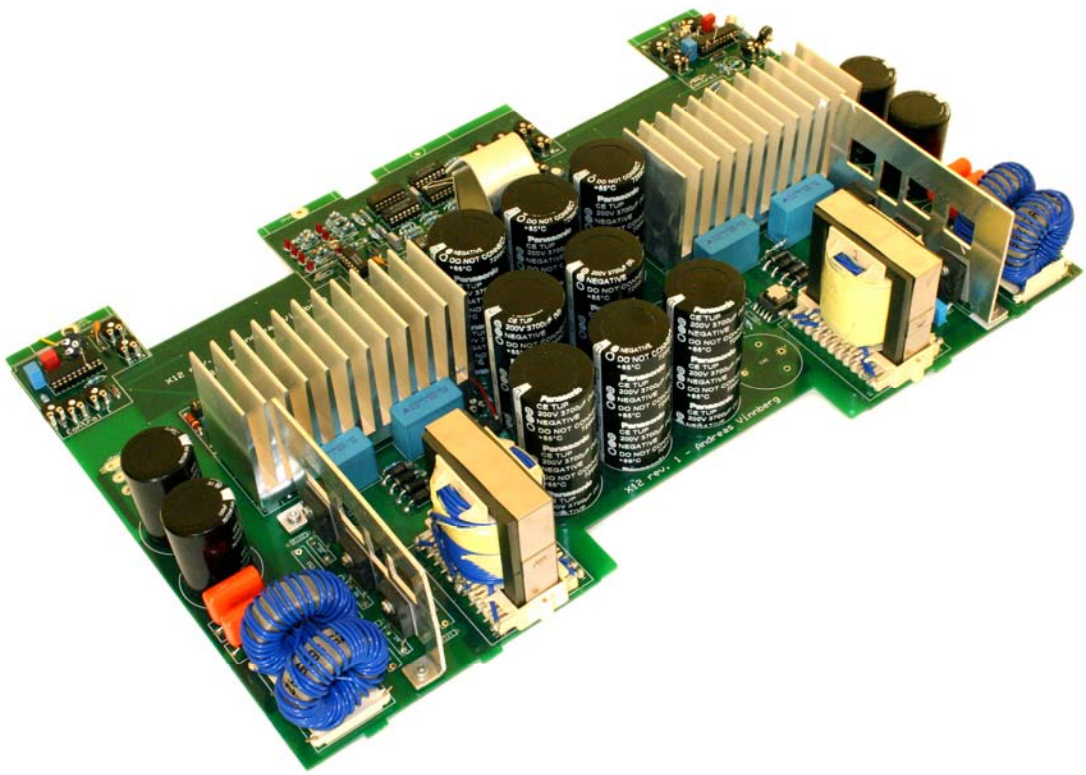
20 kW audio power supply – topology choices and construction



Abstract: This thesis highlights the differences between both high level topologies, mainly energy storage and average versus peak power transfer as well as low level topologies for both the main switches (flyback, half bridge, full bridge and so on) and output rectification from an audio amplifier perspective and find algebraic expression for which topology to choose. Functions for capacitor and transistor ratings compared to their price were calculated and research on optimum capacitor and transistor selection was carried out. A prototype was built to verify the results. For a DC/DC converter operating from a PFC front regulator with 20 kW designed program power, a peak power phase shifted double full-bridge operating 180 degrees out of phase with about all energy storage on the primary side is proposed.

1 Introduction

1.1 Background

With hundreds of millions switch mode power supplies manufactured each year, much work have been done in trying to decrease size, improve efficiency and reduce cost for every available topology. The most recent published work done in the area seems to lean towards DSP-driven power supplies using highly complex drive strategies. However, with the increased cost, this seems to be an uneconomical solution at the moment. Out of the traditional analogue driven topologies, much have been written about soft commutation in the last decade.

For power levels about 1 kW, two topologies dominate completely among the ones mentioned in recent publications – half and full bridges. Half bridges often for lower power levels and full bridges for higher, but the overlap between the two is large. If one investigates a bit further and look at what the power supplies are used for, one finds that half-brides are highly utilized in arc welders and full-bridges are far more popular in more critical fields of applications such as telecom supplies. Looking further, above 10 kW, full-bridges dominate completely.

Power supplies where high power and cost are more important than clean output such as those for arc welders seem to utilize half bridges with soft commutation to a high degree.

New possibilities with PFC

With the introduction of PFC front end converters for improving the power factor of modern audio amplifiers, the possibility to minimize the power supply by building a peak power converter with nearly no energy storage on the secondary side is now possible. This is because the PFC would allow a large capacitor bank to be used on the primary side and discharged deeply without top-wave rectification to occur. The author has not seen any examples where taking full advantage of the second order effects of a PFC is investigated, either in literature or in a commercialized product. More about the background of this is found in chapter 2.

1.2 Purpose of project

The scope of this project, named X12, was to investigate different topologies for audio amplifier power supplies and build a prototype of the most conformable candidate where size and efficiency where the ruling demands. The power supply should be electrically isolated from the mains and in a platform that could be able to deliver 40 kW of peak power to an intended 20 kW program power amplifier. The power supply should also be designed to be interfaced from a PFC pre-regulator and scalable up and down in power. Although the prototype is never intended for mass production, the topology must be suited for fitting inside a standard 2U 19”-enclosure. As an underlying design goal, the possibility to discard the secondary capacitance by using a peak power converter was to be tested.

1.3 Layout of thesis

The report starts with a background section, where previous work is explained.
The sections “energy storage” and “components available – capacitors” might seem similar but “energy storage” only focuses on why and how the power supply needs energy storage whereas “components available – capacitors” only focuses on what can be done with existing and soon available components.

2 Development of kW-scale audio amplifier power supplies

Audio power supplies come in several different configurations depending on price range, power level and designer preferences. The traditional method to for solving galvanic isolation and AC/DC-conversion is to use 50/60 Hz transformers followed by rectification and smoothing by using capacitors. This solution has many drawbacks as the transformer is very large and heavy compared to the switch-mode solution and the power factor is low due to top-value rectification and since the power factor is inversely proportional (although nonlinear) to the amount of capacitance. Thus, the amount of capacitance have to be limited which leads to the mains will experiencing amplitude modulation with the beat of the bass, which can be severe at high power. The limited amount of capacitance in conventional audio power supplies can be seen in figure 1 below where a Crown K2 is displayed.

Figure 1: Interior of a Crown K2

Please note the large toroid transformer marked with the red arrow compared to the four small 8200 μF 110 V capacitors marked with green arrows (the other two capacitors are located right under the first two and cannot be clearly seen in the picture.

Analogous to low power factor in this case is capacitor abuse due to the capacitors being charged for a very short while each cycle. This requires high ripple capacitors to be used which can affect their energy density in a negative manner. DC-rail voltage drops cannot be avoided without the use of an infinite capacitor bank and thus infinitely low power factor. Other forms of regulation would be possible with highly added complexity and cost and the author and his colleagues have not seen any such solution with the exception of the Carver M1.5t which uses triacs on the mains in order to reduce the size of the conventional transformer. This might have been used to achieve some regulation as well, although this was not the primary objective.

In order to get rid of the heavy transformer, switch mode power supplies are used. Several different switch topologies have been utilized and different brands have been keener for some than others. Regardless of which topology used, the power factor issue is not longer tied to the amount of capacitance used as long as it is located on the secondary side. Primary capacitance is still normally required here in order to reduce the transformer ratio needed for staying within regulation and this is still inversely proportional to the power factor. In this case, it is obvious why a designer let the power supply only transfer the average power needed, or at least near that value, and chooses a large secondary capacitor bank for maintaining the voltage on the rails.

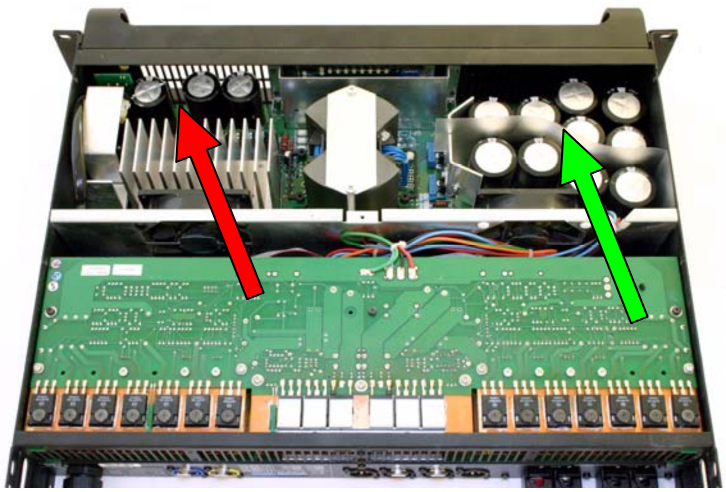


Figure 2: Interior of a Lab.gruppen fp+ 13000

An example of this is shown in figure 2 above where four 2200 μF 200 V capacitors are used on the primary side (red arrow) and ten 2700 μF 200 V on the secondary side (green arrow).

In the case where a separate PFC front regulator is used to increase the power factor, the option for primary energy storage has now become valid. This would allow for a *peak power* design where it is the converter which performs all or nearly all the regulation for the secondary by increasing the power to the secondary if it drops from the designed voltage value. Except for the benefit of having a higher voltage rail in times of high load and thus higher program and peak power, this would also allow a much more space efficient energy storage design since the primary capacitors can be discharged much more deeply, in contrast of the secondary, and compensated for in the turn ratio of the transformer. The Powersoft K20 (featuring PFC, not yet available for sale when this theses was presented) seen in figure 3 below comes a bit on the way where at least some portion of the energy reservoir is located on the primary side (red arrow) but still the vast majority is stored on the secondary (green arrow).

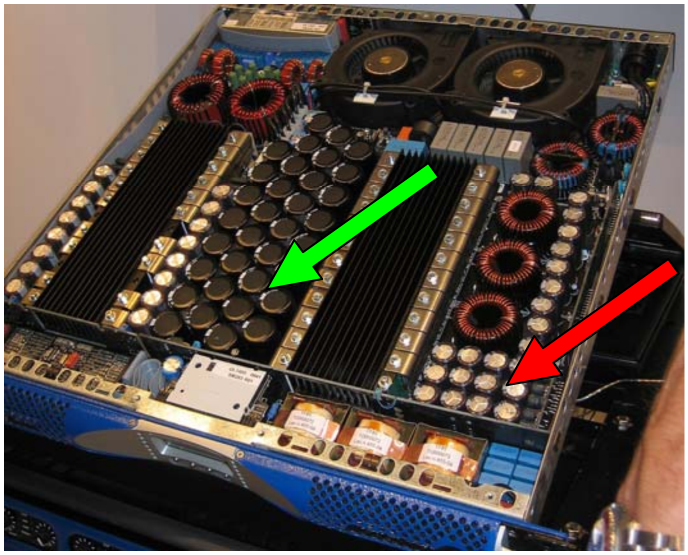


Figure 3: Interior of a Powersoft K20

This could be to the fact that a boost-only PFC is used and since the amplifier is rated for input voltages up to 265 V and the boost voltage is likely to be 400 V, only 12% of the primary capacitor energy can be used and therefore it is more economical to allow the secondary to ripple. This limitation could quite easily be overcome by redesigning the PFC stage and thus allow a much more compact DC/DC-converter to be used.

3 Topologies explained

In every switch mode power supply design there are both *low level topologies* to consider such as “How to interconnect the semiconductors?” and *high level topologies* such as “What is the optimum energy storage solution?” or “What control strategy is the best for my design?”.

3.1 Low-level topologies

Since the converter is required to have galvanic isolation, only topologies incorporating a transformer are considered here and for isolated power supplies, there are several switch topologies to choose from, all with their own benefits and drawbacks. Some could be discarded early without further investigation based on previous knowledge and some had to be calculated before their inherent problems proved them unsuited for the application.

When selecting a switch topology, cost is most often the ruling limitation, followed by losses and EMI (electromagnetic interference). In general, more advanced topologies are generally more expensive in terms of semiconductors to build but with less losses and lower EMI. The breakpoint where one topology will be better than another with a given set of demands can be very difficult to calculate and there can be a large overlap between them. In order to understand where the differences come from one needs to understand where the costs and losses come from. To begin with, there is an initial cost for all key components. A transformer, no matter how small, will at least cost the tooling and packaging price from the manufacturer even if there would be next to none copper or iron on it. In the same manner, a transistor or diode with infinitely small current and voltage capability will always cost at least as much as the package does. For control of the circuit, some kind of control IC is often used, which usually starts at $1. By only using this knowledge, we can see why a power supply manufacturer wants to minimize the number of semiconductors. Going up in power, the added cost for a slightly larger transistor or capacitor is often lower than for an extra diode or transistor. *At low powers, the number of components in the topology sets the cost.* Low power topologies would therefore be those which require few semiconductors, starting with one switch flyback and moving up to two switch flyback and the various forward converters.

When we move up further in power, each topology has it disadvantages when it comes to voltage and current stress. Therefore, there are turn-over points where the exponential added cost for a transistor with higher voltage rating is higher than the cost for another transistor with lower voltage (and current) rating. For MOSFETs, this is due to the fact that the variable cost for the silicon is proportional to the area, and as the area needed for a given voltage and on-resistance is

$$\mathrm{A}_{\mathrm{die}} = k\frac{\mathrm{V}_{\mathrm{ds}}^{2.6}}{RdsON} + m \qquad [1]$$

, where the 2.6 exponent is found in reference [1]. More about transistor die area and cost in section 4.1.2. This calls for topologies with less voltage stress on the transistors rather than as few transistors as possible. As for passive components, the wind has changed here so the amount of capacitance and copper on the transformer costs more than packaging and control – hence the dynamic cost is ruling. *At high powers, the component stress in the topology sets the cost.* High power switch topologies would therefore be low stress ones such as push-pull (for low voltage applications), half bridge and last full bridge for the most power demanding applications.

Below is a brief walkthough of the most popular low level topologies, all of them possible solutions for an audio power supply, followed by a comparison table.

3.1.1 Flyback

Flyback converters are one of two topologies which work with a single transistor and diode and this comes with the price of very high voltage and current stress on the switching transistor along with the lowest transformer utilization of the available topologies. A flyback converter works by charging a current though the primary winding of the transformer, which acts as an inductor here. When the transistor opens again, the inductor discharge though the secondary winding via the transformer core and is rectified by a diode. Flyback converters are therefore somewhat different from other types of converters in the sense that instantaneous power transfer does not take place since the primary and secondary winding does not conduct current at the same time.

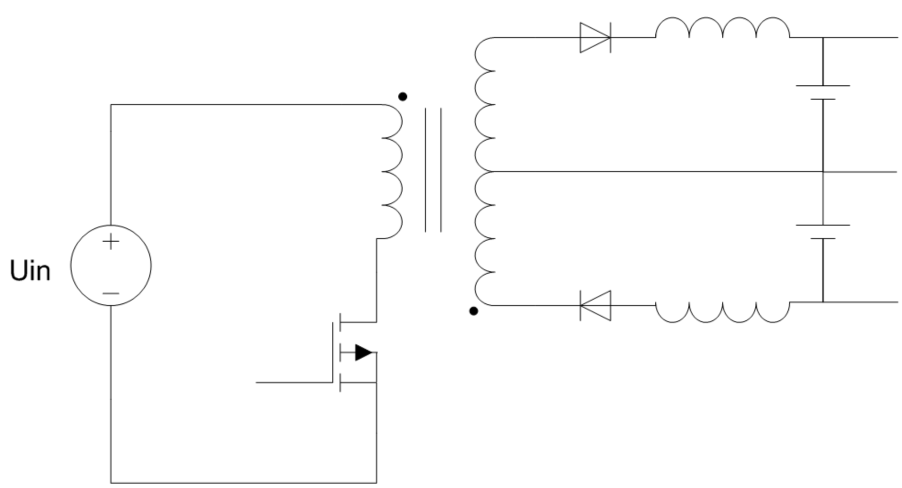


Figure 4: Single transistor flyback converter

Due to the way the input connects to the output when the primary is conducting, the output voltage is reflected back to the primary (via the transformer ratio) so the voltage stress on the transistor is very high in flyback converters. If a demagnitizling winding is not used, like in figure 4 above, all energy stored in the leakage inductance in the transformer has to be dissipated either in the transistor or in an external snubber over it. Other than semiconductor stress there is the problem with transformer utilization since the transformer is used as a coupled inductor and current is charged though it for each cycle, the resulting waveform is triangular whereas the top value still sets the saturation limit needed for the core. Except for Lab.gruppen’s ~4 kW main power supply, a 1+ kW Hewlett Packard laboratory power supply dating back to the 1970’s and a Göfab arc welder from the 1980's, the author have not been able to find any >1 kW single transistor flyback power supplies, for reasons explained above.

3.1.2 Two transistor flyback

The two transistor flyback converter, as seen in figure 5, works in a very similar way with the benefit of half the voltage stress over the single transistor version. Moreover, since the two diodes make a current path from the transformer, less or no snubbing, are needed for the transistors, which in turn do increase efficiency [2].

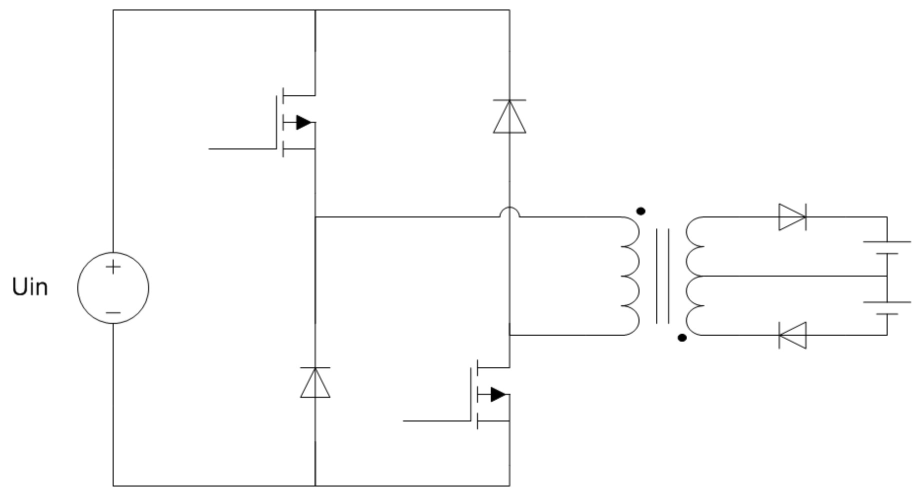


Figure 5: Two transistor flyback converter

3.1.3 Forward

A forward converter works in a similar way as that of the flyback as seen below in figure 6. The forward provides slightly better transformer utilization and can reduce the voltage stress, depending on which transformer ratio that is needed at the cost of a demagnetization winding or zener diode across the switch to conduct the reflected magnetization current. Forward converters are normally found in low power applications (<300 W) and often with low input voltage since the voltage stress is twice the input voltage, thus more popular in the US than Europe. Telecom equipment (48 V) and battery powered devices would also be suitable here.

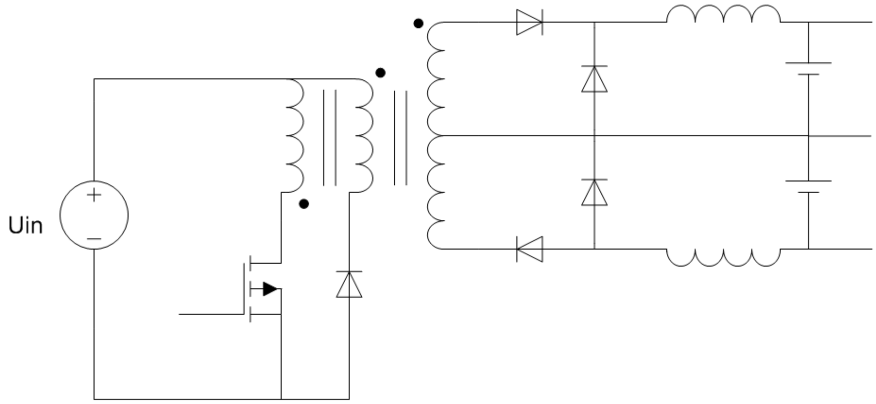


Figure 6: One transistor forward converter

3.1.4 RCD forward

A variation of forward is RCD forward where the reset winding is replaced with an RCD network which dissipativly discharge and resets the core. The benefit besides a simpler transformer is higher maximum duty cycle. The non-RCD version can accomplish 50 % maximum duty-cycle whereas the RCD version can go up to about 75 % [3]. The increase in maximum duty cycle is analogous to higher Vin/Vout ratio which is often used in both non-PFC and simpler PFC universal power supplies ranging from 85-265 Vac. Wide input range is also needed in the case where a PFC is applied if there is a desire to discharge the capacitor bank deep for space efficiency reasons. Voltage stress is even higher here so the universal input equipment which uses this topology have paid a high die area price from the voltage stress already so high power is unlikely to be found here. RCD forward converters are seldom seen today.

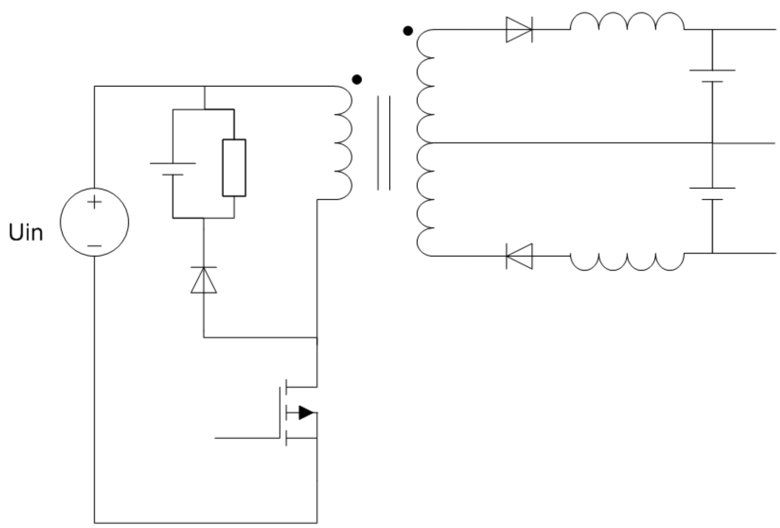


Figure 7: RCD forward converter

3.1.5 Two switch forward

In the two switch forward topology seen below in figure 8, the benefits are very similar to that of the flyback – voltage rating on the switches goes down by 50 % and the path for the reflected magnetization current obsoletes the reset/demagnetization winding. Lower voltage stress is analogous for increased probability of this topology being used in more high power applications. Two switch forward converters are popular today in applications just above the one transistor flyback power level.

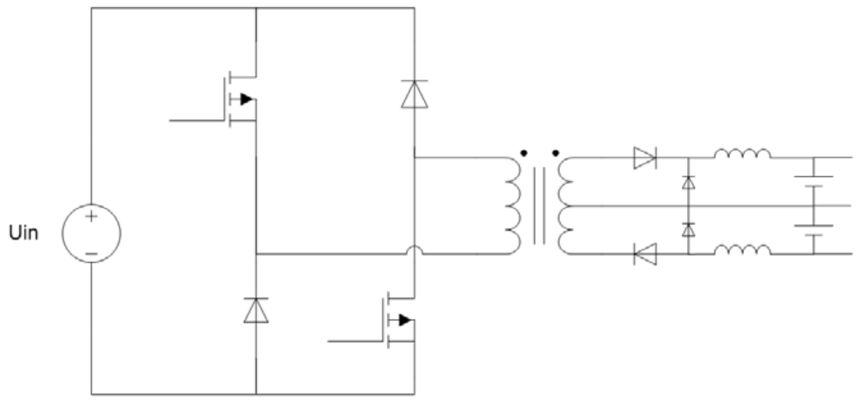


Figure 8: Two switch forward converter

3.1.6 Active clamp forward

The active clamp forward converter seen in figure 9 below is another variation of the forward converter whereas this one is aimed to replace the RCD-block in the RCD-version with a switch in order to enable it to recycle the current back to the primary capacitor which would otherwise be dissipated in the resistor. Lower losses would again increase the likeliness of a high power application using this topology. Active clamp forwards are quite uncommon today.

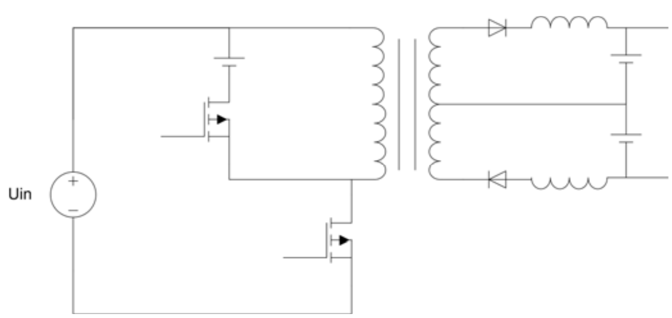


Figure 9: Active clamp forward converter

3.1.7 Push-pull

This is the first topology in this list which has bidirectional transformer excitation. This comes with several benefits, mainly a reduction of transformer size with up to 50 % in the ideal. A downside with push-pull is that they require very careful transformer design where the primaries have to couple hard to each other and timing of the switches as the reflected EMF can quickly become a problem. Since switch-mode power supplies' transformers are often thermally limited and transformer cores have high hot spot to ambient thermal impedance by nature, reduction of the transformer size is very beneficial at high power levels. In figure 10 we can see an example of a push-pull converter. As push-pull experience high voltage stress on the transistor, they are most often found in low voltage/battey powered equipment.

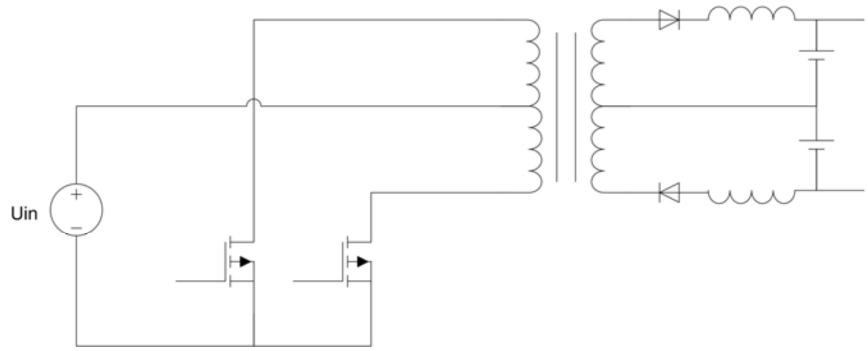


Figure 10: Push-pull converter

3.1.8 Half bridge

Another way to use two transistors is to build a half-bridge which is very similar to an active clamp forward, although the transformer is here center tapped and the connection slightly different. The half bridge requires a voltage divider of some sort, usually made up of capacitors which are an increasing problem with current and lower frequency. Furthermore, a capacitor in series with the transformer is needed unless some advanced control method for not DC-biasing and saturating the core is used. Half-bridges are popular in high frequency and relatively low power applications. Arc welders using half-bridges with IGBTs seems more and more popular in the high power range.

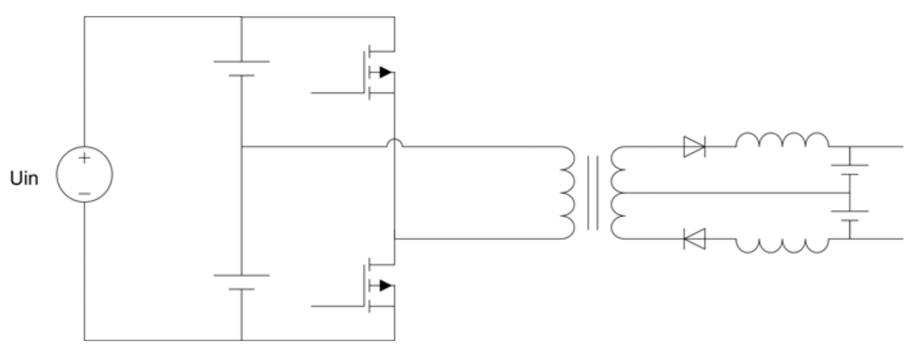


Figure 11: Half-bridge converter

3.1.9 Full bridge

The full bridge is the full blown version of the half bridge where the voltage divider is replaced with another leg (half bridge). This does not only eliminate the capacitors but also serves doubles the voltage delivered to the transformer without increased voltage stress on the transistors, i.e. half the voltage stress compared to output power. There are two major control strategies for full bridges, duty cycle control or phase shift control where the later is used to facilitate soft switching. Full-bridges is as far as it goes for normal switch topologies and they range from about 500 W up to the very high power range of trains and industrial applications in the MW-range where the structure is the same but with recover circuits for reflected energy from leakage inductances for example.

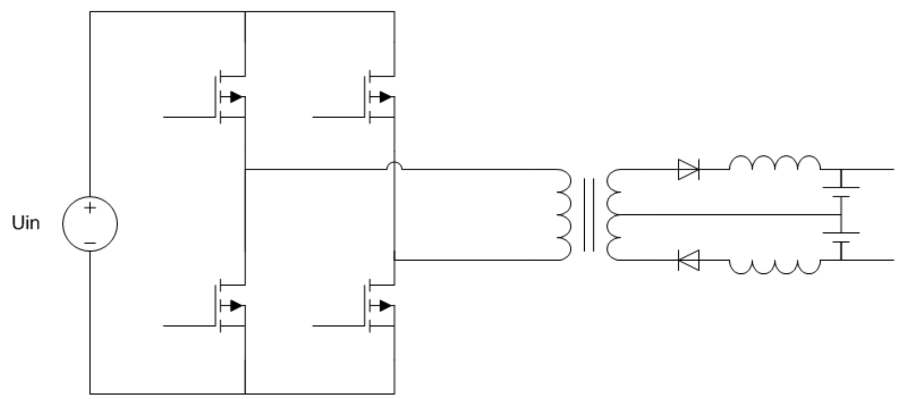


Figure 12: Full bridge converter

With dropping prices on semiconductors (a variation of Moores law applied) and increasing prices on copper [4] and aluminium [5] as well as problems with EMI, half and full bridges are taking market share in the lower power ranges which was previously dominated with simpler topologies. High power converters (1+ kW) have traditionally utilized full bridges and still continue to do so.

3.1.10 Comparison table

Topology	Flyback	Two transistor flyback	Forward	RCD forward	Two switch forward	Active clamp forward	Push-pull	Half-bridge	Full-bridge
Transistor voltage stress	$V_{in}+V_{out}\left(\frac{N_p}{N_s}\right)$	$V_{in}+V_{out}\left(\frac{N_p}{N_s}\right)$	$2V_{in}$	$>2V_{in}$	V_{in}	$V_{in}\left(\frac{1}{1-D}\right)$	$2V_{in}$	V_{in}	V_{in}
Transistor current stress	$\frac{V_{in}t_{on}}{L_p}$		$>\left(\frac{N_s}{N_p}\right)I_{out}$	$\left(\frac{N_s}{N_p}\right)I_{out}$		$\left(\frac{N_s}{N_p}\right)I_{out}$	$\left(\frac{N_s}{N_p}\right)I_{out}$	$\left(\frac{N_s}{N_p}\right)I_{out}$	$\left(\frac{N_s}{N_p}\right)I_{out}$
Output voltage								$\frac{V_{in}}{2}\left(\frac{N_s}{N_p}\right)$	$V_{in}\left(\frac{N_s}{N_p}\right)$
Relative transformer utilization	$\approx\frac{3}{8}$	$\approx\frac{3}{8}$	$\approx\frac{1}{2}$	$>\frac{1}{2}$	$\approx\frac{1}{2}$	$>\frac{1}{2}$	$\frac{1}{\sqrt{2}}$	1	1
Transistors needed	1	2	1	1		2	2	2	4
Diodes needed before rectification	0	2	1	1	2			2	4

3.1.11 Paralleling

At medium or high power levels it can sometimes be the case that two small transistors are cheaper and easier to cool than one large one. This would require paralleling and it can be done in two ways.

Parallel transistors

This is the obvious no-brainer solution and requires only one transformer, one output rectification and filtering unit and one regulation circuit. Transistors that can be paralleled will and have to share the current equally in a passive way, so no precautions have to be taken here. Problems with this solution are several - mainly that PT IGBTs cannot be used. In high power and/or high frequency conversion, the dynamic course of events cannot be neglected. If one transistor for instance would have a lower gate threshold than the others, it would take more abuse at both turn off and turn on every switch cycle. This effect varies from negligible to catastrophic failure. Paralleled transistors does also have secondary effects, especially on the transformer since one large transformer will have a high hot-spot to surface thermal impedance and relatively low surface area which can be a problem as transformers are normally thermally limited. Parallel transistors is a cheaper solution if the cost of control and tooling are still notable and the transformer size have not yet become a problem – i.e. medium power applications.

Parallel converters

This would be the exclusive solution which comes at a higher price in many areas but also with great benefits. Here, any transistors can be used, including PT IGBTs. Since parallel converters can be driven out of phase by 180 degrees in the case with two and 120 with three and so on, the output ripple can be reduced to half or a third and so on with a given output inductor without any drawbacks. If space or cost is of greater concern, the output inductance can of course be reduced with unaffected output ripple instead. This could very possibly be a cheaper solution at high powers as very large inductors, transformers, semiconductors and so on can be hard to source and in many cases one large component can often be more expensive than two half the size. Two converters instead of one would provide at least some redundancy. If one would stop switching for some reason without breaking anything vital, the other converter could carry on and work up to half the rated power. Parallel converters would be more economical once the static costs have become neglectable, the transformer design become very complicated from sourcing (due to size) and/or thermal reasons or when PT IGBTs are required to run in parallel.

3.1.12 Output rectification

In all bridge topologies, there are several options for output rectification and filtering, especially in audio applications where two separate DC-rails are commonly used. Regarding rectification, given the output voltage(es) and current, there is an optimum number of diodes and ways to interconnect them in order to minimize losses and ripple. As we shall see, minimum losses and/or ripple do not necessarily mean minimum cost instead rather the contrary. The most economical solution would therefore take in account for transistor and transformer stress, which can be very complicated to solve analytically. For filtering, there are fewer options but they should be considered too.

The simplest solution for a two winding secondary as most amplifier power supplies have, uses two diodes which give the lowest number of components but the highest losses due to current only being transferred for 50 % of the time. A two diode output configuration can be seen in figure 13.

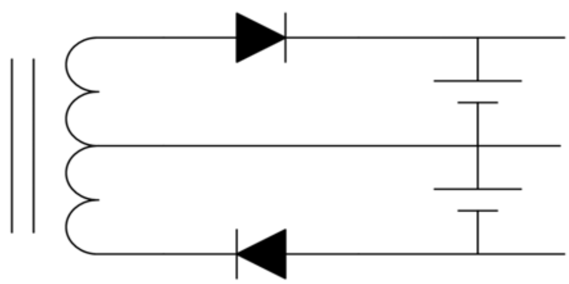


Figure 13: Half-wave output rectification using two diodes

With one more diode, one can connect it between the plus and minus rail (before filtering, not seen in the picture) in order to allow the current to freewheel between the rails without going though the transformer which in this case acts as a series resistor. An example of this can be seen in figure 14.

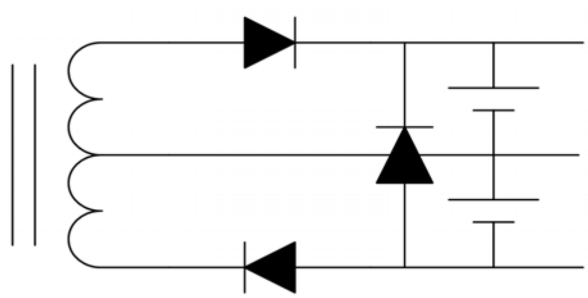


Figure 14: Half-wave output rectification with a freewheel path using three diodes

With four diodes, one has the option to do a full wave rectification as in figure 15 or to do normal half-wave and provide two freewheel paths, one for each rail. Full-wave has the advantage of less ripple with the cost of current freewheeling though the transformer and the half-wave solution will have twice the ripple and does not freewheel.

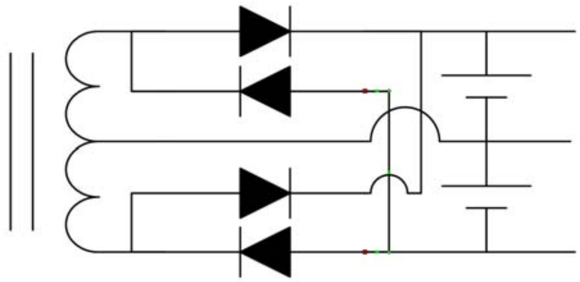


Figure 15: Full-wave output rectification using four diodes

Another solution with four diodes is half-wave rectification and two separate current paths, one for each rail, for the freewheeling current as seen in figure 16. This solution is unlikely to be found in a production amplifier.

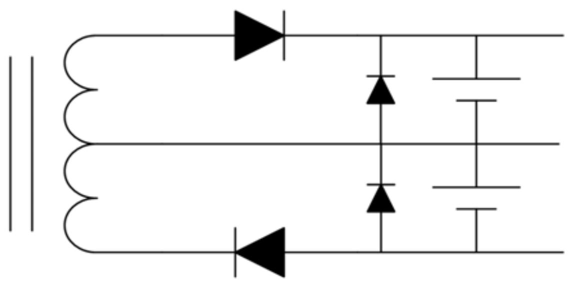


Figure 16: Half-wave output rectification with two freewheeling paths using four diodes

Additionally one more diode would give us full wave rectification and a freewheel path from the minus to the plus rail, just like the case with three diodes. This is exampled in figure 17.

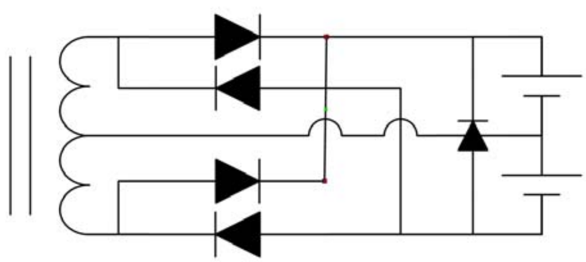


Figure 17: Full wave output rectification with a freewheel path using five diodes

Six diodes would do the above but with separate paths for the plus and the minus rail, like the case with tree versus four diodes. This would be the most expensive solution but also the one which puts the lowest stress/demands on the output capacitor, transformer and main transistors which could all in all be the cheapest solution for a given set of demands. A schematic of a six diode rectifier solution is seen below in figure 18.

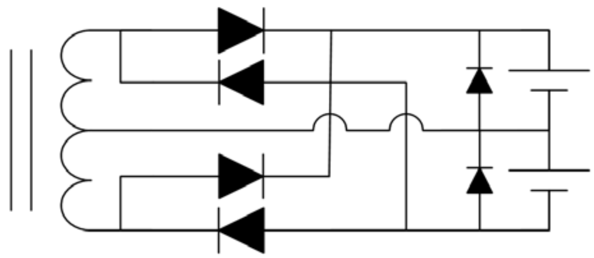


Figure 18: Full wave output rectification with two freewheel paths using six diodes

Using only one diode for freewheeling would decrease the stress on the transformer only when both rails can freewheel whereas one diode for each always provides one path for every situation, therefore two freewheel paths should put less stress on the transformer. Weather this is needed or not in a high powered class D design or not is beyond the scope of this project.

In just about all fast switching power supplies the transformer is thermally limited so reducing the losses in it can be very beneficial unless the transistor cannot take any more ripple. As described in the beginning of section 3.1, component stress sets the cost for high power applications and this along with the thermal limitation calls for better and better output rectification methods for increasing power.

3.1.13 Current/voltage fed

In every design with a few exceptions, the power must be transferred from a voltage stiff source (the mains) to a current stiff middle link before it can provide a voltage stiff output. This can be solved in two major ways, current feeding or voltage feeding. Voltage feeding is the simplest and most common solution and is realized with an inductor on the secondary side, in which case the main switches are voltage fed as seen in figure 19.

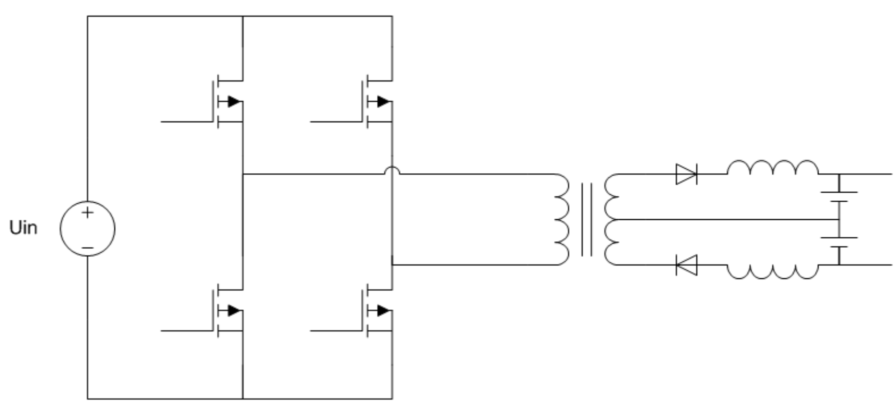


Figure 19: A voltage fed full bridge

As long as the output inductor can be considered large, the current though the primary switches as well as output rectification and smoothing can be considered fixed which simplifies calculations and control.

In the other case the inductor is instead placed after the primary capacitor and just before the main switches with no inductor on the secondary side. Current fed conversion is seldom seen today as it requires a more complex gate drive strategy since legs in bridge designs or the corresponding in other topologies have to be shorted in order to build up a current in the main inductor and the power transfer will instead take place when the current then freewheels though the circuit as seen in figure 20.

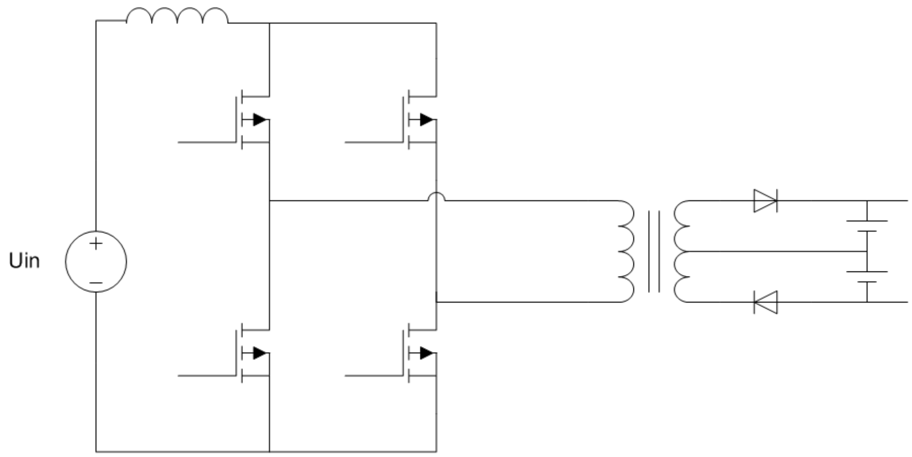


Figure 20: A current fed full bridge

There are of course exceptions to this rule, especially when the output current is very high compared to the input current, often in the case with high input voltage and low output voltage where the inductor design can be very simple and cheap when placed on the primary side compared to the secondary.

A very interesting exception is the power supply in Crown's I-tech series which is a current fed "poor mans PFC". Here, top value rectification have to be avoided and instead the inductor makes sure a constant current is drawn from the mains on a switch cycle time basis and by adjusting the duty cycle to track the input voltage waveform, PFC operation as achieved.

3.1.14 Resonant converters

In all topologies there is the option to make it fully resonant in order to reduce EMI and switching losses. This is done by using a resonant tank in which the energy is transferred between a coil and capacitor which can be connected in different configurations. The problem with resonant converters is that they make the current waveform more or fully sinusoidal which means lower transformer utilization. Furthermore, adding series components will undeniably lead to more resistive losses which are further increased by the added amount of reactive power being transferred back and forth. For a peak power converter where the transformer size can be a problem, resonant operation is not a good option.

Another option is to only make the transition resonant in order to reduce the switching losses and reduce the EMI somewhat, called soft commutation, soft switching, ZVS or

ZCS, depending on operation principle. In this case, either the leakage inductance in the transformer or in some cases an external series inductor will form a resonant circuit with the output capacitance of the transistors and the transformer capacitance. This will unlike the full resonant solution interfere very little with the transformer utilization and not add any series resistance unless a series inductor is needed. For high frequency conversion, soft commutation is often very favorable.

3.2 High-level topologies

High level topologies refer to methods for power conversion which are of a more abstract nature.

3.2.1 Primary or secondary energy storage/peak or average power

One important aspect of designing an amplifier is energy storage in cases where the power supply cannot deliver the peak power, which is the case for most amplifiers today. As high-end power amplifiers can deliver several kW of power, although only for a short while, and the grid will pass a few kW before the fuse trips, the need for a secondary power resource is obvious. Whereas all energy is being delivered from the mains as usual, the pulsed power comes from an energy storage formed of capacitors. The conventional way to do this is on the secondary side, since primary energy storage is impossible when using conventional mains transformers.

Benefits of secondary storage

With newer SMPSes in today's amplifiers, the energy storage have remained on the secondary side although the conventional transformer has now been eliminated. Reasons for this could be several, including easy regulation as the output ripple is quite low (and needs to be so!) and pumping in the same beat as the music, thus allowing the feedback loop to be quite slow. If already working output modules which can handle ripple on the DC-rails in one way or the other are available, a lazy and/or ignorant engineer might find it a pragmatic approach to leave it that way. A peak power solution in an amplifier which does not feature PFC would not be possible as mentioned before in section 2. Allowing the DC-DC-converter to only deliver the average load, puts less stress on it and allows more topologies to be chosen from, during the design process.

Benefits of primary storage

The problem with secondary energy storage is ripple and voltage stress. If one intends to use the energy stored in a capacitor, one must discharge it to a lower voltage according to $E = CU^2$. If your application will work with 10 % ripple, you can at most use 21 % of the stored energy. At the same time, the output transistors must be able to withstand the maximum voltage the capacitors will be charged to. These two facts are quite counterproductive for a good amplifier design.
In a MOSFET topology, transistor rating costs

$$\$_{si} = k_{chip}\left(\frac{U^{2.6}}{RdsON} + A_{bond}\right) + k_{package} \qquad [2]$$

as suggested in section 4.1.2 and capacitor rating costs

$$\$_{cap} = k_{chem}\,(U^{\alpha} C^{\beta}) + k_{package} \qquad [3]$$

as suggested in section 4.2.1.

Suppose one wants to build an amplifier capable of delivering x W into y Ω without clipping (defined as x W for 25 ms and 0.063x W (-12 dB) for 375 ms), a $\pm\sqrt{2xy}$ V rail would be needed. The power supply for this amplifier would need to deliver 0.125x W continuous. If it is designed that way, the secondary would need to ripple. If the voltage ripple is allowed to be a factor z higher than the working voltage, the new minimum transistor voltage rating would be $z\sqrt{2xy}$. To dimension the capacitor, it would need to store enough energy to allow all the excess power to be taken from it during the entire burst. A total of $(1-0.125)x*25m = E$ joules need to be stored. With the total stored energy being

$$E_{cap} = C*\left(z\sqrt{2xy}\right)^2 \qquad [4]$$

, only

$$E_{cap,use} = C*\left(\left(z\sqrt{2xy}\right)^2 - \left(\sqrt{2xy}\right)^2\right) \qquad [5]$$

can be used in the application.

Since E is set, C can now be quantified.

$$C_{needed} = \frac{25m*0.875}{2xy(z^2-1)} \qquad [6]$$

Before calculating the cost, RdsON have to couple to current which it does via the thermal resistance R_{js}, the RdsON increase due to temperature rise and ambient temperature. The thermal impedance could also save a design if the time constant is longer than the burst period, which is likely for a high power design.

$$I_{trans,\max} = \frac{\left(T_{j\max} - T_{amb}\right)}{k_{thermimp}k_{Rdsrise}k_{saftey}R_{jc}R_{ds\,ON}} \qquad [7]$$

The cost of transistors and capacitors would therefore be:

$$\$_{total} = \left(nk_{chip}\left(\frac{(k_{Usaftey}z\sqrt{2xy})^{2.6}}{\dfrac{\left(T_{j\max}-T_{amb}\right)}{k_{thermimp}k_{Rdsrise}k_{saftey}R_{jc}\sqrt{\dfrac{x}{y}}}} + A_{bond}\right) + k_{package}\right) + m\left(k_{chem}(z\sqrt{2xy}^{\alpha}\frac{25m*0.875}{2xy(z^2-1)}^{\beta}) + k_{package}\right) \qquad [8]$$

Suppose that one 2500 W @ 2 ohm channel uses two MOSFETs in TO-247 capsules and requires three 35*50 mm capacitors, a safety-factor of 1.5 for both current and voltage, an ambient temperature of 50 degrees, Tjmax of 125 degrees, a thermal impedance factor of two, RdsON rise of 2.3, Tjs of 1 K/W, we find:

$$\$_{total} = \left(nk_{chip} \left(\frac{\frac{(k_{Usaftey} z\sqrt{2xy})^{2.6}}{\left(T_{j\max} - T_{amb}\right)}}{k_{thermimp} k_{Rdsrise} k_{saftey} R_{jc} \sqrt{\frac{x}{y}}} + A_{bond} \right) + k_{package} \right) + m\left(k_{chem} (z\sqrt{2xy}^{\alpha} \frac{25m*0.875^{\beta}}{2xy(z^2-1)}) + k_{package} \right) \qquad [9]$$

This comes from the fact that the energy stored in a capacitor is

$$E_{total} = \frac{C\hat{U}^2}{2}$$

But the useful energy is only the

$$E_{useful} = \frac{C\hat{U}^2}{2} - \frac{C(\hat{U} - \Delta U)^2}{2}$$

Part. In a secondary capacitance design, the $\Delta U / \hat{U}$ -ratio must be kept small for economic reasons and hence the capacitors will have a very low utilization. In a peak power converter this would no longer be a problem and the now primary capacitors can now be discharged as deep as one can accept high transformer ratio and duty cycle loss.

3.2.2 Current/voltage control

In every regulated design some kind of feedback is required and this can be solved in several more or less advanced solutions. The two most obvious physical quantities to measure for regulation purposes are voltage and current. It should be noted that not all power amplifiers have regulation and both solutions comes at a price. No regulation would mean rail voltage dips analogous to lower power and increased cost in component stress just like the case with average power/secondary storage. Regulated designs does of course cost more in terms of components needed but in high power amplifiers this is quickly recapitalized with lower voltage stress compared to power output.

Voltage control

Traditionally audio power supplies have relied on either no regulation at all or regulation using voltage feedback. In a secondary storage solution, voltage feedback makes sense as the output voltage moves slowly and by design does not rely on fast feedback. Current mode control could still be beneficial here but since a large capacitor bank does not only discharge slowly but also charges slowly, speeding up the control loop would soon saturate/max out the converter.

Current control

Current control aims at minimizing the output error by creating an inner control loop based on the current. Unlike output error which takes some time to detect depending on load and capacitor bank size the primary current reacts faster as an increased output current will directly lead to a higher primary current. How much current is taken from the primary depends on the amount of capacitance and ESR of the secondary as well as total impedance formed of the transformer, semiconductors and PCB. If the capacitor bank would be large, much of the instantaneous current would be taken from the there the current control loop would therefore provide less increase regulation speed/stability to the system than if it where small, unless double sensing would be used which is discussed in the next paragraph. This could very likely be the reason why current control loops are seldom seen in audio power amplifiers.

There are several methods for implementing a current control loop in a power supply. The most popular would be to connect a current transformer in series with the primary winding of the main transformer. This puts no DC-bias on the transformer in many topologies which simplifies its construction. This allows one to measure consumed current and detects any transformer saturation but not cross-conduction in for example bridge topologies. In order to detect any cross-conduction a Hall effect or otherwise DC-capable sensor could be used on the primary DC-voltage side of the converter. If one wants to compensate fully for the current portion being taken from the secondary capacitors one could use a Hall effect/shunt sensor on the output as well in a double sensing configuration and regulate the input current to match the output (multiplied by the transformer ratio). Voltage feedback would in this case only be needed for getting rid of DC-drift and the regulation speed would in theory only be limited to the switching frequency and inherent slowness of the converter, which in turn is considered far faster than the output ripple swings.

4 Components available

A very important factor for determining which topology to choose is how good the available semiconductors are. If one could lay his/her hands on an ideal transistor, there would be very little reason to choose an especially advanced topology. Such components do of course not exist, thus a power supply designer has to calculate which topology will be optimal for the given application using existing components.

4.1 Transistors

4.1.1 BJT

BipolarJuction Transistor where the first transistors to be mass produced [6] and therefore, the first to be put in switch mode power supply applications. They are more seldom seen today due to their somewhat complicated drive technique for switch mode application. They do however have excellent high voltage and high current abilities. Apart from output transistors for audio amplifiers, most of the high powered BJTs “end up” in IGBTs, which are explained later on.

4.1.2 MOSFET

When it comes to MOSFETs, they are heavily used in low power and/or low voltage applications since they are easy to drive and only have a series resistance with no fixed forward voltage drop. The parasitic reverse diode is also advantageous in many configurations where freewheeling is needed, further increasing their popularity. They do come in both N- and P-types, whereas the P-type is almost exclusively used in audio output power stages due to their intrinsic 2.3 times higher resistance for a given die area. MOSFETs have two different interests for X12, both as candidates for the actual converter and, together with capacitors, set the function for an optimum balance between the power supply and output modules which are made with MOSFETs.

Comparing MOSFETs is time consuming as there are tens of thousands available, all with their advantages and disadvantages. One approach to solve this was to parse the search results from Farnell’s webpage after a search for suitable components and use MATLAB to determine the relation between price, voltage rating and current capability. To eliminate at least some variables, all test candidates came in uninsulated TO-220 capsules. The major remaining problem would then be speed, as all of them have different gate, output and miller capacitances. Ignoring this fact, MATLAB suggested a relation of XXX.
Another way to compute this was to simply ask a semiconductor manufacturer to provide the information and IXYS proved kind enough in this matter. The two MOSFETs IXTH230N085T and IXTH220N75T come in the same package and the price differs as follows:

Transistor	Voltage [V]	RdsON [mohm]	Relative price
IXTH230N085T	85	4,4	128
IXTH220N075T	75	4,5	100

Although the RdsON is slightly different we can see that the voltage increase of 13.33% comes with a 28% price increase, hence a relation between voltage an price as $ = U^1.97230652

Transistor	Voltage [V]	RdsON [mohm]	Relative price [marshmallows]
IXFK120N25P	250	24	100
IXFK140N30P	300	24	146

In the case of transistors with slightly higher voltage, a 20% voltage increase costs 46% more and here the relation is $ = U^2.07565381

When it comes to RdsON, two transistors from IXYS HiPerFET was used, namely IXFH120N20P with 22 mohm and IXFH74N20P with 34 mohm, both having a 200 V voltage rating. The 35.3% decrease in resistance increases the price with 81.83%. The relationship is therefore $ = (1/R)^1.97788634.

As we can se here, MOSFETs will be less and less favorable for higher and higher voltages. At some point, other factors will come in as well, limiting all MOSFETs operational voltage.
MOSFETs can often be paralleled due to their resistance positive temperature coeficient, which is very useful if one would need to spread out the heat generated or have larger thermal mass to handle transients.

Although MOSFETs dislike high voltages and suffers from I^2R-losses, 12 different candidates was selected, all of them being best in their range or series.

4.1.3 IGBT

IGBT stands for Insulated Gate Bipolar Transistor and are BJTs with a modified MOSFET for driving its base. There has been heavy research in this field since their introduction in 1983 due to their dual nature of easy driving due to MOSFET gate and very high power capabilities thanks to the BJT power stage. They are currently working their way down in power with the introduction of PFC front regulators which often demand higher voltage rating on the DC/DC-converter than without PFC and also up in power to the drive inverters for trains, trams and hybrid/electric vehicles, replacing SCRs and thyristors.

The two most dominating characteristics to look for in IGBTs in SMPS applications like X12, is the forward voltage drop as a function of temperature and current as well as output capacitance. The voltage drop should of course be as low as possible at elevated temperatures and the dynamic resistance should also be as low as possible. Depending on how wide current range the DC/DC converter should work with, one can trade high voltage drop for low resistance in the case of high power and vice versa. The temperature variance though, comes in two different categories, explained later on. The capacitance should be low, unless the leakage inductance in the transformer is high and cannot be reduced in which case the capacitance should match the inductance for the given switching frequency. More about ZVS switching in section 3.1.14.

IGBTs comes in two major categories, punch through (PT) with have negative resistance temperature constant and non punch through (NPT) with have positive. This little fact leads to NPT IGBTs being parallelable since they share current in a way to minimize temperature difference among them. This process is quite simple. If one NPT IGBT would somehow run hotter than the other, its resistance would increase and therefore the other IGBT would start conducting more current, heating up the second one and cooling down the first until they have reached equilibrium. PT IGBTs on the other hand, when paralleled, will very likely experience thermal runaway of the IGBT with the best conduction at high temperature and hence does not allow paralleling. There is nothing other inherent than the temperature coefficient of PT IGBTs which prevent them from being paralleled, consequently if one could just make sure that their temperature coefficients are well matched and that the actual chips would thermally couple to each other good enough, they could in fact be paralleled with great power savings as a results, compared to NPT IGBTs. The problem with NPT IGBTs is that they are manufactured with extra steps in the factory in order to make the temperature constant positive which have very bad effects on both forward voltage drop and dynamic resistance. If one would still need to use several IGBTs and NPT is not an option, one can always use paralleled converters at the cost of added complexity.

The author did comb though the market for suitable IGBTs in September-October 2007 and 18 candidates where found. Most of them had inconclusive datasheets so some parameters had to be made up with a touch of educated guess. Furthermore, few of them had any information about thermal transient response and for this sake, the one for an APT where used for all and normalized with the thermal resistance for each IGBT.

4.2 Capacitors

4.2.1 Wet aluminium electrolytic

Capacitors come in different sizes and colors with various voltage and capacitance ratings, internal resistance and maximum working temperature. Regardless of how much capacitance is needed, a designer should take the time to understand what current capacitors can and cannot do. When designing with capacitors, there is trade-offs between voltage and capacity, as well as for losses (ESR) and capacity. Starting with voltage, there is a clear relation between energy storage and voltage for a given capacitor size.

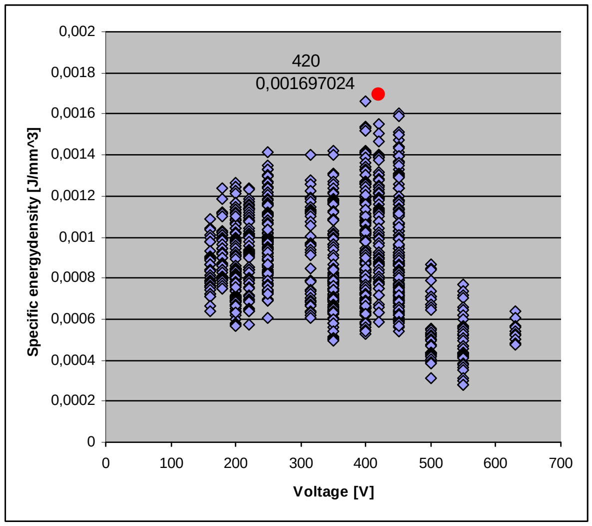


Figure 21: Specific energy density as a function of voltage for various capacitors

From the above chart, we can see that the best commercially available capacitor was at the time rated for 420 V. This could be due to the fact that most PFC front end regulators boost the voltage to about 400 V and hence the demand for good ~400 V capacitors. Above that, it goes steeply down, probably due to some governing property of how modern wet aluminium capacitors are made.

Moving on to capacity, we can see below in figure 22 that for electrolytic capacitors there is an optimum capacitance of 680 μF. This will of course vary with voltage but one should remember to try out different combinations if such options are available.

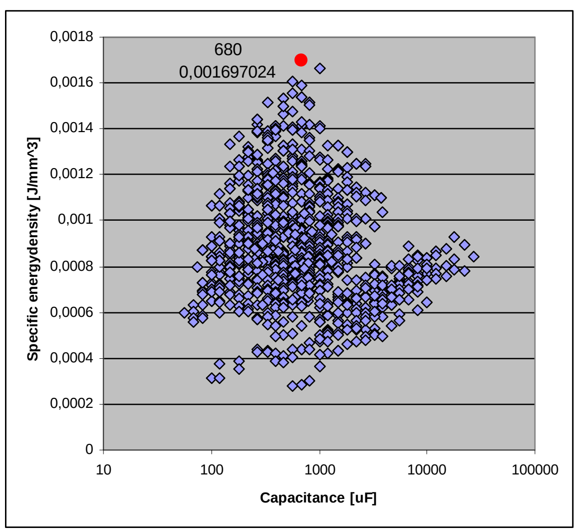


Figure 22: Specific energy density as a function of capacitance

Furthermore, the physical size of the capacitor comes with a price. Very tall or very flat capacitors will cost in terms of energy density as we can see in figure 23. The physical size needed is often quite fixed in existing designs but when made from scratch, one should take into account how stacking capacitors affects performance.

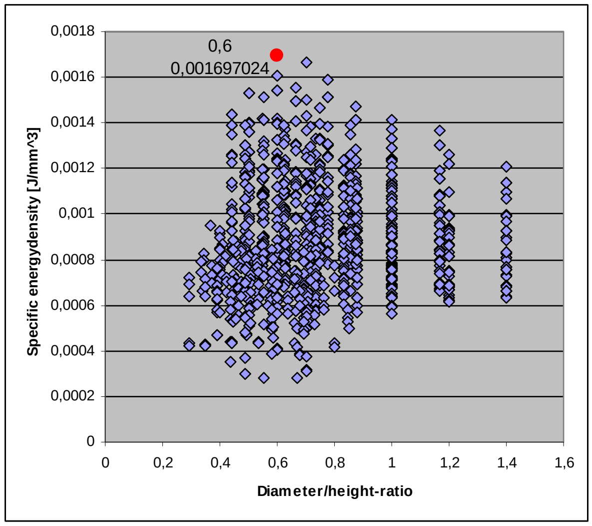


Figure 23: Influence of diameter to height-ratio on energy density for electrolytic capacitors

The other side of the coin is power density. In some cases, a given design might provide enough energy storage whereas ESR will restrict the maximum power level too much with double layer capacitor a prime example. A given series of capacitors will often have the same tan δ for all capacitors so with higher specific energy density comes higher specific power density as well as seen by the quite clear lines in the variation of a Ragone chart in figure 24.

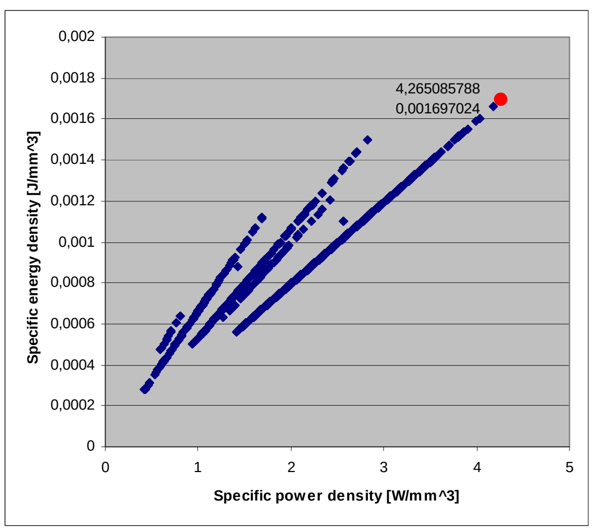


Figure 24: Specific energy density as a function of specific power density

The author expected a bigger trade-off between power and energy density. In the low end rage there exist some trade-off but not in the high end range, which is beneficial for good amplifier designs.

In all four charts the clear winner was at the time Panasonic UQ-series 420 V, 680 µF.

4.2.2 Double layer electrolytic (Gold caps)

A new interesting field for capacitors is double layer caps, also known as gold caps. Instead of the traditional dielectricum, they utilize a layer of activated carbon which gives them far higher capacity at the cost of higher internal resistance and very limited breakdown voltage. Interested readers should look at the references provided.

The problem in using them for an amplifier power supply would mainly be the ESR. For a 20 kW load with the PFC only supplying the average power, using the currently best commercial capacitor and discharging them to 50% voltage, the bank would be about 450*450*70 mm large to be able to supply the power whereas only 65*45*70 mm is needed for the energy to be stored.

With the demand from hybrid car manufacturers for double layer capacitors, the author expects great progress in the field compared to standard electrolytic capacitors.

5 Design

5.1.1 Limitations

Since no PFC or otherwise high powered DC-supply available at the time, the prototype had to be designed for 230 Vac input. This was a major limitation which required a much larger capacitor bank than otherwise needed and due to a lower working voltage, put higher current stress on the main transistors. From an electrical point of view, severe amplitude modulation and extremely low power factor accounted for a calculated 65 Arms input current consumption if run at full power. The input rectifier did for this reason have a lot to put up with.

5.1.2 Scalability

Scalability was one of the demands for X12. This could be realized in several ways but since PT IGBTs seemed the best solution, the most reasonable solution was to use several converters in parallel. This would make it very easy to scale in power since a third converter could be added if more power was needed and one could be dropped if there was demand for less. The converters were run out-of-phase in order to reduce input and output ripple, at little added cost and an overall reduction.

5.1.3 Floorplanning

Since quite bulky components where used for the prototype, some kind of plan had to be thought out in order to make it fit. Several requirements had to be fulfilled due to physical laws. First, the transformers would produce fairly little heat compared to the transistors and their cooling was limited with only the surface area of the transformer itself available so the only possibility was to place them as close to the air intake as possible. The same applied to the output inductors. A common input in the middle with two half-circles around the PCB for the power flow was therefore proposed as seen in figure 25 where the numbers represent each converter.

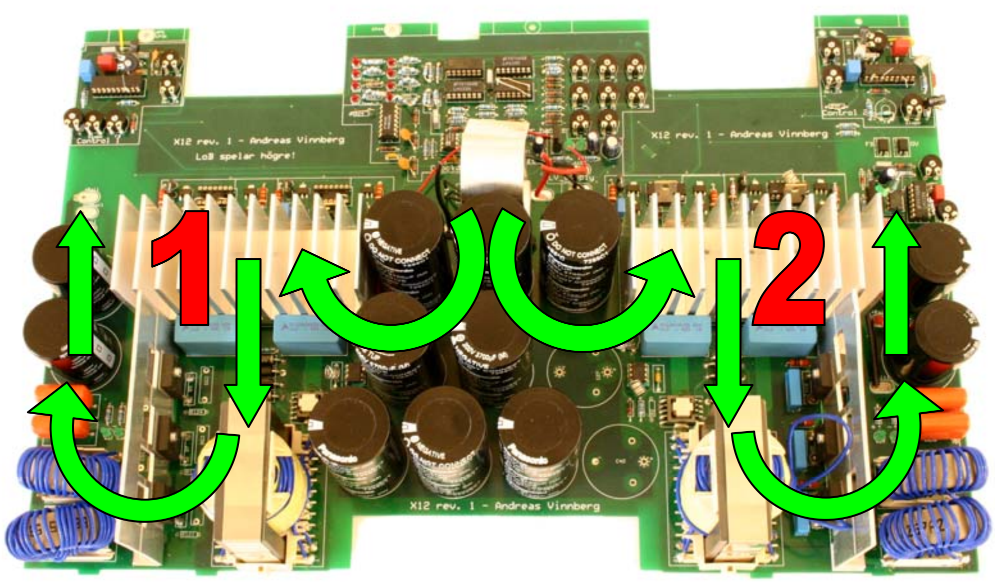


Figure 25: Power flow on the PCB

Next, the leakage inductance between the transformer and transistors needed to be as small as possible and hence placed as close as possible. In retrospect, they should have been placed even closer, but with limited time and a level of decent "prototypeability", the longest path became about 100 mm. The error handler, low voltage supply and clockdrive where of less importance and therefore placed where they seemed fit. The gate drive circuits needed to be placed in close proximity of the main switches and the non-power side (facing the back) of the heatsinks was found a good alternative.

5.1.4 Energy storage

In order to miniaturize the design and to test the concept of peak power and primary energy storage, nearly all capacitance was located on the primary side. Since the converter were to be tested directly from the mains without any PFC, the capacitor bank had to be oversized compared to a boost PFC solution. The energy consumption from the capacitor bank was simulated in SPICE, where the phase of the input voltage compared to the audio burst was tweaked until the most ungrateful situation was found and then series stacks of in-house Panasonic 200 V, 3700 μF capacitors were added until the bank voltage did not drop under half the rectified input voltage. This proved to require ten capacitors. If a PFC solution would have been used, the capacitors could have been charged more often and not just when the input voltage exceeded the bank voltage and the higher voltage would had packed more energy in the bank which would reduce its size. Using 400 V capacitors directly and not two 200 V in series would also have had great benefits as evidenced in section 4.2.1. A picture of the capacitor bank can be seen below in figure 26. Missing in the picture are two capacitors.

Figure 26: Close up of primary capacitors

5.1.5 Transformer design

The transformer sole function in X12 is to provide galvanic isolation and reduce the voltage. In order to do this, several issues needs to be solved namely core selection, bobbin selection and winding configuration. The core selection had several requirements. Saturation had to be avoided so either the window area had to be large enough to provide enough turns to avoid it or the core itself needed to withstand high magnetic flux. Depending on which core that was chosen and configured, the hysteresis losses would be different and this could not cause the core to overheat. Here, either a core with high maximum allowed temperature and/or low thermal impedance would be needed or the losses had to be minimized from the beginning. The windings had to provide a transformer ratio close enough to the specifications, fit in the given winding area, provide enough turns in order to divide the voltage enough not to saturate the core, couple hard enough to keep the leakage inductance within reason and not become hotter than the cooling can take without melting the wires. As we can see, several factors couple to each other and form an equation system. To solve it without any personal preferences or guesses, everything was calculated in a large spreadsheet.

First, thirteen different core materials was selected from the two major manufacturers, EPCOS and Ferroxcube. There are of course several other manufacturers but differences where quite small among them and since Ferroxcube provided the datasheets with the most comprehensive curves and figures for the task, little effort was spent on finding them all. Since the prototype is reaching a midpoint where it is just between thermally and saturation limited, only very high end materials were considered. Then, sixteen different geometries were selected, mainly from the ETD series as they where the most recommended for large transformers. From there 43 different winding ratios was calculated which was within reason from the specified. This gave 8944 combinations to be evaluated. As the losses core losses would be the hardest ones to cool and deal with they had to be calculated and minimized. The losses for a ferrite core is given by the equation

$$P_{fe} = kf^{\alpha} B^{\beta} \qquad [10]$$

where k, α and β are material constants for each core. To find out the three constants, the "Specific power loss as a function of peak flux density with frequency as a parameter" curve was used. As we can se in figure XX, β can in fact be slightly dependent on f as well. In order to limit this report from being a guide in transformer cores, this effect was neglected and minimized by calculating β from the frequency closest to the assumed end frequency of the converter.

Next up to add was saturation and geometry for each core which was given in plain text in the datasheets. From the geometry and number of turns, it could be calculated whether the given wire would fit or not and with saturation data if the core would be saturated or not. Maximum voltage-time product was chosen to the full input voltage and 100 % duty cycle in order to have margins for annoying and spontaneous malfunctions which could potentially blow the main transistors if the current sense would fail and lead to spontaneous combustion which in turn would be annoying. A production version could be made smaller by making sure that either the current loop being tight enough to naturally run the transformer within margins or have/trust the over current protection (to be) fast enough to prevent a saturation situation going havoc. In order to meet safety requirements and keep the complexity down, triple insulated AWG 16 wire was used. If more degrees of freedom would be allowed, one could use a forth dimension for the equation matrix and try various wire/taping configurations here as well.

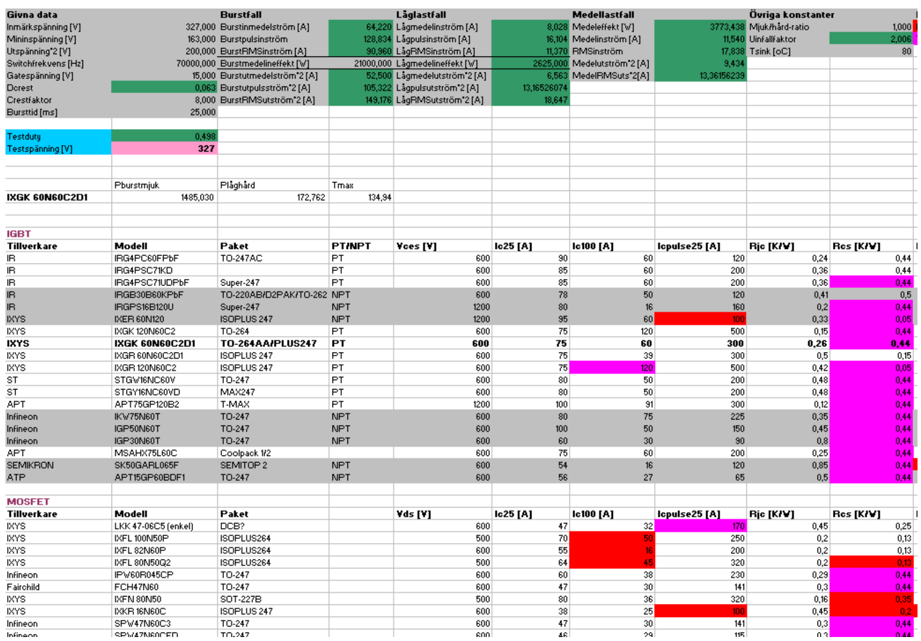
Figure 27: Close-up of main transformer

As for a bobbin to wind on, the German manufacturer Norwe's standing ETD59 was selected as it was the only standing bobbin in ETD59 size the author could find. Standing bobbins seemed unpopular in the industry for reasons unknown, but very beneficial for a 2U-design where PCB-space comes at a premium and the height do not. All this, along with the electrical data from the main circuit of the converter provided 8944 solutions. Most of them could be ruled out early due to saturation or amount of wire not fitting inside the window. The EPCOS N97 in ETD59 size with a turn ratio of 29:2*45 provided the lowest losses. This configuration was however not available at the time from the distributor so Ferroxcubes 3C94 in the same size, which was the second best choice with only a few percent more losses, was selected.

When the actual winding took place, two problems were encountered. To begin with, the author was unable to handwind the transformer hard enough to squeeze in all 119 turns and more alarmingly, the leakage inductance was very high with 5.4 µH. The solution was to revert from the idea to have safety margin for full duty cycle at full voltage and make it more productionlike and assume that any voltage-time area too high for the core will be sensed and solved by the current limiter in the control circuit. After some more discussion, the single primary turn was found to run very hot and a new turn ratio of 2*22 : 2*33 was selected. This would still cope with most over duty-cycle phenomenon. Also, changing the winding layout to trifilar lowered the leakage inductance to 0.75 µH, which was found close to the theoretical value of $(1/2)^N$, where N is number of interleaving sections. As a consequence, the transformer capacitance increased from 120 pF to 383 pF which is normally an additional loss factor but with soft commutation, the added capacity was now closer to what was needed to compensate for the leakage inductance. The assembled transformer can be seen above in figure 27.

5.1.6 Main circuit

Each main circuit consisted of transistors, thermal management and decoupling capacitors. Choosing topology and transistors proved to be a difficult job as there where many factors to consider and the resulting equation system were sometimes underdetermined. The goal was however quite easy to determine, find the optimum topology and transistor combination which could supply the power without breaking and run as cool as possible. To solve the problem, a very large spreadsheet was created with all electrical requirements of the converter, 19 IGBTs and twelve MOSFETs (where the MOSFETs where stacked as single, double and quad configurations, giving 36 configurations), configured as half-bridge, full-bridge and double-full-bridge for IGBTs and half-bridge and full-bridge for MOSFETs; all in all 129 combinations. There would of course be more configurations possible here but the author soon found out that it would not be worth the CPU time calculating them. Selecting IGBTs was more difficult than MOSFETs as differences among IGBTs are easier to cover up for creative datasheet artists than for MOSFETs. Apart from normal voltage rating, thermal resistance, turn on and turn off times, various charges, RdsON for MOSFETs and the exponential voltage drop as a function of current and temperature for IGBTs was calculated manually as they where not given.



Givna data		Burstfall		Låglastfall		Medellastfall		Övriga konstanter	
Inmärkspänning [V]	327,000	Burstinmedelström [A]	64,220	Lågmedelinström [A]	8,028	Medeleffekt [W]	3773,438	Mjuk/hård-ratio	1,000
Mininspänning [V]	163,000	Burstpulsinström	128,834	Lågpulsinström [A]	16,104	Medelinström [A]	11,540	Uinfallfaktor	2,006
Utspänning*2 [V]	200,000	BurstRMSinström [A]	90,960	LågRMSinström [A]	11,370	RMSinström	17,838	Tsink [oC]	80
Switchfrekvens [Hz]	70000,000	Burstmedelineffekt [W]	21000,000	Lågmedelineffekt [W]	2625,000	Medelutström*2 [A]	9,434		
Gatespänning [V]	15,000	Burstutmedelström*2 [A]	52,500	Lågmedelutström*2 [A]	6,563	MedelRMSuts*2[A]	13,36156239		
Dcrest	0,063	Burstutpulsström*2 [A]	105,322	Lågpulsutström*2 [A]	13,16526074				
Crestfaktor	8,000	BurstRMSutström*2 [A]	149,176	LågRMSutström*2 [A]	18,647				
Bursttid [ms]	25,000								
Testduty	0,498								
Testspänning [V]	327								
	Pburstmjuk	Plåghård	Tmax						
IXGK 60N60C2D1	1485,030	172,762	134,94						

IGBT

Tillverkare	Modell	Paket	PT/NPT	Vces [V]	Ic25 [A]	Ic100 [A]	Icpulse25 [A]	Rjc [K/W]	Rcs [K/W]
IR	IRG4PC60FPbF	TO-247AC	PT	600	90	60	120	0,24	0,44
IR	IRG4PSC71KD		PT	600	85	60	200	0,36	0,44
IR	IRG4PSC71UDPbF	Super-247	PT	600	85	60	200	0,36	0,44
IR	IRGB30B60KPbF	TO-220AB/D2PAK/TO-262	NPT	600	78	50	120	0,41	0,5
IR	IRGPS16B120U	Super-247	NPT	1200	80	16	160	0,2	0,44
IXYS	IXER 60N120	ISOPLUS 247	NPT	1200	95	60	100	0,33	0,05
IXYS	IXGK 120N60C2	TO-264	PT	600	75	120	500	0,15	0,44
IXYS	**IXGK 60N60C2D1**	**TO-264AA/PLUS247**	**PT**	**600**	**75**	**60**	**300**	**0,26**	**0,44**
IXYS	IXGR 60N60C2D1	ISOPLUS 247	PT	600	75	39	300	0,5	0,15
IXYS	IXGR 120N60C2	ISOPLUS 247	PT	600	75	120	500	0,42	0,05
ST	STGW16NC60V	TO-247	PT	600	80	50	200	0,48	0,44
ST	STGY16NC60VD	MAX247	PT	600	80	50	200	0,48	0,44
APT	APT75GP120B2	T-MAX	PT	1200	100	91	300	0,12	0,44
Infineon	IKW75N60T	TO-247	NPT	600	80	75	225	0,35	0,44
Infineon	IGP50N60T	TO-247	NPT	600	100	50	150	0,45	0,44
Infineon	IGP30N60T	TO-247	NPT	600	60	30	90	0,8	0,44
APT	MSAHX75L60C	Coolpack 1/2		600	75	60	200	0,25	0,44
SEMIKRON	SK50GARL065F	SEMITOP 2	NPT	600	54	16	120	0,85	0,44
ATP	APT15GP60BDF1	TO-247	NPT	600	56	27	65	0,5	0,44

MOSFET

Tillverkare	Modell	Paket		Vds [V]	Ic25 [A]	Ic100 [A]	Icpulse25 [A]	Rjc [K/W]	Rcs [K/W]
IXYS	LKK 47-06C5 (enkel)	DCB?		600	47	32	170	0,45	0,25
IXYS	IXFL 100N50P	ISOPLUS264		500	70	50	250	0,2	0,13
IXYS	IXFL 82N60P	ISOPLUS264		600	55	16	200	0,2	0,13
IXYS	IXFL 80N50Q2	ISOPLUS264		500	64	45	320	0,2	0,13
Infineon	IPW60R045CP	TO-247		600	60	38	230	0,29	0,44
Fairchild	FCH47N60	TO-247		600	47	30	141	0,3	0,44
IXYS	IXFN 80N50	SOT-227B		500	80	36	320	0,16	0,35
IXYS	IXKR 16N60C	ISOPLUS 247		600	38	25	100	0,45	0,2
Infineon	SPW47N60C3	TO-247		600	47	30	141	0,3	0,44
Infineon	SPW47N60CFD	TO-247		600	46	29	115	0,3	0,44

Figure 28: Screendump from spreadsheet calculation of different transistor/topology combinations

From there, each effect contributing to the total losses were calculated for different load conditions and summed up. A thermal impedance model for one of the IGBT was used for calculating the maximum temperature rise during the burst. Different IGBTs do of course have different thermal impedance responses but the manufacturers considered the importance of putting it in the datasheets differently and the thermal model was assumed to be near identical and only normalized with the thermal resistance for each IGBT. As the MOSFETs suffer from I^2R-losses, they proved very unsuited for peak power conversion and no thermal impedance model was extort for them.

When all parameters where given, the only real variable to tweak was frequency, which was adjusted until the best topology-transistor combo had reached just under the maximum allowed temperature. The best solution from the chart partly displayed in figure 28 above was a double full-bridge using IXYS IXGR60N60C2D1 IGBTs at 70 kHz which also became the solution used in X12. The transistors mounted on their heatsink can be seen in figure 29 below.

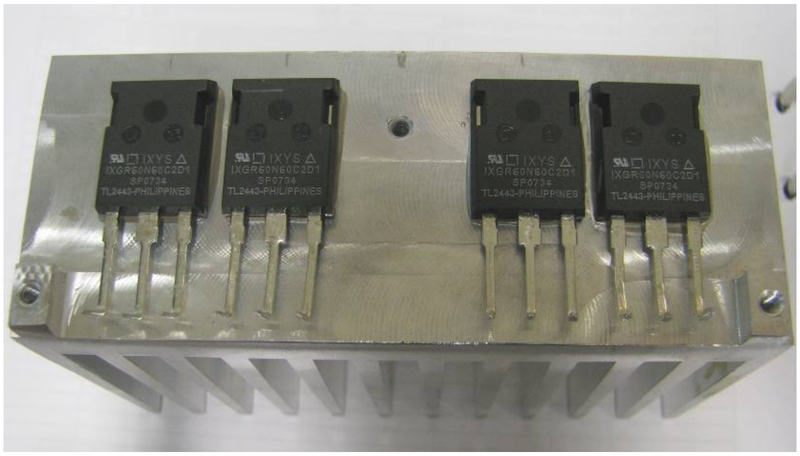


Figure 29: Close-up of main transistors

As for thermal management, it was a question of a press contact using thermal grease or gluing since TO-264 capsules did not fit onto existing heatsinks so holeless TO-247 had to be used. An important reason why the transistor in question won, was because of its thermal structure using internal isolation. Although this gives a higher junction to case thermal impedance, the total junction to sink thermal impedance was substantially lower than using the best uninsulated transistor and best available isolation pad. Interested readers are referred to the document [7] found in the references. Gluing was preferred since a very thin layer could be used and the best available glue to the authors' knowledge had very high thermal conductivity, over 7.5 K/mK, which gives about 4 K temperature difference during max load bursts. The result can be seen in figure 30.

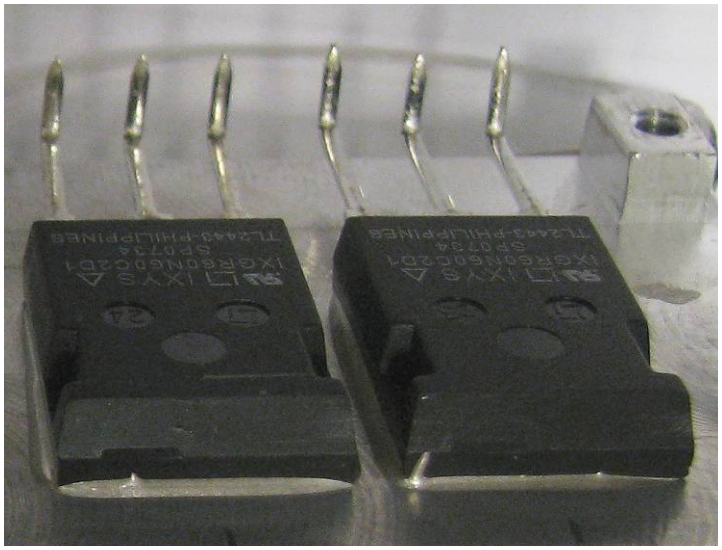


Figure 30: Transistors glued in place on their heatsink

Decoupling was done using one EPCOS 2.2 uF polypropylene capacitor per leg. There were virtually no differences among manufacturers here. They where placed as close to the transistors as the heatsink allowed to in order to reduce stray inductance.

5.1.7 Output stage

The output stage of X12 consists of rectification, inductance and capacitance. To begin with rectification, it had effects on several other parts of the converter so they had to be considered. As the transformer is assumed to be thermally limited or close to, freewheeling of the secondary current though it had to be designed away. With the problem of the output inductors suffering very much in size for a peak power converter, the ripple had to be minimized in the rectification stage as well in order to keep the inductor size down. These two facts called for a six diode solution to minimize all ill-effect at the cost of number of diodes. Nowadays when EMI is a major concern designers look for quiet diodes which turns of gently with low di/dt and therefore low EMI emissions. Just about all recently marketed diodes says “quiet”, “soft recovery”, “ultrasoft recovery” and so on with some manufacturers measuring how soft or quiet in various ways. As this was beyond the scope of the thesis, a range of fast, low drop and low resistance diodes with low thermal resistance and marked as “quiet”/”soft” was picked out. As for the selecting the actual diodes, an again large spreadsheet was created for the candidates where power loss and temperature rise for various conditions where calculated. The temperature rise was about 30 K for the best candidate which is much less than needed but since this was a prototype where proving the concept where more important than cost, state of the art IXYS DSEI120-12A was selected as it was the one with the least losses that could cope with both current and voltage stresses which some degree of safety margin of the considered candidates.

The output inductors are flux density limited rather than thermally limited so very high flux cores were selected. Output inductors are very much affected by the selection of primary/secondary storage and hence peak power as they have to stay unsaturated at all times. There was a trade-off between transistor losses and inductor size here as with just about any design and as the peak current problem already led to large inductors, XX% current ripple was chosen to design for. For this reason, Magnetics Inc. 58257 HiFlux cores where used with 34 turns of wire, providing 26 μH. There would be some great benefits [8] in winding all four inductors on one core such as cross-regulation and total size reduction but with the drawback of having one quite large core where all four outputs needs to physically connect to and the problem of placing one large item on the PCB rather than two or four small ones. Again, making an optimum layout was not the task so four separate ones where chosen. Interested readers are encouraged to experiment with coupled inductors.

As for output capacitors, two different types where used. First, one polypropylene capacitor was used for each rail and converter to filter out high frequency ripple and secondly there was a high power density electrolytic capacitor to take care of faster disturbances. In figure 31 below we can see the different components which make up the output stage.

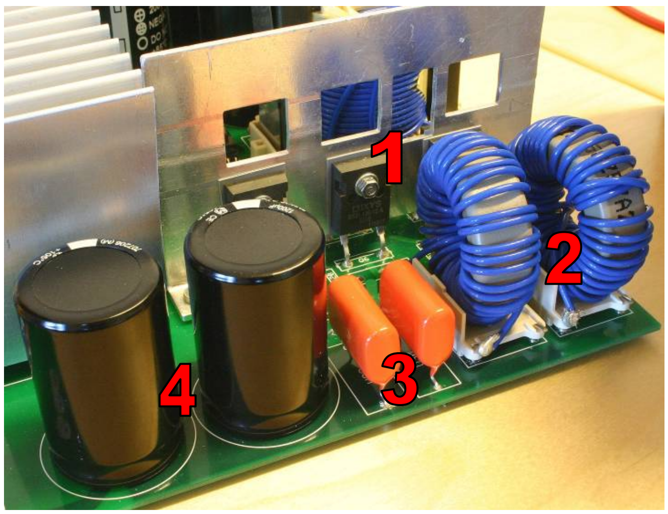


Figure 31: Close-up of output stage

1 is the six diodes mounted on a heatsink (only one of them is fully visible in the picture)
2 is the two output inductors
3 is the polypropylene capacitors for high frequency filtering
4 is the electrolytic capacitors for low frequency filtering

5.1.8 Gate drive

Since the outputs from the regulation circuit can only provide 400 mA which would lead to slow turn on and offs and thus high switching losses, a separate drive stage was implemented for driving the gates of the IGBTs. Furthermore, the high side IGBT in each leg needs its gate offset to the legs center voltage, which can vary rapidly. This was done by using a high and low side gate drive circuit followed by a fast high current BJT complimentary pair. Since the gate drive has high dv/dt and di/dt in its nature, the drives stages power supply was heavily decoupled and filtered though beads and no plane was used for reference and instead separate paths for each collector where used. A picture of the gate drive circuit is provided in figure 32 where we can see:

1 is the IR2113 gate drive circuit,
2 is the two complementary BJT pairs
3 are the gate resistors
4 is the high speed diode for the high side bootstrap
5 is a “pigtail” for high frequency measurement
Only partly seen in the picture is the bead on the diode leg for EMI reduction.

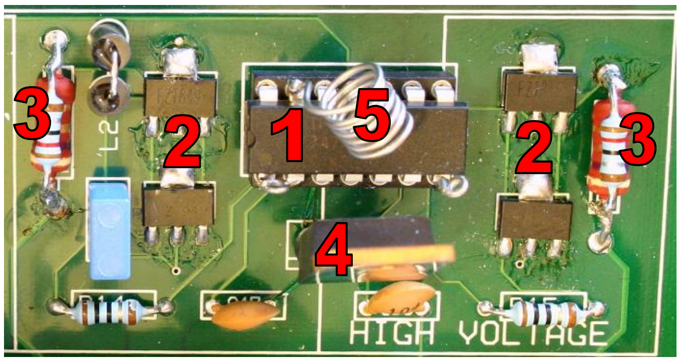


Figure 32: Close-up of gate drive circuitry

5.1.9 Error handling

Since there where several different failure modes for the converter, a number of safety features was implemented. At high power levels, temperature is a potent killer for semiconductors so precision temperature sensors where placed on the presumed hottest part of the heatsinks. In a shorter time perspective, overcurrent is the major criminal for killing the mentioned semiconductors. The control IC used had built in protection for overcurrent with two preset safety levels for the current sense. At 2 V on the current input, it would cycle by cycle current limit the regulator by dropping the duty cycle to 0 % until the overcurrent sense sensed under 2 V. At 2.5 V, the IC would shutdown the regulator and do a full soft start cycle. Since the current sense is located in series with the transformer, any crossconduction in each leg would go undetected but the control IC's internal dead time was assumed to be adequate.

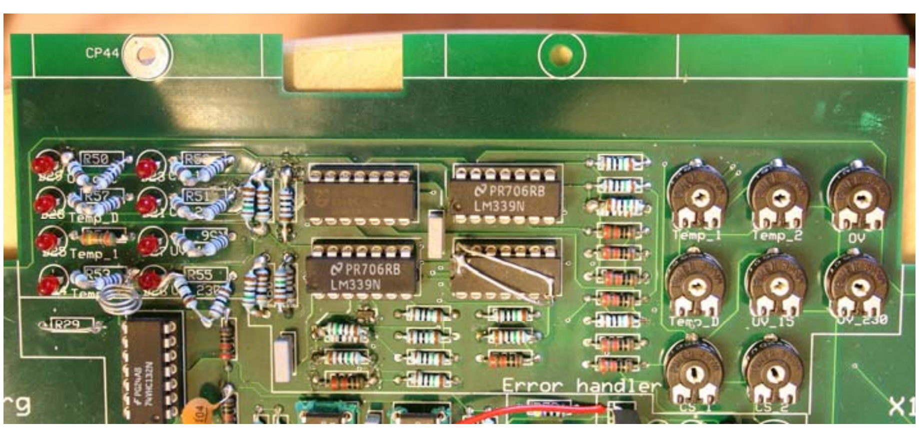


Figure 33: Close-up of error handler

Overvoltage was detected on both input and output by a simple comparator circuit with the addition of an optocoupler for the output to keep the galvanic isolation. Undervoltage was in a similar manner implemented for the input and low voltage supply.

All detected failure modes was logically “ORed” together and the output signal was used to shut down all gate drive circuits and dump the soft start capacitor in order to force the control IC to shut down and standby until the failure had disappeared. A picture of the error handler is provided in figure 33.

5.1.10 Control

In order to control the converter, several things are needed. First, some kind of PWM circuit is needed in order to switch the semiconductors and secondly a feedback loop is needed to compensate for load changes, input voltage variations and other things affecting the output voltage.

The first part was realized with two Texas Instruments UC3879 circuits. It had all the features needed (adjustable dead time, current control, phase shifted operation and cycle-by-cycle current limiting), was the most well documented of the candidates and its predecessor was widely used in the industry. As for feedback, a traditional operational amplifier with PI effect on the secondary side and an optocoupler for connecting to the primary side was opted for. Since the regulation also consisted of an inner current control loop, two current transformer or possibly resistors would be needed on the primary side, one for each converter. As the current would reach over 100 A peak for each converter, a current transformer was the only solution and for this, Pulse engineering PB0027NL was opted for. The transformer output was terminated by a resistor and connected to an operational amplifier for adjusting the voltage to an appropriate level and the signal was low-pass filtered in order to get rid of switching ripple.

6 Conclusions

6.1.1 Testing

With an input voltage of 70.8 volt and output of 145.7 V, 1650 W of burst power from one converter was achieved with a load regulation error of about 2.5 %. This was 16.5 % of the designed power and five times higher load regulation error than designed for. However, the author finds this as a proof of the concept – primary energy storage is a viable solution for size miniaturization and is overall cost effective.

6.1.2 Problems

The testing of the prototype proved harder than expected. Many sections could be run separately, stimulated from external sources or run while the converter itself was turned off but several had to interact in order to work. Since the full voltage could not just be applied but rather ramped up, several protection mechanisms had to be reconfigured or bypassed in the startup process. The output rectifier diodes caused a lot of interference which disturbed the various measurements used for regulation which was solved by snubbing the diodes and placing beads on their legs for HF-suppression. This reduced the problems a lot but the design was still not entirely stable. HF-suppression on the high side gate drive diodes as well reduced the problems slightly. In order to get the prototype up and running the feedback loop speed was reduced and the slope compensation was set for far more voltage control than initially planned for. This was a somewhat viable solution as long as the output/input remained fairly constant.

6.1.3 Lessons learned

Groundplane, EMI

7 Glossary

BJT –	Bipolar Junction Transistor, a type of transistor
MOSFET –	Metal-Oxide Silicon Field Effect Transistor, a type of transistor
IGBT –	Insulated Gate Bipolar Transistor, a type of transistor
Program power –	Measurement method for audio amplifiers which exists in many forms, where AES and EIAJ are the most commonly used. Usually consists of a burst of power, 5-30 ms long follows by a silent or lower amplitude “pause” for 20-400 ms, which is repeated until the amplifier have reached equilibrium.

1. miniBLOC, the polyvalent power package, IXYS 2002
2. Mohan, Undeland, Robbins, Power Electronics, p 310

3. Active clamp and reset technique enhaces forward converter performance, Bill Andreycak, October 1994, http://focus.ti.com/lit/ml/slup108/slup108.pdf
4. London Metal Exchange (http://www.lme.co.uk/copper_graphs.asp), retrieved on 2008-05-15
5. London Metal Exchange (http://www.lme.co.uk/aluminium_graphs.asp), retrieved on 2008-05-15
6. Wikipedia, http://en.wikipedia.org/wiki/Transistor, retrieved on 2008-04-14
7. ISOPLUS™ - The Revolution in Discrete Isolation Technique, http://ixdev.ixys.com/DataSheet/e9d94d3c-3573-44d2-b066-00bb75832e81.pdf
8. Coupled filter inductors in multi-output buck regulators, Lloyd Dixon, http://focus.ti.com/lit/ml/slup082a/slup082a.pdf