

Article

Overvoltage Suppression Filter Development for GaN Inverter-Fed Electrical Drive with Long Cable Based on Impedance Measurement

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Abstract

Wide-bandgap transistors have short voltage rise and fall times, thus leading to overvoltage at the end of the cable connecting the inverter and the motor. In this paper, the overvoltage reduction possibilities have been investigated analytically, experimentally, and based on a simulation model. High-frequency models of the motor and the cable have been created based on impedance measurements. Different solutions for overvoltage reduction have been compared and an improved combined filter for the inverter with high switching frequency has been proposed. The overvoltage that was initially 80 percent has been reduced to below 10 percent by applying the filtering solution.

Keywords: GaN transistors; overvoltage reduction; overvoltage mitigation; inverters; impedance measurement; long transmission effect; motor drive; PMSM drive

1. Introduction

Permanent magnet synchronous motors (PMSMs) are used in many application areas due to high power density and efficiency. To control the speed and the torque of this kind of motor, variable frequency drives are used. Higher switching frequency allows for the development of faster control loops which gives the opportunity to improve the performance of servo drives. One potential rapidly developing field of application of such servo drives is robotics [1]. The efficiency of the fast-switching inverter can be increased by implementing Gallium Nitride (GaN) transistors [2], mainly due to faster turning on and off and thus reduced switching losses.

The schematics of a GaN transistor-based inverter connected to the three-phase motor through the cable can be seen in Figure 1. Fast turning on and off (high dv/dt) of GaN transistor switching action can lead to undesired oscillations and overvoltage in the cable connecting the motor and the inverter, and in the motor itself [3,4]. This can be explained with a parasitic capacitance and inductance of the cable and motor windings. The current in the bearing is also caused by the high dv/dt and this reduces the bearing lifespan [5]. Magnetic bearings, shielding or insulated bearings can be used to prevent bearing damage, but that complicates the design of the motor.

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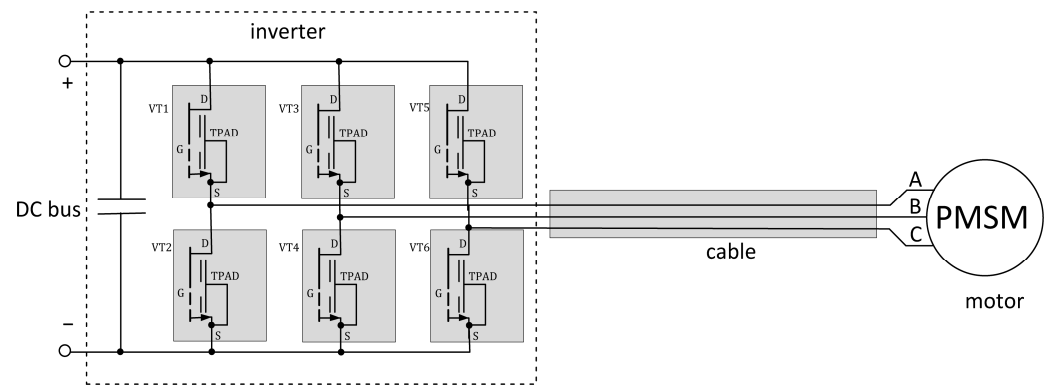


Figure 1. The schematics of a GaN transistor-based inverter connected with a three-phase motor.

Overvoltage is mainly caused by the impedance mismatch between the cable and the motor. This is often called a reflected wave phenomenon [6], and can be characterized by the reflection coefficient:

$$\Gamma_L = \frac{Z_M - Z_0}{Z_M + Z_0}, \quad (1)$$

where Z_M is impedance of the motor;

Z_0 —characteristic impedance of the cable per-unit-length.

The cable characteristic impedance considering lossless transmission can be calculated as follows:

$$Z_0 = \sqrt{\frac{L_0}{C_0}}, \quad (2)$$

where L_0 —inductance per-unit-length;

C_0 —capacitance per-unit-length.

To prevent this overvoltage, this fast voltage rise time should be limited by slowing the switching action of the transistors, thus increasing switching losses, or an additional dv/dt reduction technique is required [7], which usually adds some additional losses as well. Applications of sinusoidal filters have been analyzed in the scientific papers [8,9] and others. These filters have the potential to increase the system efficiency because the motor is fed with filtered, almost sinusoidal voltages, which increases the efficiency of the motor itself [10], but the filter increases the size of the system, adds losses, and limits the torque response.

A filter placed between the drive and the motor that reduces just the dv/dt at the motor terminals is another option, which is used quite often. In [11], the dv/dt filter and sine wave filter options were compared, and it was concluded that the dv/dt filter has a higher efficiency and smaller size. Usually, dv/dt filters have passive damping to prevent resonant oscillations. Although different solutions are proposed in the literature, with active control-based damping such as that presented in [12–14], they are not that easy to implement and therefore classical damped filter is most often used in the industry.

In the literature, different methods have been proposed for a passive filter with passive damping design [15–18]. In [15,16], it was proposed to use simplified equations for filter component selection without accurate estimation of overvoltage in case of a long cable. Ref. [17] provides the selection of RLC filter components based on simulations. In [18], approximate value of motor impedance was used to design the filter. In [19], the voltage reflection theory was used for optimal filter value selection. The cable itself can also be used as a part of the filter; such approach has been demonstrated in [20,21]. Different models of the high-frequency impedance of the cable and the motor have been presented in the literature, for example, in [22,23], models for motors are available. In most

cases, these impedances have been used for electromagnetic emission estimation like in [24]. At the same time, high-frequency impedance of the motor and cable can be used not only to estimate overvoltage, but also to investigate the filter influence on the overvoltage process. Such an approach will be used in this article.

In this article, different models from the literature are combined to get acceptable accuracy and to avoid very complex models. High-frequency model parameters will be estimated based on impedance measurements. The paper describes GaN transistor-based inverter prototype and suggests how to simulate the switching process of such an inverter. Next, it provides a concise, practical comparison of common filtering approaches to show that passive filtering is a competitive solution. Subsequently, it analyzes the high-frequency impedances of the cable and the motor to obtain information on overvoltage at the end of the cable. Improved solutions are proposed for distributing filter components across both ends of the cable (inverter-side and motor-side), aiming to enhance the damping performance, reduce overvoltage and oscillations, and improve overall system efficiency. To address the challenges of compactness and integration in high-frequency GaN transistor-based drive systems, this paper proposes the design of planar inductors for filter applications. Practical experimental results will be shown to verify the effectiveness of the proposed solution.

2. The Model and the Experimental Prototype of GaN Transistor-Based Three-Phase Motor Inverter

By replacing traditionally used Si transistors with GaN transistors, the efficiency of the inverter can be improved, mainly due to lower switching losses. One of the main differences from Si transistors in terms of gate control is that the gate threshold voltage $V_{GS(th)}$, the gate plateau and the maximum gate voltage $V_{GS(max)}$ are lower. To turn off the GaN semiconductor, the gate voltage should be maintained below the minimum gate threshold voltage, which is around 1 V. This can be a challenge in the applications where the GaN transistor drain is exposed to a high dv/dt , as undesired turning on of the transistor can occur. Gate drivers must have high common-mode transient immunity. The integrated circuit of the driver should be placed near the transistor gate to limit stray inductances and capacitances. High dv/dt causes high common-mode currents; this is the reason why it is important to minimize coupling capacitances of an isolated power supply and use a gate driver with high common-mode transient immunity.

The electrical circuit of the three-phase inverter can be seen in Figure 1. It consists of six transistors that are connected to each motor phase and powered by a DC voltage source or rectified AC voltage. To control the motor, a digital controller is used, which measures the currents and voltages and controls the speed and the torque. Considering the previously mentioned experimental inverter design, isolated driver circuits have been selected. Part of the inverter's printed circuit board (PCB), which includes one half-bridge of the transistors and the driver, can be seen in Figure 2. To isolate the gate-driving voltages, low-power switched-mode power supplies (2) with high isolation voltage and low capacitance between primary and secondary side have been selected. The driver circuit (3) and half-bridge of the GS-065-004-1-L GaN transistors (Infineon, Neubiberg, Germany) (4) can be seen in Figure 2. For current measurement, hall effect-based current sensor (1) is also placed on the board.

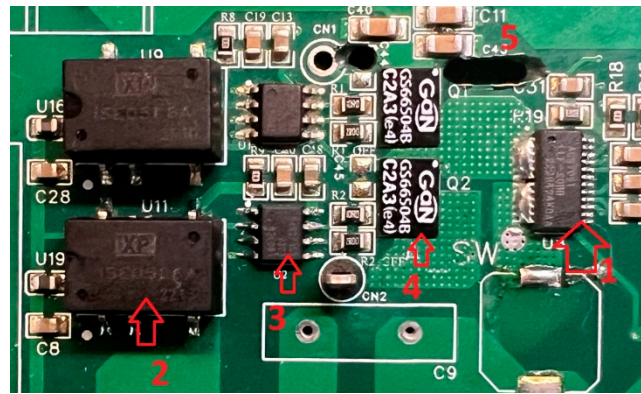


Figure 2. One inverter leg of a GaN transistor-based three-phase inverter PCB: 1—current sensor; 2—isolated power supply for drivers; 3—driver integrated circuit with isolation of driving signal; 4—GaN transistor, 5 – drills to insert current probe of an oscilloscope.

The developed PCB of the three-phase motor inverter can be seen in Figure 3. A STM32G411 microcontroller (STMicroelectronics, Geneva, Switzerland) has been used to control the inverter. This microcontroller offers features like filter math accelerator (FMAC), high-resolution timer (HRTIM), with the ADC and processor working at a frequency up to 170 MHz and are available for relatively low cost. This microcontroller is suitable for controlling power electronics converters and electrical drivers due to advanced driver capability.

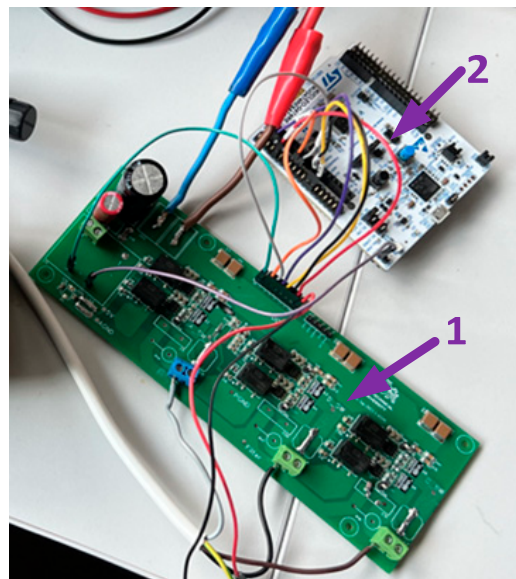


Figure 3. PCB of the inverter (1) with microcontroller control board (2) connected to it.

To investigate the voltage reflection and to test different filter designs, a simulation model has been created. The simulation model consists of one leg of the inverter that is very similar to synchronous buck converter. The simulation was created in LTSpice XVII simulation software. This simulation model can be seen in Figure 4. To verify the accuracy of the Spice model of the transistors, simulation-based results were compared with experimental ones. The voltage can be seen in Figure 5 between the drain and the source of one of the transistors, which is the switching node voltage V_{sw} . The curve was obtained from the LTSpice simulation model shown in Figure 4. Additional inductance was placed on the DC bus to adjust the waveform to be like the experimental one. In the simulation, the duty cycle is 0.5 and the driving pulse was supplied to the driver at the 50th nanosecond.

The oscillations are caused by parasitic inductance of the PCB traces, capacitor parasitic inductance, and the package of the transistors. This inductance can be measured, calculated or adjusted to fit the experimental waveform as in this case. If a proper inductance value is added to the simulation model, then it describes the transient process with high accuracy, and this simulation model can be used to analyze the overvoltage and different filter solutions. Only an accurate enough high-frequency model of the cable and the motor itself are needed.

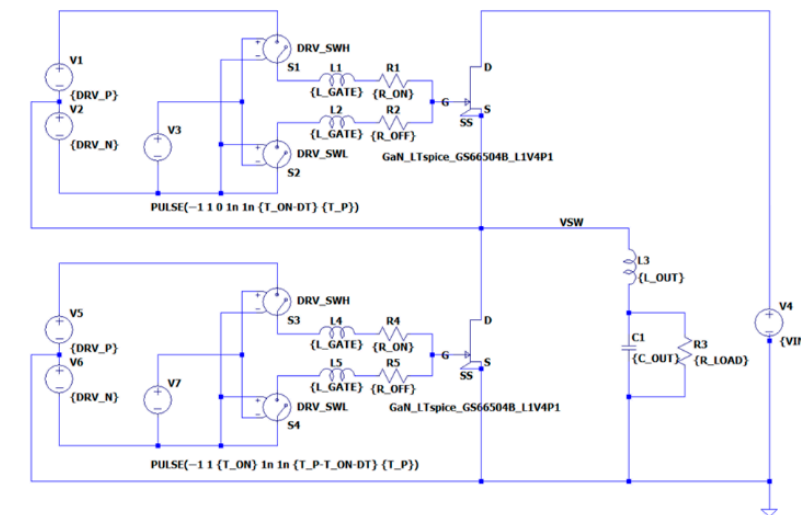


Figure 4. LT Spice simulation model of GaN transistors in half-bridge connection.

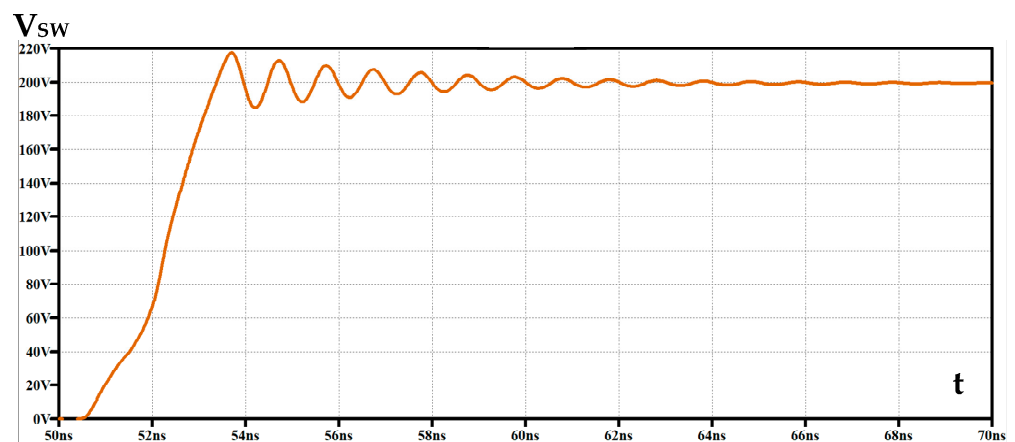


Figure 5. Switching node voltage (V_{sw}) during switching action obtained from simulations by adjusting parasitic inductance from experimental measurements.

3. High-Frequency Model of the Cable and the Motor

To understand the process of reflection and to develop solutions to mitigate these effects, it is important to obtain the high-frequency models of the electrical motor and the power cable connecting the inverter and the motor. The parameters of the equivalent circuit can be determined by measuring the differential-mode (DM) and common-mode (CM) impedance of the cable and the motor in the frequency domain [25]. The connections to measure such impedances can be seen in Figures 6 and 7. In a similar way, the cable was connected for the measurements of the impedance.

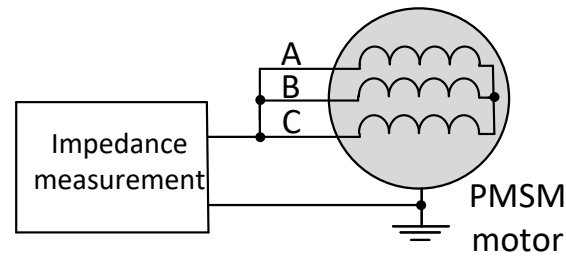


Figure 6. Connections of the motor phases for differential mode impedance measurement.

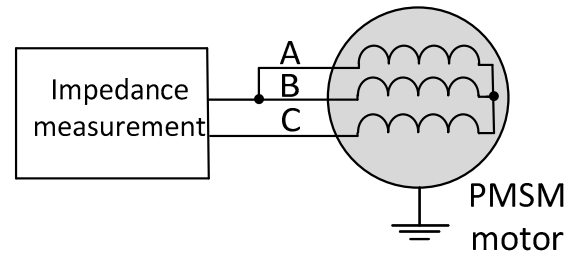


Figure 7. Connections of the motor phases for common mode impedance measurement.

The experimental measurement setup can be seen in Figure 8. To measure the impedance, an AP310 (Ridley Engineering, Camarillo, CA, USA) frequency response analyzer (2) was used, and the data were transferred and stored on the laptop (1). For the experiments, a 1 kW 3000 rpm MXL-10A0830FA222 (Trio Motion Technology, Tewkesbury, United Kingdom) PMSM motor (4) was used. A 33 m long cable consisting of 4 wires with a 0.75 mm² cross-sectional area (3) was used for this research.

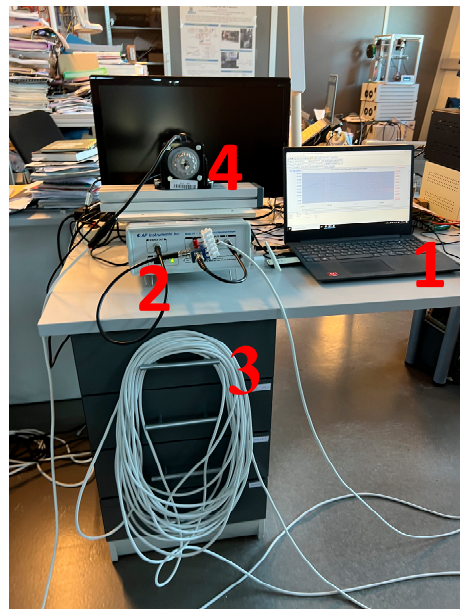


Figure 8. Experimental setup to measure cable and motor impedances: 1—laptop for storing measured results; 2—AP310 analyzer; 3—cable used in experiments; 4—PMSM motor.

The measurement results can be seen in Figures 9–12. From these measurement results, it is possible to obtain the values of the equivalent circuit representing the motor and the cable for high frequencies. In the literature, several equivalent circuits have been proposed that describe motor or cable impedance with good accuracy. In Ref. [26], a

circuit has been proposed that models the AC motor, including the high-frequency model and the dynamic dq model; other equivalent circuits have been proposed in [24] for electromagnetic emission investigation. In Ref. [27], a lumped-parameter cable model has been proposed. The review of different motor models can be found in [3], where it is stated that in the case of a PMSM motor, the position of the rotor can influence impedance graph, but in this particular case such an impact has not been observed by practical measurements. A more complex model can be found in [28]. In this case, a similar model was used as in [29], where a universal high-frequency model of induction motor was proposed. The values of model elements can be obtained from the measured impedance curves.

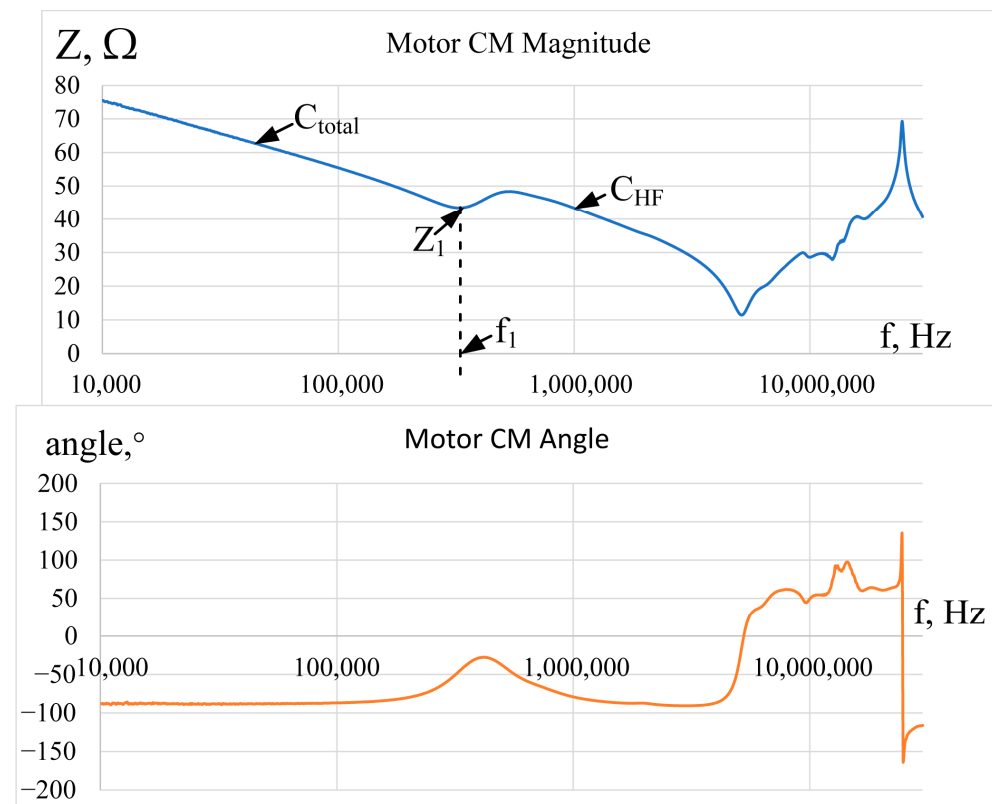


Figure 9. Motor common-mode impedance frequency response.

In Figure 13, high-frequency differential mode and common models of the motor can be seen. The values of model elements can be obtained from the measured impedance curves [29]. The measured DM and CM impedance magnitudes and phases and the most important points for the model characterization are shown in Figures 9 and 10. The total winding-to-ground capacitance C_{total} and the high-frequency winding-to-ground capacitance C_{HF} can be extracted from the slopes of common-mode impedance shown in Figure 9. C_{total} characterizes the low-frequency range of curve shown in Figure 9, but C_{HF} is the straight section of the impedance curve at a high frequency. From that graph, for motor impedance measurements, values of C_{total} and C_{HF} can be estimated by selecting random point on these straight sections of curve where the phase is close to -90° , as shown in Figures 9 and 10, and calculating the capacitance using the well-known equation ($C = 1/2\pi fZ$). Capacitances C_{g1} and C_{g2} used in the motor are as follows [30]:

$$C_{g1} = \frac{1}{3}C_{HF}, \quad (3)$$

$$C_{g2} = \frac{1}{3}(C_{total} - C_{HF}). \quad (4)$$

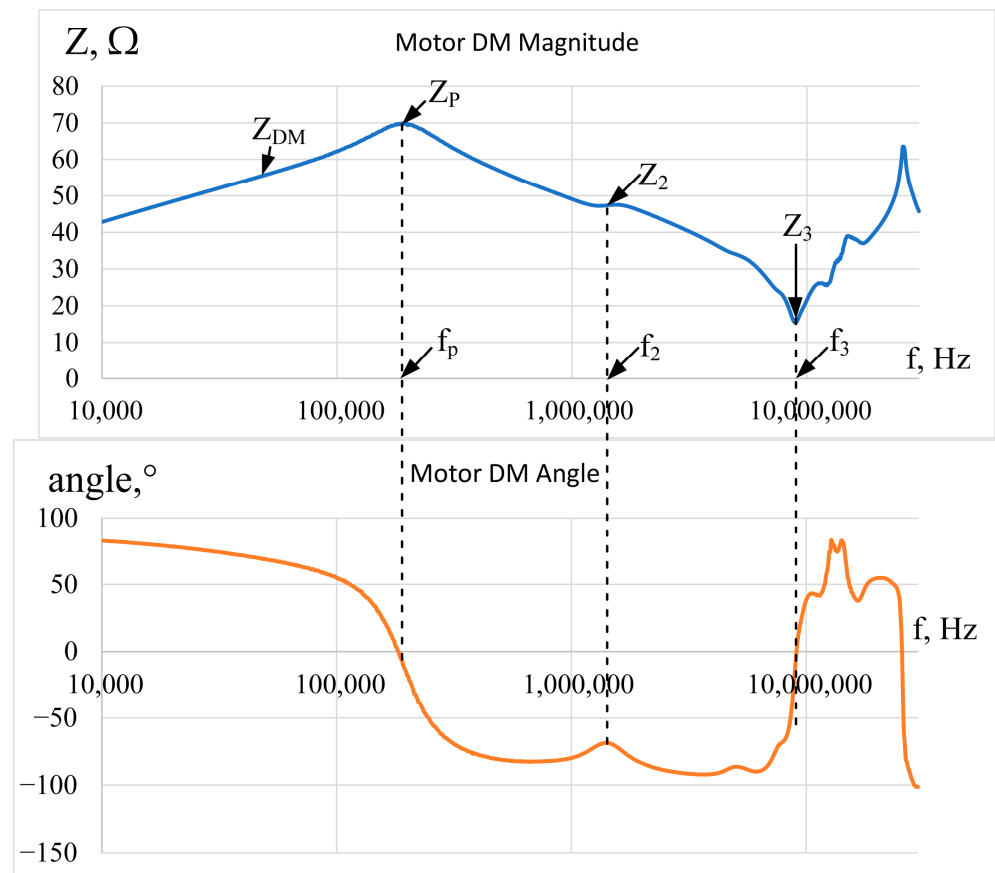


Figure 10. Motor differential-mode impedance frequency response.

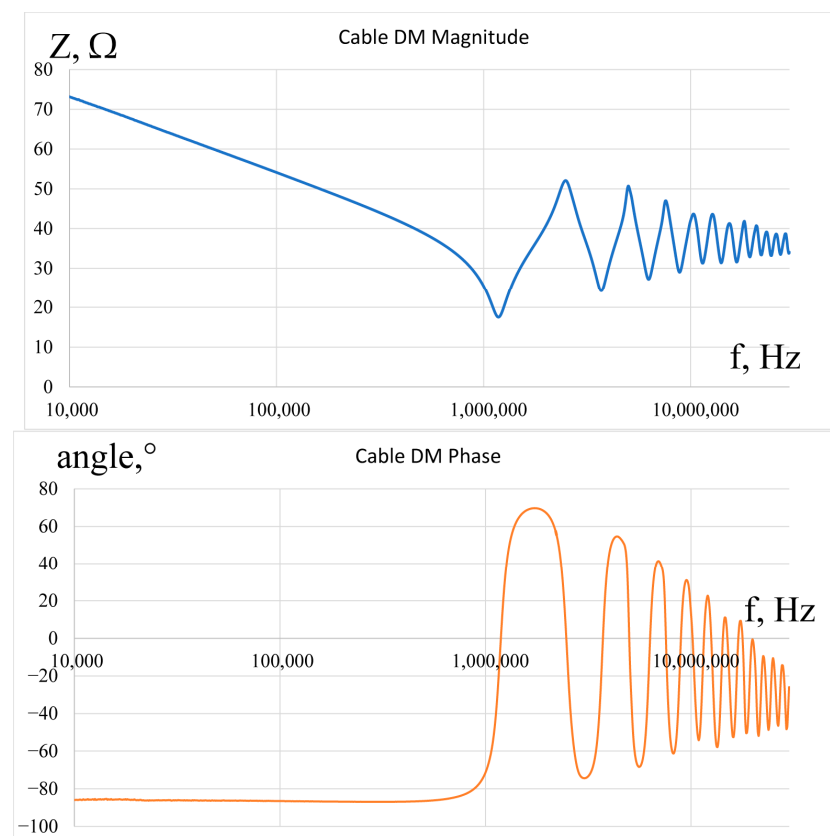


Figure 11. Cable differential-mode impedance frequency response.

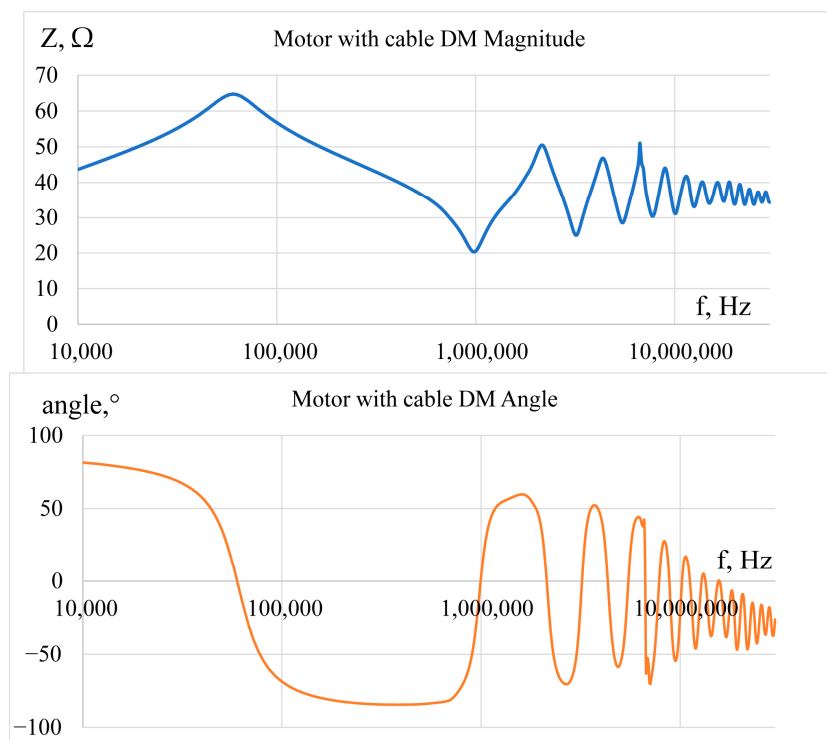


Figure 12. Motor with cable differential-mode impedance frequency response.

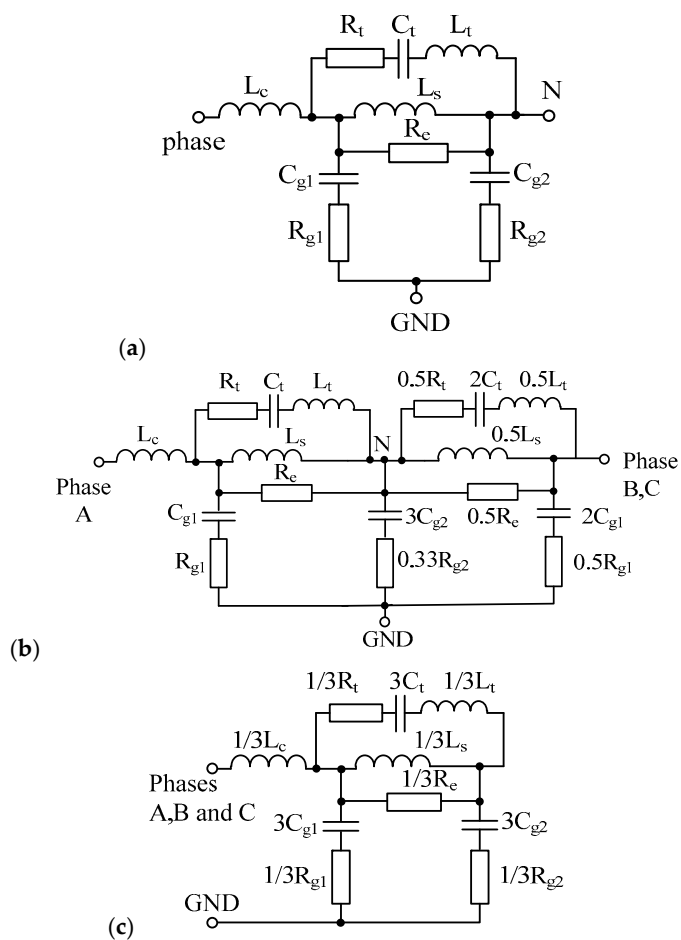


Figure 13. High-frequency model of PMSM motor: (a) per-phase equivalent circuit; (b) three-phase differential mode circuit model; (c) three-phase common-mode circuit model.

The winding-to-ground equivalent resistances, R_{g1} and R_{g2} are calculated from the impedance Z_1 resonant point shown in Figure 9 and the impedance Z_3 resonant point in Figure 10. These resonance points can be recognized at the place where the straight curve changes direction and the phase is close to 0° . At these resonance points, inductive impedance compensates for capacitive impedance and impedance is equal to resistance. Corresponding resistances in the motor equivalent circuit can be calculated as follows:

$$R_{g1} = \frac{2}{3}Z_3, \quad (5)$$

$$R_{g2} = 3Z_1. \quad (6)$$

The differential-mode inductance L_{DM} can be calculated from measured differential impedance shown in Figure 10 by selecting some random point on the low-frequency slope where the phase is close to 90° and calculating using the well-known equation ($L = Z/2\pi f$). In this case, this inductance is equal to 2 mH. Common-mode inductance L_{CM} can be calculated using the impedance Z_1 at the resonant point, at which capacitive impedance is equal to inductive impedance and thus it can be calculated as follows:

$$L_{CM} = \frac{1}{12\pi^2 C_{g2} f_1^2}. \quad (7)$$

Stator leakage inductance L_s can be obtained from previously obtained inductances as follows [30]:

$$L_s = L_{CM} + \frac{4}{9}L_{DM}. \quad (8)$$

The high-frequency loss resistance R_e , similarly to R_{g1} and R_{g2} , can be determined from the resonant point impedance Z_P in Figure 10 as follows:

$$R_e = \frac{2}{3}Z_P. \quad (9)$$

The capacitance C_t can be determined as follows [26]:

$$C_t = \frac{1}{10}(C_{g1} + C_{g2}). \quad (10)$$

The inductance L_t and resistance R_t for the series resonance are calculated from the impedance Z_2 at the resonant point as follows [23]:

$$L_t = \frac{1}{4\pi^2 C_t f_2^2}, \quad (11)$$

$$R_t = Z_2 \cdot \cos(\theta_{Z_2}), \quad (12)$$

where θ_{Z_2} is the phase angle at frequency f_2 from measured differential mode impedance shown in Figure 10.

Inductance L_c can be calculated from the resonant point Z_3 at which capacitive impedance is equal to inductive impedance as follows from [29]:

$$L_c = \frac{2}{3} \left(\frac{3}{8\pi^2 C_{g1} f_3^2} \right). \quad (13)$$

The values of the elements of the high-frequency model of the motor calculated by Equations (3)–(13) can be seen in Table 1.

Table 1. Component values of the high-frequency motor model.

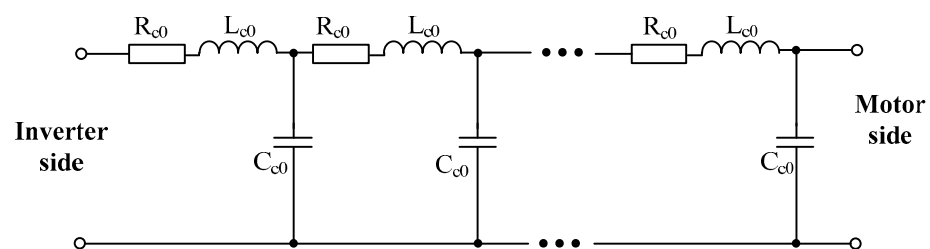
Circuit Element	Estimated Value
C_{g1}	38.6 nF
C_{g2}	51.1 nF
R_{g1}	0.84 Ω
R_{g2}	445.7 Ω
L_s	1.1 mH
R_e	2039 Ω
C_t	89.78 pF
L_t	2.7 mH
L_c	82.4 μ H
R_t	149 Ω

From the cable differential-mode impedance shown in Figure 11 capacitance, inductance and resonant frequency can be determined in a similar way, as in the case of the motor model.

Simple models consisting of RLC branches are mostly used for cable modelling. The electrical circuit of such a model can be seen in Figure 14. The impedance of the cable can be calculated from the geometrical parameters of the cable; a simplified equation has been presented in [31]:

$$Z_0 = 90 \cdot \ln \left(\frac{2(z+d)}{d} \right), \quad (14)$$

where z —thickness of the insulation of the cable strands; d —diameter of the copper of the cable strand.

**Figure 14.** A simplified model of the cable consisting of per-meter resistance, capacitance and inductance.

From the impedance curve, the determined capacitance is $C_0 = 110$ pF/m, and the inductance is equal to $L_0 = 0.53$ μ H/m, while resistance $R_0 = 20$ m Ω /m. The corresponding cable impedance, calculated by (2) and (14), is approximately equal to 67 Ω . The propagation delay in the long cable can be calculated as follows:

$$\tau = l \sqrt{L_0 C_0}, \quad (15)$$

where l is the length of the cable. In this case, by using (15), it can be determined that propagation delay is equal to 250 ns.

Also, there is a possibility to use more complex cable models such as those presented in [25,26,32]. One such more complex model, based on Figure 12 and the approach given in [29], has been created, and the obtained simulation results gave just less than a 15 percent difference compared to the simplified model, meaning that the use of simplification is acceptable for the overvoltage filter design.

The accuracy of the model also depends on the number of branches employed, although this factor has only a minor influence. To evaluate the impact of the filter components on overvoltage reduction, the equivalent circuit shown in Figure 15 was used. Kirchhoff's voltage law was applied to derive the governing equations, which were subsequently solved using numerical methods. Kirchhoff's voltage laws have been applied to

obtain equations that can be solved by using numerical methods. In this case, Wolfram Mathematica software was utilized. To reduce the complexity and time required for the calculation, the cable model was limited to four branches.

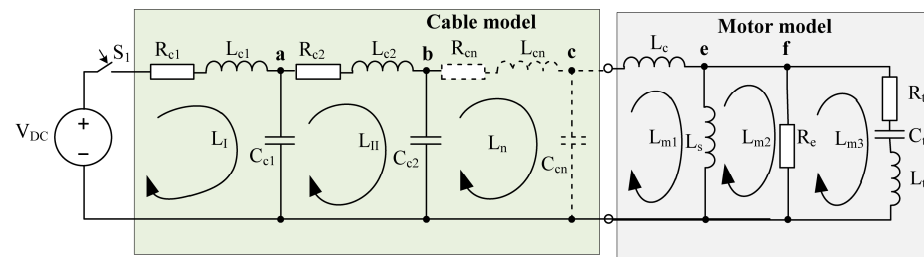


Figure 15. Equivalent circuit to analyze overvoltage mitigation filter analytically, arrows show loops (L_i) for writing Kirchoff's loop rule equations for this circuit.

4. Overvoltage at the Motor Terminals

The long cable combined with high dv/dt of GaN transistors causes reflections which are leading to overvoltage. The reflection coefficients at the motor side (Γ_{mot}) and at the inverter side (Γ_{inv}) are often used to describe the process in the cable; the reflection coefficient at the motor side can be calculated as follows:

$$\Gamma_{mot} = \frac{Z_m - Z_c}{Z_m + Z_c}, \quad (16)$$

where Z_m is the motor impedance; Z_c is the impedance of the cable. The reflection coefficient at the inverter side can be calculated as follows:

$$\Gamma_{inv} = \frac{Z_{inv} - Z_c}{Z_{inv} + Z_c}, \quad (17)$$

where Z_m is the motor impedance.

As the impedance of the motor is usually higher than the impedance of the cable, the reflection coefficient is positive and, in the worst case, it can be equal to 1. In that case, the voltage will reach two times higher voltage than the DC voltage, which can destroy the isolation of the motor. The impedance of the inverter is low and, therefore, the reflection coefficient at the inverter side is usually negative. This is the reason why the third reflected wave is negative and is going to reduce the further growth of the voltage. This reflection process can be seen in Figure 16.

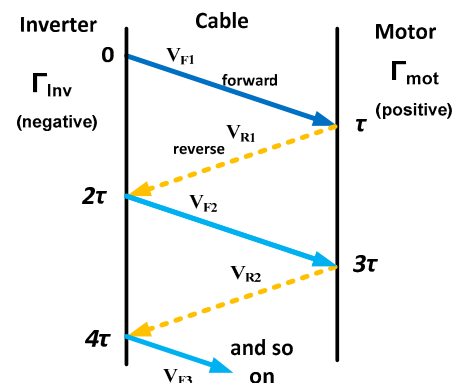


Figure 16. An illustration of inverter signal reflections if the transistor rise time is faster than the propagation delay in the cable (arrows show travelling wave: blue – forward, yellow reverse wave).

The simulation model shown in Figure 4 has been extended with the cable and motor equivalent circuits obtained previously. The cable was simulated in the LTSpice software

with 33 branches connected in the series connection. As a simulation result, it is possible to obtain the voltage at the motor terminal, which can be seen in Figure 17.

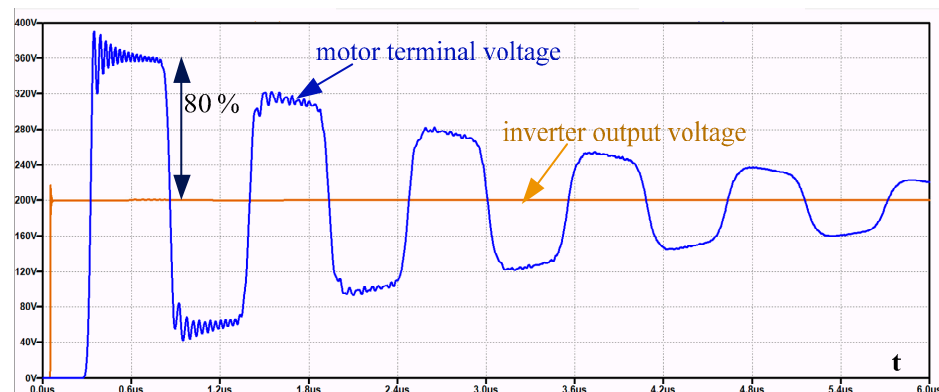


Figure 17. A simulation-based waveform of the voltage at the motor side and inverter side.

Similar results can be measured experimentally by supplying the PMSM motor through the long cable with high dv/dt signals generated by GaN transistor-based three-phase inverter. An experimental setup can be seen in Figure 18. The signals measured by the oscilloscope can be seen in Figure 19. Around 80 percent overvoltage was observed in simulations and experiments. Experimental results confirm the previously determined propagation delay, which can be determined from the interval between two peaks which takes four reflections. Since the simplified model of the cable is used, the waveform is not fully accurate, but describes the overvoltage and the waves travelling into the cable with good enough accuracy.

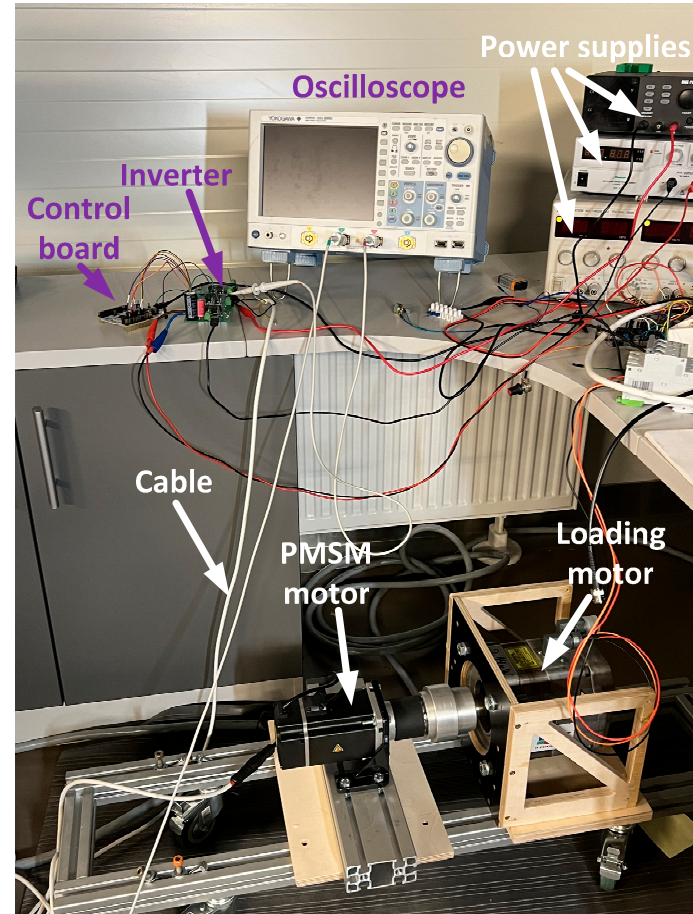


Figure 18. An experimental setup with the PMSM motor connected to the inverter via a long cable.

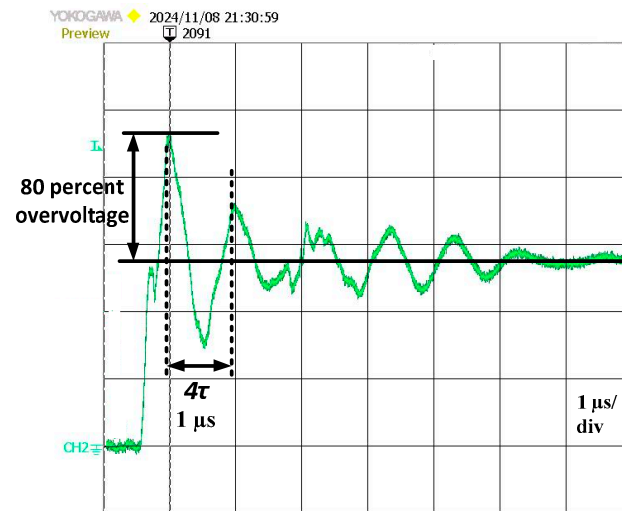


Figure 19. The voltage at the end of the cable measured by an oscilloscope.

For the reflection process and overvoltage analysis, different approaches can be used. The traditional theory of long lines with distributed parameters can be used for analysis, but for practical applications, these equations are too complex. A simplified expression for overvoltage calculation can be found in [33] and in other sources:

$$U_{max} = \frac{3l \cdot V_{DC} \Gamma_{mot}}{v \cdot t_r} + U_{DC}, \quad (18)$$

where v —propagation speed of the wave in the cable; t_r —rise time of the voltage of the transistor.

As can be seen from (18), overvoltage depends on several parameters: the DC bus voltage U_{DC} which is fixed, and the propagation speed v that depends on the cable parameters and is constant as well. Mainly, two approaches can be used to reduce overvoltage—influencing the reflection coefficient or changing the rise time of the voltage.

5. Overvoltage Reduction at the Terminals of the Motor

There are several ways to deal with overvoltage at the motor terminals. One solution could be to increase the dielectric strength of the insulation of the motor, but this is expensive and does not solve the problem with common-mode voltage that creates the current in the bearings of the motor. Therefore, solutions that reduce overvoltage should be implemented.

5.1. Reduction in Switching Speed of the Transistors

The simplest way to reduce dv/dt is to slow down the switching process of the transistor by influencing the gate voltage of the transistor. As one of the best solution is artificial increasing of C_{GD} (gate to drain capacitance) [34]. In the LT Spice model, it is possible to add additional C_{GD} capacitance and change the gate resistor, obtaining the desired dv/dt . The result can be seen in Figure 20.

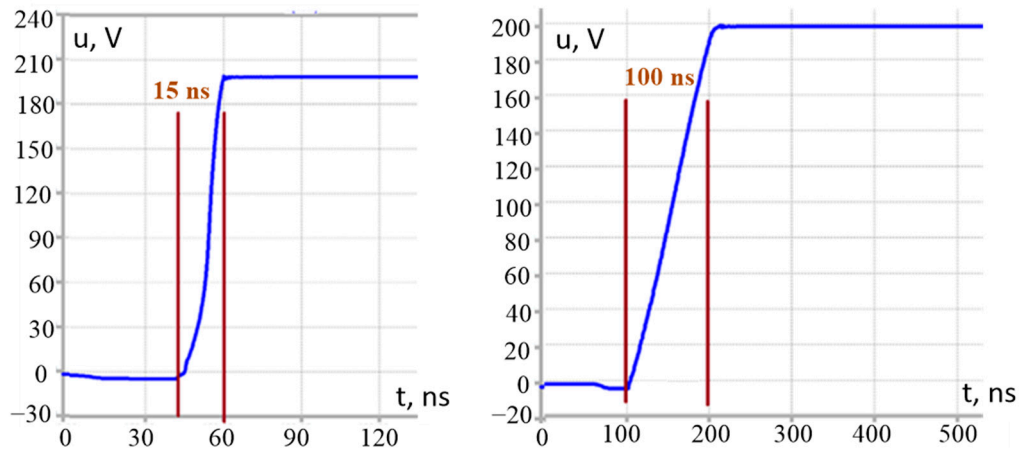


Figure 20. Voltage rise time without and with increased C_{GD} .

Figure 21 shows a transistor half-bridge with an additional 50 pF capacitor. As can be seen in Figure 22, this solution allows for a reduction in dv/dt . The results are close to the simulation model, in which case parasitic inductances of PCB are considered. This is a very simple and effective method to limit dv/dt . As a drawback, it can be mentioned that slower transient process is going to increase the switching losses, thus destroying overall efficiency. GaN transistors will require better thermal solutions to cool down these losses, and the efficiency will be decreased.

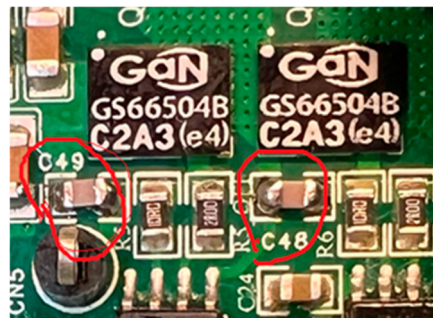


Figure 21. Experimental PCB of GaN half-bridge with additional C_{GD} , red lines mark additional components to influence switching speed.

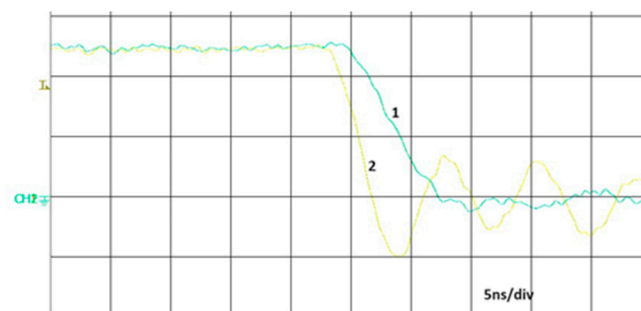


Figure 22. An experimental turning off process of the transistor without the additional C_{GD} (2) and with increased C_{GD} (1).

5.2. Application of Undamped Passive Filter

In [11], an undamped LC filter was mentioned as the most efficient solution for overvoltage reduction by comparing different solutions, similar to that proposed in [35]. For such a filter, there is a need to generate an additional pulse to charge the capacitor to avoid

oscillations. This makes the control of the inverter more complex as in addition to the sinusoidal PWM signal; this pulse should be added. This additional control signal allows us to achieve a voltage transition without any overshoot. Moreover, in contrast to the passive filter, the energy stored in the filter capacitor is not dissipated but can be recovered. As a drawback two additional switching actions that are going to increase switching losses can be mentioned.

The precharge signal length can be calculated as follows [35]:

$$t_{rise} = 2.094 \cdot \sqrt{LC}. \quad (19)$$

By using a microcontroller, there is a need to use several timers to implement such a pulse. A practically obtained waveform of the pulse width modulated (PWM) signal can be seen in Figure 23.

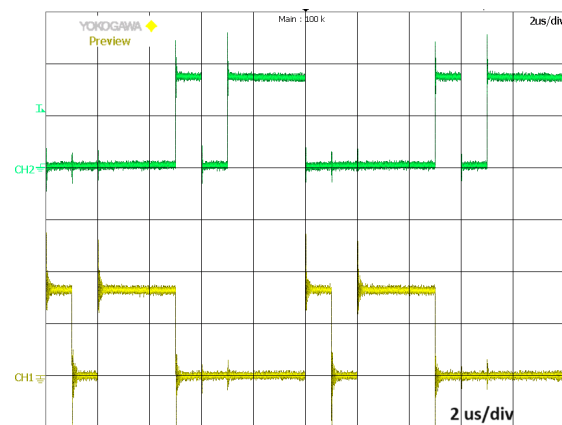


Figure 23. Pulse width modulated signal with filter precharge signals: green - high side transistor control signal; yellow – low side transistor control signal.

The filter was implemented into inverter design shown in Figure 24. Inductance was equal to 3.3 μ H and capacitance to 100 nF. The filter charge pulse calculated by (18) equal to 1 μ s was generated. Experimental waveforms can be seen in Figure 25. Although it is possible to generate such pulses with a microcontroller, it makes it more difficult to generate sinusoidal PWM; therefore, more advanced controllers are preferable. Moreover, there are additional switching actions of GaN transistors and therefore, switching losses are increased. Some oscillations can be seen, which can be explained with load influence, since it is mentioned in [35] that an additional pulse should be generated to compensate for the load influence. These are the reasons why such kinds of filters have not found wide application in industry today.

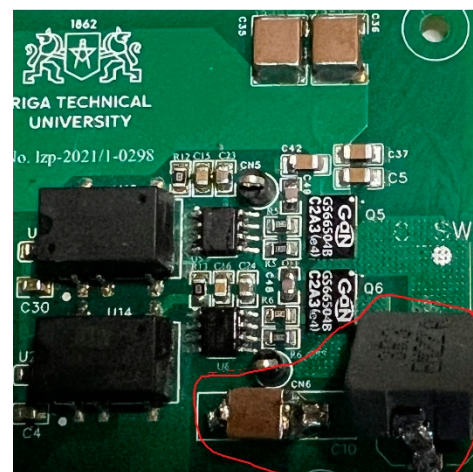


Figure 24. One half-bridge of the inverter with an implemented LC filter with active control: filter components wrapped with red line.

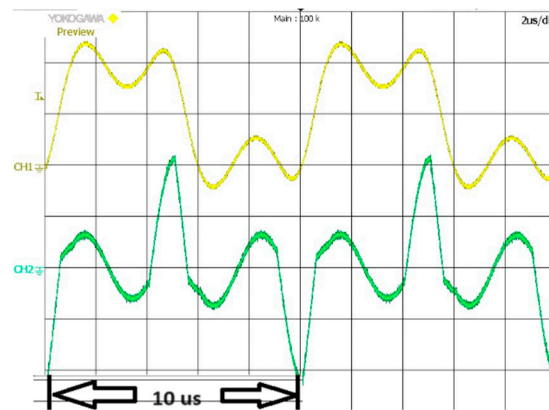


Figure 25. Experimental waveforms of an inverter with an LC filter with active control—output voltage and current: yellow – voltage; green - current.

5.3. Passive Filter with Damping

Still the most widely used approach to reduce dv/dt remains the use of passive filters with damping resistance. Most widely used filter types are RLC, RC and RL filters that can reduce overvoltage. In [31], an RL filter is mentioned as a good solution for overvoltage mitigation. In [21], it is stated that the best resistance value for overvoltage reduction is equal to Z_0 . To recalculate the filter component values to the three-phase system, the following equations should be used:

$$L_{fa} = \frac{2}{3} L_f, \quad (20)$$

$$R_{fa} = \frac{2}{3} R_f, \quad (21)$$

$$C_{fa} = \frac{3}{2} C_f. \quad (22)$$

The inductor value influences the operation of the RL filter significantly. High inductance provides higher impedance up to lower frequencies, and therefore current through the resistor starts to flow at lower frequencies; thus, overvoltage is eliminated. Low inductance is going to eliminate overvoltage, but not that effectively. The same is true for RC filter. In the case of the GaN transistor-based inverter, it is beneficial to increase the switching frequency to obtain smoother current waveform and eliminate audible noise. This is going to limit the maximum value of inductance, since the fundamental switching frequency should mainly flow through the inductor and not the resistor, because there will be high losses in the resistor otherwise, and this is unacceptable. The possibility of improving the performance of the damped passive filter will be further analyzed.

6. Improved Filter for Overvoltage Mitigation of High Switching Frequency Inverter

As was mentioned previously, the high inductance of the inductor in the case of the GaN inverter with high switching frequency cannot be used, and full overvoltage reduction with RL filter is challenging. In Figure 26, a case can be seen when a reduction RL filter with 25 μ H inductor and 67 Ω resistor has been used for overvoltage. As can be seen, overvoltage still exceeds 40 percent.

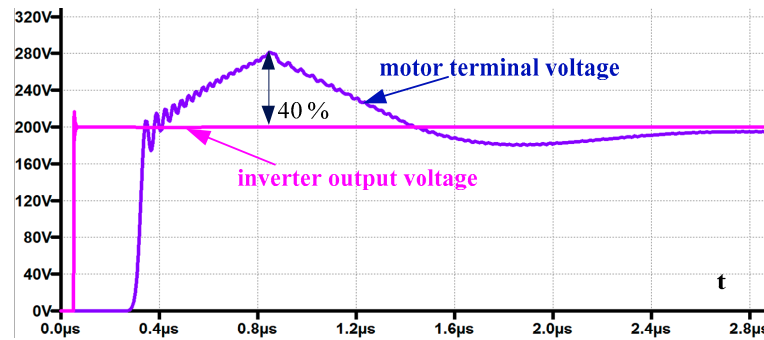


Figure 26. Simulation-based waveforms of inverter with RL filter.

Another option is to use the RC filter at the motor side. To calculate the values of RC filter, the methodology presented in [19] can be used. In Ref. [19], developing the filter at the motor side has been proposed, which creates negative reflection coefficient. The value of R_{f2} is selected equal to $Z_0/2$ and capacitor value is calculated by the following equation:

$$C_f = \frac{4}{3} \times \frac{l \cdot C_0}{\ln\left(\frac{1}{\Gamma_{mot}}\right)}, \quad (23)$$

where l is the length of the cable, C_0 is the equivalent capacitance of the cable per metre and Γ_{mot} is the reflection coefficient at the motor side, which can be calculated by (1) or determined from Figure 19 to be approximately equal to 0.8.

Figure 27 shows the curves for the case if R_{f2} is equal to $Z_0/2$ at different values of filter capacitor, obtained by analytically solving the circuit shown in Figure 15 with the added RC filter. As can be seen, the overvoltage is being reduced, but overvoltage reduction is similar to the case of the RL filter, since it is not possible to use the high capacitance for the filter to maintain the losses in the filter resistor in the decent range.

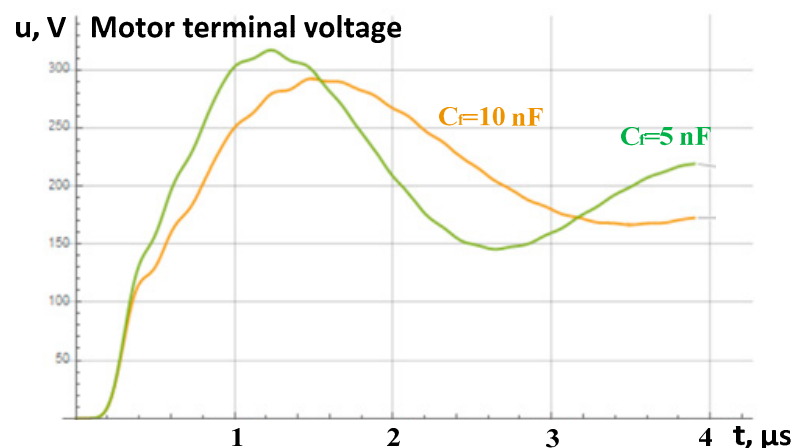


Figure 27. Overvoltage at different values of filter capacitance if R_{f2} is selected equal to $Z_0/2$ (33.5 Ω).

The second option is to use R_{f2} equal to Z_0 and select a capacitor value to provide significant impedance at the switching frequency and reduced impedance above this frequency. The results can be seen in Figure 28; the overvoltage reduction is slightly better than in the case when R_{f2} is equal to $Z_0/2$.

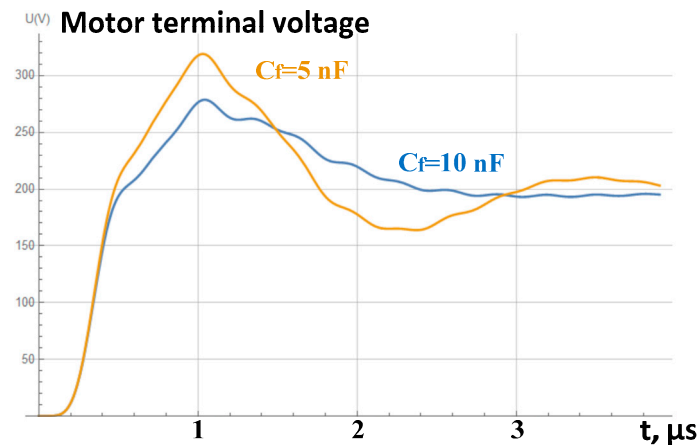


Figure 28. Overvoltage at different values of filter capacitance if R_2 is selected equal to Z_0 (67Ω).

Figure 29 shows experimental results with an added RC filter to the end of the cable. As can be seen, the overvoltage is reduced, but still, it is quite high—in the range of 40 percent. The result is quite like the simulation and the analytically obtained results. To reduce overvoltage further, the filter should be improved.

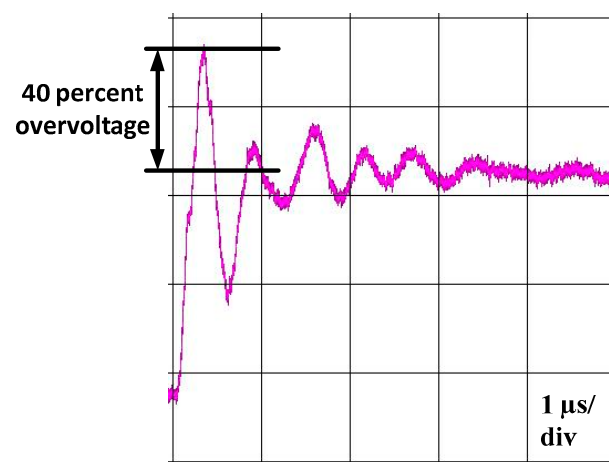


Figure 29. The experimentally obtained voltage at the end of the cable with an added RC filter.

To reduce overvoltage more, the RL filter can be combined with an RC filter at the motor side. The circuit can be seen in Figure 30. Since the desired switching frequency of the inverter has been selected to be equal to 50 kHz, the inductor is selected equal to 25 μH . Higher inductance will increase the impedance, and a significant part of the current is going to flow through the resistor and is going to increase the losses. To reduce the size of the RL filter inductor in this case, planar inductors have been developed. The manufacturing process of such inductors can be automated, and the size is being reduced due to the planar core. In Figure 31, a developed planar inductor can be seen; in the design of inductor ELP 43/10/28 N87, material core has been used. The inductor consists of three PCB-based windings on the top side and three windings on the bottom side of the PCB. Six PCBs are soldered together to create a series connection of 36 windings.

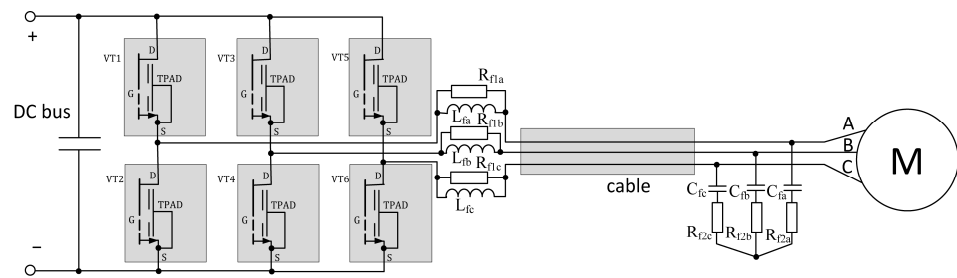


Figure 30. The schematics of the inverter powering the motor through the long cable with a filter at both ends of the cable.

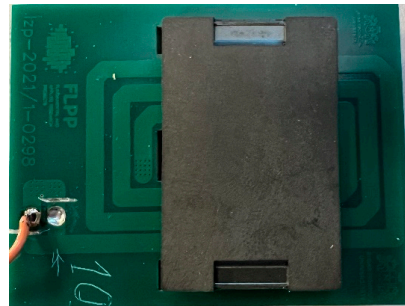


Figure 31. Developed planar inductor with PCB-based windings.

In Figure 32, a comparison can be seen between different values of filter resistance R_f , with and without RL filter. As can be seen in Figure 33, the best overvoltage reduction is achieved by combining the RL and RC filters and selecting R_f equal to Z_0 which, in this case, is 67Ω . These values were selected as the final solution for implementation.

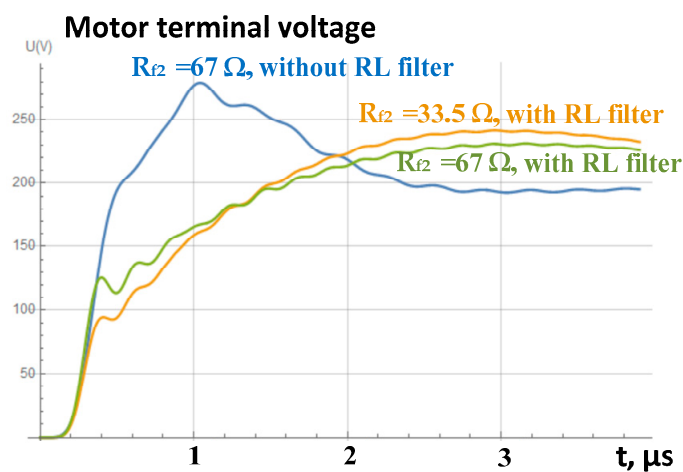


Figure 32. Overvoltage comparison at different values of filter resistance R_f , with and without RL filter: $C_f = 10 \text{ nF}$, $L_f = 25 \mu\text{H}$, $R_f = 67 \Omega$.

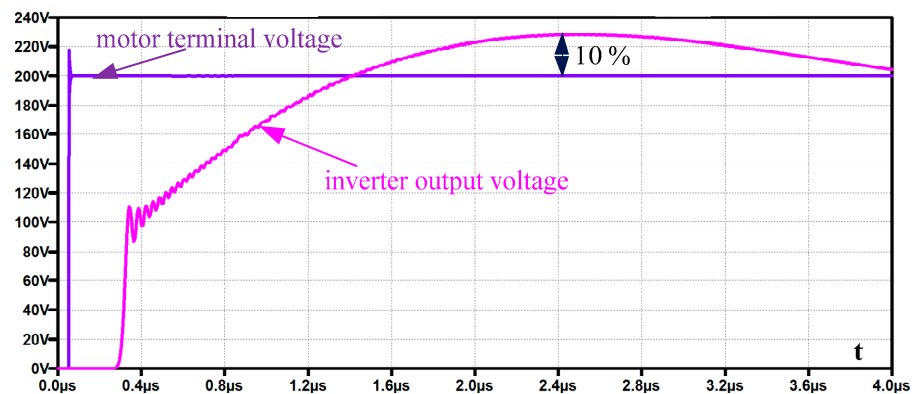


Figure 33. Simulation-based result, implementing the RL and RC combined filter: $C_f = 10 \text{ nF}$, $L_f = 25 \text{ } \mu\text{H}$, $R_{f1} = 67 \text{ } \Omega$, $R_{f2} = 67 \text{ } \Omega$.

Figure 34 shows experimental results with the improved filter. As can be seen, the overvoltage is small, thus eliminating the problem with high dv/dv caused by the GaN transistor-based inverter. As can be seen, the overvoltage has been reduced to just ten percent. The losses in the case of just the RL filter are 7 W in the case of 50 kHz switching frequency, but in the case of a combination of the RL filter at the input and the RC filter at the input, the losses are 10 W. In the case of higher switching frequency, the losses are going to increase proportionally.

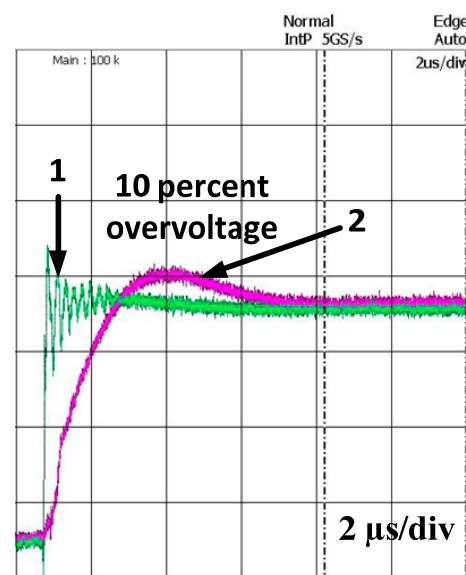


Figure 34. The experimental result implementing the RL and RC combined filter: $C_f = 10 \text{ nF}$, $L_f = 25 \text{ } \mu\text{H}$, $R_{f1} = 67 \text{ } \Omega$, $R_{f2} = 67 \text{ } \Omega$. The voltage at the inverter side and motor side at the end of the cable with the proposed filter: 1—voltage at the inverter side; 2—voltage at the motor side.

7. Conclusions

Wide bandgap transistors such as GaN transistors have short turn-on and turn-off transition times, thus leading to overvoltage in the cable connecting the inverter and motor. The overvoltage occurs due to reflections in both ends of the cable. The parameters of the high-frequency model of the motor and cable have been determined by measuring the impedance in a wide frequency range. The results shows that the model allows us to accurately determine overvoltage with and without the filtering solution.

Several overvoltage mitigation techniques were experimentally evaluated. The limitation of switching speed of the GaN transistor is an effective solution, but it creates

additional losses in the transistor. Since the GaN transistors are small, cooling is not easy, and this solution is preferable only in case the switching speed needs to be limited just slightly. Active filtering is a promising solution, but it is not that easily implementable, since a high-performance control system is required. Traditional passively damped filters remain the most practical and effective solution for most applications. However, conventional RL filters alone are insufficient when dv/dt is very high and cable lengths are long, as they fail to adequately suppress the peak overvoltage.

Therefore, it has been proposed to combine the RL filter with the RC filter connected to the motor terminals. For the selection of filter component values, the equivalent circuit of the cable and motor model has been used, and the electrical circuit of this model has been solved numerically to obtain accurate overvoltage values. Overvoltage is evaluated at different filter values by using numerical solutions and simulations. Results show the effectiveness of the proposed filtering solution. An experimental prototype of the GaN transistor-based three-phase inverter has been built and experiments with a long cable carried out. The experimental result verifies the accuracy of simulation results and results obtained by solving the model circuit numerically. By applying the proposed filter, the overvoltage has been reduced from the initial 80 percent value to 10 percent with losses below 10 W at the 50 kHz switching frequency.

With further development and wider application of wide bandgap semiconductors into motor drives, the problem of overvoltage will have to be addressed more often. Limiting the dv/dt of the transistor is the easiest way but at the same time it reduces benefits from GaN transistor application. The active filter seems to be an effective way to reduce overvoltage, but it requires complex control system and reduces reliability. Therefore, passive filters are still going to play an important role in overvoltage reduction. For proper filter component selection and overvoltage estimation, the high-frequency model of the motor and the cable is important. Complex impedance models do not allow us to obtain equations analytically or even simulate or properly solve circuits numerically. This article has shown that even simplified models can be useful to develop the filter. Further artificial intelligence-based methods to simplify complex high-frequency models and to automate the component selection of the passive filter and topology can lead to more optimal results.

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