SOME DESIGN CONSIDERATIONS FOR THE 6.5 KV IGBT-BASED HALF-BRIDGE DC/DC CONVERTER

Dmitri Vinnikov
Department of Electrical Drives and Power Electronics, Tallinn University of Technology
Ehitajate tee 5, Tallinn, 19086, Estonia. Tel. (+372) 6203705, Fax. (+372) 6203701
dm.vin@mail.ee

Abstract – **The 6.5 kV insulated gate bipolar transistor (IGBT) is a new promising power device for high-voltage high-power applications. Due to its increased voltage blocking capability, the new HV IGBT can serve as a good replacement for GTO and IGCT thyristors in medium or medium-to-high power applications. One of its general application fields is railway traction. The paper discusses design and development considerations of the 50-kW auxiliary power supply to be used in 3.0 kV DC commuter trains. Focus is on the implementation of 6.5 kV IGBTs in the primary inverter of APS to improve power density, integrity and reliability of the whole system. For instance, analysis and simulation of main functional blocks of a converter for different operation points are described. General design considerations mainly concerning the inverter stage, power transformer and output rectifier are specified in the paper. Some design optimization possibilities are proposed.**

Keywords - **DC/DC converter, design, efficiency, high frequency transformer, power converter**

I. INTRODUCTION

The 6.5 kV IGBT modules (EUPEC, ABB, IXYS, DYNEX, etc.) recently implemented are basically designed for 3.0 kV DC rolling stock applications with their high demands on reliability concerning thermal cycling capability. A single IGBT has the voltage blocking capability two times the nominal catenary voltage level, which copes with the requirements for the rolling stock power electronics. Such transistors give an attractive possibility of using a simple 2-level inverter topology, achieving better efficiency, power density and reliability compared to the older designs.

Today's state-of-the-art 6.5 kV IGBT modules are available in three basic configurations: with 200 A, 400 A and 600 A collector current capabilities. While for the traction drives with their MW-power ranges it is essential to use the parallel connection of HV-IGBTs, for such middle-power applications like onboard auxiliary power supply (so-called static voltage converters), the use of single 6.5 kV IGBT modules becomes economically feasible.

Basically, onboard auxiliary power supply (APS) is the power interface between the main supply system of the railway vehicle (catenary) and the low-voltage secondary onboard systems (lighting system, braking system, compressors, blowers, air conditioning system, etc.). According to the rugged railway norms and requirements, the main technical objective of an APS can be formulated as follows: a smooth operation despite voltage fluctuations on the primary (2000...4000 V DC, [1]) and variation of loads (40...150 A) on its secondary side. Taking into account such end-user criteria like maintainability, availability and reliability, power circuit simplification by means of new 6.5 kV IGBTs seems to be very attractive.

II. OPERATION LIMITS FOR 6.5 KV IGBTs

Despite the described advantages of 6.5 kV IGBT transistors in the primary inverter of the catenary fed APS, the overall design of it involves several limitations. They are mostly related to the specific properties of 6.5 kV transistors and need to be taken into account during the development routine. The switching dynamics of IGBT generally depends on such parameters as stray inductance and parasitic capacitance of the module as well as on the resistance of the gating circuit. The average per pulse turn-on (E_{on}) and turn-off (E_{off}) energy losses of IGBTs with the different voltage blocking capabilities in the hard-switching mode are compared in Fig. 1.

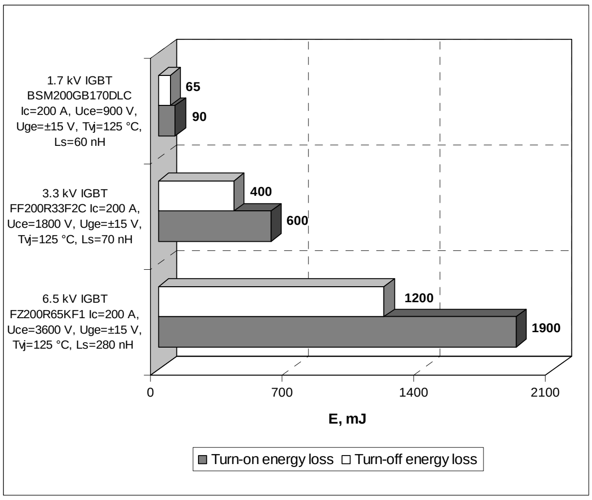
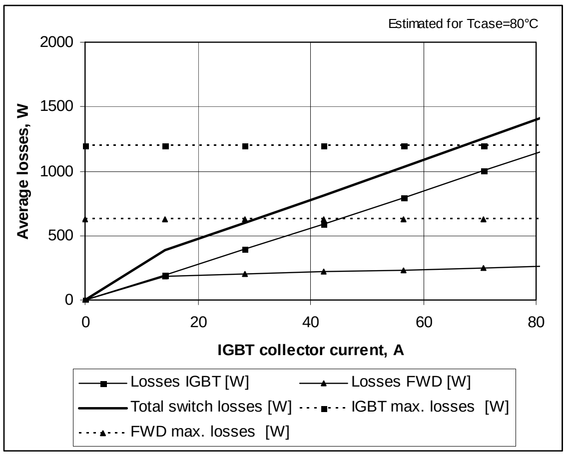


Fig. 1. Average per pulse turn-on and turn-off energy losses of different IGBTs (produced by EUPEC).

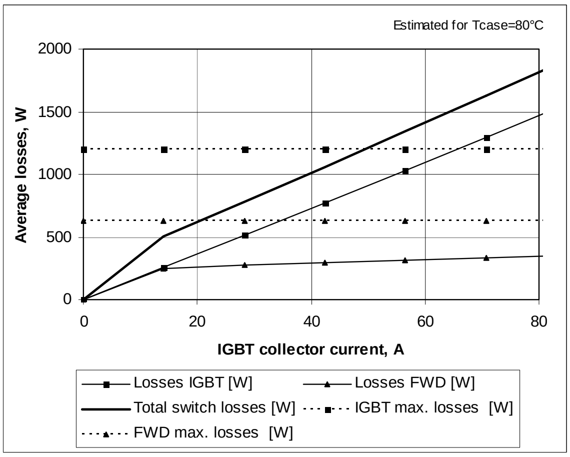
As it seen from Fig. 1, the switching energy losses of the 6.5 kV IGBT are dramatically increased compared to the 1.7 kV IGBT and even to another representative of HV-IGBT class - the 3.3 kV IGBT. The typical turn-on transient of the 200 A 6.5 kV IGBT takes about 1.1 μs. Since hard turn-on transients are basically determined by the turn-on transient of the IGBT internal MOSFET the occurring switching times, the *dI/dt*, and the *dU/dt* can be adjusted by the gate drive [2]. The minimum gate resistances for turn-on transients depends essentially on the DC-link voltage and the stray inductances of the circuit.

Slower current rise and higher DC-link voltage result in a nearly negligible voltage drop for the 6.5 kV IGBT. Therefore, the turn-on loss of the 6.5 kV IGBT is relatively high, but the voltage spike at the freewheeling diode is very low. Actually, the IGBT turn-on speed is limited by the peak power of the diode. Comparing the single 6.5 kV IGBT with *n*-series connected low-voltage IGBTs (series combination of low-voltage IGBTs for the desired voltage blocking capability is the topology of choice of the recent rolling stock converter designers), the lower current droop for a 6.5 kV device results in less stored energy in the stray inductance and a smaller relative overvoltage at the turn-off.

Basically, the operating frequency of the switched-mode power supply should be selected to obtain the best balance between switching losses, total transformer losses, size and cost of magnetic components and input/output capacitors. Since 6.5 kV IGBTs have relatively high switching losses, the switching frequency must be carefully adjusted. Fig. 2 demonstrates average switching loss curves for the 200 A 6.5 kV IGBT module (EUPEC FZ200R65KF1) with the different switching frequencies and collector currents, which have been produced by the *Iposim 6.0* software. The maximum permissible currents for IGBT and FW diode can be read at the crossing of the corresponding average losses curves with the maximum losses curves (dotted lines). Analyzing losses for the different switching frequencies, the increased switching frequency has the result of the 10 % decrease in ruggedness limit for the selected IGBT type. Further, increasing switching frequency from 1 kHz to 2 kHz we will have the result of 30 %-increased switching losses.



(a)



(b)

Fig. 2. Average losses of 200 A 6.5 kV IGBT with the different switching frequencies: 1 kHz (a) and 2 kHz (b).

The bottom line is that the switching dynamics of the 6.5 kV IGBT is significantly reduced due to the effect of the absolute voltage vs. current. However, switching can be controlled to achieve a good compromise between allowable switching frequency and switching power loss.

III. POWER CIRCUIT DESIGN

For the APS with output power up to hundred kW, the conventional half-bridge topology and two 200 A 6.5 kV IGBTs (one for the TOP and one for the BOT switch) is sufficient to fulfill all the design requirements (Fig. 3).

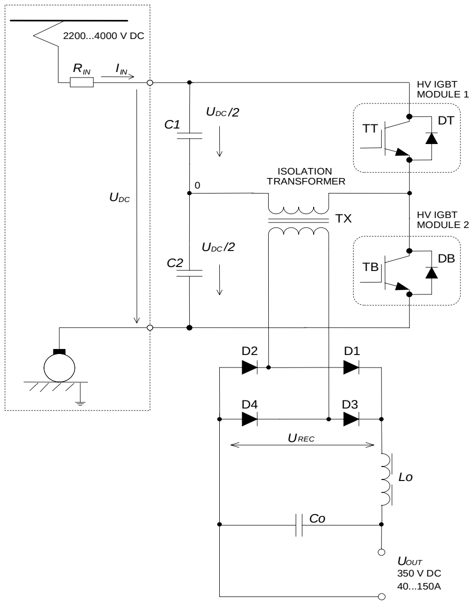


Fig. 3. Power circuit layout of the converter.

The half-bridge DC/DC converter configuration consists of two equal capacitors *C1* and *C2* connected in series across the DC input voltage source, providing a constant potential of one-half U_{DC} at their junction (Fig. 3). [3] The HV IGBT switches TT (top switch) and TB (bottom switch) are turned on alternately and are subjected to a voltage stress equal to that of the input voltage. Concerning another two-switch DC/DC converter topology, a double-ended forward converter, an andvantage of the half-bridge is that its secondary produces a full-wave output rather than a half-wave output. Thus, the square-wave frequency in the half-bridge converter is twice that of the forward converter, and the associated output filter components will be smaller. In addition, considering specific railway demands for better reliability and maintainability of the power electronic converters, the half-brigde topology with its reduced component number looks like more preferable choice compared to the full-bridge topology, for instance.

A. Primary Inverter

The input side of the converter is assumed to be connected directly to the traction supply grid with the voltage tolerances from 2200 V DC ($U_{DC(min)}$) up to 4000 V DC ($U_{DC(max)}$). The most demanding operation point is at the minimum input voltage and at the rated load conditions (i.e., maximum duty cycle operation). It is essential to prevent even short-time simultaneous conduction of TT and TB switches in these demanding conditions - it leads to the short circuit across the supply voltage and to the destruction of the converter. It means that the maximum on-state time $t_{on(max)}$ of each switch in the half-bridge must be set at 80% of a half-period to ensure that this does not happen. Pulse width at the maximum input voltage is minimal and may be determined as

$$t_{on(\min)} = \frac{U_{DC(\min)}}{U_{DC(\max)}} t_{on(\max)} = \frac{2200}{4000} t_{on(\max)} = 0.44\left(\frac{T_{sw}}{2}\right) \quad (1)$$

Operation voltage ranges and inverter switch on-state times of the half-bridge topology in this application are presented in Table I. Simulated voltage and current waveforms of primary inverter IGBTs with the different input voltages and at rated load are presented in Fig. 4.

TABLE I
Operation voltage ranges vs. inverter switch on-state time

Converter input voltage U_{DC}	*2200 V*	*4000 V*
Switch on-state time t_{on}	$0.8(T_{sw}/2)$	$0.44(T_{sw}/2)$

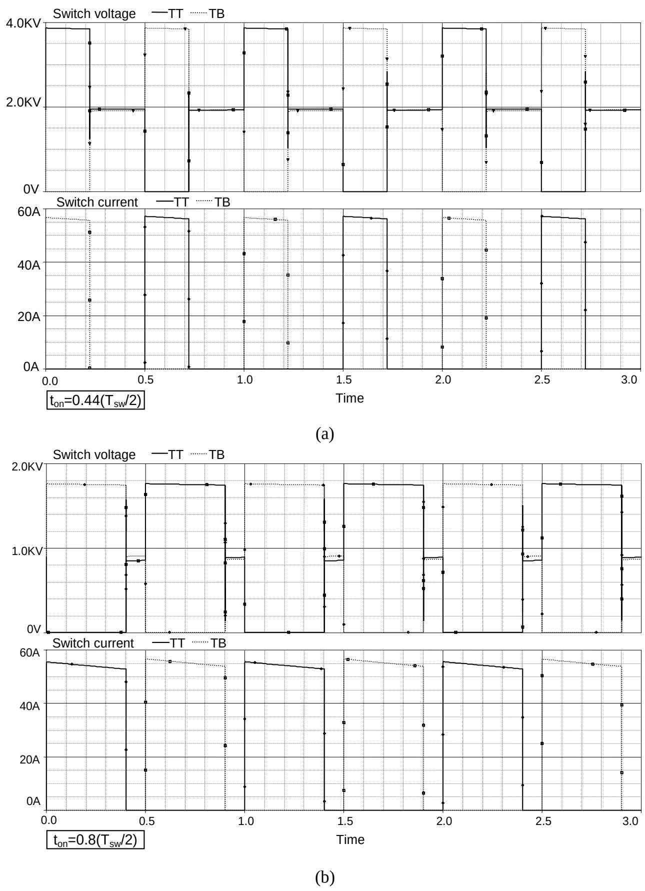


Fig. 4. Inverter switch voltage and current waveforms for minimum (a) and maximum (b) duty cycle operation.

In the half-bridge topology the midpoint of input capacitors charges up to an average potential of $U_{DC}/2$. Thus, IGBT peak collector current (neglecting transients) can be described as

$$I_{Cpeak} \geq \frac{2 \cdot P}{\eta \cdot U_{DC(\min)}}, \quad (2)$$

where P is converter output power and η is converter efficiency. Thus, with the twice the collector current for the half-bridge (compared to the full-bridge topology) the utilization of 200A 6.5 kV IGBT switch seems more optimal.

In the design of switchmode converters with the high-voltage IGBTs, the determination of switching frequency should be considered in practical realization including size and weight of passive components (inductors, isolation transformer, capacitors) and, other hand, cost, size and availability of cooling system for primary switches. For low and low-to-middle power applications with 6.5 kV IGBTs, a switching frequency around 1.0...1.5 kHz is a preferred choice in real design. Fig. 5 gives an overwiev of total switch losses of the 50-kW half-bridge inverter based on the 200 A 6.5 kV IGBTs. The approximation was done for hard-switching mode with the maximum converter input voltage (4000 V DC) and at rated load. Remarkably, the total switch losses at frequencies higher than 500 Hz are mostly caused by the dynamic losses of IGBT and frewheeling diode.

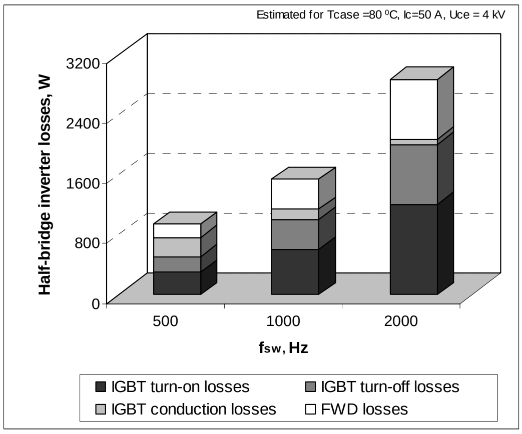


Fig. 5. Average losses of half-bridge inverter (2 x IGBT FZ200R65KF1) at different switching frequencies.

With the dramatically increased power dissipation of HV IGBTs the only high-performance heatsinks should be used for the improvement of converter dimensions. Fig. 6 demonstrates a comparison between general specifications of the different cooling tehniques. For the power loss range of 1000-2000 W the hollow-fin cooling aggregates with blowers seems to be the most attractive solution. In spite of slightly decreased cooling properties the forced cooling about three times outpace the fluid coolers in terms of cost and complexity.

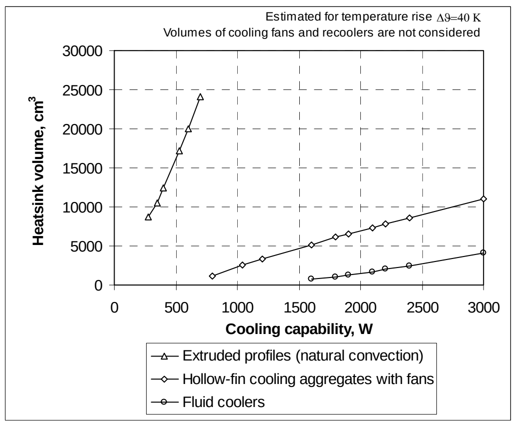


Fig. 6. Cooling capability comparison of different cooling techniques.

For this project 1 kHz switching frequency was implemented in the primary HV IGBT inverter. Hollow-fin cooling aggregate with the radial blower and improved fin structure was used for the cooling of HV IGBTs. In the selected configuration its nominal cooling capability is 2500 W with the temperature rise $\Delta\vartheta$ of 40 K. Fig. 7 shows developed 50 kW HV IGBT inverter assembly with the associated control and measurement circuits.

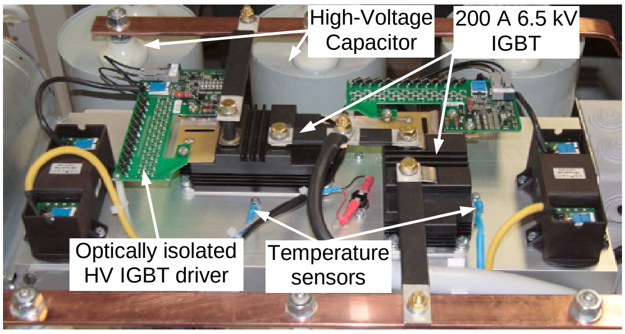


Fig. 7. High-voltage half-bridge inverter assembly.

B. Isolation Transformer

The isolation transformer of a rolling stock APS is responsible for providing the necessary I/O galvanic insulation required by special safety norms. For the 3.0 kV DC-supplied converters the required isolation barrier is 10.2 kV/1min. Due to the capacitors providing a mid-voltage point, the isolation transformer during the operation sees a positive and negative voltage with the amplitude value of only half the input voltage (i.e., 1100...2000 V, see Fig. 8). This results in twice the desired peak flux value of the core, because the transformer core is operated in the first and third quadrant of the B-H loop and experiences twice the flux excursion of a similar forward converter core. This is an advantage of the half-bridge topology over the double-ended forward topologies, where the half-bridge primary transformer winding has half the turns for the same input voltage and power. Being compared to full-bridge transformer the half-bridge one still has the result of half the primary turns for the same input voltage and power.

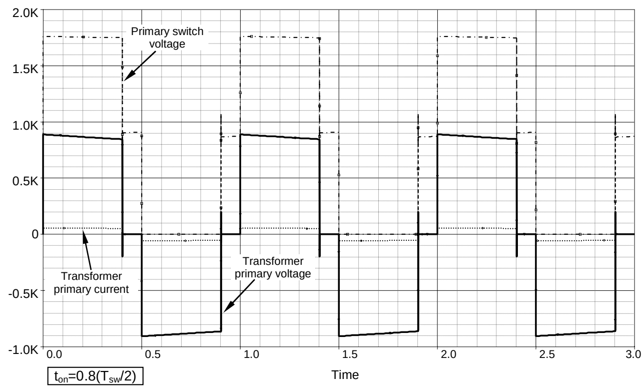


Fig. 8. Simulated isolation transformer primary voltage and current waveforms for the maximum duty cycle operation.

With the comparably low switching frequencies available with the 6.5 kV IGBTs the isolation transformers in such applications can be regarded as the bulkiest components in the whole converter stack. For the such operation conditions (i.e., 500...2000 Hz) the higher possible values of permeability (μ_H) and saturation magnetic induction (B_{max}) of the magnetic core material is essential, because it gives an additional advantage of transformer weight-space optimization. Table II provides a comparison of the different soft magnetic materials. The basic field of choice in such operation frequency range is the iron steel and the Permalloy. However, the last one despite the good magnetic properties has a drawback of increased magnetostriction. Another drawback of Permalloy type alloys is associated with their softness in that flat particles are liable to deform by stresses induced during milling to form a coating composition, also resulting in a loss of magnetic properties. Ferrites are not considered for such high-power low-frequency operation because of their relatively low permeability and saturation magnetic induction.

TABLE II
Magnetic properties of some soft magnetic materials

	B_{max}, T	μ_H	μ_{max}
Permalloy	0.70...0.75	14000...50000	60000...300000
Silicon steel	2	200...600	3000...8000
Ferrites	0.18...0.40	100...6000	3000...10000
Iron	2.16	250	7000
Gammamet	0.8...1.12	7000...20000	40000...600000

Considering the design, in this demanding application, toroidal transformers outpace the laminated ones. But using the toroidal-form transformers for high power densities, the minimization of core losses by design optimization and material development is important. For this project, the Gammamet toroidal magnetic cores, made from 25 µm thick ribbon of soft magnetic nanocrystalline alloy on Fe-basis were investigated. The real advantages of the selected magnetic cores Gm14DC are their high initial permeability and very low core losses [4], [5]. The rate of specific core power dissipation for the selected core type (Gm14DC) can be found by

$$P'_{D(core)} = 0.75 f^{1.4} B_M^{1.7} \tag{3}$$

Fig. 9 gives an overview of core volume required and core losses for a 50 kW isolation transformer with the Gammamet Gm14DC core estimated for the different operation frequencies.

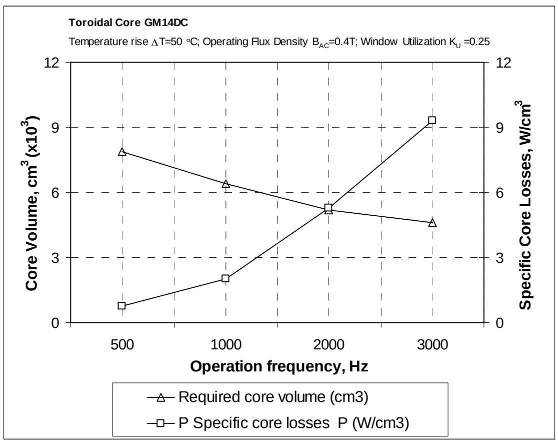


Fig. 9. Average core volume and core losses of 50 kW isolation transformer for the specified operation frequency range.

Fig. 10 demonstrates isolation transformer power loss distribution (50 kW isolation transformer with the Gammamet Gm14DC core) for the different operation frequencies. With the assumed maximum flux density of 0.6 T the average core power dissipation for the 1 kHz 50 kW isolation transformer is about 30 W. Estimated winding losses for this particular design (special litz wire was used in both windings) are not higher than 150 W. With the total power dissipation of 165 W the efficiency of developed 1 kHz 50 kW transformer is 99.6 %, which is quite good result for the such demanding application.

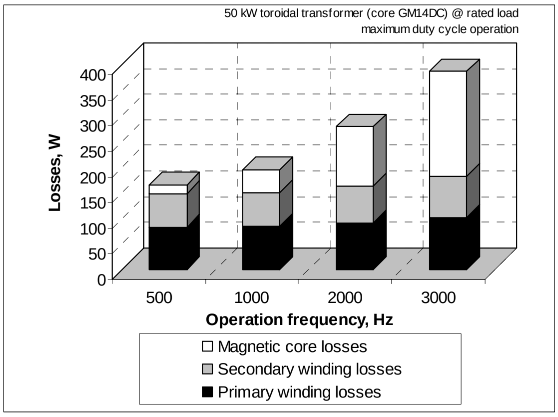


Fig. 10. Isolation transformer power loss distribution for the different operation frequencies.

Leakage inductance spikes, which are so troublesome in some switchmode converter designs, do not exist in the half-fridge topology. This is so because any such spikes become clamped to U_{DC} by integrated frewheeling diodes of HV IGBT transistors. Anyway, toroidal transformers, properly wound, must have all windings distributed uniformly around the entire core. Thus the winding breadth is essentially the circumference of the core, resulting in the lowest possible leakage inductance. Fig. 11 shows the developed 1 kHz 50 kW isolation transformer with the Gm14DC toroidal core. The external diameter and the height of the developed transformer are 440 mm and 140 mm respectively. Estimated total leakage inductance is not higher than 1 μH.

Fig. 11. Developed 1 kHz 50 kW toroidal transformer.

C. Output Rectifier

The output rectifier diodes for the converters with such a wide input voltage range must be selected to satisfy the two basic criteria:

- maximum possible repetitive reverse voltage U_{RRM} (operation with high input voltage level is considered here),
- maximum possible average forward current I_{FAVM} and minimized forward voltage drop, U_F (considering operation with minimal input voltage and maximal duty cycle).

Half-bridge converter output voltage defining equation [6], [7]

$$U_{OUT} = U_{REC} \cdot \frac{2 \cdot t_{on}}{T_{sw}}. \tag{4}$$

Thus, rectifier output voltage amplitude (U_{REC} in Fig. 3) can be estimated as

$$U_{REC} = U_{OUT} \cdot \frac{T_{sw}}{2 \cdot t_{on}}. \tag{5}$$

For the described application, the converter output voltage U_{OUT} must be carefully regulated to 350 V DC despite the voltage fluctuations on the input side. Boundary switch on-state times ($t_{on(min)}$ and $t_{on(max)}$) and corresponding amplitude values of U_{REC} are shown in Table II. Simulated waveforms of U_{REC} for the different converter operation points are presented in Fig. 12.

Table II
Amplitude Values of U_{rec} for Different Operation Points

Converter input voltage U_{DC}	*2200 V*	*4000 V*
Switch on-state time t_{on}	0.8(T_{sw}/2)	0.44(T_{sw}/2)
Amplitude value of U_{REC}	438 V	795 V

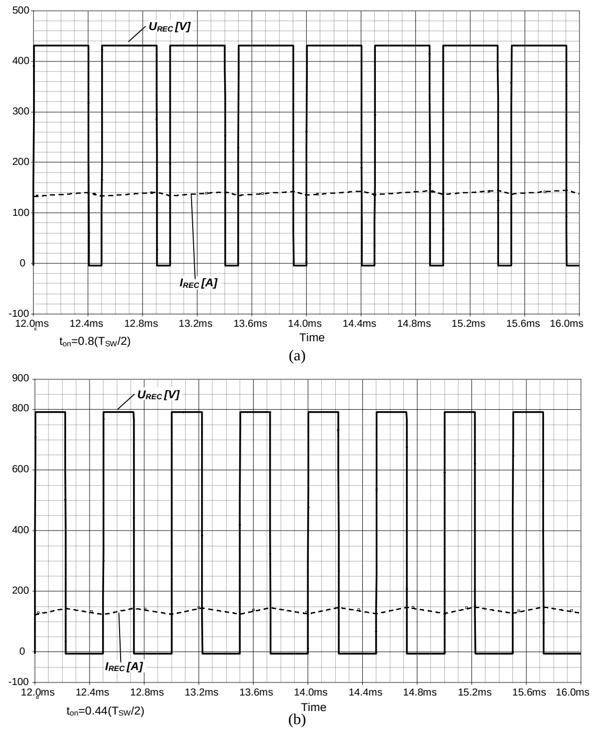


Fig. 12. Simulated waveforms of U_{REC} for different operation points.

Considering transient overvoltages, the rectifier diode repetitive reverse voltage U_{RRM} must be about 1100...1200 V. Conventional fast recovery diodes for such operation voltage have significant forward voltage drop (1.9...3.2 V), which leads to increased diode conduction losses. Otherwise, low-voltage power Shottky rectifiers with minimal (0.4...0.55 V) forward voltage drop can be used in parallel-series connection to achieve the required voltage and current demands. Thus, the selection of output rectifier diodes and overall configuration of an output rectifier for such applications is always a tradeoff between efficiency, compactness and reliability.

Another problem is a temperature mode of the output rectifier. Schottky barrier diode is extremely sensitive to temperature because of its high leakage characteristics over its operating range. The PN junction diode, typically Fast Recovery Epitaxial Diode (FRED) has comparatively lower leakage at higher temperature than the Schottky diode, but the forward voltage is considerably higher. The high forward voltage translates to a higher power loss in the diode, thus creating a larger amount of heat within the diode. This also lowers the overall efficiency of the power supply. So, by exchanging the reduced chances of thermal runaway cause by using Schottky diodes by replacing it with PN junction diodes, the designer reduces the overall efficiency of the system. For power supply designers, efficiency is a very undesirable trade-off for thermal stability. [8], [9], [10]

Fig. 13 shows the developed compact 50 kW FRED-rectifier assembly with associated measurement and protection circuits. Maximum possible power dissipation of output rectifier is assumed to be 500 W and a compact hollow-fin heatsink with the natural convection is enough to ensure the proper thermal mode of the rectifier.

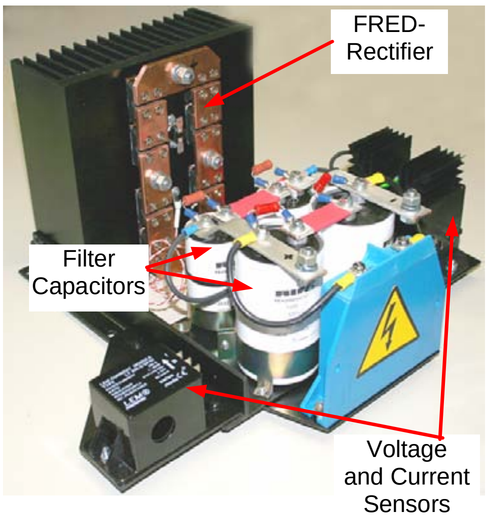


Fig. 13. Compact 50 kW FRED-rectifier with the output filter and associated voltage and current sensors.

IV. CONCLUSIONS

This paper has discussed the development problems of a half-bridge isolated DC/DC voltage converter to be used in 3.0 kV DC commuter trains. The implementation of new 6.5 kV IGBT transistors as a primary-side switches helps to solve common problems of converter power scheme complexity and reliability. Analysis and simulation of main functional blocks of a converter for different operation points have been presented in the paper. General design considerations mainly concerning the inverter stage, power transformer and output rectifier are specified in the paper. Some design optimisation possibilities are proposed.

In the design of high-power converters with high-voltage IGBTs the main attention must be paid to a proper selection of a switching frequency of IGBTs. Here must be analyzed such items like compactness, EMI level, audible hum, etc. In the majority of applications designers use the hard-switching mode of HV IGBT, which is more optimal in terms of power circuit complexity and maintainability. However, implementing soft-switching the IGBT energy losses can be reduced significantly down to 8% (turn-on) or 30% (turn-off) of the hard-switching level. [11] Overally, this gives an effect of 20-30% reduced IGBT losses, which means that the cooling effort and therefore costs can be sufficiently reduced. Otherwise, the switching frequency of IGBTs can be increased to obtain the same power dissipation per IGBT as with the hard-switching.

ACKNOWLEDGEMENT

This work has been supported by the Enterprise Estonia under research and development contract EU23764 “Development of Auxiliary Power Supplies for Electric Trains”.

REFERENCES

[1] *EN 50163:1995, Railway applications - Supply voltages of traction systems*, CLC/TC 9X, 1996-02-19.

[2] Bernet, S., „Recent developments of high power converters for industry and traction applications”, IEEE Transactions on Power Electronics, Volume 15, Issue 6, November 2000, pp.:1102 - 1117.

[3] Dieckerhoff, S.; Bernet, S.; Krug, D. „Evaluation of IGBT multilevel converters for transformerless traction applications” in *Proc. 34th IEEE Annual Power Electronics Specialist Conference, 2003. PESC '03.* pp. 1757 - 1763, vol. 4.

[4] *Magnetic Properties of Tape Wound Encapsulated Cores GAMMAMET®14DC*, GAMMAMET Research & Production Enterprise, 2006.

[5] Starodubtsev Y. N.; Keilin V. I.; Belozerov V. Y. “Cores of fast-quenched GAMMAMET alloys with a linear magnetization curve”, Russian Electrical Engineering , ISSN 1068-3712, ETATS-UNIS, 1993.

[6] K.H. Billings, *Switchmode Power Supply Handbook.* McGraw-Hill, 1989.

[7] A.I. Pressman, *Switching Power Supply Design*. Second Edition, McGraw-Hill, 1998.

[8] D. Vinnikov, „Research, Design and Implementation of Auxiliary Power Supplies for the Light Rail Vehicles”, Ph.D. dissertation, Dept. El. Drives Pow. Elec., Tallinn Univ. Tech., Estonia, 2005.

[9] Brown, Jess; Davies, Richard; Williams, Dilwyn and Bernacchi, Jerry. “High-Efficiency Half-Bridge DC-to-DC Converters with Secondary Synchronous Rectification,” PCIM 2001, Nuremberg, Germany.

[10] *High Temperature Performance of Diodes*, Applications Department, APD Semiconductor, San Jose CA, 2004, rev. 1.

[11] Schwarzer, U.; De Doncker, R.W. „Characterization of 6.5 kV IGBT modules for hard- and soft-switching operation in medium voltage applications”, in *Proc. Twentieth Annual IEEE Applied Power Electronics Conference and Exposition, 2005. APEC 2005.* pp.:329 - 335, vol. 1