A Power-Efficient Audio Amplifier Combining Switching and Linear Techniques

Ronan A. R. van der Zee and Ed (A. J. M.) van Tuijl

Abstract—Integrated class-D audio amplifiers are very power efficient but require an external LC reconstruction filter, which prevents further integration. Also due to this filter, large feedback factors are hard to realize, so that the load influences the distortion and transfer characteristics. The 30-W amplifier presented in this paper consists of a switching part that contains a much simpler filter and a linear part that ensures a low distortion and flat frequency response. The switching part of the amplifier was integrated in a BCD process. Combined with a linear part and with a loudspeaker as load, it has a flat frequency response ±0.3 dB, a dissipation that is up to five times lower than a traditional class-AB audio amplifier, and a distortion of <0.02% over power and frequency range.

Index Terms— Analog circuits, audio amplifiers, distortion, parallel architectures, power amplifiers, switching amplifiers.

Fig. 1.   Typical class-D audio amplifier.

I. INTRODUCTION

DURING the last few years, there has been a growing market demand for audio amplifiers with a high output power and a large number of channels. This is a result of the increasing dynamic range of audio equipment and the movement toward multichannel sound systems (surround sound). At the same time, however, dimensions have become smaller. Minisets, car radios, and PC multimedia equipment have only a little space available. This limitation requires integrated audio amplifiers with a small number of external components and a low power dissipation. Traditional class-AB amplifiers, which are still largely used, are not suited to meet these demands. The efficiency of a class-B audio amplifier can theoretically be 78.5% for a rail-to-rail sinewave, but for real-life music and speech signals, it is much lower: 15–20% [1].

Class-G and class-D audio amplifiers have a higher efficiency than class AB. Class-G amplifiers are operated from multiple supply voltages. Small signals use the lower supply voltage, thus reducing the average voltage drop across the output transistors. This principle works well for music and speech signals because of the approximately Gaussian amplitude-probability-density function of those signals [1], [2]. Similar techniques lift the power supply by means of an electrolytic capacitor [3], or put two loudspeakers in series for small
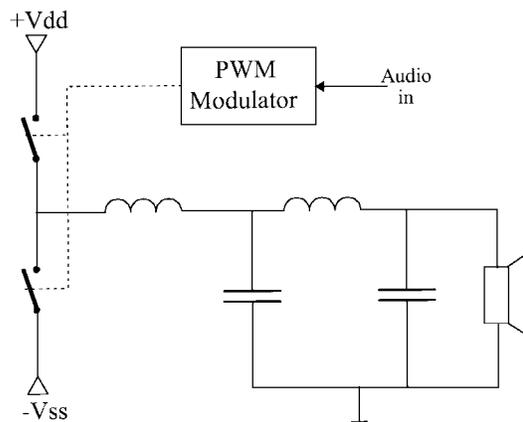
signals, and drive them separately during high output levels [4]. Class-D amplifiers [5]–[9] have a higher efficiency. A typical class-D amplifier consists of a modulator that converts an analog or digital audio signal into a high-frequency pulse-width-modulated (PWM) signal, followed by a half bridge power switch (Fig. 1). The low-frequency audio signal is reconstructed by means of an LC filter. Ideally, the switches do not dissipate and neither does the filter, so the efficiency can be very high. The widespread use of class-D amplifiers, however, has been prevented by a number of limitations, which are discussed in the next section.

II. CLASS DESIGN

Like any audio amplifier, a class-D amplifier should at least satisfy the following demands:
- flat frequency transfer;
- low distortion.

A purely feedforward class-D amplifier (Fig. 1), as is still often used, has problems with both specifications. First, the load seriously affects the frequency transfer. Fig. 2 shows the impedance of a random three-way loudspeaker system. The class-D filter, however, is designed for a real and constant load impedance. The result of connecting the loudspeaker is shown in Fig. 3. 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 30 kHz, loaded with the specified load impedance of 4 Ω. The other line shows what happens when the loudspeaker is connected. The transfer deviates several decibels from the flat line. This will color the sound impression. Concerning distortion, it is obvious that
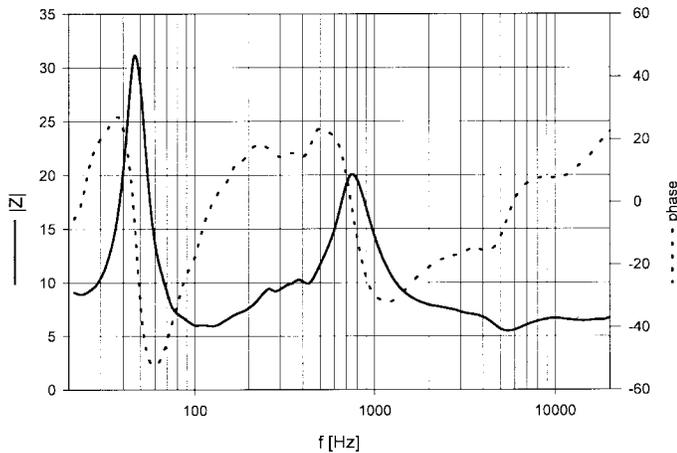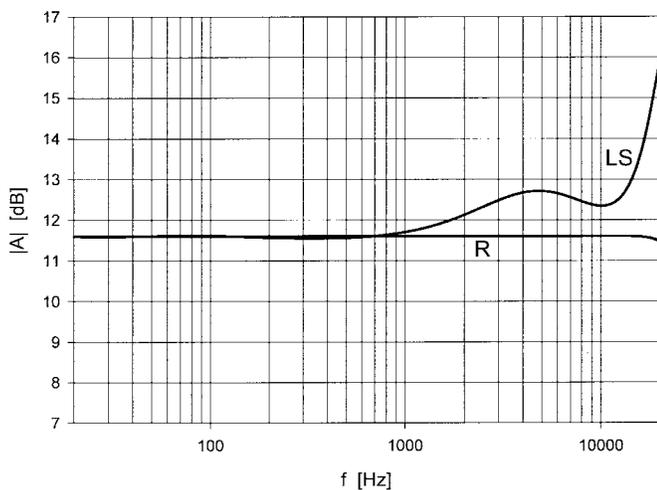
Fig. 2. Loudspeaker impedance.

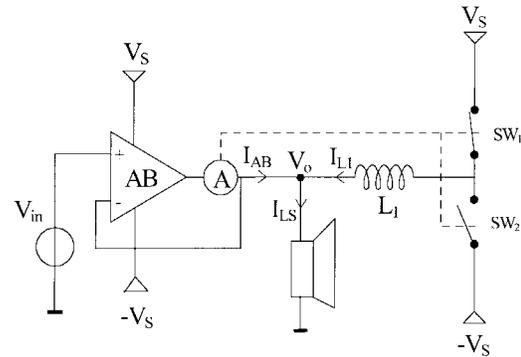Fig. 3. Simulated class-D transfer with resistor and loudspeaker load.

Fig. 4. Circuit principle.

only a small voltage across the output power transistors, the power dissipation is kept low. The class-AB amplifier controls the output voltage; the distortion is low and not sensitive to load variations. Furthermore, a simple filter can be used since the switching residue is reduced by the power-supply rejection of the class-AB amplifier. In practice, however, the power dissipation is not much lower than that of a class-G amplifier. The maximum output power is limited because the class-D and class-AB amplifiers are in series, both introducing a voltage drop. Besides, the voltage drop across the output transistors of the class-AB amplifier must be relatively large for proper functioning.

This paper describes the design and realization of an amplifier that has a switching amplifier and a linear amplifier in parallel [12], [13]. In this way, the number of external components is reduced, and the transfer is much less dependent on the load or the filter characteristics.

III. THE CIRCUIT PRINCIPLE

The circuit principle is shown in Fig. 4. 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 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 ($\pm I_{thr}$), 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. Fig. 5 shows the typical waveforms when a sinusoid is applied to the input.

IV. DIMENSIONING

A. 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. Fig. 6 shows a detail of the typical currents in the coil and in the loudspeaker (refer to

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any output-stage imperfections or filter nonlinearities worsen the distortion figures.

Feedback can reduce these problems considerably. However, applying feedback before the final filter [5], [6] does not eliminate these problems. Feedback after the filter is much more difficult, and high feedback factors can not be realized [8], [9]. Even when these problems are overcome, the filter prevents further integration because of a conflict between two demands of a practical nature:

1) a simple filter to reduce the external component count in integrated designs;
2) a small switching residue to minimize electromagnetic interference.

For sufficient suppression of the carrier frequency, typically a fourth-order filter is necessary. Apart from the problems of applying feedback, 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 will limit the application area of the amplifier.

One way to circumvent the conflicting design criteria of class-D amplifiers is to let a class-D amplifier generate the supply voltage of a class-AB amplifier [10], [11]. By leaving
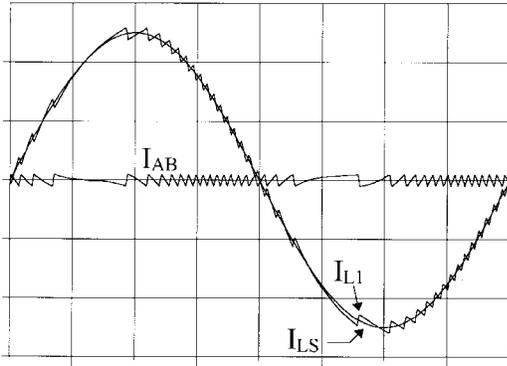
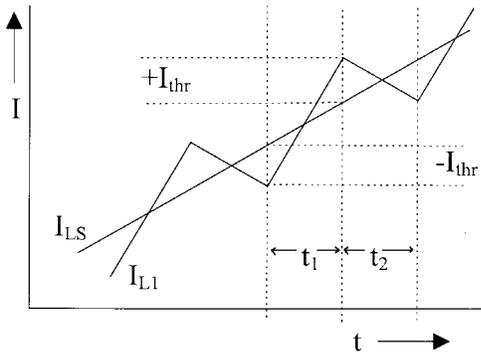Fig. 5.   Typical currents when amplifying a sinusoid.

Fig. 6.   Typical currents (detail).

Fig. 4 for the meaning of the symbols). Suppose the thresholds of A are $+I_{\mathrm{thr}}$ and $-I_{\mathrm{thr}}$, as shown in Fig. 6. The following equations hold during t_1 and t_2, respectively:

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

$$= \left(\frac{V_S - V_o}{L_1} - \frac{dI_{\mathrm{LS}}}{dt}\right) \cdot t_1$$

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

$$= \left(\frac{-V_S - V_o}{L_1} - \frac{dI_{\mathrm{LS}}}{dt}\right) \cdot t_2.$$

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

$$f_{\mathrm{switch}} = \frac{1}{t_1 + t_2} = \frac{V_S^2 - \left(V_O + \frac{L_1}{R_{\mathrm{LS}}} \cdot \frac{dV_O}{dt}\right)^2}{4I_{\mathrm{thr}}L_1V_S}. \qquad (1)$$

Note that the highest switching frequency is at $V_o = 0$. For large output voltages and/or slew rates, f_{switch} goes to zero. This happens when the amplifier clips or when D's power bandwidth is exceeded. The second important parameter is the power bandwidth of D. The slew rate of I_{L_1} 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

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

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_{L_1} is

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

If L_1 has to deliver the main load current, the maximum slew rate of I_{L_1} 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

$$f_{\mathrm{audio,\,max}} = \frac{R_{\mathrm{LS}}}{2\pi L_1}\sqrt{\frac{1}{\alpha^2} - 1}. \qquad (2)$$

When D is not able to provide the full load current, AB could supply the rest. This is the approach in [12], resulting 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.

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

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

This expression is a good starting point for the dimensioning of the amplifier. By making $I_{\mathrm{thr}} = 100$ mA, the quiescent power dissipation is raised by 0.9 W, which is approximately half of the total quiescent power. Other practical values for a realization are $\alpha = 0.85$, $V_S = 18$ V, and $R_{\mathrm{LS}} = 4$ Ω. According to (2), for a power bandwidth of 20 kHz, L_1 should be 20 μH. With (1), it can be calculated that the resulting free-running switching frequency at $V_o = 0$ would be 2.25 MHz. Since this leads to unacceptable switching losses in the power switches, L_1 is chosen to be 80 μH. With this value, the amplifier can deliver a full-power sinusoid up to 5 kHz, decreasing to 20% of full power at 20 kHz. This makes the amplifier a suitable candidate for transient intermodulation distortion (TIM). However, audio signals have a limited high-frequency content [14], so this should not be a problem. Indeed, computer simulations with a behavioral model of the amplifier rarely showed any distortion of this type.

B. AB's Output Impedance

At first, the prototype 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 nonzero
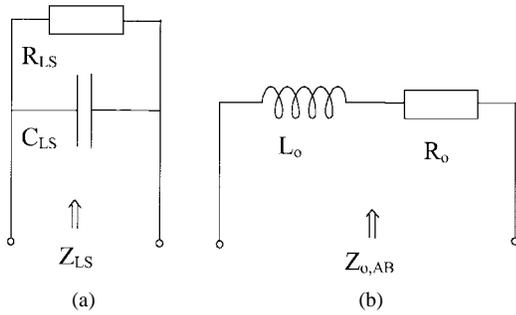
Fig. 7.   A model of (a) the loudspeaker and (b) the output impedance of AB .

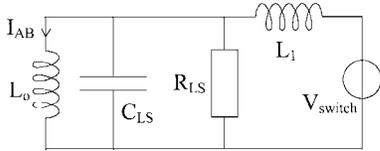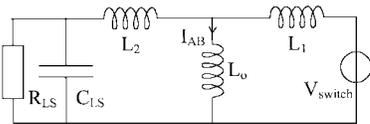Fig. 8.   Behavioral model of the amplifier at f_{switch}.

Fig. 9.   Model of the amplifier at f_{switch} with stabilizing inductance L_2.

output impedance. Furthermore, a loudspeaker may have a complex impedance, and especially its capacitance may lead to instability. Assume that the loudspeaker and the output impedance of AB can be modeled as in Fig. 7. R_{LS} is the loudspeaker's dc resistance and C_{LS} its parallel capacitance (this could, e.g., be the cable capacitance). The amplifier AB uses feedback to reduce distortion. Its loop gain has a first-order behavior above 0 dB, and its dominant pole lies within the audio range. At dc, the loop gain reduces the output resistance to R_o (Fig. 7). For frequencies higher than the pole frequency, the loop gain 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. The stability of the total amplifier is analyzed with the behavioral model shown in Fig. 8. V_{switch} represents the common node of SW$_1$ and SW$_2$. 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 Fig. 6. For this, we calculate the admittance $Y = I_{\text{AB}}/V_{\text{switch}}$

$$Y = \frac{1}{sL_1} \cdot \frac{1}{1 + s\frac{L_o}{R_{\text{LS}}} + s^2 L_o C_{\text{LS}}}.$$

The first term describes the desired behavior, 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 inductance in series with the loudspeaker. The resulting model then looks like Fig. 9. When calculating Y for this
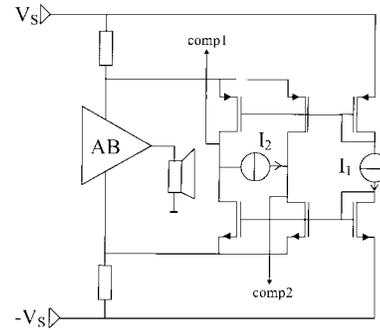

Fig. 10.   Output current sensing circuit.

model, we find

$$Y = \frac{1}{sL_1} \cdot \frac{1 + s\frac{L_2}{R_{\text{LS}}} + s^2 L_2 C_{\text{LS}}}{1 + s\frac{(L_o + L_2)}{R_{\text{LS}}} + s^2 (L_o + L_2) C_{\text{LS}}}.$$

By choosing $L_2 \gg L_o$, the nominator and the denominator of the second term cancel, leaving only the desired first term. The measured virtual output inductance L_o was 15 nH, so for $L_2 \gg 15$ nH, the amplifier is stable. The series inductance in most loudspeakers will have the same stabilizing effect as L_2.

The output resistance of AB also determines the amount of switching residue at the output. It is easy to see (Fig. 9) that when $L_1 \gg L_o$, the switching residue relative to the class-D half-bridge output is L_o/L_1, which is in our case -74 dB.

V. REALIZATION

The modular structure of the amplifier is reflected in the experimental realization. The linear part is at present still external, and is built with a commercially available power op-amp. It must source and sink 100 mA. The switching part of the amplifier was realized in two modules in a BCD process (a process that allows bipolar, CMOS, and DMOS devices on the same chip). Fig. 10 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 200 μA. The measuring resistors are 0.1 Ω and also integrated. Two scaled copies of the output current are made, each of which receives an opposite offset by means of I_2 (5 μA). 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.

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 adds up to a deviation in $+I_{\text{thr}}$ and $-I_{\text{thr}}$, changing dissipation and switching frequency. With the process and transistors used, deviations smaller than 3σ result in a switching frequency between 300 kHz and 1 MHz and an extra dissipation of less than 1.2 W. In Fig. 11, the copies of the output current are connected to the comparators comp1 and comp2. Regenerative comparators offer the lowest power-delay products [15], but since no clock signal is available, a multistage amplifier design is chosen. The amplifiers are inverters with a feedback resistor.
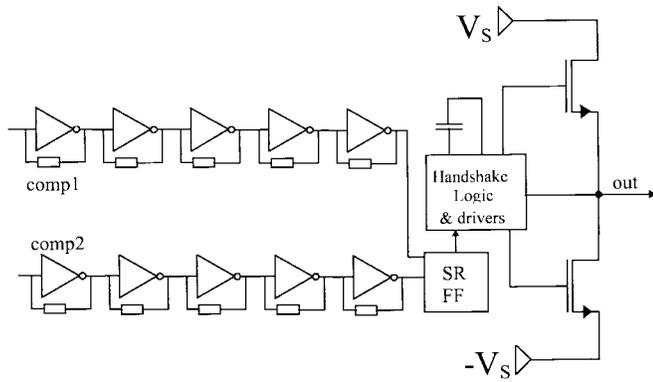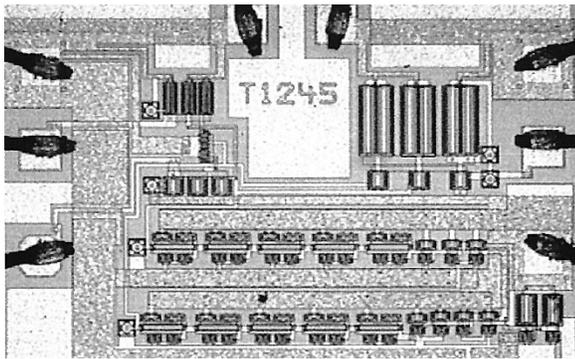
Fig. 11.   Comparators and power switches.

Fig. 12.   Chip photo of sensing and comparator section.

Six 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 Fig. 12. This comparator module is followed by the power switches, realized 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.

VI. MEASUREMENTS

A 30-W version of the amplifier was measured. The maximum efficiency for sinusoidal signals is 85%. 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 [16]. The signal was used up to an output power of 15 W 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 Fig. 13. 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
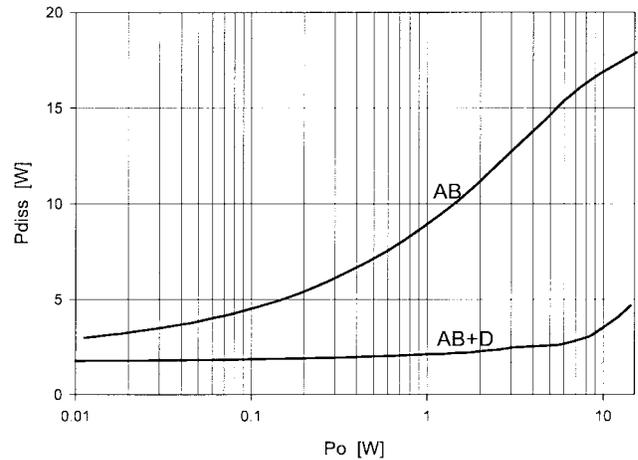

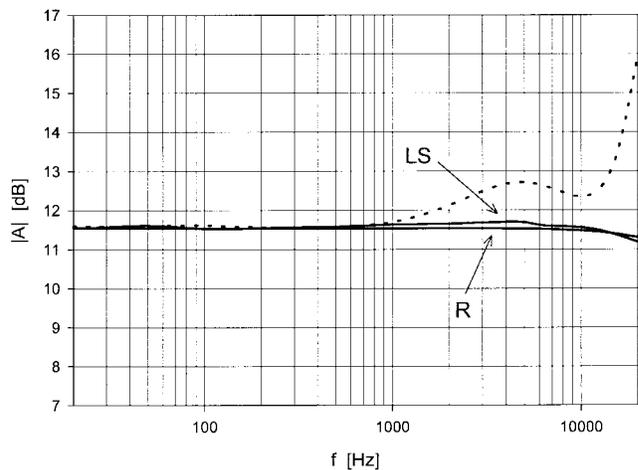Fig. 13.   Power dissipation for the IEC 268 test signal.

Fig. 14.   Frequency transfer with resistor and loudspeaker. Dotted line is simulated class-D frequency transfer with loudspeaker (Fig. 3).

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 [see the comments below (2)]. 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.

The load hardly influences the frequency transfer of the amplifier, as is shown in Fig. 14. Fig. 14 displays the frequency transfer for both a resistor and a loudspeaker connected to the output. The decrease at high audio frequencies is a result of the output inductor, but it is still a major improvement over Fig. 3, displayed as a dotted curve in Fig. 14.

Fig. 15 shows that the distortion with a resistor load at 1 kHz remains less than 0.02% up to the clipping point at 30 W. The distortion at different frequencies is shown in Fig. 16. 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 analyzer. The distortion with resistor load is virtually frequency independent, although it is expected to increase with frequency due to rolloff of the loop gain. Probably, noise caused by switching of the output current is
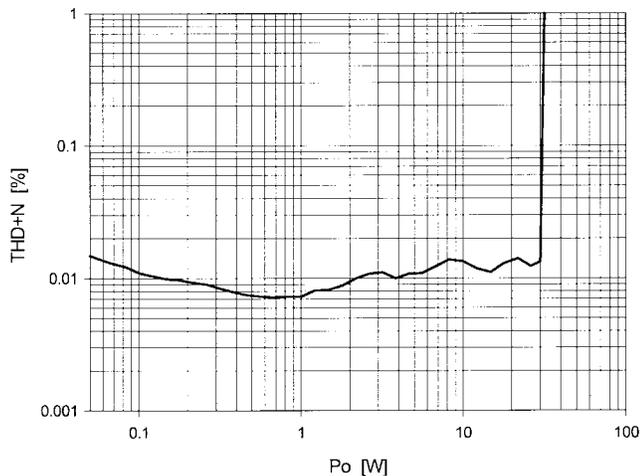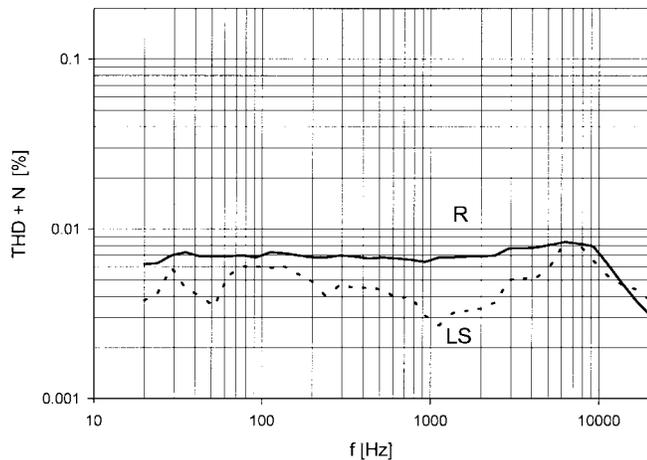
Fig. 15. THD + N at 1 kHz, filter 22 Hz–22 kHz.

Fig. 16. THD + N at 1 W, filter 22 Hz–22 kHz.

TABLE I
AMPLIFIER SPECIFICATIONS

Parameter	Condition	
Switching frequency	V_o=0	550kHz
Max. output power	R_{LS}=4Ω,THD=0.1%	30W
Maximum efficiency	P_o=30W	85%
Quiescent power		1.8W
Power dissipation	P_o=10W IEC268 test signal	3.5W
THD+N (filter 22Hz-22kHz)	20Hz-20kHz,1W-30W R or LS load	0.003-0.02%
Frequency response	20Hz-20kHz R or LS load	+/- 0.3dB
f_{switch} attenuation	f=500kHz, R_{LS}=4Ω	77dB

influencing the linear part, because the distortion is lower at frequencies where the loudspeaker impedance is higher. The power-supply rejection ratio is also a significant specification for audio amplifiers. In this amplifier, it is mainly determined by the external linear part, but for a fully integrated realization, it is a point of attention. The measured residual switching noise at the output is 3 mV, which is equal to a 77-dB reduction. In a standard class-D amplifier, a fourth-order output filter would be needed to achieve this value. Table I summarizes

the main specifications of the amplifier. A comparison with 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 [3], [4], [10], [11]. 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 [6]–[8]. Overall, the amplifier combines good audio performance with low dissipation.

VII. 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 distortion and frequency transfer of the amplifier are less dependent on the connected load. This is a new step toward highly integrated, power-efficient audio amplifiers. Future work will concentrate on reducing the quiescent power dissipation, designing the linear amplifier in the same process, and merging the different modules on one chip.

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