A practical investigation of a high power, bidirectional charging system for electric vehicles

J. Tritschler, S.Reichert, B.Goeldi
Fraunhofer Institute for Solar Energy Systems,
Heidenhofstr. 2, 79110 Freiburg, Germany
Tel.: +49/(0)–761 4588-5088
E-Mail: Johannes.Tritschler@ise.fraunhofer.de
URL: http://www.ise.fraunhofer.de

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Abstract

In the last few years, there has been a very high interest in inductive charging systems for electric vehicles, as they have many advantages compared to cable connected solutions. The power rating of most of these systems is kept around 3.3 kW, which leads to very long charging times.
The following paper presents an inductive charging system rated for a high power level of 22 kW. A series-series resonant circuit proves to be very efficient, as high circulating currents, present in parallel compensated systems, are avoided. To reduce switching and conduction losses, but also the size of the whole system, SiC-MOSFETs are used. For the control of the inductive transmission path, several methods are presented and a pulse density modulation (PDM) is investigated and experimentally evaluated.

Introduction

There are an increasing number of different solutions for charging electric vehicles. Most of them are conductive charging systems, which can be separated into stationary fast charging DC and onboard AC charging systems. The possibility of contactless charging is gaining more and more attention due to its excellent user friendliness. Until now, the main disadvantages of contactless charging systems are the lower power ratings and lower efficiencies compared to conductive systems. Also the problem of electromagnetic fields and their interaction with metal objects or human bodies is not yet satisfactorily solved. The variety of charging power for conductive charging systems ranges from 3 kW systems up to more than 120 kW for DC-fast charging stations.
For the inductive charging systems, the power is mostly kept at around 3.3 kW which results in a charging time of more than 6 hours for a conventional 20 kWh battery. In this work an inductive charging system with a power rating of up to 22 kW is presented and investigated. To compensate the reactive power in the two coils, series-series compensation is chosen (Fig. **3**). For high power levels, the advantages of a series-series compensated resonant network have already been shown in [1] and [2]. With a good coil design, most of the losses occur in the series capacitors and in the primary and secondary power semiconductors mainly as conduction losses [3]. Switching losses only play a minor role, as on primary side soft switching can usually be achieved and on the secondary side, the reverse recovery losses of SiC-Diodes are very low. To keep the overall losses and so especially the conduction losses very low, the system was designed to operate at high voltages (up to 800 V) and low currents at nominal operation [4]. On both sides SiC-MOSFETs are used, which enables the system to be bidirectional. The switches can also be used to actively rectify the secondary DC-link voltage in a synchronous manner in order to lower the conduction losses and further increase the efficiency. At nominal operation very high efficiencies of up to 97.4% could be reached. As the infrastructure for inductive charging systems is not yet available, the idea was to combine inductive and conductive charging in one system (Fig 1). This ensures that the users of such a system can plug in their car, if no inductive charging system is available.

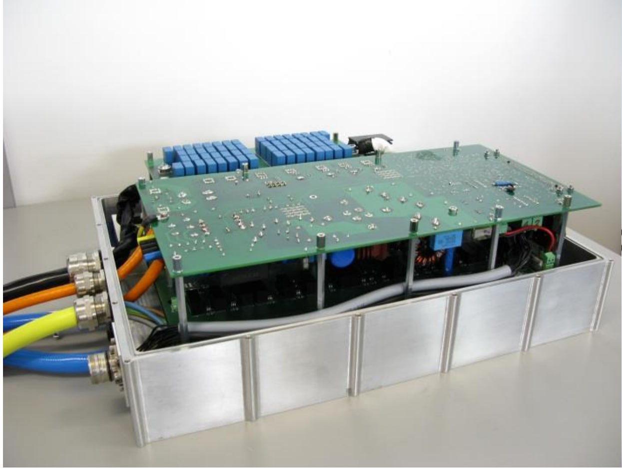
Fig 1: Combined charging system

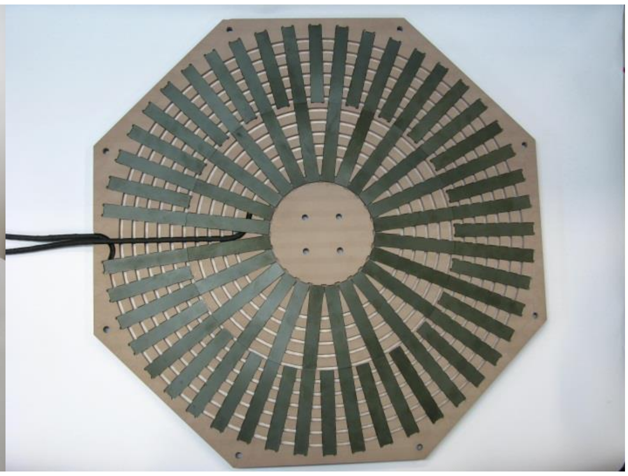
Fig. 2: Transmission coil (60x60cm)

The system is able to charge 22 kW inductively, as well as conductively if necessary. An improved prototype of the transmission coils is shown in Fig. 2.
In this paper, the focus is kept on an optimized control strategy for the operation at partial load. Several methods are presented and two of them experimentally compared to each other.

Partial load operation

To control the transferred power and so the current charging the battery, several methods are possible. At resonant operation, the secondary current is directly proportional to the primary voltage. This gives the possibility to control the power flow by adjusting the input and/or output DC-voltages of the IPT system. This leads to the necessity of added power stages and higher hardware effort. Other possibilities to control the power flow are to change the operating frequency, to phase shift the primary gate signals to generate a PWM output voltage or to use a pulse density modulation (PDM) [6]. The heart of the system consists of two active full-bridges with SiC-MOSFETs and the series-series compensated resonant network (Fig. 3). The series capacitors are dimensioned so that they compensate the reactive power occurring in the self-inductances of the coils (Equation 1) and both the primary and secondary resonant tanks are identical.

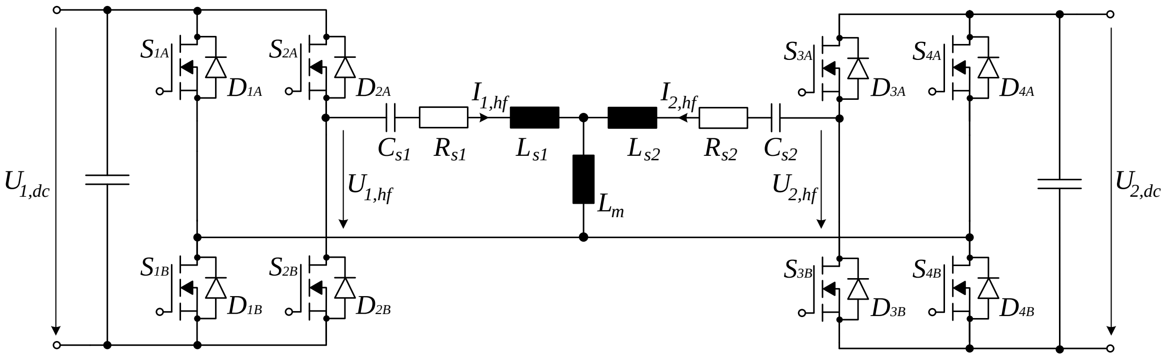

Fig. 3: Inductive power transmission system with t-equivalent circuit

$$\boldsymbol{C_{S1} = C_{S2} = \frac{1}{\omega^2 \cdot (L_{S1} + L_m)}} \qquad (1)$$

The highest efficiency of this IPT-system will be reached, when the primary side is driven slightly above resonant frequency so that ZVS (Zero Voltage Switching) is possible and when $U_{1,dc} = U_{2,dc}$ for a given power [4].

Use of DC/DC-converters

In order to control the IPT system, DC/DC converters can be used between the grid-side AC/DC rectifier and the stationary HF-inverter or the mobile HF-inverter and the battery.
It can be seen in Fig. 4, that the gain $|G_{IU}|$ at resonant frequency between the primary voltage $u_{1,hf}$ and the secondary current $i_{2,hf}$ at resonant frequency is constant and independent of the load condition. So, the transferred power is directly proportional to the amplitude of the input voltage $u_{1,hf}$ when the secondary voltage is fixed. The same calculation can be done for the primary current i_{2hf} and the secondary voltage $u_{2,hf}$. The gain of these two variables is also constant at resonant frequency and independent of the load. So changing the input and/or output voltage of the IPT system changes the transferred power. The first power stage, able to control the transmission line is the active AC/DC grid-rectifier, which can ideally adjust the first DC-link voltage between 650-800 V.

$$\boldsymbol{G_{IU}(j\omega)} = \frac{\boldsymbol{i_{2,hf}(j\omega)}}{\boldsymbol{u_{1,hf}(j\omega)}} \tag{2}$$

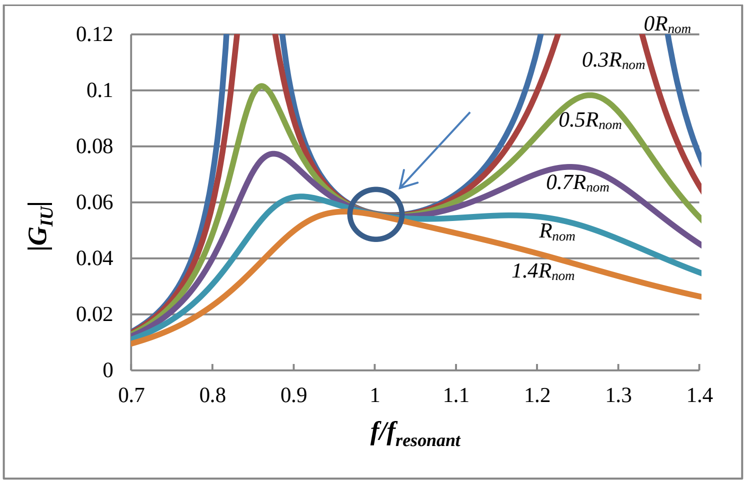

Fig. 4: Gain of G_{IU}

To further reduce the power flow, lower voltages are needed. One first approach is to use two DC/DC-converters, one on each side of the IPT-system Fig. 5(a) operating as step-down converters in the direction grid to battery.

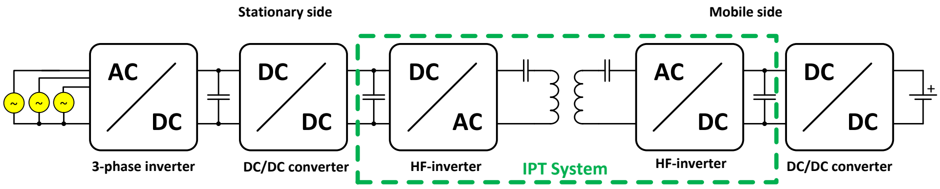

Fig. 5(a): Use of stationary and mobile DC/DC-converters

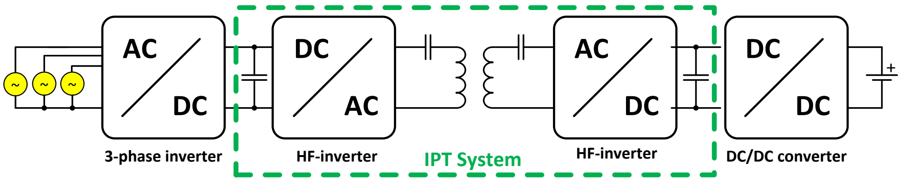

Fig. 5(b): Use of mobile DC/DC-converter

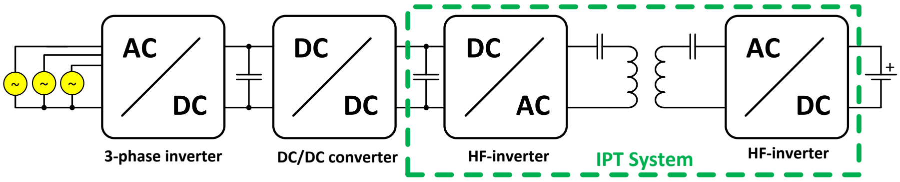

Fig. 5(c): Use of stationary DC/DC-converter

With the configuration in figure 5(a), the IPT-system itself can be driven at a very high efficiency for a wide load range. In addition, the tolerance against position misalignment of the IPT-system is excellent, as a change of the inductive values which will automatically change the current can be compensated by adjusting the voltages on both sides. There are a lot of possible voltage/current combinations until the maximum electrical ratings of the IPT system are reached.
Fig. 6 shows measured efficiencies of the IPT system with fixed secondary voltage $U_{2,dc}$ (blue, yellow and green curves) and a curve when the condition $U_{1,dc} = U_{2,dc}$ is fulfilled (red). By looking on the red curve, the efficiency is nearly constant over a wide power range. This characteristic can only be obtained, when two DC/DC-converters are used (Fig. 5(a)). The IPT-system can then be driven at maximum efficiency, until $U_{2,dc}$ equals the battery voltage. Disadvantage of this configuration is the very high hardware effort.

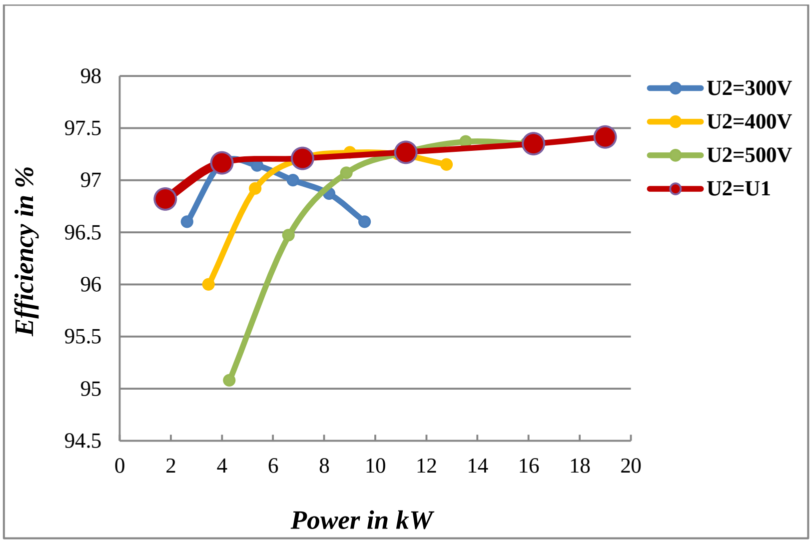

Fig. 6: IPT efficiency from DC-link to DC-link with fixed secondary voltages (blue, yellow, green) and equal voltages on primary and secondary (red)

Other possibilities are shown in fig. 5(b) and 5(c) where only one DC/DC-converter is used, once on the stationary and once on the mobile side. The system presented in fig. 5(b) can only control the power transfer for secondary voltages higher than the battery voltage. When the mobile DC/DC-converter is a buck converter and its input reaches the battery voltage, it's not possible to lower the power level at partial load. So the control has to be done by the primary HF-inverter.
The system shown in fig. 5(c) is able to decrease the input voltage $U_{1,dc}$ and proportional to this the secondary current down to zero. The efficiency of the IPT-system then depends on the battery voltage and moves along the blue, yellow or green curve in fig. 6.

Phase shifting and frequency control

The power flow can be controlled by detuning the circuit out of the resonant frequency. In fig. 4 it can be seen that the output current $I_{2,HF}$ and so the transmitted power strongly depends on the frequency. The disadvantage of this control method is that the soft switching behavior on the primary side will be lost [6] and very high switching losses occur. Another method is to phase shift the gate signals of the primary two legs to change the duty cycle and therefore the average value of the input voltage $U_{1,HF}$. As described in [7] this method also produces high switching and conduction losses.

Pulse density modulation

The pulse density modulation is well known from induction heating. The converter is driven at resonant frequency, modulated by a lower frequency, which adjusts the duty cycle. The converter is able to operate at soft switching conditions over nearly the whole power range. This method was practically evaluated. In fig. 7 the pulse pattern of the primary voltage $U_{1,HF}$ is presented. The time T represents the resonant period and D equals the duty cycle of the modulated lower frequency which is n times lower than the resonant frequency.

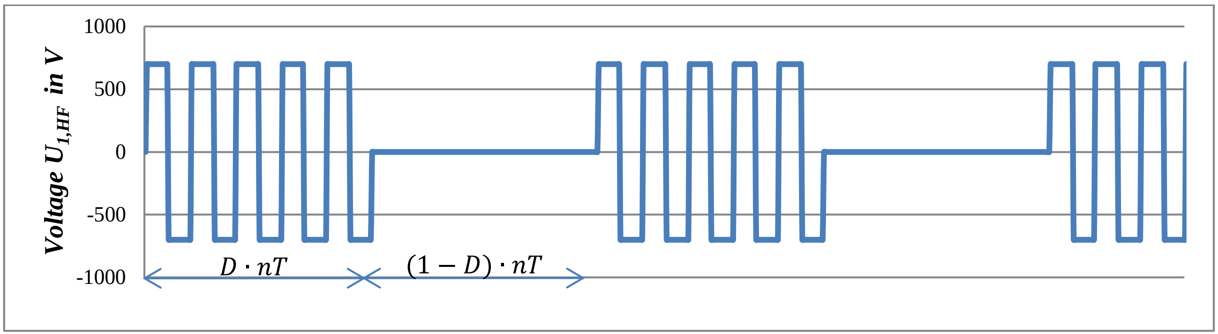

Fig. 7: Pulse pattern of a PDM primary voltage

The transferred energy is regulated by interrupting the power flow to the secondary side and allow the primary current to circulate through the resonant network and the primary switches S_{1A} and S_{2A} or S_{1B} and S_{2B}. In order to get balanced current stresses on all switches, the upper and the lower switches should be changed in each period nT. The period of the low frequency signal can be PWM-modulated and n can be chosen as a tradeoff between the control speed, the size of the output capacitors which have to buffer the low frequent ripple and the efficiency.

Measurement Results

To validate the resonant switching behavior, measurements were conducted. Fig. 8 & 9 show the typical waveforms of the primary pulse density modulated voltage and the sine-wave current. During the time when either both upper or both lower switches are on, the primary current circulates and nearly no damping can be seen. During this free-wheeling period, conduction losses are generated. Other than with a frequency modulated or a phase–shifted modulation, no switching losses appear during this freewheeling period. Another method of PDM control is not to short the two upper or the two lower switches, but to block all the switches. In this case, the primary current is not freewheeling, but fed back into the input capacitors. Measurements showed that better results can be obtained with a PDM with free-wheeling rather than with pulse stop.

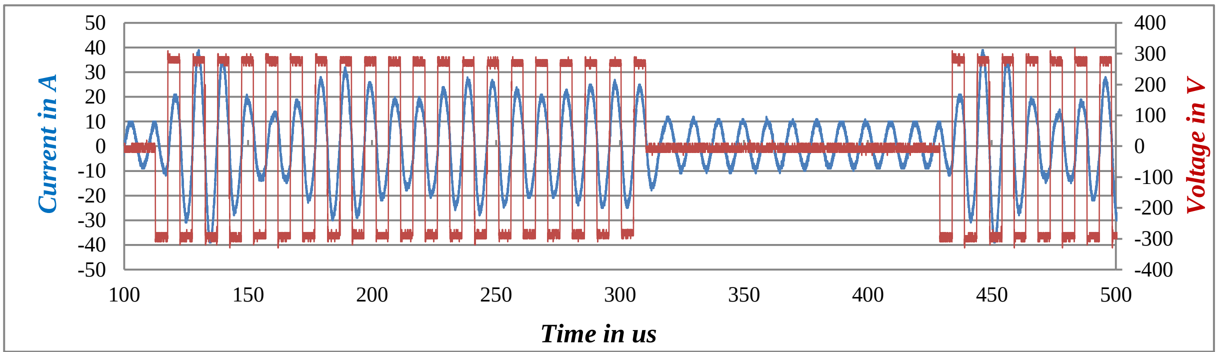

Fig. 8: Measured waveforms of primary voltage and current

During the time when power is transferred to the secondary (fig. 9), soft switching can be achieved in most switching moments.

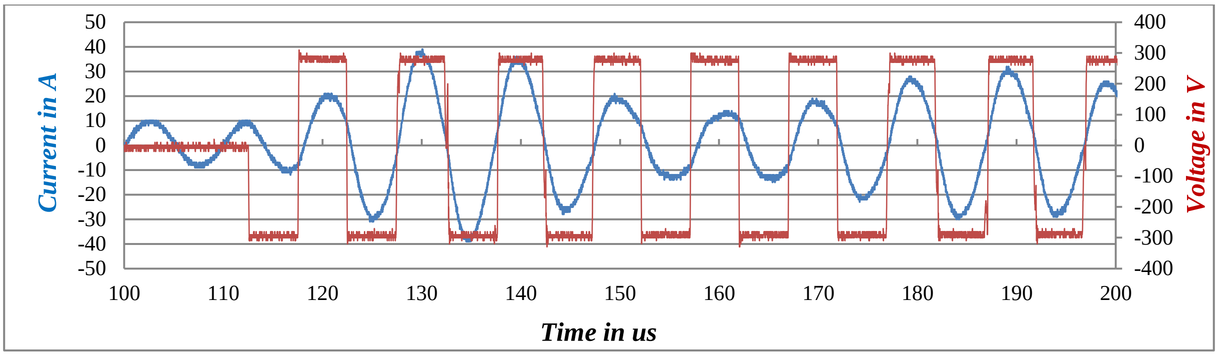

Fig. 9: Measured waveforms of primary voltage and current

It can be seen, that current and voltage are out of phase at startup of the pulse pattern. To improve the switching behavior at startup of each pulse pattern, further research will have the goal to detect the phase shift and to begin the pulse pattern of the voltage in phase with the current.

To verify the assumptions that the pulse density is a real alternative in terms of efficiency, compared to the control by voltage adjustment, measurements were taken at a fixed output voltage of 300 V (fig. 10). At this fixed output voltage, the maximum transferrable power of the system is 11 kW and can be higher at increased voltage $U_{2,DC}$. The general efficiencies are lower than the values shown in fig. 6 due to the fact that in Fig. no active rectification was used.

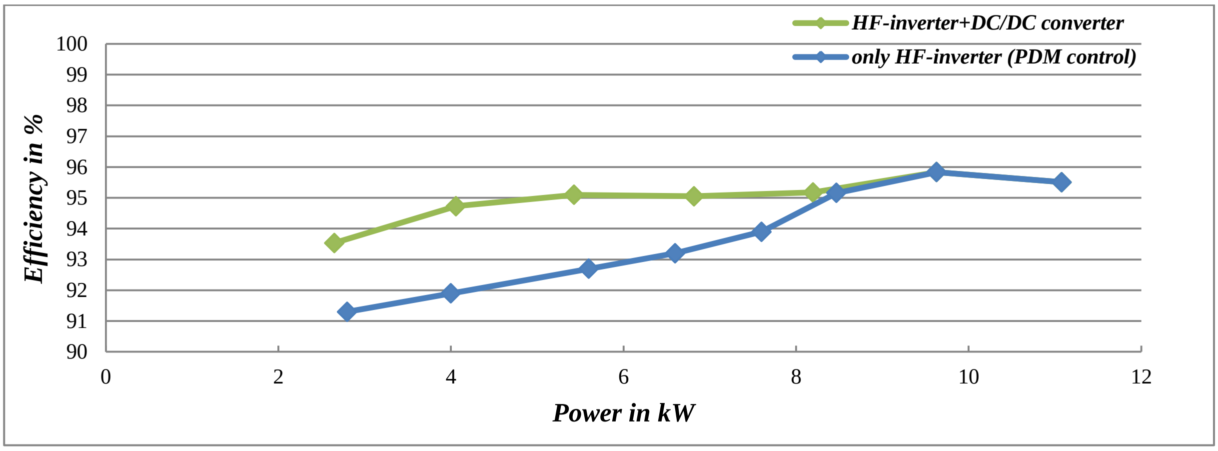

Fig. 10: Measured efficiencies for PDM control (blue) and control by adjustment of $\boldsymbol{U_{1,DC}}$ with an extra DC/DC-converter

With both control methods, it was assumed that the active AC/DC grid rectifier can adjust the DC-link voltage between 650 V and 800 V. Below a desired average voltage of 650 V in the first case (green curve) a DC/DC converter like in the system configuration of fig. 5(c) was used to reduce the input voltage. A second measurement was done without any DC/DC-converter, just by controlling the system via the presented pulse density modulation. In both cases, the efficiencies were measured without the grid connected AC/DC-rectifier. For full power operation at 11 kW, both methods give nearly the same result. For partial load operation the efficiency of the PDM controlled inverter is up to 3% lower than the version with DC/DC-converter.

Conclusion

Several control methods were presented and two of them are experimentally compared to each other. One was with an added DC/DC-converter on the primary side and the other without any additional hardware and PDM-control. As already shown in previous works [2][4], high power inductive power transmission systems based on a series-series compensation can reach very high DC-link to DC-link efficiencies at nominal operation. For the nominal operation both methods get nearly the same efficiency result. At partial load, the solution with DC/DC-converter gets higher efficiencies, at the drawback of additional hardware, control and size of the system. Both solutions are useful, depending on the application.

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