Adaptation of MW-capable DC Supplies to Application Requirements and Resulting Impacts

Abstract

In view of the transition to an emission-free society, applications such as high-performance battery charging, hydrogen electrolysis, and data centers supporting artificial intelligence (AI) require above all powerful, stable, efficient, and reliable supply systems in the MW range. Well-established solutions start colliding with growingly restrictive grid-codes. Various alternatives can be considered, each offering distinct advantages and disadvantages. This paper provides an overview, highlights benefits and drawbacks at a system-level, and compares different solutions regarding critical parameters such as efficiency, complexity, and effort necessary to comply with grid-codes. For infrastructure-style systems, efficiency is on top of the list as even fractions of 1% may lead to GWh of losses during their lifetime.

1. Boundry Conditions

It is important to respect the application's dedicated requirements when choosing and optimizing a topology. The differentiating feature within the three applications compared is the output voltage swing. Additionally, regenerating energy to the grid needs to be considered. Table 1 gives a simplified overview.

Table 1. Application-specific requirements

	Data Center [1]	Battery Charging		Electrolysis
		MCS	Utility	
Output voltage	400 V, ± 400 V, ± 600 V	700–1250 V		640–1000 V
Bidirectionality	No	No	Yes	No
Power Factor	> 0.95	> 0.95	-1…1	> 0.95
Current ripple	< 5%	< 5%		< 7%

Three additional aspects for large-scale, infrastructure-like applications need to be considered that also impact the choice of power semiconductor components. These are availability of uninterrupted uptime, reliability of the design, and its longevity. Though an increase in reliability and uninterrupted uptime can be achieved by integrating redundancies, it is also a consequence of system complexity, and the number of components included within a single system.

Within semiconductors, power modules in solder-bond structure coexist with press-pack or disc-shaped devices. While each of these technologies has its own set of pros and cons, disc-devices offer a simple way of adding redundancy and are more robust with regards to thermal cyclic load.

The available converter topologies implicate different challenges related to the generated harmonic distortion. To evaluate the topologies in terms of harmonics, multiple standards are relevant. The manufacturer of power electronics needs to satisfy the IEC EN 61000-3-4 standard, which defines limits for harmonic currents in low-voltage power supply systems. As these current harmonics encounter the grid impedance of the feeding grid consisting of the impedance of transformers and cables and the short-circuit power of the feeding medium voltage, voltage harmonics are generated. The operator in an industrial plant needs to satisfy the voltage limits in the IEC EN 61000-2-4 standard. As high-power power supplies are directly connected to the medium voltage grid, the power supplier requires that the operator meet the harmonic voltage limits defined in EN 50160. Therefore, the relevant voltage limits are $THD_V < 8\%$ with the individual harmonic voltage limits of 6% for the 5th harmonic, 5% for the 7th harmonic, 3.5% for the 11th harmonic, and 3% for the 13th harmonic.

2. Technical Approach

Power supplies in the power range well below 1 MW can typically take their power from the low-voltage grid directly. In contrast, MW-capable applications need access to the medium-voltage grid by means of transformers, independent of the energy conversion strategy followed. This is an important prerequisite, as examinations done on lower power levels often compare transformer-less topologies to those that inherently need the transformer, thus adding cost and build-volume [2].

MW-arrangements in which DC currents are used to provide high power are now widely equipped with rectifier circuits based on diodes and thyristors. Since a static diode rectifier only provides an output voltage that is neither controlled nor stabilized, further steps need to be considered to provide a stable voltage at a suitable magnitude. At the same time, it is necessary to fulfill the grid code requirements while remaining within the given limits for harmonic distortions. Several potential scenarios can be considered based on the application's requirements:

- Option 1: Use a transformer with an output voltage lower than the lowest voltage needed in the application and apply a boost-converter to adjust the voltage level.
- Option 2: Use a transformer that provides a voltage higher than the highest voltage needed in the application and add a step-down converter.
- Option 3: Apply a standard transformer and use a boost-converter to adapt the input to a step-down converter.
- Option 4: Implement a variable transformer to adapt the output roughly and add means of power electronics for precise adjustment.

In all these categories, various topologies have been developed and optimized in recent years. As option 3 ends up with the highest losses and option 4 is typically installed in applications >10 MW, only options 1 and 2 are considered for this work.

2.1. Boost-type Converters

Boost converters in general allow the DC-output to be controlled by stepping up the voltage to a level that is at least slightly higher than the voltage obtained through simple rectification. This, in turn, enables implementation of power factor correcting schemes (PFC) and reduction in filtering effort, as the grid currents remain sinusoidal. The low-frequency problems caused by the static rectifier are replaced by high-frequency issues, defined by the switching frequency of the boost converter. Two schemes have received a growing interest in recent years, namely, the Vienna Rectifier and the active front end (AFE). The Vienna Rectifier is depicted in Figure 1.

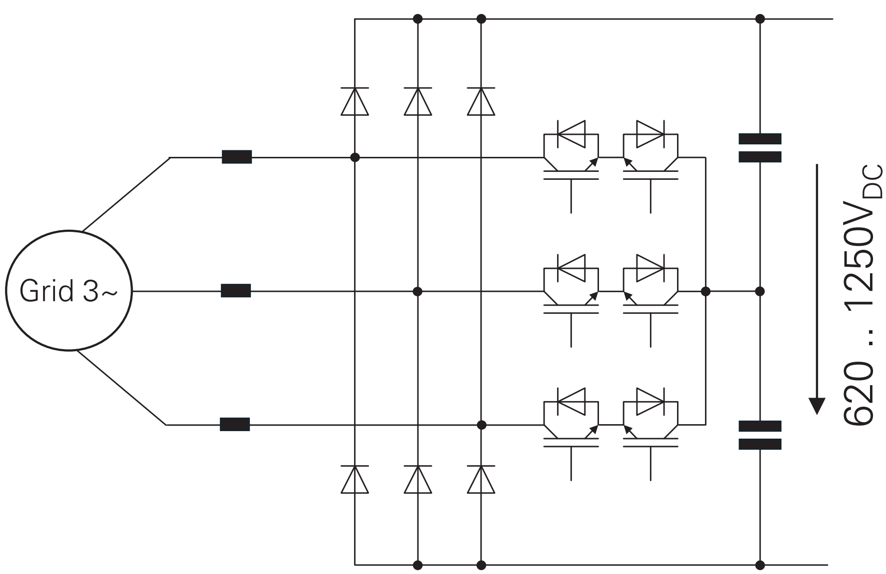

Figure 1. Vienna rectifier as a PFC stage

The topology benefits from using switches connected to the center-tapped DC-Link and thus are only subject to a lower voltage. With an input voltage of 400 Vac, components rated no less than 750 V can be considered, opening the path to using 900 V Si– or SiC-MOSFETs. Another option exists in the form of an AFE. As sketched in Figure 2, a common full bridge in B6C-configuration is implemented.

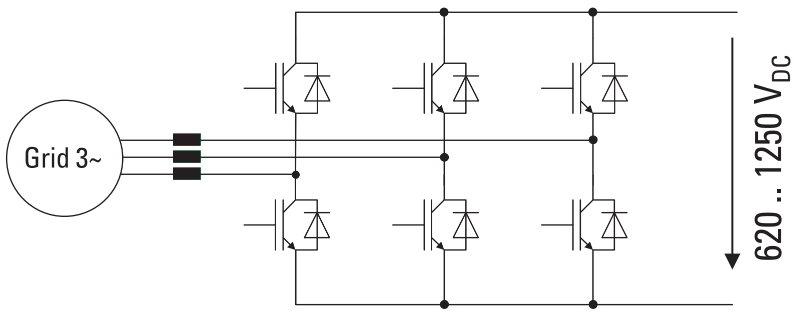

Figure 2. IGBT-based B6C as PFC stage

In contrast to the Vienna Rectifier, the power semiconductors must withstand the full DC-Link voltage. For the Battery Charging application with a projected cut-off voltage of 1250 V in the mega-watt charging standard (MCS), components rated 1200 V are insufficient; the step to using 1700 V devices becomes a necessity. AFEs gain the feature of being bidirectional which enables them to work in grid-scale battery storage systems. The remaining dilemma exists because higher switching frequencies lead to smaller magnetic components, while at the same time increasing the switching losses and thus reducing efficiency.

A countermeasure lies within the utilization of wide band gap semiconductors like silicon carbide (SiC) MOSFETs as these inherently feature lower switching losses than IGBTs. Though this eases the situation regarding efficiency, the high number of individual dies needed to support MW-scale applications increases the probability of failure [4]. Additionally, the high switching speed in both di/dt and dv/dt necessitates close examination of EMI-behavior of these devices and installation of properly dimensioned filters.

Current IGBT-technology offers a maximum rated chip current of 1200 V, 200 A for a single die. A bidirectional switch for higher current must therefore be constructed from multiple parallel branches of two IGBTs in either common emitter or common collector configuration. Today, no compact power module is available to build such a high-current phase leg; therefore, a fictitious component comprised of available power semiconductors is considered. Based on values known from similar components, the characteristic parameters for 200 A/1200 V devices can be estimated to be:

IGBT:

- forward voltage of the IGBT $V_{CE(sat)} = 2$ V
- sum of switching losses $E_{on} + E_{off} = 35$ mJ
- thermal resistance $R_{th(j\text{-}c)} \sim 0.15$ K/W

Diode

- forward voltage of the diodes $V_F = 1.7$ V
- recovery energy of the diodes: $E_{rec} = 15$ mJ
- thermal resistance $R_{th(j\text{-}c)} \sim 0.26$ K/W

Operation in the rectifier is subject to certain restrictions. The die is specified to switch a maximum current of twice its rated value, so exceeding 400 A is not allowed. Simulation reveals that a 200 A-setup is sufficient to support about 250 A of DC current before the peak current in the boosting IGBT exceeds 400 A, as depicted in Figure 3.

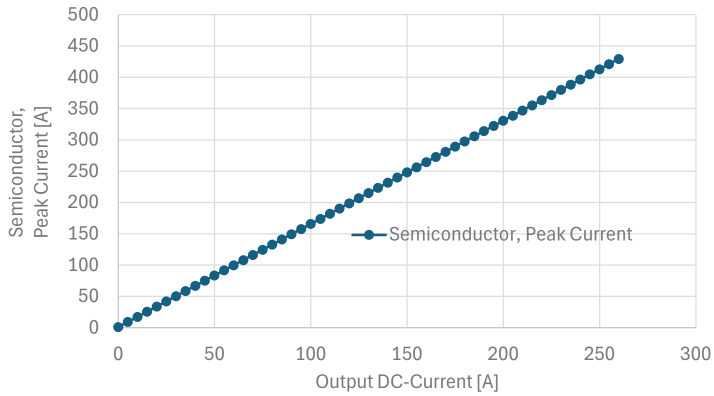

Figure 3. Peak current vs. DC-Output current, Vienna rectifier

Consequently, paralleling a minimum of five individual IGBT chips in each position is required. The 1 MW Vienna Rectifier would thus need to be constructed of no less than 3·4·5 = 60 diodes and at least 3·2·5 = 30 IGBT chips or 30 individual dies per phase leg. Though this may sound excessive, consider that current high-power modules already feature up to 24 chips in parallel. To achieve comparable efficiency across all topologies by minimizing switching losses, the switching frequency was set to just 337 Hz. Loss calculation leads to 95 W for the IGBTs, 145 W for the diodes in the bidirectional switch, and 140 W for the diodes connected to the DC-link. This totals 5700 W of losses at 1 MW of output power and achieves an efficiency of 99.4%.

Following the same approach, each switch in the AFE must be constructed of several chips rated 200 A each. As these need to withstand the full DC-link voltage, it is necessary to use components rated for 1700 V. The drawback here is that components with higher blocking voltage inherently feature higher forward voltage and switching losses. Taking an available 200 A-component for reference, the determining parameters are:

IGBT:

- forward voltage of the IGBT $V_{CE(sat)} = 2.3$ V
- sum of switching losses $E_{on} + E_{off} = 130$ mJ
- thermal resistance $R_{th(j\text{-}c)} \sim 0.12$ K/W

Diode:

- forward voltage of the diodes $V_F = 1.9$ V
- recovery energy of the diodes: $E_{rec} = 45$ mJ
- thermal resistance $R_{th(j\text{-}c)} \sim 0.16$ K/W

Simulating the setup, the correlation between the DC output current and the peak current in the individual devices can be found. For IGBT and diode, this is displayed in Figure 4.

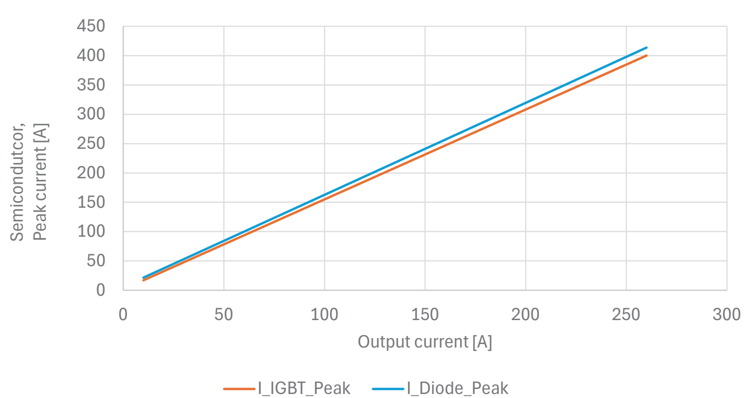

Figure 4. Peak current vs. DC-output current, AFE

2.2. Thyristor-based Converter

Thyristor-based topologies allow for stepping down the voltage applying phase-control techniques. In contrast to systems built on fast-switching semiconductors, neither very high switching speed nor high switching frequencies are necessary.

The B6C bridge commonly used on the three-phase grid is a building block, often as B12C or even higher pulse rates to reduce the voltage ripple in combination with appropriate transformers. The resulting circuit is depicted in Figure 5.

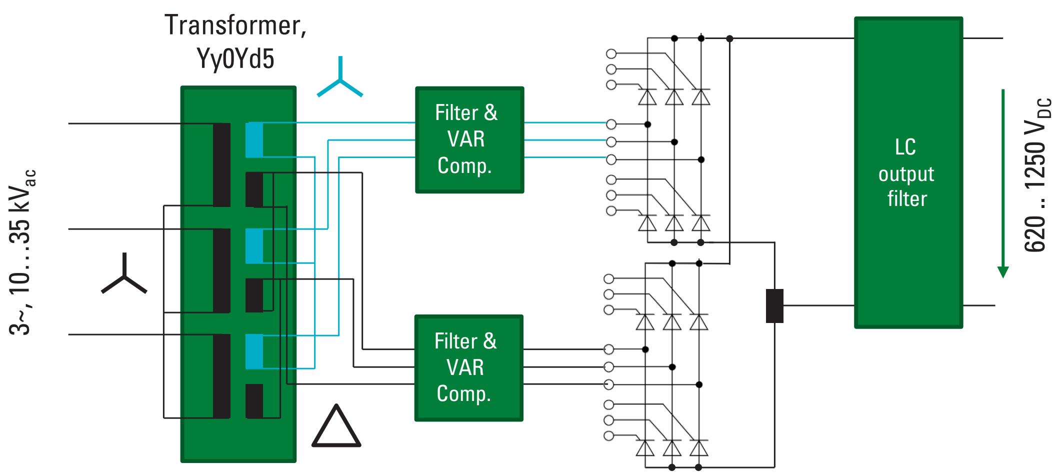

Figure 5. Thyristor-based B12C circuit

From a simulation based on a suitable phase control thyristor [8], the losses per switch were determined. As seen in Figure 6, each switch contributes 417 W of losses, which leads to 5 kW of losses within the B12C arrangement and an efficiency of 99.5%.

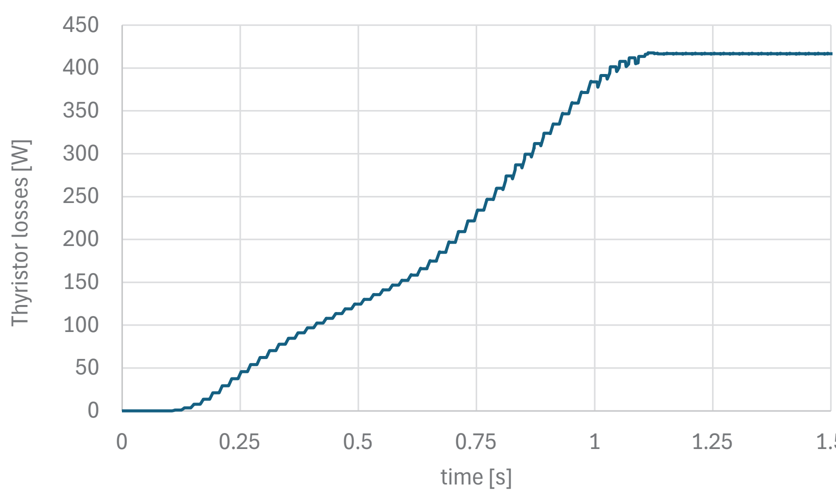

Figure 6. Thyristor losses

This efficiency matches that of the Vienna rectifier, but the dramatically reduced number of individual devices and the lower complexity level still make thyristors the more favorable choice. Given the importance of availability, closely examing strategies for integrating redundancy is recommended. While a Vienna rectifier structure would have to be doubled to be redundant, adding a second thyristor in series to the existing ones would be sufficient. While this reduces efficiency by 0.5%, it does not increase the necessary build-volume by a lot, thus offering a much better ratio regarding build-space. However, the B12C approach necessitates reactive power compensation and harmonic filtering which are considerations as well. To get the most efficient setup, setting up reactive power compensation in a way that accomodates harmonic filtering is recommended.

3. Filter and Compensation Effort

Generally, it is possible to install the filtering and reactive power compensation devices on the low voltage side or directly in the medium voltage grid. Both approaches have pros and cons. Because of higher isolation requirements for components like circuit breakers, inductors, and capacitors, medium-voltage installations tend to be more expensive.

With increasing power or filter demand, medium voltage installations become competitive at higher ampacities. The breakpoint typically lies between 2.5 and 5 MVA, depending on the installation environment and possibilities for heat removal. A further disadvantage is that grid operators may restrict access to medium-voltage installations because the transformer is the predefined interconnection point to the network. In such cases, it is not possible to install power quality devices on the medium-voltage side.

Especially for the thyristor rectifier, it is more challenging to install the power quality devices on the low-voltage side. The advantage is that low-voltage components such as transformers and circuit breakers are directly relieved from the harmonic currents and the resulting additional losses. Two challenges need to be considered. First, for the 12-pulse thyristor rectifier, the 5th and 7th harmonic current are still present and not yet eliminated. The filter design needs to respect that the filter does not absorb these currents.

Additionally, the thyristor rectifier generates commutation notches. These deep and steep voltage notches lead to voltage and current harmonics with a frequency > 2 kHz and present high stress for all components. The impact of commutation notches rises with increasing firing angle α. A common objection to the 12-pulse thyristor rectifier is that it requires high amounts of reactive power depending on the firing angle α. However, proper filter design enables using the harmonic filters for reactive power compensation. The current levels for the different topologies are depicted in Figure 7.

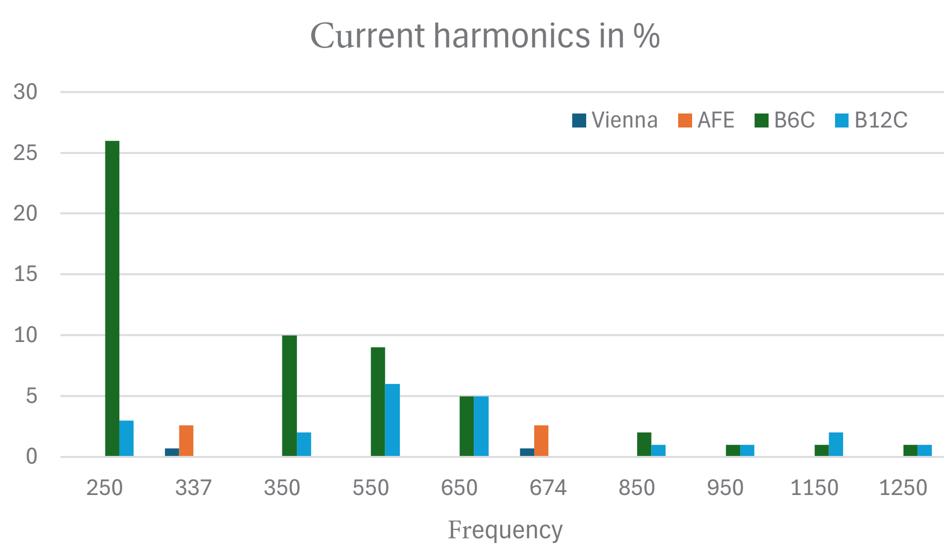

Figure 7. Current harmonics for the topologies considered

Depending on the different current and voltage frequency spectra, the required filter components can be derived. The Vienna rectifier as well as the AFE use a switching frequency of 337Hz. This is quite low, given that the 1200 V-IGBTs could support several kHz. However, to achieve similar efficiency values in all topologies, higher switching losses would not be tolerated.

Both the step-down converters draw a nearly sinusoidal current from the grid, so that filter measures become necessary only for harmonics caused by the switching frequencies. The switching actions not only lead to voltage harmonics around the switching frequency itself but also around twice the switching frequency. It is assumed that the line inductor is designed to limit the current ripple to 20% of the peak current [2].

The current harmonics of the B6C thyristor bridge can be derived from the waveforms of a diode bridge B6U, but phase-shifted with the firing angle α. In theory, the current harmonics of an ideal rectifier with a large inductance on the load side result in a current of THD_I = 31%, the odd non-triple harmonics decreasing with the order 1/n. In practice, the current THD is around 36 % [6]. As the harmonic currents generated by the power electronics topologies encounter the grid impedance, harmonic voltage distortion is generated. To evaluate the voltage distortion, it is assumed that the feeding transformer has a power rating of S_T = 2 MVA.

With a short-circuit voltage of 6 %, the impedance is $X_{T\sigma}$ = 0.5 mΩ + j5 mΩ. In Figure 8, the resulting voltage harmonics can be seen.

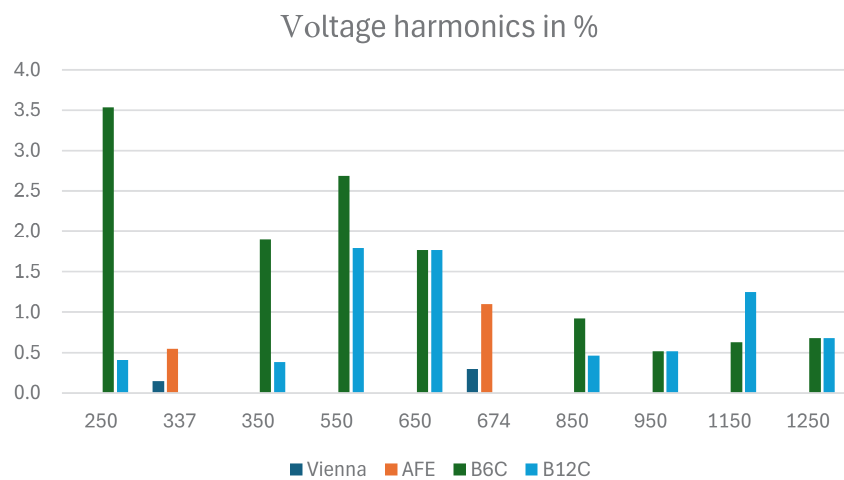

Figure 8. Voltage harmonics created by the topologies considered

The approach for the filter design is to consider the rectifier circuit and its harmonic content as a current source. By installing passive filters, the grid impedance can be shaped so that the harmonic currents get absorbed by the filter. The harmonic trap filter is depicted in Figure 9a.

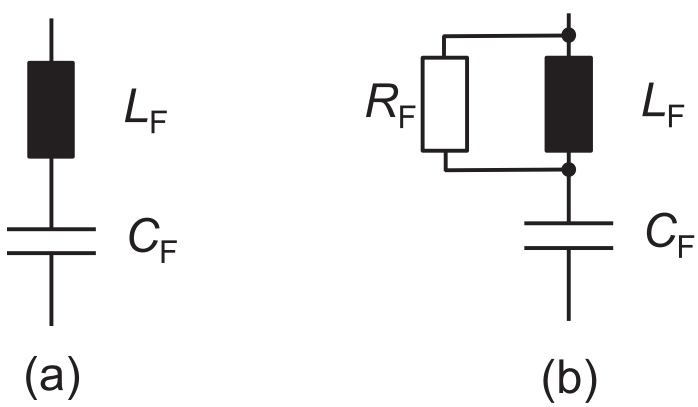

Figure 9. General structure of harmonic trap filters

An inductor and a capacitor are connected in series and form series resonance. The series resonance of the filter needs to be placed just below the harmonic frequency which should be filtered. As an example, a harmonic trap filter targeting the switching frequency is derived.

With an inductor L_F = 700 µH and a capacitor C_F = 343 µF, the resonant frequency is fixed to 325 Hz, which is 12 Hz below the switching frequency to be filtered. The resulting grid impedance can be seen in Figure 10.

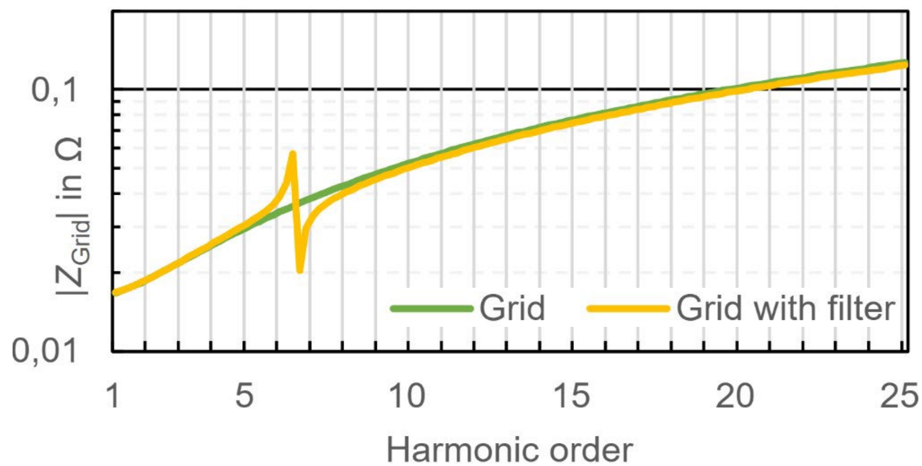

Figure 10. Grid impedance shaped by the filter

As the filter inductor has a larger value than the grid impedance, the harmonic trap filter has no impact for higher frequencies. The same process can be repeated for the doubled switching frequency at 674 Hz. In general, it is important to consider that harmonic amplification does not occur.

For example, lays the double switching frequency close to the 13th harmonic.

Another constraint in filter design may be the limitations on installing capacitive reactive power. To ensure power quality in a thyristor rectifier, the reactive power demand must first be determined. The relation of the output voltage range leads to a maximum firing angle of:

$$\alpha = \arccos\left(\frac{700\,V}{1250\,V}\right) = 56°$$

Reactive power demand depends on the output voltage and thus correlates to the firing angle. The correlation is illustrated in Figure 11.

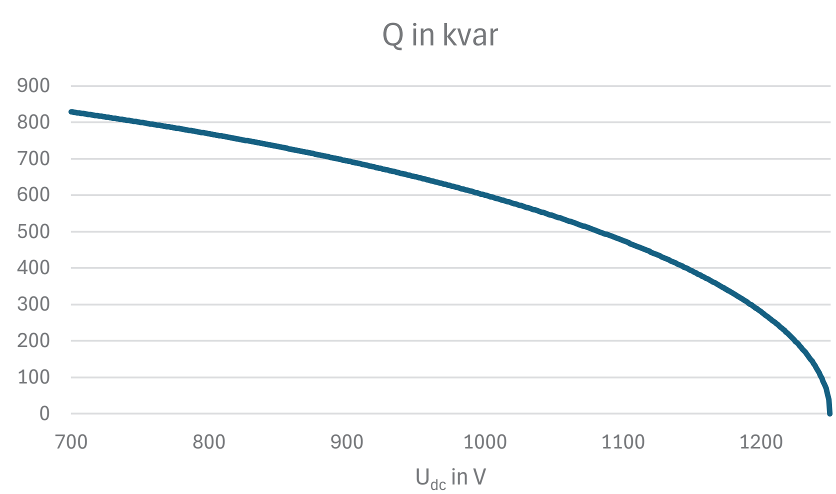

Figure 11. Reactive power demand vs. output voltage

The maximum reactive power demand at 700 V is around 800 kVAr. As in the 12-pulse rectifier, the outputs are connected in series and the reactive power is shared by both rectifiers. For an exemplary filter design for each thyristor rectifier, one damped harmonic trap filter as depicted in Figure 9b is derived.

The filter must be designed with the requirement to also compensate for the reactive power as a secondary use. For 400 kVAr and tuned for the 11th harmonic, the values recommended are C_F = 6.1 mF and L_F = 14.8 µH. The value of the damping resistor mainly depends on factors such as losses, filter effect, and undesired amplification effect for the lower frequency harmonics. The goal is to identify the best possible compromise among the discussed aspects. It must be acknowledged that, in this approach, the filter leads to an increase in impedance for the 7th harmonic, around 350 Hz. There are essentially two options to solve this problem. The first is to install an additional filter for the 7th harmonic, which converges the resulting grid impedance back to inductive grid impedance without filters. This filter would just absorb a small current correlating to low losses. In Figure 12, the resulting grid impedance with and without damping resistor can be seen.

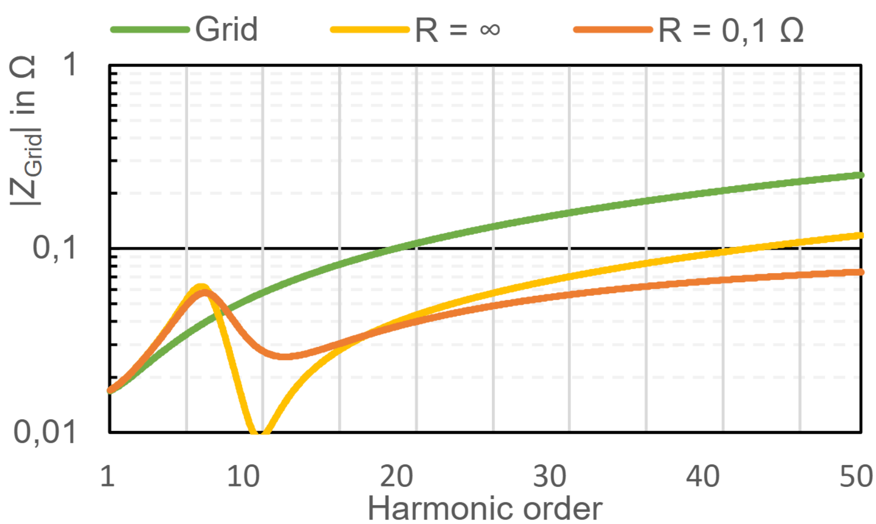

Figure 12. Grid impedance including compensating filter

The second option is to divide the total reactive power demand into several smaller filter steps, each tuned to different frequencies, with or without damping resistors. The total number of steps also depends on the possible operating points of the connected DC load. The higher the thyristor firing angle, the more filter steps are needed and switched on. The dependence on the operating point is related to both the capacitive reactive compensation power demand and the filter current drawn by the corresponding filter step, since a higher thyristor firing angle typically results in increased harmonic current load.

Furthermore, the filter currents absorbed by the filter must be considered. The designed filter is likely to absorb a high filter current, which reinforces the need to use several smaller filters tuned to different resonant frequencies. In addition, the detailed filter design is highly dependent on grid and load parameters and goes beyond the scope of this paper. However, a general approach to thyristor filter design is presented. From extensive experience, the efficiency of the compensation and filtering can be estimated to be ~10 W/kvar in the trap filter and ~25 W/kvar in the damped filter due to the damping resistor.

4. Conclusion

As in most cases, there is no "one-fits-all"-solution when it comes to designing MW-capable supply systems. Depending on the application's requirments, different solutions must be considered. Based on the presented work, the thyristor-based solution with reactive power compensation is preferred for applications that operate within a narrower bandwidth as it offers the highest efficiency, robustness, and longevity at the lowest system cost and complexity. When a wider bandwidth, from 10% to 100%, is required, solutions featuring active front ends and DC-DC-converters become viable. For applications requiring bidirectional operation, IGBT-based solutions compete with thyristor-based alternatives. From a failure probability perspective, the thyristor-based solution is also favored due to its lower complexity and reduced number of components, even though it requires reactive power compensation.

Acknowledgement

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