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An asynchronous switching high-end power amplifier

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ABSTRACT
Switching power amplifiers need an analogue low-pass filter between the power transistors and the speaker to prevent HF switching noise to enter the speaker. Cancellation of the non-linearities, as well as the (load dependent) frequency transfer of this filter is desired, but traditionally very difficult because of stability problems with filter-output feedback. A switching power amplification topopogy is presented which offers total control over the output filter dynamics and cancellation of filter characteristics. The proposed amplifier uses a hysteresis circuit with a high-bandwidth capacitor-current feedback as PWM modulator. The output filter is thus an integral part of the modulator which dramatically improves control of the output. The result is a very high fidelity amplifier which is unconditionally stable, regardless of the load.

INTRODUCTION
During the last few years, switching amplifiers have become increasingly popular in all segments of the market. This ranges from single-chip filterless amplifier modules in notebook computers up to several hundred watt audiophile quality amplifiers. This recent increase in popularity is mostly the result of improved modulation and noise-shaping techniques [1][2][3].

In the medium and high power range, a passive analogue low-

pass filter between the switching power stage and the speaker is required to reduce RF radiation and switching noise entering the speaker. This output filter is responsible for two major problems. First, the filter is an additional low-pass filter in series with the speaker, thus it decreases amplifier control over the speaker, especially at high frequencies. The frequency transfer of the amplifier becomes load dependent; for low load impedances the high-frequency transfer will drop, whereas for high load impedances the high-frequency transfer will peak due to increased filter-Q. The amplifiers' transfer becomes increasingly unpredictable for loads which are not purely resistive, which is the case for practical speakers. In such cases the crossover filter of the speaker can be detuned because of interaction with the amplifier's low-pass filter. To reduce the negative effects of the output filter, the crossover frequency of the filter is chosen rather high, typically 60-80kHz [4], which is a compromise between RF attenuation and output impedance.
Secondly, the filter components suffer from non-linearities. For example: hysteresis- and saturation effects in ferrite and iron core inductors will cause harmonic distortion and thus are often replaced by air-core alternatives [4]. In general, the filter has to be designed with extreme care, ensuring highest possible linearity up to the maximum power and over the full audio frequency range. This makes the output filter one of the more expensive components in the amplifier.

Attempts have been made to improve control and reduce distortion by applying filter-output feedback. Due to the delay of the LC filter, such amplifiers are prone to stability problems [6] which in practice limits the gain of the feedback loop. As a result, most commercial switching amplifiers have an uncompensated filter.

ERROR SOURCES IN SWITCHING AMPLIFIERS

Switching amplifiers have very different distortion mechanisms compared to analogue amplifiers, due to the switching actions. Linearity in the signal frequency band is directly compromised by incorrect timing and slopes of the high-frequency switched waveform. The distortion sources can be subdivided in modulation errors, switching errors and filter errors:

1. Modulation errors.
   (a) PCM to PWM conversion in digital-input amplifiers is by definition uniform sampled and therefore non-linear[7].
   (b) non-linearities in the triangular waveform in analogue-input PWM amplifiers.
2. Switching errors
   (a) due to the necessary dead-time, effective pulse-width of the PWM waveform becomes load- and signal dependent [8].
   (b) the output impedance of the DC supply results in load- and signal dependent pulse-heights[4].
   (c) ringing effects in the switching bridge[5].
3. Filter errors
   (a) load dependent frequency transfer (and corresponding signal delay) due to interaction between the passive output filter and the complex impedance of the speaker.
   (b) linearity of the filter is difficult to maintain up to full power. This requires special (expensive) capacitors and (air) coils[4].

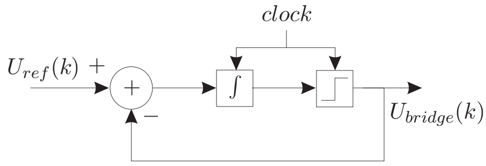

Fig. 1: Basic sampled ΣΔ modulator circuit.

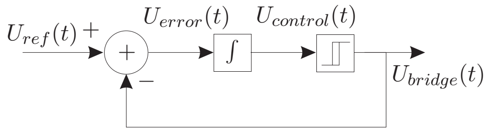

Fig. 2: Basic hysteresis controller circuit.

These are the main issues to adress when designing a switching amplifier [9]. Many of these issues have been subject of detailed studies. For example; the distortion due to uniform sampling can be reduced by applying quasi-natural sampling [1] or non-linearity compensation [2]. However both require additional calculation, thus increasing complexity.

Switching errors can be minimised with short dead times, down to 10ns, but timing accuracy of much less than 1ns is required for accurate CD reproduction, not to mention the new DVD and SACD standards. Fast switching times require carefull circuit layout and RF screening. Another option is bridge-output feedback, where the generated high power waveform is fed back to the modulator input. This reduces switching errors as well as pulse-height errors resulting from a modulated DC supply.

The filter remains problematic. The passive LC output filter is generally tuned at a rather high frequency of typically 60-80kHz to reduce the load dependency of the signal transfer and the output impedance. In return, the attenuation of HF switching noise is reduced, which might be unpleasant for the the speakers and results in increased electro-magnetic interference (EMI).
This paper presents a modulator and controller that solve all these problems in a unique way.

ANALOGUE ΣΔ: HYSTERESIS MODULATION

Much experience has been gained with sigma-delta (ΣΔ) modulators over the last 15 years for AD and DA conversion. These modulators are discrete-time or clocked modulators such as displayed in the example circuit in figure 1. The ΣΔ modulator can also be implemented as a continuous-time circuit by removing the clock input from the comparator. The result is a self oscillating circuit which oscillates at an infinitely high frequency. The switching frequency will be infinitely high because the differential gain (around 0) of the comparator is ∞. By replacing the comparator with a hysteresis circuit, the oscillation can be slowed down. The resulting circuit is shown in figure 2.

The hysteresis circuit lowers the oscillation frequency to a

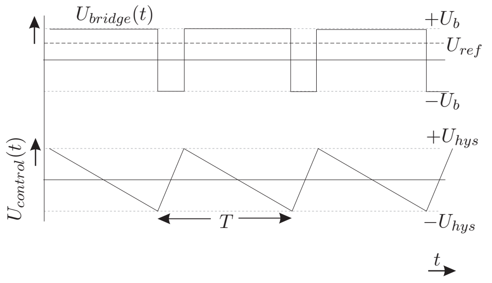

Fig. 3: Main signals in a hysteresis controller circuit with DC input.

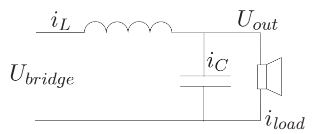

Fig. 4: LC output filter for switching ripple suppression.

more sensible value of choice, depending on the width of the hysteresis window. A typical value for audio application is 300 to 400kHz which is sufficiently high for high fidelity, while still low enough to prevent exessive dead-time errors. The self-oscillating loop of figure 2 is commonly known as a hysteresis controller. Although the hysteresis controller is the workhorse of industrial power electronics, it seems to be unfamiliar to audio engineers. This is probably due to its seeming lack of accuracy; the hysteresis controller allows errors to exist within a certain window. If the error is ultrasonic however, the hysteresis controller can be put to practical use even in high end audio.

The behaviour of the hysteresis controller differs fundamentally from its discrete-time counterpart. Figure 3 shows the most important signals, assuming DC input. The behaviour of the hysteresis modulator is as follows: the hysteresis circuit holds its state if its' input signal $U_{control}$ is within the hysteresis band. Assume that the state of the hysteresis circuit is low. Through the negative feedback and the integrator, $U_{control}$ will be rising. If $U_{control}$ rises above the upper hysteresis limit, the output U_{bridge} will switch high. Again, through the negative feedback and the integrator, $U_{control}$ will start to fall until the lower hysteresis limit is reached. Then the output U_{bridge} will switch low again.

Because of the analogue nature of the hysteresis modulator, the modulator doesn't suffer from typical digital artefacts such as quantisation errors and idle tones. Furthermore, because there is no quantisation, no noiseshaping is necessary, thus a first-order integrating feedback loop suffices (compared to 5^{th} or 7^{th} order feedback in noiseshaping feedback loops).

Integration by output filter

As shown in figure 2, a hysteresis control loop requires a comparator with hysteresis as well as an integrator in the feedback loop. The RF suppression output filter of a switching amplifier is usually a 2^{nd}-order LC filter such as shown in figure 4. Both the inductor and the capacitor in this filter have integrating properties. The question now is: is it possible to use one of these filter integrators as loop integrator?

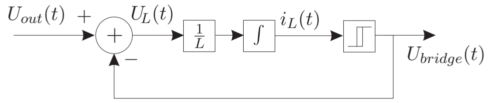

Fig. 5: Using the filtercoil as loop integrator in a hysteresis controller.

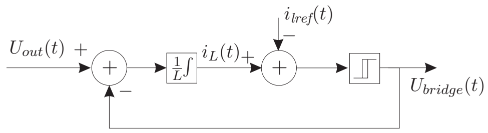

Fig. 6: Hysteresis controller with filtercoil integration and additional coil-current setpoint.

At the oscillation frequency, the filter impedance is dominated by the inductor. The current through the inductor i_L is the integral of the bridge voltage U_{bridge} (minus the nearly constant output voltage U_{out}), thus i_L is a proper signal for feedback. The corresponding block model is shown in figure 5. The block model in figure 5 is very similar to the model of the hysteresis modulator in figure 2, only no input is present. The modulator in figure 5 attempts to make the inductor signal current 0A. With addition of an explicit setpoint i_{Lref} for the inductor current to the loop (figure 6) it is possible to make a non-zero inductor signal current i_L and thus, indirectly, an output voltage U_{out}. Unfortunately, this setup is of limited use for amplification purposes because the coilcurrent i_L is load dependent and therefore unknown without explicit measurement.

Capacitor current feedback

Because the capacitor impedance is much lower than the load impedance at the oscillation frequency, the triangular high-frequency component in i_L flows through the capacitor. The difference between i_C and i_L is that whereas i_L contained an unknown audio component, the desired audio component in i_C is known exactly: i_C is the mathematical derivative of the output signal U_{out}.

$$i_C = C_{filt} \cdot \frac{dU_{out}}{dt} \tag{1}$$

Thus it is easy to define a capacitor current setpoint i_{Cref}, for it is the derivative of the reference voltage U_{ref}.

$$i_{Cref} = C_{filt} \cdot \frac{dU_{ref}}{dt} \tag{2}$$

Figure 7 shows the model of the hysteresis output stage including the LC output filter and an i_C setpoint i_{Cref}. The hysteresis circuit tries to keep the capacitor current error around zero. If the realised i_C has the correct value, the correct capacitor voltage U_{out} is automatically obtained.

CONTROL LOOP

The PWM modulator, constructed around the hysteresis circuit and the output filter, delivers the correct voltage at the output when applying a capacitor current setpoint i_{Cref} which is the derivative of the desired output voltage. At high loads (low load impedance) or high output voltage, the shape of the triangular capacitor current error waveform starts to deteriorate due to increased ripple voltage on U_{out}. In addition to

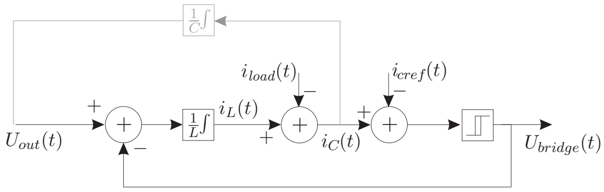

Fig. 7: Hysteresis controller with capacitor current feedback. Filter coil is used as loop integrator.

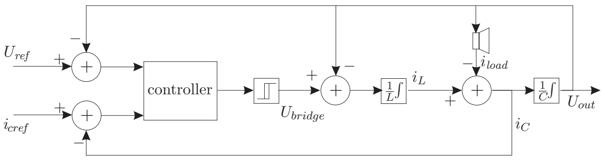

Fig. 8: Final block model of hysteresis amplifier with 2nd order output filter, capacitor current feedback and additional proportional feedback.

this, current measurement is less accurate for low-frequency signals. In order to improve performance, an additional proportional error signal $U_{ref} - U_{out}$ is fed to the controller. Just as with conventional switching amplifiers with output feedback [6], the maximum gain of the proportional feedback loop is limited. The reason for this is that the current feedback loop should stay dominant at high frequencies in order to guarantee the high frequency oscillation condition. If the high-frequency gain of the proportional loop is too high, the amplifier will start to oscillate at a lower frequency, due to the additional delay of the 2$^{\text{nd}}$ order output filter.
With the filter-output and capacitor current feedback loops, the controller is effectively a PD controller (a controller with a proportional plus a differentiating action). The low frequency performance can be further improved by adding an integrating action to the controller. The model of the hysteresis modulated amplifier with capacitor current feedback and output voltage feedback is shown in figure 8.

AMPLIFIER PROPERTIES

Conventional modulators make a transparent signal to PWM translation, either explicitly (like sine-triangle modulation or PCM-PWM conversion) or implicitly (for example $\Sigma\Delta$ modulators with bridge output feedback). In other words: they require a bridge output setpoint at the modulator input. Usually the input signal is used as modulator setpoint, which leaves the output filter and dead-time errors un-compensated. A control loop with filter output feedback can be applied to make a modified modulator setpoint which accounts for these errors, thus compensating the output filter. However, this is difficult because of the 180° phase-shift of the output filter, the delay of the modulator and the presence of remaining switching ripple. This delay causes stability problems in the control loop.

On the other hand, the hysteresis amplifier focuses on input-output behaviour (*after* the LC filter). In contrast to conventional modulators, the hysteresis modulator *exploits* the filter delay in order to deliberately start oscillation. The modulator makes any bridge voltage without a bridge voltage setpoint, its sole interest is to maintain the correct filter output. Because only signals for which a correct setpoint is available are used in feedback (there is no bridge output feedback), the modulator always has a zero-error.
Other features of the hysteresis amplifier are:

- Integrated solution; one circuit combines the functions of PWM modulator, error correction and switching ripple filtering in one inseparable circuit.
- Inherently very good power supply rejection thanks to hysteresis modulator.
- Load insensitive thanks to capacitor current feedback (a differentiating controller action). Output impedance is less than 1mΩ at 1kHz, 14mΩ at 10kHz and 30mΩ at 20kHz.
- No clock and no quantisation, therefore no high order noise shaping is required.
- Varying switching frequency.
- Signal bandwidth can extend the output filter bandwidth because the filter transfer is removed from the system transfer by the control loop, without having explicit knowledge of the filter parameters. The response of the power stage including the filter (figure 8) is completely flat, the overall response is determined solely by the transfer function of the differentiating setpoint filter at the input.
- Timing requirements in the power stage are relaxed compared to other topologies. The prototype had a dead time of over 100ns and a delay of 20ns from controller output to bridge output.
- Amplifier power efficiency is better than 90% for all frequencies at rated power. Idle power is approximately 3W per channel.

Measurements
The modulator/controller topology described in this paper has been named 'Mueta' and an international patent has been applied for [11]. Several pre-production prototypes have been built by Mueta b.v.[1]. Measurements on one of these prototypes demonstrating the amplifiers' performance will be presented now.
The device under test is shown in figure 9. It has a full H-bridge output stage and is designed for 200W per channel into 4Ω, so the maximum output is $40\mathrm{V_{peak}}$ or $28.3\mathrm{V_{RMS}}$. Vertical scales in the plots are in dBV, so maximum output is 29dBV. The measurements were performed with Audio Precision system 2. In order to correctly asses the performance of a switched-mode system like the Mueta amplifier, high-frequency switching ripple was removed with an internal Audio Precision 22Hz-22kHz bandpass filter. No brick wall (AES17) or weighing filters were applied.

The first 2 measurements (figures 10 and 11) show the total harmonic distortion (THD) of the amplifier under various conditions. Figure 10 shows THD versus frequency at 10W and 100W. The distortion is extremely low for a switching amplifier. This result shows the effect of the control loop which cancels any errors in the power stage, as well as the output filter. Furthermore, the distortion also doesn't rise with increasing power. This is confirmed by figure 11 which shows the THD versus power with an input signal of 1kHz.

[1]The invention will commercially be exploited by Mueta b.v., the Netherlands

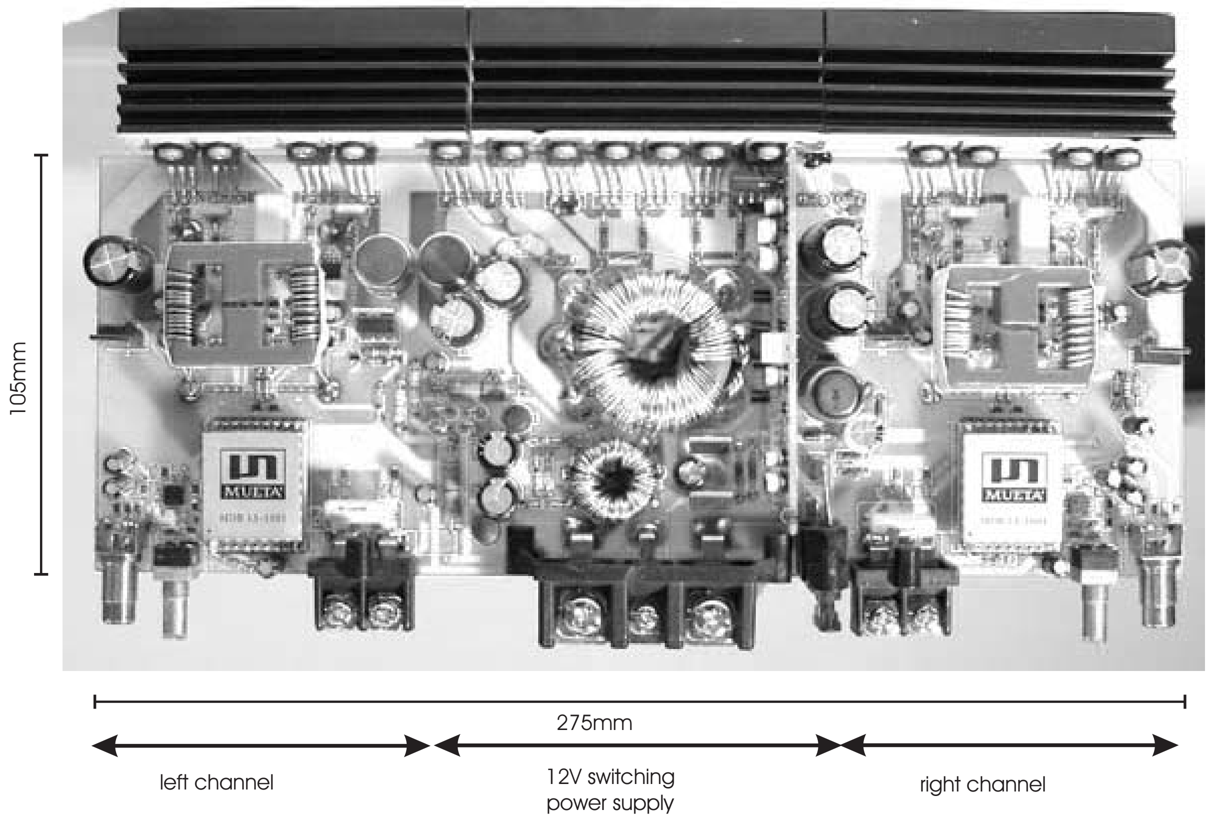

Fig. 9: Printed circuit board of the Mueta prototype amplifier.

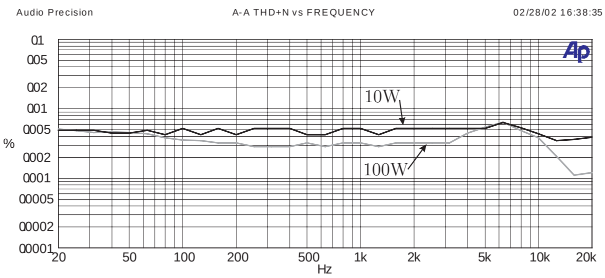

Fig. 10: THD (unweighted) versus frequency of Mueta prototype amplifier at 10W and 100W.

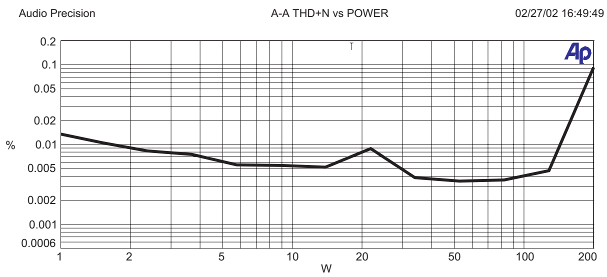

Fig. 11: THD (unweighed) versus power of Mueta prototype amplifier. Signal frequency was 1kHz.

Another measurement, possibly correlating more with subjective audio quality is the inter-modulation test. First in figure 12 we see a conventional inter-modulation test with 19kHz and 20kHz input signals of equal amplitude. The 1kHz inter-modulation product does not rise above the noise floor at -110dB, which is unprecedented. The 18kHz inter-modulation product does exist, but is well below -90dB. This is an exceptionally low value, even for a linear (class A) amplifier.

A more elaborate test is a multi-tone test over the full audio frequency spectrum. The amplifier was supplied with a multi-tone waveform (a short signal fragment stored on hard disk) input of 6 tones per octave from 20Hz to 20kHz. The frequency spectrum of the output is shown in figure 13. The inter-modulation products are below -100dBV over the full audio frequency band. The degradation at lower frequencies is an artefact of Audio Precision. The stored waveform is simply too short to accurately represent these low frequencies.

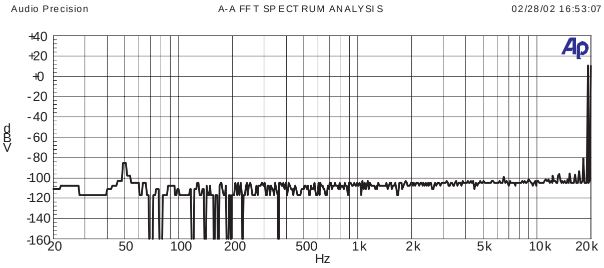

Fig. 12: Inter-modulation distortion measurement of Mueta prototype amplifier with 19kHz and 20kHz input at 5W.

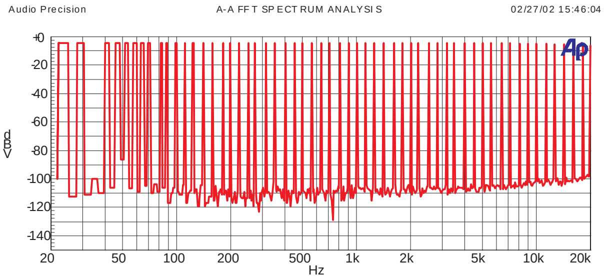

Fig. 13: Inter-modulation with multi-tone (6 tones per octave) input of Mueta prototype amplifier Total RMS power is 5W.

Figures 12 and 13 also show the exceptionally large dynamic range of the amplifier. The noise floor is below -100dBV while maximum RMS output is +29dBV. The dynamic range is therefore over 130dB.

A convincing demonstration of the amplifiers' low output impedance is an experiment to assess the amplifiers suitability as AC line voltage supply. The amplifier is connected to a bridge rectifier which is DC loaded with a 4700μF capacitor and a 8Ω resistor. The AC frequency is 60Hz and the total rectified output power 73W.
Figure 14 shows the output voltage and current of the amplifier, as well as the frequency spectrum of the voltage. The load current pulses are 16A_{peak} high but the output voltage is not measurably distorted. The spectrum of the output voltage shows the 2^{nd} harmonic to be 50dB below the fundamental, but since this is the same in the no-load condition, the distortion can largely be attributed to the distortion of the function generator at the amplifiers' input. This measurement shows the effectiveness of the capacitor current feedback in realizing a stiff voltage source.

CONCLUSIONS

The general approach to class-D amplifier building has been to optimize the different components (the modulator, the power bridge and the output filter) seperately. According to this approach, much effort has been put into design of low-distortion PWM modulators. Combined with a bridge with almost nonexistent dead-time and an expensive filter, adequate performance was achieved, but only at high production costs.

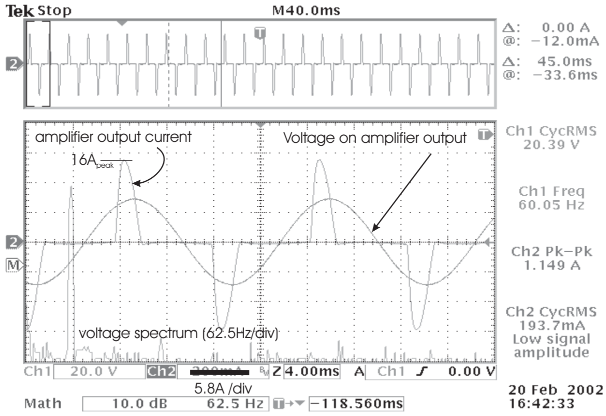

Fig. 14: Mueta prototype amplifier output voltage and -current when loaded with a rectifier.

The proposed amplifier design focuses on input-output behaviour instead, meanwhile solving existing modulation problems. The result of this approach is a class-D power stage which offers performance far beyond the current state of the art.
The advantages of applying feedback are obvious: the qualitative demands on the modulator, the power bridge and the filter are much more relaxed because in the end the control loop takes care of the errors. This is only possible if the loop gain of the control loop is high enough, which is ensured by the hysteresis comparator.

Digital switching amplifiers can be considered 'power DA converters'. Because of the digital nature of their modulator and controller they will never be able to exploit feedback to the extent that analogue amplifiers can, simply because feedback of any signal from behind the power bridge is impossible because these signals are by definition analogue. Digital amplifiers will therefore always suffer more from a non-ideal power stage than analogue amplifiers. We therefore advise to perform DA conversion in a low-power circuit.
Analogue technology has evolved over the last 20 years, even more so than digital technology. Perhaps this is due to analogue application of 'digital' components; analogue circuits can do more with the extended bandwith of 'digital' switches. We have shown that analogue circuitry is now capable of performance far better than its digital equivalent with very modest requirements for component specifications and finishing (such as screening).

The proposed switching (non-linear) power amplifier yields excellent power efficiency while surpassing specifications of conventional linear (inefficient, class A and AB) amplifiers.

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