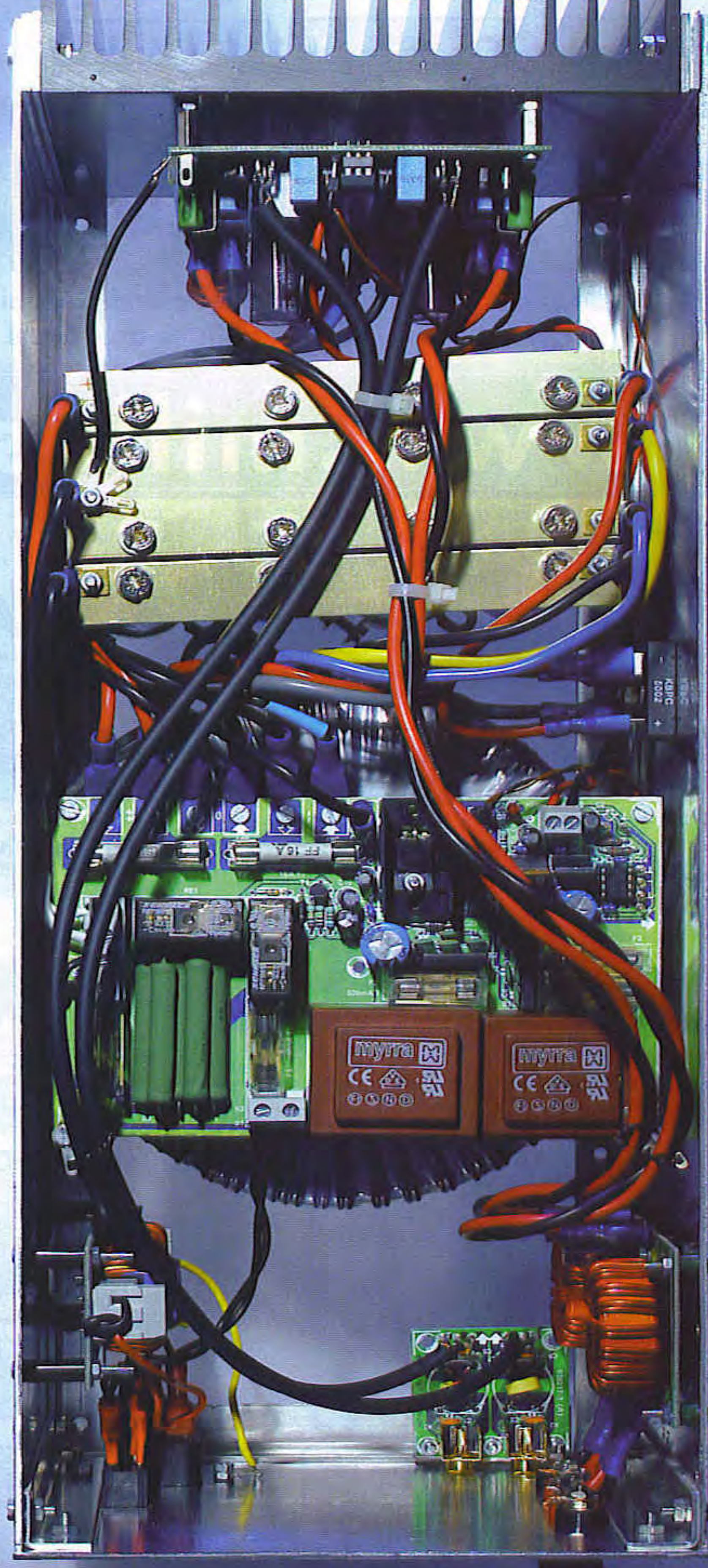
ClariTy 2x300W

Ton Giesberts

This top-end amplifier proves that high power does not have to mean a large, heavy design. Although this amplifier is highly efficient (and thus compact), its specifications easily surpass those of quite a few conventional designs.



Class-T amplifier

If we've given you the idea that the fully assembled amplifier is as light as a feather, perhaps we should qualify our statement somewhat. After all, 2 × 300 watts of *true* power naturally requires a rather substantial power supply. But that's the only aspect of this amplifier that is comparable to other types of amplifiers. Thanks to clever use of pulse-width modulation, this amplifier is so efficient that a heat sink with quite modest dimensions can be used, which means that the enclosure can also be kept relatively small. What's more, this amplifier is not an ordinary pulse-width amplifier. This design, which is based on the Tripath TA3020 Class-T digital audio driver, has excellent specifications and can easily hold its own against other top-end amplifiers. For more information about pulse-width modulation in audio final amplifiers, please see the article 'That's class...' elsewhere in this issue.
The design is largely based on the standard application example and the manufacturer's reference PCB layout. This is because the board layout largely determines the quality of the overall amplifier. Besides this, the nature of this design (with high switching frequencies and large currents) imposes severe requirements on various components. That means that special types of electrolytic capacitors and decoupling capacitors are used in most locations. Even for the thermal coupling between the output transistors and the heat sink, ordinary mica or Kapton washers are not satisfactory. Instead, ceramic washers with a thickness of several millimetres must be used. The IC also needs two auxiliary supply voltages, for which a separate printed circuit board has been developed. This board also includes a mains voltage switch-on delay for the main transformer and two hefty fuses for the main supply voltages. To suppress electromagnetic interference (EMC), extra filters are also included at the inputs and outputs. This should already give you an idea of what to expect, but in this first part of the article we primarily concentrate on how the Tripath IC works.
In **Figure 1**, you can clearly see that the IC essentially consists of three sections for each channel: an analogue input stage (inverting amplifier), a processing and modulation unit, and the driver stages (level shifters) for the external power MOSFETs. The IC also provides overcurrent protection, overvoltage and undervoltage protection, and a connection for an external mute signal. All of these collectively determine whether the amplifier outputs are active.

Input stage

The analogue input stage is implemented as an inverting amplifier for convenient dimensioning of the gain and bandwidth. According to the IC specifications, the maximum allowable signal level for fully driving the modulator is 4 V_{pp}. With the dimensioning used here, assuming an input sensitivity of 1.13 V_{eff} for maximum output amplitude, the output of the input stage can be driven to 3.2 V_{pp}. The ratio of R3 and R2 (R24 & R23 for the second channel) determines the gain of the input stage. Here the ratio is 1, as can be seen from the schematic diagram in **Figure 2**. Capacitor C2 (C15) increases the stability of the input amplifier and suppresses RF noise by limiting the bandwidth to approximately 240 kHz. C1 (C14) sets the lower corner frequency, which in this case is around 2.5 Hz. The gain for frequencies in the audio band is thus as flat as possible. C1 and C14 are standard MKT capacitors, since as a matter of principle we try to avoid using electrolytic capacitors in the signal path.
R4, R5 and P1, in combination with decoupling capacitor C3 (R25, R26, P2 & C16), allow the output offset voltage of the amplifier to be adjusted to a minimum.

Modulator

The modulator amplifies the signal from the input stage to the output level. It is the second part of the overall amplification, or better said, the actual gain stage. The processor provides a switching waveform that depends on the level and frequency of the signal. With no input signal, the average value of the switching frequency is approximately 700 kHz. It can vary over a maximum range of 200 kHz to 1.5 MHz. Two complementary MOSFET drivers with

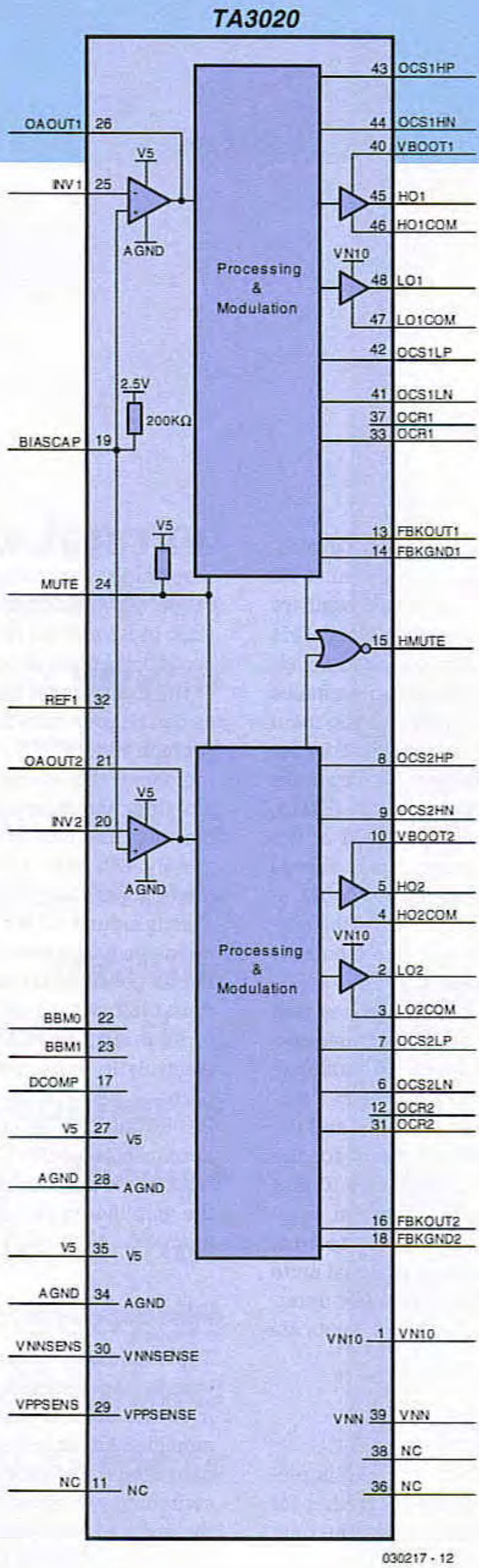


Figure 1. The internal structure of the Tripath TA3020.

level shifters convert this signal to the proper level for driving the gates of the MOSFETs. The supply voltage for the drivers (10 V) is provided via pin VN10. It is first decoupled as much as possible by C38 and C39. These capacitors must therefore be placed as close as possible to the associated pin of the IC. On the negative side, LO1COM (connected to the source of T2) and VN10 provide the supply connections for the driver. On the positive side, bootstrap capacitor C7 (C20) is charged via D5 (D12) to nearly 10 V when the output is negative, and it 'rides up' on top of the output voltage when T1 starts to conduct. This voltage is fed to VBOOT, which together with HO1COM (the source of T1 is also the bridge output) forms the other supply connection for this driver. At the clipping level, C8 (C21) provides an extra buffer, since the switching frequency is lower at the clipping level. R13 (R14) limits the charging current of C8 (C21) when the amplifier is switched on.

MOSFETs

Two n-channel MOSFETs (T1 & T2 or T3 & T4) form a half-bridge circuit for each channel. The level-shifter outputs alternately drive each of the MOSFETs into conduction. A 'dead time' (break-before-make) is maintained to ensure that the two MOSFETs can never both be conducting at the same time (no 'shoot-through' current). This time can be set using two jumpers (JP1 and JP2). We strongly recommend against experimenting with the selected setting. It might be possible to reduce the dead time if you use MOSFETs with considerably smaller gate capacitance (smaller amplifier power), but certainly not here! Gate resistors R78 and R9 (R28 & R30) limit the slew rate, and thus limit the amount of overshoot due to switching. They also somewhat reduce the amount of power that would otherwise be dissipated in the drivers (these are 1-W resistors) for charging and discharging the gate capacitances of the MOSFETs. D1 and D2 (D8 and D9) decrease the gate discharge time. This reduces the fall time of the pulses, and thus reduces the chance that both T1 and T2 may both be conducting at the same time. R8 and R10 (R29 & R31) are added for safety reasons. If no IC is fitted, they insure that the gates of the MOSFETs remain discharged. Without these resistors, leakage currents and noise voltages could occasionally have disastrous consequences. R6 and R11 (R27 and R32) are low-inductance resistors that are necessary for current limiting, which is described later on. R12 and C4 (R33 & C7) form a snubber network that eliminates high-frequency spikes in the output signal. This network is thus placed as close as possible to T2 (T4). Diodes D6 and D7 (D13 & D14) are connected between the source and drain of each of the transistors to suppress overshoots. Such overshoots are primarily caused by the coil in the output filter when large currents flow. These diodes (in SMD packages) are also placed as close as possible to the associated leads, primarily to protect the IC. D3 and D4 (D10 & D11) are additional diodes connected between the sources and drains (respectively) of the MOSFETS to suppress overshoots. All of these diodes (D1–D14) must be ultra-fast-recovery types. C5 and C6 (C18 & C19) decouple the half-bridge circuit and are especially included to suppress spikes on the supply voltage lines. This also has a beneficial effect on the operation of the MOSFETs. These capacitors must also be placed as close as possible to the leads of the MOSFETs. C6 (C19) must be an electrolytic capacitor with extremely low ESR and very good HF characteristics. Here you should not use a substitute for the prescribed type unless its specifications are just as good or better. A normal electrolytic capacitor would probably explode or have a very short life. The pulse-width modulated signal at the output of the half bridge is fed to the output terminals via the LC filter L1/C9 (L2/C22).

Output filter

Thanks to the high switching frequency, here it is only necessary to use a second-order filter with a relatively high corner frequency (resonant frequency 101 kHz). To dampen the Q factor of the filter, which is primarily important if no load is connected, a Zöbel network is placed at the output, since otherwise resonance currents and ringing signals at the output could reduce the reliability of the amplifier. As the corner frequency of the filter is higher than for conventional Class D amplifiers, the connected speaker system has a considerably smaller effect. In light of the large currents involved here, an ordinary noise-suppression choke cannot be used for the output filter. A special core material must be used to maintain low distortion and high efficiency. We have more to say on this subject in Part 2 of this article, which will appear in the September 2004 issue.

Amplifier configuration

The gain of the modulator is set using feedback resistor R15 (R36) and voltage divider R18/R20 (R39/R41). These components must be dimensioned according to the value of the supply voltage that is used. This is necessary to make the amplifier independent of the behaviour of the power supply (such as voltage fluctuations due to the output amplitude, mains voltage variations, etc.). Additional reverse feedback to counter 'ground bounce' is provided by R16 (R37) and voltage divider R17/R19 ((R38/R40). These two networks must be identical! The resistor values can be calculated quite easily. A value of 1 kΩ is generally used for R17 and R18, so the value of the other resistors only depend on the value of the supply voltage VPP (assuming a purely symmetric power supply) and the value selected for R17:

$$R19 = R17 \times VPP \div (VPP - 4)$$

This yields an E96 resistance value of 1.07 kΩ. This value is reasonably independent of the supply voltage. If a maximum supply voltage of 51 V is used, it only increases to 1.10 kΩ.
Finally, the value of R15 determines the gain of the modulator:

$$R15 = R17 \times (VPP \div 4)$$

We have assumed a maximum supply voltage of 62 V (the special decoupling electrolytics on the amplifier board are 63-V types). This yields a value of 15.4 kΩ for R15. The gain of the modulator can then be calculated in the same manner as for a standard non-inverting amplifier:

$$A_{modulator} = (R15 \div Rp) + 1$$

where R_p is the parallel resistance of R18 and R20.
Capacitors C11 and C24 filter and delay the feedback signal to the modulator. They have different values. These capacitors prevent RF noise with very high frequencies from penetrating the feedback network, and using different values causes the modulators to have different switching frequencies. This prevents mutual interference between the modulators. The values are chosen to cause the difference to be greater than 40 kHz.

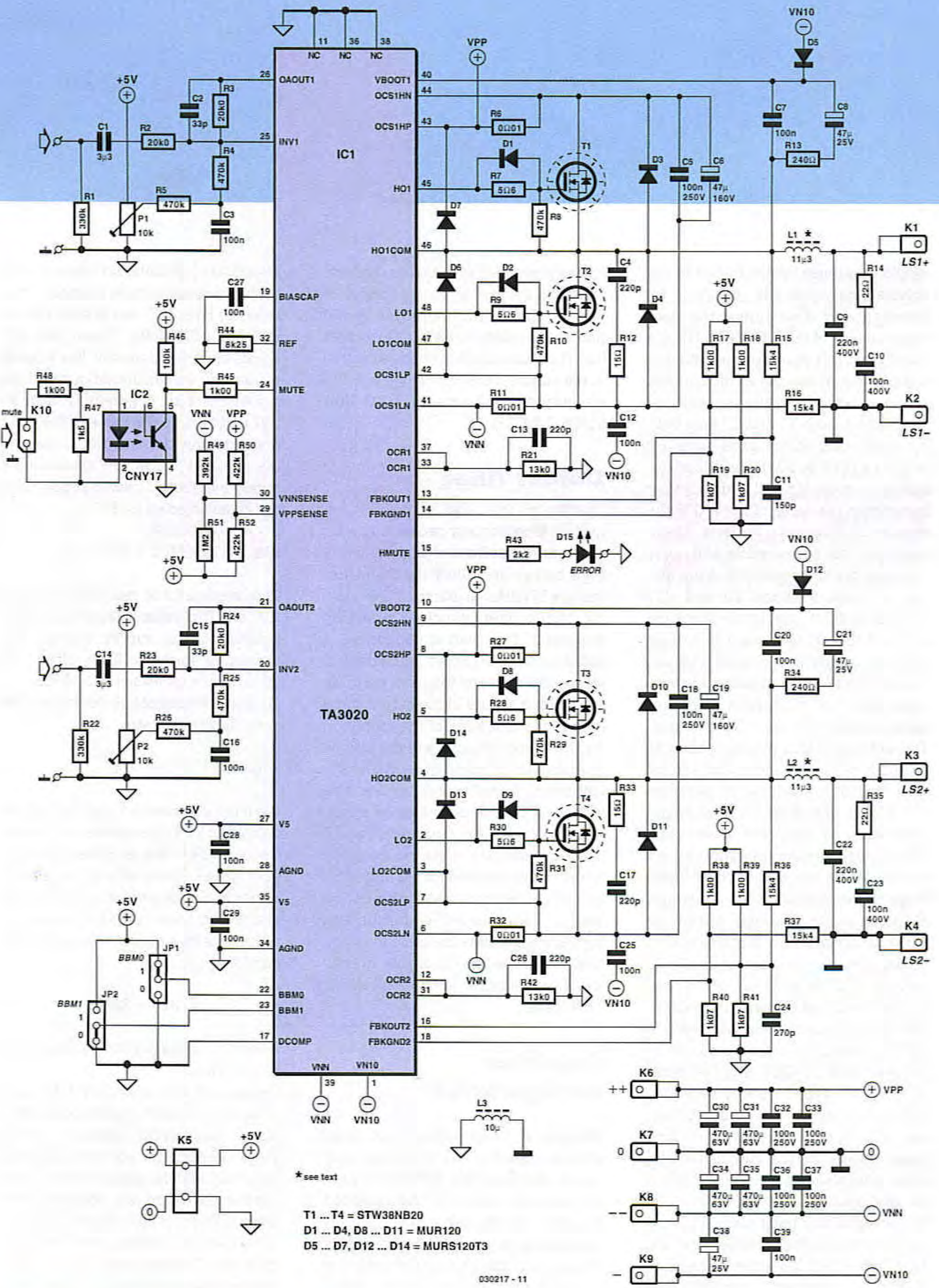


Figure 2. The circuitry around the amplifier IC.

Protection

To protect the amplifier, the driver IC monitors the supply voltages and the currents through the transistors. The VPPSENSE input is used to monitor the main positive supply voltage for overvoltage and undervoltage; the VNNSENSE input is used in the same way for the main negative voltage. If the supply voltage is outside the allowable limits, the output stage is disabled (mute mode). If the supply voltage returns to within the allowable limits, the output is again enabled. For the calculation of the associated component values, please refer to the data sheet. Theoretically, the amplifier could become 'stuck' in situation in which it constantly detects an overvoltage. However, that is very unlikely, since both detection inputs need more than roughly 68 V before they respond. This thus primarily amounts to protection for the IC itself, since several of the power supply capacitors are only rated for 63 V.

The calculations for overcurrent protection are certainly more interesting that those for voltage protection, since they determine the minimum load impedance the amplifier can handle at maximum output power. Since the output stage operates in switch mode, the MOSFETs used in the circuit determine the maximum load capacity of the amplifier. Here we have selected a relatively heavy-duty ST Microelectronics type, the STW38NB20. This transistor, which is housed in a TO-247 package, can handle up to 38 A and has a maximum drain–source voltage of 200 V. The maximum channel resistance with a gate–source voltage (U_{GS}) of 10 V is 0.065 Ω (I_D = 19 A). A disadvantage of MOSFETs with this sort of specifications is that their input capacitance (C_{iss}) is rather large, in this case as much as 3800 pF. That explains why the drivers in the IC must be able to deliver rather substantial currents in order to switch the MOSFETs sufficiently quickly. We primarily chose these transistors in order to reduce the risk of unpleasant surprises when using speaker systems with unknown impedances. Naturally, the break-before-make time could be made shorter if transistors with significantly smaller gate capacitance are used, which would reduce the distortion level. However, our choice was in favour of a design that can tolerate low impedances.

Overcurrent detection is provided by the two low-inductance resistors R6 and R11 (R27 & R32), which are connected in series with the transistors as sense resistors. R6 is used for positive half-cycles in series with the drain of T1, while R11 is used for negative half-cycles in series with the source of T2. The response threshold of the protection circuit is set in combination with R21. The IC directly measures the voltages across the sense resistors and uses these voltages to generate a current through R21. The maximum output is determined by comparing the voltage across R21 with the overcurrent threshold voltage V_{TOC}. C13 (C36) filters the voltage from the rectifier. The relationships between these components are given by the following two equations:

$$I_{max} = 3580 \times (VTOC - (Ibias \times R21)) \div (R21 \times R6)$$

$$R21 = (3580 \times VTOC) \div (Imax \times R6 + 3580 \times I_{bias})$$

Here VTOC is the threshold voltage for overcurrent detection (typically 0.97 V) and I_{bias} is 20 µA.

The first equation can easily be rearranged to allow the component values to be calculated. The second equation can be used to determine the value of R21 (R42). We have chosen a maximum output current of nearly 20 A, so that a load of somewhat less than 3 Ω just avoids triggering the mute mode.

The mute mode can only be reset by briefly switching the level at the Mute input or briefly switching off the amplifier. When the mute mode is active, the HMUTE output is High, and this signal drives a LED that can be fitted to the front panel if desired. A red high-efficiency LED should be used for this purpose, since reducing the value of R43 would overload the output.

Power supply

The supply voltages for the amplifier board are provided by a second printed circuit board. This board includes, among other things, the +5-V and VN10 supplies, as well as fuses for the main supply voltages. It also supplies a delayed 'un-mute' signal that prevents switch-on 'plopping'. To avoid creating an earth loop and prevent ripple currents from flowing though the input stage ground path, the mute signal is fed to the IC via an optocoupler. It is located on the amplifier board. The input of the optocoupler is thus fully isolated from the amplifier, but an active signal is required to switch the amplifier Mute input.

The main supply voltages (VPP and VNN) for the TA3020 are decoupled as well as possible using special electrolytic capacitors (C30, C31, C34 and C35) and MKT capacitors (C32, C33, C36 and C37). A simple decoupling network is used for the 5-V supply voltage for the input amplifiers.

To suppress possible interference from the output circuit as well as possible, analogue ground and modulator ground (which is also the ground for the rest of the circuit) are kept separate and coupled on the solder side of the board at a single location using an SMD inductor.

Layout

As already mentioned at the beginning of the article, the amplifier relies on a carefully designed layout. The layout thus forms an essential part of the overall amplifier. Tripath emphatically recommends copying the reference layout, since otherwise the large high-frequency currents could give rise to unexpected effects. Naturally, some of the components we have selected differ from those used on the reference board, primarily with regard to their dimensions. This is because we have given special attention to the availability of the components (preferably in single quantities). Some of the tracks have been shifted slightly in some places, and a few components have been added, but by and large we still succeeded in maintaining the recommended layout. If you take the trouble to look at the photo of the reference board in the data sheet, you will see the resemblance to the photos of the prototype. For some of the components shown in the schematic that are fitted on the solder side of the board in the Tripath layout, we have put them on the component side instead. This is why the circuit board is placed parallel to the heat sink in our version, with the transistors mounted below the board. This produces an attractively compact and robust module, but we'll save further comments for the construction description in the second part of the article. What we can tell you already is that although the module looks very simple at first glance, on closer examination you will notice that compared with the schematic, a few things seem to be missing. Many of the components are SMD types, and they are fitted on the solder side of the board. This yields the lowest likelihood of interference problems and results in an amplifier board with very modest dimensions for a 2 ¥ 300 W amplifier. Most of these SMD devices come in 0603 'shapes', which in all honesty are nasty little things to work with. To make things easier for you, we will try to supply the circuit board in the near future with the SMD components already fitted.

In the second part of this article (September 2004 issue), we give detailed attention to the construction of this unusual amplifier.

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Web pointers

TA3020 data sheets & application note:
www.tripath.com/downloads/TA3020.pdf

TA3020 reference board --
www.tripath.com/downloads/RB-TA3020.pdf