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MW-Capable DC Supplies, their Adaptation to Application's Needs and Resulting Consequences

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Abstract

In view of the transition to an emission-free society, applications such as high-performance battery charging, hydrogen electrolysis or 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 providing individual pros and cons. The present work gives an overview, highlights benefits and drawbacks on system-level and compares different solutions regarding critical parameters such as efficiency, complexity, and effort necessary to comply to 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 the lifetime.

1 Boundary conditions

Choosing and optimizing a topology has to respect the application's dedicated requirements. 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.

	Data Center [1]	Battery charging		Electro-lysis
		MCS	Utility	
Output voltage	400 V, ±400V, ±600V	700-1250 V		640-1000 V
Bidirec-tionality	No	No	Yes	No
Power Factor	>0.95	>0.95	-1…1	>0.95
Current ripple	<5%	<5%		<7%

Table 1 Application specific requirements

Three further aspects for large-scale, infrastructure-like applications need to be considered that also impact the choice of power semiconductor components. These are availability or 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.

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

The available converter topologies implicate different challenges regarding the generated harmonic distortion. To evaluate the topologies in terms of harmonics different standards are relevant. The manufacturer of power electronics needs to satisfy the standards IEC EN 61000-3-4, 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 standard IEC EN 61000-2-4. As high-power power supplies are directly connected to the medium voltage grid, the power supplier demands from the operator to meet the harmonic voltage limits defined in EN 50160. Therefore, the relevant voltage limits are $THD_V < 8\%$ with the voltage limits: 6% for fifths harmonic, 5% for sevenths harmonic and 3.5% for elevenths harmonic and 3% for thirteenths harmonic voltage.

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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 on 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.
As 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 in a suitable magnitude.
At the same time, fulfilling the grid code requirements and remain within the given limits for harmonic distortions needs to be achieved.

Regarding the application requirements, several potential scenarios can be considered:

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.
2) Use a transformer that provides a voltage, higher than the highest voltage needed in the application and add step-down conversion
3) Apply a standard transformer and use a boost-converter to adapt the input to a step-down-converter
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 >10MW, 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 at least slightly higher than the purely rectified voltage would be.
This in turn allows implementing power factor correcting schemes (PFC) and reduce 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 the last years, namely the Vienna Rectifier and the Active Fron-End (AFE)
The Vienna Rectifier is depicted in **Fig. 2**.

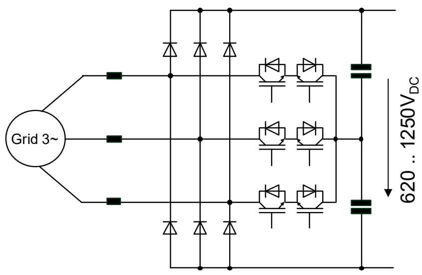

Fig. 2 Vienna Rectifier as a PFC stage

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.

A further option exists in form of an active front end (AFE). As sketched in **Fig. *1***, a common full bridge in B6C-configuration is implemented.

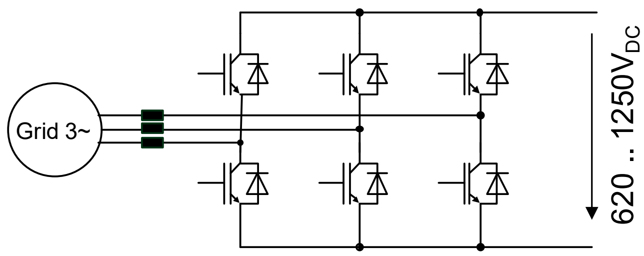

Fig. 1 IGBT-based B6C as PFC stage

In contrast to the Vienna Rectifier, the power semiconductors must withstand the full DC-Link voltage. For the application *Battery Charging* with a projected cut-off voltage of 1250 V in the megawatt 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 increase the switching losses and thus reduce efficiency.

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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 du/dt demands to closely examine EMI-behavior of these devices and install 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 thus needs to 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_{thjc}~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_{thjc}~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 **Fig. *3***.

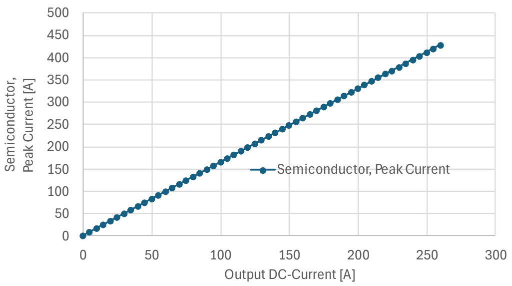

Fig. 3 Peak current vs. DC-Output current, Vienna rectifier

Consequently, paralleling a minimum of 5 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, it needs to be considered, that current high-power modules feature up to 24 chips in parallel already.
To reduce switching losses to get to comparable results regarding efficiency in all topologies, the switching frequency was chosen to be 337 Hz only.
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 sums up to 5700 W of losses at 1MW of output power and achieves an efficiency of 99.4%.

Following the same approach, each switch in the Active Front-End needs to be constructed of several chips rated 200 A each. As these need to withstand the full DC-link voltage, using components rated for 1700 V becomes necessary. This leads to the drawback, 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_{thjc}~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_{thjc}~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 **Fig. *4***.

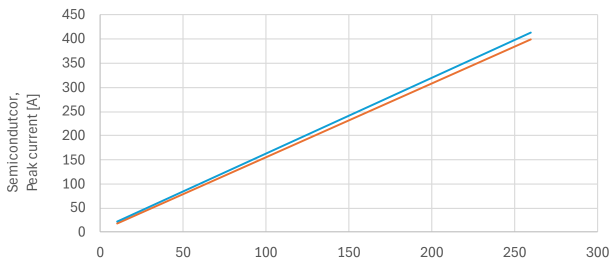

Fig. 4 Peak current vs. DC-Output current, AFE

Here too, 400 A peak in the semiconductors achieves 250 A of DC-current, making 5 dies in parallel a minimum.
The full Active Frond End thus needs to be constructed from no less than 30 individual IGBTs and 30 diodes.

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As expected, this leads to higher losses as in the Vienna rectifier. The IGBTs account for 225 W of losses and the diodes contribute 363 W each.
This sums up to 17.64 kW which reduces efficiency to 98.3%.

2.2 Thyristor-based converter

Thyristor-based topologies allow for stepping down the voltage applying phase-control techniques. In contrast to systems build 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 **Fig. *5***.

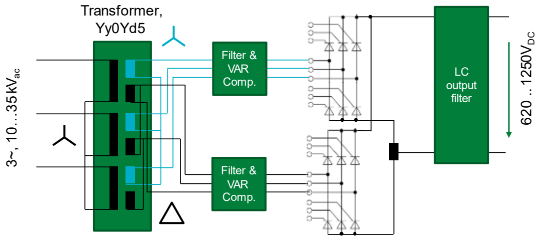

Fig. 5 Thyristor-based B12C circuit

From a simulation based on a suitable phase control thyristor [8], the losses per switch were determined. As seen in **Fig. *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%

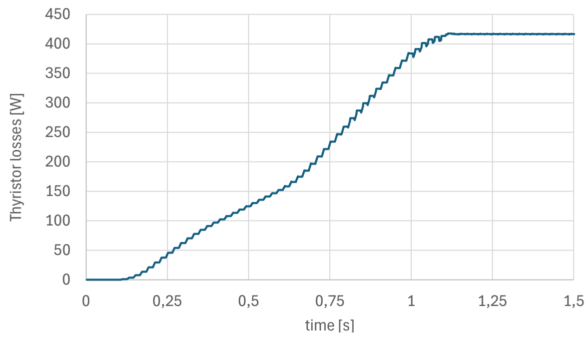

Fig. 6 Thyristor losses

This efficiency matches the Vienna rectifier, but the dramatically reduced number of individual devices and the lower complexity level still favor the thyristors.
As availability is an important point too, a closer look on how to integrate redundancy is advised. 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 a lot, thus offering the much better ratio regarding build-space.

However, the B12C-approach demands to have reactive power compensation and harmonic filtering which needs to be considered as well.

To get the most efficient setup, setting up a reactive power compensation in a way that allows to also care for 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 are in general more expensive.
With increasing power or filter demand, medium voltage installations become competitive at higher ampacities. The breakpoint lays around 2,5-5 MVA, depending on the installation environment and possibilities for heat removal.
A further disadvantage is, that grid operators might restrict the access for installations on the medium voltage side, because the transformer is the predefined interconnection point to the network.
If so, it is not even 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 like transformer 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. This deep and steep voltage notches lead to voltage and current harmonics with a frequency > 2 kHz and present a 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.

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The current levels for the different topologies are depicted in Fig. 7.

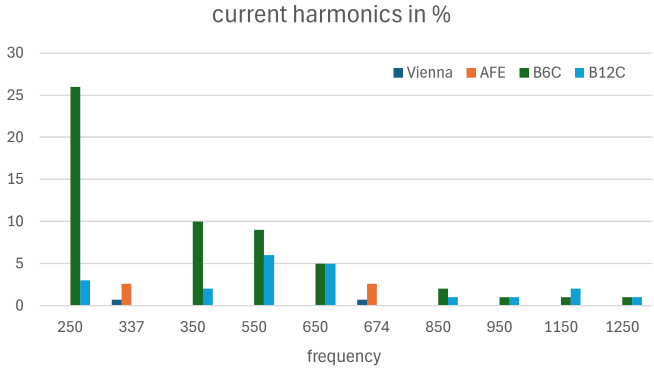

Fig. 7 Current harmonics for the topologies considered

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 could not be tolerated.

Both the step-down converters draw a nearly sinusoidal current from the grid, so that filter measures become only necessary 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 Fig. 8 the resulting voltage harmonics can be seen.

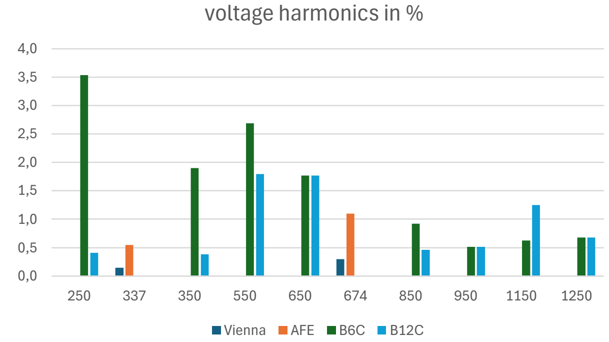

Fig. 8 Voltage harmonics created by the topologies considered

In this example the limits for harmonic voltages neither in the EN 61000-2-4 nor the EN 50160 will be exceeded.

Nevertheless, the derivation shows in general that it is not sufficient to look only at the current harmonics as the grid impedance increases with rising frequency. This leads to relatively higher distortions in the voltage harmonics compared to the current harmonics. In practice, even higher voltage levels are expected as in this approach additional impedances of cable distribution, the short-circuit power of the medium voltage grid or voltage levels superimposed from the feeding medium voltage are neglected for simplification.

For the AFE and the Vienna rectifier the switching frequency is limited below 500 Hz because of switching losses. In this example a switching frequency of 337 Hz is chosen. The odd number ensures that there won’t be resonances with multiples of either 50 or 60 Hz.

Grid-connected active rectifiers are usually connected to the grid via a third order LCL-filter.

The general approach is to choose the cut-off frequency of the LCL-filter with a factor 10 higher than the fundamental frequency and a fraction of 10 of the switching frequency, which results in an attenuation of -60 dB/decade for the switching harmonics [7]. Because of the low switching frequency this approach is not appropriate in this application. Therefore, harmonic trap filters become necessary around the switching frequency and twice its value.

The approach for the filter design is to consider the rectifier circuits 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.

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The harmonic trap filter is depicted in a) in Fig. 9.

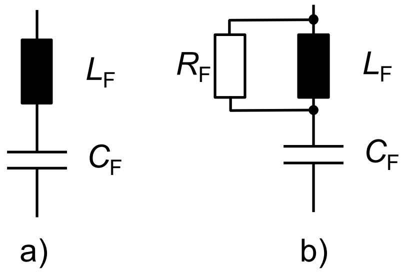

Fig. 9 General structure of harmonic trap filters

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

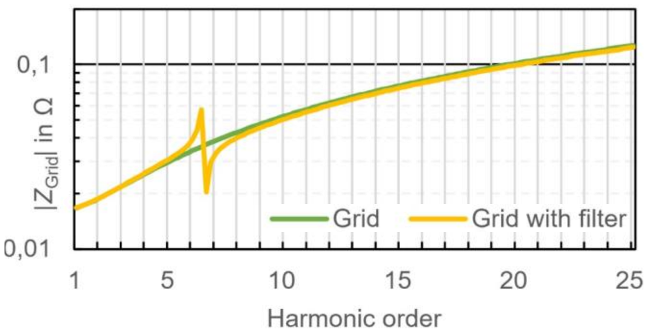

Fig. 10 Grid impedance shaped by the filter

can be seen in Fig. 10.
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 no harmonic amplification occurs.
For example, lays the double switching frequency close to the 13th harmonic.
Another constraint in the filter design might be limits for the installation of capacitive reactive power. To ensure the power quality in the thyristor rectifier first the reactive power demand needs to 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 Fig. 11.

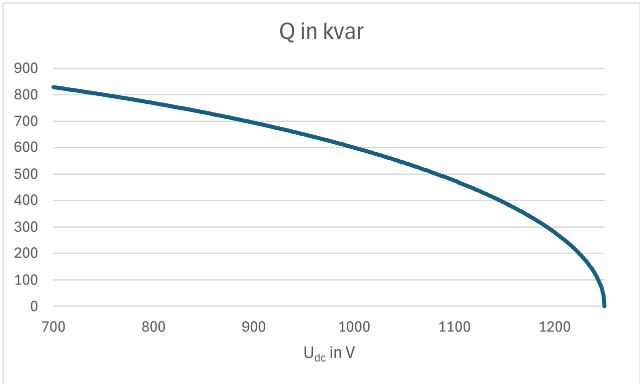

Fig. 11 Reactive power demand vs. output voltage

The maximum reactive power demand at 700 V is around 800 kVAr. As in the 12-pulse rectifier the outputs are connected in series, 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 b) of **Fig. *10*** will be derived.
The filter will be designed with the requirement to also compensate for the reactive power as a secondary use. For 400 kVAr and tuned for 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 the factors: losses, filter effect and undesired amplification effect for the lower frequency harmonics. The purpose is basically to find the best possible compromise between these mentioned aspects. It needs to be respected that in this approach the filter leads to an impedance-increase for the 7th harmonic around 350 Hz.
There are basically two options to solve this problem. The first option 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 Fig. 12 the resulting grid impedance with and without damping resistor can be seen.

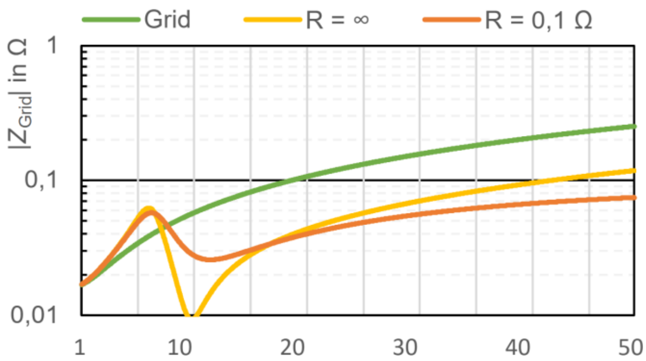

Fig. 12 Grid impedance incl. compensating filter

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Second, the total reactive power demand can be split in several smaller filter steps with different tuning frequencies with or without damping resistors. The total number of steps also depends on the possible operation points of the connected DC load. The higher the thyristor firing angle is, 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 that is drawn by the according filter step, since a higher thyristor firing angle typically also means a higher harmonic current load.
Furthermore, the filter currents, which the filter will absorb need to be considered. The filter designed will probably absorb a high filter current, which also presents the necessity to use several smaller filters, tuned for different resonant frequencies. In addition, the detailed filter design is strongly dependent on grid and load parameters and exceeds the scope of this paper. But a general approach for thyristor filter design is shown.
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.

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 demand, different solutions need to be considered. From the work presented, the thyristor-based solution with reactive power compensation is to be preferred in applications that are well-served with a less wide operation bandwidth as it offers the highest efficiency, robustness, and longevity at lowest system cost and complexity. Once a wider bandwidth from 10% to 100% is required, solutions featuring active front ends and DC-DC-converters can be considered. In case bidirectional operation is required, IGBT-based solutions compete against thyristor-based solutions.
From failure probability point of view, the lowest complexity and the lowest number of devices used favor the thyristor-based solution as well, even if these need reactive power compensation.

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