This is an electronic reprint of the original article.
This reprint may differ from the original in pagination and typographic detail.

Koppinen, Jussi; Rahman, F. M. Mahafugur; Hinkkanen, Marko

Effects of the switching frequency of a grid converter on the LCL filter design

Published in:

8th IET International Conference on Power Electronics, Machines and Drives (PEMD 2016)

DOI:

[10.1049/cp.2016.0206](https://doi.org/10.1049/cp.2016.0206)

Published: 19/04/2016

Document Version

Peer reviewed version

Please cite the original version:

Koppinen, J., Rahman, F. M. M., & Hinkkanen, M. (2016). Effects of the switching frequency of a grid converter on the LCL filter design. In *8th IET International Conference on Power Electronics, Machines and Drives (PEMD 2016)* <https://doi.org/10.1049/cp.2016.0206>

This material is protected by copyright and other intellectual property rights, and duplication or sale of all or part of any of the repository collections is not permitted, except that material may be duplicated by you for your research use or educational purposes in electronic or print form. You must obtain permission for any other use. Electronic or print copies may not be offered, whether for sale or otherwise to anyone who is not an authorised user.

Effects of the switching frequency of a grid converter on the LCL filter design

Jussi Koppinen, F. M. Mahafugur Rahman, and Marko Hinkkanen*

*Aalto University School of Electrical Engineering, Finland, *jussi.koppinen@aalto.fi*

Keywords: Area product, core material, grid converter, high-frequency inductor model, LCL filter

Abstract

The size of the LCL filter can be decreased significantly using higher switching frequencies enabled by next generation semiconductors. This paper studies the effects of the switching frequency of a grid converter on the LCL filter design. The goal in the design is to minimize the stored energy of the magnetic components. The core material of the converter-side inductor is selected based on the highest peak flux density within core-loss density limits. A laminated iron-core inductor is assumed on the grid side. For achieving higher accuracy during optimization, the grid-side inductor is modelled using an analytical high-frequency model. The proposed design method is evaluated using a 50-Hz 12.5-kVA SiC grid converter as an example case.

1 Introduction

Active two-level grid converters are replacing diode bridges in motor drives for being able to feed braking energy back to the grid. The grid converters cause switching harmonics that should be attenuated in order to comply with existing grid standards such as IEEE 519-2014 and IEC 61800-3 [1, 2]. An attractive way to attenuate the switching harmonics is to use an LCL filter between the grid and the grid converter. The LCL filter provides better attenuation than a conventional L filter, meaning that the same filtering performance can be obtained with smaller overall inductance value. This leads to a more compact filter size.

Various methods have been proposed for designing and optimizing the LCL filter [3–11]. Since the LCL filter is still one of the bulkiest components in grid converters, the most common design goal is to minimize the size of the filter. One typical method to achieve a compact size is to minimize the stored energy of the magnetic components in the filter at the rated operating point without exceeding the grid-current harmonic limits. A compact size is achieved, since the inductor energy is proportional to the inductor area product, which gives a rough estimate for the core size. The area product is defined by the product between the window area and the cross-sectional area of the core.

The selection of the inductor core material is essential since it defines key magnetic properties such as saturation flux density and core-loss density. The core material can be selected based on the peak flux density. The peak flux density should be as high as possible in order to minimize the area product. However, the peak flux density is limited either by the saturation flux density or the core-loss density. Low core-loss materials, such as amorphous or nanocrystalline, should be used in the converter-side inductor due to high converter-side current ripple. In this way, losses stay within limits defined by the acceptable temperature rise. On the other hand, the core material of the grid-side inductor are selected based on the highest saturation flux density since the grid-current ripple is low. An iron-core inductor is often a good choice for the grid side due to its high saturation flux density and low manufacturing costs.

Typical design methods assume constant parameter values in optimizing the LCL filter, which is justified if the converter operates at low switching frequencies, e.g. 4–8 kHz [3–8]. The switching frequency can be increased by using next generation semiconductors, such as SiC and GaN, without causing more switching losses in comparison to the conventional Si semiconductors. This gives the possibility to design a more compact LCL filter. The constant parameter model is usually enough for the converter side due to low core-loss density material. However, if higher switching frequencies are applied, the high-frequency behaviour of the grid-side inductor can be taken into account. This gives a possibility to achieve higher accuracy in meeting the existing grid standards.

Due to the skin and proximity effects, the equivalent series inductance of the inductor decreases and the equivalent series resistance increases with frequency. These frequency-dependent phenomena have various effects on the harmonic content of the grid current: the filtering capability of the LCL filter becomes poorer due to the lower inductance at higher frequencies, and the damping ratio of the resonance is increased due to the higher resistivity at higher frequencies. Frequency-dependent phenomena can be examined using a frequency-dependent inductor model, such as a model for laminated iron-core inductors presented in [12]. The frequency characteristics of the iron-core inductor depend on the sheet thickness. Thicker sheets increase equivalent series resistance and decrease equivalent series inductance. The sheet thickness is chosen depending on the current content. Thicker sheets (0.5 mm) are usually sufficient on the grid side due to low grid-current ripple.

This paper studies the effects of the switching frequency of a

Fig. 1: Equivalent circuit of an LCL filter. The parameters are the converter-side resistance R_{fc} , converter-side inductance L_{fc} , capacitance C_f , grid-side resistance R_{fg} , and grid-side inductance L_{fg} . The capacitor resistance and the grid impedance are neglected for simplicity.

grid converter on the LCL filter design. The stored energy of the magnetic components in the filter is minimized. An analytical model of the laminated iron-core inductors is considered during the optimization stage. The final core material of the converter-side inductor is selected based on the simulated current waveforms and datasheet values. A core material having the highest peak flux density is selected within the core-loss density and saturation limits. The proposed design method is evaluated for a 50-Hz 12.5-kVA SiC grid converter as an example case.

2 High-frequency model of the LCL filter

RMS-value scaled space vectors will be used for three-phase quantities. Both time-domain and frequency-domain vectors will be denoted by boldface lowercase italic letters. The argument reveals whether the time domain or the frequency domain is considered. The argument t represents time and the argument ω represents frequency. The imaginary unit is denoted by j . The magnitude of the vectors are denoted by plain uppercase italic letters. For example, the grid current vector is $\mathbf{i}_g$ and its magnitude is I_g .

2.1 Equivalent circuit

Fig. 1 shows a space-vector model of an LCL filter. The converter voltage is denoted by $\mathbf{u}_c$, the capacitor voltage by $\mathbf{u}_f$, and the grid voltage by $\mathbf{u}_g$. The converter current is $\mathbf{i}_c$. The frequency response from the converter voltage to the grid current can be derived from Fig. 1,

$$\frac{\mathbf{i}_g(j\omega)}{\mathbf{u}_c(j\omega)} = \mathbf{Y}(j\omega) = \frac{1}{a_3(j\omega)^3 + a_2(j\omega)^2 + a_1j\omega + a_0} \quad (1)$$

where

$$\begin{aligned} a_3 &= L_{fc}L_{fg}C_f & a_2 &= R_{fg}C_fL_{fc} + L_{fg}R_{fc}C_f \\ a_1 &= L_{fg} + L_{fc} + R_{fg}R_{fc}C_f & a_0 &= R_{fg} + R_{fc} \end{aligned}$$

2.2 High-frequency model of the inductors

Skin and proximity effects increase the resistivity and decrease the inductance of the inductor with frequency. The high frequency behaviour depends on the chosen core material. If

a low core-loss material, such as amorphous and nanocrystalline, is used, the skin and proximity effects stay negligible. Then, a constant-parameter model is sufficient. This is verified in Fig. 2(a), which shows the measured impedance of a Fe-based amorphous inductor as well as the impedance of the constant-parameter model. As can be seen, the constant-parameter model describes the high-frequency behaviour accurately below 100 kHz. In this paper, the converter-side inductor is assumed to have a low core-loss material in order to prevent overheating due to the high converter current ripple:

$$L_{fc}(\omega) = L_{fc}(0) \quad R_{fc}(\omega) = 0 \quad (2)$$

The grid-side current, on the other hand, contains only small current ripple. Then, a laminated iron-core inductor is a good choice for the grid side due to its high saturation flux density and low manufacturing costs. For laminated iron-core inductors, the equivalent series inductance and resistance of the inductor core can be modelled as

$$\begin{aligned} L_{fg}(\omega) &= L_{fg}(0) \frac{w}{\delta} \frac{\sinh \frac{\delta}{w} + \sin \frac{\delta}{w}}{\cosh \frac{\delta}{w} + \cos \frac{\delta}{w}} \\ R_{fg}(\omega) &= \omega L_{fg}(0) \frac{w}{\delta} \frac{\sinh \frac{\delta}{w} - \sin \frac{\delta}{w}}{\cosh \frac{\delta}{w} + \cos \frac{\delta}{w}} \end{aligned} \quad (3)$$

where δ is the skin depth of the iron sheet and w is the thickness of a single sheet [12]. In this paper, the sheet thickness is chosen to be $w = 0.5$ mm which is a typical value for 50-Hz inductors and transformers.

The skin depth is

$$\delta = \sqrt{2\rho/(\mu_r\mu_0\omega)} \quad (4)$$

where $\mu_0 = 4\pi \cdot 10^{-7}$ H/m, ρ is the electrical resistivity of the iron sheet, and μ_r is the equivalent relative permeability [12]. The strength of the eddy current effect depends on the relative permeability as can be seen from (3) and (4). The eddy current effect is stronger with higher relative permeability, which leads to poorer filtering capability. Typically, iron-core inductors with an air gap have relative permeability in the range of $\mu_r = 100$ –200. In order to ensure that particular LCL filter parameter combination is sufficient for the attenuation of the grid-current harmonics, the relative permeability can be modelled higher than the typical values. For this reason, the equivalent permeability is chosen to be $\mu_r = 300$ during the modelling stage.

Fig. 2(b) shows the measured impedance as well as the impedances of the equivalent core model and the ideal constant-parameter model. The equivalent core model is shown in two cases, i.e., for relative permeabilities of $\mu_r = 150$ and $\mu_r = 300$. As can be seen, both equivalent core models are more accurate than the constant-parameter model.

3 Design procedure

The LCL filter is optimized for a 12.5-kVA SiC grid converter using the switching frequency of 20 kHz. The rating of the

Fig. 2: Measured and calculated frequency responses: (a) converter-side Fe-based amorphous inductor; (b) grid-side laminated iron-core inductor.

converter is as follows: frequency of 50 Hz, line-to-line rms voltage of 400 V, rated current of 18 A. The DC bus voltage is 650 V.

3.1 Harmonic components in the converter voltage

The converter voltage reference is calculated based on the desired fundamental grid current using (1). The space-vector pulse-width modulation (SVPWM) method is used. The converter operates in the linear voltage range. Analytical expressions for the harmonic components of the converter voltage can be found in [4]. Here, however, the harmonic components were calculated numerically from the time-domain voltage waveforms. Using the SVPWM algorithm, the converter voltage was generated using a resolution time of $T_s = 10$ ns. Then, the harmonic components were determined by means of the discrete Fourier transform

$$u_c(jn\omega_1) = \frac{1}{N} \sum_{k=0}^{N-1} u_c(kT_s) \exp\left(-j\frac{2\pi kn}{N}\right) \quad (5)$$

where $\omega_1 = 2\pi f_1 = 2\pi \cdot 50$ rad/s is the fundamental angular frequency, $T_1 = 1/f_1 = 20$ ms is the period of the fundamental component, $N = T_1/T_s$ is the number of samples, k is the k th time-domain sample, and n is the n th harmonic component. Fig. 3 shows the spectrum of the resulting converter

$ n < 11$	$11 \leq n < 17$	$17 \leq n < 23$	$23 \leq n < 35$	$35 \leq n \leq 50$
4.0	2.0	1.5	0.6	0.3

Table 1: Maximum odd harmonic current components in percent of the rated current for distribution systems with 0.12–69 kV according to IEEE 519 [1]. Only the limits for the SCR below 20 are given. Even harmonics are limited to 25% of the odd harmonic limits.

Fig. 3: Spectrum of the converter voltage harmonics.

voltage at the switching frequency of $f_{sw} = 20$ kHz. The grid-current harmonics can be, then, solved from (1)–(3) for each converter-voltage harmonics.

3.2 Harmonic limits

The LCL filter is designed to meet the harmonic distortion limits of IEEE 519-2014 given in Table 1. The short circuit ratio (SCR) less than 20 is assumed. The rated grid current with unity power factor is considered. In this paper, a conservative approach is considered: all the grid current harmonics are attenuated below 0.25% of the rated current. Total harmonic distortion (THD) of the grid current is kept below 5%. Further, the converter-current THD is kept at 7% in order to limit the losses of the converter-side inductor and the converter. Both the positive and negative sequences are considered in the analyses.

3.3 Optimization

The goal is to design a compact LCL filter. The stored energy $W_{L,tot}$ of the magnetic components is minimized:

$$W_{L,tot} = \frac{3}{2} (L_{fc} I_c^2 + L_{fg} I_g^2) \quad (6)$$

The inductor energy W_L is proportional to the inductor area product (cf. Fig. 4)

$$A_p = W_a A_c = \frac{2W_L}{K_u J_m \hat{B}} \quad (7)$$

where W_a is the window area of the core, A_c is the cross-sectional area of the core, K_u is window utilization factor, J_m is the rms value of the current density in the window, and $\hat{B}$ is the peak value of the core flux density [13].

Fig. 4: Area product is the product between the window area W_a and the cross sectional area A_c of the core.

The area product gives rough estimate for the core size. For simplicity, the window utilization factor and current density are assumed as constant parameters. Thus, the area product depends on the inductor energy and peak flux density of the core. Optimal parameter values are searched through iterations.

3.4 Selection of the core material for the converter-side inductor

The converter-side inductor has to endure a higher ripple current content than the grid-side inductor. The high current ripple causes core losses which increase with switching frequency [14]. This sets higher demands on the core material, which has to be properly selected. Typical magnetic properties for different core materials can be found in Table 2. In order to minimize the area product (7), the peak flux density should be as high as possible. Thus, the peak flux density is the main criteria for choosing the material. The selection is limited either by the saturation flux density or the core-loss density.

To avoid overheating, the core losses should be kept low enough. The core-loss density gives a rough estimate for evaluating the temperature rise. For example, the maximum core-loss density is approximately 100 mW/cm³ under the natural convection condition [13].

It is assumed that all the ripple current is packed on the switching frequency. The total core-loss density can be expressed as $P_{c,tot} = P_{c,1} + P_{c,sw}$, where $P_{c,1}$ and $P_{c,sw}$ are the losses related to the fundamental component and the switching harmonics, respectively. Since the core-loss density is approximately proportional to the square of the flux density at particular frequency [14], the total core-loss density can be expressed as

$$P_{c,tot} = \left[\frac{B_1}{B_{ref}(f_1)} \right]^2 P_{c,ref}(f_1) + \left[\frac{B_{sw}}{B_{ref}(f_{sw})} \right]^2 P_{c,ref}(f_{sw}) \quad (8)$$

where B_1 is the flux density caused by the fundamental current component and B_{sw} is the flux density caused by the ripple current. The reference values for the core-loss density and flux density are $P_{c,ref}$ and B_{ref} , respectively.

The inductor is sized in such a way that the peak flux density $\hat{B}$ corresponds the peak current value $\hat{I}$ at the rated operating point. Further, it is assumed that the flux density and the current are directly proportional below the peak values, i.e.,

Fig. 5: Optimized parameters of the converter-side and grid-side inductors with switching frequency.

$B_1 = (I_1/\hat{I})\hat{B}$ and $B_{sw} = (I_{sw}/\hat{I})\hat{B}$ hold, where I_1 is the rms value of the fundamental current and I_{sw} is the rms value of the ripple current. Under these assumptions, the peak flux density can be solved from (8)

$$\hat{B} = \sqrt{\frac{P_{c,max}}{\left[\frac{I_1/\hat{I}}{B_{ref}(f_1)} \right]^2 P_{c,ref}(f_1) + \left[\frac{I_{sw}/\hat{I}}{B_{ref}(f_{sw})} \right]^2 P_{c,ref}(f_{sw})}} \quad (9)$$

If the core-loss density is the limiting factor for choosing the material, the peak flux density is determined by (9). Otherwise, the peak flux density is determined by the saturation flux density. The core material is selected based on the highest peak flux density.

4 Results

4.1 Optimized parameters

The optimized parameter values are $L_{fc} = 575 \mu\text{H}$, $C_f = 8.10 \mu\text{F}$, and $L_{fg} = 250 \mu\text{H}$. The proposed LCL filter design is compared with a conventional filter used with a grid converter having the switching frequency of 6 kHz [16]. According to the results, the inductance of the designed converter-side inductor is only 20% of the inductance of the conventional converter-side inductor, and the inductance of the designed grid-side inductor is 10% of the inductance of the conventional grid-side inductor. The capacitance is 80% of the capacitance of the conventional filter capacitor. The parameter values can be further decreased by increasing the switching frequency as can be seen from Fig. 5. However, it is worth noticing that the peak flux density might decrease with the switching frequency due to the higher core losses.

4.2 Simulation results

The proposed LCL filter design is evaluated by means of simulations. The grid converter is controlled using a two-degree of freedom state-feedback controller [16]. The grid-side inductor is modelled with two series-connected Foster elements shown in Fig. 6. The parameter values of the Foster elements

Material type	Manufacturer	Material	Saturation flux density (T)	Core loss @ 20 kHz, 0.1 T (mW/cm ³)
Amorphous	Metglas	2605SA1	1.56	70
Silicon steel	JFE	10JNHF600	1.88	150
Nanocrystalline	Vacuumschmelze	Vitroperm500F	1.2	5
Powder core	Magnetics	MPP	0.75	45
Ferrite	Ferroxcube	MnZn	0.52	5
Powder core	Arnold magnetics	High-Flux	1.5	130

Table 2: Typical magnetic properties of different core materials [15].

Fig. 6: Grid-side inductor modelled with two series-connected Foster elements.

Fig. 7: Frequency response of the analytical model and the Foster-equivalent model of the LCL filter.

are obtained by fitting the frequency response of the Foster elements with the analytical response. As can be seen from Fig. 7, two Foster elements are sufficient to capture the frequency behaviour below 100 kHz. Further, the resonance is damped due to the high grid-side equivalent resistance at high frequencies. It is worth noticing that the losses stay low due to the low grid-side current ripple. Fig. 8 shows the waveform and spectrum of the grid current. As can be seen, all the grid-current harmonics stay below 0.25% of the rated value. The THD of the grid current is 0.4%. Fig. 9 shows the waveform and spectrum of the converter current. The THD of the converter current is 7%.

4.3 Prototype inductors of the LCL filter

The core material of the inductor L_{fc} is selected according to (9). The peak value of the converter current is $\hat{I} = 27.5$ A and the ripple current is $I_{sw} = 1.3$ A. Above the switching frequencies of 20–30 kHz, the core losses are mainly caused due to the high converter ripple current. For this reason, the reference value for the fundamental part of the core-loss density is assumed to be $P_{c,ref}(f_1) = 0$ mW/cm³. The reference values for the ripple core-loss density can be found from Table 2.

Table 3 shows the area-product minimizing material for the

Fig. 8: Simulated waveform and spectrum of the grid current.

Fig. 9: Simulated waveform and spectrum of the converter current.

particular maximum core-loss density range. Under natural convection conditions, the optional materials are 2605SA1 and 10JNHF600. Under forced convection conditions, the 10JNHF600 material is the most suitable. The inductor prototypes were manufactured to be used under the natural convection conditions. The core material of the converter inductor was chosen to be 2605SA1. Fig. 10 shows the photographs of the manufactured prototype inductors.

Core material	Maximum core-loss density (mW/cm ³)
Vitoperm500F	0–23
2605SA1	23–82
10JNHF600	82–

Table 3: Area product minimizing core materials.

Fig. 10: Prototype inductors: converter-side (left); grid-side (right).

5 Conclusions

This paper studied the effect of switching frequency on the LCL filter design in grid-converter systems. The stored energy of the inductors were minimized. An analytical high-frequency model of the laminated iron-core inductor was considered. The core material was selected based on the peak flux density within saturation and core-loss density limits. According to the results, higher accuracy can be obtained using the proposed design method in comparison with the conventional design method based on the constant parameters. Based on the simulation results, the grid current harmonics stay within allowable limits defined by the grid current standards. Further, the converter current THD stays within the user-defined limits. The size of the LCL filter can be decreased significantly using higher switching frequencies, enabled by next generation semiconductors. A future work will include experimental verification of the proposed method.

Acknowledgment

The authors gratefully acknowledge ABB Oy and Walter Ahlström Foundation for the financial support.

References

- [1] “IEEE recommended practice and requirements for harmonic control in electric power systems,” *IEEE Std 519-2014 (Revision of IEEE Std 519-1992)*, June 2014.
- [2] “Adjustable speed electrical power drive systems—EMC requirements and specific test methods,” *IEC 61800-3*, Aug. 2004.
- [3] M. Liserre, F. Blaabjerg, and S. Hansen, “Design and control of an LCL-filter-based three-phase active rectifier,” *IEEE Transactions on Industry Applications*, vol. 41, no. 5, pp. 1281–1291, Sept. 2005.
- [4] K. Jalili and S. Bernet, “Design of LCL filters of active-front-end two-level voltage-source converters,” *IEEE Transactions on Industrial Electronics*, vol. 56, no. 5, pp. 1674–1689, May 2009.
- [5] Y. Jiao and F. Lee, “LCL filter design and inductor current ripple analysis for a three-level NPC grid interface converter,” *IEEE Transactions on Power Electronics*, vol. 30, no. 9, pp. 4659–4668, Sept. 2015.
- [6] K. B. Park, F. Kieferndorf, U. Drofenik, S. Pettersson, and F. Canales, “Weight minimization of LCL filters for high power converters,” in *9th International Conference on Power Electronics (ICPE-ECCE Asia)*, June 2015, pp. 142–149.
- [7] J. Muhlethaler, M. Schweizer, R. Blattmann, J. Kolar, and A. Ecklebe, “Optimal design of LCL harmonic filters for three-phase PFC rectifiers,” *IEEE Transactions on Power Electronics*, vol. 28, no. 7, pp. 3114–3125, July 2013.
- [8] B. G. Cho and S. K. Sul, “LCL filter design for grid-connected voltage-source converters in high power systems,” in *IEEE Energy Conversion Congress and Exposition (ECCE)*, Sept. 2012, pp. 1548–1555.
- [9] P. Channegowda and V. John, “Filter optimization for grid interactive voltage source inverters,” *IEEE Transactions on Industrial Electronics*, vol. 57, no. 12, pp. 4106–4114, Dec. 2010.
- [10] A. Reznik, M. Simoes, A. Al-Durra, and S. Mueen, “LCL filter design and performance analysis for grid-interconnected systems,” *IEEE Transactions on Industry Applications*, vol. 50, no. 2, pp. 1225–1232, March 2014.
- [11] E. Kantar, S. Usluer, and A. Hava, “Design and performance analysis of a grid connected PWM-VSI system,” in *8th International Conference on Electrical and Electronics Engineering (ELECO)*, Nov 2013, pp. 157–161.
- [12] G. Grandi, M. Kazimierzczuk, A. Massarini, U. Reggiani, and G. Sancineto, “Model of laminated iron-core inductors for high frequencies,” *IEEE Transactions on Magnetics*, vol. 40, no. 4, pp. 1839–1845, July 2004.
- [13] M. K. Kazimierzczuk, *High-Frequency Magnetic Components (2nd Edition)*. NJ, USA: John Wiley & Sons, 2013.
- [14] F. Dong Tan, J. Vollin, and S. Cuk, “A practical approach for magnetic core-loss characterization,” *IEEE Transactions on Power Electronics*, vol. 10, no. 2, pp. 124–130, Mar 1995.
- [15] M. Rylko, B. Lyons, J. Hayes, and M. Egan, “Revised magnetics performance factors and experimental comparison of high-flux materials for high-current DC-DC inductors,” *IEEE Transactions on Power Electronics*, vol. 26, no. 8, pp. 2112–2126, Aug. 2011.
- [16] J. Kukkola, M. Hinkkanen, and K. Zenger, “Observer-based state-space current controller for a grid converter equipped with an LCL filter: Analytical method for direct discrete-time design,” *IEEE Transactions on Industry Applications*, 2015, early access.