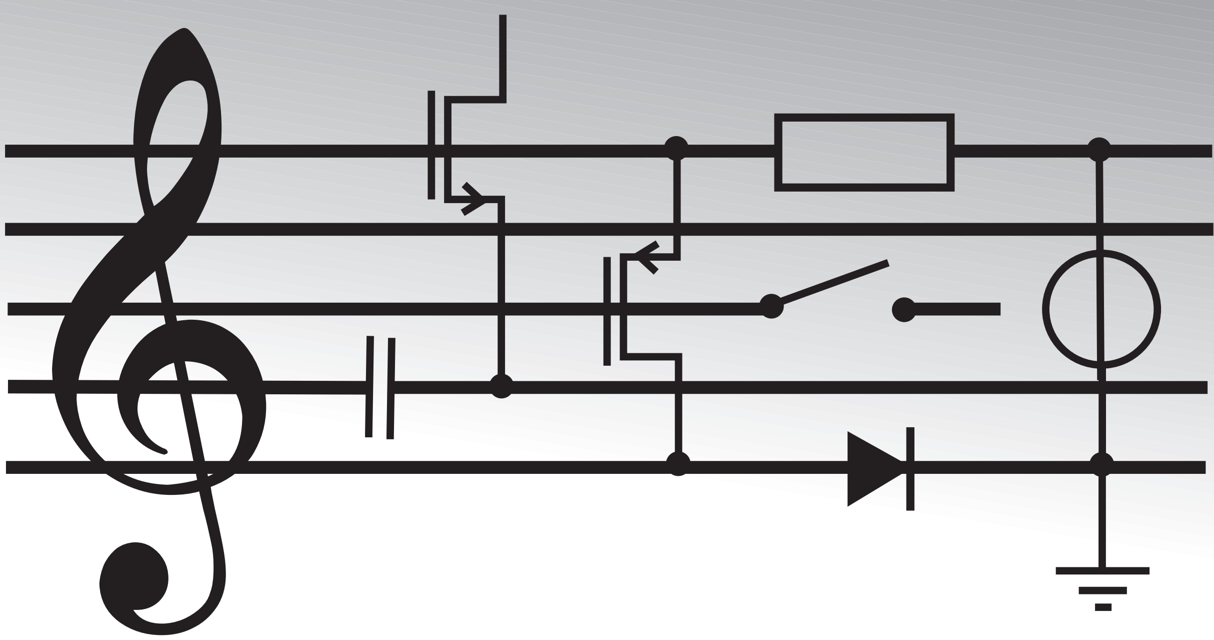
HIGH EFFICIENCY AUDIO POWER AMPLIFIERS
design and practical use

Ronan van der Zee



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HIGH EFFICIENCY AUDIO POWER AMPLIFIERS
design and practical use

Proefschrift

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“The reasonable man adapts himself to the world; the unreasonable one persists in trying to adapt the world to himself. Therefore all progress depends on the unreasonable man.”

-- *George Bernard Shaw*

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Introduction

1.1 Motivation

During the history of audio registration and reproduction, which started more than a century ago, there has been a steady improvement in quality. The first record player, which was in fact a rotating drum, used only the mechanical excitation of the needle to produce sound. The movements of the needle were transferred to a diaphragm in a horn, thus forming a true 'audio amplifier'. Later in time, the movements of the needle were first transformed into electrical signals. These signals were amplified by means of vacuum tubes and fed to a loudspeaker. With the introduction of the transistor, vacuum tubes were replaced by transistors, and later by integrated circuits. These developments led to audio amplifiers with less weight, using less power and sounding better. (Regarding this last aspect it is quite unfortunate that many people are misled by the term 'warm feeling of tube amplifiers', thinking it refers to sound quality rather than to dissipation).

At the same time, the quality of storage media improved. The Phonograph was succeeded by the Gramophone. Analogue magnetic recording developed from steel wire to tape. Noise reduction techniques increased the dynamic range. Over the past years, audio in the consumer domain has essentially become digital. Audio is stored on media like CD (Compact Disc), DAT (Digital Audio Tape), DCC (Digital Compact Cassette), MD (Mini Disk), or DVD (Digital Versatile Disc). Formats include PCM (Pulse Code Modulation) up to 96kHz 24 bit, DSD (Direct Stream Digital), multichannel sound up to 6 channels, and psycho acoustic codecs like Dolby AC-3 (Audio Coding 3), MPEG (Motion Picture Expert Group) layer 1-3 and MPEG-4 AAC (Advanced Audio Coding).

With the introduction of DAB (Digital Audio Broadcasting) and HDTV (High Definition TeleVision), the audio amplifier is one of the few remaining analogue components in the audio chain. This not only means that it has to fulfil high requirements on standard specifications points like distortion, slew-rate, power supply rejection, etc. The large dynamic range of digital signals, for instance, demands high peak power. Classical class AB amplifiers with high peak power, however, have very poor efficiency at moderate signal levels. Also, good bass reproduction is getting more and more important, requiring much power of the amplifier. At the same time, dimensions have become smaller. Mini sets, car radios and PC multimedia equipment have only little space available, leading to an increasing conflict between manageable power dissipation and market demands for high output power and many output channels. Also, more and more equipment becomes portable. In these cases, a low power consumption is necessary to lengthen battery life. To meet these demands, highly integrated, power efficient audio amplifiers are essential.

1.2 Problem definition

1.2.1 Audio Signals

Essentially, an audio amplifier is a normal voltage amplifier optimised for the amplification of audio signals. The limited frequency response of the ear sets the bandwidth limits: 20Hz - 20kHz, although most people are not able to hear 20kHz. Most power is concentrated in the mid frequencies, and occasionally in the low frequencies. Generally, the amplitude probability density function of audio signals is gaussian. This means that the ratio between maximum and average power is large: 10…20dB. In average, it is 15dB (see section 2.3.2), which is 12dB below the power of a rail-to-rail sinewave. Chapter 2 contains a much more elaborate analysis of audio characteristics, which can successfully be used in the design of high efficiency audio amplifiers.

1.2.2 Amplifier Power

The ear has a very large dynamic range. To give an example: the ratio between the acoustic power of a rock concert and the sound of breathing can be as large as 10^{11}. This makes large demands on the dynamic range of the audio amplifier. To get an idea about the order of magnitude of amplifier output powers, refer to Table 1.1. The SPL's have been taken from [57]. Table 1.1 displays some situations in which audio power amplifiers can be used. The first column gives the Sound Pressure Level (SPL) in dB's. 0dB SPL is the hearing threshold and defined as 0.00002 N/m^2. The second column shows what sound sources would produce an equivalent SPL - just to give an idea.

Now suppose we want to reproduce these SPL's with an audio amplifier and a loudspeaker. Assuming that the loudspeaker has an efficiency of 90dB/W@1m (a normal value for large loudspeakers) and that the SPL decreases with the squared distance, the needed loudspeaker power at a certain hearing distance can be calculated (third column). In practice, these values can be a little too high, because of reflections against walls or ceiling.

SPL [dB]	Sound Pressure Level (SPL) is equivalent to:	Power in LS for that SPL @ distance from LS	Necessary amplifier rating
50	Low level background music at 1m	≈ 100μW @ 1m	1.5mW
60	Normal speech at 1m	≈ 1mW @ 1m	15mW
80	Orchestra in concert hall	≈ 1W @ 3m	15W
110	Rock band	≈ 2kW @ 4.5m	30kW

Table 1.1: Amplifier power needed for different sound pressure levels

Finally, audio signals have an average power that is considerably lower than their peak power, so for undistorted sound the maximum sine power rating of an amplifier should in average be 12dB higher than the average power delivered to the loudspeaker. The resulting calculated amplifier peak power is displayed in the fourth column of Table 1.1.

From Table 1.1 we conclude that audio amplifiers must operate over a wide range of power levels. The ratings in column 4 are an indication of the amplifier powers found in transistor radios (100mW-1W), midi sets (10W-100W) and professional PA equipment (1kW-10kW). These values depend on many factors; they are mainly meant to create a feel for amplifier powers.

1.2.3 Efficiency or dissipation?

Although the term ‘amplifier efficiency’ is used numerous times both in this thesis and in literature, the efficiency of an audio amplifier is hardly important in systems that use audio amplifiers. In battery powered equipment, the dissipation should be minimal for the longest battery life time. In systems where cooling is a problem, the maximum dissipation is an important design criterion. In literature, however, the most common measurement graphs depict the efficiency of an audio amplifier as a function of output power as shown in Figure 1.1. A problem with these kinds of charts is that it is difficult to see how much the amplifier actually dissipates. The dissipation of an amplifier in relation to the output power P_o and the efficiency η is:

$$P_{diss} = P_o\left(\frac{1}{\eta} - 1\right)$$

which makes it not very easy to see that the right amplifier in Figure 1.1 dissipates 50% more than the left one at full power (which seriously affects heat sink design).

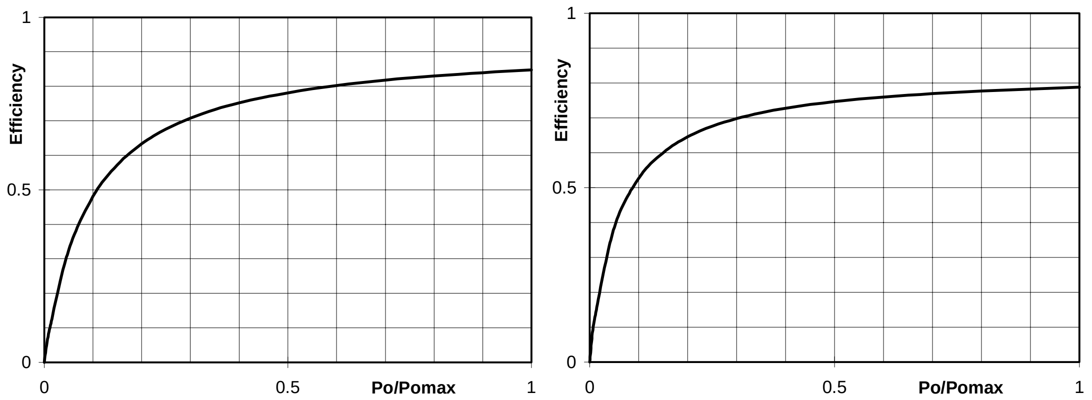

Figure 1.1: Simulated efficiency of two hypothetical audio amplifiers with different quiescent- and maximum dissipations.

The fact that the left amplifier has a 50% higher quiescent power dissipation (which seriously affects the battery life of e.g. a portable radio) is not visible at all, since the efficiency is always zero at zero output power. Apart from that, the average power of an audio signal is on average 12dB lower than a full power sinewave, so the majority of the graph displays useless information. From now on we will therefore use graphs as displayed in Figure 1.2. The dissipation for the whole power range is clearly visible thanks to the logarithmic x-axis, and also the maximum- and quiescent power dissipation can easily be observed.

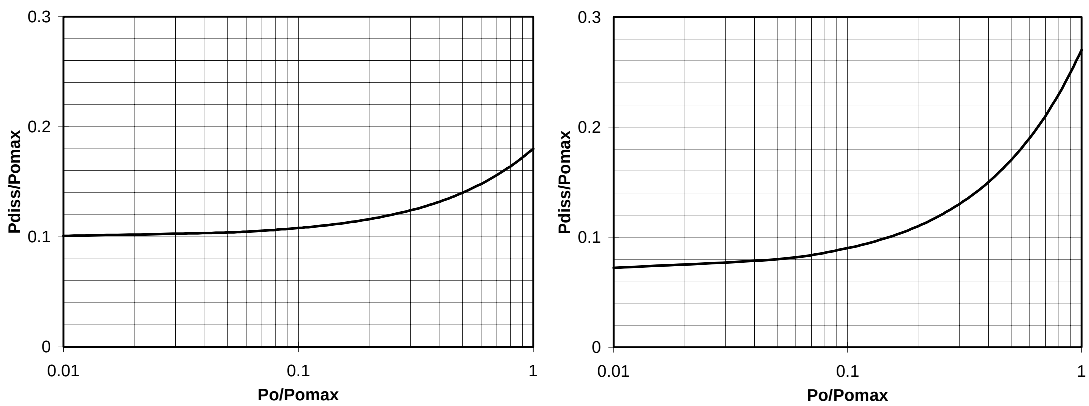

Figure 1.2: Dissipation of the two amplifiers in Figure 1.1.

1.2.4 Distortion

Making a high efficiency audio amplifier would be a lot simpler if its distortion was not important. A class D amplifier on a low switching frequency can have an excellent efficiency, but its distortion will be too high. The design of class G amplifiers is complicated by switching distortion, etcetera. Therefore, a low distortion is an important condition when judging efficiency. There are several types of distortion that can be measured:

Total Harmonic Distortion (THD)

When a sinusoidal signal is applied to a non-linear amplifier, the output contains the base frequency plus higher order components that are multiples of the base frequency. The Total Harmonic Distortion is the ratio between the power in the harmonics and the power in the base frequency. This can be measured on a spectrum analyser. Most distortion analysers, however, subtract the base signal from the amplifier's output and calculate the ratio between the total RMS value of the remainder and the base signal. This is called THD+N: Total Harmonic Distortion + Noise. Normally, the noise will be low compared to the distortion, but the noise of a noisy amplifier or the switching residues in a class D amplifier can give garbled THD figures. For a THD+N measurement, the bandwidth must be specified. For class D measurements, a sharp filter with a 20kHz corner frequency is necessary to prevent switching residues -that are inaudible- to show up in the distortion measurements.

InterModulation distortion (IM)

When two sinusoids are summed and applied to a non-linear amplifier, the output contains the base frequencies, multiples of the base frequencies and the difference of (multiples of) the base frequencies. Suppose a 15kHz sinusoid is applied to an audio system that has a 20kHz bandwidth, and the THD+N needs to be measured. All the harmonics are outside the bandwidth and will be attenuated, resulting in too low a THD+N reading. The same situation occurs when the distortion analyser has a 20kHz bandwidth. In these cases, an IM measurement can be a solution.

The first standard was defined by the SMPTE (Society of Motion Picture and Television Engineers). A 60Hz tone and a 7kHz tone in a 4:1 amplitude ratio are applied to the non-linear amplifier. The 60Hz appears as sidebands of the 7kHz tone. The intermodulation distortion is the ratio between the power in the sidebands and the high frequency tone. Another common standard is defined by the CCITT (Comité Consultatif Internationale de Télégraphie et Téléphonie), and uses two tones of equal strength at 14kHz and 15kHz. This generates low frequency products and products around the two input frequencies, depending on the type (odd or even) of distortion.

Interface InterModulation distortion (IIM)

In this test, the second tone of an IM measurement set-up is not connected to the input, but to the output (in series with the load impedance) [3],[9].

Transient InterModulation distortion (TIM)

When a squarewave is applied to an amplifier with feedback, its input stage has to handle a large difference signal, probably pushing it into a region that is less linear than its quiescent point. When a sinusoid is added to the squarewave, the non-linearity induced by the edges of the squarewave will distort the sinusoid, giving rise to TIM, also called transient distortion or slope distortion [10]. There are many ways of testing TIM and it remains unclear how much it adds to the existing measurement methods. If the maximum input signal frequency during normal operation of an amplifier is limited to 20kHz, a 20kHz full power sinusoid is the worst case situation. When that generates little distortion, TIM will not occur [7].

Cross-over distortion

Cross-over distortion is generated at the moment the output current changes sign. At that moment, the output current gets supplied by another output transistor. The process of taking over generates distortion, visible as spikes in the residual signal of a THD measurement. This kind of distortion is notorious for its unpleasant sound (a small percentage error is quickly noticeable). Because it's usually present around zero amplitude, the impact on small signals can be relatively large.

Which distortion is important?

There is no consensus as to which distortion measurements are essential. In the ongoing search for the critical attributes that determine the 'sound' of an audio amplifier, many other mechanisms can play a role like reactive harmonic distortion [2], the spectrum of the distortion [6], non-linear crosstalk [IEC268-1], memory effects [11], granularity distortion [14], and external influences like speaker cables [5], decoupling capacitors and the phase of the moon. It is unclear to what extent these concepts influence the 'sound' of an amplifier. Also, alternative measurement methods have been described, like measuring the difference between input and output of an amplifier for audio signals [4], or analyse the output signal in Volterra space [12].

Based on experience and for practical reasons, the frequency transfer characteristics and the THD+N over power and frequency range are important. Observing the residual signal in a THD measurement and an IM measurement are also good practice.

1.2.5 Integration on chip

The integration of electronic functions on one chip has several advantages. The lower number of interconnections on PCB (Printed Circuit Board) can make the circuit more reliable and -even more important- cheaper. Furthermore, it goes well with the ongoing miniaturisation of dimensions and weight. This tendency seems both inevitable and desirable. High efficiency audio amplifiers are no exception, and are integrated as much as possible. The number of external components (components that are not integrated) must be kept as low as possible. This is a major criterion for new topologies.

1.2.6 Other specifications

The gain of an audio amplifier is usually fixed, or variable in a small range. The output resistance must be low to ensure a proper control of the loudspeaker. The PSRR (Power Supply Rejection Ratio) must be high to avoid distortion caused by variation of the supply voltage, possibly induced by the circuit itself. These specifications, however, are usually not the bottleneck in realisations.

1.3 Efficiency of the class AB amplifier

Most audio amplifiers nowadays are class AB amplifiers. A low distortion class AB amplifier has a relatively low complexity and requires almost no external components if integrated. The efficiency for audio signals however, is quite low. Assume a class B amplifier with power supplies $+V_S$ and $-V_S$. Load resistance is R_L. Figure 1.3 shows how the amplifier dissipates.

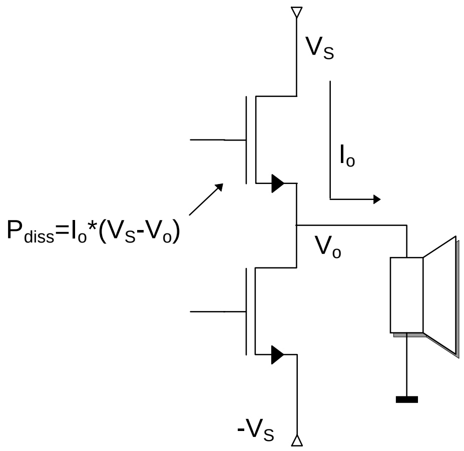

Figure 1.3: Dissipation in a class B output stage for $V_o>0$

The classic way of calculating the efficiency of a class B amplifier assumes a rail-to-rail sinewave at the output. See Figure 1.4. The efficiency over any number of periods is equal to the efficiency over a quarter of a period. The efficiency is defined as $\eta=P_o/P_i$. Over a quarter of a period:

$$P_o = \int_0^{\frac{1}{2}\pi} \frac{V_S{}^2}{R_L} \sin^2\theta \, d\theta \text{, and } P_i = \int_0^{\frac{1}{2}\pi} \frac{V_S{}^2}{R_L} \sin\theta \, d\theta$$

Which yields $\eta = \frac{1}{4}\pi = 78.5\%$. This figure does not really call for improvement, if it were representative for normal use.

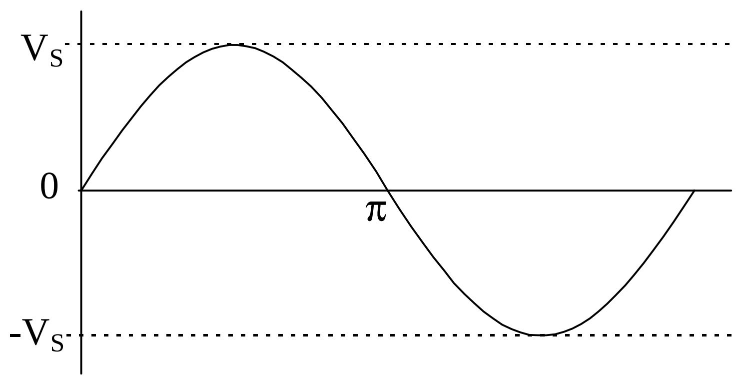

Figure 1.4: Rail-to-rail sinewave

First of all, a practical amplifier can not drive its load to the power supply lines. The voltage difference between the power supply and the maximum output voltage lowers the efficiency. Apart from that, a normal audio amplifier signal is not a sinewave, but music or speech. Finally, the average output power of audio signals is at most half the maximum sinewave power (section 2.3.2). Since an audio signal waveform is complicated and not standardised (see Figure 1.5), a calculation of the efficiency in the same way as above is hardly possible. A possible solution is provided by the use of amplitude probability density functions.

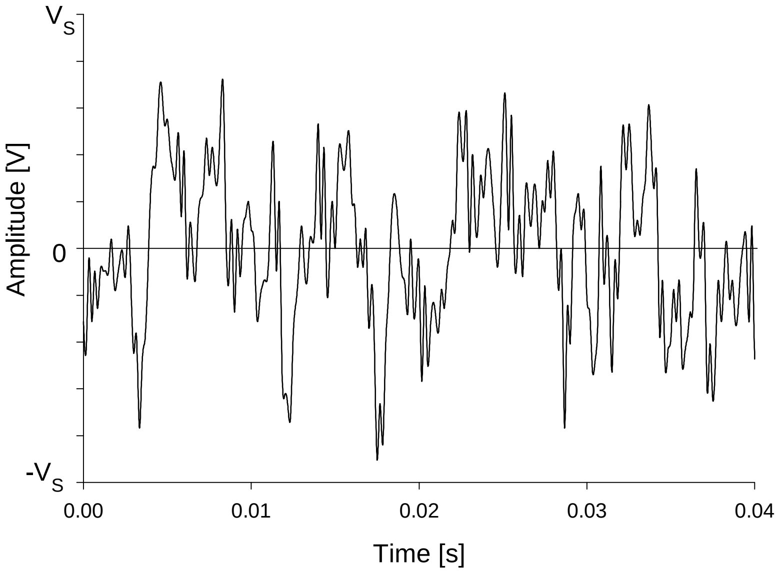

Figure 1.5: A music fragment in the time domain

Therefore an approach is used as described in [24], which uses the instantaneous input- and output power of the amplifier, and the amplitude probability-density function (PDF) of the signal. The instantaneous input power (as a function of output voltage) is defined as the input power of the amplifier at a DC voltage of that value. Idem for the instantaneous output power. The average dissipation and efficiency can be calculated if the PDF(V_o) is known:

$$P_{i,avg} = \int_0^{V_S} P_{i,inst}\left(V_o\right) PDF\left(V_o\right) dV_o \tag{1.1}$$

$$P_{o,avg} = \int_0^{1} P_{o,inst}(V_o)\, PDF(V_o)\, dV_o \tag{1.2}$$

$$\eta_{avg} = \frac{P_{o,avg}}{P_{i,avg}} \text{ and } P_{d,avg} = P_{i,avg} - P_{o,avg}$$

In the previous case of a rail-to-rail sinewave $V_S \sin(t)$, the PDF is:

$$PDF(V_o) \quad = \quad \frac{2}{\pi V_S} \cdot \frac{1}{\sqrt{1-\left(\frac{V_o}{V_S}\right)^2}}$$

Furthermore, it is easy to see that for a class B amplifier

$$P_{i,inst}\left(V_o\right) = \frac{V_o}{R_{LS}} V_S$$

$$P_{o,inst}\left(V_o\right) = \frac{V_o^{\,2}}{R_{LS}}$$

When these are substituted in Equation (1.1) and (1.2), the efficiency can be calculated, which yields of course ¼π. While this calculation was just a complicated way to obtain the same result, the method is advantageous for calculating amplifier dissipations for audio signals. Most audio signals have a PDF that is gaussian, as shown in Figure 1.6.

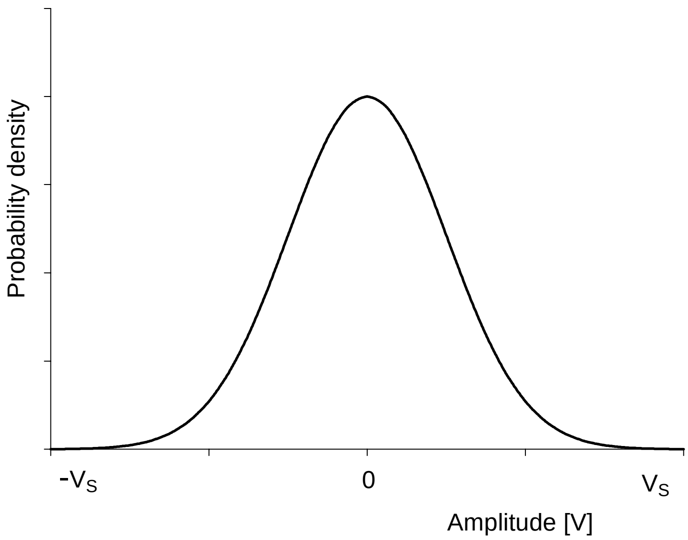

Figure 1.6: A gaussian amplitude probability density function.

By substituting this PDF in Equation 1.1 and 1.2, the dissipation for audio signals can be calculated. Although it is not possible to find an analytical expression in these cases, the equations can be solved numerically. In Figure 1.7, which shows the results of this calculation, we see that the efficiency for music signals is very low. At ½P_{omax} for instance, the efficiency is 60%. Most audio signals, however, are heavily distorted at this output power. For undistorted playback, the average output power of most audio signals must not be higher than $0.1P_{omax}$. At this power, the efficiency is barely better than 25%. Matters get even worse when we look at a realistic situation. Suppose a 100W audio amplifier in a living room, generating 1W electrical power in the loudspeaker, which could already be described as ‘loud’ (see Table 1.1).

Figure 1.7 shows us that the amplifier dissipates 10W, which boils down to an efficiency of 9%. Assuming a non-zero saturation voltage of the output stage and at least some quiescent power dissipation, it becomes clear that the efficiency of the average class AB amplifier leaves a lot to be desired.

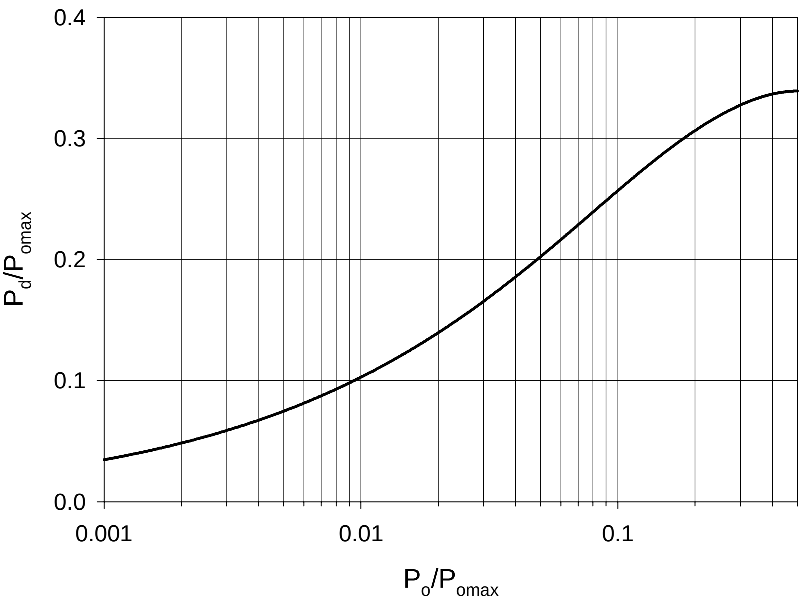

Figure 1.7: Dissipation of an ideal class B amplifier for signals with a gaussian amplitude distribution.

1.4 Scope and Outline

In this Chapter we have clarified the need for high efficiency integrated audio amplifiers. It was shown that the efficiency of a class AB amplifier is quite low and that it needs improvement. Therefore, the scope of this work is to investigate amplifier principles that have a higher efficiency than the traditional class AB amplifier.

In Chapter 2, problems with measuring and predicting amplifier dissipation are discussed. Audio signals are very inconvenient test signals and sinewaves are not representative for audio. Therefore many audio fragments are analysed with respect to amplitude- and frequency distribution. Existing and new test signals are compared to these characteristics to determine if they are suitable to compare present amplifiers, and to predict how new ones will perform.

Chapter 3 gives an overview of already existing high efficiency amplifiers. They can be divided into linear amplifiers and switching amplifiers. The advantages and crucial limitations of both types are analysed.

Chapter 4 takes a look at combinations of linear and switching amplifiers. These combinations can benefit from the qualities of both types. It appears that the combinations described in literature must have a large chip area for their output power. Other combinations are introduced that do not suffer from this.

Chapter 5 described the realisation of a class AB/D bridge amplifier. It consists of a linear amplifier and a switching amplifier in bridge configuration. The linear amplifier guarantees a low distortion. A common mode output voltage control circuit ensures that the linear amplifier has a low dissipation, resulting in a high total system efficiency. The circuit principle is discussed, as well as design choices, the realisation and the measurements.

Chapter 6 describes the realisation of a class AB/D parallel amplifier. It consists of a class AB amplifier and a switching mode amplifier, both connected to the output. The switching amplifier reduces the output current of the class AB amplifier. Thus, the new amplifier combines the high efficiency of class D designs with the low distortion of class AB amplifiers. There is a trade-off with respect to power bandwidth and switching frequency. A slow switching prototype and a fast switching prototype are realised and compared.

Chapter 7 contains the conclusions. What has been achieved? How does it compare to other work in this field? What are the future developments?

Several chapters in this thesis are based on published work. Chapter 2 is based on [19], Chapter 5 is based on [52]. Publications [53-55] have served as a basis for Chapter 6.

2

Measuring and predicting amplifier dissipation

2.1 Introduction

It is important that the efficiency of audio amplifiers is measured correctly. Good test signals and adequate measurement procedures are crucial to make fair comparisons between amplifiers and reliably predict the dissipation in practical situations. This is also a vital condition for judging the usefulness of new amplifier topologies.

The first section describes the problems associated with measuring amplifier efficiency. After that, a test-set of audio fragments is selected and the characteristics of those fragments are investigated. The next section studies how suitable the IEC-268 test signal is for measuring and predicting the dissipation of audio amplifiers for these audio fragments. In the last two sections, alternative signals are proposed that can prove valuable in testing and simulation environments.

2.2 How to measure?

In literature, the efficiency of amplifiers is usually measured with sinusoidal signals. For amplifiers based on a class D topology, this gives approximately the same results as for audio signals, as long as one bears in mind that the average output power of an audio amplifier while playing normal audio signals is much lower than its maximum sine output power. Some high efficiency audio amplifiers, however, need specific audio characteristics to obtain a high efficiency. Well known topologies in this field are the class G and class H principles. The amplifiers described in [21-26, 44, 50, 53] all use knowledge about either the amplitude- or the frequency distribution of average audio signals. For this kind of amplifiers, measurements with sinusoids can give pessimistic results.

The best signal would be a real audio signal, but this has several disadvantages. The question is which audio signal should be taken. Speech? Music? What kind of music? This is not standardised. Furthermore, at least several seconds of audio are necessary to get a good impression, which is not very practical for simulations. Also, a music signal does not give stable readings on meters. In practice, more creative ways were found. Either the efficiency was measured indirectly by measuring heat sink temperatures [21], or an ad hoc measure is defined [50]. Another possibility is to use the IEC-268 ‘simulated programme material’ [53]. In [16] and [17], the spectral distributions of programme material were measured, and the latter also investigated whether the IEC test signal is useful for evaluating the power rating of loudspeakers. There is, however, no standard test signal intended for measuring or predicting amplifier efficiency. In this chapter, we try to find such a signal.

2.3 Characteristics of audio signals

2.3.1 The test set

In order to compare test signals to realistic audio signals, it is necessary to define a test set of audio fragments. Due to the variation in volume in audio signals, the statistical parameters depend on the length of the time interval that is being analysed. Figure 2.1 shows the amplitude distribution of complete CD tracks. Compared to shorter fragments with constant volume (see Figure 2.2), we notice a somewhat larger spread and a clearly different shape which peaks around zero amplitude. This is a result of the sections with a lower volume.

Now suppose we would use the distributions of Figure 2.1 to predict amplifier dissipation. Such a signal has a certain average power that has to be delivered by the amplifier, leading to a certain (predicted) average dissipation. During the loud passages, however, the amplifier has to deliver considerably more power, and when they last longer than the heat sink’s thermal time constant, the amplifier will overheat. Therefore, we have chosen audio fragments with constant volume. Of course it should be noted that ‘constant’ is a relative measure, since the audio waveform itself is not constant. It is assumed that variations in less than seconds will not give rise to the problems described above.

We have chosen 80 fragments from various CD’s, including classical music, pop music, jazz, hard rock, house, heavily compressed music, and speech signals. The length of each fragment is between 3 and 12s. The volume during each fragment is constant. All fragments were converted to mono and normalised to full scale, with the highest sample just clipping. The number of bits per sample was reduced to 8 to get smoother amplitude distributions. Because the fragments are normalised to full scale, this barely affects the sound impression.

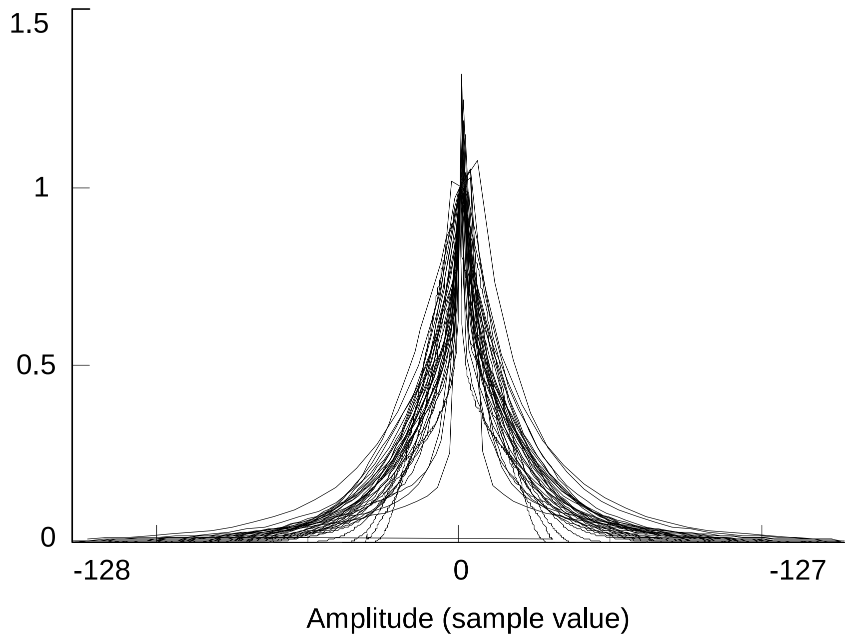

Figure 2.1: Amplitude distributions of 36 CD tracks, normalised to 1@amplitude=0 and then scaled to equal power.

2.3.2 Amplitude distribution

The amplitude distribution is determined by counting how many samples with a certain amplitude (2^8 = 256 levels) occur in one fragment. Figure 2.2 shows the amplitude distribution of all 80 fragments.

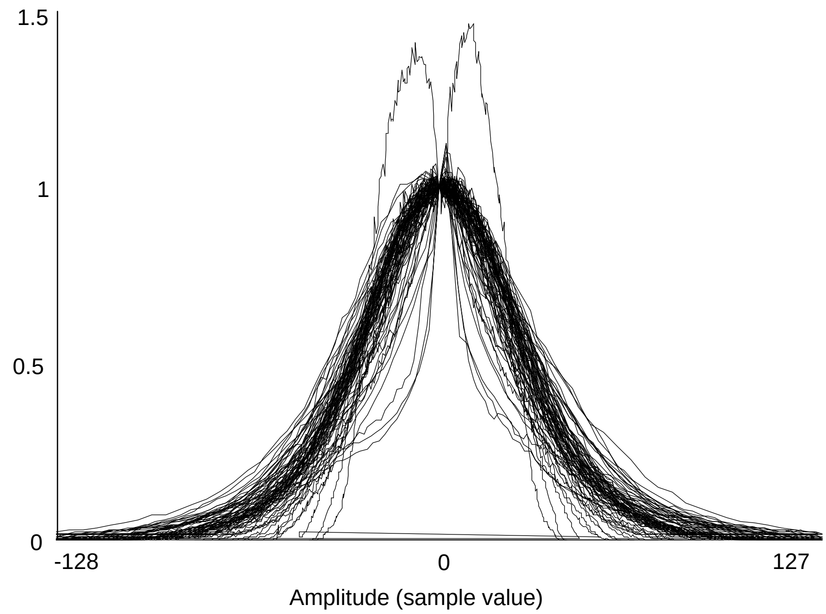

Figure 2.2: Amplitude distributions of all fragments, normalised to 1@ amplitude=0 and then scaled to equal power.

It confirms that the shape of the amplitude distribution is gaussian [16, 24]. There are a few exceptions, though. Firstly, one curve has two peaks symmetrically around zero amplitude. This is the distribution of a fragment hard-core house music, that contains purely synthesised sounds. Although this is an exceptional case, it shows the importance of realising that certain audio characteristics can differ significantly from the average case. Secondly, we see some very narrow curves. These are the distributions of speech signals. Due to the pauses inherent to spoken word, the distributions peak around zero amplitude.

When discussing amplitude distributions, it is useful to critically examine the Peak-to-Average Ratio (*PAR*) [16, 24]. It is widely acknowledged as a signal property, and identical to the traditional crest factor. Expressed in dB's, the *PAR* is defined as:

$$PAR = 20\log\left(\frac{U(t)_{\max}}{U_{RMS}}\right)$$

Figure 2.3 shows the PARs of all fragments. Roughly, it is between 10dB and 20dB, with an average of 15dB. This means that -in order to be undistorted- the average audio fragment must have a power at least 12 dB below a full power sinewave.

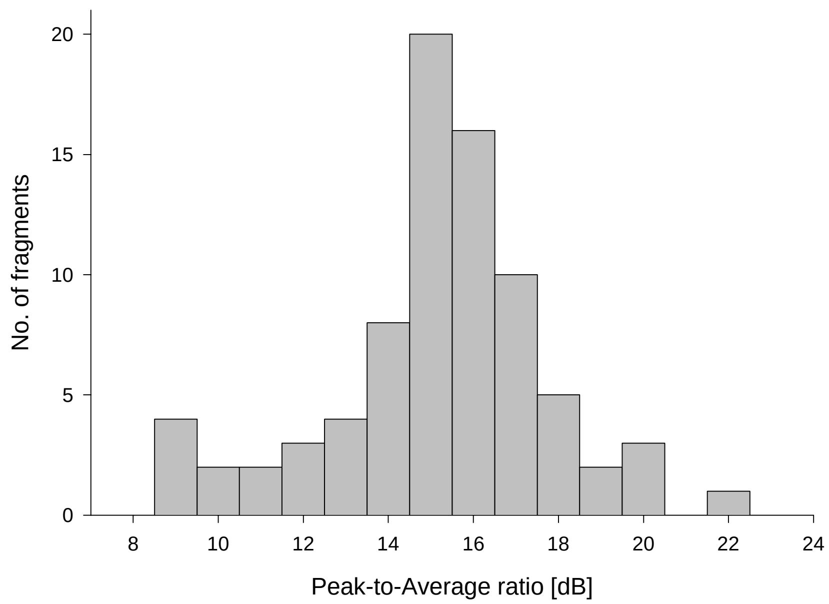

Figure 2.3: Peak-to-Average ratios of all fragments.

Often, the *PAR* is also used for calculating amplifier efficiencies, resulting in a certain efficiency for a certain *PAR* of the signal. In that case it is assumed that every fragment is amplified to a level just below clipping. The result is that the amplifier dissipation strongly depends on the *PAR*. The reason for this is, that the average power (or U_{RMS}) also varies considerably, since $U(t)_{max}$ is the clipping point of the amplifier and therefore constant. In Figure 2.2, however, it can be seen that, when scaled to equal power, the amplitude distributions are almost the same. $U(t)_{max}$ varies, but since the high amplitudes near $U(t)_{max}$ are unlikely to occur, they hardly effect the total dissipation of the amplifier. When a fragment with a large *PAR* is amplified to equal power as a fragment with a low *PAR*, there will be some clipping, but this is barely perceptible in normal listening conditions. Only when we increase the volume a lot, the sound quality degrades. Subjective listening tests show that the *PAR* can be made as small as 6dB before most fragments sound really bad through clipping. A *PAR* of 6dB means that the output power is half the maximum sine power. From the above we conclude the following: Audio fragments of constant volume generally have a gaussian amplitude distribution with an average PAR of 15dB. Concerning amplifier dissipation, average power is the most important variable, while the PAR does not play a significant role. Amplifier dissipation for gaussian signals must be tested up to half the full sine power.

2.3.3 Frequency distribution

On the same audio fragments, a Fast Fourier Transform (FFT) was performed over the full length. A normal log-log bode plot of the frequency content (Figure 2.4) does not provide very useful information.

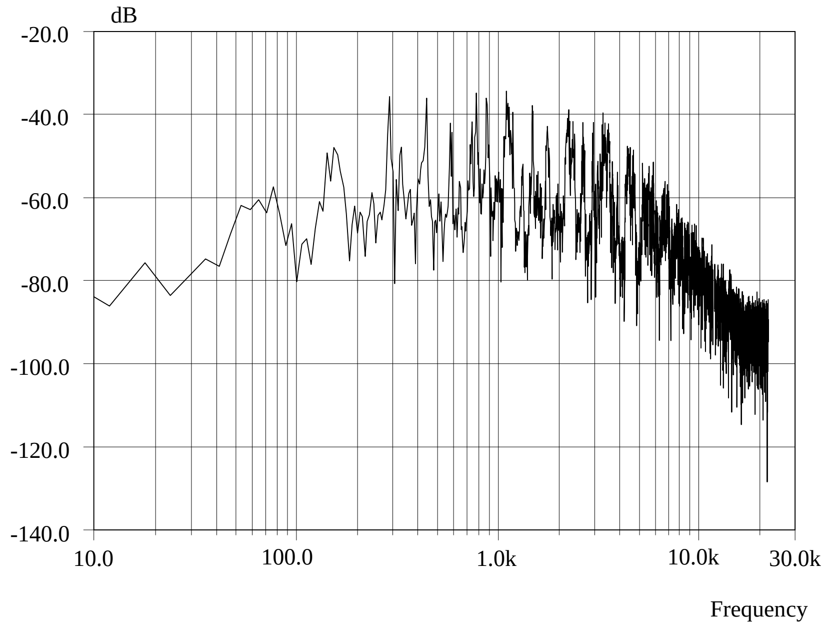

Figure 2.4: Traditional graph of a Fourier transform of a music fragment. Vertical scale dB's are relative to full scale for measurement bandwidth $2/T_{fragment}$.

Firstly, there is no need for a high accuracy, so it seems more logical to choose the vertical scale of the plot linear instead of logarithmic. Secondly, efficiency is a matter of power. When an amplifier has a better efficiency for certain frequencies, it is important to know how much power is present in those frequencies, not how much amplitude. So it's more useful to square the amplitudes. Finally, the squared

FFT gives the power of the frequencies in the signal. The frequencies are linearly spaced. With a logarithmic frequency axis, a temptation exists to overemphasise the lower frequencies because they are relatively enlarged. A linear frequency axis might seem a logical choice, but since pitch perception is logarithmic in nature (every octave higher equals a factor two), it is preferable to use a logarithmic axis, and plot the sum of the squared Fourier coefficients. An extra advantage is that the summation smoothens the curve.

Presented in this way, the frequency distribution is a line that starts at (almost) power = 0 at 20Hz, climbing to power = 1 at 20kHz. The frequency distributions of all fragments are shown in Figure 2.5. The average fragment is S-shaped, with a mid-frequency part corresponding to a straight line between (50Hz,0) and (3kHz,1). This does not come as a surprise when we realise that the notes in a musical scale are fixed factors in frequency apart, in which case a linear frequency distribution requires all notes to be equally loud. In Figure 2.5, the fragments with much power in the lower frequencies have a house beat or a contrabass. The fragments with much power in the higher frequencies mostly have electric guitars or synthesisers. One fragment in particular stands out because it contains much more high frequencies than the others. It is the intro of Melissa Etheridge's 'Like the way I do', containing a guitar and a tambourine.

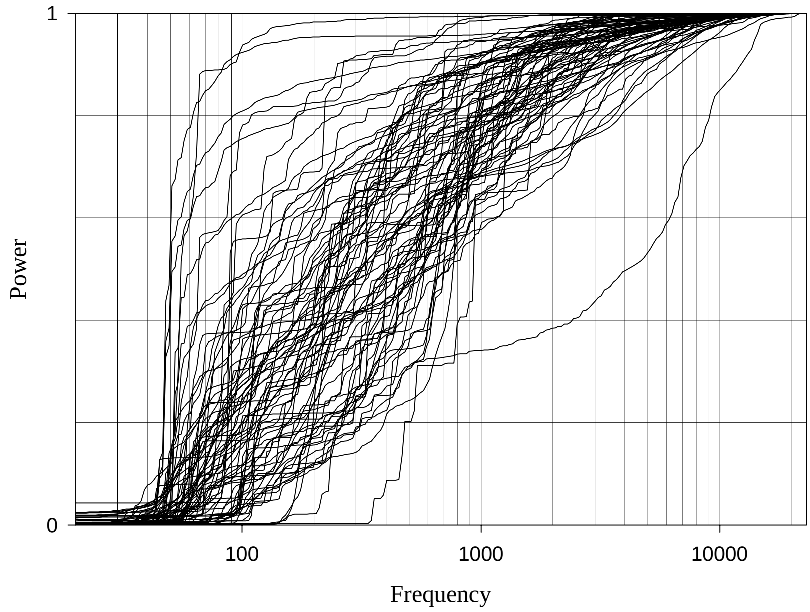

Figure 2.5: Frequency distribution of all audio fragments.

2.4 The IEC-268 test signal

The International Electrotechnical Commission (IEC) has defined a noise input signal representative for normal programme material [18]. It is generated by a pink or white noise source followed by a filter. We will refer to this signal as the 'IEC signal', and investigate if it is useful for efficiency measurements.

2.4.1 Characteristics

Figure 2.6 shows that the amplitude distribution of the IEC signal is gaussian.

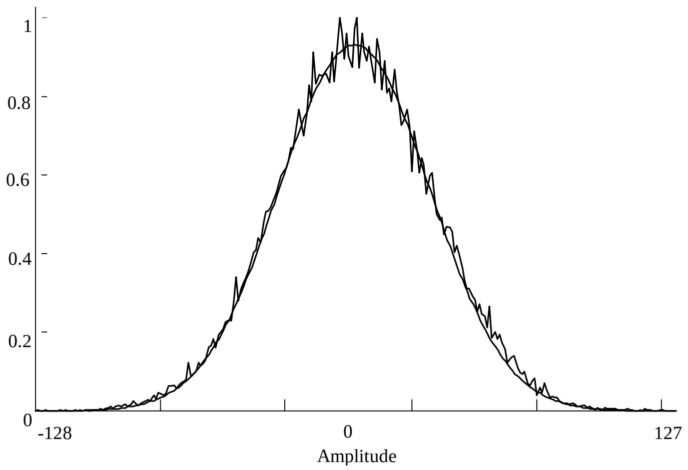

Figure 2.6: Amplitude distribution of the IEC-268 test signal and a gaussian curve as reference

Figure 2.7 shows the IEC signal frequency distribution, together with the distribution of the fragments. The IEC signal serves well as a typical audio fragment.

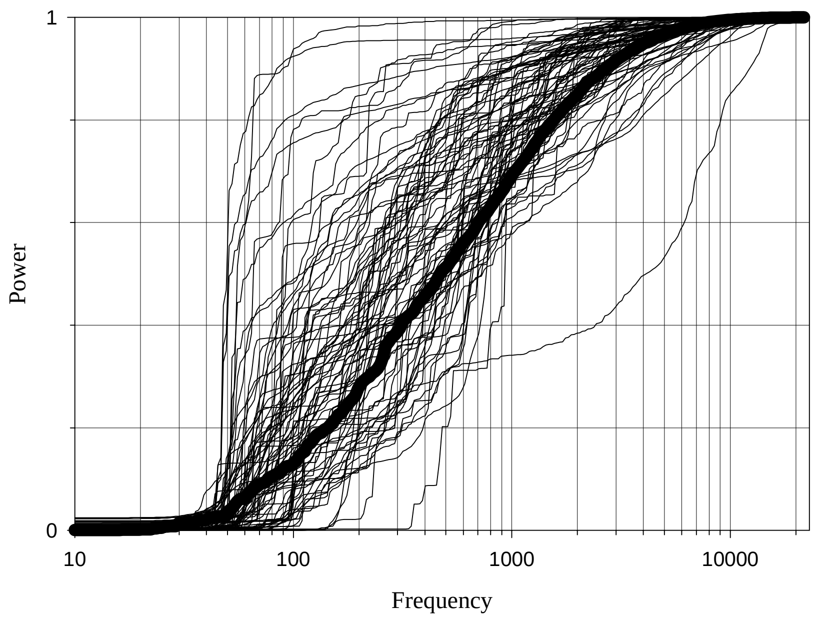

Figure 2.7: Frequency distribution of the fragments and of the IEC test signal (fat line).

2.4.2 Completeness

Despite the good characteristics of the IEC signal, one can still wonder if these characteristics are complete; do they fully determine amplifier dissipation? To answer this, three amplifiers were built: a standard class AB amplifier, a class H amplifier (an amplifier that lifts the power supply during signal peaks by means of an electrolytic capacitor [22]), and a class D+AB amplifier (an amplifier that has a class AB and a class D amplifier in parallel [53]). The class AB amplifier is only sensitive to the amplitude distribution of its output signal; the frequency is not important. The class H amplifier dissipation is -to some extent- frequency dependent, because charging and discharging the capacitor is not lossless, so the total dissipation of this amplifier depends on both the volume and the frequency of the output signal. Finally, the class D+AB amplifier is also sensitive to both characteristics. The class AB part in this amplifier has to support the output current for high $d(I_{out})/dt$, starting at 1kHz full scale signal swing. All amplifiers have a maximum output power of 30W, and have identical heat sinks.

Input to the amplifiers are both the IEC signal and a music fragment that is selected because it has almost identical characteristics (a fragment of ‘Me and Bobby McGee’ by Janis Joplin). We measured the heat sink temperatures as a function of time of all amplifiers. The results are depicted in Figure 2.8. The average output power was 2W, at which the amplifiers were clipping a negligible part of the time. The difference in dissipation between the two signals is insignificant. When the average output power is increased to 10W, the music and the test signal are clipping a considerable part of the time. Even then, there is hardly any difference between the two, as is shown in Figure 2.9. The differences that do occur can be explained by measurement inaccuracies or slight differences between the amplitude distributions.

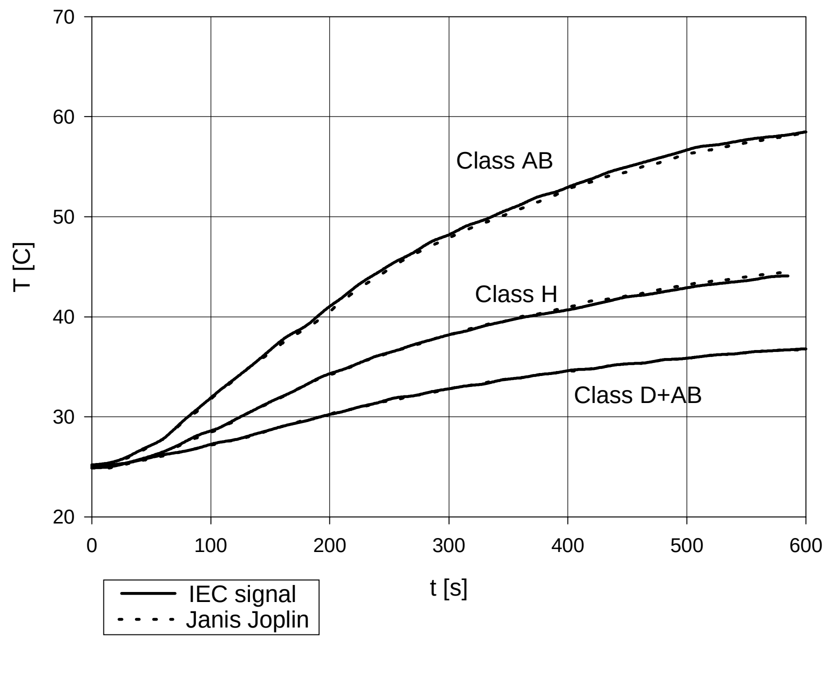

Figure 2.8: Heat sink temperatures for three amplifier classes and two signals at an average output power of 2W (no clipping).

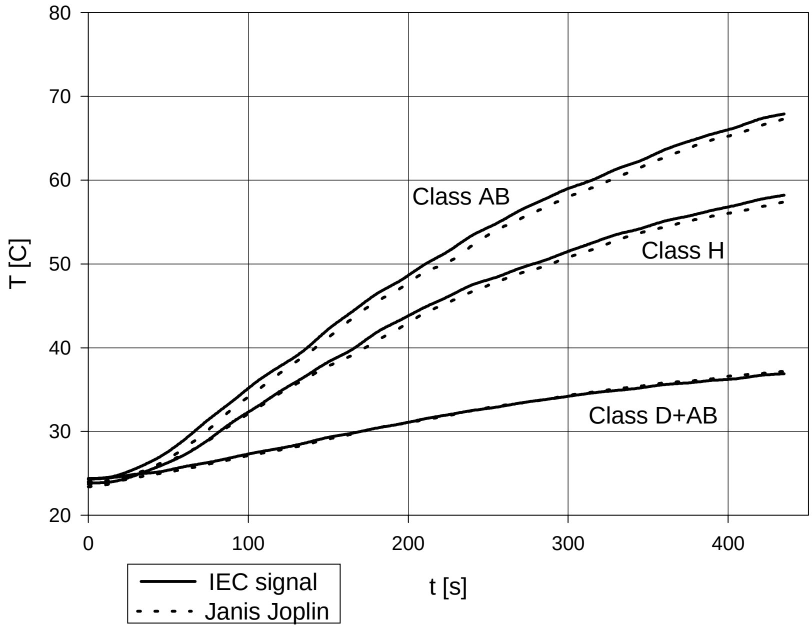

Figure 2.9: Heat sink temperatures for three amplifier classes and two signals at an average output power of 10W (heavy clipping).

With these results it seems that the amplitude- and frequency characteristics fully determine amplifier dissipation, also under clipping conditions. Thus we can trust that the dissipation of audio fragments with the same characteristics as the IEC signal will also cause the same dissipation.

2.4.3 Accuracy

Although the IEC characteristics are a good average, individual fragments can have characteristics that are quite different. The question arises if these fragments produce amplifier dissipations that are also significantly different. To answer this question, it is necessary to measure the dissipation of the three amplifier classes for all audio fragments. Direct measurement of amplifier efficiency for audio signals, however, is difficult. One possibility is measuring the heat sink temperature, as was done in the previous section. This requires a constant ambient temperature and is very time consuming. Another (complicated) possibility is sampling the output voltage and the supply current, and calculate the dissipation.

To circumvent these drawbacks, we used behavioural models of the amplifiers, and simulated the dissipation with C programs, evaluating the dissipated energy per audio sample. With the models described in section 3.2.5 and section 6.3.2, it is easy to calculate the dissipations for the various audio fragments. The models were developed with the IEC signal measurement results as reference. To demonstrate the validity for real audio signals, Figure 2.10 shows the simulated dissipation for both the IEC signal and the fragment of Janis Joplin. The dissipations are practically the same, as they should be. Furthermore, the ratios between amplifier dissipations at 2W and 10W deviate less than 15% from the ratios of the extrapolated increase in heat sink temperatures of Figure 2.8 and Figure 2.9.

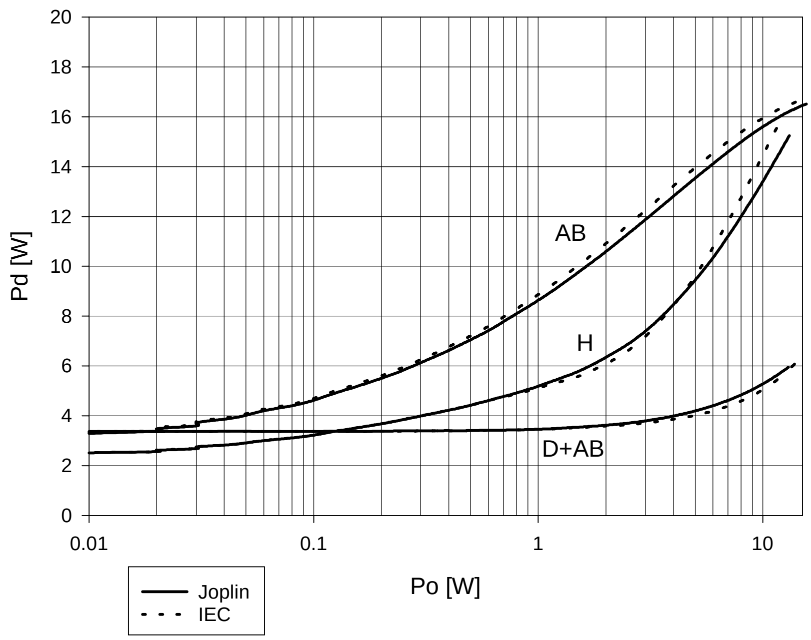

Figure 2.10: Simulated dissipation of three amplifier classes for the IEC test signal and a fragment of Janis Joplin.

After all audio fragments were scaled to equal power, the dissipation they caused was calculated for all amplifier classes. Figure 2.11 shows the results as a histogram. It has a logarithmic x-axis. The distance between the left border and the right border of each bar is a factor 1.05. The height of the bar indicates how many audio fragments cause a dissipation in that range. The vertical lines indicate the dissipation for the IEC signal.

It appears that all fragments have dissipations within +/- 20% of the dissipation predicted by the IEC test signal. One fragment stands out because it causes a high dissipation in both the class D+AB and the class H amplifier. It is the intro as discussed in section 2.3.3. The large high frequency contents decreases the efficiency of the two amplifiers. Although this is an exceptional case, it is important to realise that the good predictive qualities of the IEC signal might not be valid for an amplifier which is more sensitive to the frequency contents of its input signal. In general, however, the IEC signal is representative for a wide range of audio signals.

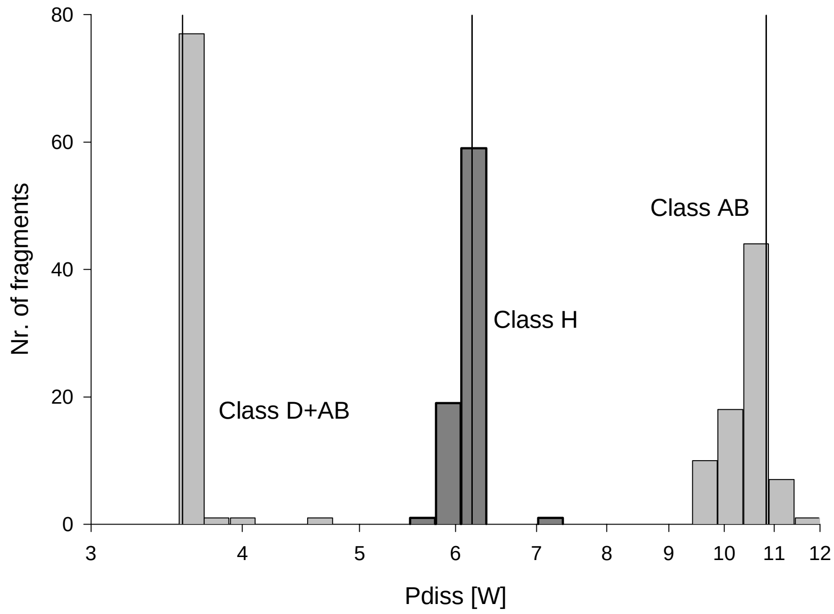

Figure 2.11: Histogram of the simulated dissipation of all audio fragments in three amplifier classes. Vertical lines indicate the dissipation for the IEC signal.

2.5 A simple periodic test signal

The IEC signal is suitable for measuring the efficiency of audio amplifiers, but for prediction purposes it is less ideal. The signal is difficult to generate when the efficiency of an amplifier has to be simulated in a circuit simulator. In circuit simulators, transient noise sources are rarely available, and usually take a long simulation time. Also, for long term testing (reliability), a noise generator is often not available. A simple, periodic test signal would be welcome. The most important quality is a controlled amplitude probability density function. Suppose the signal is V = f(t), and that it is monotonously rising on $t \in [0,t_1]$. The distribution function is the chance that f(t) is smaller than a certain value V, is:

$$F(V) \quad = \quad P(f(t) < V) \quad = \quad \frac{f^{-1}(V)}{f^{-1}{}_{\max}} \quad = \quad \frac{f^{-1}(V)}{t_1}$$

The probability density function is the derivative:

$$f(V) \quad = \quad \frac{\delta F}{\delta V} \quad = \quad \frac{1}{t_1} \cdot \frac{\delta f^{-1}(V)}{\delta V}$$

So if we want to design a signal with a gaussian amplitude probability density function, we know that f(V) is a gaussian curve. Then, $f^{-1}(V)$ is the integral of a gaussian curve, which is the normal distribution function. Thus, f(t) must be the inverse of the normal distribution function. Figure 2.12 shows a possible time-function.

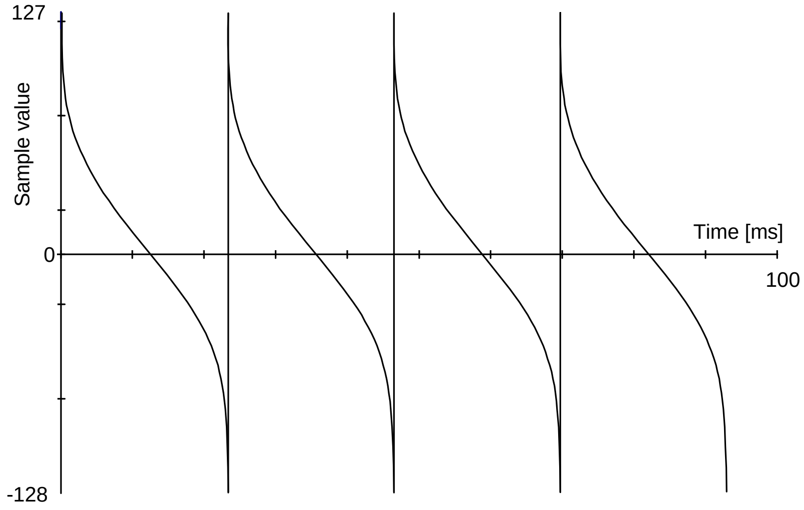

Figure 2.12: Signal with a gaussian amplitude distribution

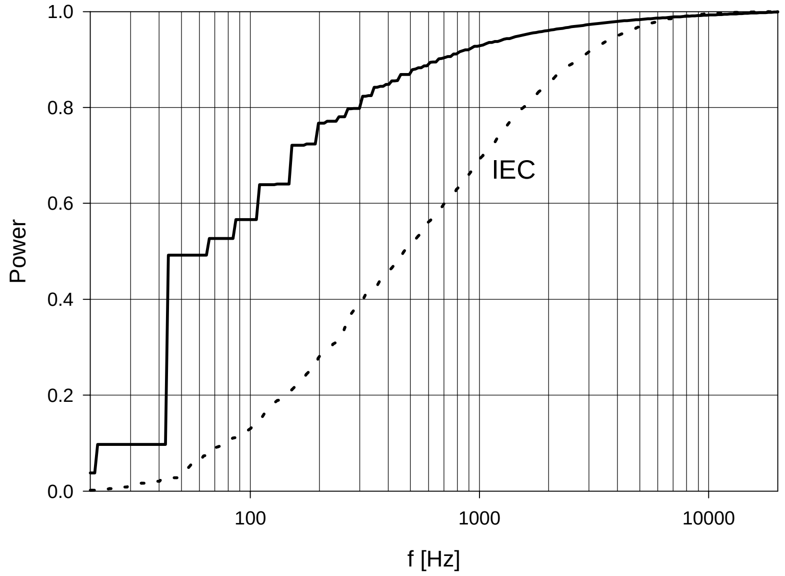

Figure 2.13: Frequency distribution of a signal with gaussian amplitude distribution (Figure 2.12), and of the IEC signal.

Unfortunately, the corresponding frequency distribution, shown in Figure 2.13, is not OK. The higher frequencies are relatively weak. Although the period time of the signal could be chosen a little shorter, the frequency distribution can never match that of the IEC signal.

Synthesising a signal that also has a controlled frequency distribution is not straightforward. The signal in Figure 2.12 or a $1/\sqrt{t}$ signal (which has a $1/f$ power distribution) can be filtered to produce the signal in Figure 2.14. This signal has a correct frequency distribution, but the amplitude distribution is too wide due to much power in the high amplitudes. Also, the frequency distribution alters when the signal clips, so testing for higher output powers is not possible. When the frequency distribution is of prime importance, the signal of Figure 2.14 can be useful; in practice its application will be very limited.

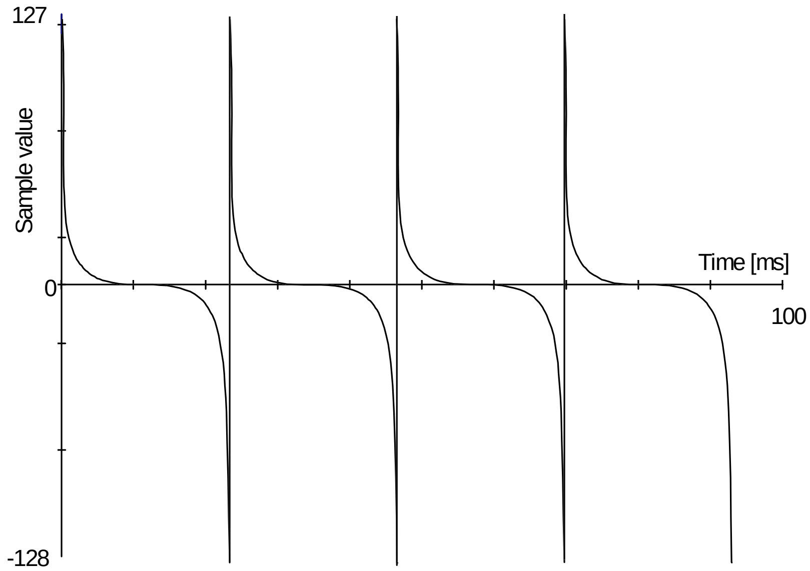

Figure 2.14: A signal with an IEC 268 frequency distribution.

Conclusion: It seems hardly possible to construct a simple periodic test-signal that has all the properties we need to simulate music and speech. Often, however, the amplitude distribution is the most important property. Class G amplifiers, for instance, are not very sensitive to the exact frequency of their output signal. In that case, the signal of Figure 2.12 is advantageous owing to its very short repetition frequency. In a circuit simulator, a single period will suffice to give a good dissipation prediction.

2.6 An IEC variant

The main problem with the IEC signal lies in its need for a noise source. Therefore, a new test signal is proposed that is equivalent to the IEC test signal. The noise source is replaced by 24 squarewaves of equal amplitude, all a factor $\sqrt{2}$ in frequency apart. In the frequency range 10Hz…28kHz, this simulates pink noise, since the energy per octave is constant. This semi pink noise is filtered to get the IEC frequency characteristics. An additional advantage is that the signal, which had only 24 possible amplitude values, now becomes continuous.

Only 100ms of simulation with this IEC variant suffice, since frequency components below 10Hz are not present. The squarewaves are easy to define in a circuit simulator, which will speed up simulations. Also, such a signal can easily be generated in hardware with binary counters or with IC's that are used as tone generators in electronic organs. To see if this IEC variant is indeed equivalent, the dissipation curves for the three amplifier classes were measured, this time for the IEC signal and its variant. The results in Figure 2.15 show that the dissipations are almost the same at low output powers. At high output powers, the results differ more. Since heat sink temperature measurements on more than one high power music fragment are not available, it remains unclear if this error is the same for all music fragments at that output power. However, the differences are rather small and only occur when the signal is heavily clipping. At lower, more usual output powers, the IEC variant gives good results.

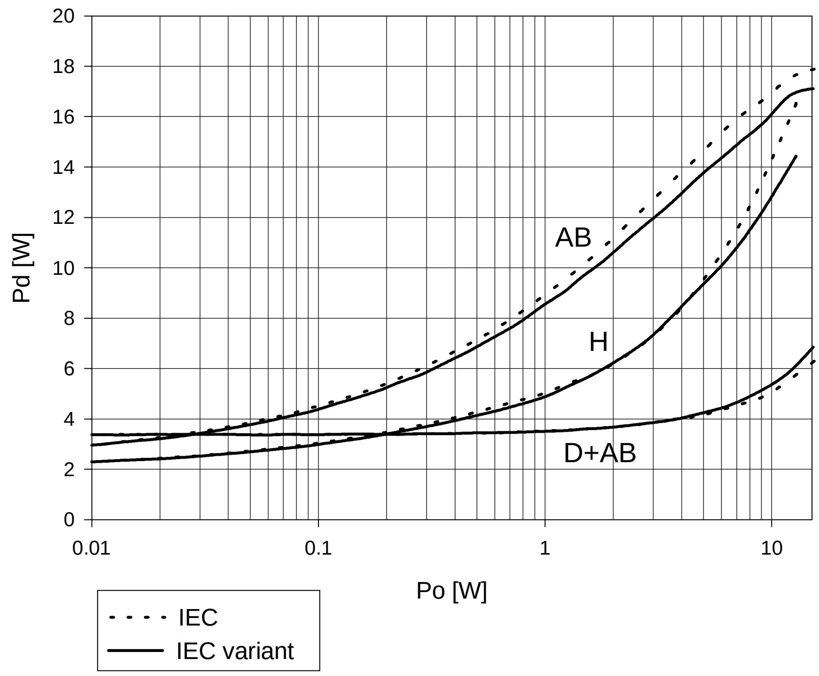

Figure 2.15: Measured dissipation of three amplifier classes for the IEC signal and the IEC variant.

2.7 Conclusions

For the tested types of high-efficiency amplifiers -a class AB, a class H, and a class D+AB amplifier- the power, the amplitude distribution and the frequency distribution of the output signal fully determine the amplifier’s dissipation. The Peak-to-Average ratio of the signal is not very significant.

The dissipation for a variety of real-life audio signals of constant volume deviates only 20% from the dissipation caused by the IEC 268 test signal at the same output power. Therefore, this signal is very suitable for measuring audio amplifier efficiency. This must be verified for new amplifiers types, that may be more sensitive to amplitude- or frequency distribution deviations.

Two alternative test signals are proposed. For simulation and test purposes, a simple test signal can be used for amplifiers with near frequency independent dissipation. When the frequency contents is also important, an IEC look-alike test signal can be used which has the same characteristics as the IEC signal, but is easier to generate in simulation and hardware.

3

Linear and switching amplifiers

3.1 Introduction

In this chapter, a summary is given of different principles that have been used to obtain audio amplifiers with a high efficiency and a low distortion. Of each principle (linear amplifiers and switching amplifiers) the theoretical aspects are investigated. Different examples of each principle are discussed, along with the limitations of practical realisations of such amplifiers.

3.2 Linear amplifiers

With linear amplifiers, we mean amplifiers with a linear output stage, in which there exists a voltage drop across the output transistors to generate the correct output voltage. Even though most of these amplifiers use some sort of switching, they are not to be confused with switching amplifiers, which will be discussed in section 3.3.

3.2.1 Class G

In Chapter 1 it was shown that a class B amplifier has an efficiency of 78.5% for a rail-to-rail sinusoid. This figure is relatively good because the output signal is close to the supply lines a considerable part of the time, with a limited voltage drop across the output transistors. The output signal for audio, however, is close to zero most of the time, with only few excursions to higher levels. Thus the average voltage drop across the output transistors is large, causing the poor efficiency figures for audio.

An amplifier in class G uses multiple supply voltages. At lower power levels, the lower supply voltage is used. When the signal becomes too large for this supply, the higher power supply takes over, and delivers the output power. In this way the average voltage drop across the output transistors is reduced and the overall efficiency can be improved [23, 24, 26]. There are two basic ways in which class G amplifiers are realised. The difference is the way of switching between the supply voltages. Figure 3.1 shows the upper half of a possible output stage. The upper transistor is switched on during signal peaks increasing the power supply of the lower transistor -that controls the output voltage- from V_{dd1} to V_{dd2}. Another way to use this circuit is opening the lower transistor totally during signal peaks, giving the higher MOST the role of output transistor.

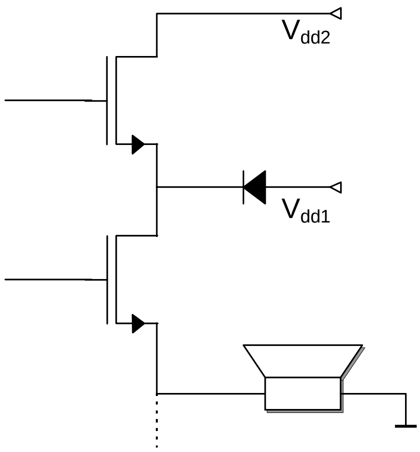

Figure 3.1: 'Serial' class G amplifier.

A disadvantage of this circuit is that there are always two elements in series. At low output voltages, the diode decreases the efficiency. During signal peaks the two transistors are in series, so that the output current has to pass two V_{DS} voltage drops. Figure 3.2 shows a 'parallel' topology that does not suffer from these problems. It needs special precautions in the driver circuitry, however, to prevent high V_{GS} reverse voltages across the upper left output transistor.

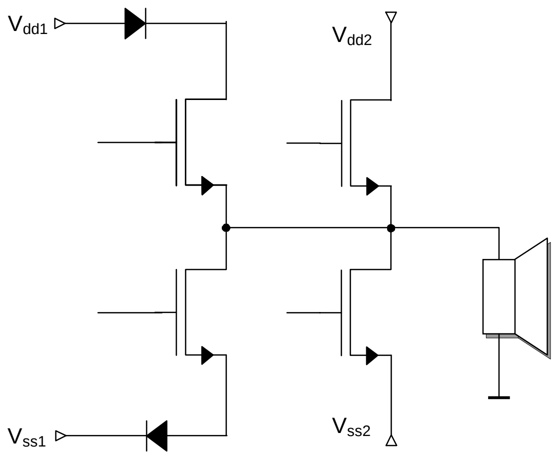

Figure 3.2: Parallel class G structure

In general, the need for multiple supplies may be a problem. If a transformer is used in the power supply, multiple taps are a good solution, but if a car battery is used, it is more problematic. Another problem with this type of amplifiers is the distortion caused by the switching between the two amplifiers. By using a comparator with hysteresis and delay to decide between the supplies [23], the number changes can be reduced, but this is a very inelegant way to reduce the total distortion. Another way to limit switching distortion is by switching between the two amplifiers gradually. However, this cuts down the efficiency a little.

3.2.2 Class H

It is not necessary to use two power supplies like in class G. Because the signal peaks generally last only a short time, the energy can be supplied by a capacitor. This technique is referred to as ‘Class H’ [22]. See Figure 3.3:

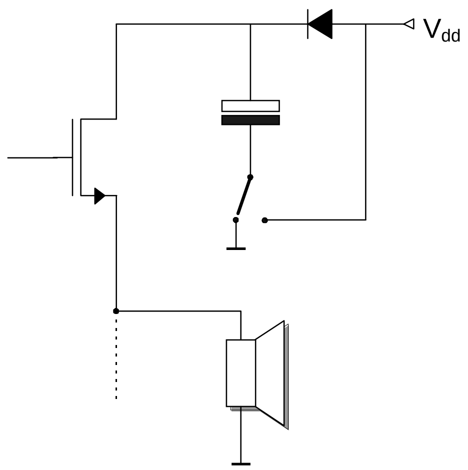

Figure 3.3: Class H amplifier

During low output voltages, the switch is in the position as drawn in Figure 3.3. During signal peaks the switch lifts the lower side of the elco to the power supply, such that the upper output transistor sees a voltage of approximately $2V_{dd}$. The time that the signal is ‘high’ should not be too long; a large elco is required for high power at low frequencies. Switching according to the envelope of the signal, as is sometimes done with class G amplifiers, is riskier as it is impossible to tell how long an envelope will last.

The advantage of class H is that only one power supply is needed. As such it is ideal for car audio applications. To prevent the need for four lifting elcos, it is then built like a bridge amplifier with a signal dependent common mode level. Figure 3.4 shows a class H bridge amplifier. The common mode level is normally half the supply voltage. When the load voltage must be higher than V_{dd}, the common mode level of the bridge is increased such that one half of the bridge remains at a constant voltage close to ground and the other half gets the lifted supply voltage. See Figure 3.5 for the waveforms of the two bridge halves for a sinusoidal output.

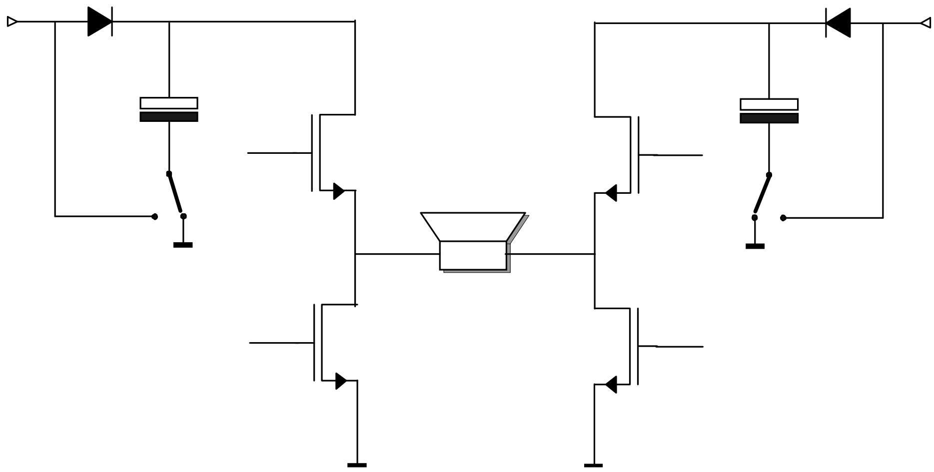

Figure 3.4: Class H as a bridge amplifier

For normal audio signals, and even for a rail-to-rail sinewave, only one lifting circuit would suffice. In practice this is not implemented, because it’s a bad habit to test audio amplifiers with near rail-to-rail squarewaves, which give the lift elco not enough time to recharge.

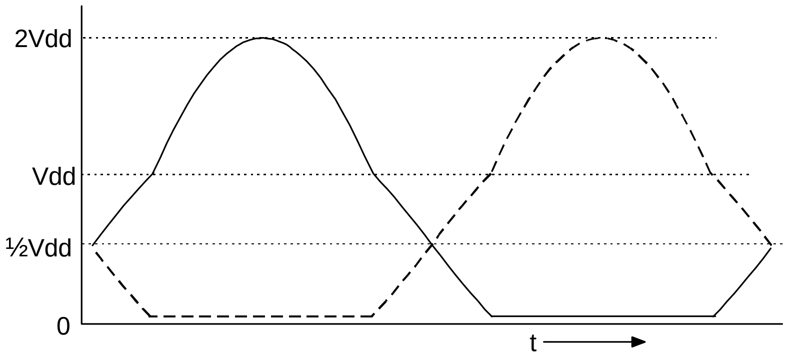

Figure 3.5: Waveforms of both bridge halves for a sinusoidal output

3.2.3 ‘Cool power’

The ‘cool power’ technique is very similar to the techniques in the last two paragraphs. The circuit normally operates single-ended: one side of the loudspeaker is connected to an amplifier, and the other end to an elco. The ‘quiescent output voltage’ of the amplifier is half the supply voltage, so it charges the elco to the same value. Because audio signals have no DC component, this voltage will hardly change. During signal peaks that are higher than half the supply voltage, the loudspeaker is disconnected from the elco, and connected to a second amplifier to work in a bridge configuration [25]. See Figure 3.6 and Figure 3.7:

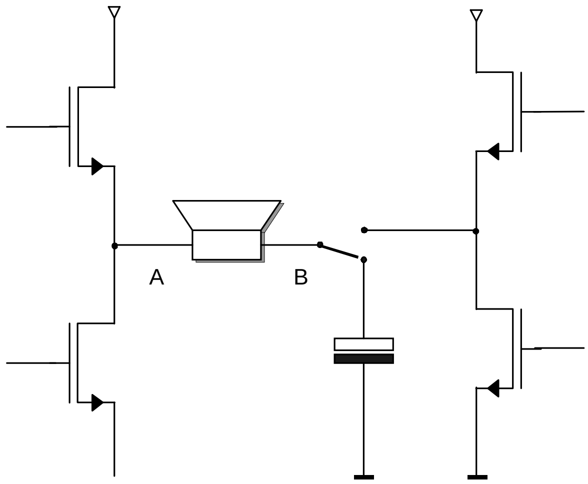

Figure 3.6: ‘Cool power’ amplifier.

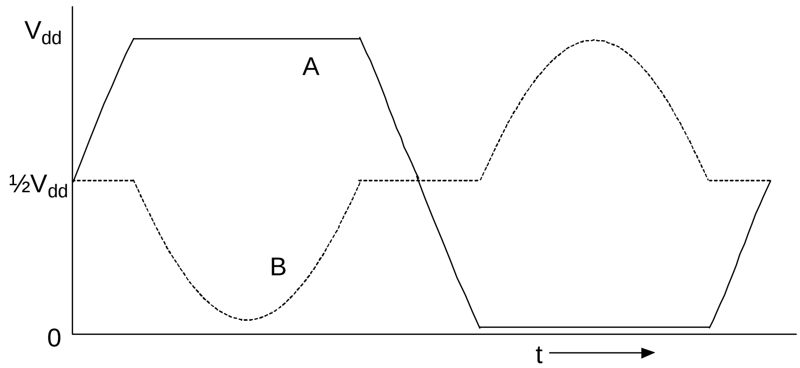

Figure 3.7: Waveforms of the bridge halves of the cool power amplifier for a sinusoidal output.

Like the class H amplifier, it provides good possibilities for automotive applications, as it needs only one power supply. Contrary to the class H amplifier, there is no limit to the duration of the high power signals. However, it is favourable to design the heat sink for average music/speech signals.

3.2.4 ‘Front/Rear’

Car audio amplifiers that have four channels (2 in front and 2 in the rear) can benefit from the correlation between the two front and the two rear signals. When these are the same, the approach in Figure 3.8 leads to the same efficiency as other class G topologies without the need for multiple supplies or external components [21].

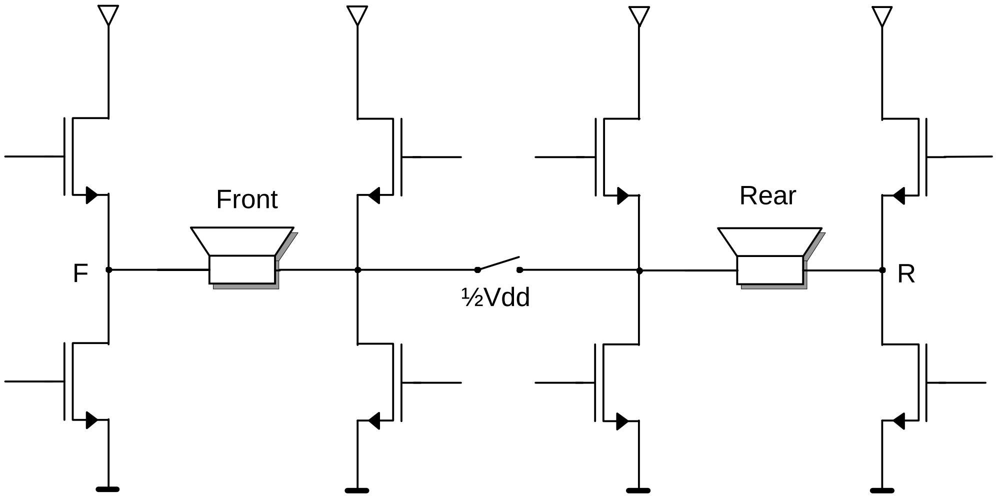

Figure 3.8: Stereo bridge amplifier

During low signals the switch is closed, and the two loudspeakers are in series. One of the amplifiers in the middle is put in high impedance while the other is set at $½V_{dd}$, supplying the difference current between the front and the rear channel. When the total supply voltage is not high enough to deliver the desired power to the two loudspeakers in series, the switch is opened, and the circuits act as two separate bridge amplifiers.

It is also possible to use the correlation between the left and the right channel of an audio signal. The weakness of this topology, however, lies also in this correlation. It is very rare that the front and rear channel or the left and right channel are the same. When the differences are very large, the amplifier acts as two separate bridge amplifiers without any improvement in efficiency.

3.2.5 Generalisation and modelling

The main dissipation in linear amplifiers is caused by the output current that has to flow from the supply voltage to the output voltage. The voltage drop times the output current is the dissipated power. This is true for all the amplifiers in the previous paragraphs, which can all be considered as amplifiers with multiple supplies. Note that it does not matter across which component the voltage drop exists. In a class H amplifier, for instance, the component controlling the output voltage might as well be the lifting circuit (switch) as the output transistor; the dissipation will be the same.

For amplifiers with multiple supplies, the actual decrease in dissipation depends on the amplitude distribution of the audio signal. All amplifiers discussed above have a dissipation that is approximately equal to a class G amplifier with two supplies with $V_{DD2}=2V_{DD1}$. In a class G amplifier, the number of supplies can be larger to achieve an even better efficiency at the cost of complexity. Also, the voltages of these supplies can be chosen freely, which influences the efficiency. Given a certain amplitude distribution, there is a optimum [24]. This optimum strongly depends on the load [20]. Because most loudspeakers are more or less reactive, the current is out of phase with the voltage. This could lead to large currents while the full supply voltage is across the output transistors. This influences the power dissipation and consequently the choice of the supply voltages.

In practice, the performance is limited by saturation voltages and quiescent currents. As an example, we will make a dissipation model for a class AB and a class H amplifier. The purpose of this model is to use it in a computer program that accepts audio files and can calculate the dissipated energy.

The class AB amplifier

Suppose a single amplifier with a symmetrical power supply. For a bridge amplifier, the situation is the same. A bridge amplifier with V_S supply voltage can be considered as a single amplifier with a symmetrical supply +/- V_S. The dissipation calculation is the same for positive or negative signals. Therefore, first the absolute value of the output voltage is taken and the loudspeaker current is calculated:

$$V_{o,abs} = |V_o| \qquad\qquad I_{LS} = \frac{V_{o,abs}}{R_{LS}}$$

The dissipated energy for that sample is a result of the voltage drop across the output transistors:

$$E_{drop} = \frac{1}{f_s} \cdot I_{LS} \cdot \left(V_S - V_{o,abs}\right)$$

with f_s the sampling frequency and V_S half the total supply voltage. The V_o range does not fully extend to V_S, but only to the clipping point of the real amplifier, V_{clip}. With P_Q the quiescent power dissipation, the average power dissipation during one audio fragment of length T_{frag} is:

$$P_{diss\_AB} = P_Q + \frac{1}{T_{frag}} \sum E_{drop}$$

The class H amplifier

The class H amplifier also works symmetrically:

$$V_{o,abs} = |V_o| \qquad\qquad I_{LS} = \frac{V_{o,abs}}{R_{LS}}$$

Although the class H amplifier is a special bridge amplifier, it can be modelled as a class G amplifier with two power supplies, $V_{S,high}$ and $V_{S,low}$. For small signals $V_{S,low}$ is used. A signal is small if V_o is smaller than a certain threshold voltage V_{switch}, which will lie somewhat below $V_{S,low}$. The dissipated energy per sample due to the voltage drop across the output transistors is:

IF (V_o<V_{switch}) THEN $E_{drop} = \frac{1}{f_s} \cdot I_{LS} \cdot \left(V_{S,low} - V_{o,abs}\right)$

ELSE $E_{drop} = \frac{1}{f_s} \cdot I_{LS} \cdot \left(V_{S,high} - V_{o,abs}\right)$

The class H amplifier also dissipates due to the fact that it has to charge and lift the electrolytic capacitor:

IF (V_o>V_{thr}) now, AND (V_o<V_{thr}) in the previous sample THEN $E_{lift} = E_{lift}$

ELSE $E_{lift} = 0$

Which results in a power dissipation of:

$$P_{diss\_H} = P_Q + \frac{1}{T_{frag}} \sum \left(E_{drop} + E_{lift}\right)$$

The used parameters are summarised in Table 2.

Class AB	V_S	half the total supply voltage
	V_{clip}	clipping point of the amplifier
	R_{LS}	loudspeaker resistance
	P_Q	quiescent power dissipation
Class H	$V_{S, low}$	supply voltage
	$V_{S,high}$	$2V_{S,low}$ minus saturation voltages
	V_{switch}	output voltage at which lifting starts
	V_{clip}	clipping point of the amplifier
	R_{LS}	loudspeaker resistance
	E_{lift}	energy per $1/f_s$ to lift the supply
	P_Q	quiescent power dissipation

Table 2: Summary of the parameters used in the class AB and class H dissipation models.

To verify the model, the IEC-268 test signal was applied to two real amplifiers with the same maximum output power, and the model parameters were fitted to the measurements. The comparison between the simulation and the measurements are shown in Figure 3.9. It is remarkable that even these simple models give quite a good description of the amplifiers. Only the dissipation of the class AB amplifier at higher output powers is in reality higher than in the measurements. Several mechanism can cause this, but without more knowledge about the interior of the used IC, there seems little point in trying to improve the model.

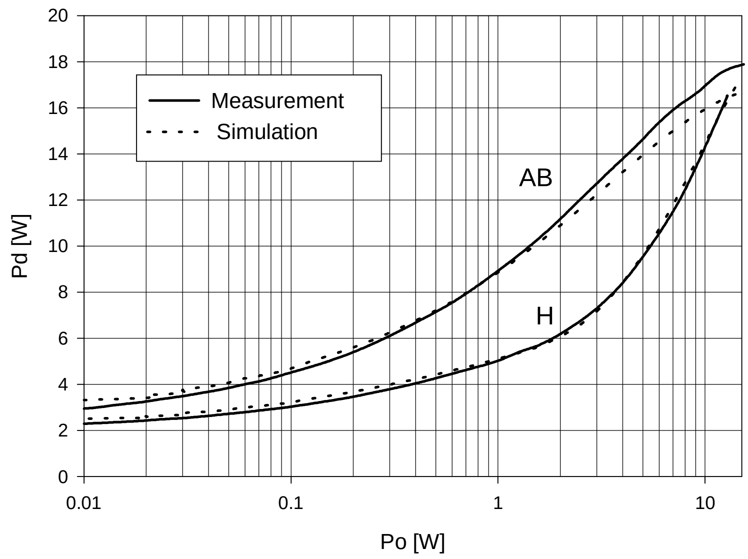

Figure 3.9: Measurement and simulation of the dissipation of a class AB and a class H amplifier (both $P_{o\text{-}max}$=30W) with the IEC-268 test signal as input.

3.2.6 Limitations of linear amplifiers

The main limitation of linear amplifiers is that increasing the number of supply voltages is complex, while the benefit is limited owing to the quiescent power dissipation. Figure 3.10 shows the dissipation of ideal amplifiers with 1,2, or 3 power supplies.

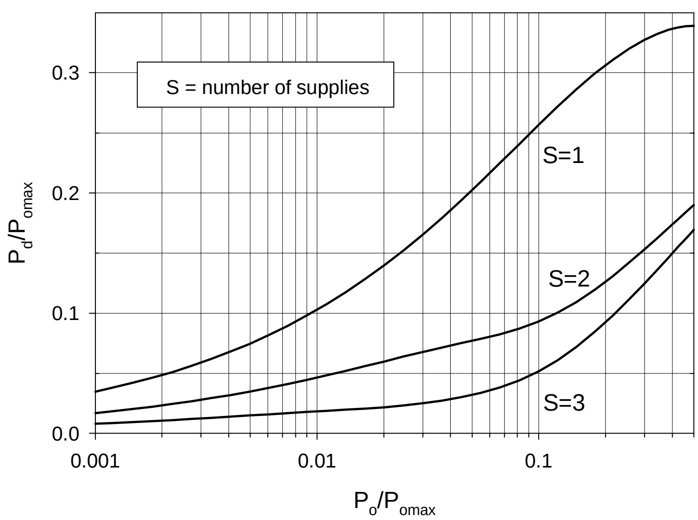

Figure 3.10: Dissipation of ideal amplifiers with 1,2, and 3 supply voltages. The second supply voltage equals half the first one, etc. S=1 represents a class B amplifier.

Saturation voltages and quiescent power dissipation are zero. In the ideal case, a class G amplifier with 2 supply voltages halves the dissipation. Adding an extra 3rd supply voltage halves the dissipation again. At present, it is not attractive to use a higher number of voltages. Every extra supply introduces at least 1 extra elco and switching becomes very complicated. When we consider non-idealities, even a third supply seems unattractive. A quiescent power dissipation of 0.02…0.05P_{omax} is quite usual for audio amplifiers. This means that at 0.1P_{omax}, approximately the maximum output power for undistorted playback of audio signals, 20…50% of the dissipation is caused by quiescent dissipation. For compressed music (that can play at a higher average output power without distortion) these figures are even less favourable. Furthermore, the maximum dissipation is barely lower than for two supplies.

3.3 Switching amplifiers

With switching amplifiers we mean amplifiers with a switching output stage. This means that the transistors in the output stage have a switch function; any simultaneous occurrence of voltage across and current through these transistors is undesirable.

3.3.1 The class D principle

A typical class D amplifier consists of a modulator that converts an analogue or digital audio signal into a high frequency Pulse Width Modulated (PWM) or Pulse Density Modulated (PDM) signal followed by the output stage, often a half bridge power switch (Figure 3.11). The output of the switches is either high or low, and changes at a frequency that is much higher than the highest audio frequency. Typical values are between 200kHz and 500kHz. The frequency spectrum of the PWM signal in the audio band is the same as the frequency spectrum of the audio signal. An LC filter filters out the high frequency switching components, so that the audio signal is available at the output of the filter. Ideally, the switches do not dissipate and neither does the filter, so the efficiency can be very high.

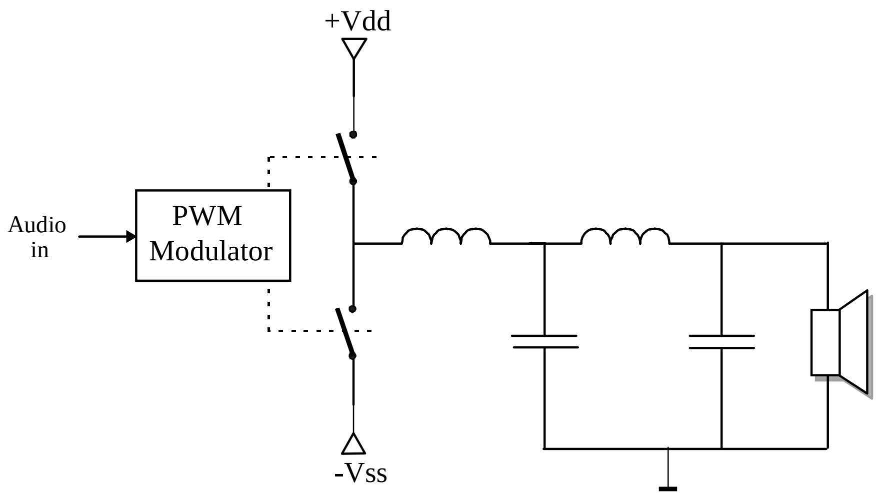

Figure 3.11: Principle of PWM amplifier

For a 10kHz sinewave, a switching frequency of 350kHz, and a filter with a 30kHz Butterworth characteristic, the signals look like Figure 3.12. In this case, the audio frequency is close to the corner frequency of the filter, so some phase shift can be observed between the PWM signal and the audio signal.

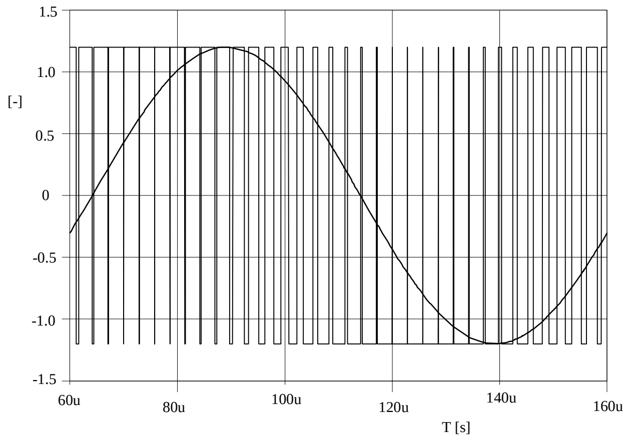

Figure 3.12: Class D output signal before and after the filter.

3.3.2 Output stage

Figure 3.13 shows a typical class D output stage. It is a class AD stage, which is used for most class D amplifiers. It is a simple inverter: when the input signal is positive, M_2 conducts. When it is negative, M_1 conducts.

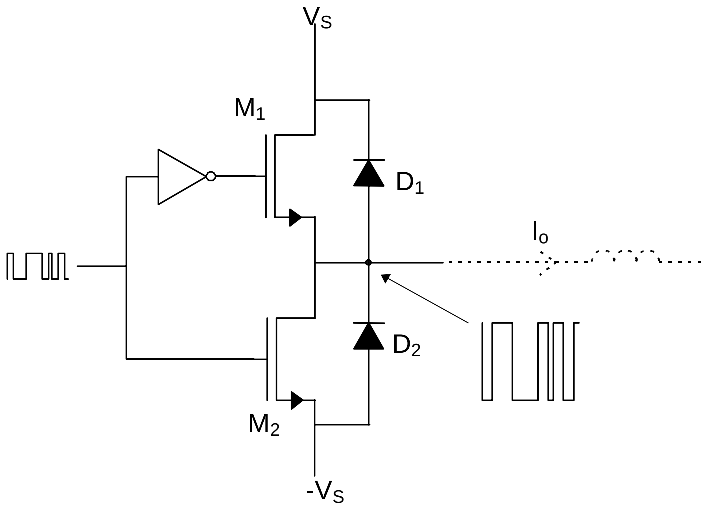

Figure 3.13: A typical class D output stage.

Diodes

The diodes D_1 and D_2 are needed because the transistors are unidirectional switches. Suppose the output signal is positive, and the output current I_o is also positive. When M_1 is switched on, this is OK, but when M_2 is switched on, the coil in the output filter still tries to keep the current I_o, forcing the output voltage below $-V_S$, causing D_2 to conduct. With DMOS transistors as switches, the intrinsic diodes can be used. However, the intrinsic diode of a DMOS transistor can have a long recovery time (several hundred ns) or cause latch-up. In that case external (shottky) diodes are a solution, although not a desirable one. It is also possible to build DMOS transistors with a fast-recovery intrinsic diode [34].

Switching speed

High switching speeds are necessary to keep switching losses small. Typical values of today's integrated designs are tens of nanoseconds. Because of the large gate-source capacitances of M_1 and M_2, this leads to large peak currents. Also, the high speed switching in combination with wires and (gate) capacitances can cause ringing, overshoot, and delays. For a low distortion it is important that the switching times of M_1 and M_2 are equal [29]. Tuneable coils between M_1 and M_2 can provide a solution. However, both the fact that these coils can not be integrated and that they need to be tuned make this an unattractive solution. With high speed switching, the risk of common conduction of M_1 and M_2 increases. The introduction of a 'dead zone' in which both transistors are turned off is a common solution, although this introduces extra distortion in the audio signal. Another option is a handshake procedure to check if the other transistor is turned off.

Power supply

In pure feedforward systems a stable power supply is extremely important, because any deviation from the nominal value shows up in the output signal. For an output signal of 16 bit accuracy, the power supply should have a 16 bit stability. Common solutions are feedback from the pulsed output or feedforward correction by referring the triangle waveform to the supply voltage (see section 3.3.5). Another supply issue arises from the use of NMOS devices that are preferable thanks to the lower R_{on} per area. The gate of M_1 needs a voltage that is higher than V_S. A bootstrap capacitor or a chargepump can provide such a voltage [56].

Cross-over distortion

M_1 and M_2 have a certain R_{on} resistance. D_1 and D_2 have a certain voltage drop when conducting. Suppose the output current is positive. During conduction of M_1, the voltage will be a little lower than V_{DD} because of R_{on1}. During conduction of D_2, the voltage will be a little lower than $-V_S$ due to the voltage drop. So all the time the voltage is lower than it should be. When the output current is negative, the same reasoning shows that the output voltage is too high. This results in cross-over distortion. It can be solved by connecting the transistors to a tap of the output inductor [29] or a separate supply voltage.

Class BD output stage

An alternative to the class AD stage is the class BD stage. In class BD there are 3 possible output voltages: positive, negative, and zero. There are several ways in which this can be implemented, but the simplest one is shown in Figure 3.14. [28].

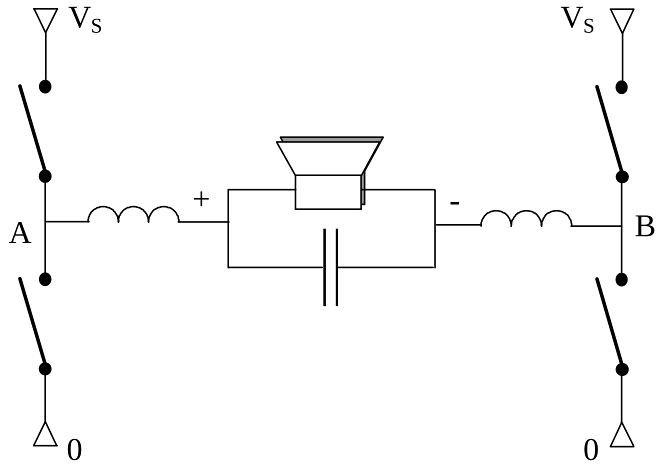

Figure 3.14: Class BD modulator with output filter.

In quiescent, the signal at A and B is the same PWM signal with 50% duty cycle. The signal A-B across the filter is therefore zero. For a positive output voltage, the duty cycle of A is increased and that of B decreased. The difference signal A-B is now a voltage that varies between 0 and V_S. Similarly, for negative output voltages, A-B varies between 0 and $-V_S$. The pulse frequency of A-B is doubled compared to A and B, which is favourable for speed requirements. Balanced current design has the same qualities [37] and the topologies are very similar.

The difference between a bridge class BD stage and a bridge class AD stage is subtle. The topologies are exactly the same. In the class AD case, however, A is always the inverse of B, so that A-B alternates between $-V_S$ and V_S.

Resonant output stage

A way to generate the high frequency pulses for a PDM modulator (section 3.3.5) is to use a quasi-resonant converter [27].

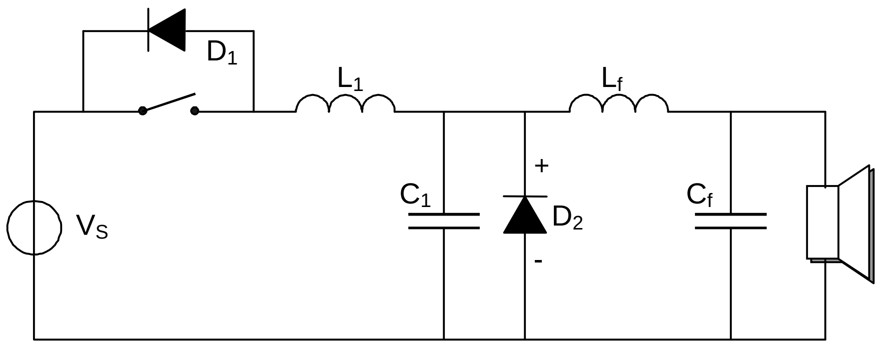

Figure 3.15: Quasi-resonating converter.

This converter gives 1 bit each time it is switched on. The bit is not a squarewave, but the positive half of a sinewave. This is irrelevant, as long as the area under the signal is the same each time. For the topology in Figure 3.15 (L_f and C_f are the output filter) this is true, virtually independent of output current and voltage. Switching occurs when the current is zero, giving better efficiency and lower switching noise. The large number of filter components make this topology not very attractive for use with integrated circuits.

Conclusion

The design of a class D output stage is not a trivial matter. In general, an output stage will not be able to preserve the exact frequency content of its input signal.

3.3.3 Output stage dissipation

Most of the dissipation in a class D amplifier is generated in the output stage. The dissipation consists of three main components: Conduction losses, switching losses, and capacitive losses. Conduction losses are the result of the on-resistance of the switches. Refer to Figure 3.13. The conduction losses can be expressed as:

$$P_{cond} = I_o^2 \cdot R_{on}$$

where I_o equals the output current and R_{on} the on-resistance of the switches. I_o has some high frequency ripple. At higher output powers it is relatively small, since it is undesirable to dimension the maximum current rating of the switches much higher than the maximum output current of the amplifier.

Switching losses are a result of the simultaneous presence of voltage across and current through the switches. Suppose the output current I_o is positive and M_1 is switched from conducting to non-conducting. The coil in the output filter tries to keep its current I_o at a constant value, bringing down the output voltage of the power stage below $-V_S$ so that D_2 starts conducting. During the time the voltage goes down, I_o has to be supplied by M_1. Figure 3.16 shows the corresponding waveforms.

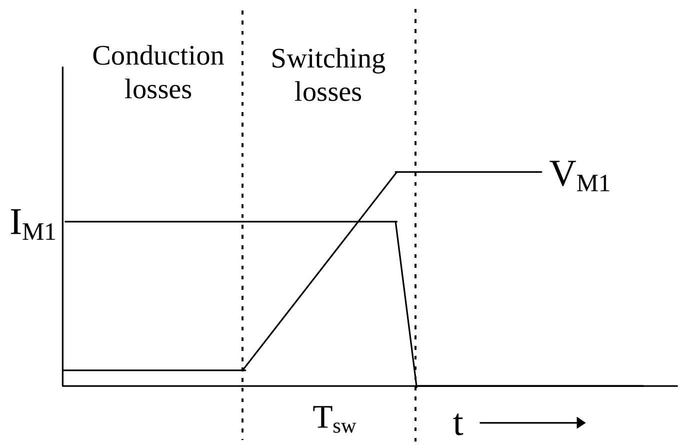

Figure 3.16: Voltage over M_1 and current through M_1 during switching from high to low output voltage.

If we assume the voltage across M_1 to rise linearly from 0 to $2V_S$, the dissipated energy during the switching time T_{sw} is approximately $½*2V_S*I_o*T_{sw} = I_oV_ST_{sw}$.

There are two transitions per period of switching, so the switching losses at a switching frequency f_{sw} can be expressed as:

$$P_{switch} = 2 V_S I_o T_{sw} f_{sw}$$

The third source of dissipation are (internal) capacitances that are charged and discharged. When a capacitance C with zero initial charge is charged to a voltage V, the final energy $\frac{1}{2}CV^2$ has also been dissipated during the charging process. This can be prevented with the aid of inductors like in the output filter, but the driving circuitry of the gate capacitances, for instance, does not work that way. This brings the capacitive losses to:

$$P_{cap} = C_{\text{int}} V_c^{\ 2} f_{sw}$$

where f_{sw} is the switching frequency, C_{int} are the internal capacitances and V_c is the voltage to which these capacitances are charged. In a practical system there is more than one capacitance present, and they may work at different voltages.

In a class D output stage, all three dissipation sources are significant. The model parameters were matched to the class D stages that were used in our experiments. This yielded 16ns switching time, a total on-resistance of 0.65Ω, and internal capacitances of 500pF. With these values the dissipation at higher output powers could be predicted with less than 10% error. Possible errors include negligence of the drain-source capacitances, which are charged and discharged at the switching frequency. If this is done by the coil current, the process is lossless. If it is done by the output transistors, it is not. The actual situation depends on the output power, the switching time and the ripple current.

The dissipation model is useful for sizing the output transistors. If very large transistors are utilised in an output stage, the conduction losses will be small, resulting in a lower maximum dissipation. The capacitive losses, however, will be considerable, resulting in a higher quiescent dissipation. This is a common trade-off in class D amplifiers, which could be circumvented by using small transistors at low output currents and larger transistors at high output currents. Reduction of the switching time T_{sw} is always favourable, as it decreases the switching losses.

3.3.4 Output filter

The output filter is a low pass filter that reduces the switching frequency. When designing the filter, the load impedance is part of the equation. Thus, the load can seriously affect the frequency transfer. For different loudspeakers, the impedance over the audio range can vary from 1Ω to as much as 30Ω, with a phase of +56...-67° [20]. Figure 3.17 shows the impedance of a 3-way loudspeaker system that was used for our listening tests.

The output filter, however, is designed for a real and constant load impedance. The result of connecting the loudspeaker is shown in Figure 3.18. The flat line is the simulated transfer of an ideal class D filter followed by a fourth order Butterworth filter with a corner frequency of 30kHz, loaded with the specified load impedance of 4Ω. The other line shows what happens when the loudspeaker is connected. The transfer deviates several dB's from the flat line. This will colour the sound impression. Another problem is that any non-linearities in the filter show up in the distortion figures.

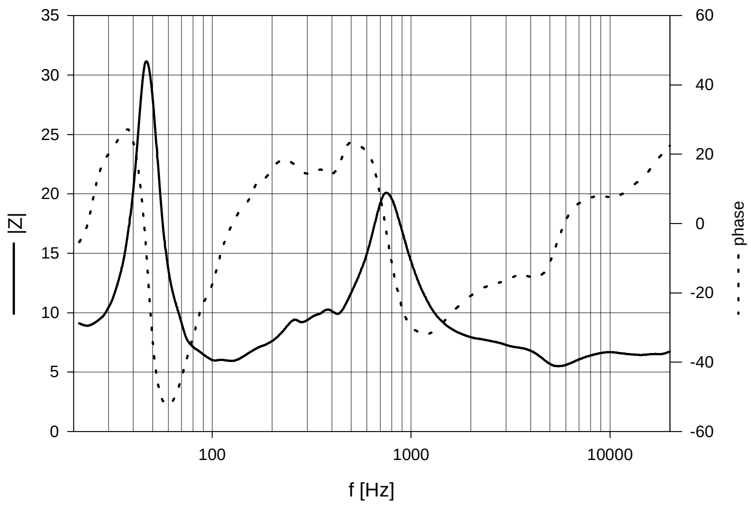

Figure 3.17: Loudspeaker impedance

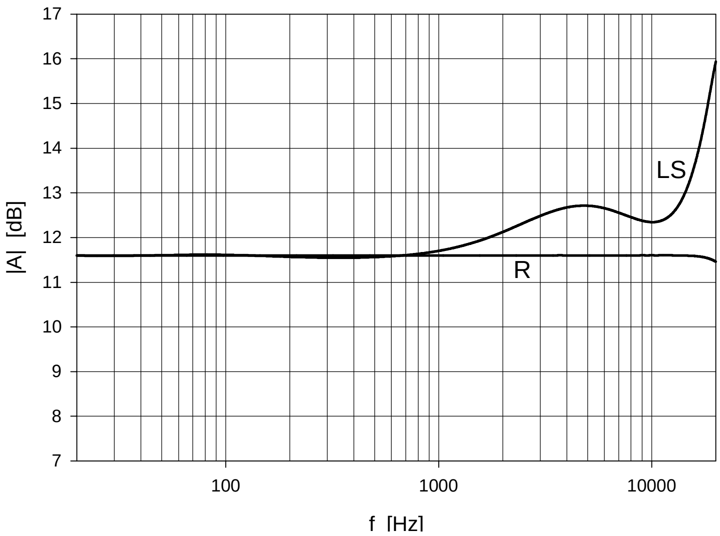

Figure 3.18: Simulated class D frequency transfer with resistor and loudspeaker load.

Feedback can reduce these problems considerably, but because of the phase shift, feedback after (part of) the filter is complicated [28, 29, 34]. In general, the filter (or the filter in combination with lead compensation) must have a first order frequency transfer at 0dB to ensure stable operation. This is extra complicated by the connected load, which is a part of the filter. High feedback factors can not be realised and feedback around a filter with more than 2nd order behaviour is very rare. Even when these problems are overcome, the filter prevents further integration because it contains elements that can not be integrated on chip. For sufficient suppression of the carrier frequency, typically a fourth order filter is necessary. In this case, the amount of filtering in practical situations is limited by parasitic capacitances and resistances [32,34]. Furthermore, two coils and two capacitors are already considered to be many external components. Using only a second order filter is a solution, but the amount of switching ripple can cause EMI problems and the application area of the amplifier will be limited.

3.3.5 Modulators and feedback

There are numerous modulators, and it is not our objective to give an extensive overview. In this section, only the basic topologies are discussed.

PDM modulators

PDM modulators have resulted from the digital signal processing domain. In more and more equipment, the signal is available in digital form. For a switching amplifier it must be converted into a 1 bit signal at a high frequency. Sometimes, as with DSD audio data, this is even the native format. The output stage acts as a 1 bit D/A converter. Because the length of each bit is constant, and only the presence or non-presence of a bit is controlled, this is called Pulse Density Modulation (PDM). To convert a multi-bit signal to a 1-bit signal, oversampled noiseshaping is used. Figure 3.19 shows a general noise shaper [62].

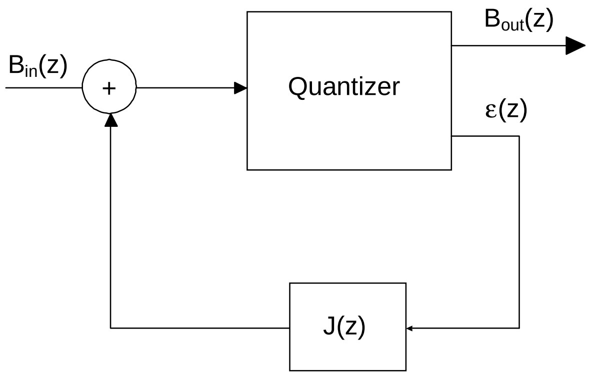

Figure 3.19: Noise shaper

The input signal $B_{in}(z)$ has a larger number of bits than $B_{out}(z)$. (When the input signal is analogue, a similar structure in the analogue domain constitutes a sigma-delta modulator [30,62]). The block called ‘Quantizer’ reduces the number of bits by simply passing only the most significant bits to $B_{out}(z)$. The least significant bits, which are the error, are added to the input after passing through a transfer function J(z). It is easy to calculate B_{out}:

$$B_{out}(z) = B_{in}(z) - \varepsilon(z)(1-J(z))$$

Suppose $J(z) = z^{-1}$, one clock delay. The system is now a first-order noise shaper. $B_{in}(z)$ is a 16 bit signal at $256f_s$ and B_{out} is a 1 bit signal at $256f_s$. In that case, the quantizer transfers 1 bit to the output. The other 15 bits are the error signal. B_{out} equals:

$B_{out}(z) = B_{in}(z) - \varepsilon(z)(1\text{-}z^{-1})$

With $z = e^{2\pi j\left(\frac{f}{256 f_s}\right)}$, we see that for low frequencies (audio) the error in the output signal approaches zero. The error reaches a maximum for f=128f_s. See Figure 3.20.

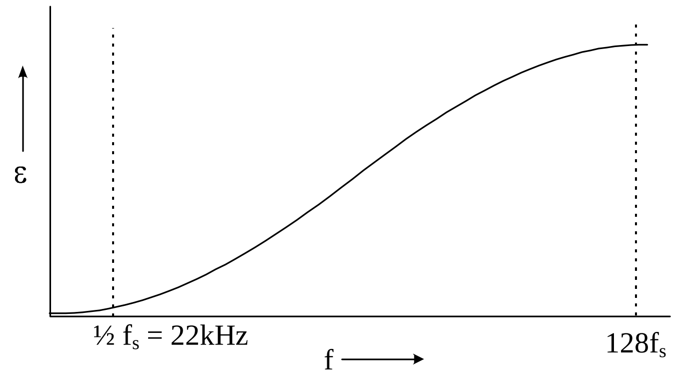

Figure 3.20: Noise distribution as a function of frequency

Applying B_{out} to a 1 bit D/A converter and filtering above 20kHz reconstructs the original signal. In the time domain such a noise shaper is a way to convert resolution in the amplitude domain to resolution in the time domain. It outputs bits at high speed in such a way that the average is the intended output (which has a higher amplitude resolution). This way it is also easy to see that although the D/A converter is only 1 bit, it should have a 16 bit accuracy.

To convert the audio signal to 256f_s, an oversampling interpolating filter must proceed the noise shaper. A two times oversampling filter works as follows [60]. Suppose the spectrum of the signal sampled at f_s looks like Figure 3.21. This signal is converted to a sampling frequency of 2f_s by inserting a sample of value zero after every original sample. See Figure 3.22. Because every sample is a Dirac-pulse of proportional height, the frequency spectrum stays exactly the same.

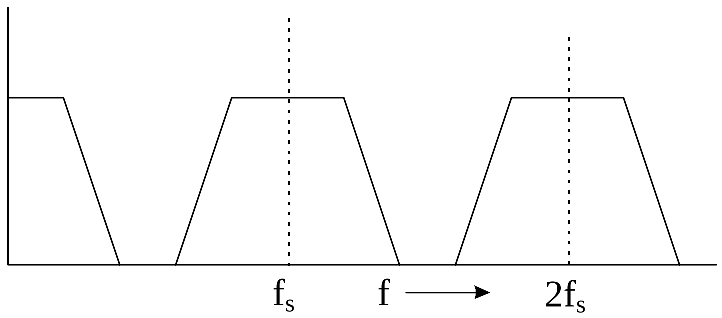

Figure 3.21: Spectrum of the signal

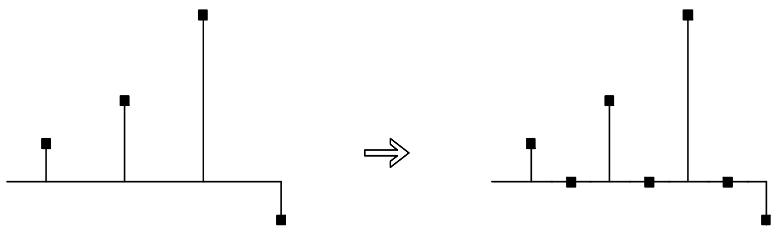

Figure 3.22:   Inserting zero samples

Next, the signal is applied to a digital filter at $2f_s$ that filters out the middle replica, see Figure 3.23. After that, the frequency spectrum of the signal looks exactly like it has been sampled at $2f_s$. These techniques, oversampling interpolating filtering and noise shaping are essential for all digital PDM systems, although the exact realisation may vary.

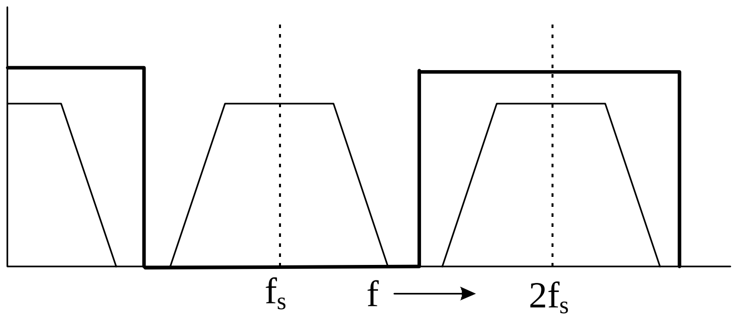

Figure 3.23:   Filtering out the middle replica.

In [62], the 256 times oversampling for a CD player D/A converter is done in two stages. A four times oversampling filter is followed by a 64 times linear interpolator. The direct use of a 256 times oversampling filter is also possible, but the filter would be very large. A linearly interpolating filter is easier to build, and at $4f_s$ the distortion that it creates has only little effect in the audio band. Then, at $256f_s$, a second order noise shaper suffices to get a 1 bit signal with 16 bit resolution in the audio band. Unfortunately $256f_s$=11MHz which is too high for power switching.

Another possibility is to use only $32f_s$ with an eighth order noise shaper [27]. Noise shapers with a higher order than three are prone to instability, and it is necessary to manipulate the system when it becomes potentially unstable [63]. Extensive simulations are necessary for evaluation. Even in this case, the switching frequency is 1.4MHz. The high switching frequencies are a general problem of PDM modulators. Bit-flipping techniques can reduce the average frequency at which the output changes somewhat [36].

Digital PWM modulators

Digital PWM modulators offer a lower switching frequency than PDM modulators. The Pulse Amplitude Modulated (PAM) samples are converted to PWM.

This could be done by giving each pulse a length that is proportional to the original amplitude. However, for CD quality the internal clock frequency would have to be 44.1kHz*2^{16}= 2.9Ghz, which is way too high. Furthermore, the frequency spectrum of the PWM signal would not equal that of the PAM signal. This can be calculated, but for a better understanding it is best to realise that natural sampling yields the best results because it does not introduce harmonic distortion. In natural sampling, the audio signal is compared to a triangle or sawtooth waveform (see the analogue modulator section below). When we convert a digital PAM signal directly to PWM, it looks as if, looking in the analogue domain, we compared the sawtooth waveform to a step-like representation of the signal instead of the signal itself. This is called uniform sampling. See Figure 3.24. It introduces harmonic distortion, which depends on many factors including the signal frequency, the switching frequency and the modulation depth [1].

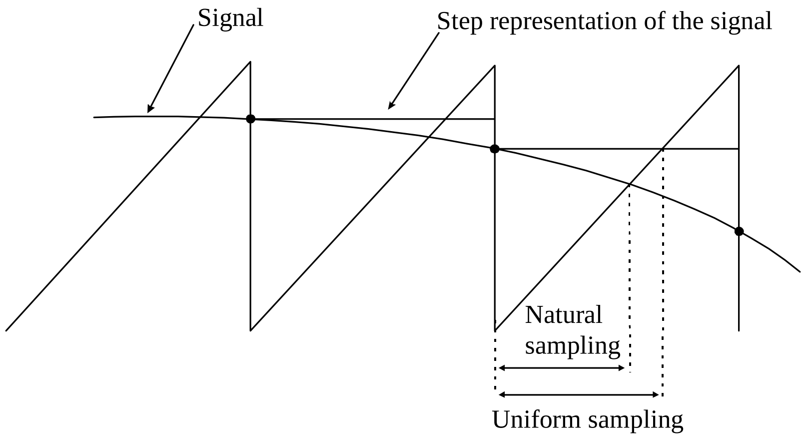

Figure 3.24: Natural sampling versus uniform sampling

To approximate natural sampling, linear or higher order interpolation between two or more samples is used to approach the natural PWM pulse width [33, 39, 41]. When the pulse width has been calculated, the sample instant can be the beginning or the end of the pulse (single sided modulation) or the middle (double sided modulation). There are more aspects that deserve attention, but a full discussion of these would be beyond the scope of this chapter.

Analogue PWM modulators

In the analogue domain a PWM signal can be generated by comparing the audio signal to a triangle or sawtooth waveform. This technique, called natural sampling, is the basis of almost all analogue modulators. See Figure 3.25. When the momentary value of the input signal is larger than the triangle, the output of the switch is high. It is easy to see that in this way the pulse width at the output is proportional to the input voltage. The modulator does not introduce harmonic distortion, only (multiples of) the carrier frequency and (multiples of) harmonics of the modulating frequency around the carrier [1].

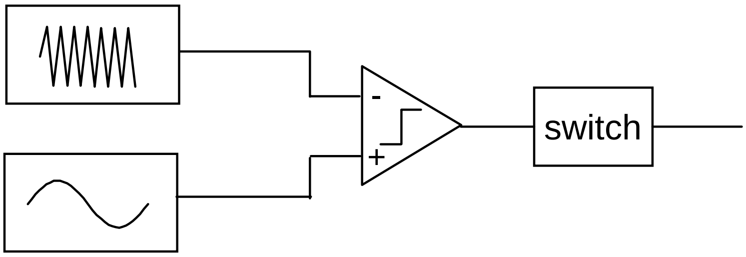

Figure 3.25: Open-loop class D modulator

The main problem is the lack of feedback. Output stage inaccuracies, non-linearities, timing errors and supply voltage variations all contribute to the distortion. We will discuss feedback here, as it is so closely related to the modulator.

Figure 3.26 shows a modulator with feedback. Both inputs to the comparator have triangular waveforms. Figure 3.27 shows the waveforms for zero and positive output voltage. At zero output voltage, the feedback signal intercepts the reference triangle in such a way that the duty cycle is 50%. When the output voltage is not zero, the rising and falling slope of the feedback triangle are different, leading to a larger (or smaller) dutycycle.

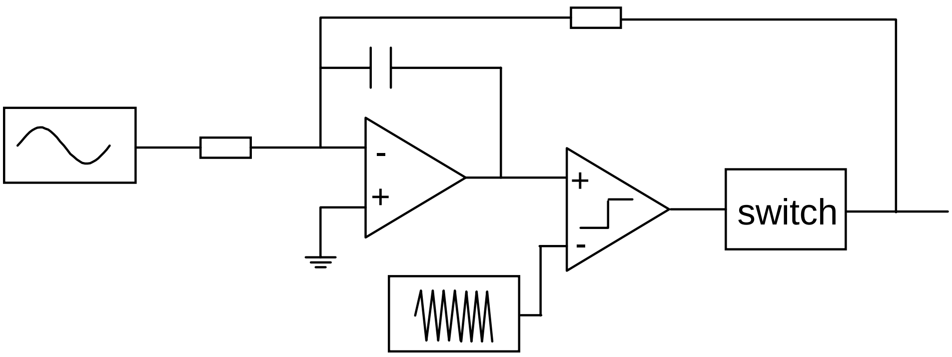

Figure 3.26: Modulator with feedback.

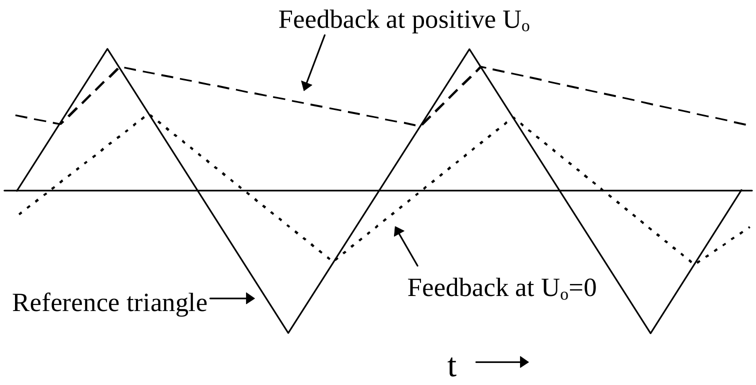

Figure 3.27: Signals at the input of the comparator of the feedback modulator.

The slew rate of the feedback signal must always be smaller than the slew rate of the reference triangle. Otherwise, the amplifier starts oscillating at a very high frequency. This constitutes a compromise between switching frequency and loopgain. The slew rate requirement can roughly be translated to the demand that the loopgain of the amplifier at the switching frequency is smaller than ½. Thanks to the integrator, the open loop frequency transfer of the amplifier is first order, so that the loopgain at a certain frequency has a maximum that is related to the switching frequency.

A way to get more loopgain at low (audio) frequencies is by introducing a range with second order frequency response in the loop. As long as the loopgain is back to first order at 0dB, stability is ensured. This can be done in the modulator by adding a second integrator before the comparator while bypassing it for high frequencies [40, 38]. In practical realisations of a feedback modulator, the triangle is generated by adding a squarewave to the input of the integrator [42]. The feedback properties of this type of modulator can also be used when the input signal is generated by a digital modulator [40]. Because in that case the bitstream is already clocked, the negative input of the comparator can be tied to ground. Other techniques, like the one cycle control technique [35] or pulse edge delay error correction [39] are similar to this modulator in their attempt to control the integral of the switched output voltage.

The high frequency oscillation that occurs in a feedback modulator when the feedback signal is too large, is exploited in the self-oscillating class D modulator [42]. See Figure 3.28. The comparator is equipped with some hysteresis to control the switching frequency. Other factors that influence the switching frequency are the integrator time constant and the output voltage. For large output voltages, the frequency approaches zero. This can cause aliasing problems that can be overcome by using a comparator with a variable hysteresis dependent on the input voltage. In that way the oscillator frequency is kept constant over a wide range of output voltages [31].

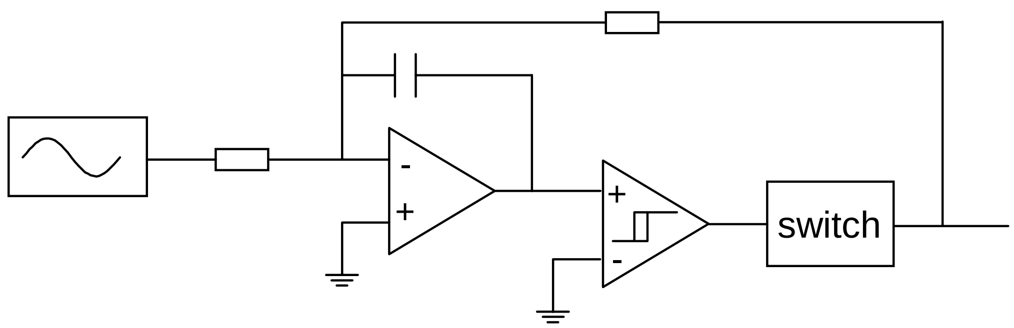

Figure 3.28: Self oscillating class D modulator

In the situations above, feedback is successfully taken before the output filter. The combination with feedback after the filter is more troublesome: see section 3.3.4.

3.3.6 Limitations of switching amplifiers

In the previous sections, the main building blocks of switching amplifiers have been discussed. To summarise the limitations that were encountered, it is easy to start with an important audio amplifier specification: low distortion. With feedback directly from the switched output, very good high power PWM signals can be generated. The output filter, however, introduces additional distortion and deviations of the specified frequency transfer when non-resistive loads are connected. Feedback after the filter is difficult, and high feedback factors can not be realised. Even when these problems are overcome, the filter prevents further integration because for sufficient suppression of the carrier frequency, typically a fourth order filter is necessary. It is not possible to eliminate the external two coils and two capacitors without introducing a much larger switching residue.

3.4 Conclusions

The objective of this chapter was twofold: give an overview of existing techniques to build high efficiency amplifiers, and at the same time analyse the main problems of these topologies.

Linear amplifiers have a low complexity, can have a low distortion, but show limited possibilities for reduction of the dissipation. To reduce the dissipation to very low values, complex switching schemes are necessary, and a large number of external elcos makes such a solution little attractive. Switching amplifiers can have a very low dissipation, but suffer from switching noise at the output, an external filter, a load dependent frequency transfer and difficulties in achieving a low distortion.

The idea that a mix of these two systems may be beneficial is not new, and the next chapter will look further into the possibilities of such combinations.

4

Combinations of linear and switching amplifiers

4.1 Introduction

The main problems with class D amplifiers are the switching ripple at the output and the difficulty to achieve a low distortion and flat frequency transfer. Class AB amplifiers do not suffer from these problems, but have a high dissipation. There are two basic ways of combining the two: by connecting them in series (section 4.2) or in parallel (section 4.3). This chapter is concerned with both existing and new topologies, and serves as a bridge to Chapter 5 and 6, that discuss the realisation of two topologies.

4.2 Amplifiers in series

4.2.1 Amplifiers with a tracking power supply

The most popular combination of switching and linear amplifiers is a topology in which a class D amplifier (D) generates the supply voltage for a class AB amplifier (AB). This is a 'series' way of connecting the two, as the output current is delivered by both amplifiers. Figure 4.1 shows one half of the basic schematic of such a combination [45, 47, 49, 51]. When the supply voltage of AB tracks the audio signal plus some headroom ΔV, the voltage drop across the output transistors is small. When the output voltage and current are in phase, the dissipation of AB is also small. The voltage headroom enables AB to correct distortion generated by D. Also, the switching ripple of D is reduced by the PSRR (Power Supply Rejection Ratio) of AB.

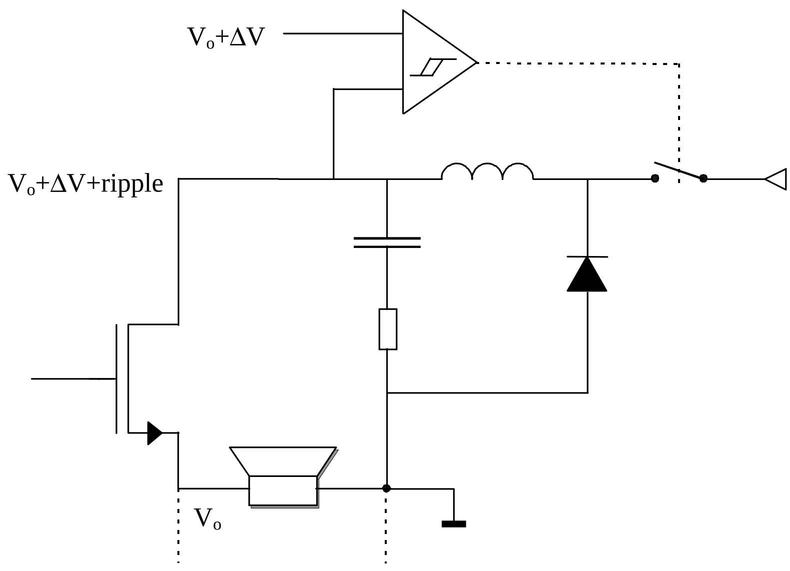

Figure 4.1: Basic schematic of an amplifier with a variable supply voltage.

The type of class D amplifier used in Figure 4.1 needs some explanation. When a standard class D output filter were used, it would introduce delay. Figure 4.2 shows the simulated phase shift caused by an ideal 4^{th} order 30kHz butterworth filter loaded with a loudspeaker (trace labelled 'LS'). The maximum of about 100° is far too much in this application. A filter with the same time response in series with the input of AB would be necessary to decrease the phase difference between AB's supply and output voltage. Figure 4.2 shows that the variation in phase caused by variation in load impedance is sufficiently small.

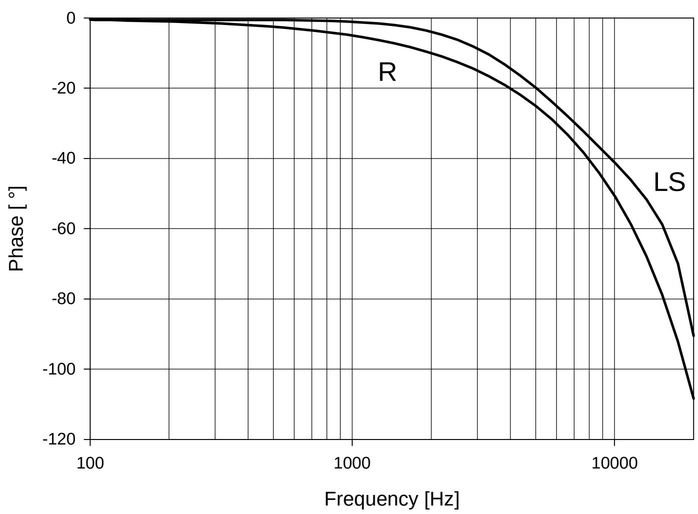

Figure 4.2: Phase of an ideal class D output filter loaded with the specified resistance and with a loud-speaker.

Another problem, however, is the fact that AB acts as a current source for frequencies higher than the audio band. This way, the filter is not loaded with its specified impedance at the switching frequency. Instead, it sees a high impedance. In a standard LC-filter, this gives rise to peaking, so that the output voltage can not easily be controlled.

A self-oscillating class D amplifier as displayed in Figure 4.1 solves these problems. In this topology, the integrator is the output filter itself. The comparator (with hysteresis) compares the desired output voltage with the actual output voltage and adjusts the switch accordingly. The output voltage ripple is determined by the hysteresis. For stable operation, the filter must behave as a first order integrator at the switching frequency. This is the function of the resistor, which is chosen somewhat larger than the impedance of the capacitor at the switching frequency. Figure 4.3 shows the loop gain of the self oscillating loop. The switching frequency is determined by L, R, the output voltage, and the hysteresis of the comparator. An extra function of the resistor in this application is to damp the LC filter.

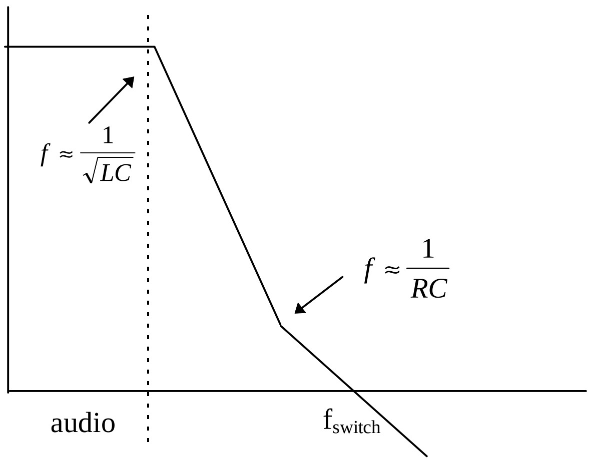

Figure 4.3: Loop gain of the self-oscillating class D amplifier.

For a low switching frequency and a low ripple, the corner frequency of the filter needs to be low. On the other hand, it should be above the audio band. When this is a problem, and when the audio signal is available in digital form, it can be stored in a shift register, giving the amplifier the ability to look ahead. A slow class D amplifier can use that time to adjust to the desired voltage [44]. A disadvantage of this approach is that for high frequency signals the efficiency decreases. On the system level, it can be advantageous to chop the supply voltage at the primary side of the power transformer, so it can be smaller than in the normal 50/60 Hz case [48].

A disadvantage of the topology in Figure 4.1 is that the full amplifier needs two class D amplifiers: one for the upper- and one for the lower output transistor. In a bridge topology, a single class D amplifier suffices. In Figure 4.4 [51], it always follows the higher of the loudspeaker terminals. The common mode level of the bridge is chosen very small and increases during signals such that the lower of the loudspeaker terminals is close to ground. The corresponding waveforms are discussed in Chapter 3, as the TDA1560 class H amplifier uses a similar kind of common mode level control.

The dissipation of series AB/D combinations is comparable to that of class G amplifiers, which is hardly worth the extra complexity. Although the theoretical dissipation can be lower, it is larger due to practical problems. It is difficult to design a class AB amplifier that can drive very close to the supply rails. The stability, for instance, can be problematic when the gain of the output transistor decreases at low drain-source voltages. The necessary voltage headroom is further increased by the ripple of the class D amplifier. Consequently, the output current has to cross a considerable voltage drop, leading to a dissipation that is relatively high.

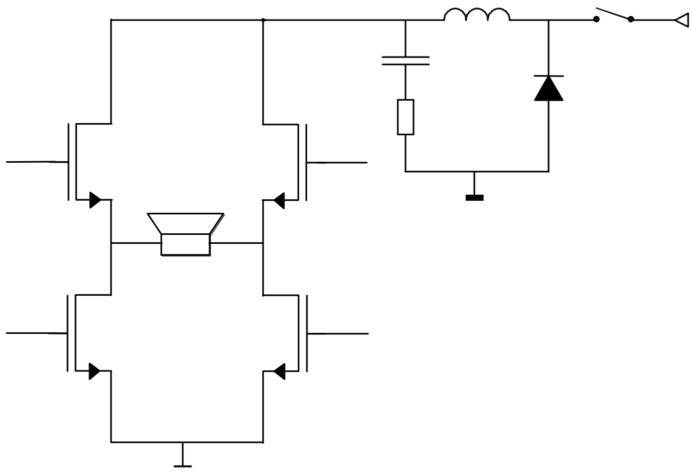

Figure 4.4: Bridge amplifier with only one switching regulator.

4.2.2 Chip area considerations

For comparison of different amplifier topologies, the chip area can be a useful criterion. IC costs are determined by several factors like design, wafer production, testing, packaging and distribution, of which the wafer production costs are the highest. Therefore reduction of the chip area results in a cheaper IC. For integrated audio power amplifiers this picture can be simplified even further, since the output transistors usually take up a big part of the chip area. Consequently it is desirable to use output transistors that are as small as possible, given a certain power rating and topology. Suppose a class AB output stage as shown in the left of Figure 4.5. The only effect that the MOSFETs have on the maximum output power is their R_{on}, which determines how close they can drive the output to the supply rail. Now suppose the supply voltage of this class AB stage is doubled, so that the maximum output power quadruples.

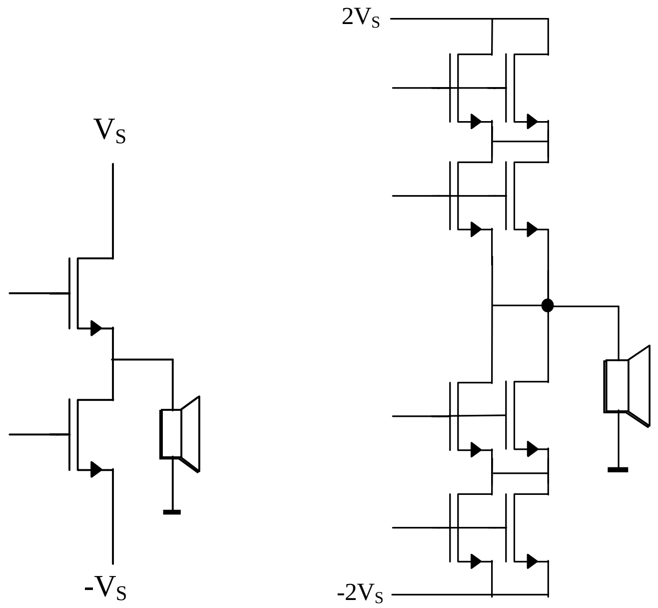

Figure 4.5: *A class AB output stage with MOSFETs (left) and one with quadruple output power (right).*

First of all, two MOS transistors in series are needed to prevent voltage break-down. This doubles the on-resistance, and two such combinations in parallel are needed to truly get an output power four times that of the original stage. From Figure 4.5 it can easily be seen that the area per output power has stayed exactly the same. Similarly, it is indifferent if the class AB stage is realised as a single ended or bridge configuration. See Figure 4.6. Transforming the bridge stage to a single ended stage, there are still two transistors in series, because of the increased voltage requirements. Therefore, the R_{on} stays the same, and the total required number of transistors stays the same.

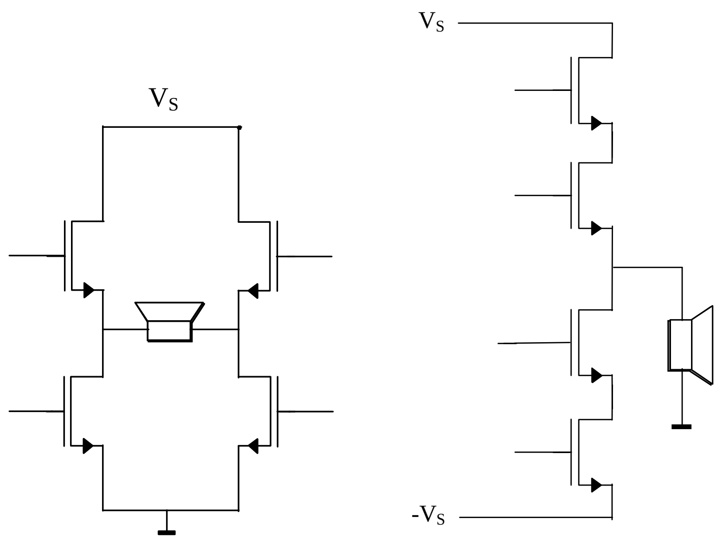

Figure 4.6: *Class AB bridge output stage (left), and single ended output stage (right).*

The simple concept of putting transistors in series and parallel to obtain the correct R_{on} and V_{BR} (breakdown voltage) has a theoretical basis. To half the on-resistance of a MOS transistor, it can simply be made twice as wide, requiring twice as much area. The breakdown voltage is a bit more complicated. There are several places where breakdown in DMOS transistors can occur. In an LDMOS (Lateral DMOS), the breakdown voltage is usually determined by avalanche breakdown of either the reversed body-epi junction or the epi-substrate junction. The latter is usually the highest, and sets the upper limit of the transistor's breakdown voltage. Below this value, the body-epi junction is the determining factor. Usually, LDMOSTs are used in this region, because otherwise a reduction of the drift region length could lower the R_{on} at no breakdown voltage penalty. In this region the breakdown voltage is proportional to the drift region length [59]. In VDMOS transistors, avalanche breakdown in the body-epi junction usually determines the transistor's breakdown voltage. Much work has been done in the field of optimising the on-resistance for a given breakdown voltage, considering cell-spacing and doping profiles. In general, the R_{on} per area rises with V_{BR}^2. [58],[61]. This is exactly what is demonstrated in Figure 4.5 and Figure 4.6. Only for lower voltages the lithography is the limiting factor.

Chip area for AB/D combinations

Let us consider a serial AB/D combination as displayed in Figure 4.7. The switches have been replaced by MOS transistors. These switch transistors must have dimensions in the same order of magnitude as the output transistors. Suppose a normal class AB amplifier has output transistors with area 1 as shown in Figure 4.7. When it is transformed to a serial AB/D combination, there are twice as many transistors, all in the signal path. In order to give this amplifier the same maximum output power as the original class AB amplifier, all transistors must have half their original R_{on}, requiring twice as much area per transistor. When we compare the necessary chip area for the class AB amplifier and the combination, we see that the needed chip area has increased with a factor 4. This makes the combination commercially very unattractive.

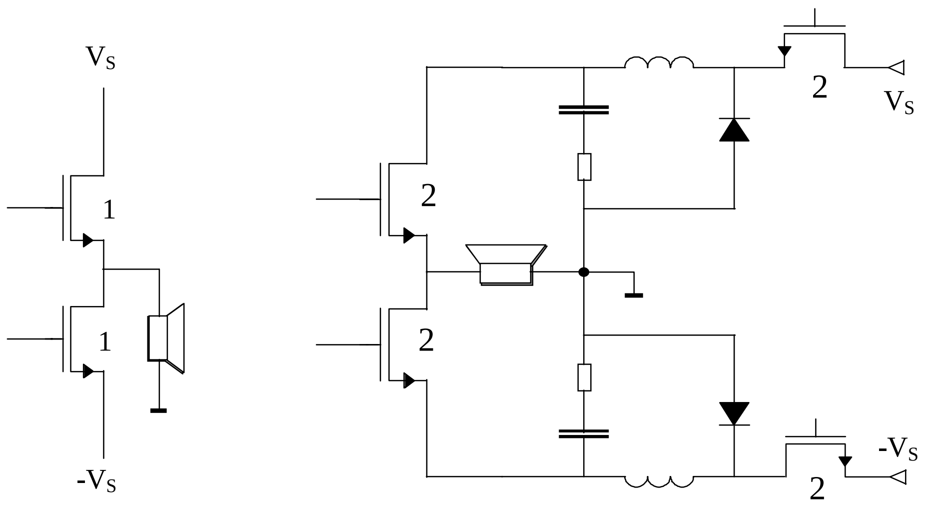

Figure 4.7: A normal class AB stage (left) and a AB/D series combination (right).

For the AB/D series bridge combination, the ratio is milder. Figure 4.8 shows that the total required area is almost a factor 2 higher. This is still a disadvantage.

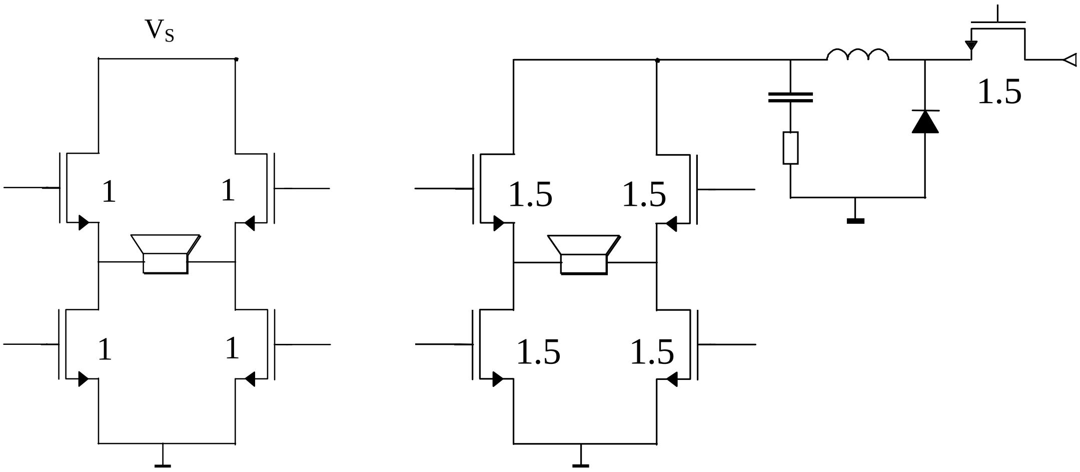

Figure 4.8: A normal class AB bridge stage (left) and a AB/D series bridge combination (right).

4.2.3 Bridge topology

In a bridge amplifier, there are already 2 transistors in series. When we replace one half of the bridge by a class D amplifier, the chip area does not have to increase. See Figure 4.9. The voltage across the output transistors of the left bridge half (the class AB amplifier) must be small. This is only possible when AB's output signal is close to the supply rail. Figure 4.10 shows the necessary waveforms for a low dissipation.

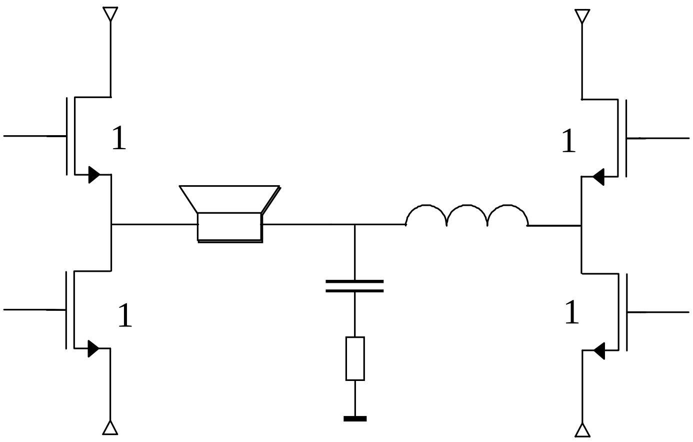

Figure 4.9: An AB/D bridge combination that requires the same chip area as a normal class AB bridge amplifier.

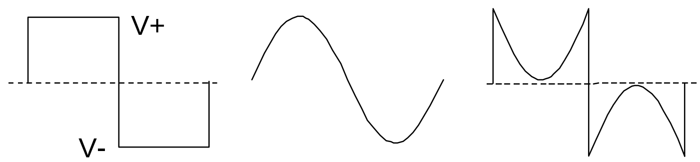

Figure 4.10: The waveforms of the class AB stage, the load, and the class D stage respectively.

The resulting amplifier can have the same dissipation as other series concepts. It is especially attractive in environments with a high integration level, where chip area and the number of external components are important. The demands on the class D amplifier are similar to those of other series concepts, so a self-oscillating class D amplifier is a practical choice. Reduction of the switching ripple, however, is only limited. This is no problem in applications where the amplifier is near to the load, like active (computer) loudspeakers, or when the class D amplifier is realised as a parallel class AB/D amplifier (see Chapter 5). Chapter 5 is concerned with a further analysis of this topology and realisation of a prototype.

4.3 Amplifiers in parallel

A parallel combination of amplifiers does not suffer from inefficient chip area usage, since every stage contributes to the output current. Only when one or more stages are not used, or not used to their maximum, the combination is not efficient regarding chip area. Often, however, the idea is to have a large amplifier that delivers the main power, and a small amplifier that corrects the errors made by the large amplifier. In that situation, at most the chip area of the small amplifier may be superfluous.

4.3.1 Problems with amplifiers in parallel

To get an idea of the difficulties associated with running more amplifiers in parallel, look at Figure 4.11. This is a very basic parallel combination. D is a complete class D amplifier, including modulator, switches, and filter. With $R_D>R_{AB}$, most output power is provided by the efficient D, while AB corrects possible errors.

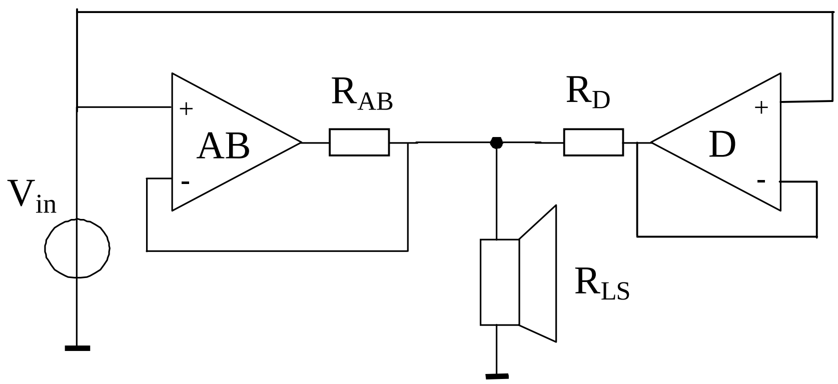

Figure 4.11: A simple parallel combination of a class AB and a class D amplifier.

The first point of attention is offset. For an efficient amplifier, R_D must be chosen much smaller than R_{LS}. In a practical situation R_{LS} is e.g. 4Ω. An arbitrary maximum value of R_D in that case would be about 0.4Ω. At that value, 10% of the output power is wasted in R_D. With a typical output voltage offset of 10…100mV, this leads to a cancellation current of 25…250mA. This current crosses the full supply voltage, causing a dissipation of several watts. Any gain mismatch between the two amplifiers creates the same problem.

A second problem is associated with the delay of the class D filter. When the output of D is not in phase with AB, large correction currents will flow, although it is not a matter of distortion or amplitude correction. Suppose the output signals of the amplifiers are $A\cos(\omega t-½\phi)$ and $A\cos(\omega t+½\phi)$ respectively. The difference voltage between the two amplifiers is: $A\cos(\omega t-½\phi)$ - $A\cos(\omega t+½\phi)$ = $2A\sin(½\phi)$ $\sin(\omega t)$. Since any difference voltage appears across R_D, a small phase shift generates a large correction current. With the practical values above, a phase shift ϕ of 1° creates a correction current that is 17% of the associated output current. It is obvious that even with a compensation filter before AB's input, the variation of the phase shift depending on the load is too large (see Figure 4.2). A self oscillating class D amplifier does not produce any phase shift, but its ripple is fairly high (practical values around several hundred mV), also resulting in large correction currents.

4.3.2 Current dumping

The problems with running amplifiers in parallel can largely be overcome by the current dumping principle. Originally it was used for linear amplifiers to achieve a very low distortion. It can be viewed in many ways [8]. For the system in Figure 4.12 it is easy to proof that if $Z_A Z_D = Z_B Z_C$, V_{LS} is independent of V_D. Thus, D is allowed to have distortion, which will be corrected by AB. It should be noted that although distortion of D is corrected, V_{LS} is not necessarily equal to V_{in}, but remains dependent on Z_{LS}.

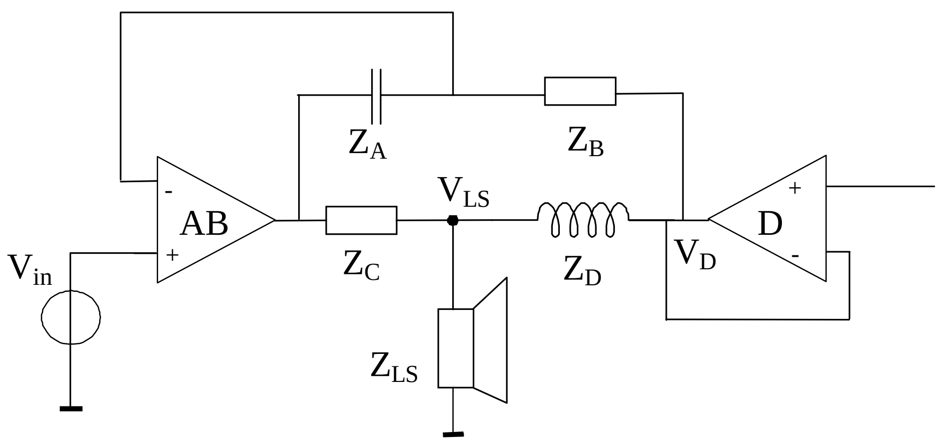

Figure 4.12: Current dumping principle with class D as a dumper.

By choosing Z_D inductive and Z_A capacitive, the power loss in Z_D can be very small thanks to its low impedance in the audio band. At the switching frequency of D, Z_D's impedance is much larger, which limits the correction currents that AB needs to supply.

The input of D can be connected anywhere, although certain nodes are advisable for efficiency reasons. When D's input is connected to V_{in}, offset problems can occur, especially since Z_B is an inductor that integrates the offset. When D's input is connected to AB's output, a feedback mechanism is constituted that activates D when AB has to take corrective action.

The exact choice of D is still free. When a standard class D amplifier is used, a compensation filter is necessary before AB's input, and even then stability problems occur [15]. D could also be a simple half bridge switch. However, on one hand Z_D should be large enough to minimise the high frequency ripple current, while on the other hand it should be small enough to deliver power in the audio band to Z_{LS}. It appears that it is not very well possible to meet both demands. Calling Z_C=R_C, Z_D=L_D, etc., calculating the frequency transfer results in:

$$\frac{V_{LS}}{V_{in}} = \frac{1 + j\omega\left(\dfrac{L_D}{R_C}\right)}{1 + j\omega\left(\dfrac{L_D}{R_{LS} // R_C}\right)}$$

constituting a pole and a zero. These should be placed above the audio band. By assuming a minimum R_{LS} of 4Ω and by choosing R_C=8Ω, L_D must be smaller than 10μH to obtain a bandwidth of 40kHz. At a switching frequency of 500kHz, the resulting current ripple in L_D is too large to obtain a good efficiency. When we choose a self-oscillating class D amplifier for D, simulations confirm that the current ripple is acceptable owing to the lower ripple at V_D. The increased complexity of the system makes it less attractive. The disadvantage that remains is the connection between L_D, Z_{LS} and the frequency transfer.

4.3.3 Reduction of AB's output current

The main goal of connecting linear and switching amplifiers in parallel, is to maximise the part of the output current provided by the efficient class D amplifier. In other words, the output current of the linear amplifier must be minimised [43]. This idea is the basis of Figure 4.13. The class AB amplifier, with its feedback taken from the output, determines the output voltage. AB's output current is sensed by 'A', and applied to one of inputs of the class D amplifier, where it is compared to zero. Consequently, D's output signal is adjusted in such a way that it minimises AB's output current.

The class D amplifier must have a current source character, as it is not wise to connect two voltage sources in parallel. When D has a voltage source character, it may be connected by means of a coupling network, which can be a resistor, a coil, or a more complex network. In the most basic form, D is a half bridge switch with a coil in series to the output.

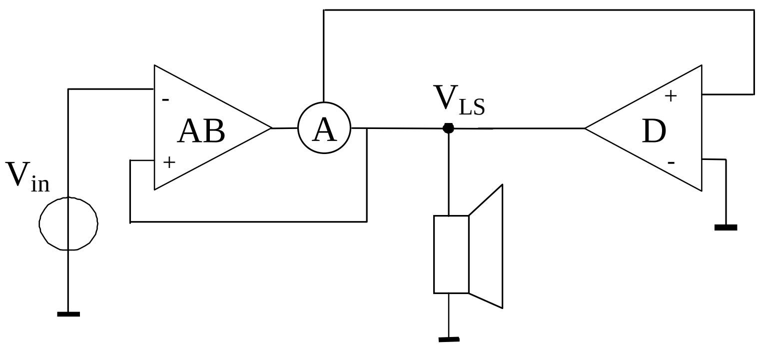

Figure 4.13: A parallel combination where D is controlled by AB's output current.

The value of the coil is only related to the power D can deliver, and not to the frequency transfer, as the output voltage is determined by AB. This is a distinct advantage over the current dumping approach. A further analysis of this principle and the realisation of two prototypes can be found in Chapter 6.

4.4 Conclusions

The supplementary qualities of linear and switching amplifiers have inspired people to design amplifiers that are a combination of those concepts. The traditional series approach, however, suffers from a chip area disadvantage compared to classic linear amplifiers with the same output power. A new series approach is a bridge amplifier which contains a linear and a switching halve. Theoretically, it can have the same dissipation as other series topologies without the increase in chip area. The usefulness of this attractive feature is further explored in chapter 5, including the realisation of a prototype.

In parallel AB/D combinations, all extra chip area can be used to increase the output power. When amplifiers are simply connected in parallel, many problems are encountered. The well-known current dumping topology is a better alternative, but prone to instability. Controlling the switching amplifier by the output current of the linear amplifier, is less complex. Therefore, chapter 6 is concerned with a further analysis of this idea, and a description of two realisations.

5

Realisation of a class AB/D bridge amplifier

5.1 Introduction

This chapter describes the design and realisation of a bridge amplifier that consists of a linear amplifier on one side and a switching amplifier on the other. This offers distinct chip area advantages over other series approaches, where a switching amplifier generates the power supply for a linear amplifier (see Chapter 4). The common mode level is adapted such that the linear amplifier has a low dissipation, giving the total system a high efficiency.

The structure of the chapter is as follows: In section 5.2 the circuit principle is explained. Section 5.3 deals with the analysis of the circuit in more detail and with some design choices. Section 5.4 and 5.5, finally, are concerned with the measurement results and the conclusions.

5.2 Circuit principle

The circuit principle is based on the well known technique of keeping the voltage drop across the output transistors of a linear amplifier small. The circuit principle is shown in Figure 5.1. For simplicity, let us consider the feedback factor β=1. The amplifier consists of a class AB amplifier (AB) on one side, and a class D amplifier (D) on the other side. AB is represented as a power OPAMP connected as a non-inverting amplifier. The + input of AB is the audio signal. The - input is the voltage over the load, so that the output voltage between the bridge outputs is determined by AB. D is a complete class D amplifier, including modulator, switches, and filter. A control circuit creates the input signal for D. The output of the control circuit is the inverted audio signal added to a common mode voltage.

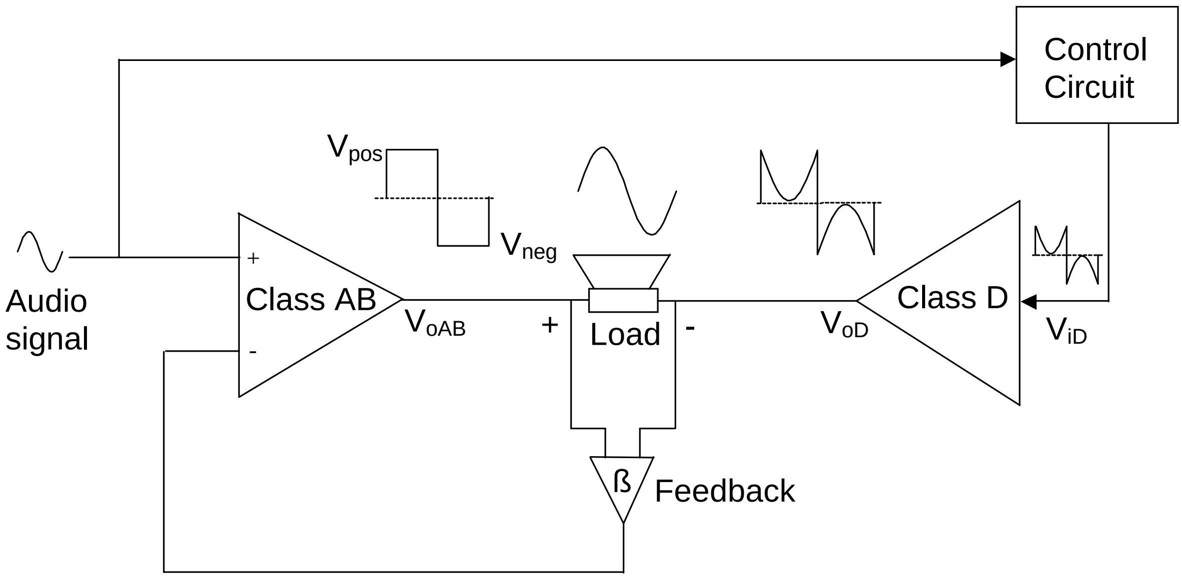

Figure 5.1: The circuit principle

When the audio signal is positive, the common mode voltage equals V_{pos}, a voltage close to the upper clipping point of AB and D. When the audio signal is negative, the common mode voltage equals V_{neg}, a voltage close to the negative clipping point of AB and D. Figure 5.1 shows the corresponding waveforms when the input is a sine wave. Note that the possible output voltage ripple and distortion of D are small and therefore not visible in Figure 5.1.

Independent of the output of D, created by the control circuit, AB will try to keep the load voltage equal to the input signal. This will force the output of AB to be either V_{pos} or V_{neg}. This is close to either of the supply lines, so the voltage drop across the output transistors of AB is small, resulting in a low power dissipation. Since AB determines the load voltage, and since there is still some room between V_{pos} and V_{neg} and the clipping point of AB, AB can correct most of the distortion of D. Since the output of D is derived directly from the input signal, the stability of the amplifier is only determined by AB and its feedback.

5.3 Circuit analysis and design considerations

5.3.1 Timing of bridge switching

In the explanation of the circuit principle, the moment the common mode level of the bridge changes (from now on referred to as bridge switching) is controlled by the input voltage. This will ensure that the output voltage of AB is always close to one of the supply lines. For a resistive load impedance, this is no problem. For complex load impedances, however, it does not guarantee that the output voltage is always near the *preferred* supply line. As an example, suppose that a slightly capacitive load is connected (phase -30º). The left column of Figure 5.2 shows what happens. Most of the positive half wave before t_1, the output voltage of AB is high, and AB is sourcing current. A little time before t_1, AB's output current becomes negative.

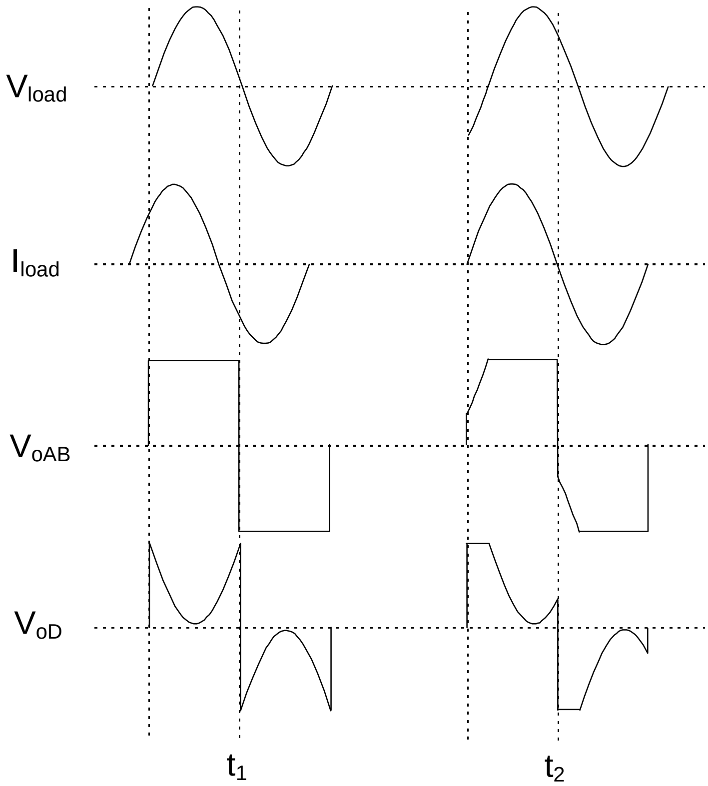

Figure 5.2: Wave forms with bridge switching as a result of load voltage (left) or -current (right).

The output voltage of AB and D are still high, however. The load current is still sourced by the upper output transistor of AB, which now has the full supply voltage across its terminals. This will lead to extra dissipation. This situation is not cancelled until t_1, when the common mode level changes because the input voltage becomes negative.

In the situation described above, even a small phase shift of the load current gives rise to short periods with almost the full supply voltage across AB's output transistors. This is avoided when the moment of bridge switching is determined by AB's output current. This is shown in the right column of Figure 5.2. Most of the half wave before t_2, the output voltage of AB is high, and it must source current. At t_2, the load current (equal to AB's output current) becomes negative, and the common mode level changes. The output voltage, however, is still positive, so that D should have an output voltage of V_{neg}-V_{in}, which would be below its clipping point. This is not possible, so D's output goes to approximately V_{neg}. AB, in its attempt to make a positive output voltage, does not follow D to V_{neg}, but stops a little higher. At this point, AB starts sinking current when there is a voltage across its lower output transistor. The magnitude of this voltage, however, is much lower than in the previous case, namely the output voltage at t_2 instead of the full supply voltage minus the output voltage at t_1.

The difference in dissipation between the two switching strategies also holds for larger phase shifts. With the practical values used in the final realisation, we can make Figure 5.3, that shows the simulated dissipation in AB as a function of the phase of the load current at full power (100W). In this plot, the supply voltage is 40V, and the load impedance is 8Ω. The class AB amplifier is ideal, with an output voltage swing from 0 to 40V. The dissipation is plotted for both switching methods. As can be seen, a considerable amount of power is saved by output current controlled switching when the load impedance is complex.

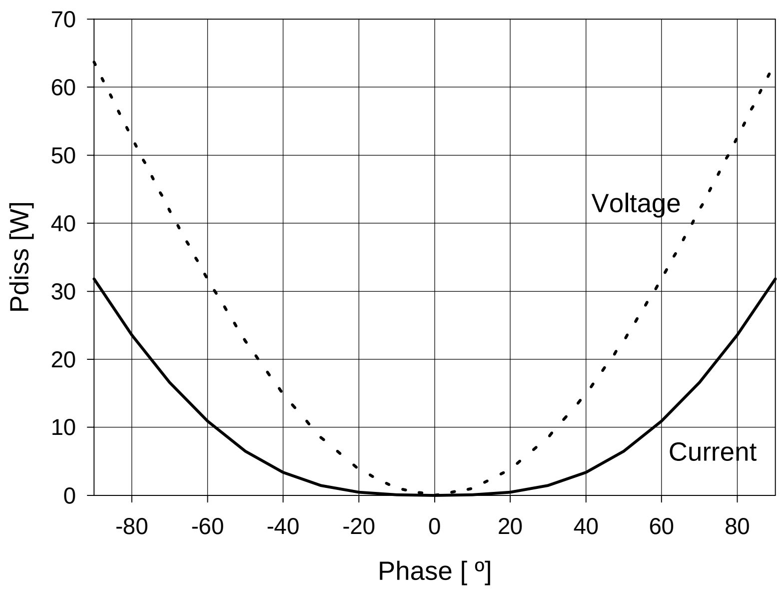

Figure 5.3: Simulated power dissipation of AB caused by a non-zero phase of the load current for bridge switching as a result of load voltage or -current.

One could wonder if AB is always able to produce the desired output voltage if we use the output current to control bridge switching. Suppose the common mode level is randomly changing. Table 3 shows the maximum possible V_{out} for various common mode levels. Every row is a possible combination of input voltage V_{in} and common mode level CM. E.g. the first row describes the case that $V_{in}>0$ and $CM=V_{pos}$. The output voltage of D is $V_{pos}-V_{in}$, and the maximum output voltage of AB is V_{pos}. Thus, the maximum voltage over the load is V_{in}, which is just enough. In the same way, it can be seen that for all possible combinations of CM en V_{in}, the desired output voltage can always be generated.

On the basis of Figure 5.3 and Table 3, bridge switching as a result of output current is chosen as the basis for a realisation. A small offset of 200mA is added to prevent the circuit from bridge switching when there are only small signals present.

V_{in}	CM condition	V_{oD}	max/min V_{oAB}	max/min (V_{oAB}-V_{oD})
$V_{in}>0$	$CM=V_{pos}$	$V_{pos}-V_{in}$	$V_{oABmax} = V_{pos}$	$V_{o,max} = V_{in}$
	$CM=V_{neg}$	V_{neg}	$V_{oABmax} = V_{pos}$	$V_{o,max} = V_{pos}- V_{neg}$
$V_{in}<0$	$CM=V_{pos}$	V_{pos}	$V_{oABmin} = V_{neg}$	$V_{o,min} = V_{neg}- V_{pos}$
	$CM=V_{neg}$	$V_{neg}-V_{in}$	$V_{oABmin} = V_{neg}$	$V_{o,min} = V_{in}$

Table 3: Maximum possible V_{out} for various Common Mode (CM) levels.

So far, we have assumed that the output of D is an instantaneous copy of its input. D’s output filter, however, may cause a phase shift at the higher audio frequencies. It is easy to see that this can lead to situations in which AB can not create the desired output signal. This occurs at high output powers and high frequencies, so it is a possible cause of transient intermodulation distortion. It is important to keep this in mind when designing D.

5.3.2 Dissipation

The power dissipation of the total amplifier is the sum of the following components:

- Dissipation of the class D amplifier, consisting of conduction losses, switching losses and capacitive losses.
- Power dissipation caused by AB’s quiescent current. The switching of the common mode level can be regarded as a high frequency, high power signal for AB. Depending on the design, this can lead to a dynamic increase in quiescent current. In the class AB amplifier used in the realisation of section 5.4, this leads to an extra 1W power dissipation at 1kHz.
- Dissipation as result of a capacitive or reactive load impedance, as discussed in the previous section and shown in Figure 5.3. When we bear in mind that a normal loudspeaker rarely has a phase shift beyond -50º…50º [20], and that we use bridge switching as a result of current, the power dissipation due to this phenomenon will probably not exceed 5W. Please note that a normal class AB (bridge) amplifier also dissipates more when the load impedance is complex [20].
- Losses caused by the minimum voltage drop across AB’s output transistors. For the same practical values as in the previous section (V_{supply}= 40V, R_{LS}= 8Ω), Figure 5.4 shows the power dissipation of the class AB amplifier for several values of the minimum voltage drop across AB’s output transistors. This voltage drop is important for the total dissipation. The class AB amplifier we use can swing to within 4V of either power supply at maximum current.

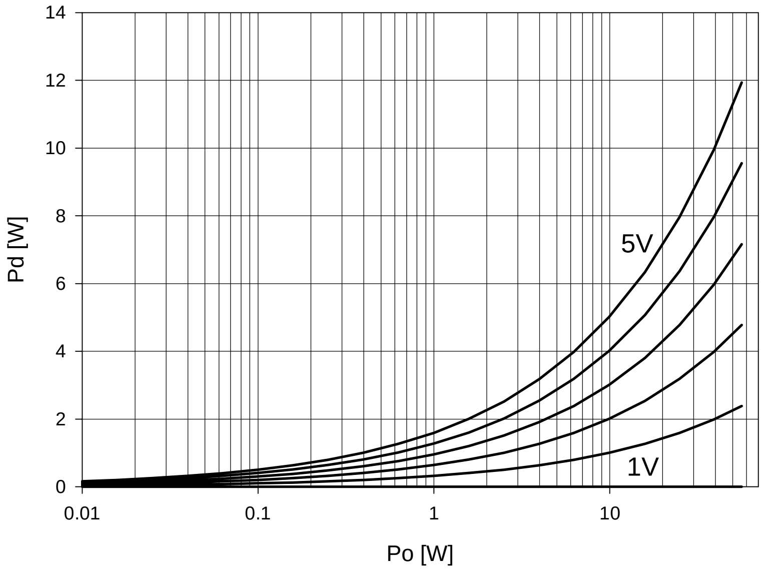

Figure 5.4: Simulated power dissipation of AB caused by the minimum required voltage drop across the output transistors (0-5V).

5.3.3 Common mode rejection

The common mode level of the bridge changes abruptly during bridge switching. The change is initiated by the class D amplifier after which the class AB amplifier follows. This makes high demands on the Common Mode Rejection Ratio (CMRR) of the feedback network and on the ability of AB to correctly follow the sudden change.

First, let us consider the CMRR of the feedback network, assuming that AB is perfect (infinite gain, infinite bandwidth, zero output impedance). The change in common mode level of the bridge is not instantaneous. In the control circuit of Figure 5.1, the common mode voltage to D’s input is filtered with a 10kHz first order low pass filter before the inverted audio signal is added. When the output of D changes V_{pos}-V_{neg}, the feedback network will generate a voltage

$$V_\beta = \frac{\beta\left(V_{pos} - V_{neg}\right)}{CMRR@10kHz} \qquad (1)$$

Because the feedback network is inside the loop, this voltage will appear across the load. This voltage is independent of the actual output voltage, so it is important that the bridge does not switch at low output voltages, since this would cause a high distortion. In the practical realisation, the feedback network is an audio line receiver with a CMRR of 90dB at 10kHz. V_{pos}-V_{neg} is 32V, and the output current at which the bridge switches is 200mA (V_{switch}=1.6V with an 8Ω load). The maximum distortion caused by the feedback network can now be calculated as V_β/V_{switch}= 0.006%. This is low enough not to be the limiting factor in the practical realisation.

Secondly, let us consider the problem that *is* the limiting factor, namely the limited bandwidth of AB. This can be illustrated with Figure 5.5.

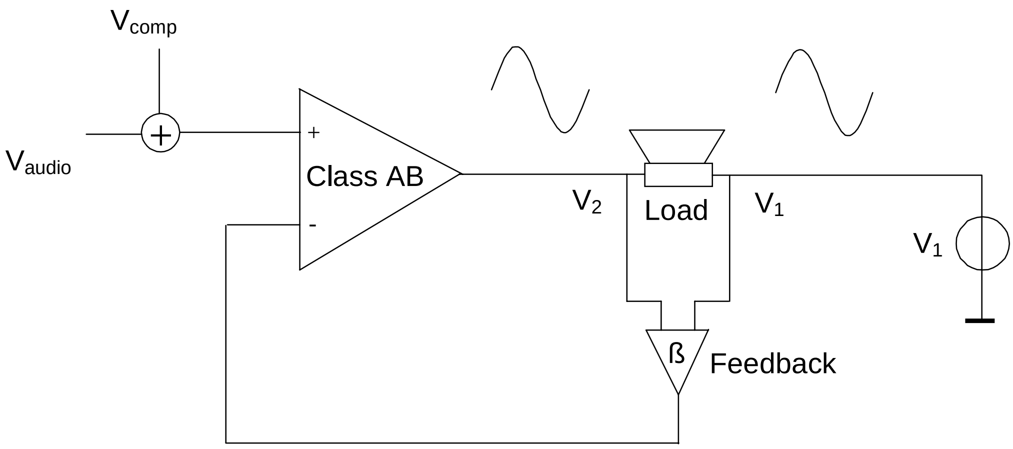

Figure 5.5: Common mode swing handling capability of AB.

The class D amplifier is represented by V_1, generating a sinusoid. Suppose $V_{audio}=V_{comp}=0V$. Ideally, AB generates $V_2=V_1$. In practice, however, V_2 will lag behind, causing a voltage difference $(V_2-V_1) \neq 0$. When AB has a first order behaviour above 0dB, the difference V_1-V_2 is:

$$A_{AB} = \frac{A_0}{1+s\tau_1} \qquad\qquad V_1 - V_2 \approx \frac{V_1}{1+\beta A_0} \cdot \frac{(1+s\tau_1)}{\left(1+s\frac{\tau_1}{1+\beta A_0}\right)} \tag{2}$$

This is represented in Figure 5.6 as a bode plot.

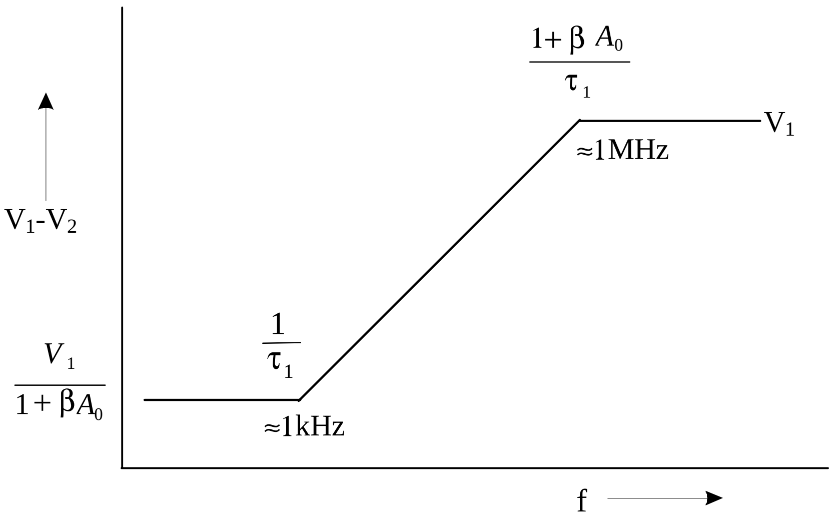

Figure 5.6: V_1-V_2 in the circuit of Figure 5.5.

The practical values for A_0, β and $1/2\pi\tau_1$ are approximately 10,000, 0.1 and 1kHz (measured from AB). A V_1 of 32V@10kHz now causes a V_1-V_2 of 0.32V, which is far too high (THD = 20%). Since the distortion is exactly known, a solution to this problem is feedforward error correction.

In Figure 5.6, we see that between 1kHz and 1MHz, the difference increases with a first order slope. A simple high pass filter with a similar frequency response can add the difference to V_2 via V_{comp} (see Figure 5.5 and Figure 5.7). For ‘exact’ compensation it is necessary that $R_cC_c=\tau_1/A_0$, which means that the corner frequency of the compensation network must be equal to the unity gain frequency of AB. With this technique it is possible to reduce the distortion significantly. Like any feedforward compensation systems, the compensation depends on the matching of components. When R_c, C_c, or the unity gain frequency of AB deviate from their nominal value due to component spread or temperature changes, the distortion of the total amplifier increases. The next section discusses how this affects the measurements.

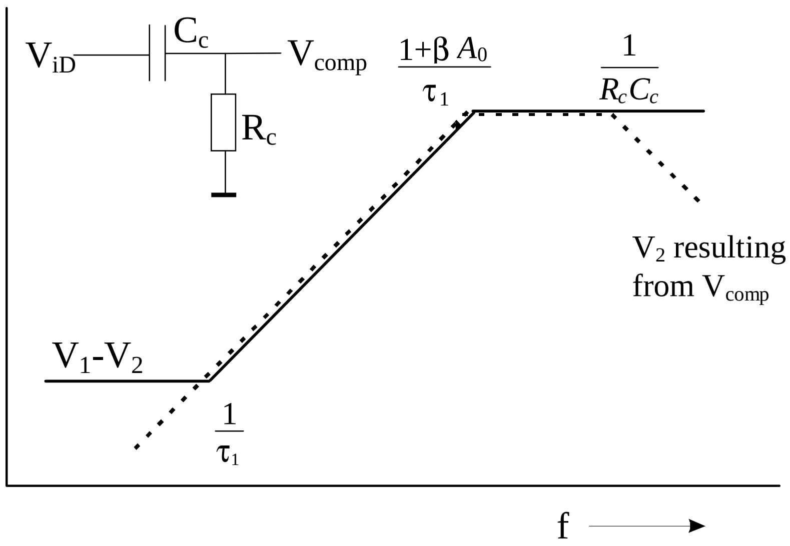

Figure 5.7: Feedforward compensation

5.4 Implementation of a prototype

5.4.1 Realisation

The amplifier was realised with integrated subcircuits. AB is a standard audio amplifier IC, and the feedback network is a audio line receiver. Two different realisations were tested. The first has a self oscillating class D amplifier as used in [47, 49, 51]. It was chosen because it does not exhibit any phase shift between input and output, which is essential in the AB/D bridge amplifier. It also has a considerable output voltage ripple. Compensation of the ripple by the class AB amplifier is limited because of the low loop gain at high frequencies. This configuration can be useful in applications where EMI on the loudspeaker cables is not a big problem (active loudspeakers).

We will discuss the second realisation more elaborately here. For the class D part, the amplifier presented in [53] is used. This amplifier is a combination of class AB and class D itself, but can successfully be used in the bridge configuration presented here to eliminate the need for a second class D part. Like the self-oscillating class D amplifier, it has no phase shift between input and output. The symmetrical supply voltage is +/- 20V. The common mode level swing V_{pos}-V_{neg} is set at 32V (+/- 16V), leaving 4V across AB's output transistors. A smaller voltage drop is possible, but increases the distortion of the amplifier because the region near the clipping point is strongly non-linear. The load is 8Ω.

5.4.2 Measurements

The dissipation for sinusoidal signals is shown in Figure 5.8. The maximum output power is 75W. At this power the distortion is 1%, and the output of AB is pushed below V_{neg} (and above V_{pos}) at the moments of the largest output voltage. For comparison, the dissipation of a normal class AB bridge amplifier is also plotted. The AB amplifiers used are the same as those used in the AB+D bridge amplifier. Notable is the 'bump' at 0.2W. Only above this power there is bridge switching, giving rise to a dynamic increase in AB's quiescent current. For output powers >3W, the power dissipation of AB+D in bridge is approximately half as much as AB in bridge.

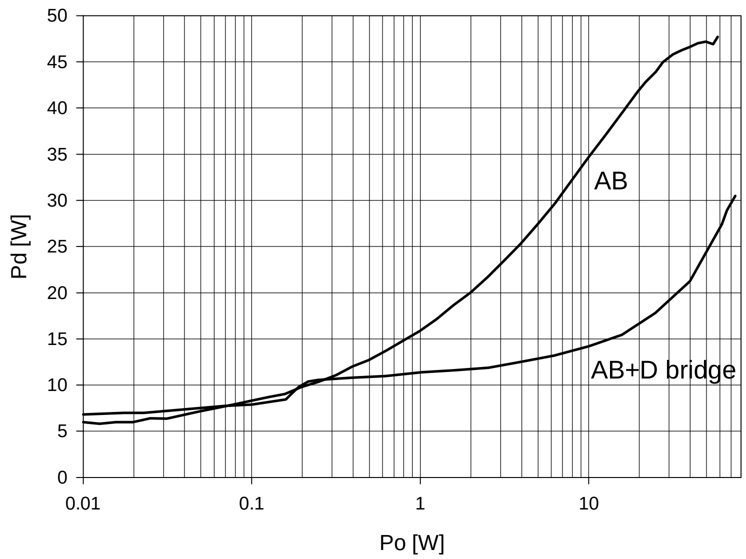

Figure 5.8: Measured dissipation for a 1kHz sine wave.

Figure 5.9 shows the dissipation for the IEC-268 test signal. This signal has the amplitude- and frequency characteristics of an average audio signal, and is well suitable for predicting the efficiency of audio amplifiers (Chapter 2). It is used up to an output power of 35W. At that power, audio signals (and consequently the IEC 268 test signal) clip a large percentage of the time. We see that at low power levels, the advantage of using the AB+D bridge amplifier is marginal. At higher power levels, the dissipation is approximately 50-60% of the class AB bridge amplifier. At moderate power levels <10W, the quiescent current of the AB amplifiers are an important part of the total power dissipation.

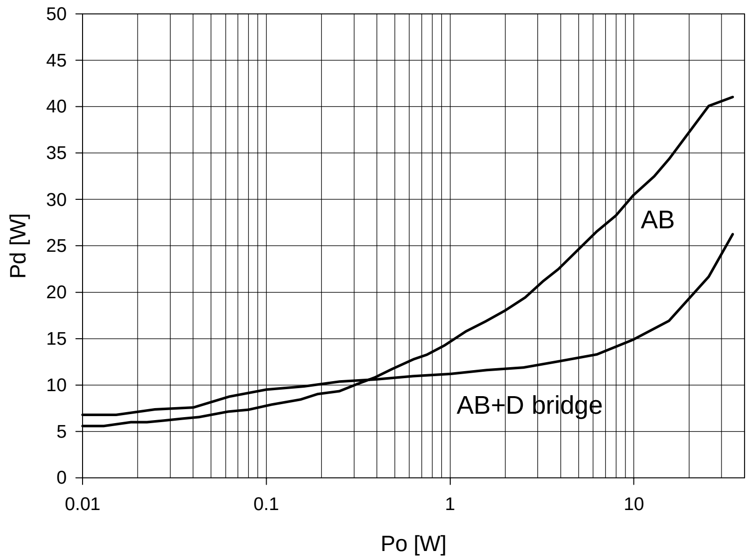

Figure 5.9: Measured power dissipation for the IEC 268 test signal.

Figure 5.10 shows the Total Harmonic Distortion plus Noise (THD+N) as a function of output power. Here, too, the point where bridge switching begins is clearly visible. The measurements of Figure 5.10 were performed on amplifiers with the feedforward error correction. When the error correction is exactly trimmed, the distortion is even lower. For practical applications this is not very useful, as there is a certain spread in the unity gain frequency of the class AB amplifiers which would require trimming of every amplifier.

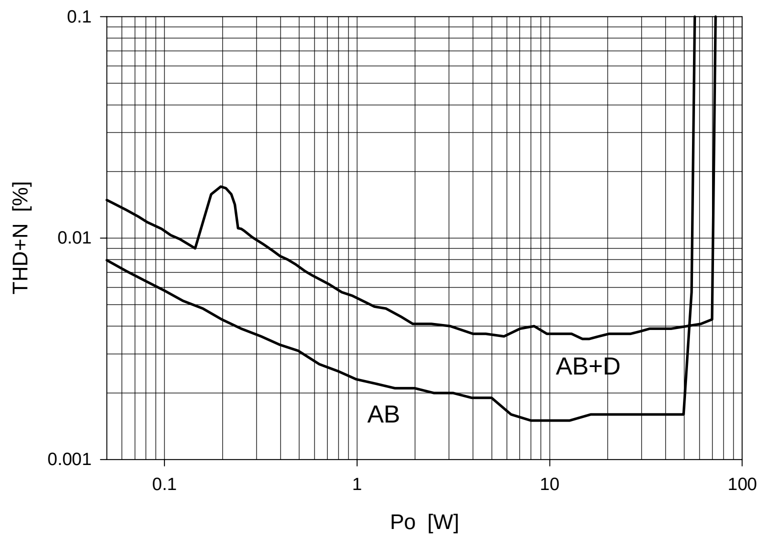

Figure 5.10: THD+N vs. output power @ 1kHz, filter 22Hz-22kHz

More important, temperature changes have an effect that is even larger. Experiments with a number of class AB IC's at temperatures between 20-80 ºC revealed that the distortion in Figure 5.10 was deteriorated with at most a factor 3, so that the worst case distortion was 0.05%. The distortion stays low at other frequencies, as is shown in Figure 5.11.

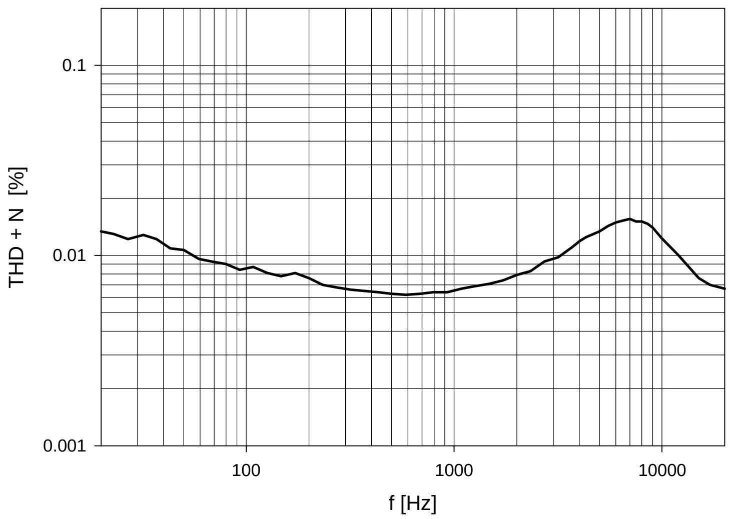

Figure 5.11: THD+N vs. frequency @ 1W, resistor load, filter 22Hz-22kHz.

At frequencies higher than 1kHz (the open loop pole of AB) the distortion rises, and seems to decrease above 7kHz, which is a result of the filter of the distortion analyser. Figure 5.12 shows that when this measurement is repeated with the load disconnected, the distortion is higher.

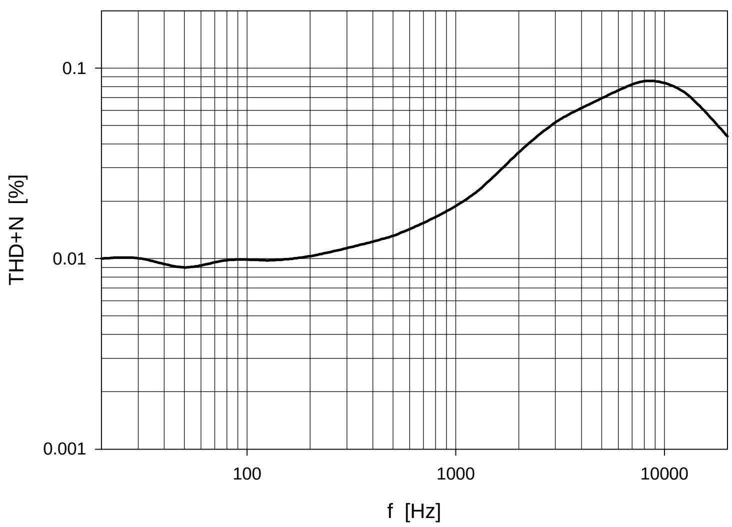

Figure 5.12: THD+N vs. frequency @ Uo=2.89V, no load, filter 22Hz-22kHz.

Due to the absence of bridge switching (no output current), the common mode level is at a constant V_{neg}, causing D to clip at positive output voltages. Although this is corrected by AB, the input and output of D are not equal, so that the error correction generates the wrong signal. In future realisations this can be solved by taking the input of the compensation network from the output of D instead of the input.

5.4.3 Discussion

Although the dissipation of the AB/D bridge amplifier is lower than the dissipation of a class AB bridge amplifier, it is hardly worth the complexity of the realisation. The relatively high dissipation is mainly caused by the necessary voltage drop across the output transistors of AB. At an output power of 60W, 10W of the 27W dissipation is caused by the 4V voltage drop (Figure 5.4). There is clearly room for improvement by designing a class AB amplifier optimised for rail-to-rail output. At low output powers, a reduction of the quiescent current of the class AB amplifier is favourable to reduce the dissipation. Nowadays, class AB amplifiers can easily have a quiescent power dissipation that is a factor 2 lower. At higher output powers, this also decreases the power dissipation of the AB/D combination relative to class AB.

Concerning distortion, the accuracy of the error correction is the main point for improvement. By matching the unity gain frequency of the class AB amplifier to the corner frequency of the compensation network, the distortion can be made insensitive to temperature and process spread. It is also possible to use a different error feedforward mechanism by adding D's output signal to AB's output signal. This way the feedback signal does not contain abrupt changes. The common mode changes of the output signal, however, still put high demands on the CMRR of the feedback network. Realising high CMRR figures in an IC realisation is not trivial.

5.5 Conclusions

A class D amplifier can be used in bridge configuration with a linear amplifier. This offers a chip area advantage over other serial AB+D combinations. Switching the common mode level of the bridge results in a low dissipation (approximately 2 times lower than class AB) and a low distortion (<0.05%). The prototype demonstrates the functionality of the approach, but it is clearly not the best achievable realisation. However, there are good possibilities to optimise the design. The number of realisation problems that have to be overcome like common mode rejection and error correction, make this amplifier not an easy candidate for integration.

6

Realisation of a class AB/D parallel amplifier

6.1 Introduction

This chapter deals with an amplifier topology that contains a switching amplifier and a linear amplifier in parallel. The linear part controls the output voltage, while most of the output current is provided by the switching amplifier. Thus, the dissipation can be low.

In the next section (6.2), the circuit principle is explained. A further analysis and design considerations are discussed in section 6.3. Sections 6.4 and 6.5 describe two implementations of the circuit. Finally, section 6.6 and 6.7 are concerned with future possibilities of the circuit and the conclusions.

6.2 Circuit principle

The circuit principle is shown in Figure 6.1. A class AB amplifier (AB) is directly connected to the output, controlling the output voltage. A switching part (D), consisting of the two switches $SW_{1,2}$ and the coil L_1, is connected to the same output. The control signal to D is derived from AB's output current. The output current of AB is measured by A.

The system is self-oscillating: if switch SW_1 is closed, and SW_2 is open, the current through L_1 increases linearly with time, and the part not needed as load current flows right into AB. This is measured by A, and when I_{AB} exceeds a certain (small) value, SW_1 is opened, and SW_2 is closed. Then, the current through L_1 decreases, etc. Because I_{AB} oscillates between two small threshold currents (+/- I_{thr}), the power dissipation in AB is small, while D delivers the main load current.

L_1 does not have to be very linear, because AB determines the output voltage. Figure 6.2 shows the typical waveforms when a sinusoid is applied to the input.

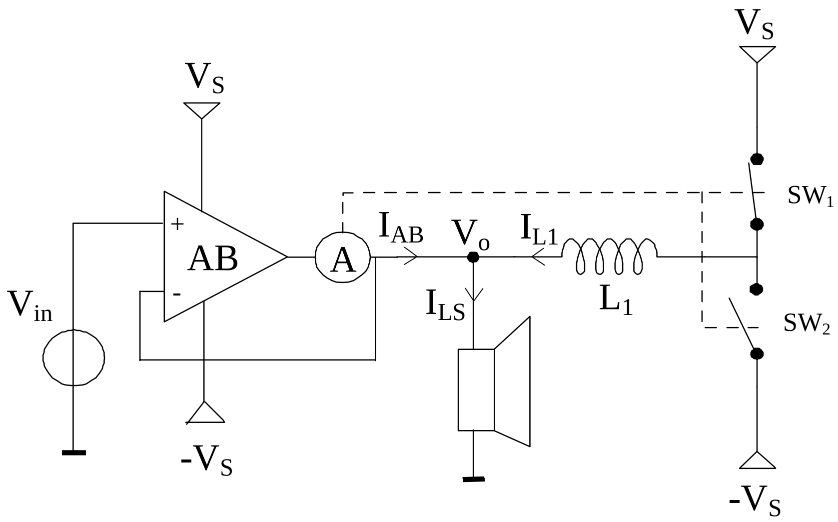

Figure 6.1: Basic amplifier structure

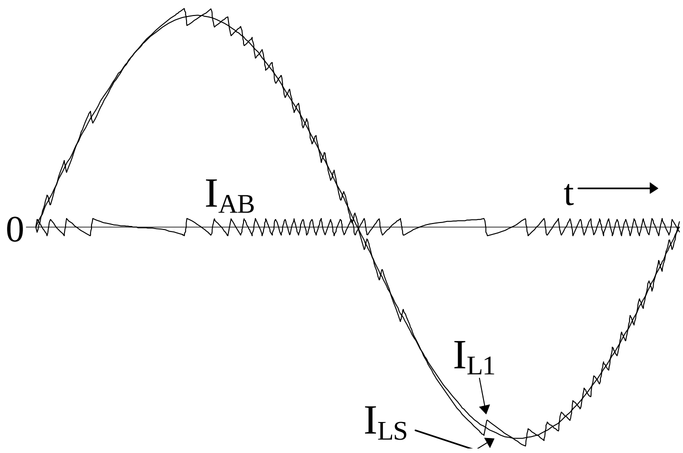

Figure 6.2: Typical currents when amplifying a sinusoid.

In the current dumping approach (Chapter 4), the value of the coil L_1 also influences the frequency transfer of the amplifier. This topology does not exhibit such a relation. When L_1 is chosen large, D can not deliver much power at high frequencies, but AB determines the output voltage and delivers the remaining part. Figure 6.3 shows the corresponding waveforms in that case.

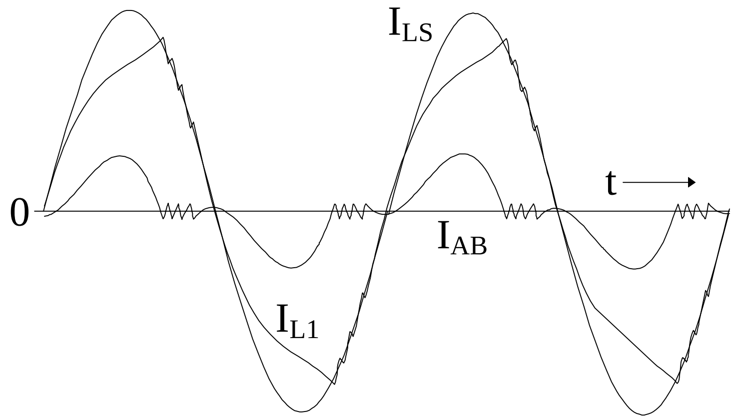

Figure 6.3: Typical waveforms if AB must supply part of the current.

6.3 Circuit analysis and design considerations

6.3.1 Current ripple and coil

The choice of the threshold current I_{thr} and the coil L_1 depends on three important parameters. The first is the intended switching frequency. Figure 6.4 shows a detail of the typical currents in the coil and in the loudspeaker during normal operation where D supplies all output current. (refer to Figure 6.1 for the meaning of the symbols).

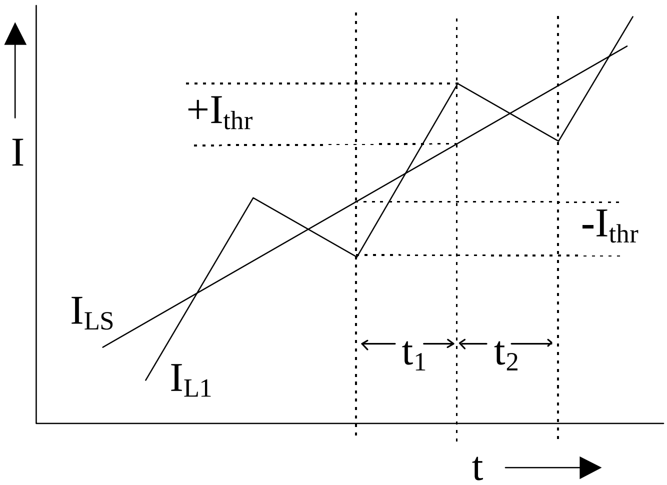

Figure 6.4: Typical currents (detail)

Suppose the thresholds of A are $+I_{thr}$ and $-I_{thr}$, as shown in Figure 6.4. The following equations hold during t_1 and t_2 respectively:

$$2I_{thr} = \left(\frac{dI_{L1}}{dt} - \frac{dI_{LS}}{dt}\right) \cdot t_1 = \left(\frac{V_S - V_o}{L_1} - \frac{dI_{LS}}{dt}\right) \cdot t_1$$

$$-2I_{thr} = \left(\frac{dI_{L1}}{dt} - \frac{dI_{LS}}{dt}\right) \cdot t_2 = \left(\frac{-V_S - V_o}{L_1} - \frac{dI_{LS}}{dt}\right) \cdot t_1$$

From this the oscillating frequency for any V_o and dV_o/dt can be calculated:

$$f_{switch} = \frac{1}{t_1 + t_2} = \frac{V_S{}^2 - \left(V_O + \frac{L_1}{R_{LS}} \cdot \frac{dV_O}{dt}\right)^2}{4 I_{thr} L_1 V_S} \tag{1}$$

Note that the highest switching frequency is at V_o=0. For large output voltages and/or slew rates, f_{switch} goes to 0. This happens when the amplifier clips or when D's power bandwidth is exceeded.

The power bandwidth of D is the second important parameter. The slew rate of I_{L1} should always be larger than the slew rate of I_{LS}. Suppose the output signal is a sinewave:

$$V_o = \alpha V_S \sin(2\pi f_{\sin} t)$$

with α the amplitude as fraction of the power supply voltage V_S. Then, the slew rate of I_{LS} is:

$$SR_{I_{LS}} = \frac{1}{R_{LS}} \cdot \frac{\delta V_o}{\delta t} = \frac{2\pi\alpha V_S f_{\sin} \cos(2\pi f_{\sin} t)}{R_{LS}}$$

For the analysis it is necessary to consider the first quarter of a period. It is not sufficient to consider the point where V_o=0, because the slew rate problems occur at a different output voltage. The slew rate of I_{L1} is:

$$SR_{I_{L1}} = \frac{V_S - V_o}{L_1} = \frac{V_S(1 - \alpha \sin(2\pi f_{\sin} t))}{L_1}$$

If L_1 has to deliver the main load current, the maximum slew rate of I_{L1} should be larger than the slew rate of I_{LS}. In that way, the maximum frequency at any given amplitude can be calculated. After some calculation we find:

$$f_{\sin,\max} = \frac{R_{LS}}{2\pi L_1} \sqrt{\frac{1}{\alpha^2} - 1} \tag{2}$$

When D is not able to provide the full load current, AB could supply the rest, as demonstrated in Figure 6.3. Of course, this deteriorates the efficiency.

The third important parameter is the quiescent power dissipation of the amplifier, which consists mainly of three components: the quiescent power of AB, the quiescent power of D, and the class D current ripple that is dissipated by AB. It is this last component we can influence significantly by choosing I_{thr}. The total quiescent power dissipation of the amplifier can be expressed as:

$$P_Q = P_{Q(AB)} + P_{Q(D)} + \tfrac{1}{2} V_S \cdot I_{thr} \qquad (3)$$

We have to make a choice for L_1 and I_{thr}, considering the output resistance of AB and the power dissipation of the total amplifier. A smaller L_1 will result in a higher power bandwidth of D, but also in a higher switching frequency. A high switching frequency will lead to a higher power dissipation due to higher switching losses. It can also lead to problems with AB's output resistance, which has to be low compared to the load resistance. A larger L_1, on the other hand, reduces the power bandwidth of D, leading to a higher dissipation at high output voltages at high audio frequencies. A large value for I_{thr} decreases the switching frequency but increases the quiescent power dissipation due to the larger ripple current.

6.3.2 Dissipation

The power dissipation in the amplifier consists of the following elements:

- Dissipation of the class D part consisting of switching losses, conduction losses, and capacitive losses in the switches $SW_{1,2}$ and their drivers. These occur in any normal class D amplifier, and can be quite small (< 10% of P_{omax}).
- The quiescent current of AB. The quiescent current depends on the design. The trend towards lower quiescent currents in linear amplifiers can be favourable in this case, although AB's output impedance must be low at the switching frequency. Generally, these are conflicting demands.
- Dissipation due to the current ripple in AB. The current I_{AB} alternates between $-I_{th}$ and $+I_{th}$, leading to a dissipation ½*I_{th}*V_S (see Equation 3). This dissipation can be kept small by choosing I_{th} small, although this increases the switching frequency. Together with the quiescent power dissipation of AB, this is the main factor that determines the quiescent power dissipation of the total amplifier.
- Extra dissipation in AB during transients that I_{L1} cannot follow. This depends on the output signal and the power bandwidth of D. Fortunately audio signals have a limited high frequency contents, so this dissipation can be low even if D's power bandwidth does not comprise the full audio band.

An advantage compared to linear amplifiers appears when the amplifier has a capacitive or reactive load. In that case, the load current and the load voltage are not in phase. In normal class AB and class G amplifiers, this leads to a higher dissipation [20]. The AB+D bridge amplifier of the previous chapter also exhibits a higher dissipation for complex load impedances. In the amplifier discussed here, a non-resistive load hardly changes the power dissipation.

In chapter 2, the dissipation of a parallel class AB+D amplifier is simulated to avoid long measurements. This concerns the slow switching parallel amplifier presented in section 6.4. The dissipation model for this amplifier consists of all components mentioned above, except the class D switching losses and capacitive losses. The model is quite straightforward like the models for linear amplifiers in section 3.2.5. Only the for the extra dissipation of transients above D's power bandwidth, a separate simulation with a small time step is used. The deviation between measured and simulated dissipation is less than 5%.

6.3.3 AB's output impedance

At first, the prototype of the amplifier was not stable for all loads, so a stability analysis was done. The system is guaranteed stable as long as the load is purely resistive and the amplifier AB is ideal. In practice, however, AB is not ideal; it has a non-zero output impedance. Furthermore, a loudspeaker may have a complex impedance, and especially the capacitive part may lead to instability.

Assume the loudspeaker and the output impedance of AB can be modelled as in Figure 6.5. R_{LS} is the loudspeaker's DC-resistance, and C_{LS} its parallel capacitance (this can also be the cable capacitance). The amplifier AB uses feedback to reduce distortion. Its loop gain has a first order behaviour above 0dB, and its dominant pole lies within the audio range. At DC, the loopgain reduces the output resistance to R_o (Figure 6.5). For frequencies higher than the pole frequency, the loopgain rolls off, leading to an apparent inductive output impedance L_o. Above the audio range (at f_{switch}) the inductive part of the output impedance dominates, so we can ignore R_o.

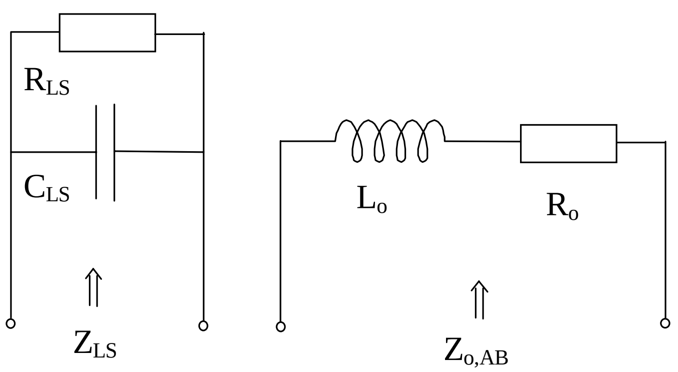

Figure 6.5: A model of the loudspeaker (left) and the output impedance of AB (right).

The stability of the total amplifier is analysed with the behavioural model shown in Figure 6.6. V_{switch} represents the common node of SW_1 and SW_2.

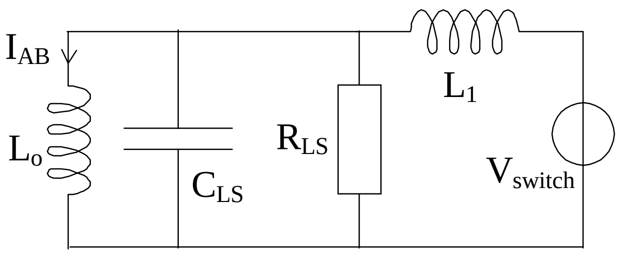

Figure 6.6: Behavioural model of the amplifier at f_{sw}.

To obtain the first-order transfer from V_{switch} to I_{AB} that our system needs, it is essential that this transfer is determined by L_1. Only then, the oscillation will be well defined as in Figure 6.4. For this, we calculate the admittance $Y=I_{AB}/V_{switch}$:

$$Y = \frac{1}{s\,L_1} \cdot \frac{1}{1+s\dfrac{L_o}{R_{LS}}+s^2\,L_o C_{LS}}$$

The first term describes the desired behaviour, the second term causes the trouble. It can add the extra phase shift that causes instability. The solution to this problem is the insertion of a small coil in series with the loudspeaker. The resulting model then looks like Figure 6.7.

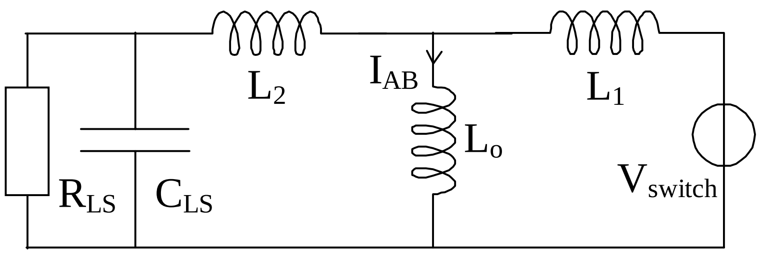

Figure 6.7: Model of the amplifier at f_{sw} with stabilising inductance L_2.

For Y we find:

$$Y = \frac{1}{s\,L_1} \cdot \frac{1+s\dfrac{L_2}{R_{LS}}+s^2\,L_2 C_{LS}}{1+s\dfrac{(L_o+L_2)}{R_{LS}}+s^2(L_o+L_2)\,C_{LS}}$$

By choosing $L_2 >> L_o$, the nominator and the denominator of the second term practically cancel, leaving only the desired first term. The measured virtual output inductance L_o in the two final prototypes varied between 15-150nH, so L_2 can be implemented with only a small air coil. The series inductance in most loudspeakers will have the same stabilising effect as L_2.

The output resistance of AB also determines the amount of switching residue at the output. It is easy to see (Figure 6.7) that when $L_1 >> L_o$, the switching residue relative to the class D half bridge output is L_o/L_1.

6.4 Implementation of a prototype with a low switching frequency

6.4.1 Realisation

Basically, the amplifier can be realised in two ways. The first has a low switching frequency and a limited power bandwidth of D at the cost of some dissipation. The second has a high switching frequency. In this section, the first kind is described, in which AB must be able to provide the full load current for high frequency high power signals.

The dimensioning of the amplifier depends on many factors. Let us first consider the output resistance of AB. Measurements show that up to 100kHz the output resistance of AB is smaller than 0.1Ω, which is still considerably smaller than the load impedance of 4.7Ω. We choose f_{sw}=80kHz. At this frequency, switching losses are virtually non-existent. Together with I_{th} =100mA, it follows from (1) that L_1=0.6mH. Now, the maximum power that D can deliver at 20kHz is $0.075 \times V_S$, or 0.2W. At the clipping point of AB ($0.85 \times V_S$ = 17V), D can deliver all output power up to 930Hz.

If the audio signal has a higher frequency, AB will supply a bigger part of I_{LS}, as was shown in Figure 6.3. This does not affect the output voltage, although some distortion is introduced due to AB's non-zero output impedance. The efficiency is not degraded very much, since the high frequency content of audio signals is limited. To test this, a behavioural model of the amplifier was implemented in a C-program. The program accepts standard audio files, and calculates this extra dissipation. Many different audio signals were tested. In fact, the test set of Chapter 2 was used. The simulations show that the extra dissipation is negligible for almost all signals. Only for one or two fragments at high output levels the extra dissipation is 1…2W (see Figure 2.11).

A prototype of the amplifier was built with integrated subcircuits. For AB, a standard audio power amplifier IC was used. The switches were realised with a DMOS-IC. The supply voltages are +/- 20V. L_1=0.6mH, wound on a ferrite core. This requires fewer turns than an air coil, so it has less conduction losses. The virtual output inductance of AB was measured, and is equal to 150nH. Without the output inductor L_2, the amplifier is unstable if loaded with more than 10nF. L_2=5μH, a small air coil, proved to be sufficient to keep the amplifier stable for any load that was tested, including long cables, loudspeakers, and heavy capacitive loads (5μF). The maximum output power is 30W in a 4.7Ω load.

6.4.2 Measurement results

The power dissipation for sinusoidal signals is plotted in Figure 6.8. The quiescent power dissipation is about 3.5W, of which 2.5W is AB's quiescent dissipation, and 1W is the result of dissipation of the current ripple. At 1kHz, D supplies the full load current. Only at higher output powers there is some dissipation, mostly conduction losses. At frequencies higher than 1kHz, AB has to deliver part of the load current, resulting in higher dissipations at higher power levels. The dissipation of AB only is plotted for comparison. For low frequencies, the dissipation of AB+D is up to 3.5 times lower than that of AB only.

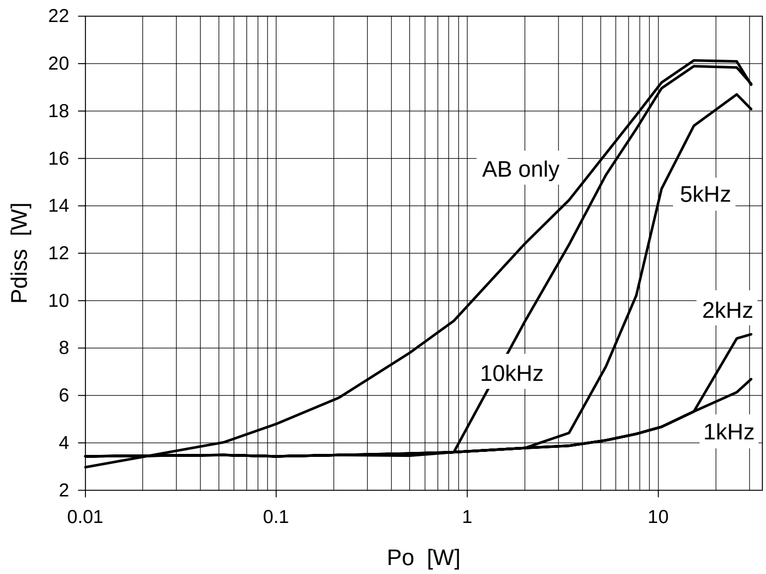

Figure 6.8: Power dissipation AB/D for sinusoidal signals.

The performance of the amplifier for audio signals is shown in Figure 6.9.

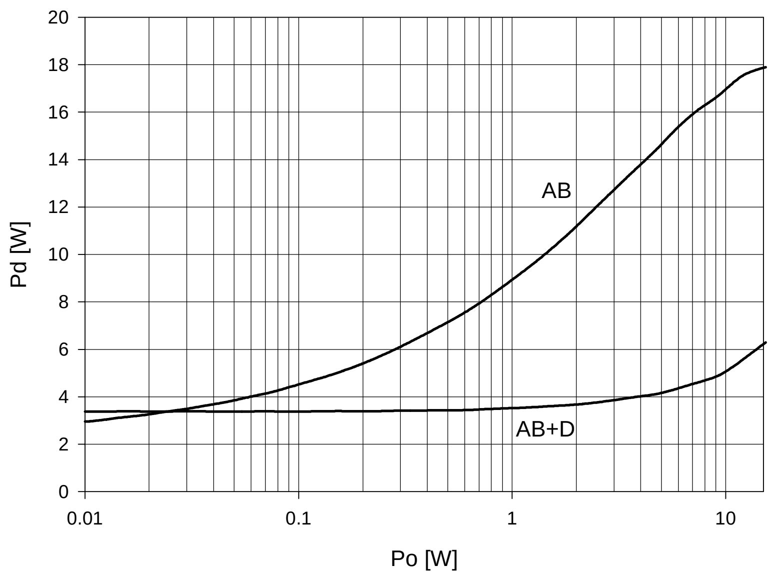

Figure 6.9: Power dissipation for the IEC268 test signal.

The problem with audio is that it is not possible to determine the power dissipation easily. Instead of real audio signals, the IEC-268 test signal was used (see Chapter 2). The signal was used up to an output power of 15W. Above this power, audio signals are severely distorted, clipping for more than 10% of the time. Compared to AB, the power dissipation of AB+D is at least 2.5 times lower at powers between 1 and 15W. At average listening levels (1-3W), 85% of the power dissipation is quiescent power dissipation. Since the majority of this results from the quiescent power dissipation of the amplifier IC, this a point of attention for future designs. The results of Figure 6.9 were verified by amplifying audio signals and measuring the heat sink temperature of the total amplifier compared to the AB amplifier only. The ratios of dissipations that can be calculated from these measurements confirm these results.

At first it might seem a good idea to have both switches $SW_{1,2}$ open for small output powers, since this would eliminate the dissipation of the current ripple needed for self-oscillation. Figure 6.9, however, shows that this is only slightly favourable for output powers below 25mW. This is not worth the extra complexity.

Figure 6.10 and Figure 6.11 show the distortion of the amplifier for different output powers. At 1kHz there is hardly any difference between AB and AB+D. At 10kHz, the distortion of AB is higher than at 1kHz. For the AB+D this is the same, except at low output powers, where the distortion is lower than that of AB. This can be explained with the help of Figure 6.8: At 10kHz, we see that D supplies the load current up to 1W. Above 1W, AB supplies part of the load current, leading to a higher dissipation. AB's distortion increases with increasing output current, so in Figure 6.11 we see a rising distortion above 1W.

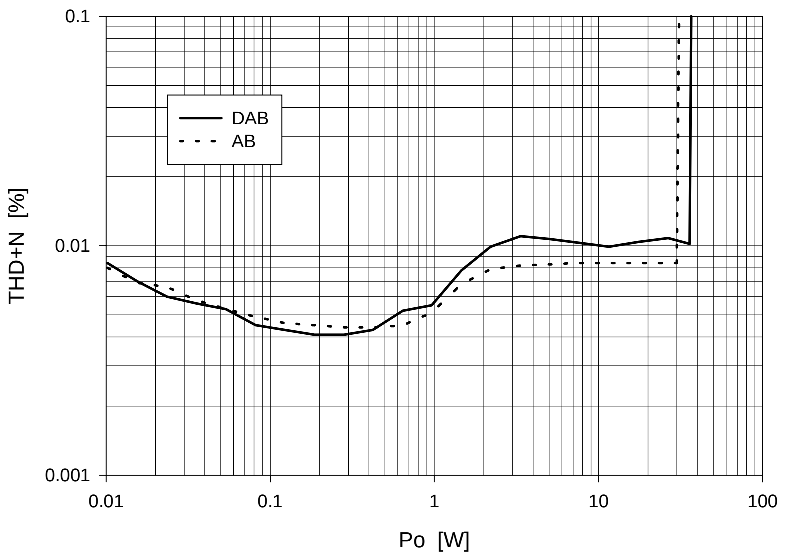

Figure 6.10: THD+N vs Pout @ 1kHz, filter 22.4Hz-22.4kHz.

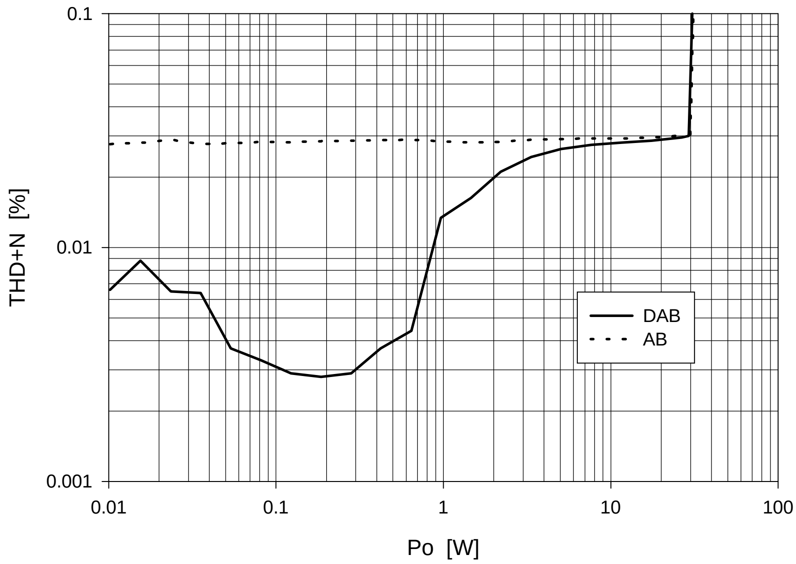

Figure 6.11: THD+N vs Pout @ 10kHz, filter 22.4Hz-22.4kHz.

The advantage of this amplifier over class D amplifiers is the fact that AB almost directly controls the output voltage. It is expected, therefore, that the distortion characteristics do not change when connecting a loudspeaker. The impedance of the used loudspeaker is shown in Figure 3.18. Figure 6.12 and Figure 6.13 show the distortion as a function of frequency, with a resistor load and a loudspeaker load respectively. At frequencies below 1kHz, the distortion of AB+D is higher. This is mainly due to L_2. When AB was also tested with L_2 present, the values were also higher. Above 1kHz, the combination of AB+D has a lower distortion, caused by the same phenomenon as discussed above. The apparent decrease in distortion above 10kHz is a result of the filter in the distortion analyser.

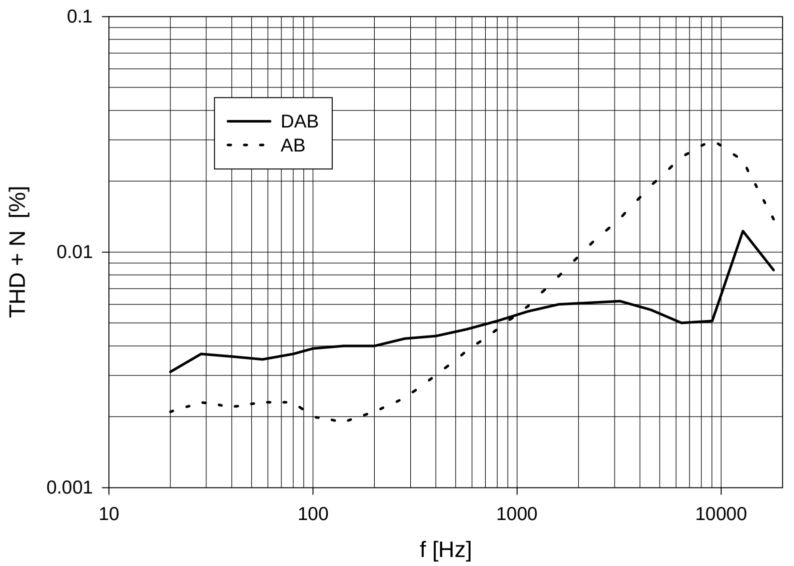

Figure 6.12: THD+N vs f @ P_o=1W, resistor load, filter 22.4Hz-22.4kHz.

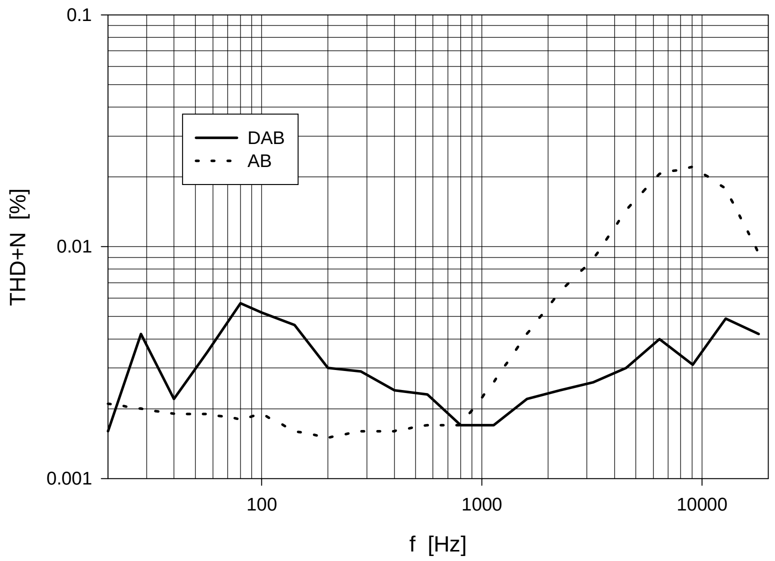

Figure 6.13: THD+N vs f @ P_o=1W, loudspeaker load, filter 22.4Hz-22.4kHz.

The frequency transfer function of the amplifier is flat +/- 0.1dB up to 10kHz, and +/- 0.3dB up to 20kHz, also with the loudspeaker as load. The phase varies between 2° and -9°, which is mainly due to, again, L_2. The main specifications of the amplifier are summarised in Table 4.

Parameter	Condition	
Maximum output power	R_{LS}=4.7Ω, THD=1%	30W
Frequency response	20Hz-10kHz, R or LS load 20Hz-20kHz, R or LS load	+/- 0.1dB +/- 0.3dB
THD+N (filter 22Hz-22kHz)	20Hz-20kHz, 1W-30W	0.003%-0.03%
Switching frequency	V_o=0V	80kHz
Quiescent power dissipation		3.5W
Power dissipation	IEC-268 signal, P_o=10W	5W

Table 4: Amplifier specifications

6.4.3 Discussion

The class AB+D parallel amplifier features a high efficiency (up to 3.5 times better than class AB), and a low distortion (0.01%@1kHz) that is highly independent of the connected load. The efficiency is a little lower than that of class D amplifiers, but better than that of class G amplifiers or amplifiers with a tracking power supply. The quiescent power of AB is something that needs improvement, since it is the major source of dissipation at normal output levels. An advantage of the new amplifier over class D amplifiers is the better control over the output voltage, especially with varying loads. Furthermore, the filter is simpler than in a class D amplifier, and L_1 needs not be very linear. However, L_1 is larger than in an average class D filter.

A number of the drawbacks of this realisation can be overcome by using a higher switching frequency. In that case, L_1 could be smaller (both electrically and physically). The switching losses increase, but are not expected to be a major source of dissipation below approximately 1MHz. The virtual output inductance of AB has to decrease proportional to the switching frequency. As this puts extra demands on the high frequency behaviour of AB, it is best to use AB only to provide the current ripple. This in possible, provided that D's power bandwidth is larger than the audio bandwidth. Additional advantages of a smaller and faster AB include a smaller chip area and a lower quiescent current.

6.5 Implementation of a prototype with a high switching frequency

6.5.1 Realisation

Allowing AB to provide part of the output current can results in a very low switching frequency. In that case, however, AB must be able to supply the full load current for large amplitude high frequency audio signals. In the design presented here, we want D to provide the full load current for all audio frequencies, so that AB can be small, requiring less chip area.

Equation (3) is a good starting point for the dimensioning of the amplifier. By making I_{thr}=100mA, the quiescent power dissipation is raised by 0.9W, which is approximately half of the total quiescent power. Other practical values for a realisation are: α = 0.85 (α denotes the clip voltage of AB as fraction of the power supply), V_S = 18V and R_{LS} = 4Ω. This results in a 30W amplifier. According to equation (2), for a power bandwidth of 20kHz, L_1 should be 20µH. With equation (1), it can be calculated that the resulting free running switching frequency at V_o=0 would be 2.25 MHz. Since this is expected to lead to unacceptable switching losses in the power switches, L_1 is chosen 80µH. With this value, the amplifier can deliver a full power sinusoid up to 5kHz, decreasing to 20% of full power at 20kHz.

A power bandwidth of 5kHz makes the amplifier a suitable candidate for Transient InterModulation distortion (TIM). However, since audio signals have a limited high frequency contents, a power bandwidth of 5kHz should be enough to avoid this type of distortion [7]. A behaviour model of the amplifier was implemented in a C-program. The program accepts an audio file as input and generates the simulated audio output. By comparing the input and output file, it was confirmed that the amplifier rarely showed any slew rate distortion.

The modular structure of the amplifier is reflected in the experimental realisation. The linear part is at present still external, and built with a commercially available power OPAMP. It must source and sink 100mA. The switching part of the amplifier was realised in two modules in a BCD process (a process that allows Bipolar, CMOS and DMOS devices on the same chip). Figure 6.14 shows the circuit diagram of the output current sensing circuit. The output current of AB is measured by sensing its supply lines. I_1 is a bias current source of 200μA. The measuring resistors are 0.1Ω, and also integrated. Two scaled copies of the output current are made, each of which receives a opposite offset by means of I_2 (5μA). The values of I_1 and I_2 are not very critical. I_1 influences I_{thr} through the g_m of the sensing circuit, and I_{thr} is proportional to I_2.

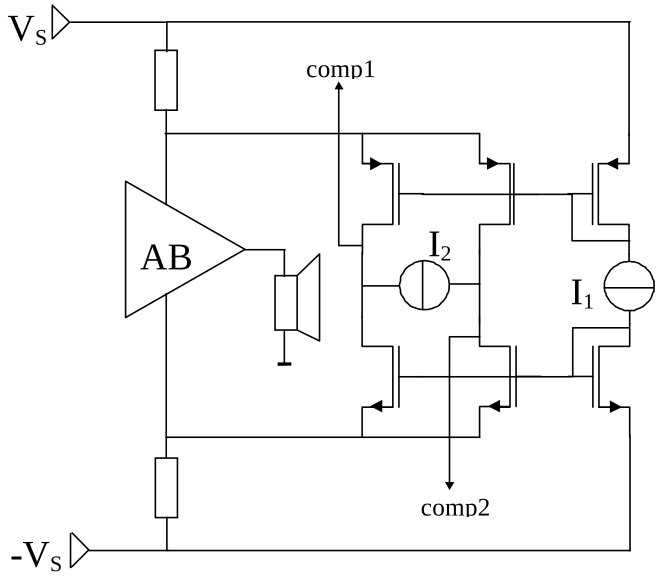

Figure 6.14: Output current sensing circuit.

A more important issue is mismatch. Mismatch between the mirror ratios gives rise to an offset in I_{thr}, increasing the dissipation. Mismatch between transistors within the mirrors add up to a deviation in $+I_{thr}$ and $-I_{thr}$, changing dissipation and switching frequency. With the process and transistors used, deviations smaller than 3σ result in a switching frequency between 300kHz and 1MHz and an extra dissipation of less than 1.2W.

In Figure 6.15, the copies of the output current are connected to the comparators comp1 and comp2. Regenerative comparators offer the lowest power-delay products [64], but since no external clock signal is available, a multistage amplifier design is chosen. The amplifiers are inverters with a feedback resistor. 6 stages offer a total gain of more than 10,000. An SR flip-flop combines the comparator outputs such that they behave like one comparator with hysteresis. A chip photo of this part is shown in Figure 6.16. This comparator module is followed by the power switches, implemented as DMOS transistors. A bootstrap capacitor provides the upper gate voltage, and a control circuit avoids common conduction of the two transistors by means of a handshake procedure.

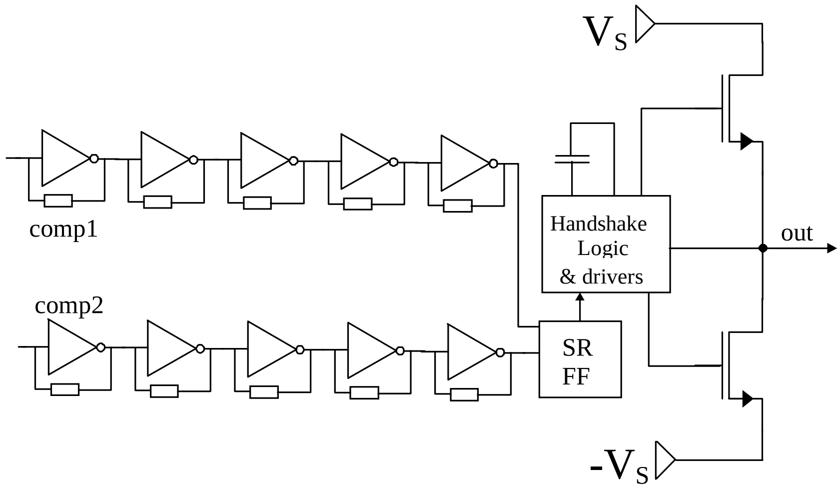

Figure 6.15: Comparators and power switches.

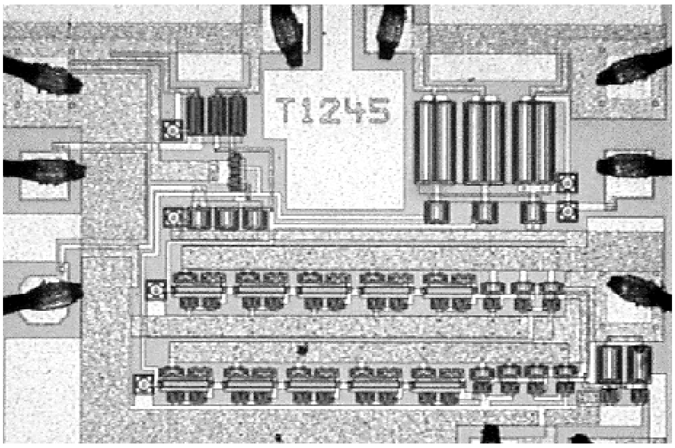

Figure 6.16: Chip photo of sensing and comparator section.

6.5.2 Measurements

A 30W version of the amplifier was measured. The maximum efficiency for sinusoidal signals is 85%. This is slightly lower than in a normal class D amplifier (90% is possible [28]) due to the extra dissipation of AB (quiescent power and current ripple dissipation). A better choice would be to measure with audio signals. Audio signals, however, are very inconvenient test signals, therefore the IEC 268 test signal was used. This signal has a gaussian amplitude-probability-density function, and a frequency distribution that is average for normal audio material. It gives a good prediction of audio amplifier dissipation in practical situations [19]. The signal was used up to an output power of 15W RMS. Above this power, audio signals (and consequently the IEC test signal) get severely distorted, clipping for more than 10% of the time. This is caused by the much larger crest factor of these signals compared to a sinewave. The dissipation of the amplifier for the IEC 268 test signal is shown in Figure 6.17. To give an indication of the efficiency improvement over class AB amplifiers, the dissipation of a standard (arbitrary) class AB amplifier is also displayed. Note that for the new amplifier, most of the dissipation at normal listening levels is due to the quiescent power dissipation; this is a point of attention for future designs. In principle, the dissipation is independent of the frequency of the input signal. This is a result of our choice to use D to provide all output current. In practice, when the amplifier is used above D's power bandwidth, AB supplies some current, but since AB is current limited at a small value, it will not cause much extra dissipation. As for the load: it hardly influences the frequency transfer of the amplifier, shown in Figure 6.18.

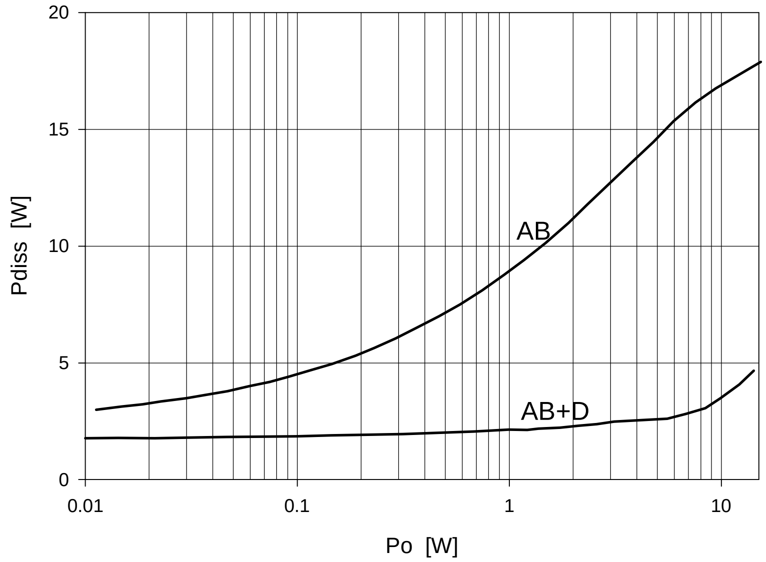

Figure 6.17: Power dissipation for IEC-268.

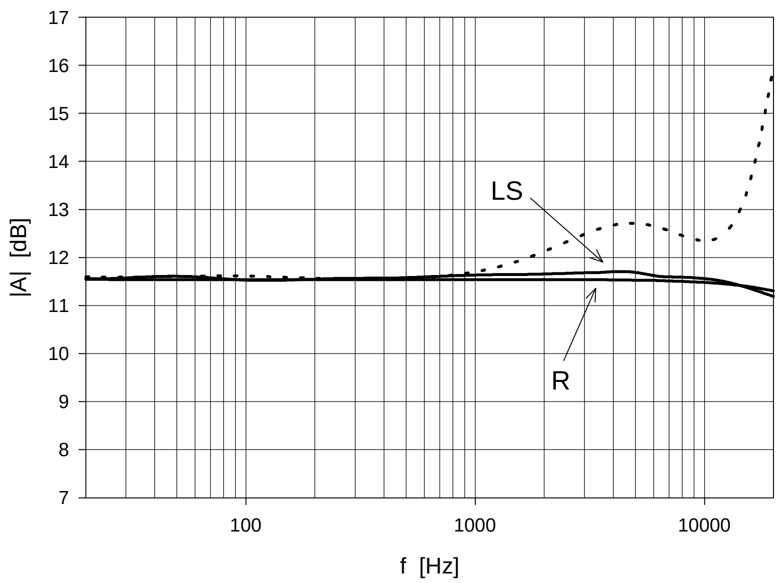

Figure 6.18: Frequency transfer with resistor and loudspeaker. Dotted line is the simulated class D frequency transfer with loudspeaker (Figure 3.18).

Figure 6.18 displays the frequency transfer for both a resistor and a loudspeaker connected to the output. The decrease at high audio frequencies is the result of the output inductor, but it is still a major improvement over the frequency transfer of a normal class D amplifier (Figure 3.18, displayed as a dotted curve in Figure 6.18).

Figure 6.19 shows that the distortion with a resistor load at 1kHz remains less than 0.02% up to the clipping point at 30W. The distortion at other frequencies is shown in Figure 6.20.

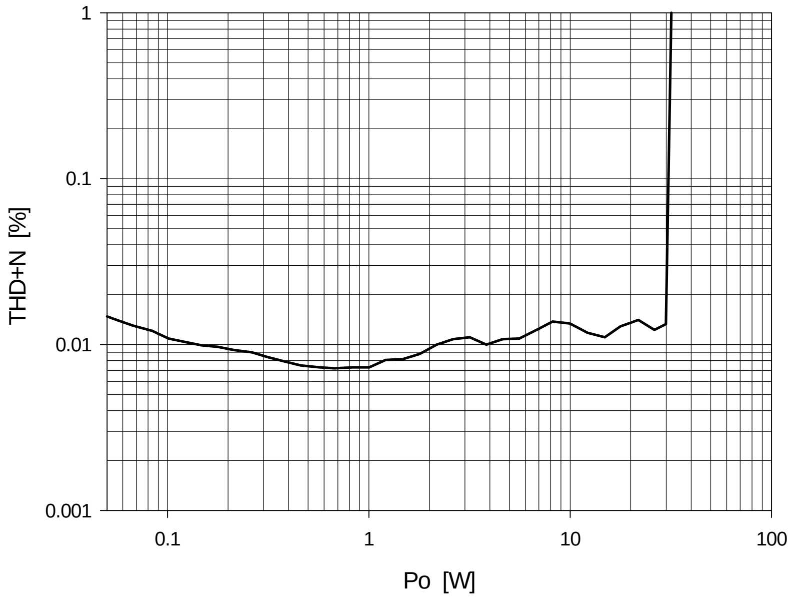

Figure 6.19: THD+N @ 1kHz, filter 22Hz-22kHz.

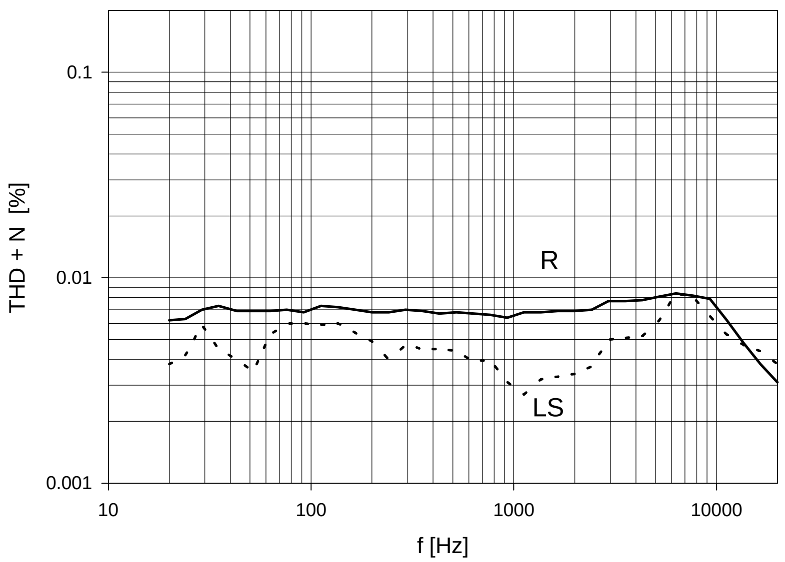

Figure 6.20: THD+N @ 1W, filter 22Hz-22kHz.

As expected, the values do not increase with a loudspeaker as load. The apparent decrease at higher frequencies is a result of the filter of the distortion analyser. The distortion with resistor load is virtually frequency independent, although it is expected to increase with frequency due to roll off of the loop gain. Probably, noise caused by switching of the output current is influencing the linear part, because the distortion is lower at frequencies where the loudspeaker impedance is higher.

The measured output inductance L_o of AB is 15nH. The switching residue relative to the class D half bridge output is L_o/L_1, which is in our case -74dB. The measured residual switching noise at the output is 3mV, which is equal to -77dB. In a standard class D amplifier, a 4th order output filter would be needed to achieve this value. Table 5 summarises the main specifications of the amplifier.

Parameter	Condition	
Switching frequency	V_O=0V	550kHz
Max. output power	$R_{LS}=4\Omega$,THD=0.1%	30W
Maximum efficiency	P_{out}=30W	85%
Quiescent power		1.8W
Power dissipation	P_o=10W, IEC-268 test signal	3.5W
THD+N (filter 22Hz-22kHz)	20Hz-20kHz, 1W-30W	0.003-0.02%
Frequency response	20Hz-20kHz, R or LS load	+/- 0.3dB
f_{switch} attenuation	f=500kHz, $R_{LS}=4\Omega$	77dB

Table 5: Amplifier specifications

6.5.3 Discussion

It is interesting to compare this realisation with the previous one. We see that the main specifications concerning frequency response and distortion have stayed approximately the same. The dissipation at 10W output power has decreased with 30% and the quiescent dissipation with 50%. L_1 has an 8 times smaller volume, and AB's maximum power is much lower. This clearly demonstrates the benefits of a higher switching frequency.

A comparison with other existing amplifiers is not straightforward because it is only fair to compare amplifiers in practical operating conditions with audio signals (or the IEC test signal) and loudspeaker loads. In most articles, however, amplifiers are measured with sinewaves and resistive loads. Compared to class G amplifiers and class AB amplifiers with a tracking power supply, the amplifier introduced in this paper has comparable frequency transfer and distortion figures, but a much lower dissipation under realistic test conditions [22, 21, 49, 47]. Compared to class D amplifiers, the amplifier has a lower distortion and a better frequency response for loudspeaker loads. The output filter is simpler, but the dissipation is a little higher [28, 31, 34]. The main drawback is the high quiescent power dissipation. Methods to reduce it are discussed in the next section.

Another objective is to design the linear amplifier in the same process, and merge the different modules on one chip. This is useful for the high frequency behaviour. The output inductance of AB in this realisation is 15nH. It is quite a challenge to make this value dominate the breadboard parasitics. In a one-chip system, this would be a lot easier. Other problems like distortion, however, are likely to get more prominent. Also, the design of AB is in no way a trivial matter. A class A amplifier is easier to design, but its quiescent power dissipation would be too high. Again, the next section will discuss this further.

6.6 Future possibilities

The fast-switching prototype of section 6.5 has two drawbacks: the high switching frequency when D has a high bandwidth, and the relatively high quiescent power. The variations in this section are topologies that can eliminate these drawbacks. The amplifiers have not been realised, but have successfully been tested in a simulator that uses a behavioural model. The class AB amplifiers are voltage amplifiers with 1 pole, a high DC-gain and 10MHz unity gain frequency. The 1Ω open loop output resistance is reduced to 0.1Ω at 1MHz thanks to unity gain feedback. Switches are simulated as resistors that vary between 0.5Ω and 100kΩ. MOSFETs are controlled current sources, V_S=18V, α=0.85, R_{LS}=4Ω. The results of these simulations look promising. Obviously, more detailed simulations and practical realisations may reveal other aspects of these circuits. Any way they are a good candidates for future research.

6.6.1 A higher order coupling network

The inductor L_1 in the previous realisation determined the relation between the switching frequency and the power bandwidth of D. These two variables become independent to some extend by choosing a coupling network between AB and D that has a higher order than one. Figure 6.21 shows a circuit with a 3^rd^ order coupling network. A 3^rd^ order network has a current source character again.

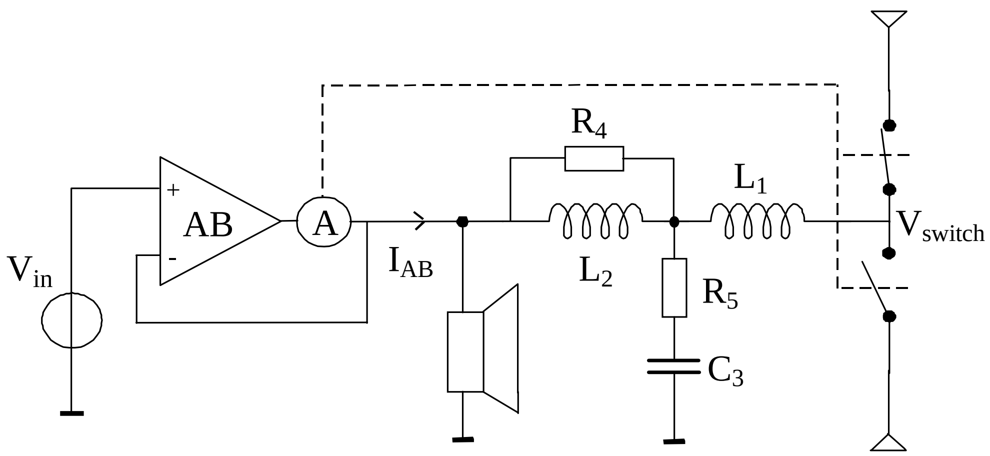

Figure 6.21: A parallel AB/D combination with a 3rd order coupling network.

The resistors R_4 and R_5 introduce zeros in the transfer from the half bridge output to AB's output current, necessary for the first order frequency response of the coupling network at the switching frequency. When we suppose AB's output is grounded, the admittance $Y=I_{AB}/V_{switch}$ for the practical values mentioned below is plotted in Figure 6.22. For low frequencies, the response of the network approaches L_1+L_2. For high frequencies, the response approaches $L_1(1+R_4/R_5)$. The dotted line in Figure 6.22 illustrates the benefit of this approach. Assuming $L_1=L_2$, the maximum switching frequency is reduced by a factor $\frac{1}{2}(1+R_4/R_5)$ compared to a single inductor of value (L_1+L_2). A simulation with $L_1=L_2=10\mu H$, $C_3=5.6\mu F$, $R_4=3\Omega$ and $R_5=0.6\Omega$, confirms that the maximum switching frequency is now 750kHz, a factor 3 lower than in the case of a single 20μH inductor, while the power bandwidth of D is still 20kHz.

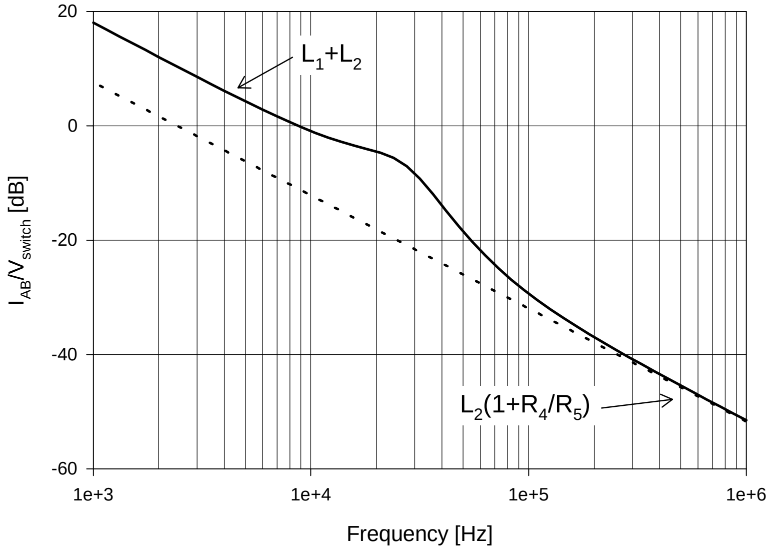

Figure 6.22: Frequency response of I_{AB}/V_{switch} for the coupling network in Figure 6.21.

Further simulations show that there are several limitations to the topology. As described in section 6.3.3, an extra output inductor is needed to stabilise the oscillating loop in case a capacitive load is connected. Also, the delay in the comparator and switches must not exceed 100ns to avoid instability. Note that with a first order coupling network this limitation does not exist. The reduction of switching frequency is limited, because of the phase shift just below the LC resonance frequency. In the example above, it is 145º, leaving a phase margin of 35º when the oscillating frequency is in that region. Although a clocked system may prevent this, any input signal in that frequency range can also cause ill-defined waveforms. Finally, a 20kHz full power sinewave causes too much dissipation in the resistors. With the IEC268 test signal, it is a few Watt. All in all, a practical implementation should not be considered unless a low switching frequency is very important.

6.6.2 A clocked version

It might be undesirable that the switching frequency of the class D part is variable. The frequency spectrum of the pulsed output has energy in a wide frequency band. A constant switching frequency would facilitate e.g. interference measurements owing to the smaller frequency band. It is important to see that with respect to the switching frequency, the self-oscillating approach realised above is just an option; the main goal is to realise a reduction of the output current of AB. Figure 6.23 shows how a system with a fixed switching frequency would look. The inverting input of the comparator acts as the input of a class D amplifier with an analogue modulator discussed in section 3.3.5. Feedback is accomplished by L_1 which acts as an integrator. AB determines the loudspeaker current I_{LS}. D is controlled by $I_{AB}=I_{L1}-I_{LS}$. Thus, the system can be viewed as a feedback loop in which I_{L1} follows I_{LS}. In a practical system, the triangle can be injected as a current in the output.

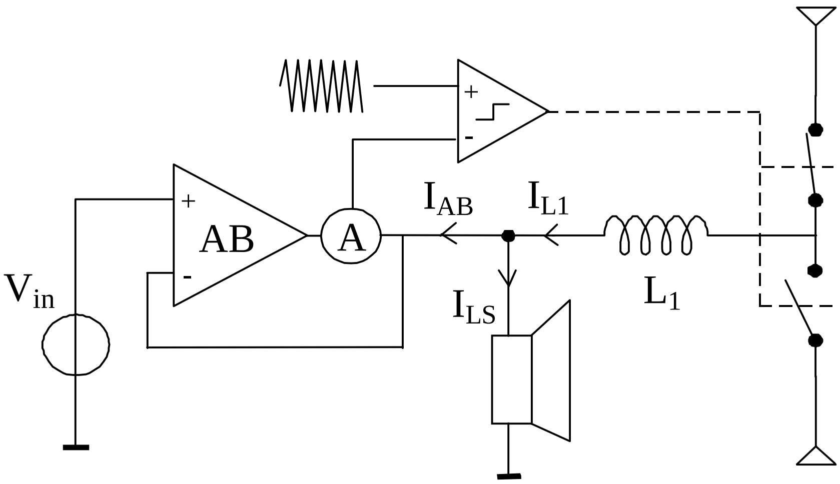

Figure 6.23: A clocked parallel AB/D amplifier.

As in any feedback loop, an input signal (I_{AB}) inversely proportional to the forward gain exists. In this case it means that I_{AB} consists of both a ripple current and a small version of the audio current I_{LS}. When the amplifier is designed for the same quiescent current as the self-oscillating amplifier, I_{AB} must oscillate between +100mA and -100mA in quiescence, which requires a triangle between -200mA and +200mA. This means that at maximum output voltage, I_{AB} has a very small ripple and is virtually 200mA. This might seem to degrade the efficiency, but all current contributes to I_{LS} and is delivered by the upper output transistor of AB, causing no extra dissipation. Simulations were done with L_1=20µH and a triangle of 2.25MHz and 400mA_{pp}. Other than the differences discussed above, the circuit behaves identical to its self-oscillating counterpart.

6.6.3 A parallel amplifier in bridge

The parallel AB/D amplifier in bridge configuration offers a lower quiescent power dissipation when the common mode level of the bridge is chosen close to ground. Figure 6.24 shows the topology, and Figure 6.25 the output voltage of the two bridge halves when the output signal is sinusoidal. One side of the bridge amplifies the positive half of the input signal, the other side the negative half. The quiescent common mode level is e.g. 1 volt. This gives the possibility to realise the linear amplifiers as class A amplifiers instead of class AB amplifiers (the MOSFETs in Figure 6.24). It is much easier to design a fast class A amplifier than a fast class AB amplifier.

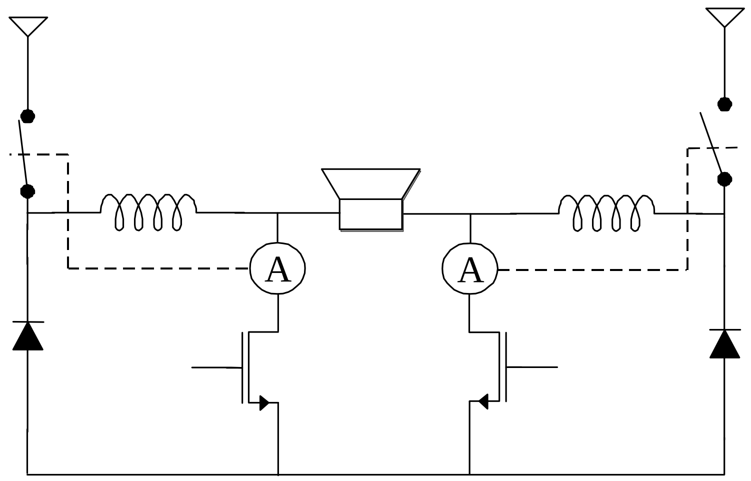

Figure 6.24: An AB/D bridge amplifier with a low common mode level.

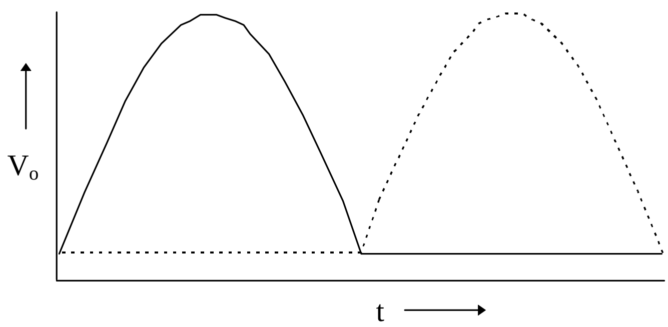

Figure 6.25: Output of the two bridge halves when the output (difference) is a sinewave.

A consequence of choosing a class A amplifier is that the current ripple should not become negative. In quiescence this is not important thanks to the low quiescent common mode level. An average ripple current of 200mA contributes only 0.2W. At non-zero output voltages, however, the DC component in the current ripple increases the dissipation.

A second advantage of the low common level is that in a self-oscillating system, the switching frequency is very low at V_o=0, and the maximum switching frequency is reduced by a factor 2. A clocked system may be necessary, though, since a self-oscillating topology can create audible difference signals when no input signal is present. As external components 2 coils are needed, of which it is uncertain if they can share the same core. The generation of the bridge signals could use differential feedback across the load and a feedforward common mode level control circuit, comparable to the TDA1560 class H amplifier (see Chapter 3).

The amplifier was simulated with L_1=L_2=20µH, a self-oscillating switching part with current thresholds of 100mA and 300mA, and a quiescent common mode level of 1V. In that case, the current ripple adds 0.4W to the quiescent power dissipation. The main advantages, however, will show up in practice: the use of class A amplifiers and the reduced switching frequency.

6.6.4 A balanced current output stage

In a balanced current output stage [37], the output current is provided by two coils that are both connected to a switch/diode pair. See Figure 6.26. The control signals for the switches are generated by comparing the input to a triangle. In quiescence, V_{D1} is the inverse of V_{D2}. Thus, the ripple current of the left coil flows into the right coil, leaving AB with no ripple current. When the output voltage is positive, the duty cycle of V_{D1} increases and the duty cycle of V_{D2} decreases. AB has to supply the ripple, which is proportional to the duty cycle. Thus, dissipation of the current ripple is related to the output power and zero in quiescence.

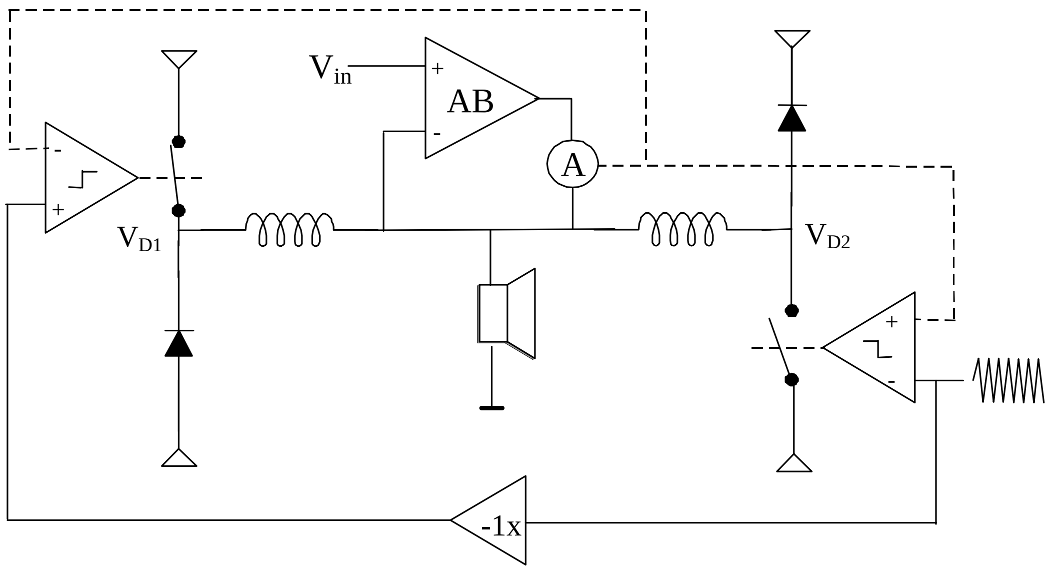

Figure 6.26: A class AB/D combination with a balanced current class D part.

Again, the usage of two coils is a disadvantage, as both are only used half of the time. With respect to the quiescent power dissipation, however, this topology remains highly attractive. Simulations with both L's 20μH, and a $400mA_{pp}$ triangle did not contradict this judgement.

6.7 Conclusions

It is possible to use a linear amplifier to do most of the filtering of a class D amplifier with only a little extra power dissipation. This way, the external filter has less components, and does not have to be very linear. Furthermore, the transfer of the amplifier is less dependent on the connected load. The main disadvantage is a quiescent power dissipation that is significantly higher than in class D amplifiers. The trend towards higher switching frequencies and the further development of new topologies contribute to elimination of this problem. Thus, parallel AB/D systems are a new step towards highly integrated power efficient audio amplifiers.

7

Conclusions

7.1 Introduction

In the previous chapters, test signals have been defined, an overview of high efficiency audio amplifiers has been given, and some new amplifier topologies have been introduced. Especially in a practical research area like this, it is important to know how this work relates to other work in the field, both in industry and at universities. The next section deals with such a comparison. In the last section, the tendencies are extrapolated, and some educated guesses are made concerning the future developments of high efficiency audio amplifiers.

7.2 This work in relation to other work

This thesis is concerned with 3 main subjects: an overview and analysis of existing amplifiers and their efficiency (chapter 1 and 3), test signals for measuring and predicting amplifier efficiency (chapter 2), and the analysis and design of new amplifiers (chapter 4, 5, 6). It is useful to structure a comparison with other work during the last four years along these lines.

Description of amplifier principles

We can be brief about the overview of high efficiency amplifiers given in chapter 3. It can serve as a reference for other people working in this field. Unfortunately, books about audio amplifiers are very rare, so chapter 3 is a good beginning.

Test signals

Concerning test signals, the situation at the beginning of this research (1994) was disastrous. Not dissipation was measured, but efficiency, always with sinewaves. Luckily, there has been some change in this area. Nowadays, the datasheets of

Philips' TDA1560 [22] and TDA 1561 [25] and of ST's TDA7454 [21] contain dissipation information for gaussian output signals. Texas Instruments offers a freely available computer program on their website that calculates the dissipation of their amplifier IC's for any audio signal. The power rating of loudspeakers was investigated with the IEC-268 test signal [16]. These are promising signs that things are changing. They have to change more however, as there are still manufacturers who specify the maximum efficiency of their class D IC's for sinewaves at 20% THD…

New amplifier topologies

The design of high efficiency amplifiers has concentrated on new concepts. The three linear amplifiers mentioned above have all become available the last few years. The ideas are often a little older, since the design of these amplifiers is quite complicated. At present, they offer approximately half the dissipation of a class AB amplifier, combined with a very limited number of external components and no EMI problems. For very low dissipations, switching amplifiers are inevitable.

Publications on standard class D amplifiers and the generation of high quality PWM signals used to be common, but only the past year fully integrated versions have become available that include modulator, driving circuitry and power switches (TI: TPA005D02, ST: TDA7480, Philips: TDA8920). At the same time, at the research level, efforts have been made to circumvent the drawbacks of class D amplifiers. The generation of high-quality PWM power signals gained attention in [35], [38] and [39]. Problems associated with the output filter are eliminated in [28], where the filter is incorporated in a new feedback loop. All in all, this has made class D amplifiers much more competitive than they used to be. The main problems are currently EMI and the external filter. The class AB/D combination presented in section 6.5 performs better on these points, but suffers from a higher quiescent dissipation and higher speed requirements. Therefore, it is still not clear which topology will be most successful, but it is clear that both concepts are quickly overcoming their limitations.

Other work in the field of class AB/D amplifiers, has showed less progress. The use of a class D amplifier for the subwoofer of a loudspeaker system [50] is only practical in active loudspeakers. Other published amplifiers are mainly variations on amplifiers with a variable supply voltage, including the chip area disadvantage and a relatively high dissipation. Very recently, a circuit similar to the one in chapter 6 has been presented [46]. This confirms our idea that this concept is a promising choice that deserves further exploration.

7.3 Future developments

A few years from now, a section like this will probably generate some good laughs, as scientific prophecies have often proven to be highly inaccurate. Still, one has to try.

Going digital

As mentioned in the beginning of this thesis, the audio amplifier is one of the few remaining analogue components in the audio chain. At this time, the input of audio amplifiers is analogue. A digital input would simplify circuit design, as it eliminates the need for a separate D/A converter. Especially the combination with class D amplifiers is advantageous, because currently, the digital audio signal is converted to analogue by an D/A converter, en then back to digital by the class D modulator. A DSD bitstream can, after decimation, serve as a class D signal. For the parallel class AB+D amplifier, the situation is a little more complicated. The strong point of this amplifier is the small output ripple. To keep this advantage for a digital input signal, strong filtering is unavoidable. Big advantage compared to class D is that this can happen in the small-signal domain.

Measurements

The developments in the measurement field probably depend on the acceptance of high efficiency amplifiers and the use of class G concepts. When those amplifiers are used often, the use of gaussian (IEC-268) signals is likely to become a standard. Class D amplifiers are usually specified by their maximum efficiency. This is not a real problem, as long as the quiescent dissipation is also specified. Therefore this way of measuring is not expected to change, even though a graph displaying dissipation as a function of output power for the IEC signal is much more desirable.

Speed

Because of the shrinking dimensions of IC-processes, the channel length of transistors decreases, and the amount of g_m per parasitic capacitance increases. This is also favourable for power applications because it enables faster switching. Faster switching diminishes switching losses, creating possibilities for very high frequency class D amplifiers. For a moderate increase in speed, both class D and parallel class AB/D amplifiers benefit. In class D amplifiers, a simpler output filter will suffice to reach sufficient ripple rejection. In class AB+D the increased speed can be used to either increase the ripple rejection or decrease the current ripple dissipation and AB’s quiescence current.

For very high frequencies, the advantage for class D amplifiers is limited, because the capacitive losses will increase with frequency as the output transistors are not getting smaller. Tricks like using only part of the output transistors in quiescence will be necessary to profit from the higher speed. For class AB+D, the situation is better, because its quiescence dissipation is dominated by AB’s quiescence current and the current ripple dissipation. An increase in speed can be used to diminish these components to the same order of magnitude as the capacitive losses in the class D part of this amplifier. Thus, it becomes more competitive to standard class D amplifiers.

For linear amplifiers, the increased speed will be of limited importance. Possible advantage is that it will be easier to avoid switching distortion.

Finally

During the last four years, there has been much progress in the design of high efficiency amplifiers. When this is an indication for the future, nothing stands in the way of high efficiency audio amplifiers that are a true alternative to class AB amplifiers in many applications.

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Summary

(Chapter 1) Audio amplifiers need to supply ever more output power thanks to the increasing dynamic range of digital media and the movement towards multichannel sound systems. As a result, the dissipation rises, whereas physical dimensions are getting smaller. This leads to heat problems, creating a need for high efficiency amplifiers. In portable systems, a higher efficiency is favourable to lengthen battery life.

In existing amplifiers, class AB amplifiers are used almost exclusively. Although the theoretical efficiency of class AB amplifiers can be high for a rail-to-rail sinewave, they suffer from a very low efficiency for audio signals, thanks to the large crest factor of these signals.

(Chapter 2) Still, in most literature the efficiency of audio amplifiers is measured with sinewaves. For several amplifier types this does not give a good indication for the dissipation in real-life situations with audio signals. Measuring with audio signals, however, is slow and complicated. A test signal representative for audio is essential to compare amplifier topologies.

To make such a test signal, the frequency distribution and amplitude distribution of many audio samples were measured. Of the existing test signals, the IEC-268 test signal represents audio. Simulations show that the dissipation caused by audio fragments scaled to equal power has only a small spread, and that the average is equal to the dissipation caused by the IEC signal. Thus, this signal is very suitable for measuring and predicting the dissipation of amplifiers for real audio signals.

A problem with the IEC signal is that it can not easily be generated in circuit simulators and in hardware. Two alternative test signals are proposed. The first one, a very simple periodic function, has an amplitude distribution equal to that of the IEC signal, and is suitable for class G amplifiers. The second one is an IEC variant that has the same properties as the IEC signal, but it is much easier to generate in simulators and hardware.

(Chapter 3) With a useful test signal it is interesting to look at known amplifiers with a high efficiency and their limitations. Basically there are two types of amplifiers. The first type are called linear amplifiers (class G, class H, etc.). They operate similar to class AB amplifiers by creating a voltage drop across their output transistors to generate the right output voltage. The different classes of linear amplifiers can be generalised to switching techniques between several supply voltages such that the average voltage drop across the output transistors decreases. The main drawback of linear amplifiers is that for very low dissipation extremely complex switching schemes are necessary, while the maximum dissipation hardly decreases.

The second type of high efficiency amplifiers are the switching amplifiers. The power part of a switching amplifier consists of switches, so ideally there is no dis-

sipation. The high frequency pulse width modulated signal produced by the switches is filtered by a lossless LC filter to reconstruct the audio signal. The main limitations of switching amplifiers are the need for an external filter, the residual switching ripple, and the influence of the load on distortion and frequency transfer. Due to the high order output filter, high feedback factors are not possible.

(Chapter 4) The combination of switching and linear amplifiers has only been realised for amplifiers in which a switching amplifier generates the supply voltage for a linear amplifier. This combines the good control over the loudspeaker with a higher efficiency. It can not reach a very high efficiency, though, and needs large output transistors because the two amplifiers are in series.

(Chapter 5) A ‘serial’ topology that does not need larger output transistors has a switching amplifier and a linear amplifier in bridge. A common mode control circuit ensures that the output voltage of the linear bridge halve is close to one of the supply lines, reducing the average voltage drop across the output transistors. Design choices involve the common mode circuitry and distortion compensation. A prototype of the amplifier demonstrates the validity of this idea. The efficiency is not optimal, but there are good possibilities for improvement.

(Chapter 6) A ‘parallel’ topology is introduced that has a switching amplifier with a current source character in parallel with a linear amplifier. The linear amplifier controls the output voltage, ensuring a low distortion and a flat frequency transfer. Most load current is supplied by the switching part, ensuring a low dissipation. Several aspects like power bandwidth, quiescent power dissipation, and stability must be considered to set up a properly working system. A prototype of this amplifier shows very promising results like a low dissipation, a low external component count, low distortion, and only little switching ripple. Main drawback is the quiescent dissipation; several improved topologies offer potential solutions for this.

(Chapter 7) The last years show a renewed interest in high efficiency amplifiers, and the prototypes presented in this work fit well in. They offer competitive specifications compared to other topologies that have were developed elsewhere during the time of this research. With the introduction of ever faster processes and a further integration of power circuits, it is expected that high efficiency amplifiers can become a true alternative to class AB amplifiers in many applications.

Samenvatting

(Hoofdstuk 1) Audio versterkers moeten steeds meer vermogen leveren dankzij het toenemend dynamisch bereik van digitale geluidsdragers en het toenemende aantal kanalen per versterker. Het gevolg is dat de dissipatie stijgt, terwijl de fysieke afmetingen van veel apparaten juist afneemt. Hierdoor ontstaan warmteproblemen. Audio versterkers met een hoog rendement zijn daarom noodzakelijk. In draagbare apparatuur is een hoog rendement wenselijk om de levensduur van de batterijen te verlengen.

In bestaande versterkers worden bijna uitsluitend klasse AB versterkers gebruikt. Het theoretische rendement van een klasse AB versterker is weliswaar hoog voor een sinus op vol vermogen, maar voor audiosignalen is het een stuk lager dankzij de grote crest factor van deze signalen.

(Hoofdstuk 2) Toch wordt in de meeste literatuur die over dit onderwerp beschikbaar is het rendement gemeten met behulp van sinussignalen. Voor verscheidene versterkertypen geven deze metingen geen goede indicatie van de dissipatie voor audiosignalen in normale gebruiksomstandigheden. Metingen met behulp van audiosignalen zijn echter traag en gecompliceerd. Het is belangrijk om een testsignaal te hebben dat representatief is voor audio, zodat verschillende versterker topologieën makkelijk vergeleken kunnen worden.

Om zo'n testsignaal te maken, zijn van veel audiofragmenten de amplitude- en frequentieverdeling gemeten. Van de bestaande testsignalen, is het IEC-268 testsignaal representatief voor audio. Door middel van simulaties is vastgesteld dat de dissipatie die veroorzaakt wordt door verschillende geluidsfragmenten van hetzelfde vermogen slechts een kleine spreiding heeft, en dat het gemiddelde gelijk is aan de dissipatie die door het IEC testsignaal veroorzaakt wordt. Daarom is dit signaal goed geschikt om de dissipatie van versterkers in realistische omstandigheden te meten en te voorspellen.

Een probleem van het IEC signaal is dat het moeilijk te genereren is in circuit simulatoren en in hardware. Hiervoor worden twee alternatieve testsignalen geïntroduceerd. Als eerste een simpele periodieke functie die dezelfde amplitudeverdeling heeft als het IEC signaal, en dus goed bruikbaar is in klasse G type versterkers. Als tweede een variant op het IEC signaal die dezelfde eigenschappen heeft als het IEC signaal, maar veel gemakkelijker te genereren is in simulatoren en in hardware.

(Hoofdstuk 3) Met een zinvol testsignaal is het nuttig bestaande versterkers met een hoog rendement en hun beperkingen onderzoeken. Er zijn grofweg twee soorten versterkers. Het eerste type zijn de lineaire versterkers (klasse G, klasse H, etc.). Ze maken hun uitgangsspanning op dezelfde manier als een klasse AB versterker: door een spanning over hun uitgangstransistoren te laten vallen. De verschillende klassen lineaire versterkers kunnen gegeneraliseerd worden tot slimme schakeltechnieken die ervoor zorgen dat de gemiddelde spanningsval over de uit-

gangstransistoren afneemt. Het belangrijkste nadeel van lineaire versterkers is dat voor een zeer lage gemiddelde dissipatie zeer complexe schakelconfiguraties nodig zijn, terwijl de maximum dissipatie nauwelijks minder wordt.

Het tweede type hoog rendementsversterkers zijn de schakelende versterkers. Het vermogensgedeelte van een schakelende versterker bestaat uit schakelaars die theoretisch niet dissiperen. Het door de schakelaars geproduceerde hoogfrequent pulsbreedte gemoduleerde signaal wordt door een verliesvrij LC filter gefilterd om het audiosignaal te reconstrueren. De belangrijkste beperkingen van schakelende versterkers zijn het niet integreerbare filter, het schakelresidu aan de uitgang en de invloed van de belasting op de vervorming en de frequentieoverdracht. Door het hogere orde LC filter zijn hoge terugkoppelingsfactoren niet mogelijk.

(Hoofdstuk 4) De combinatie van schakelende en lineaire versterkers is alleen gerealiseerd in systemen waarbij de schakelende versterker de voedingsspanning levert voor een lineaire versterker. Hierdoor wordt een goede controle over de luidspreker gecombineerd met een hoog rendement. Een echt hoog rendement kan echter niet gehaald worden, en de uitgangstransistoren moeten erg groot zijn omdat de twee versterkers in serie staan.

(Hoofdstuk 5) Een ‘serie’topologie die geen grote uitgangstransistoren nodig heeft bestaat uit een lineaire en schakelende versterker in brug. Het common-mode niveau wordt op een dusdanige manier gestuurd dat de uitgangsspanning van de lineaire versterker altijd dicht bij één van de voedingsspanningen ligt. Hierdoor blijft de gemiddelde spanningsval over de uitgangstransistoren klein. Er moeten diverse afwegingen gemaakt worden betreffende de sturing van het common-mode niveau en de vermindering van de vervorming. De resultaten met een prototype van deze versterker laten zien dat de opzet goed functioneert. Het rendement laat nog te wensen over, maar er liggen voldoende mogelijkheden tot verbetering.

(Hoofdstuk 6) Een ‘parallelle’ topologie die geïntroduceerd wordt bestaat uit een schakelende versterker met een stroombronkarakter parallel aan een lineaire versterker. De lineaire versterker bepaalt de uitgangsspanning en draagt zorg voor een lage vervorming en een vlakke frequentieoverdracht. De meeste belastingsstroom wordt geleverd door de schakelende versterker, zodat de dissipatie laag is. Voor een goed werkend systeem moeten aspecten als vermogensbandbreedte, rustdissipatie, en de stabiliteit geanalyseerd worden. Een prototype van deze versterker geeft goede resultaten te zien, zoals een lage dissipatie, weinig externe componenten, weinig vervorming en een klein schakelresidu. Het belangrijkste nadeel is het vrij hoge ruststroomverbruik, maar verscheidene verbeterde topologieën lijken hiervoor een oplossing te kunnen bieden.

(Hoofdstuk 7) Hoog rendementsversterkers staan de laatste jaren opnieuw in de belangstelling, en de prototypes die hier gepresenteerd zijn passen goed in deze ontwikkeling. De specificaties kunnen zich meten aan die van versterkers die in de tijd van dit onderzoek elders ontwikkeld zijn. Snellere IC-processen en een verdere integratie van vermogenselektronica zullen ertoe bijdragen dat hoog rendementsversterkers in veel applicaties een echt alternatief kunnen worden voor klasse AB versterkers.

Selected symbols and abbreviations

η	Efficiency
P_o	Output power
P_{omax}	Maximum sinewave output power
P_i	Input power
P_{inst}	Instantaneous power
P_{avg}	Average power
P_d,P_{diss}	Dissipated power
V_i	Input Voltage
V_o	Output voltage
V_S	Positive or negative supply voltage (equal)
V_{DD}	Positive supply voltage
V_{SS}	Negative supply voltage
V_{thr}	Threshold voltage
f	Frequency
f_{sw},f_{switch}	Switching frequency
f_{sin}	Frequency of a sinewave
Z	Impedance
α	Amplitude as fraction of the power supply
R_o	Output resistance
R_{on}	On-resistance of a MOS transistor
L_o	Output inductance
A	Gain
t	Time
T_{sw}	Switching time
I_o	Output current
I_{thr}	Threshold current
E	Energy
SPL	Sound Pressure Level
THD	Total Harmonic Distortion
IM	InterModulation

TIM	Transient InterModulation
PDM	Pulse Density Modulation
PWM	Pulse Width Modulation
PDF	Probability Density Function
CM	Common Mode
PSRR	Power Supply Rejection Ratio
CMRR	Common Mode Rejection Ratio
LS	LoudSpeaker
AB	Class AB Amplifier
D	Class D amplifier
PAR	Peak-to-Average Ratio
FFT	Fast Fourier Transform
IEC	International Elektrotechnical Committee
DSD	Direct Stream Digital

Nawoord

Promoveren is een afwisselende bezigheid. Ideeën komen soms eenvoudigweg aanwaaien, soms moet je er hard voor werken, maar ze krijgen altijd pas echt betekenis door je omgeving: Ed, bedankt voor je begeleiding; ik heb vaak behoorlijk moeten puzzelen om tot een oplossing te komen die volgens jou ‘misschien wel zou kunnen werken’. De vrijheid die je me gegund hebt, hebben de afgelopen jaren een leuke tijd gemaakt. Eric, bedankt voor alle uren die ik van je gevraagd hebt terwijl je het zo druk had. Je hulp bij het schrijven van artikelen is onmisbaar geweest. Marco, de discussies in Nijmegen waren uitdagend en het maken van een chipje was een stuk minder vlot gegaan zonder jouw hulp. Wilfred, zo weinig als er van je werk in dit boekje terug te vinden is, zo veel is er over gepraat. Marjolein en Sape, bedankt voor jullie bijdragen aan het realiseren van AB/D versterkers en het testen van de robuustheid van klasse D IC’s. Inès, heel fijn dat je altijd wat op m’n engels aan te merken had.

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De universiteit is een mooie plek om je gedachten de vrije loop te laten. Een speeltuin, waar je spannende reizen in wetenschap en kunst kunt maken. Waar je balanceert op de grenzen daartussen. Waar je spartelend onderzoek kunt doen en in de zon bij het buitenbad kunt liggen. Waar ideeën kunnen sprankelen en er ook wat van gemaakt wordt. Ik zal er met plezier aan terugdenken!

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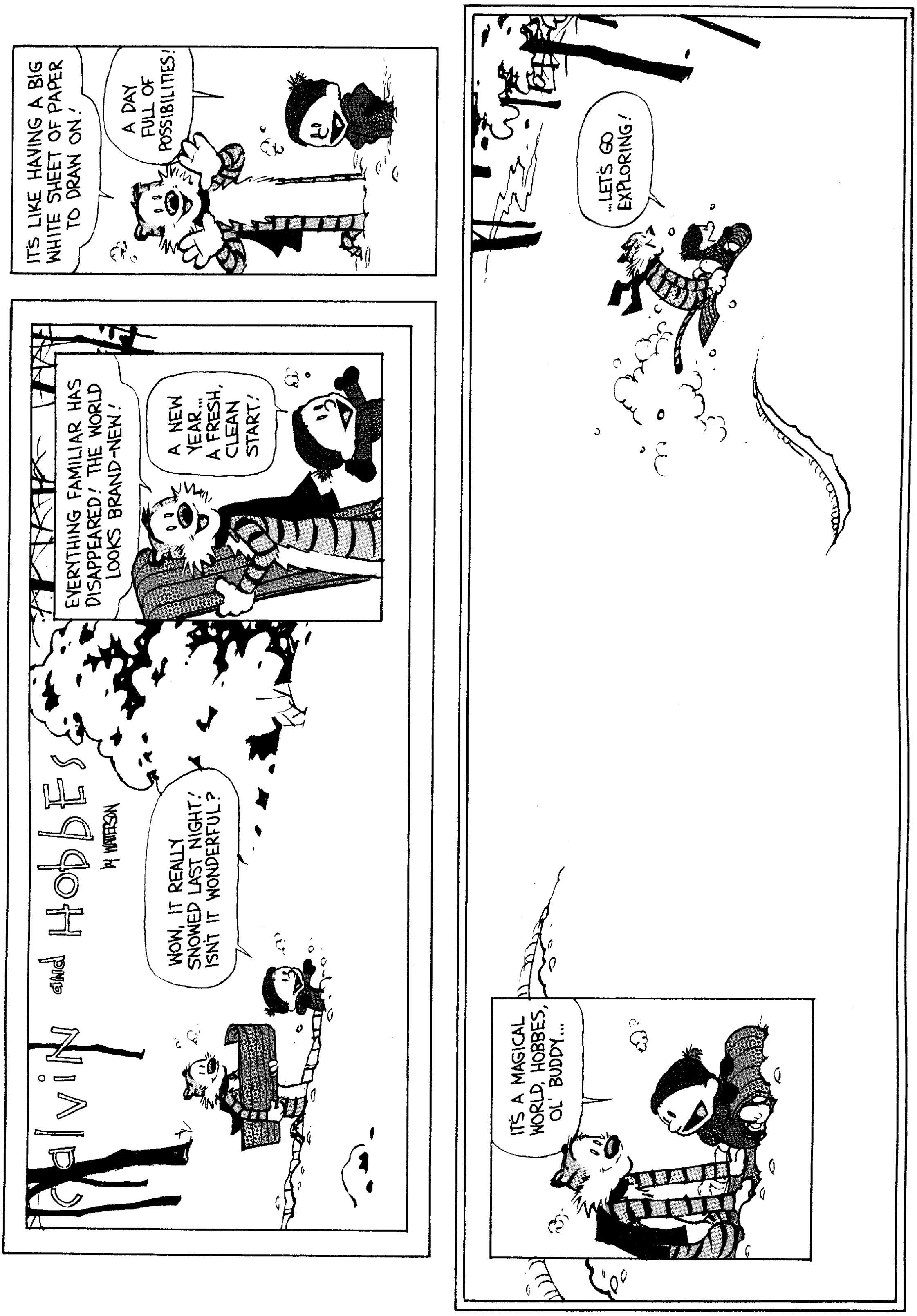
Calvin and Hobbes
by WATTERSON
WOW, IT REALLY SNOWED LAST NIGHT! ISN'T IT WONDERFUL?
EVERYTHING FAMILIAR HAS DISAPPEARED! THE WORLD LOOKS BRAND-NEW!
A NEW YEAR... A FRESH, CLEAN START!
IT'S LIKE HAVING A BIG WHITE SHEET OF PAPER TO DRAW ON!
A DAY FULL OF POSSIBILITIES!
IT'S A MAGICAL WORLD, HOBBES, OL' BUDDY...
...LET'S GO EXPLORING!