Passive LC Filter Design Considerations for Motor Applications

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***Abstract*—This paper provides design guidelines for the passive *LC* filters. Based on these guidelines, a method to design a new type of a passive filter, called hybrid *LC* filter, is proposed. A filter design example accompanies the considerations; simulation and test results of the proposed filter in time and frequency domains are shown.**

***Index Terms*—high-power drives, filters, common mode, differential mode, overvoltage.**

I. Introduction

PROGRESS in power electronics has brought new methods for the electrical machines' converter control. These converters are usually based on the pulse-width modulation (PWM) technique. A PWM sequence of pulses contains a useful base frequency and a set of high-frequency harmonics which are not required for common control purposes and all together produce electrical noise. Electrical noise is always undesirable for the electro-magnetic compatibility reasons. Besides, some of the high-frequency harmonics of the PWM are harmful for the electrical machines, basically because of two phenomena: reflections in a long cable and the rather low high-frequency impedance of the electrical machine via its phases through the bearings to the earth.

Reflections in a cable are known in the range 50 kHz – 2 MHz. They contribute to the so-called "differential mode" noise and result in overvoltages at the terminals of the electrical machines [1]. These overvoltages are harmful for the insulation of the windings [2].

At frequencies higher than 1 – 2 MHz the internal impedances of the electrical machine start to dramatically decrease since the stray capacitive couplings become dominating [3, 4, 5]. According to [3], part of the current flows through the motor bearings and leads to their accelerated wear out and, in large motors, to essential heating. This phenomenon is related to so-called "common mode" noise.

Electrical d*u*/d*t* filters are known as the universal and effective solution for differential and common mode noises attenuation. The existing solutions can include either only passive [1, 3–12] or passive and active components [13, 14]. The price of passive filters is rather low and their reliability is high, whereas active filters can have a significantly improved performance. Because of the low price passive filters are currently more popular. A typical filter topology is shown in Fig. 1.

A novel construction of the d*u*/d*t* filter called "hybrid *LC* filter" (HLCF) was proposed in [3, 7–9, 11, 12]. It was shown in [7] that, used in standard topologies, such a filter can compete with the traditional ones in performance, manufacturability and cost. The work presented in [8] proposes the way of understanding the electrical nature of the hybrid *LC* filter. The work reported in [11] discovers how to achieve the best usage of the internal capacitance of the HLCF. The work reported in [12] deals with the HLCF of the small power rating for switched-mode power supplies.

The specificity of the HLCF described in [3, 7–9, 11, 12] requires a new design procedure, which can be based partly on existing procedures. The design technique should take into account the fact that electrical and geometric properties of the HLCF are inseparably interconnected. Such an approach is not applied to the conventional *LC* filters. On the other hand, the design technique should include a complex of other factors that are usually taken into account in conventional *LC* filter and inductor designs. Some of these questions were already addressed in [1, 3, 5–10], and are combined into a complete method in this paper. The emphasis of the proposed design is mostly on d*u*/d*t* filters.

In literature we can find a number of references on the electrical design of conventional passive filters. Although the usage of *LC* filters without resistors is quite common, the authors prefer to consider *RLC* filters with overdamping resistors in series with the filter capacitances. However, such an approach leads to electrical and dimensional oversizing of the filter components. This results in extra expenses.

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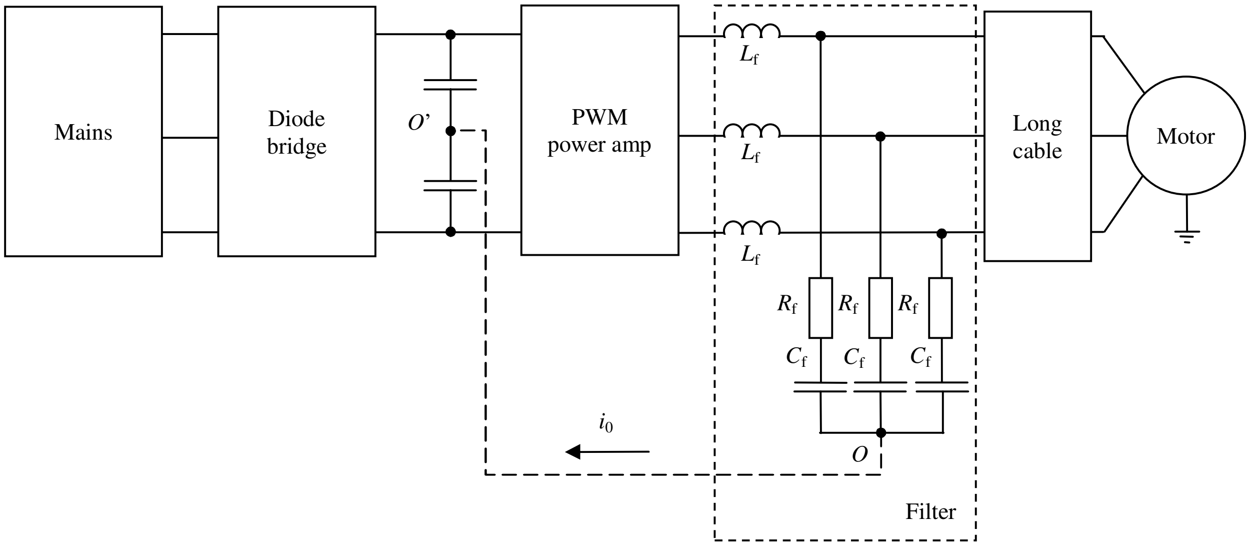


Fig. 1. A typical converter-based electrical drive topology with filter.

Quite a simple practical approach is presented in [1]. Inherently, the desired filter parameters are calculated from the motor cable parameters, the desired overshoot and the reflection coefficient. In [10] separate equations for the calculation of the filter inductance and capacitance are proposed. These methods do not take into account the stray impedances of the filter, even though [5] shows that the filter high-frequency performance may be still acceptable for a medium-power drive.

A modern engineer has a plenty of advanced tools for the versatile computer calculations and simulations, (just to mention free software programs such as LTSpice, Scilab, Octave, Maxima, Student's Quickfield, etc.), so that the filter parameters can be actually found using a simple trial-and-error method. Therefore, an engineer needs a good starting point and guidance for his investigations. An attempt to provide such guidance is done in this work with the accent on the hybrid *LC* filter design. Practice shows a good applicability of the method proposed for motor applications.

Many existing design procedures are intended for inductors with magnetic cores and cannot be applied as such to the HLCF design, in particular, because the selection of the

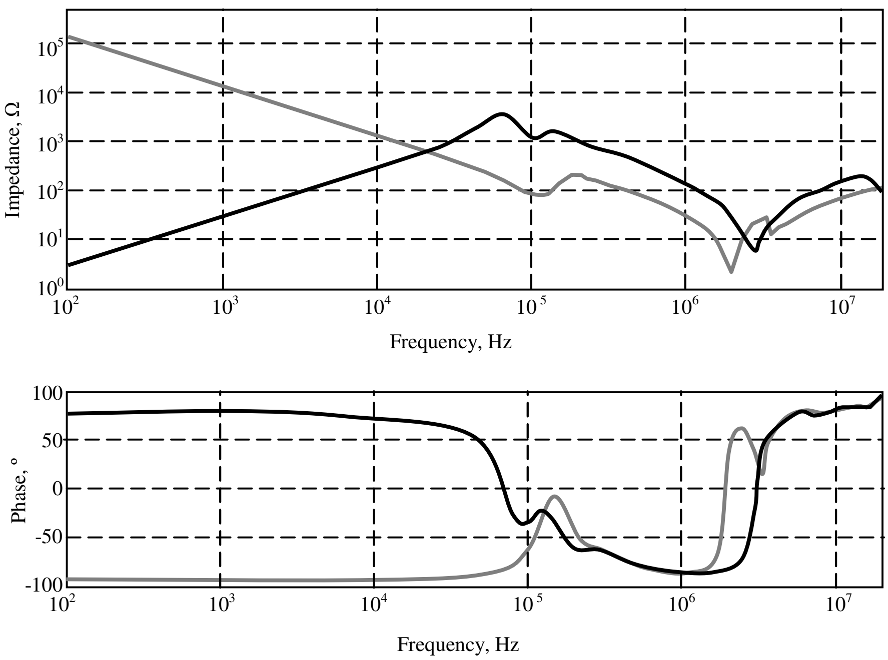


Fig. 2. Differential-mode (—) and common-mode (—), impedances of the Strömberg three-phase 22 kW induction motor used in the test setup.

magnetic core is a central part of these designs. In fact, magnetic materials could be successfully used in HLCF, but the subject requires separate considerations and is outside the scope of the current paper. On the other hand, a filter designer should pay special attention to the stray impedances, which along with the properties of the materials of the components determine the high-frequency performance of the filter. In particular, the parasitic distributed capacitance ("intra capacitance") as an element in parallel with the inductance of the coil is known [15, 16] and has a big influence, especially, in foil coil inductors.

The use of foils for inductor winding is considered, for example, in [17]. A large number of equations for inductors of different configurations was proposed in [18]. The equations are presented in a general form making them flexible for a wide range of applications.

The key idea of the current paper is to provide the designers with the basic guidelines for the *LC* filter design and, based on these, to propose a method of the HLCF design. To validate the design, a practical example is reported. Previously presented considerations in [7] and [8], are complemented with the considerations of the stray intra capacitance of the foils. The philosophy of the HLCF design, proposed in [3], is reworked so that a designer can synthesize the filter dimensions from the input electrical parameters. Test results measured for the new prototype are presented. They indicate that the HLCF technology is a good alternative to traditional *LC* filters.

II. General LC Filter Design Considerations

The initial data for the calculations are voltage rating of the drive U_{nom}, current rating of the converter I_{nom}, cable length l_{cab}, cable characteristic inductance L_{cab} (H/m) and capacitance C_{cab} (F/m), pulse rise time at electrical machine's terminals t_r without a filter. In our case $U_{\text{nom}} = 400$ V, $I_{\text{nom}} = 50$ A, $l_{\text{cab}} = 90$ m, $L_{\text{cab}} = 0.26$ µH/m, $C_{\text{cab}} = 0.55$ nF/m, $t_r = 0.12$ µs.

Because the *LC* filter has no extra resistive components, it is usually significantly cheaper than an *LCR* filter. However, resistance decreases overshoots and oscillations at the terminals and is typically recommended for implementation in applications with particular requirements on audible noise, speed and torque ripple. On the other hand, resistors dissipate significant losses, generate high temperatures, which sometimes means additional expenses for the cooling systems. Some guidelines on resistors selection are presented in [1] and [3]. Sometimes resistors can be avoided by implementing special control algorithms [14]. Anyway, damping methods lead to pulse rise times decreasing, and require searching for an optimum at the design. Reminding the designer the possible damping methods and disadvantages of the undamped circuits, this work concentrates on the *LC* filter design.

It has been shown in [2] that fast pulse rise times are stressful for the windings of the electrical machines, since up to at least 70 % of the voltage can be applied at the first coil. Slow enough rise times over a couple of µs permit equal voltage distribution across the winding, which mitigates the operating mode of the electrical machine. This tendency is also indicated in standards: the larger the rise time is, the higher are the permissible overvoltage levels. In accordance with rather recent IEC60034-25:2007 standard, 0.12 µs is not a problem for modern motors, since their windings are better insulated than the previous ones. However, for many existing motors, such as the Strömberg induction motor used in our test setup, 0.12 µs rise time might be destructive. For example, overvoltage allowed at rise-time 0.12 µs in the old IEC 60034-17:1998 standard is only 1 kV, which is not enough in our case, since 1.08 kV voltage peaks at motor terminals are expected. The evaluation of these peak voltages can be done by critical cable length estimation, as proposed in [19]. It is known that the voltage amplitude at the motor terminals, when compared to the voltage amplitude at the converter terminals, is doubled if cable length exceeds the critical value defined as

$$l_{\text{cr}} = \frac{t_{\text{r}}}{2\cdot\sqrt{L_{\text{cab}}C_{\text{cab}}}} = \frac{0.12\cdot10^{-6}}{2\cdot\sqrt{0.26\cdot10^{-6}\cdot0.55\cdot10^{-9}}} = 5 \text{ m} \quad . \quad (1)$$

Thus, cable length l_{cab} in our example exceeds the critical value l_{cr} at given voltage rise time and cable parameters. Since the frequency converter is fed by 400 V AC grid, the phase to phase PWM voltage amplitude measured at converter terminals equals to DC link voltage, i.e. 1.35 · 400 V = 540 V. In motor terminals doubled voltage compared to the DC link voltage, i.e. 1080 V, can be observed, which exceeds the margins presented in IEC 60034-17:1998 standard for the rise time 0.12 µs. Thus, a filter is required to protect the motor.

It is worth mentioning that if the pulse rise time values exceed 0.5 µs, both IEC 60034-17:1998 and IEC60034-25:2007 standards allow at least double voltages at the motor terminals. Therefore, *LC* filters without extra damping resistors can be implemented for the protection of the windings in applications with soft requirements on audible noise, speed and torque ripple.

Let us consider now the differential and common-mode impedances of the Strömberg 22 kW motor measured with HP 4194a analyser, which allows impedance measurement in the frequency range from 100 Hz to 40 MHz (Fig. 2). The differential-mode impedance was measured between two phase terminals connected together and the third phase terminal (winding is delta-connected internally). The common-mode impedance was measured between all phases combined together and the motor frame.

The motor differential-mode impedance starts on the inductive side. At 100 Hz the impedance is 2.8 Ω and increases as the frequency increases. At a frequency of about 70 kHz the impedance changes to the capacitive side. After that, between 1 MHz and 4 MHz the impedance phase angle changes from the capacitive side to the inductive side again. Thus, in the range from 70 kHz to approximately 4 MHz, the harmonics, generated by the frequency converter, tend to flow through the internal stray capacitances of the motor. Therefore, the filter should be capable to effectively attenuate these harmonics.

The motor common-mode impedance at 100 Hz starts at 130 kΩ and decreases as the frequency increases. The phase angle remains on the capacitive side up to about 2 MHz. At 100 kHz, the impedance is about 100 Ω, and the phase approaches zero after 100 kHz, but drops back close to –90º at higher frequencies up to about 2 MHz. The minimal impedance is about 2.8 Ω at 2 MHz. At higher frequencies the phase angle shifts to the inductive side. As the common-mode impedance is rather low at the whole range observed, it is easy to understand that common-mode currents can be high if large common-mode voltages are supplied to the motor. After approximately 2 MHz the common-mode impedance is on the inductive side, and the PWM harmonics are quite small, so that the motor impedance is large enough to damp the high-frequency harmonics.

Thus, the *LC* filter in the current design example should be capable to attenuate harmonics at least in the range of frequencies from 70 kHz to 4 MHz. Therefore, the attenuation at 70 kHz should be at least 3.03 dB or more. The resonance frequency of a filter can be obtained, for example, graphically, remembering that the slope in the attenuation range of the traditional *LC* filter is –40 dB/dec. Thus, the resonance frequency should be equal to f_{res1} = 44 kHz. The pulse rise time at the motor terminals of the drive equipped with the converter output *LC* filter can be approximately defined as:

$$t_r \approx \frac{1}{4 f_{\mathrm{res1}}} = \frac{1}{4 \cdot 44\mathrm{kHz}} = 5.7\ \mu\mathrm{s}, \tag{2}$$

which satisfies standard requirements as well as the approximate recommended value of the pulse rise time 5 µs provided in [20].

The cable model at high frequencies is a distributed parameter circuit, which contains a number of small elementary component inductances, capacitances and resistances [3, 4]. Therefore, a good criterion for converter output *LC* filter design is to keep the filter inductance and the capacitance equal or higher than the total inductance and capacitance of the cable. This way the filter parameters are dominating in the system.

An exact relation between the filter inductance and capacitance is difficult to define. Generally, the sum phase inductance of the filter and cable should not be more than 5% of the phase inductance of an electrical machine (considered as connected in “star”). Otherwise, the voltage drop at the sum inductance between the frequency converter and the electrical machine can be too high and the efficiency of the drive would be significantly decreased if the machine voltage is significantly lowered by the filtering.

On the other hand, too large a capacitance can cause the converter earth fault detector triggering in the frequency converter (the situation is rather similar with a failure caused by straight electrical connection of the cable conductor to ground). Moreover, boosting the capacitance in the hybrid *LC* filter is economically unjustified [7]. If resistors are implemented in the damping circuit, larger capacitance increases their power rating.

Thus, based on the discussions above, the filter inductance L_{m} is chosen equal to 120 µH and filter capacitance C_{b} is 110 nF.

III. Basics of the Hybrid LC Filter Design

The structure of the hybrid *LC* filter is described in [3, 7–9, 11, 12]. The filter contains two foil layers, separated by insulation and coiled on a supporting core. The layer, which is placed between the frequency inverter and an electrical machine, is called “the main foil”. Another foil, called “the auxiliary foil” is connected to a neutral, forming a useful capacitance between foils.

The target of the HLCF design is to find dimensions which result in the chosen inductance L_{m} and capacitance C_{b} of the HLCF.

The capacitance between the foils (“main capacitance”) of the HLCF can be calculated by equation

$$C_{\mathrm{b}} = \varepsilon_{\mathrm{r}} \cdot \varepsilon_0 \cdot \frac{\pi \cdot D_{\mathrm{mid}} \cdot N \cdot h}{d_{\mathrm{ins}}}, \tag{3}$$

where ε_{r} is the relative permittivity of the insulation material, ε_0 is the permittivity of vacuum, D_{mid} is the average diameter of the middle of the HLCF winding, N is the number of turns, h is the height of the HLCF and d_{ins} is the thickness of the insulation layer.

The inductance calculation is described, for instance, in [3, 18]. For a round air-cored solenoid with a rectangular shape of the winding cross-section, the following expression can be given:

$$L_{\mathrm{m}} = \frac{\pi}{4} \cdot \mu_{\mathrm{r}} \cdot \mu_0 \cdot N^2 \cdot \frac{D_{\mathrm{mid}}^2}{h} \cdot \left(K_{\mathrm{a}} - k\right), \tag{4}$$

where μ_{r} is the relative permeability of the core material ($\mu_{\mathrm{r}} \approx 1$ for air), $\mu_0 = 4\pi \cdot 10^{-7}$ H/m is the permeability of the free space and K_{a} and k can be found in special figures in [3] and tables in [18] that are based on the coil geometric dimensions. Alternatively, K_{a} and k might be found with the help of equations, which approximately describe the curves in [3] and provide a bit lower accuracy of calculations (the error is typically up to 10%):

$$K_{\mathrm{a}} = \frac{2}{\pi} \mathrm{atan}\left[2\left(\frac{h}{D_{\mathrm{mid}}}\right)^{\frac{\pi}{4}}\right], \tag{5}$$

$$k = 0.35 \cdot \tanh\left\{1.146 \frac{d_{\mathrm{w}}}{D_{\mathrm{mid}}} \mathrm{atan}\left[3\left(\frac{d_{\mathrm{w}}}{h}\right)^{-0.85}\right]\right\}, \tag{6}$$

where $d_{\mathrm{w}} = N \cdot (d_{\mathrm{m}} + d_{\mathrm{aux}} + 2d_{\mathrm{ins}})$ is the thickness of the HLCF winding, d_{m}, d_{aux} and d_{ins} are respectively the thicknesses of the main and auxiliary foils and insulation layer respectively.

Notice that equations (4), (5) and (6) are also valid for typical air-core inductors.

The number of turns N of the HLCF should be suitable for manufacturing. It is good to keep $N \leq 30$ at the design. In our case we selected $N = 16$.

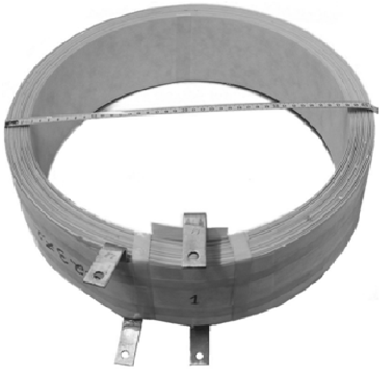


Fig 3. Designed hybrid LC filter, one phase.

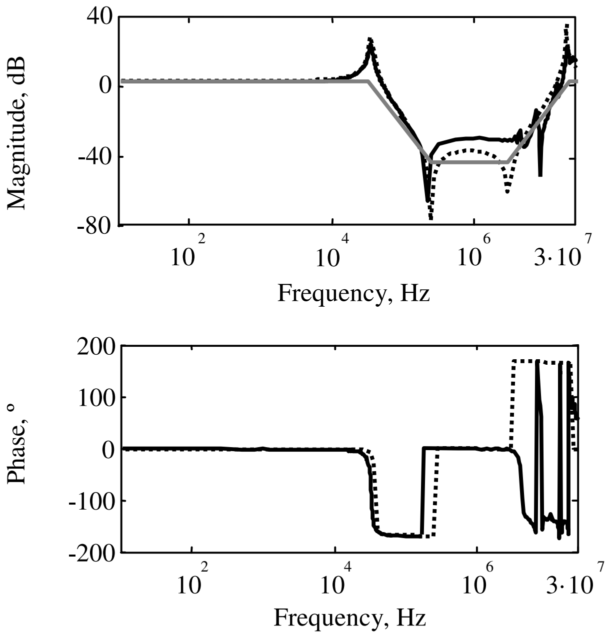


Fig. 4. Frequency response comparison: ▬ – measured, ▪▪▪▪ – simulated, ▬ – calculated using method presented in [8].

The foils of the HLCF should be mechanically firm enough to be capable of winding them on a core tightly. For the aluminum, which is a good material for HLCF manufacturing, thicknesses less than 0.1 mm can be critical. We used $d_{\mathrm{m}} = 0.5$ mm and $d_{\mathrm{aux}} = 0.1$ mm.

Relative permittivity ε_{r} depends on the insulator properties and is limited to the maximum value ε_{r}=7 given by mica. In order to provide a high capacitance, the insulation thickness d_{ins} should be as small as possible. In practice the reduction of the insulation thickness is restricted by the mechanical properties and dielectric strength of the insulation material. For the case under consideration we selected $d_{\mathrm{ins}} = 0.5$ mm Nomex insulator ($\varepsilon_{\mathrm{r}} \approx 4$).

The equation for the inductance is complex. This fact troubles the searching for the unique dimensions of the HLCF which would be suitable for both the desired inductance L_{m} and the capacitance C_{b}. However, HLCF dimensions can be found by iterations. For that purpose computer calculations are very useful and we should rewrite equation (3) as follows:

$$D_{\mathrm{mid}} = \frac{C_{\mathrm{b}} d_{\mathrm{ins}}}{\pi \cdot \varepsilon_{\mathrm{r}} \cdot \varepsilon_0 \cdot N \cdot h}. \qquad (7)$$

Increasing h in (7) step by step within a reasonable range of values, the calculation program obtains D_{mid} suitable for the required capacitance C_{b} and then calculates L_{m}. If L_{m} is equal to a desired value, D_{mid} and h are found. The outside diameter then equals to

$$D_{\mathrm{out}} = D_{\mathrm{mid}} + d_{\mathrm{w}}, \qquad (8)$$

whereas the inside diameter equals to

$$D_{\mathrm{in}} = D_{\mathrm{mid}} - d_{\mathrm{w}}. \qquad (9)$$

To save in the amount of materials and to decrease cost, it is recommended in [3, 7] to make an HLCF with as large diameter D_{out} as possible, whereas the height h of the HLCF is proposed to be kept as small as possible. With that, the height of the HLCF determines the current density in foils, which can be expressed as:

$$J = \frac{I_{\mathrm{nom}}}{d \cdot h}, \qquad (10)$$

where d is the thickness of the foil. It is recommended to keep the current density less than 2.5 A/mm^2.

For our example, equations (3) – (10) give $D_{\mathrm{mid}} = 0.373$ m; $d_{\mathrm{w}} = 0.0264$ m; $D_{\mathrm{in}} = 0.346$ m; $D_{\mathrm{out}} = 0.4$ m, $h = 0.08$ m. A photograph of the one phase of the hybrid LC filter designed is given in Fig. 3.

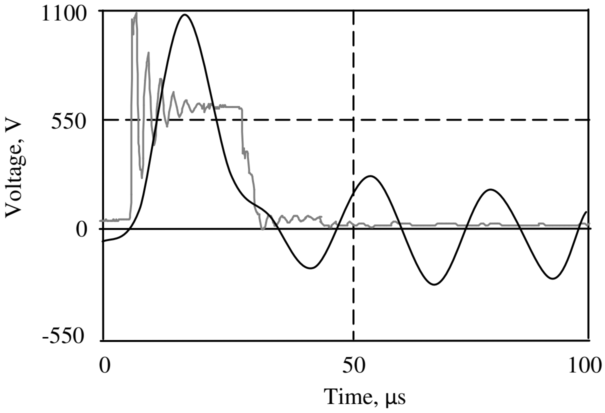
IV. Frequency Behaviour Estimation

Frequency behaviour estimation is highly recommended post-design process for any filter type. Let us estimate the frequency behaviour of the designed HLCF. For this purpose the intra capacitance C_{i} is taken into consideration. This capacitance is a set of capacitances between the turns of each foil connected in series; this is the reason why its value is much lower than the value of the capacitance between the foils C_{b}. The capacitance between the middle turns of winding is equal to $\frac{C_{\mathrm{b}}}{N-1} \cdot \frac{d_{\mathrm{ma}}}{d_{\mathrm{tt}}}$, where d_{ma} and d_{tt}, respectively, are the distances between the main and auxiliary foils and between the nearest turns of the same layer. The capacitances between other turns differ from the capacitance between the middle turns proportionally to their radii. Thus, if the insulation between all the layers has the same thickness, the intra capacitance of the foil can be estimated with the help of the main capacitance and the geometric relations as

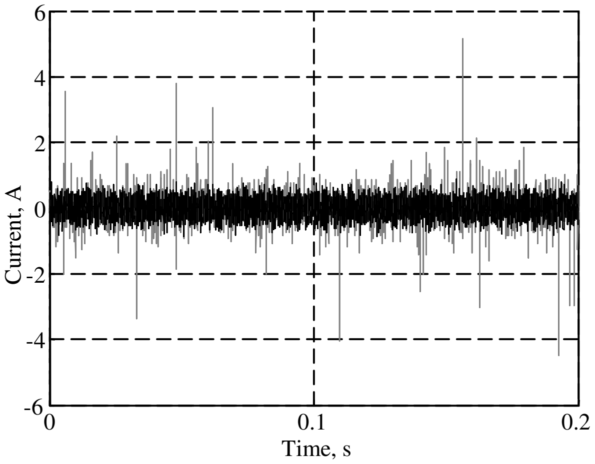
$$C_{\mathrm{i}}=\frac{1}{\sum_{N_i=1}^{N-1}\frac{N-1}{C_{\mathrm{b}}}\cdot\frac{d_{\mathrm{tt}}}{d_{\mathrm{ma}}}\cdot\frac{0.5D_{\mathrm{mid}}}{\left[0.5D_{\mathrm{in}}+\frac{d_w}{N-1}\cdot\left(N_i-0.5\right)\right]}}. \tag{11}$$

A reasonable rule of thumb is that earthing of the auxiliary foil reduces the intra capacitance of the main foil C_{i} approximately to $1/15^{\mathrm{th}}$. Thus, in our case the intra capacitance is $C_{\mathrm{i}}\approx 150\mathrm{pF}/15 = 10$ pF. The intra capacitance of the auxiliary foil can be neglected at the design or assumed unchanged [8].

It was also shown in [8] that the mutual inductance M between the foils of the HLCF plays a very important role in the attenuation at high frequencies. The larger the mutual inductance is, the better will be the attenuation level achieved and also the frequency range of attenuation will be larger.



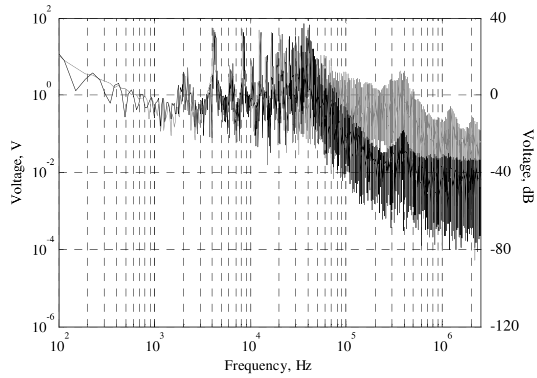
(a)



(b)

Fig. 5. (a) Phase-to-phase voltage and (b) common mode current in real drive: ▬ – with the hybrid LC filter, ▬ – without hybrid LC filter.

Typically, the mutual inductance between the main and the auxiliary foils even with nonprofessional winding rolling



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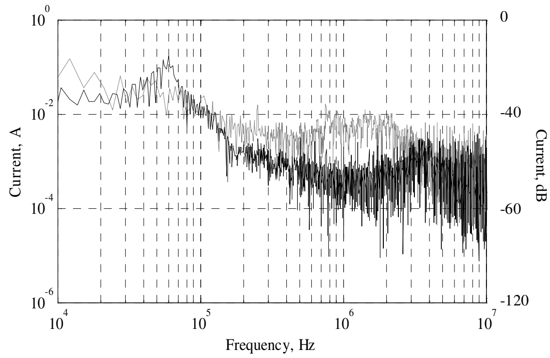

(b)

Fig. 6. (a) Phase-to-phase voltage spectrum and (b) common mode current spectrumin real drive: ▬ – with the hybrid LC filter, ▬ – without hybrid LC filter.

obtains values in the range $M = (0.9\div0.98)\cdot L_{\mathrm{m}}$. With the help of the equations and the model presented in [8], it is possible to conclude that the filter attenuation range is up to about 30 MHz, which suites the desired frequency range defined in Section II.

Fig. 4 shows the calculated, simulated and measured frequency responses of the designed filter. The attenuation of the noise in the operating frequency range is mainly about 35 dBs, or 1/56.

V. Experimental Results

In order to verify passive filter design considerations presented in this work, experimental investigations were done in Lappeenranta University of Technology. A 22 kW, 400 V, three-phase induction motor is driven by an ABB inverter with a PWM switching frequency of 2.14 kHz. The motor is connected to the inverter by a 90 m symmetric three-phase cable. The main parameters of the cable are: the DC resistance at 20 ºC is 0.524 Ω/km, the phase AC resistance at 50 Hz is 0.63 Ω/km, the phase inductance is 0.26 mH/km, and the operating capacitance is 0.55 µF/km. The inductance of the HLCF main and auxiliary foils is $L_{\mathrm{m}} = L_{\mathrm{a}} = 126$ µH and the capacitance between the main and auxiliary layers is C_b = 158 nF. The topology proposed in [5] was implemented and a wire connecting the filter star point to the DC link midpoint was used (Fig. 1). The measurements were made with a Yokogawa PZ4000 power analyzer, which provides a measurement frequency bandwidth up to 2 MHz. Differential voltage probes Testec TT-SI9002 with a frequency bandwidth up to 25 MHz were used for voltage measurements. High-frequency Rogowski coils PEM CWT075xB with bandwidths up to 15 MHz and Tektronix TDS3014B scope with a bandwidth up to 100 MHz were used for current measurements.

Fig. 5(a) shows the line-to-line voltages measured at the motor terminals. Without filtering the line-to-line pulse rise rates are about 5200 V/μs. The pulse rise rate was decreased to 81 V/μs, that is, to 1/64th of the original, when the hybrid *LC* filter was introduced into the system. The pulse rise time with HLCF is 5.3 μs versus 0.12 μs without HLCF. Again, in accordance with results measured in [3, 7] the HLCF influence considerably resembles the effect of an *LC* filter without damping, in other words, signals have significant oscillations.

Fig. 5(b) shows the common-mode current (current sensor embraces all three phase wires) with and without the filter. Again, the topology shown in Fig. 1 is used and the extra link between points *O* and *O*' is activated. Without HLCF the peak current is about 5.3A. When HLCF is introduced, peaks are decreased to 0.85 A, i.e. the common-mode noise is damped by a factor of 6.2 (about 16 dBs).

Fig. 6(a) shows the spectrums of measured voltage in the system with and without the HLCF. Notice that rightward axis is in dBs for evaluation purpose. This figure corresponds to the frequency responses presented in Fig. 4(a). When the HLCF is introduced into the system, a resonance at frequency about 40 kHz appears, which corresponds to the lowest resonance frequency of HLCF. After that, voltage harmonics are damped with the slope –40 dB/dec. Approximately at 350 kHz attenuation slope is zero, which also corresponds to Fig.3(a).

Fig. 6(b) shows the common mode current spectrum. Again, rightward axis is in dBs for evaluation purpose. When the HLCF is introduced between the frequency converter and cable, there is a resonance at 40 kHz, i.e. at HLCF resonance frequency. From about 150 kHz up to about 3.5 MHz filter provides attenuation of about 20 dB. At frequencies from 3.5 MHz to 10 MHz the attenuation is approximately 5 dB.

VI. CONCLUSION

This paper considers the passive *LC* filter design questions. Electrical design comprises following procedures: need in filter evaluation at given inverter PWM pulse rise time and cable length; required attenuation range evaluation; filter resonance frequency and desired pulse rise time at motor terminals estimation subjected to standards and, finally, desired inductance and capacitance finding. The filter design example accompanies the considerations.

Special accent in paper is invested in the new type of the *LC* filter called 'hybrid *LC* filter'. The filter's stray intra capacitance is taken into account in the design.

Simplified equations for the inductance estimation are presented in the paper. These equations can be used for the air-cored inductors calculation as well as for the hybrid *LC* filter design.

The design guidelines are reinforced by experimental results. With that, conclusions of the previously published papers are confirmed with the new hybrid *LC* filter prototype. The hybrid *LC* filter can replace the traditional ones in motor applications, effectively damping common-mode noises and decreasing the pulse rise-rates to the values permissible for the windings of the electrical machines.

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