Above 800 Hz, the performance of filter A is getting worse than filter B because the phase mismatch between the digital filter and the analog reference filter is getting worse.

5.4.1 Comparison to the analog system

The comparison between the analog system and the digital systems (filter A and filter B) can be seen from figures 5.19 and 5.20:

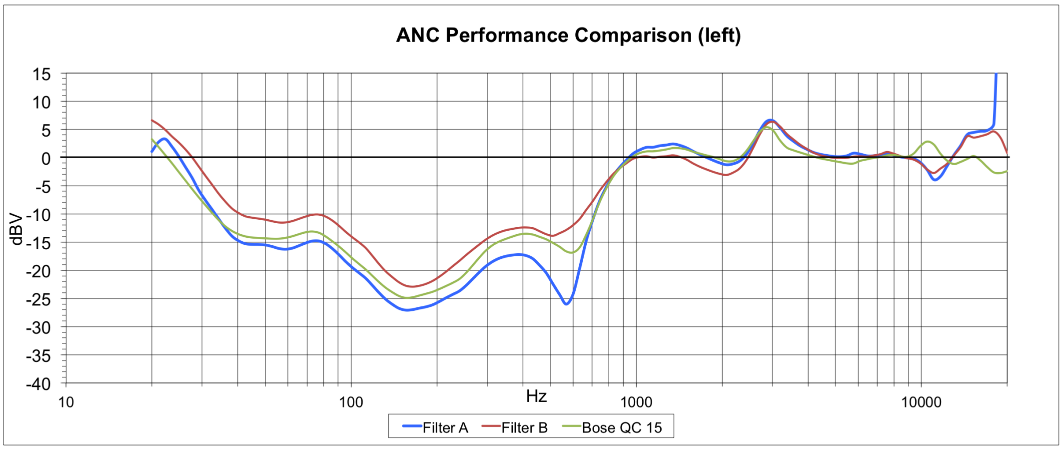

Figure 5.19: ANC performance comparison left

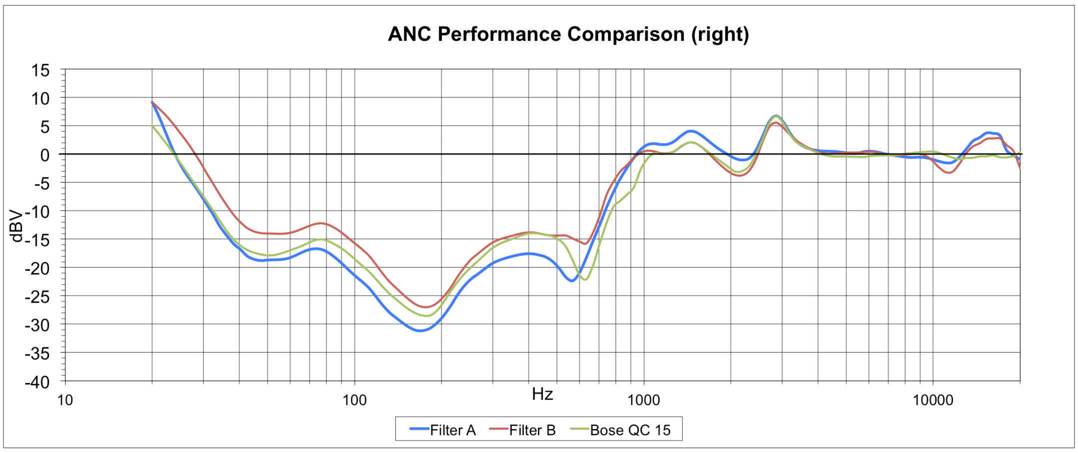

Figure 5.20: ANC performance comparison right

***NOTE**: The ANC performance of the analog system was measured again because the microphones for the feed-forward system were changed, as described in chapter 5.2. Although the microphones are the same types, the new feed-forward microphones are not as sensitive as the original microphones. Because the microphone gain of the analog system cannot be changed, the ANC performance with the new feed-forward microphones is not as good as the original ANC performance (comparison: figure 5.4 / 5.20 and figure 5.3/5.17).*

The characteristic of the ANC performance curve of the analog system is very similar to the ANC performance curve of the digital system. Both systems are working within a frequency range from 25 Hz to 1000 Hz which is a typical range for ANC headphones.

The best ANC performance in this frequency region is achieved by using filter A, although filter A produces the most overshoot for frequencies above 1000 Hz because of the propagation delay of the DSP.

We assume that it would be possible to achieve the same ANC performance with the Bose QC 15 filter if it would be possible to adjust the gain of the microphone amplifier.

The ANC performance of Filter B is worse than filter A and the Bose QC 15, but there is not as much overshoot like with filter A because the phase mismatch was compensated with the help of additional filtering (see figure 5.14 and 5.16).

5.5 Listening Test

For the listening test 15 test persons were asked to do a blind comparison between the analog and the digital system and to answer several questions afterwards. The subjects were able to switch directly between the two systems (A and B) but not knowing whether they have selected the analog or the digital system. A wooden box was built around the two ANC controllers to avoid any visual feedback to the test people (see figure 5.21). The listening test was done in the listening room of the *JOANNEUM RESEARCH*.

Figure 5.21: Listening room

In order to create a repeatable situation for each test person, the ANC performance of the digital system was adjusted to match the performance of the analog system, as seen in figure 5.22 and 5.23:

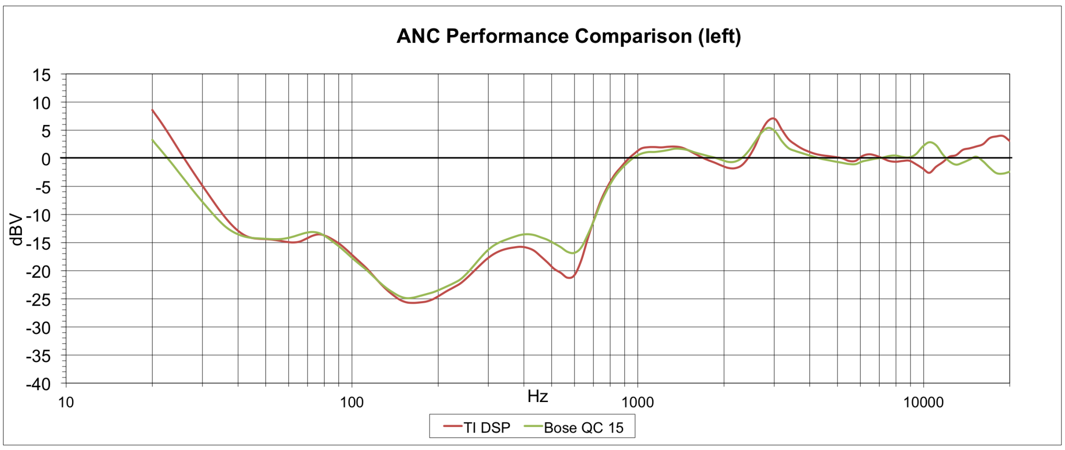

Figure 5.22: ANC performance comparison left

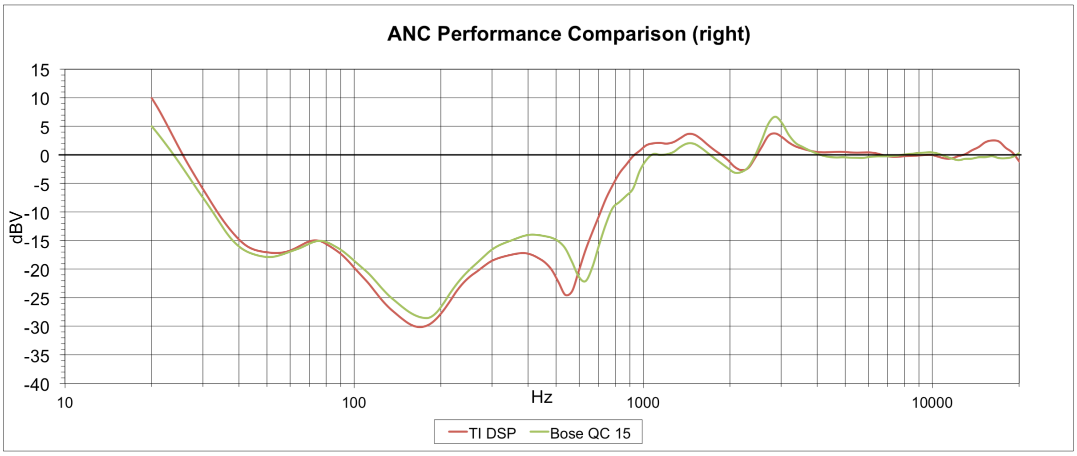

Figure 5.23: ANC performance comparison right

The subjects were confronted with four questions, for each question there was a slider to evaluate the system "A" and the system "B".
The first two questions included playing back audio files of three different sound sources (pink noise, airplane cabin noise and traffic noise) and the test person had to answer for each sound file separately.

Before the listening test started, each test person was instructed carefully in order to understand the following questions:

1. "Evaluate the active noise cancellation of both systems for different noise sources." (Lower value: rather bad – higher value: rather good)

2. "Evaluate the accentuation of the high frequency range (overshoot) of both systems for different noise sources" (Lower value: less disturbing – higher value: more disturbing)

3. "How comfortable do you feel when wearing the headphone of both systems?" (Lower value: rather uncomfortable – higher value: rather comfortable)

4. "How loud do you perceive the noise floor of both systems?" (Lower value: rather quiet – higher value: rather loud).

To answer the question, the test persons had to choose between ten values, from -5 to +5. The aim of question one was simply to find out if one of the two systems was able to cancel out certain noise scenarios better than the other.

Question two was about finding out whether the test person was able to hear the overshoot produced by the ANC headphone and which system produced less disturbing overshoot for various noise sources.

The aim of question three was about evaluating the "comfort level" of each system: In particular for the low frequency region, ANC headphones sometimes generate a feeling of pressure to the ear.

Question four was about the perceived loudness of the noise floor for both systems.

5.5.1 Analysis

To analyze the answers of the subjects, the Wilcoxon test was used, which is a non-parametric statistical hypothesis test for paired samples.
First, the (two-sided) Wilcoxon test proofs whether there is any significant difference between system A (analog) and system B (digital):

Hypothesis A0: There is no significant difference between system A and system B

If this hypothesis is rejected, another (one-sided) Wilcoxon test proofs whether the value of system A is lower, equal or higher than the value of system B:

Hypothesis B0: The value of system A is lower/ equal than the value of system B

For each question, the test statistic V and the p-value were calculated. A p-value under 0.05 indicates that the hypotheses A0 (and B0) have to be rejected.

Question one: "Evaluate the active noise cancellation of both systems for different noise sources."

The result of the Wilcoxon test **for all playback sound files** is:

- V: 291
- p-value: 0.04

=> **Barely no significant difference** between the two systems for all sounds.

The result of the Wilcoxon test **for playback sound files one (pink noise)** is:

- V: 30
- p-value: 0.73

=> **No significant difference** between the two systems for sound one

The result of the Wilcoxon test **for playback sound file two (air plane cabin)** is:

- V: 39
- p-value: 0.03

=> **Significant difference** between the two systems for sound one, system A (analog) is performing better than system B (digital).

The result of the Wilcoxon test **for playback sound file three (traffic noise)** is:

- V: 29
- p-value: 0.27

=> **No significant difference** between the two systems for sound three

The following figure 5.24 shows a bar diagram of the answers for question one over all sound files:

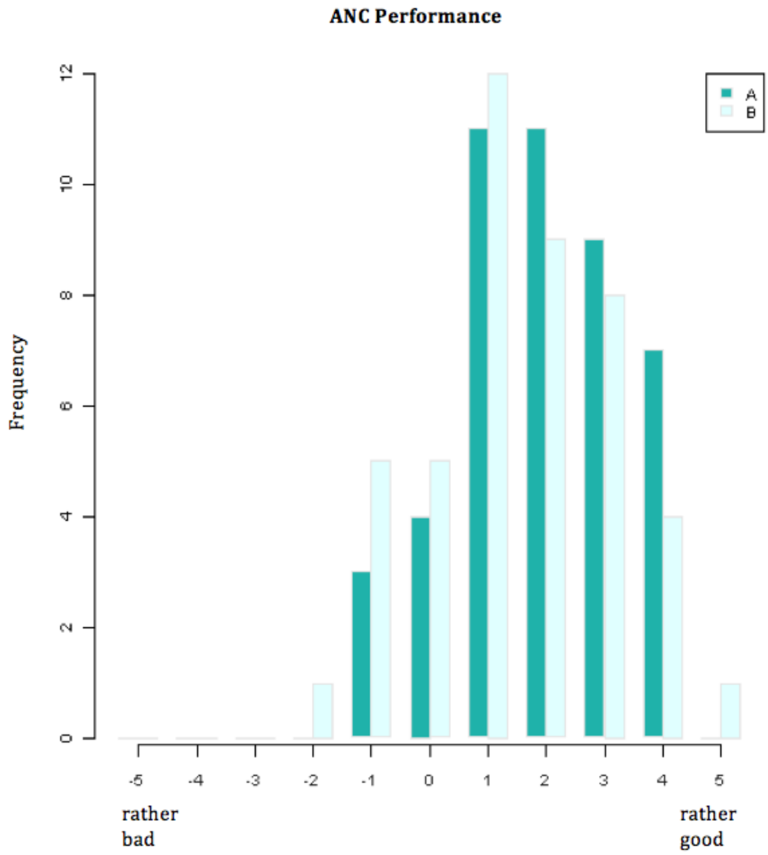

Figure 5.24: ANC performance (all sound files)

Figure 5.25 shows a box plot of the answers to question one for all sound files:

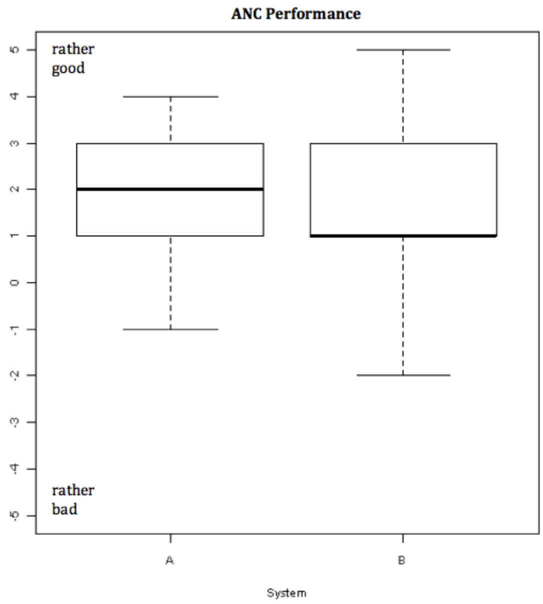

Figure 5.25: ANC performance (all sound files)

Figure 5.26 shows a box plot of the answers to question one over the different sound files. (The bold line inside the boxes always represents the median).

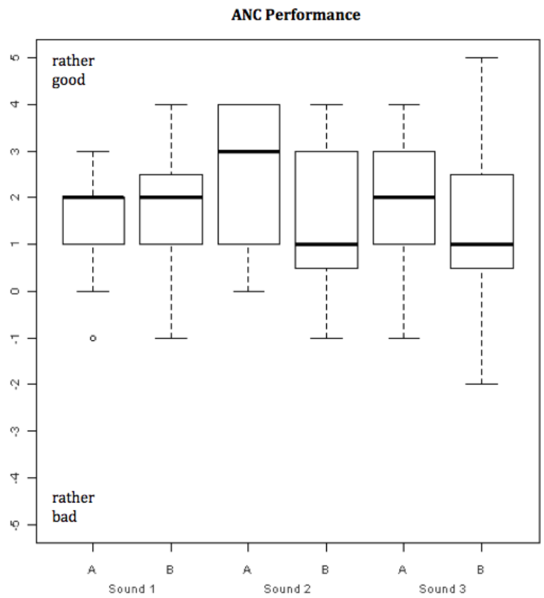

Figure 5.26: ANC performance (different sound files)

Question two: "Evaluate the accentuation of the high frequency range (overshoot) of both systems for different noise sources"

The result of the Wilcoxon test **for all playback sound files** is:

- V: 73.5
- p-value: 0.01

=> **Significant difference** between the two systems over all sounds, system A is less disturbing than system B.

The result of the Wilcoxon test **for playback sound files one (pink noise)** is:

- V: 13.5
- p-value: > 0.05

=> **No significant difference** between the two systems for sound one.

The result of the Wilcoxon test **for playback sound file two (air plane cabin)** is:

- V: 39
- p-value: 0.03

=> **Significant difference** between the two systems for sound one, system A (analog) is less disturbing than system B (digital).

The result of the Wilcoxon test **for playback sound file three (traffic noise)** is:

- V: 10.5
- p-value: 0.10

=> **No significant difference** between the two systems for sound three.

The following figure 5.27 shows a bar diagram of the answers for question two for all sound files:

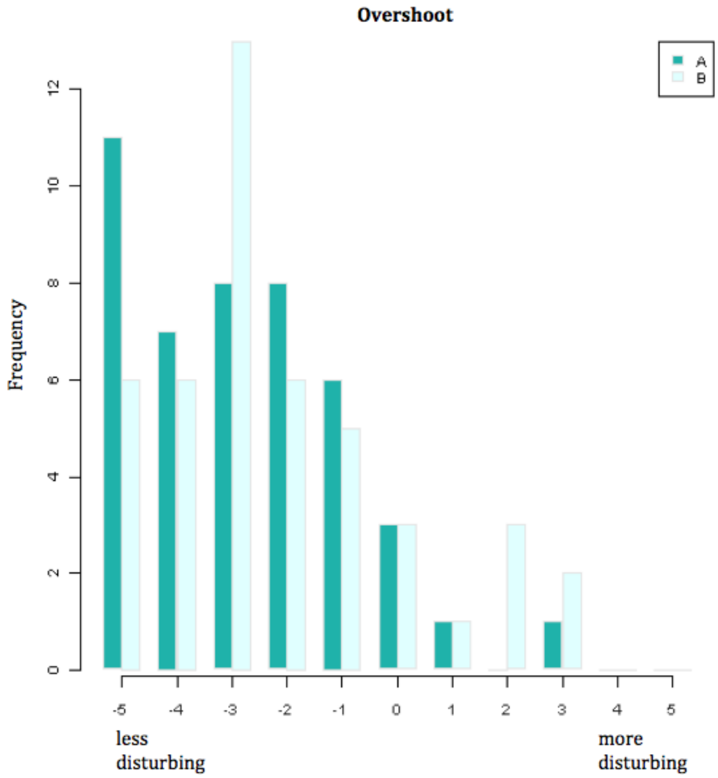

Figure 5.27: Overshoot (all sound files)

Figure 5.28 shows a box plot of the answers of question one for all sound files:

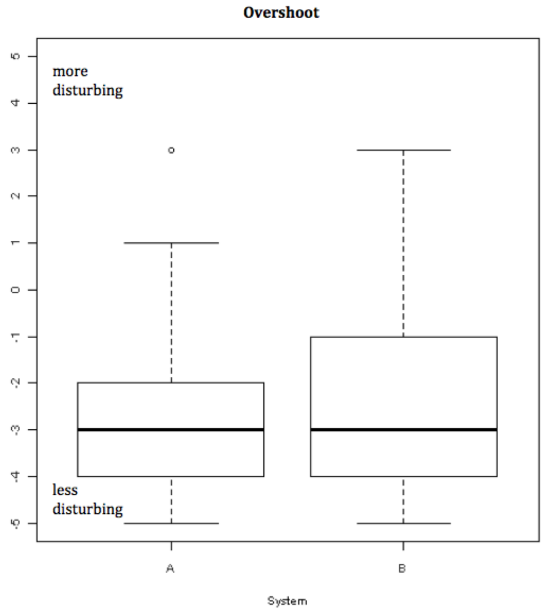

Figure 5.28: Overshoot (all sound files)

Figure 5.29 shows a box plot of the answers on question one for the different sound files. (The bold line inside the boxes always represents the median):

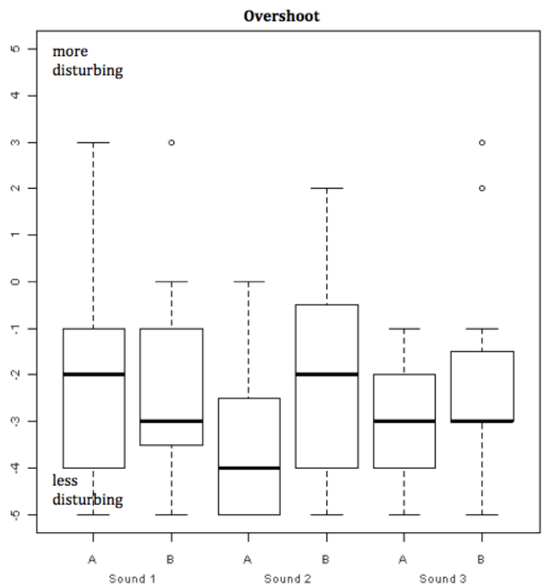

Figure 5.29: Overshoot (different sound files)

Question three: "How comfortable do you feel when wearing the headphone for both systems?"

The result of the Wilcoxon test is:

- V: 34.5
- p-value: 0.09

=> **No significant difference** between the two systems

The following figure 5.30 shows a bar diagram of the answers to question three:

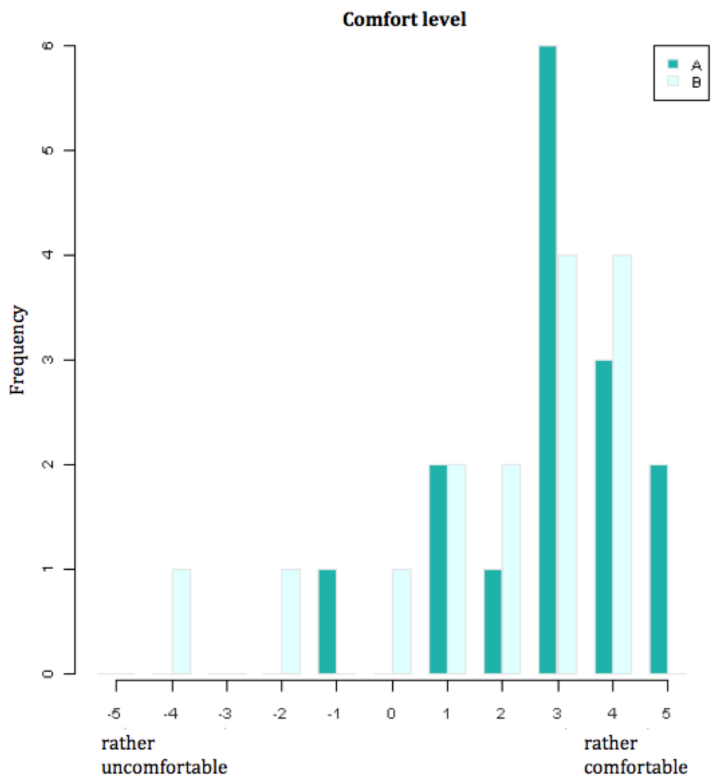

Figure 5.30: Comfort level

Figure 5.31 shows a box plot of the answers to question three:

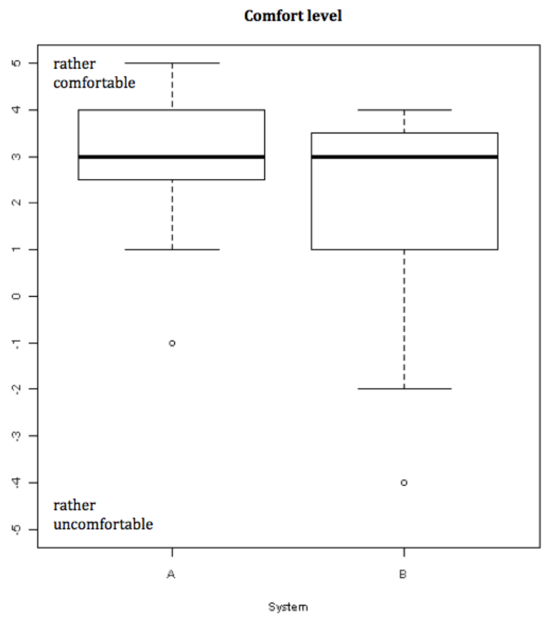

Figure 5.31: Comfort level

Question four "How loud do you perceive the noise floor for both systems?"

The result of the Wilcoxon test is:

- V: 0
- p-value: <0.001

=> **Significant difference** between the two systems, system A (analog) is more quiet than system B (digital).

The following figure 5.32 shows a bar diagram of the answers to question three:

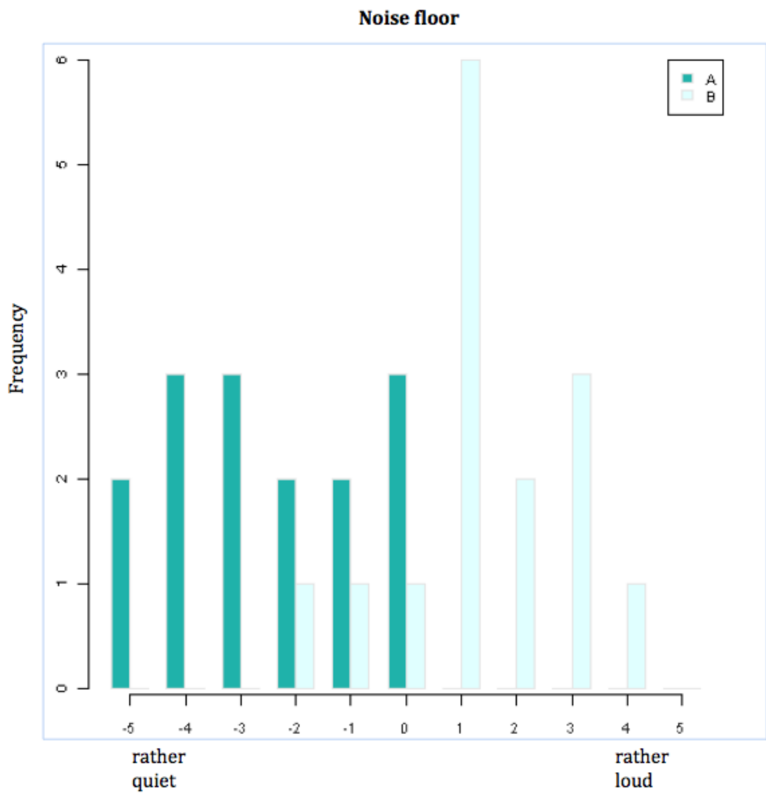

Figure 5.32: Noise floor

Figure 5.33 shows a box plot of the answers to question three:

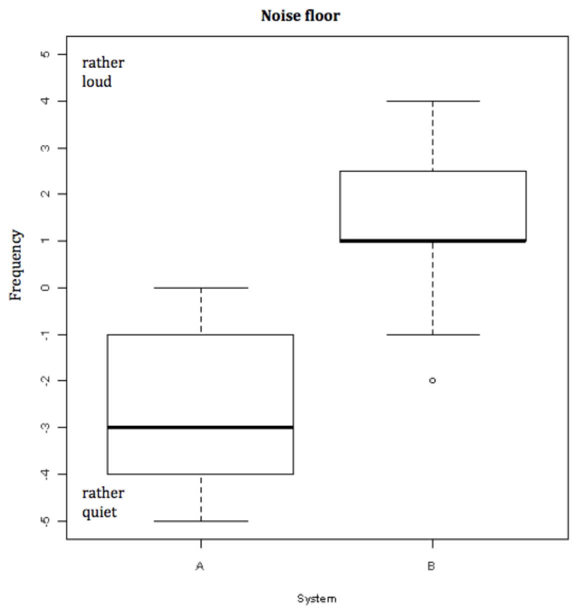

Figure 5.33: Noise floor

5.5.2 Summary

The listening test shows that there is barely any significant difference between the analog system and the digital system in terms of ANC performance (question one). If we look at the results for the different sound files, it is noticeable that a difference between the two systems only exists for sound file two (airplane cabin noise). For the other two sound files, no significant difference could be found.

For the overshoot (question two) a difference between the two systems also exists only for sound file two.

The test also has shown that there is no difference regarding the comfort level of the two systems (question three).

The noise floor (question four) of the analog system is less disturbing than the noise floor of the digital system. We can assume that the reason for the better noise floor is that the analog system by Bose is using higher quality microphone and headphone amplifiers than the digital system by Texas Instruments.

Chapter 6

Conclusion

This thesis shows that it is possible to build a hybrid ANC headphone using a digital signal processor, which performs very similar to a high end analog solution. This was proven by performance measurements as well as subjective listening tests.

The computation time of the DSP, when using high sample rates, is small enough to get a phase response which does not degrade the ANC performance of the headphone in a significant way. The benefit of digital filters (chapter 3.3) is usually a shorter development time compared to an analog filter with discrete components.

In terms of power consumption, the developed digital solution is not able to compete with the analog solution.
To build a hybrid ANC headphone using the TLV320AIC3254, three audio codecs are needed: ANC left, ANC right and one DSP for music playback equalization. This results in a total power consumption of about 73 mW. The analog reference system has about 24 mW of total power consumption, which is about a third compared to the digital solution (see figure 6.1).
The high power consumption of the DSPs makes it difficult to design a battery- powered portable headphone which could compete against an analog solution.

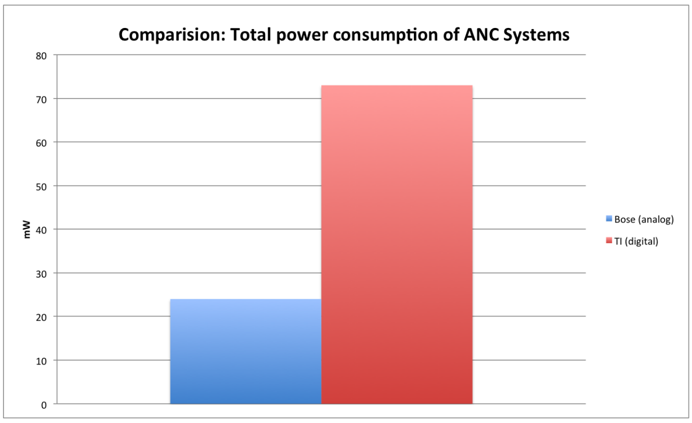

Figure 6.1: Power consumption of both systems

6.1 Outlook

Digital signal processors are constantly getting faster and more efficient. We can assume that in the near future, the power consumption of DSPs will be reduced to a level where digital ANC headphone system will show the same or even less power consumption than a comparable analog ANC headphone.
Also the computation time of the DSPs will be faster which makes it possible to design a system which can process audio signals in near-to-real time.
In addition to the benefits of programmability some additional audio processing, like "3D sound" or reverberation can be implemented very easily within the digital system.

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Chapter 7

Appendix

This appendix does include additional figures of the listening test which have not been used in the thesis.

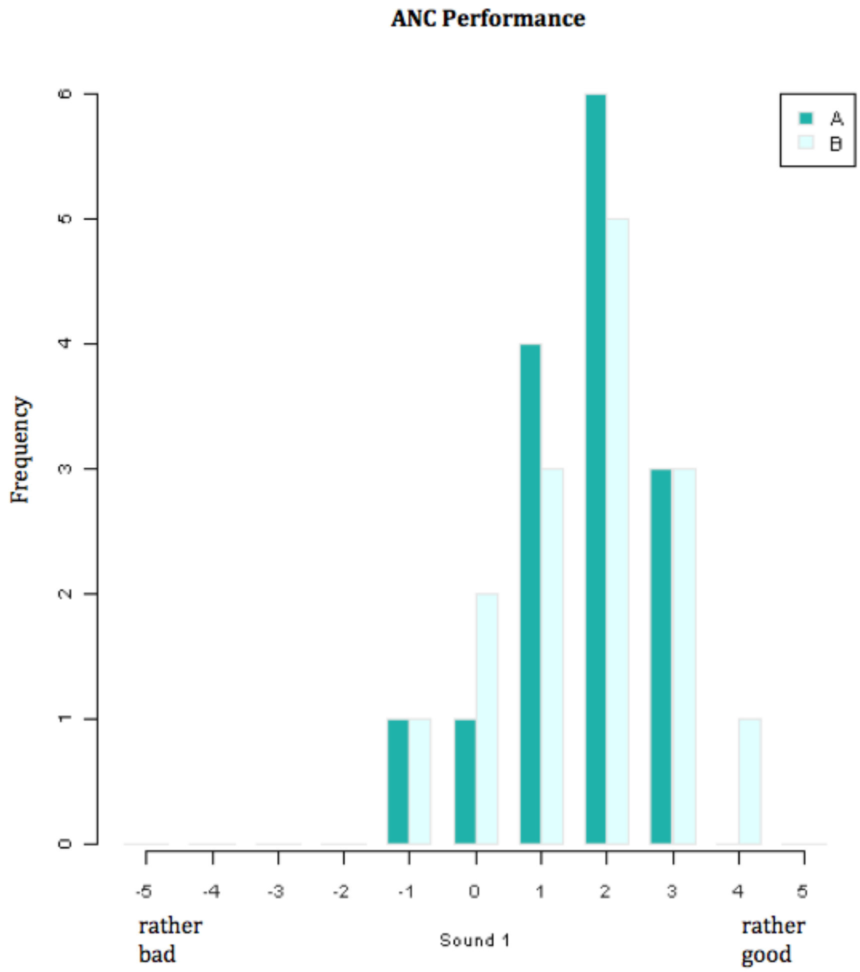

Figure 7.1: ANC performance: Sound 1

 Chapter 7 Appendix

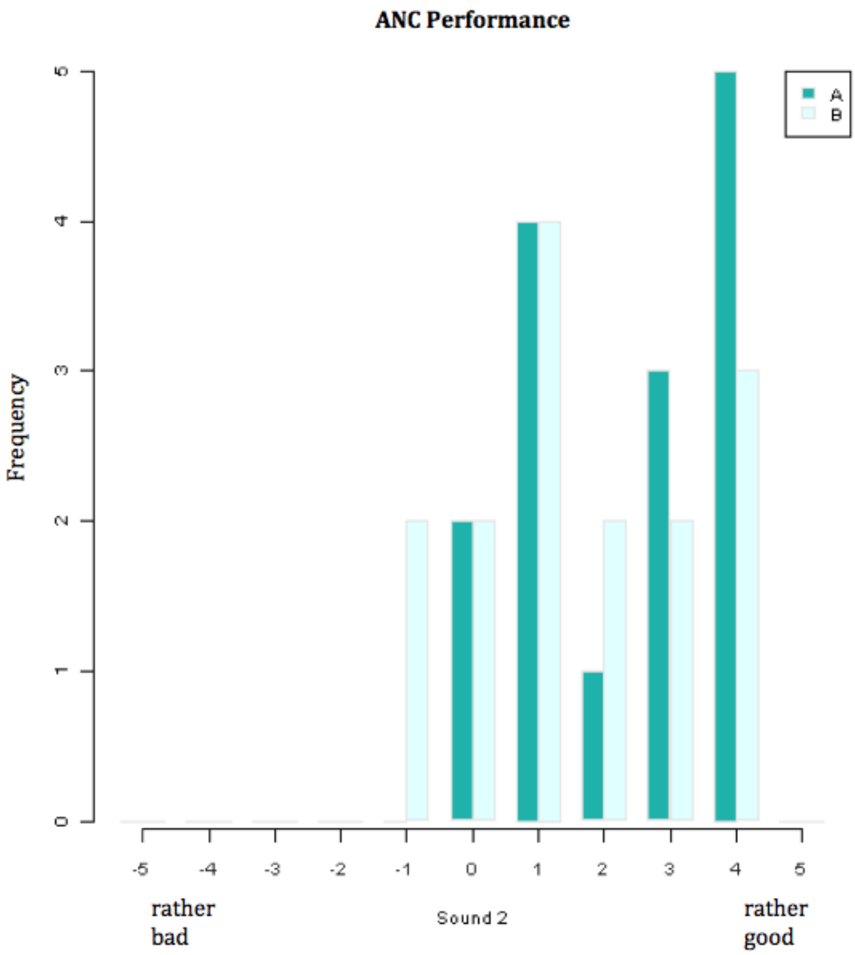

Figure 7.2: ANC performance: Sound 2

Chapter 7 Appendix

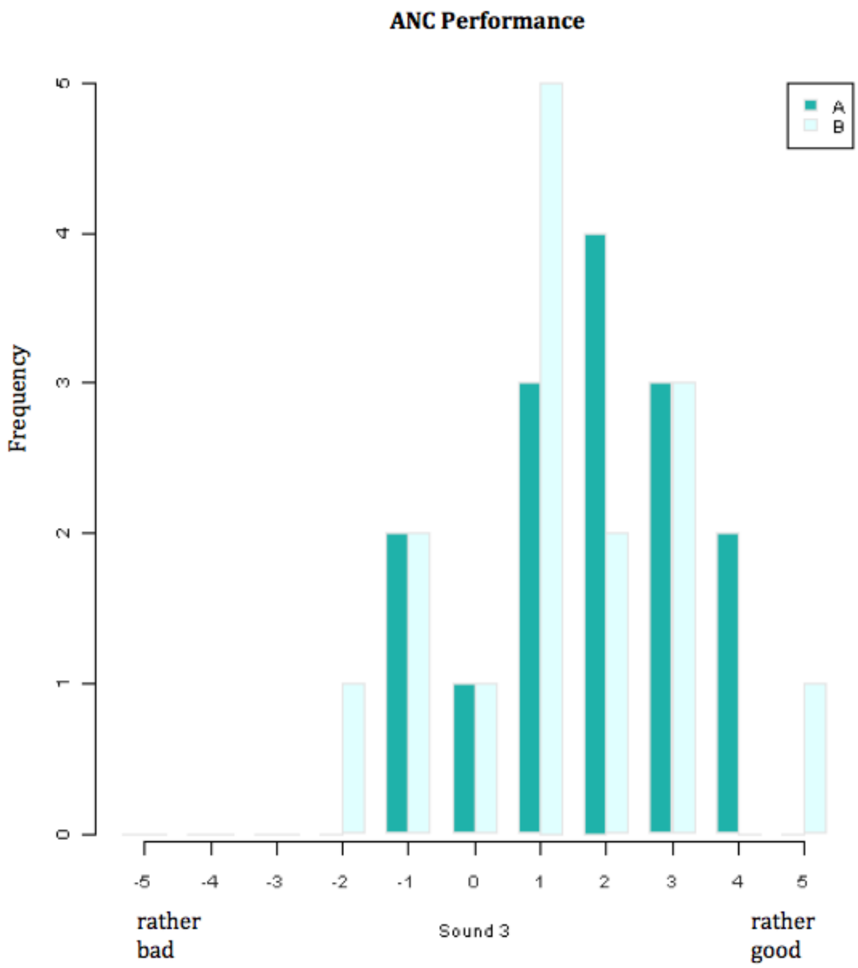

Figure 7.3: ANC performance: Sound 3

 Chapter 7 Appendix

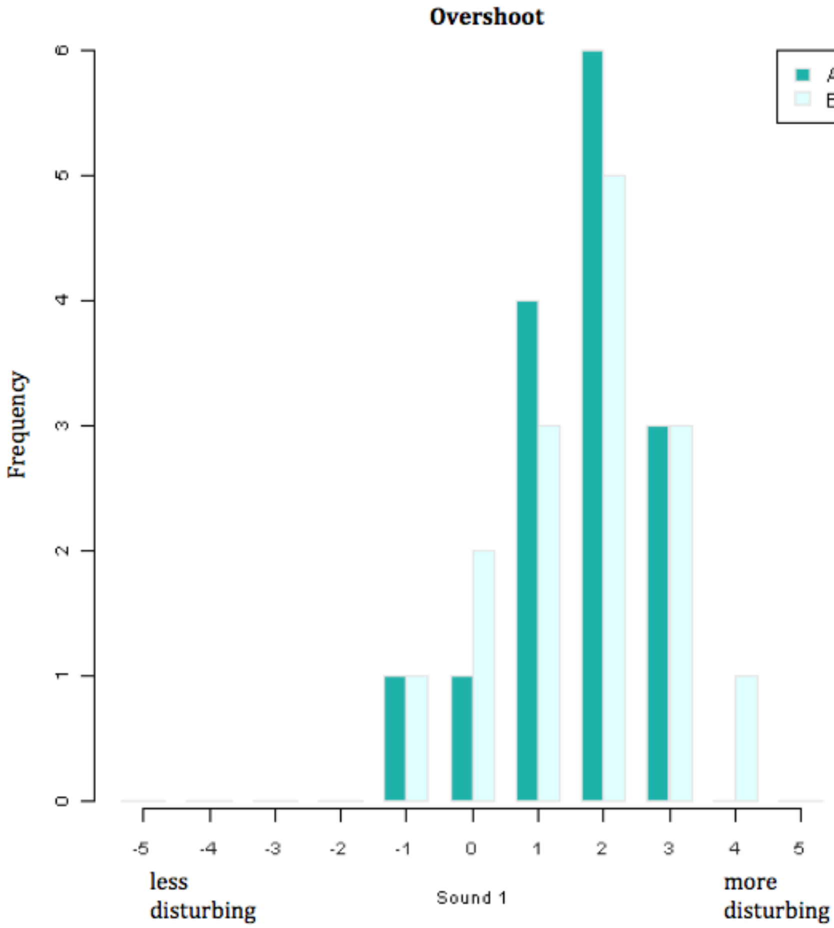

Figure 7.4: Overshoot: Sound 1

 Chapter 7 Appendix

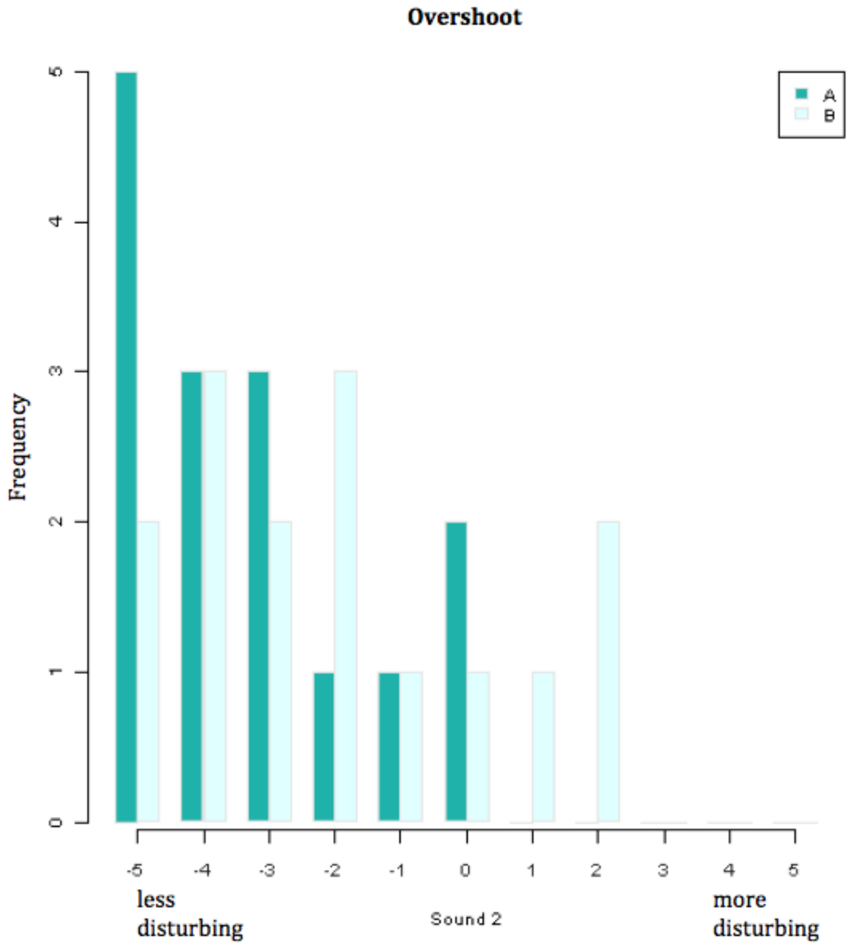

Figure 7.5: Overshoot: Sound 2

 Chapter 7 Appendix

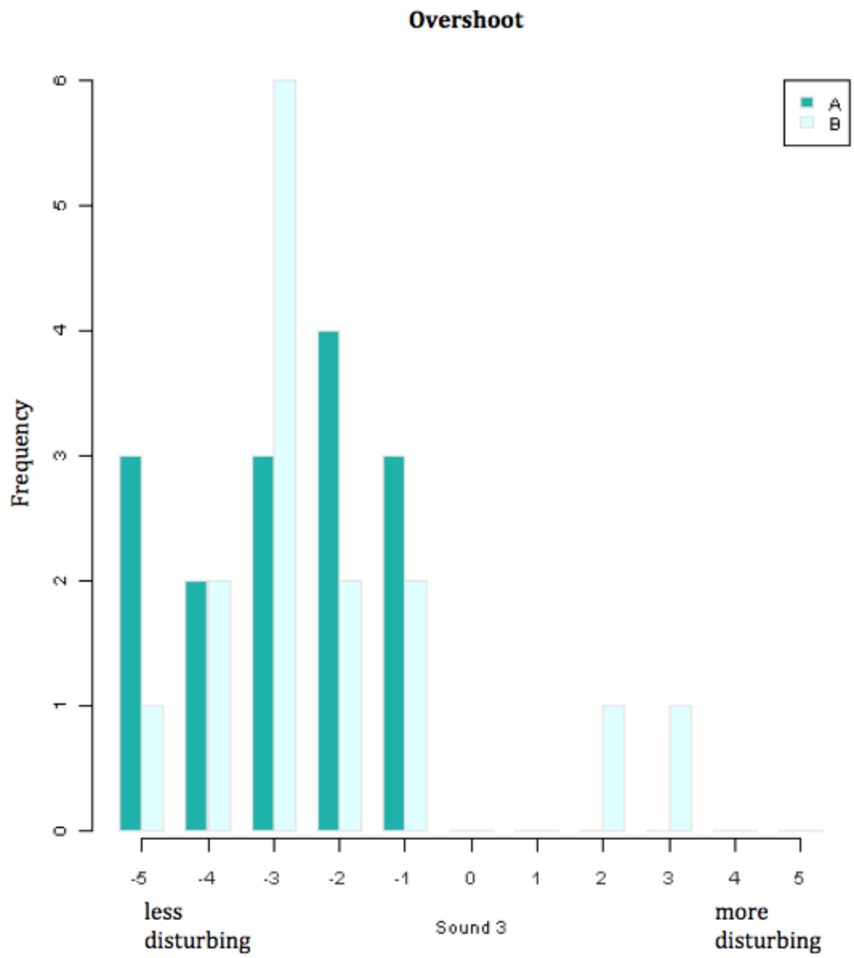

Figure 7.6: Overshoot: Sound 3