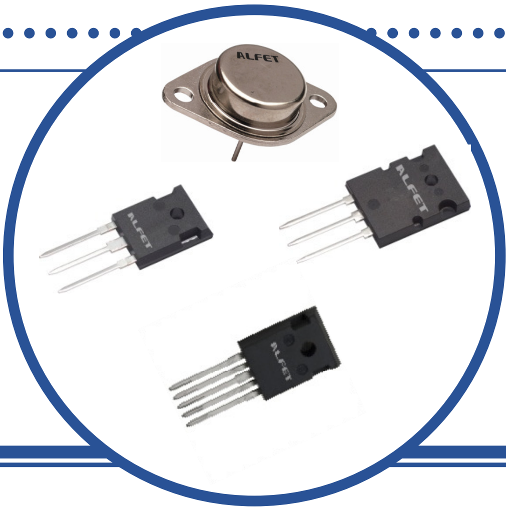
APPLICATION NOTE

LATERAL MOSFET DESIGN RECOMMENDATIONS FOR AUDIO AMPLIFIERS



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Introduction

Semelab's range of lateral mosfets has been specifically designed for audio power amplifiers and there are a number of advantages that they exhibit over bipolar transistors in these applications. This application note is intended to highlight these differences and give design guidelines in using the mosfets for high power audio products.

Semelab's new range of Alfet devices are available in both single (8Amp) and double-die (16Amp) versions with complementary N-channel and P-channel parts rated at both 160V and 200V. Table 1 outlines the currently available parts.

Table 1

Part Number	Pol	Voltage	Current	Package
ALF08N16V/ALF08P16V	N/P	160V	8A	TO-247
ALF08N16K/ALF08P16K	N/P	160V	8A	TO-3
ALF08N20V/ALF08P20V	N/P	200V	8A	TO-247
ALF08N20K/ALF08P20K	N/P	200V	8A	TO-3
ALF16N16W/ALF16P16W	N/P	160V	16A	TO-264
ALF16N16K/ALF16P16K	N/P	160V	16A	TO-3
ALF16N20W/ALF16P20W	N/P	200V	16A	TO-264
ALF16N20K/ALF16P20K	N/P	200V	16A	TO-3
ALF08NP16V5	N&P	160V	8A	TO-247-5
ALF08NP20V5	N&P	200V	8A	TO-247-5

Double-die devices incorporate 2 parallel connected die that are taken from adjacent positions on the silicon wafer and this guarantees a very high degree of matching of characteristics such that the 2 pieces of silicon behave as a single device with twice the current capability. Other highlights in this range include a new N and P-channel complementary part in a 5-pin power package that has a symmetrical pinout. This device allows a 100W-150W amplifier to be achieved with a single output device resulting in a very small footprint that can prove beneficial in applications requiring multi-channel output in a small box. Figure 1 shows this new 5-pin version of the TO-247 package.

As can be seen in the associated internal diagram, the pinout is totally symmetrical allowing an optimised pcb layout to be achieved.

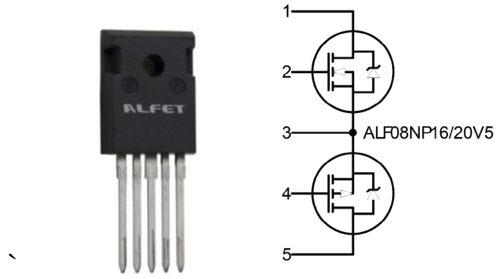

Figure 1 – Complimentary N & P channel device

Lateral Mosfet Advantages

The key advantages of lateral mosfets over bipolar junction transistors (BJTs) and vertical mosfets are:

- Complete absence of secondary breakdown when compared to BJTs. This results in very rugged performance and simplified protection.
- Self limiting current characteristic. No matter how much drive is given to a lateral mosfets they will get to a point where they will not deliver further current. This factor, contributed with the previous advantage, gives the devices their bullet-proof reliability.
- In rare cases that a device does fail the resulting damage in the amplifier is usually limited to the output stage and does not tend to cause the typical chain reaction back through the amplifier that is often seen in bipolar designs.
- Simple and stable bias because of the low current point where the temperature characteristics cross from positive to negative coefficient. This is an advantage when compared to both vertical mosfets and bipolars that many people are not fully aware of!
- No charge storage effects like bipolar transistors meaning that they are capable of higher frequency operation and do not exhibit crossover distortion switching at higher frequencies.
- No beta droop characteristic of gain at higher currents that BJTs exhibit. This contributes to consistent distortion performance across wide ranges of load impedance.
- Easy to drive compared to BJTs due to the high impedance gate. This not only simplifies the drive stages of the amplifier that are required (smaller devices and lower cost) but also results in the voltage amplifier stage generating less distortion.

- Integral anti-parallel protection diode. The intrinsic mosfet parallel body diode means that an external diode does not need to be added as in BJT amplifiers.

This application note will look at some of these advantages in a lot more detail as well as outlining some good practices to use when designer audio power amplifiers with lateral mosfets.

Biasing Considerations

It is commonly known and quoted that bipolar transistors have a positive temperature coefficient (ptc) and that mosfets have a negative temperature coefficient (ntc) in their relationship of drain current to temperature. Whilst this ntc characteristic of mosfets is certainly true, it does, however, only occur above a particular operating drain current. For lateral mosfets from Semelab this current level is around 100mA, which it just so happens is the ideal point for optimum biasing of a class-AB audio amplifier! Figure 2 shows a typical transfer characteristic for the ALF08N20V and demonstrates a thermal coefficient crossover point at around 110mA. At this point the thermal coefficient is in fact zero and therefore completely stable with fluctuations in temperature as the output stage warms up at turn on or is hot after high power usage.

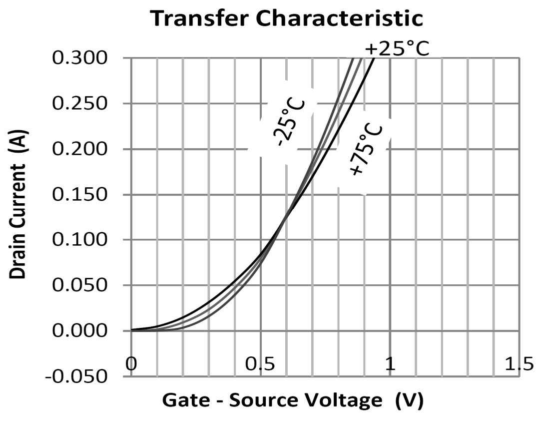

Figure 2 – ALF08N20 Transfer Characteristics around temperature coefficient crossover point

Although this is the optimum bias point for stable thermal operation some high power designs that need to minimise the power dissipated under quiescent bias conditions often run the devices at a lower current. Although there will be a small shift of the operating point with temperature it is still quite small and acceptable and 50mA will work well. Taking the bias too low will eventually result in crossover distortion (most readily observed with a 20kHz audio signal into load at high amplitude) and therefore the designer needs to make tradeoffs based on the requirements of the design.

This is one of the key characteristics that make lateral mosfets so well suited to class-AB audio amplifier applications. Stable bias against temperature results in a number of significant advantages:

- Complete absence of thermal runaway risk
- Simple bias circuitry. An adjustable or fixed value resistor is adequate and no thermal coupling between bias devices and the heatsink is required. Transistor based diode multipliers with thermal coupling are not needed.
- No requirement for source resistors for single device or parallel device operation (more on this in a later section)
- Devices cool faster and return to a stable temperature faster after high power use
- Much less critical bias point setting than bipolar amplifiers that does not drift with temperature allowing easier setup and consistency in production

There are other mosfets on the market aimed at audio applications that are not lateral types but instead of a vertical structure. These do not exhibit the same thermal characteristics and a transfer curve for a typical device in shown in Figure 3.

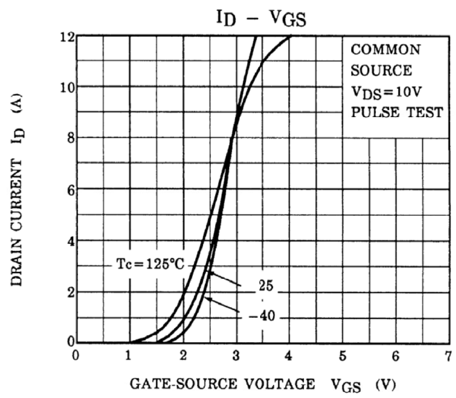

Figure 3 – Transfer characteristic of typical vertical mosfet

As can be seen from this curve, the temperature crossover point does not occur until the drain current is at around 8A! Therefore, for operation at bias levels of 100mA, source resistors and thermally coupled transistor bias circuits like those used for bipolar amplifiers are necessary for stable thermal operation to control runaway and for paralleled applications. Switching mosfets also used in some amplifier designs also exhibit similar characteristics (with thermal crossovers often at even higher currents) and similar steps in the design to provide stable thermal operation are also required. As a result lateral mosfets amplifiers are far simpler to set-up with reduced component requirements for stable bias operation that results in a smaller footprint and the eradication of these source resistors. Simplified bias circuitry and reduced pcb area mean a cost reduction in both materials and labour that should be considered in the total solution cost comparison. This exceptional thermal stability also enhances reliability.

Figure 4 shows a typical lateral mosfet output stage and the recommended bias arrangement. The values of RBF and RBA are dependent on the rest of the design and the current that is present in the preceding stage.

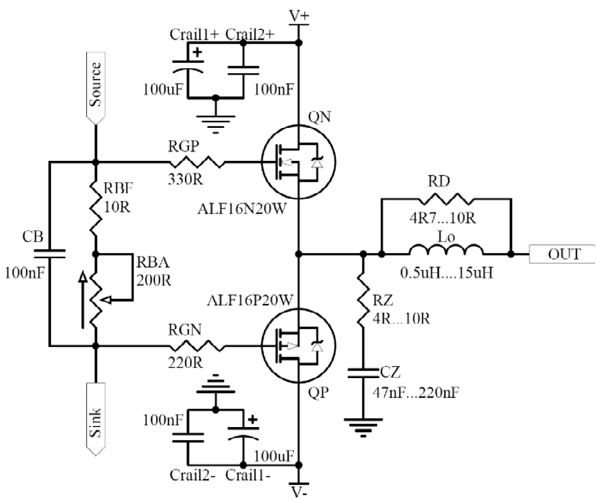

Figure 4 – Typical Lateral Mosfet Output Stage

As a guideline to the nominal value of this resistance the following equation can be used:

$$R_{BF} + R_{BA} = \frac{V_{bias(total)}}{I_{source}} \ (\Omega)$$

Where: $V_{bias(total)}$ = the required bias voltage of the N & P channel mosfets added together (typically 2V) and I_{source} = the driver stage quiescent current. For example, if the driver stage has 20mA quiescent current then the total bias resistance should have a central, nominal value of 100Ω.

The Clipping Mechanism in Mosfet Amplifiers

The output voltage swing capability and resulting clip point is a result of different parameters than the $V_{ce(sat)}$ induced clipping that occurs in bipolar power amplifiers. There are 2 potentially contributing factors with a mosfet:

- $R_{ds(on)}$ induced clipping. When the device is fully turned on it behaves as a resistor with an approximate value of 1 Ω per die. Therefore 6A of current per die would mean that it exhibits a saturation voltage of 6V and the amp would only swing within 6V of the power supply rail.
- Gate voltage (V_{gs}) induced clipping. To be able to deliver a give drain current the mosfet will need a corresponding value of gate-source voltage. For the 6A example, approximately 5V would be required for a single die device at 25°C (see figure 5).

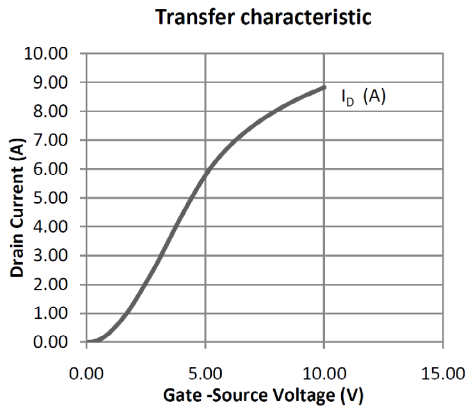

Figure 5 – ALFET transconductance characteristic

Which of these mechanisms happens first depends on a number of factors in the overall design of the amplifier circuit. In the typical output stage shown in figure 6 the driver stage can swing to within about 1V of the rail (the driver stage bias current plus driver transistor saturation). This means that the voltage resulting from the $R_{ds(on)}$ and the voltage required to drive the mosfet are around the same so the clip point would be about 6V from the power rail.

It should also be noted that both these parameters, the $R_{ds(on)}$ and the required V_{gs} to deliver a given current, will increase with temperature. This means that the onset of clipping will happen earlier at higher operating temperatures. These mechanisms form part of the action of self protection in the devices where a point will be reached on a given heatsink where the amplifier will not deliver more than a certain amount of power.

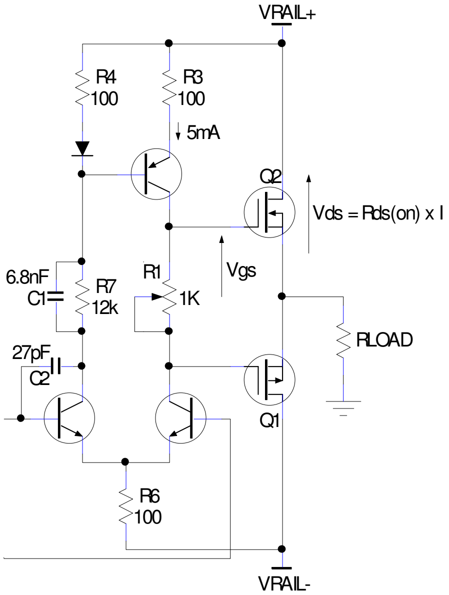

Figure 6 – Typical Output Stage

Elevated Power Rails

Because of this voltage driven characteristic and the fact that it can become the controlling factor in output voltage swing, some amplifier designs utilise higher voltage driver stage rails about 10V high than the main power rails. This is to ensure that there is plenty of available gate drive to take the output devices all the way to their R_{dsON} limit and use all the available voltage swing of the power supply rails. Ultimately the limiting factor on the output voltage swing will need to be determined for a given design but Semelab's Amplifier Power Calculator spreadsheet will allow simulation of various scenarios to determine what limits the output swing. It is often stated that mosfets *must* have a higher voltage drive rail to utilise the full output capability but this is not the case with laterals that have a relatively high R_{dsON} and will only become important in designs with lots of parallel devices (resulting in low total combined resistance) and topologies which do not allow the drivers to swing close to the power rails.

Lateral Mosfet Linearity

It is often stated that mosfets are not as linear as bipolar transistors. In simple terms this statement is true and mostly can be attributed to the lower gain exhibited by the devices (in terms of their transconductance). However, as is often the case with comparing different technologies, it is not quite as clear cut as this!

Yes, the effective gain is lower resulting in a higher apparent distortion figure of a source follower compared to a bipolar emitter follower in an open loop configuration but there are some other factors that come into effect where the mosfet has advantages:

- Mosfets do not exhibit a roll-off of gain at higher currents beta drop) as bipolars do. Figure 7 shows a graph of the transconductance of a single die lateral mosfet where this can be clearly seen. As a result the amplifier distortion specification is very similar into different impedances whereas a bipolar amplifier will always see a significant distortion increase into lower impedance loads. This make lateral mosfets especially suitable for amplifiers designed to drive 2 & 4Ω loads.

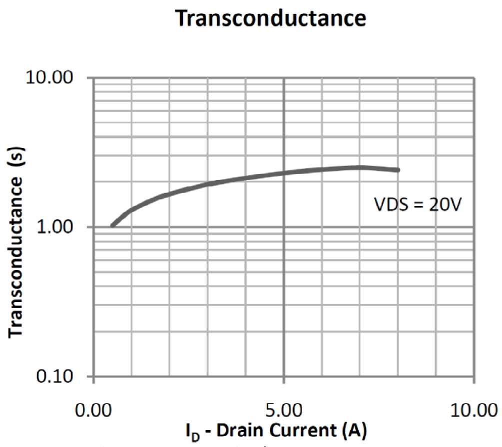

Figure 7 – Typical Output Stage

- Because of the stability of the bias point with temperature and less critical setting, lateral mosfets typically exhibit better crossover distortion characteristics. Bipolars can exhibit excellent specifications in this respect but because of the difficulties of stabilising this against temperature, the optimum bias point is usually only there under quiescent conditions.
- As a result of the mosfet's high bandwidth capabilities both high open loop gain and wide open loop bandwidth are achievable. This not only makes a mosfet amplifier easier to compensate but means that the high open loop bandwidth typically achievable when compared to bipolar amplifiers means that the resulting closed-loop distortion figure can be similar.
- The absence of stored charge that occurs in BJTs means that they do not exhibit any switching distortion that means high frequency crossover distortion is lower.

Heatsink Design Considerations

The 'on' resistance of mosfets increases with temperature and although this is part of the mechanism that allows them to be very rugged in high power audio applications it will also decrease the obtainable output power at elevated junction temperatures. Lateral mosfets will tend to self limit at high temperatures through this phenomenon which is beneficial from a protection point of view but good thermal design is required to ensure that the required output power can be delivered with an optimal number of paralleled mosfets.

The absence of source resistors creates some additional extra benefits for designers when combined with another factor that is specific to lateral mosfets. Vertical structure mosfets' silicon die back connection is the drain and it is that part of the silicon that is electrically bonded to the metal back of the package. This means that the power devices back metal is at opposite rail polarities for the N and P devices and that if a common heatsink is used the devices must be electrically isolated from it. Live heatsinks can be used but a separate heatsink for positive and negative sections of the amplifier are required. This is an identical situation to amplifiers using power bipolar transistors.

With lateral mosfets, however, the back of the die is the source connection and therefore that back metal of the package is also electrically connected to the source. This factor combined with the fact that source resistors are not required for stable biasing and sharing with paralleled devices means that all the sources of the devices directly connect together and therefore the metal backs can as well. This is significant and means that a single live internal heatsink can be used in the product (one for each amplifier channel) that is at the potential of the amplifier output. By doing this the isolating pad of mica or silicon material can be removed and thermal resistance is considerable lowered resulting in both potentially smaller heatsinks and improved reliability. The speed of response of die temperature increase to a high power transient is significantly improved without an electrically isolating pad and many existing lateral mosfets customers exploit this benefit of features within their designs. There is also another potential benefit in this technique in that the individual source connections do not need to be routed on the pcb and can instead be taken from a single point on the heatsink allowing for simplified pcb routing.

The maximum junction temperature of the mosfet silicon is rated at 150°C. Therefore, the heatsink size and any fan cooling must be calculated to keep the junctions temperatures inside this rating. To properly calculate the required heatsink and indeed how many devices should be selected a number of design parameters need to be known for the results to be valid in the real world and therefore ensure reliable operation. The key parameter is in understanding the worst case power dissipation of each output device. This is not always easy to calculate on a theoretical basis and experienced designers tend to use rules of thumb they have gained over the years A number of factors will effect this figure.

- The regulation of the power supply rails. Depending on the size of smoothing capacitors and rating of the mains transformer, the rails will move up and down with power output.
- The nature of the load reactance. An inductive load (as is the case with a loudspeaker) will always result in increased dissipation over that calculated for a purely resistive load due to phase shift. In addition a drivers impedance rating will always be lower at low frequency as the inductive reactance decreases where the impedance will become the dc resistance of the voice coil. Many commercial loudspeaker cabinets will also present dips in the impedance through the crossover points and these factors should also be included in calculations and safety margins.
- It should always be remembered in these calculations that the maximum output device power dissipation does not occur at maximum power output!
- Remember to include the bias current induced power dissipation. This figure is always in addition to that dissipated by delivering power into the load. ±50V power rails with 50mA of current results in 2.5W per output device at idle.

If we look at an example of a class-AB amplifier designed to deliver 150Wrms output power into a 4Ω resistive load with a bias current of 50mA and a power supply regulation of 6% (typical for a toroidal supply), we can plot the power dissipated against output power:

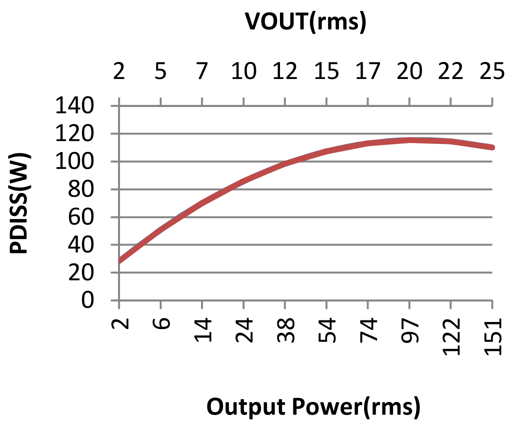

Figure 8 – Output stage total dissipated power plotted against output power for 150W rms into 4Ω

The P_{DISS} is the total output stage dissipation and includes the bias power. For the single pair of output devices this was simulated for the dissipation in each one is half the figure on this graph. This theoretical plot shows us that we would have a worst case power dissipation of 115W (57.5W per device) that occurs around 100W output. OK, now we understand the behaviour into a purely resistive load, what will happen if we hook up a real loudspeaker that includes some inductance? As an example if we run the same simulation with a typical loudspeaker (2 x 8Ω units in parallel), we can produce the following results at 50Hz input:

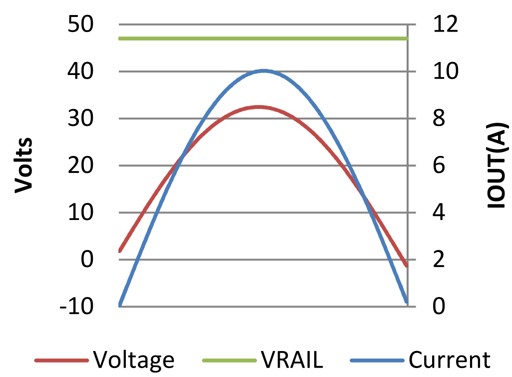

Figure 9 – Output stage Voltage and current at 50Hz

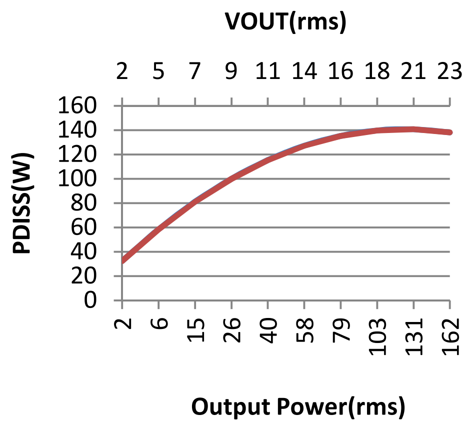

Figure 10 – Power dissipated at 50Hz

The same setup again but at 1000Hz input:

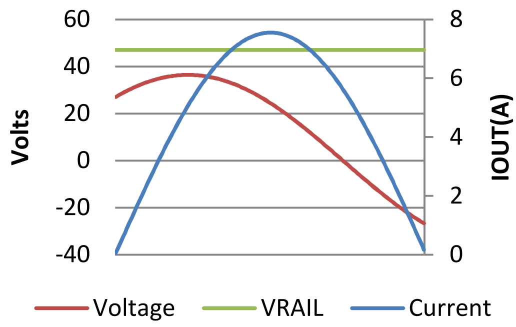

Figure 11 – Output stage Voltage and current at 1kHz

The phase shifted relationship between the output stage voltage and current can clearly be observed and although this results in lower power output than the resistive load, the power dissipation is higher.

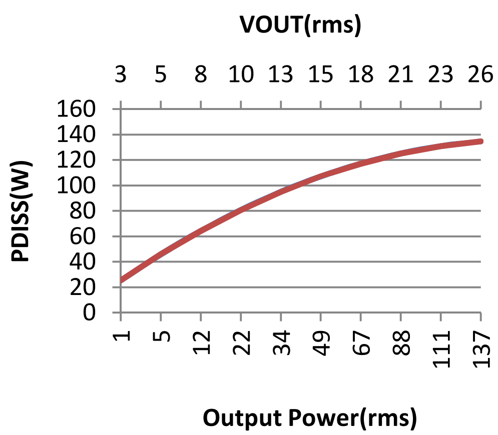

Figure 12 – Power dissipated at 1000Hz

Please note that the maximum output power achieved in each case varies as the simulation calculates the maximum voltage swing that will be achievable using real Alfet characteristics on the same power rails. Therefore, the conditions differ in each case compared to the nominal 150W rms output with a resistive load. It can be seen that at both low and mid frequencies the power dissipation is higher than for the resistive load for this nominal 4Ω rating. This drive unit had a 6.5Ω dc resistance and 1.15mH of inductance but clearly this situation just becomes worse with increasing inductance. These examples have demonstrated the need to weigh up these different considerations in calculating the heatsink size and number of devices.

In most commercial amplifier designs aimed at music reproduction most engineers make some assumptions about the nature of the signal source. This assumption is based on the fact that music does not represent a continuous full power sine wave and that through both the peak to average ratio and duty cycle of typical music the resulting average power (and therefore heating effect) of the signal is lower. Traditionally this figure was assumed to be 1/8th rated power output and long term thermal tests were measured at this level. There is, however, a strong argument with modern music with high bass content to use a higher figure of average power.

Therefore, let's assume we want to rate our heatsinks such that the junction temperature of the mosfets does not go above 150°C with continuous operation at ¼ rated power output (in this case 37.5W. If we look at our 3 previous examples we can see the power dissipation at this power point is:

Condition	Dissipation at 37.5W output power
Resistive load	98W (49W per device)
Typical speaker at 50Hz	112W (56W per device)
Typical speaker at 1000Hz	99W (49.5W per device)

Table 2 – Power dissipated at 1000Hz

So assuming we want to take this worst case 112W figure for continuous average quarter power output dissipation we can now work out our required heatsink thermal resistance.

Figure 13 shows the thermal parameters of a heatsink mounted power semiconductor and the 3 thermal resistances that make up the total coupling from the silicon junction to the ambient air: θ_{jc} – junction to case, θ_{cs} – case to sink and θ_{sa} – sink to air. The classic thermal equation used in heatsinks is:

$$\theta_{ja} = \theta_{jc} + \theta_{cs} + \theta_{sa} = \frac{(Tjmax - Ta)}{P_{DISS}} \ (^{o}C/W)$$

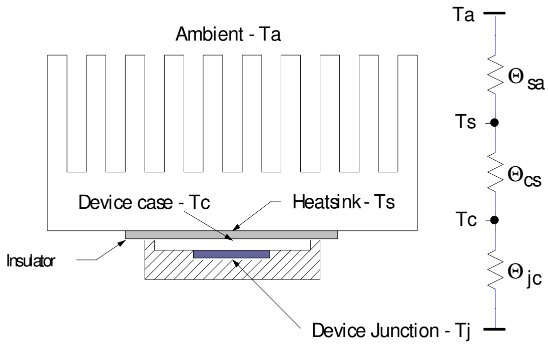

Figure 13 – Semiconductor mounting thermal resistances

The θ_{jc} of the mosfets is 1 °C/W and we have 2 so the total result is 0.5 °C/W. If we assume a heatsink insulator (mica or silpad) has a θ_{cs} of 0.4 °C/W, again we have 2 so their total is 0.2 °C/W. We want our T_{jmax} to be 150°C, our P_{DISS} is 112W and we will assume the ambient temperature, T_a to be 25°C (although for heatsinks mounted internally and allowing for operations in hot countries, 50°C may be a more realistic figure). Therefore:

$$0.5 + 0.2 + \theta_{sa} = \frac{(150 - 25)}{112}$$

$$0.5 + 0.2 + \theta_{sa} = \frac{(150 - 25)}{112} - 0.5 - 0.2 = 0.42 \ ^{o}C/W$$

So we can now select a suitable heatsink or sink/fan combination. Interestingly, if we take the approach outlined previously of running a heatsink live and we assume a thermal mounting, θ_{cs}, of 0.05 °C/W then the required heatsink thermal resistance would be 0.57 °C/W – a significant difference. Alternatively, using the previously calculated 0.42 °C/W heatsink without device insulators this would yield a junction temperature of:

$$T_j = P_{DISS}(\theta_{jc} + \theta_{cs} + \theta_{sa}) + T_a = 133.6 \ (^{o}C)$$

Taking into account that for approximately every 10°C rise in temperature that silicon lifetime is halved, this is a significant improvement in junction temperature and potential long term reliability.

Often it is difficult with fan cooled heatsinks and fabricated designs to truly know the figure of heatsink thermal resistance. As a result it can be useful to work backwards on a prototype design to verify the junction temperature under a given set of conditions. Using our previous example, we know that if we set our amplifier up with 37.5W output then the output stage is dissipating 98W. Under these conditions, if we measure the heatsink temperature close to the output device we can approximate the junction temperature and heatsink thermal resistance:

$$T_j = P_{DISS}(\theta_{jc} + \theta_{cs}) + T_s \ (^{o}C)$$

So if the heatsink measured 70 ^{o}C

$$T_j = 112(0.5 + 0.2) + 100 = 148.4 \ (^{o}C)$$

And the θ_{sa} in this example would be:

$$\theta_{sa} = \frac{(T_s - Ta)}{P_{DISS}} = \frac{(70 - 25)}{112} = 0.402 \ ^{o}C/W$$

This is a useful technique to validate thermal resistance and junction temperature in a real design.

Effects of Increasing Number of Output Devices

Clearly it can also be stated that increasing the number of output devices will also result in lower junction temperatures at a given power or a smaller heatsink. In our example, doubling up on the output stage devices (i.e. 2 pairs of output devices rather than one) will half the θ_{jc} in our equations to 0.25 and the θ_{cs} to 0.1. In addition to this the equivalent total $R_{ds(on)}$ of the mosfets will half (and the required gate drive will reduce) so on the same power supply rails the amplifier will be able to swing more output voltage. In our previous ±50V rail example, adding an extra pair of output devices results in the amplifier being able to deliver 200Wrms output power – a significant increase. With the same 0.42 ^{o}C/W heatsink the mosfet junction temperatures would now be around 122 ^{o}C, even allowing for the increased dissipation that results through the increased rail swing and currents to deliver the higher 200W power.

Therefore, there is always a trade-off in the decisions a designer makes against amplifier efficiency, heatsink size, junction temperature and the costs associated with the various options.

Parallel Operation Considerations

For higher power operation adding extra output devices in parallel is an easy approach to scaling power. One of the advantages of lateral mosfets is the ease of parallel operation. However, there are a number of points that should be considered to ensure both optimum performance and reliable operation:

- Current Sharing. We obviously want devices to share current as equally as possible. The best approach to achieving this is to make sure that they share under bias conditions. Therefore, we ideally want parts that have the $V_{gs(th)}$ characteristic matched. Again, the best approach to this is to make sure that the V_{gs} of the devices are as close as possible at the amplifiers intended bias current. Therefore, if the bias is 50mA per device, then match them at that current. Luckily the turn-on characteristic of lateral mosfets is quite soft at these operating points and it is quite easy to find devices which match sufficiently well. In addition, because of the ntc characteristic of the devices above typical bias points no single device can runaway with the current and they will balance themselves to some extent. However, we need to make sure that none of the devices are off and not contributing at bias.
- This threshold does not need to be a set voltage, it just needs to be matched in the set of devices in a particular amplifier. In addition, it is not important that N-channel devices are matched to Ps but rather that all Ns are matched to each other and all Ps are matched to each other. Many customers find that matching of 10-20% is quite sufficient for reliable operation because at higher currents the devices will very much balance each other.
- As devices are added in parallel the capacitance presented to the driver stage increases. Therefore, this driver stage needs to be able to deliver sufficient current to operate at the required bandwidth and slew rate of the amplifiers target specification.
- Semelab can supply devices pre-matched in sets and colour coded to threshold value to a customer's specification. Please enquire about this service if it is required.

Thermal Protection Considerations

When used at their limits the mosfets will get hot very fast and this will result in an increase in R_{dsON} as well as a fall off in transconductance so a point will be reached where it is difficult to get further power from the devices and a self limiting effect will happen. Although this characteristic contributes to the ruggedness of the devices the junctions will, however, get very hot under these conditions and potentially exceed their maximum rating. Therefore it is always recommend to include thermal protection on the heatsink when relying on protection by this mechanism.

The example in the previous section demonstrates that if the average power were any high than 112W or if the product was exposed to higher ambient temperatures or any cooling fans ceased to function that the junction temperature would exceed its 150°C rating and long term reliability would be compromised. Therefore, the techniques outlined previously can be used to determine the junction temperature where this occurs and set the thermal trip at this point. This ensures the amp will both run continuously at its intended power level and be protected under adverse operating conditions for any period of time.

Many power amplifier failures can be attributed to these high temperatures that occur in designs with insufficient heatsinking or thermal protection set at the wrong point.

Effects of Power Rail Variations

In these previous calculations we assumed the power rails were constant but dipped under load. In reality, most commercial amplifiers do not have regulated power supply rails and in some countries the mains may be as much as 10% higher than its nominal specification. This increase will affect the dissipation in the output stage but if the calculations have already been done on temperature to protect the amplifier at a given heatsink temperature that limits the junction then this will continue to be an effective safety measure. It may however, trip out earlier when running ¼ power tests because of the elevated voltage. A later section in this paper discusses the other implications of power rails in terms of their effect on device voltage withstand capability.

Gate Drive Power Requirements

Although a lateral mosfet does not require gate current to operate it does require current to charge and discharge the input capacitance (C_{iss} is the lumped combination of C_{gs} & C_{gd}). This capacitance does vary with gate drive voltage and drain-source voltage but the datasheet parameter does indicate a worst-case real value to use.

In addition the p-channel parts use a larger die size than the n-channel in order to match the on resistance ratings and therefore exhibit a larger capacitance. As a result, to ensure we have sufficient current available and that the driver stage is scaled to deliver this, it is wise to use this higher figure. Therefore if we assume around 700pF per single die device, we will need a peak current of:

$$I_{g(pk)} = 2\pi C_{iss} V_{gs} f$$

This equation assumes delivering enough current to reproduce a sine wave of frequency f without generating slewing distortion rather than simply charging the capacitance in a time of $1 / f$. Therefore, if we want to drive the gate to 12V and operate to a frequency of 100kHz we can calculate the required gate drive current:

$$I_{g(pk)} = 2\pi(700 \times 10^{-12}) \cdot 12 \cdot (100 \times 10^{3})$$

$$I_{g(pk)} = 2\pi(700 \times 10^{-12}) \cdot 12 \cdot (100 \times 10^{3})$$

$$I_{g(pk)} \approx 5.3mA$$

The driver stage needs to be capable of delivering both the required peak current and resulting power dissipation that is required for a given bandwidth capability. The details of the implementation of this are beyond the scope of this application note as particularly the power dissipation will vary depending on the actual topology of the amplifier design.

If the amplifier is designed to meet these figures and then operated at higher frequencies then both distortion of the sine wave and roll-off of output are likely to happen. When driving multiple devices in parallel the current of the driver stage will increase as the loading capacitance it sees is increased. However, a point to note is that in a given design, simply adding an extra pair of output devices with a given power rail will not significantly change the drive requirements. This is because when an extra stage is added the resulting total mosfet transconductance increases and the required gate-source voltage to deliver the current will decrease and the demands of the driver stage decrease. Also note that these currents are peak required currents and the average power dissipated with music will be far lower than when assuming these are continuous (although this may be the case with class-A driver stages).

Another way of looking at the date drive requirements is from the point of slew-rate. Assuming that the previous amplifier stages can all slew to the required rate, then we can ensure that the output stage can also slew at the required rate. For example, if we used our previous example with 50V rails and wanted the amplifier to slew from 0V-45V in 1uS and our maximum gate drive was to be 12V then our required current would be:

$$I_g = \frac{CV}{t} = \frac{(700 \cdot 10^{-12}) \cdot 12}{1 \cdot 10^{-6}} = 8.4mA$$

Gate Resistor Recommendations

Because of the extremely high bandwidth capability of lateral mosfets it is recommended to place a series resistors in the gate to each output mosfet to control this bandwidth and minimise any tendency for the amplifier to oscillate. Figure 4 shows the position of these resistors in the circuit and they should be physically placed as close as possible to the gate pin of the device on the pcb. If multiple, paralleled output devices are used then a separate gate resistor should be placed in the design for each transistor. The frequency of the pole of the roll-off that this resistor adds to the forward transfer function of the amplifier can be calculated using:

$$f_{-3dB} = \frac{1}{2\pi C_{iss} R_{gate}} \ (Hz)$$

Because P-channel parts have a high capacitance than N-channel, it is also advised to calculate the values for each separately. This addresses two potential issues; firstly it ensures that the amplifier frequency response is symmetrical which can help to alleviate possible oscillation issues and secondly in cases where high gate resistor values are used it alleviates any risk of the devices cross conducting. Although lateral mosfets do not exhibit any stored charge phenomena in the way that bipolar transistors do they are never the less capable of very high speed and unmatched gate timings could therefore result in both conducting at the same time. This time would nevertheless be very short and would not destroy the mosfet but could cause higher running temperatures that can be avoided by matching the timings. For N and P parts, C_{iss} values of 500pF and 730pF are typical and with a 330Ω/220Ω gate resistor this results in 3dB points at 965kHz and 1.01MHz respectively. It may be desirable to have different values on the N and P sides to equalise this frequency. Typically values of 150 to 470Ω are used in designs but some experimentation may be required to optimise the performance in a given topology and pcb layout and these resistors should be placed as near to the mosfet gate pin as possible to minimise the inductance directly at the gate and minimise any chance of oscillation occurring.

Gate Protection & Safe Operating Area Limiting

Semelab's previous family of lateral mosfets included internal gate-source zener diodes to protect the junction against overvoltage situations. On the new family of Alfets these zeners have deliberately been omitted for a number of reasons to enhance the performance and as a result some protection does need to be added externally to the devices (in addition care should be taken in handling the devices because of the high impedance sensitive gate and static handling precautions should always be observed).

In fact, having to add external zeners rather than simply relying on the internal ones gives the designer a greater degree of flexibility in deciding what characteristic zener and voltage rating is chosen. The zeners perform two main critical functions within any design:

1. To protect the gate-source junction from being damaged by an overvoltage condition. The rating of this junction is ±20V.
2. To limit the maximum current. Because a mosfet is a voltage driven device, clamping the gate voltage to a given level will also limit the current. Figure 14 illustrates the typical transfer characteristics for a double die device and demonstrates that about 9V corresponds to 14A (7A per die) at 75C. As can be seen the transconductance drops at increased temperature and therefore to allow for sufficient drive under all conditions

a gate clamping voltage of around 12V is recommended.

Although a simple zener clamp will serve to protect the gate of the mosfet and limit the current to a fixed maximum level there are still further precautions that are recommended to be taken in high power amplifiers that use rails above 50V. If a short circuit is applied to the amplifier output, although the zener clamp will limit the current to a set level where large power dissipation will occur and the mosfets will get hot fast and to some extent try and self protect, because of the high voltage seen across the devices the instantaneous dissipation will be extremely high and destruction of the devices is possible before thermal limiting has time to take place. Therefore, a protection network that senses a high voltage across the output devices and limits the current to a lower level is recommended. Figure 15 shows an example implementation of such a circuit that has been successfully used in amplifiers in excess of 1000W rms output. This circuit activates when only a very low voltage is present at the amplifier output and switches in a lower voltage zener diode of perhaps 3V across the gate to limit the current in situations where a large voltage is across the output devices such as a short circuit or highly inductive load.

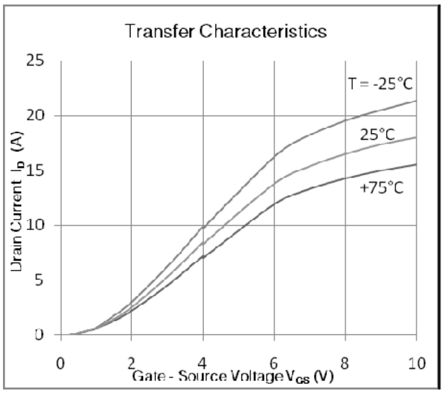

Figure 14 – Typical single die transfer characteristic

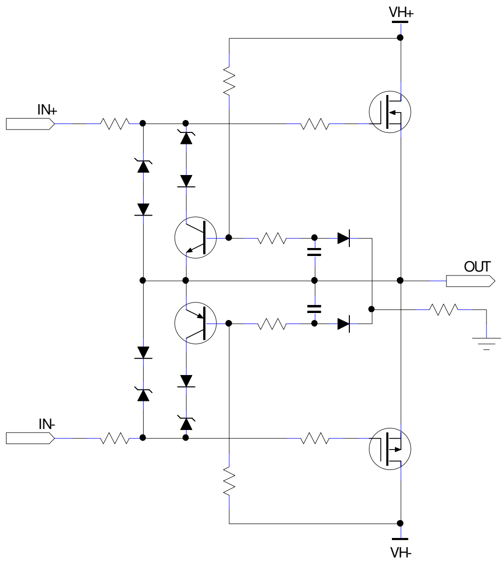

Figure 15 – Example Output Protection Network

Power Supply Rail Decoupling

As with any audio power amplifier, power supply decoupling local to the output stage devices is essential to guarantee optimised performance. In the case of a bad design, where the power supply decoupling is some distance from the power devices, the tendency for amplifier oscillation may be significantly increased.

It is normal and recommended to have local capacitors on each output rail close to the power devices with a parallel combination of an electrolytic and high frequency ceramic. The ground return of these capacitors should always be routed back to the ground star point and not share any other signal ground tracks. The recommended minimum local capacitance value is 100uF in addition to the power supply bulk smoothing capacitors which could be some distance away on another pcb.

Output Stability Networks

The majority of commercial amplifier designs include output stability networks as shown in Figure 4. The series RC network of R_z and C_z is generally referred to as a Zobel network and its purpose is to aid amplifier loop stability with inductive loads. L_o is to enhance stability with capacitive loads with R_D in parallel to damp the resonance of the inductor.

Although these networks are generally recommended for any design they will also be dependent on the specific details of the amplifier topology and it's intended application and it is therefore beyond the scope of this application note to give generic guidelines although the value ranges indicated will be typical and should provide a good starting point. An important note in choosing components is to avoid wirewound power resistors for R_z. The inductance of a wirewound can cause problems here and a film type resistor is a better choice. In addition a stacked film type capacitor is a better choice for C_z than ceramic as stability with temperature is far better.

Output Power Capability and Amplifier Classes

1) Power Limits

The maximum achievable power output of an amplifier design using lateral mosfets will always be from a combination of the voltage rating of the devices and the specified load impedance.

The worst case voltage stress across the drain and source of the device is always seen when the amplifier is off-load and mains is at it's highest specified limit (assuming a transformer, non-regulating power supply). When full output swing signal is applied under these conditions the voltage stress seen on the device will be approximately twice the power supply rail voltage. Therefore, for 200V rated lateral mosfets, the absolute maximum voltage rail used should be 100V when the amplifier is off load and the mains voltage is 10% high. This implies that for a power supply with 5% load regulation we will get an on-load voltage of:

$$VRail(onload)nominal = \frac{Vrail(offload)}{1.1} x\ 0.95$$

$$Vrail(onload)nominal = \frac{100}{1.1} x\ 0.95 = 86.4V$$

Assuming a voltage drop of around 5V across the devices for R_{dsON} drops and required gate drive this means our output power capability into a 4 ohm load at nominal input mains will be:

$$Pout = \frac{\left(\frac{Vpeak}{\sqrt{2}}\right)^2}{Rload} = \frac{\left(\frac{81.4}{\sqrt{2}}\right)^2}{4} = 828Wrms$$

And, obviously, twice this into a 2 ohm load assuming the same power supply regulation into the lower load. In practice, halving the load impedance will not double the output power as the power supply rail voltage will dip down further because of the transformer regulation and increased ripple on the smoothing capacitors.

2) Breaking the Voltage Limit Barrier

Going beyond these powers is desirable in many professional amplifier applications and may be achieved through either using a higher voltage output device or through a different amplifier topology. Because power dissipation in a class-AB amplifier becomes very high at these power levels a higher efficiency amplifier design is often chosen to both allow more power output for a given device voltage rating but also to reduce the amount of power that needs to be dissipated as heat. Two main topologies are in common use today and although there are varying definitions of the details of these we will define them as follows in this application note:

Class-G.

A two power rail system employing a lower voltage rail to reduce dissipation at lower output voltages. When this lower voltage is approached with increasing output voltage the rail starts to track the output until the main power rail is reached and clipping occurs. Figure 16 shows a simplified schematic demonstrating a typical implementation and waveforms showing the operation of this approach. Class-G has the advantage of a gradual following of supply rather than an abrupt change and is often perceived as potentially achieving higher audio performance than Class-H. The downside to this, however, is that the dissipation is spread across the top and bottom tier devices and therefore more attention to the design needs to be taken and potentially a larger number of devices in the top rail are required over class-H. There are some advantages to this as well though as this means that some of the power dissipation is taken away from the output stage of the amplifier, lowering those device's junction temperatures. This can be observed in figure 16 where as well as the total power dissipated being lower at average music levels, the power dissipated by the main output devices is significantly lower as some is moved to the rail devices. These power dissipation examples are based on the same 50V rail example from the previous section but with an additional low voltage rail set at 25V. Varying the setting of this lower rail will optimise the dissipation for different kinds of music material and should be chosen through experimentation and design goals of the end product application.

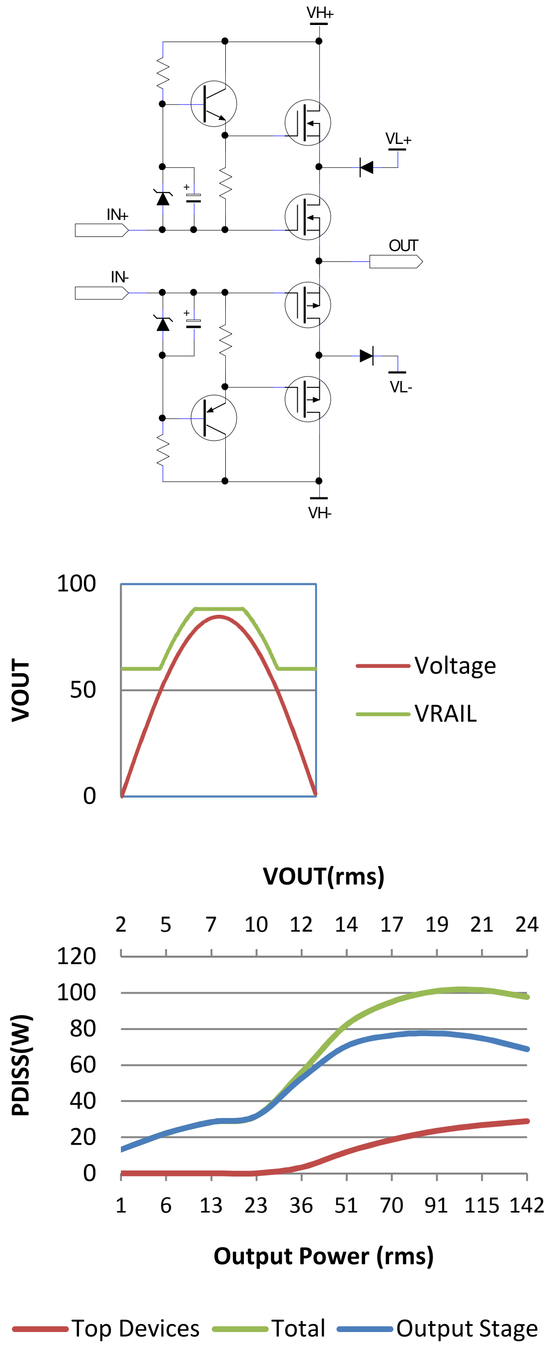

Figure 16 – Class-G simplified Schematic, Operational Waveforms & Power Dissipation

Class-H.

A 2 or more power rail system where the rails are switched on passing a threshold to improve dissipation of the output devices. Figure 16 demonstrates a 2 rail class-H system with a simplified schematic and operational waveforms.

Traditionally Class-H designs have been plagued by problems of spikes appearing on the output waveform as the rails switch in and out but modern components such as high voltage Schottky rectifiers and good design practices can realise a very high performing system with audio performance approaching that of Class-AB designs.

Despite the differences in operation of Class-G & H amplifiers, they both provide a similar improvement in output stage dissipation for a given number of rails but in practice Class-H is cheaper to implement and more easily extended to beyond just 2 power rails. There are a number of commercial professional amplifiers in production today using 3 rail class-H topologies.

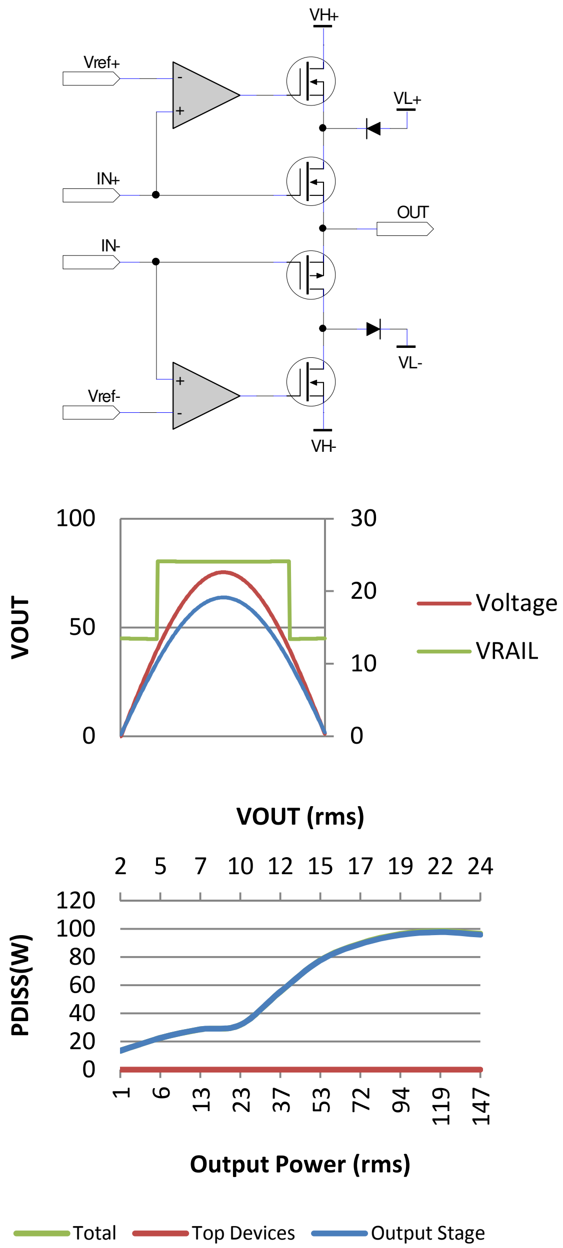

Figure 17 – Class-H Simplified Schematic, Operational Waveforms & Power Dissipation

The other significant advantage of both Class-G & H amplifiers is that the dissipation at quiescent due to the bias current will be significantly lower than for an equivalent power output Class-AB design. This is due to the lower voltage that the output stage sees on their drain terminals with zero voltage output from the amp and this equates to a lower running temperature and reduced overall power consumption which is often extremely beneficial in many applications.