

Article

# Modeling of a Quasi-Resonant DC Link Inverter Dedicated to Common-Mode Voltage and Ground Current Reduction

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**Abstract:** In this paper, the modeling methodology of the AC drive system with a Parallel Quasi-Resonant DC Link Inverter (PQRDCLI) is described. A presented modeling approach is an attractive tool used for the effective evaluation of a common-mode (CM) voltage and grounds current reduction methods. Designed models of inverter, induction machine (IM), and cable are simple, thus the methods for parameter extraction are not complicated. Verification of the proposed modeling approach was realized with the use of the the Synopsys (Mountain View, CA, USA) SABER simulator, while simulation results were experimentally verified. Operation principles of the proposed PQRDCLI converter topology are also described. Based on simulation and experimental results, it was confirmed that the proposed PQRDCLI solution represents required performance within the reduction of common-mode voltage and ground current in electric drives. Moreover, comparisons from a simulation complexity point of view have been performed to the existing methods. The evaluation is being shown at the end of the paper. It is confirmed that the presented method is simple, fast, accurate, and robust as well.

**Keywords:** quasi-resonant DC link inverter; common-mode voltage; AC drive; circuit simulation; modeling; voltage gradient; ground leakage current

## 1. Introduction

The use of the modern power semiconductor devices enables the operation of variable-frequency drives with carrier frequencies up to 200 kHz [1]. However, an increase of switching frequency results in EMI (Electromagnetic Interference Emissions) problems appearance; thus, the level of generated conducted EMI disturbances is one of the main evaluation criteria of AC drive inverters. High frequency EMI disturbances are propagated by magnetically and capacitively coupled parasitic circuits [2,3]. It should be noted that common-mode (CM) voltage at motor terminals is partially transferred through capacitive coupling to a non-grounded motor shaft, which results in shaft voltage appearance [4,5]. As a consequence, the probability of occurrence of destructive electrostatic discharge machining (EDM) bearing currents increases according to the growth of CM voltage amplitude [6]. EDM currents are the result of a breakdown of insulating lubricating grease films in rotating bearings. This is caused by overshoot the maximum breakdown value of the machine shaft voltage. As a result of the EDM current's influence, pits, craters, or stripes appear on rolling surfaces of machine bearings, which leads to faster degradation of bearings. Finally, bearings are destroyed and electric drive becomes out of order. This problem grows accordingly to the dissemination of electric drives fed by inverters, what shall be reflected also within the drive system reliability. Some reduction methods of bearing

currents occurrence are proposed, for example the use of conductive greases, application of insulated bearings or motor shaft grounding. However, these methods do not fully eliminate the bearing's currents problem, because these solutions do not affect the CM voltage. Possible elimination of the discussed problem can be realized through the use of hybrid bearings with ceramic rolling elements. The disadvantage of this approach is that the market cost of this solution is very high.

Due to the propagation mechanism, the level of CM disturbances is strictly connected with  $dv/dt$  value resulting from the transistor's commutation processes. Hence, high  $dv/dt$  CM voltage slopes generate large peaks of leakage current circulating in a protective ground wire, what reflects in excitation of the circulating bearing currents [7,8]. It is worth mentioning that a large amplitude of leakage currents may cause undesirable operation of residual current circuit-breakers, the wrong activation of fire alarms, or various sensors operation disturbances.

Various methods for reduction of these negative effects are proposed, including installation of filters [5,9] applying modified modulation methods [10] or using modified DC/AC inverters [8,11]. It should be noted that in some cases, the effectiveness of these solutions is questionable. Hence further research is required. Hence, the problem of methods development focused on bearing's currents elimination, and CM voltage influence reduction must be evaluated as still valid.

Modeling and simulation may be a useful tool for the evaluation of CM disturbance reduction methods, which is especially helpful at the early design stage of drive. Considering CM disturbance propagation paths, the models of inverter, cable, and motor should be included within the overall model of AC drive. Proposed solutions, dedicated to EMI analyses, ensures high accuracy of simulation results in a range of frequency up to tens of MHz [12]. These models take into account many aspects, e.g., the impact of parasitic capacitances (including nonlinear capacitances of semiconductors), resistances and inductances of paths and cables, skin effects, etc. [13–15]. However, models become complex, which leads to an increase of the computational time of numerical calculations. As a result, the simulation process is time-consuming, and it often cannot be successfully finished due to numerical problems. This problem is especially distinguishable when more complicated topologies of inverters are simulated, for example, resonant or quasi-resonant inverters or multilevel inverters with an increased number of switches. It should also be noted that most of the proposed models dedicated to EMI analysis require complicated methods of parameter extraction, e.g., based on Wheeler/Schneider formulas [16], finite-element calculations [17], or PEEC (Partial Element Equivalent Circuit) methods [18]. Therefore, a trade-off between model complexity, availability of parameters, and accuracy is the main criterion of model usability.

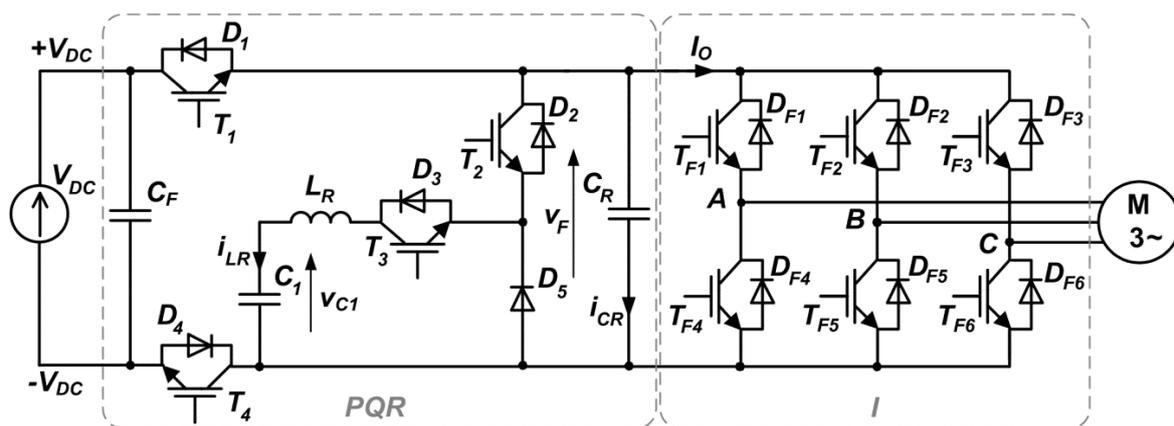
Effective CM voltage level reduction and limitation of ground currents peak values may be achieved by using Parallel Quasi-Resonant DC Link Inverter (PQRDCLI) [8,19]. Simulation enables evaluation of PQRDCLI properties at the beginning of the designing process when a target topology is developed. However, this task requires to use the models of semiconductor devices reflecting the dynamics of their switching process. Moreover, the impact of parasitic components such as resistances and inductances of paths and cables or capacitive couplings cannot be omitted.

In this paper, a modeling approach of AC drive fed by PQRDCLI is presented. Compared to other existing solutions, the models of the inverter, cable, and induction motor are not complicated, and their parameters are easy to extract. Presented approach enables realizing the simulation of the overall AC drive system with satisfactory accuracy. Simulated results were compared with measured ones to confirm the accuracy of the proposed models. Models presented in this paper were developed and validated using the SABER simulator; however, the proposed solution may be adapted without significant modifications to other circuit simulators, e.g., PSpice, LTSpice, etc. The presented modeling approach may be successfully used for the evaluation of reduction of CM voltage and  $i_{pE}$  current.

## 2. Model of AC drive with PQRDCLI

### 2.1. Model of PQRDCLI in Saber Simulator

If Parallel Quasi-Resonant DC-Link Inverters (PQRDCLI) are considered, the oscillations in a quasi-resonant circuit are excited once per commutation of inverter switches. Hence, voltage pulse-time modulation methods, including Space Vector Pulse Width Modulation (SVPWM), may be effectively adapted to PQRDCLI. A scheme of considered PQRDCLI is presented in Figure 1 [8]. During the first part of the resonant cycle, inverter input voltage  $v_F$  is reduced to zero, enabling switching of inverter transistors  $T_{F1}$ – $T_{F6}$  under Zero Voltage Conditions (ZVS). After the transition of the inverter vector to the new state, voltage  $v_F$  is rebuilt to the supply voltage  $V_{DC}$ . By controlling inverter input voltage gradients  $dv_F/dt$  during the resonant process, inverter output voltages gradients  $dv/dt$  are limited, what enables the reduction of overvoltage spikes. Due to the use of two transistors  $T_1$ ,  $T_2$  included in the DC link circuit, a full separation of induction motor from supply source  $V_{DC}$  may be realized. It results in reduction of CM voltage levels during inverter zero voltage state. Thus, compared to hard-switched inverter (considered to be conventional three-phase, two-level bridge inverter) probability of EDM currents occurrence of motor bearings fed by proposed PQRDCLI is significantly reduced. Moreover, the reduction of CM voltage  $dv/dt$  gradients leads to attenuation of ground leakage current pulses and mitigation of circulating bearing currents.



**Figure 1.** Parallel quasi-resonant DC link inverter: PQR—parallel quasi-resonant circuit, I—inverter, M—induction motor.

Operational waveforms of PQRDCLI are presented in Figure 2. Voltage  $v_{C1}$  (across the capacitor  $C_1$ ) is assumed to be constant during all operational cycles. For time intervals  $t < t_0$  or  $t < t_7$  the inverter operates at steady state. Transistors  $T_1$  and  $T_2$  are turned on, while the resonant circuit is inactive. Load current  $I_O$  flows through transistors  $T_1$ ,  $T_2$ , and voltage  $v_F$  is equal to  $V_{DC}$ . The resonant process is initiated by switching on transistor  $T_2$  under Zero Current Conditions (ZCS) at the moment  $t_0$ . Resonant Inductor current  $i_{LR}$  starts to increase linearly within the circuit  $V_{DC}$ – $T_1$ – $T_2$ – $D_3$ – $L_R$ – $C_1$ – $T_4$ . This period ends at the moment  $t_1$  when  $i_{LR}$  current reaches  $I_{LR(max)}$  value, and sufficient energy is accumulated in a resonant inductor  $L_R$  to ensure discharging of resonant capacitor  $C_R$ . It should be noted that transistor  $T_3$  should be switched-off during period  $<t_0, t_1>$  if  $v_F$  voltage zero states are applied to form a zero voltage vector of the inverter. If the inverter is switched between two active states, transistor  $T_3$  remains turned on during the full resonant cycle.

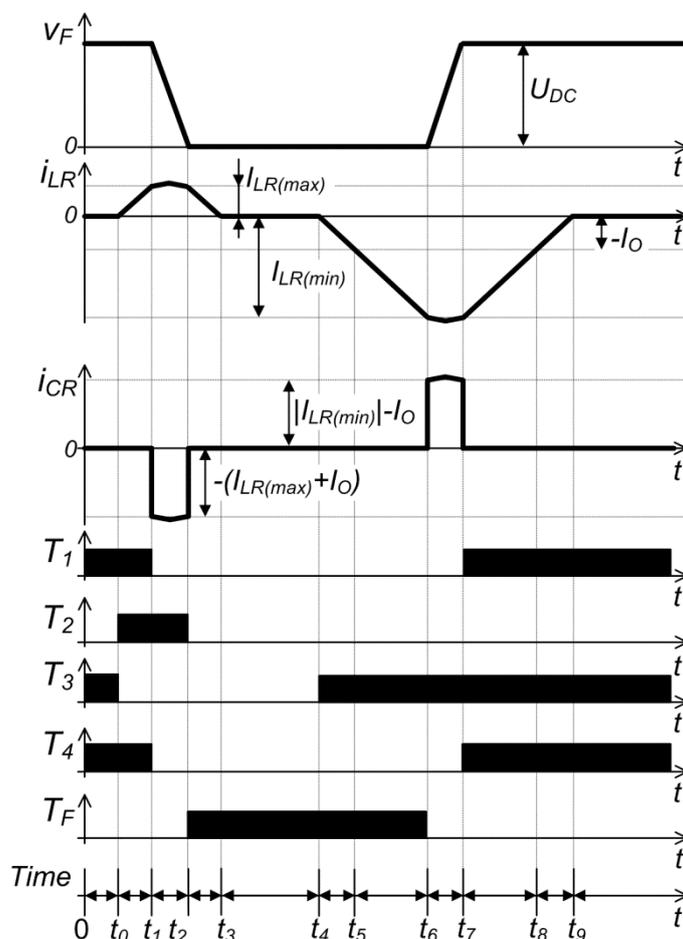


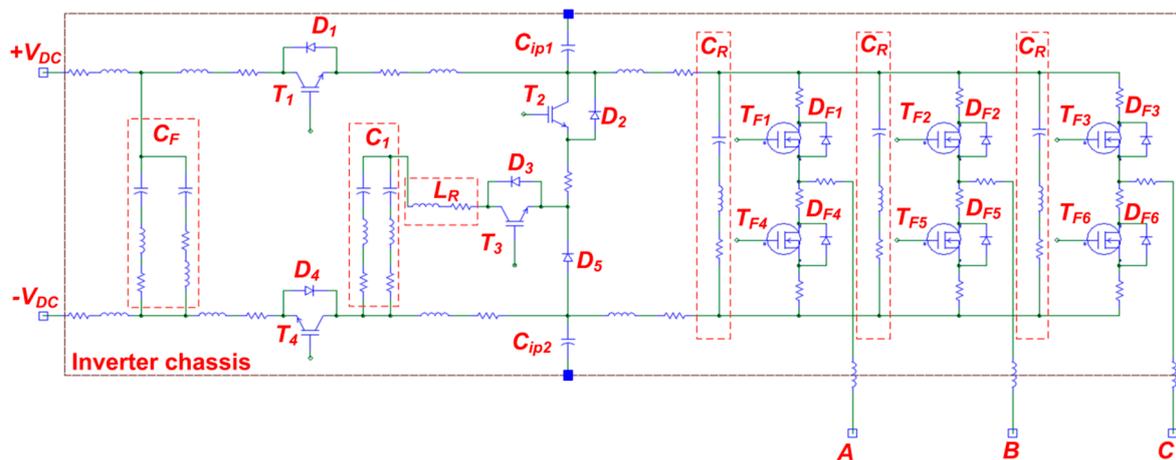
Figure 2. Transient waveforms at positive load current  $I_O$ .

At the moment  $t_1$ , transistors  $T_1, T_4$  are turned off at ZVS conditions. As a result, a resonant inductor current  $i_{LR}$  flows through the circuit  $L_R-C_1-C_R-T_2-D_3$ , which forces resonant discharge of resonant capacitor  $C_R$  by the current, which is a sum of  $i_{LR}$  and  $I_O$  currents. During period  $\langle t_1, t_2 \rangle$  voltage  $v_F$  is reduced to zero, which ensures the turn-on of inverter transistors  $T_F$  under ZVS conditions since moment  $t_2$  (signal  $T_F$  represents all of the inverter transistors at on- state). Transistor  $T_2$  is turned off under ZVS conditions at the moment  $t_2$ , thus resonant inductor current starts to decrease within the circuit  $L_R-C_1-D_5-D_3$ . Meanwhile, the load current  $I_O$  flows through the short-circuited inverter. At the moment,  $t_3$  current  $i_{LR}$  falls to zero. Period  $\langle t_3, t_4 \rangle$ , when  $v_F$  voltage is reduced to zero, it shall be maintained for an arbitrary time interval, which allows forming a zero voltage vector of the inverter. If the inverter is switched between two active states, this period is omitted. At the moment,  $t_3$  transistor  $T_3$  is turned on at ZCS conditions. As a result, the resonant inductor current starts to decrease in the circuit  $L_R-T_3-D_2-T_F$ . When current  $i_{LR}$  falls to  $I_{LR(min)}$  value, inverter transistors are switched to the new active state. Current  $i_{LR}$  starts to circulate in the circuit  $L_R-T_3-D_2-T_F$ , which forces the charging of capacitor  $C_R$  and rebuilding of voltage  $v_F$ . It should be noted that  $i_{CR}$  current should be positive, hence  $|I_{LR(min)}| > I_O$ . After rebuilding of  $v_F$  voltage to  $V_{DC}$  value, during period  $\langle t_6, t_7 \rangle$  the excess of energy accumulated in the resonant inductor is returned to the supply source  $V_{DC}$  by the current circulating within the circuit  $L_R-T_3-D_2-T_1/D_1-V_{DC}-T_4/D_4-C_1$ . At the moment  $t_7$ , the resonant inductor current reaches zero, and the resonant cycle ends.

The modulator signal initiates every operation cycle. If signal initialization is considered, the commutation of transistors is delayed due to the start of the resonant cycle. Moment  $t_2$  (Figure 2) is detected by respective comparators at the instant when voltage  $v_F$  falls below reference value  $v_{F(ref)} = 5 \text{ V}$ . Switching of transistors  $T_2$  and  $T_F$  is then possible. The second comparator detects

the moment when voltage  $v_F$  is rebuilding into  $V_{DC}$  value at the time  $t_6$ . It enables the turn-on of transistors  $T_1, T_4$ . Adopted PQRDCLI control strategy with the controlled length of resonant periods ensures the stabilization of capacitor  $C_1$  voltage. It enables control of gradient  $|dv_F/dt|$  regardless of the amplitude and direction of the load current  $I_O$  [8]. However, an estimation of  $i_{LR}$  current is required to calculate the length of period  $\langle t_0, t_1 \rangle$ , when resonant circuit transistors  $T_1, T_2, T_4$  are turned on. Similarly, this applies for determination of the length of period  $\langle t_2, t_5 \rangle$ , when turning-on  $T_F$  transistor short-circuits the inverter.

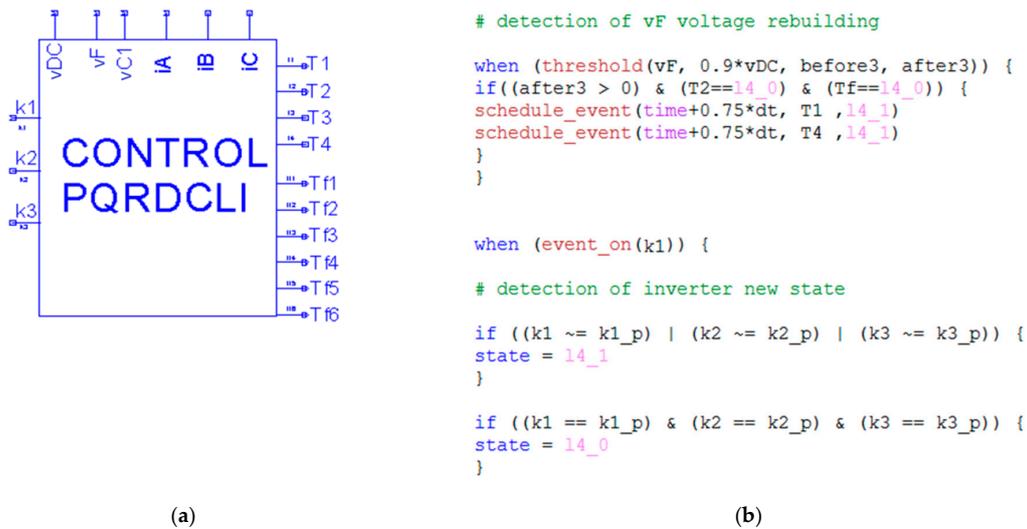
A simulation model of experimental PQRDCLI is presented in Figure 3. For the model of capacitors  $C_F, C_1$ , and  $C_R$ , the equivalent schematic with a series connection of capacitance, resistance ESR and inductance ESL is adopted [20]. The model of resonant inductor  $L_R$  consists of an inductance connected in series with resistance [21]. Simplified formulas were used to calculate resistances and inductances of main paths and buses in dependency on their geometrical dimensions [22]. This approach does not take into account skin effects, but it enables decreasing the PQRDCLI model complexity and simplifies the procedure of model parameters extraction. Capacitors  $C_{ip1}$  and  $C_{ip2}$  model a capacitive coupling between DC buses and inverter chassis/heatsink, and their values could be directly measured using an impedance analyzer ( $C_{p1} = C_{p2} = 260$  pF).



**Figure 3.** A simulation model of proposed PQRDCLI in SABER@Sketch editing window of SABER simulator.

Models of semiconductor devices should take into account the influence of parasitic nonlinear capacitances and dynamic behavior during switching processes. Within the simulation model of PQRDCLI, behavioral models of Insulated Gate Bipolar Transistors (IGBTs), and Metal-Oxide Semiconductor Field-Effect Transistors (MOSFETs) are applied based on the modeling approach presented in Ref. [23]. Moreover, the model of the diode, enables simulating a reverse recovery effect during the turn-off process [24]. The model of the control system is realized as a “user block” named “CONTROL PQRDCLI” and programmed using the mixed-technology language for electromechanical design and analysis MAST programming language (Figure 4). Initially, the values of load current  $I_O$  before and after the commutation of transistors are estimated in dependency on modulator signals [8]. It refers to the measurement of the actual value of output currents.





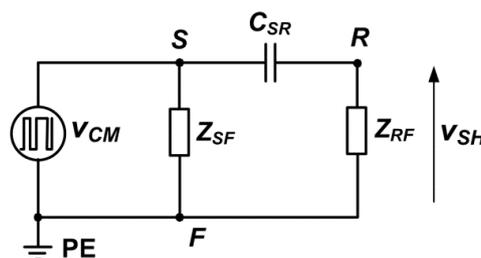
**Figure 4.** Model of the PQRDCLI control system: (a) SABER@Sketch symbol; (b) exemplary part of code written in MAST programming language.

Consequently,  $I_{LR(min)}$  and  $I_{LR(max)}$  values are calculated as a function of voltages  $V_{DC}$ ,  $v_{C1}$ , and function of the values of load current  $I_O$  and required value of gradient  $|dv_F/dt|$ . Finally, the length of switching sequences for each transistor is individually calculated. Required current and voltage sensors are adopted from Saber@Sketch parts library. Moreover, models of transistor drivers (defined as “user blocks” written in MAST) reflect gate voltage rise and fall times and gate current level.

2.2. Model of an Induction Motor Common-Mode Impedance

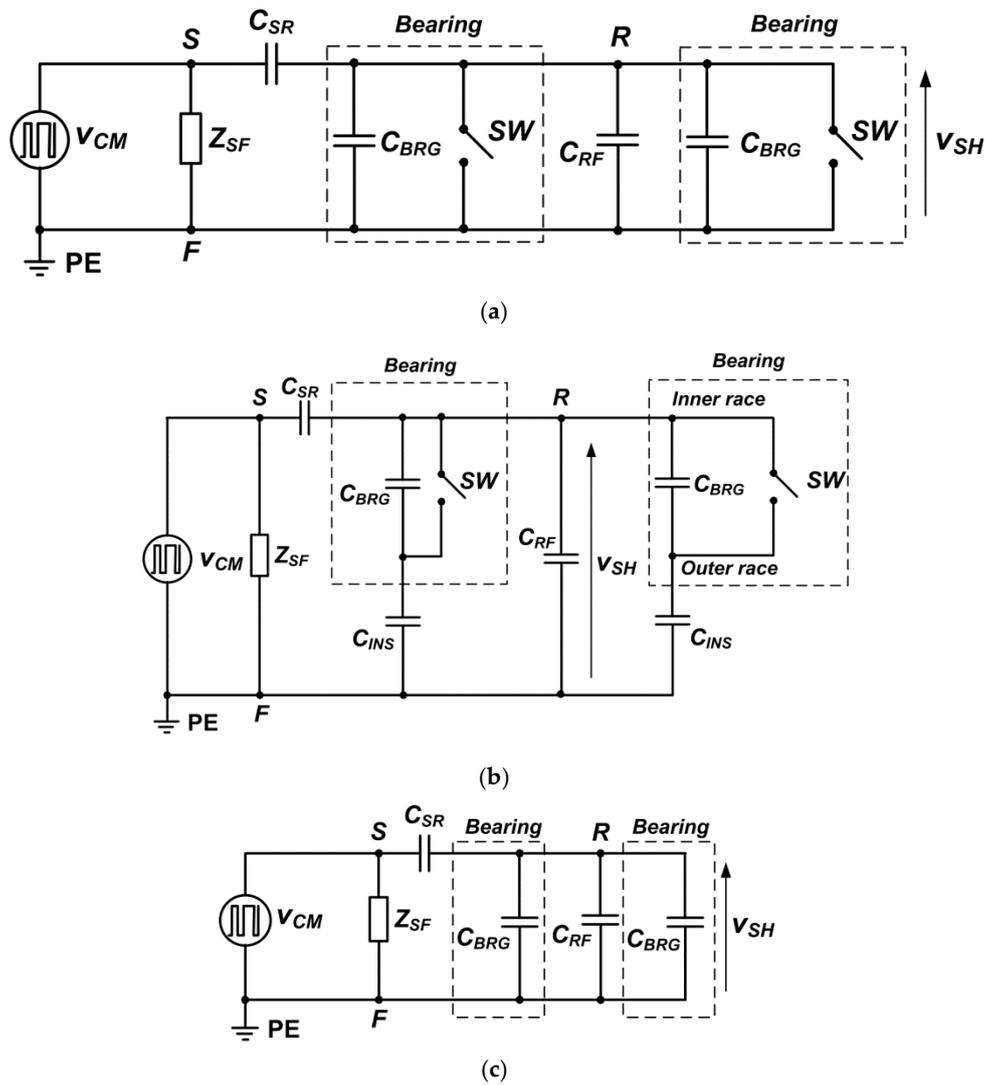
The appearance of CM voltage  $v_{N-PE}$  in the electric drive is caused by an inverter operation [25,26]. This voltage applies to a motor CM impedance, which generates a current flow in a protecting wire connecting the inverter and motor. It should be noted that  $v_{N-PE}$  acts on a stator winding and capacitive couplings between stator windings and motor frame or shaft and additionally between the shaft and grounded frame [25]. Hence, an equivalent scheme of a motor common-mode impedance may be considered (Figure 5), whose main components are:

- $v_{CM}$ —common-mode voltage source,
- $Z_{SF}$ —impedance between stator windings and grounded frame,
- $C_{SR}$ —capacitance between short-circuited stator windings terminals and motor shaft,
- $Z_{RF}$ —impedance between the motor shaft and frame,
- S—short-circuited input terminals of star-connected stator windings,
- F—motor frame
- R—motor shaft.
- $v_{SH}$ —shaft voltage.



**Figure 5.** A model of an induction machine common-mode impedance.

The model of impedance  $Z_{RF}$  between the motor shaft and frame is composed from inherent motor capacitance  $C_{RF}$  between shaft and frame and from bearings determined by the type of applied bearings. If standard bearings are used, a model of a single bearing is composed of capacitor  $C_{BRG}$  representing capacitance between the inner and outer race of bearing (Figure 6a). Additionally, a switch SW is used to model a breakdown of insulating lubricating grease films in a rotating bearing, which results in EDM current appearance [4,9]. If insulated bearings (standard bearings with additional insulation layers) are used, a model of single bearing presented in Figure 6a is modified by adding a capacitor  $C_{INS}$ . It represents the capacitance of an insulating layer between the motor frame and the outer race of bearing, as it is presented in Figure 6b [9]. If hybrid bearings with ceramic rolling elements are used, a model of a single bearing is only composed of a capacitor  $C_{BRG}$  without switch SW. (Figure 6c).



**Figure 6.** An equivalent scheme of an induction motor common-mode impedance: (a) with standard bearings; (b) with insulated bearings; (c) with hybrid bearings.

As is presented in Figure 7a, IM common-mode impedance is measured between input terminals of stator windings and a motor frame. Moreover, during the measurement, the input terminals of stator windings should be connected. Capacitive couplings distinctly dominate obtained frequency characteristic  $Z_{Cm}(\omega)$  with a small impact of inductive components in frequency bands “b” and “d” (Figure 7b) [15,27]. The typical value of capacitance  $C_{SR}$  is small - about 10 pF to 100 pF, hence current  $i_{PE}$  circulating through the ground wire is mainly determined by impedance  $Z_{SF}$  [28]. However,

bearings are affected by a shaft voltage  $v_{SH}$ , which arises from  $v_{N-PE}$  voltage value and parameters of capacitors forming impedance  $Z_{RF}$ . Considering previous relations, it is possible to propose a lumped parameter CM impedance model of IM with hybrid bearings (Figure 8).

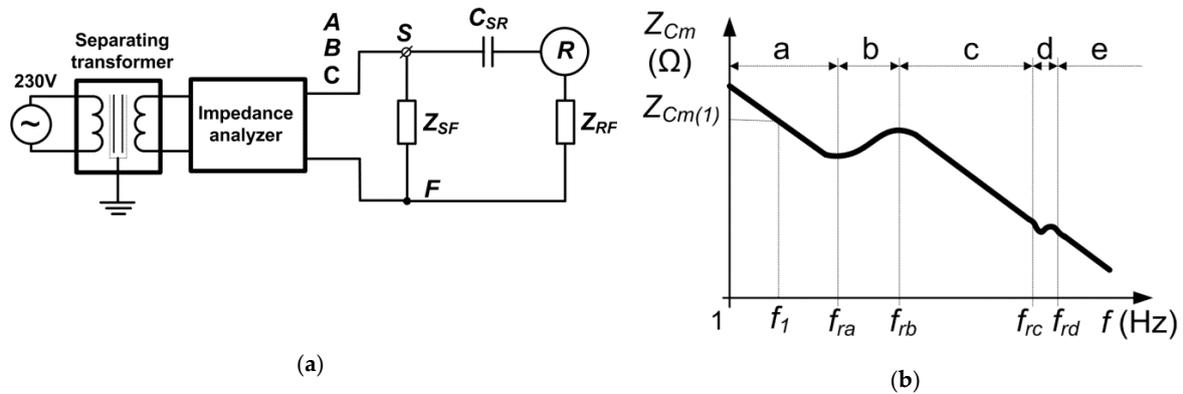


Figure 7. Measurement of IM common-mode impedance  $Z_{Cm}$ : (a) measurement setup; (b)  $Z_{Cm}(\omega)$  characteristic.

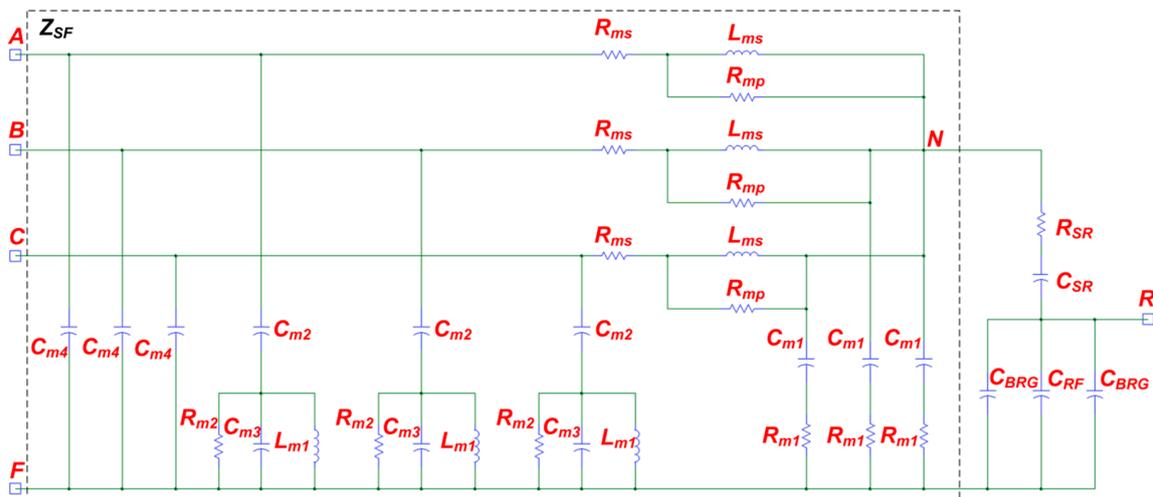


Figure 8. A model of IM common-mode impedance in the editing window of SABER simulator.

Impedance  $Z_{SF}$  is modeled by a set of resistances, capacitors, and inductors, whose parameters are extracted from the measured  $Z_{Cm}(\omega)$  characteristic. Omitting an influence of capacitance  $C_{SR}$ , in frequency band “e” (Figure 7b), impedance  $Z_{Cm}$  is determined by capacitor  $C_{m4}$  together with a series connection of capacitors  $C_{m2}$  and  $C_{m3}$ :

$$|Z_{Cm}(\omega)| = \frac{C_{m2} + C_{m3}}{3\omega(C_{m2}C_{m3} + C_{m2}C_{m4} + C_{m2}C_{m4})} \tag{1}$$

Moreover, analogously in section “c,” it can be assumed that:

$$|Z_{Cm}(\omega)| = \frac{1}{3\omega(C_{m2} + C_{m4})} \tag{2}$$

In the section “a,” when frequency  $f \ll f_{ra}$ , value of impedance  $Z_{Cm}$  results from a parallel connection of capacitors  $C_{m1}$ ,  $C_{m2}$ , and  $C_{m4}$ :

$$|Z_{Cm}(\omega)| = \frac{1}{3\omega(C_{m1} + C_{m2} + C_{m4})} \quad (3)$$

At frequency  $f_{ra}$ , a series resonance between inductor  $L_{ms}$  and capacitor  $C_{m1}$  occurs. Hence it can be assumed that the resonant frequency  $f_{ra}$  is given by:

$$f_{ra} \approx \frac{1}{2\pi \sqrt{L_{ms}C_{m1}}}. \quad (4)$$

On the boundary between bands “b” and “c,” a parallel resonance between inductor  $L_{ms}$  and capacitors  $C_{m2}$  and  $C_{m4}$  are recognized. Similarly, in-band “d” series and parallel resonances between inductors  $L_{m1}$  and capacitors  $C_{m2}$  and  $C_{m3}$  are observed. In bands “a” and “b,” absolute value of impedance  $Z_{Cm}$  may be described as follows:

$$|Z_{Cm}(\omega)| = \frac{1}{3} \sqrt{\left(\frac{Z_1(Z_1^2+Z_2^2)}{Z_1^2+(\omega(C_{m2}+C_{m4}))(Z_1^2+Z_2^2)-Z_2^2}\right)^2 + \left(\frac{(Z_2-\omega(C_{m4}+C_{m2}))(Z_1^2+Z_2^2)}{Z_1^2+(\omega(C_{m2}+C_{m4}))(Z_1^2+Z_2^2)-Z_2^2}\right)^2} \quad (5)$$

where

$$Z_1 = R_{m1} + R_{ms} + \frac{(\omega L_{ms})^2 R_{mp}}{(\omega L_{ms})^2 + R_{mp}^2} \quad (6)$$

and

$$Z_2 = \frac{\omega L_{ms} R_{mp}^2}{(\omega L_{ms})^2 + R_{mp}^2} - \frac{1}{\omega C_{m1}} \quad (7)$$

Similarly, for frequency  $f \gg f_{rB}$  in sections “c,” “d” and “e,” the following formulas are applicable:

$$|Z_{Cm}(\omega)| = \frac{1}{3} \sqrt{\left(\frac{Z_3(Z_3^2+Z_4^2)}{Z_3^2+(\omega C_{m4})(Z_3^2+Z_4^2)-Z_4^2}\right)^2 + \left(\frac{(Z_4-\omega C_{m4})(Z_3^2+Z_4^2)}{Z_3^2+(\omega C_{m4})(Z_3^2+Z_4^2)-Z_4^2}\right)^2} \quad (8)$$

where

$$Z_3 = \frac{(\omega L_{m1})^2 R_{m2}}{(\omega L_{m1})^2 + R_{m2}^2(\omega^2 L_{m1} C_{m3} - 1)^2} \quad (9)$$

and

$$Z_4 = \frac{(1 - \omega^2 L_{m1} C_{m3})\omega L_{m1} R_{m2}^2}{(\omega L_{m1})^2 + R_{m2}^2(\omega^2 L_{m1} C_{m3} - 1)^2} - \frac{1}{\omega C_{m2}} \quad (10)$$

Model parameters are extracted from the measured CM impedance characteristic (Figure 7b). Firstly, parameters of capacitors  $C_{m2}$ ,  $C_{m3}$ , and  $C_{m4}$ , inductor  $L_{m1}$ , and resistance  $R_{m2}$  may be obtained based on experimental measurement of  $Z_{Cm}(\omega)$  characteristic within bands “c,” “d” and “e.” Calculations are done using Equations (8)–(10) and curve fitting method, e.g., `fminsearch` function of MathWorks (Natick, MA, USA) MATLAB/GNU Octave programs [29]. Capacitor  $C_{m1}$  may be parameterized by applying modified Equation (3):

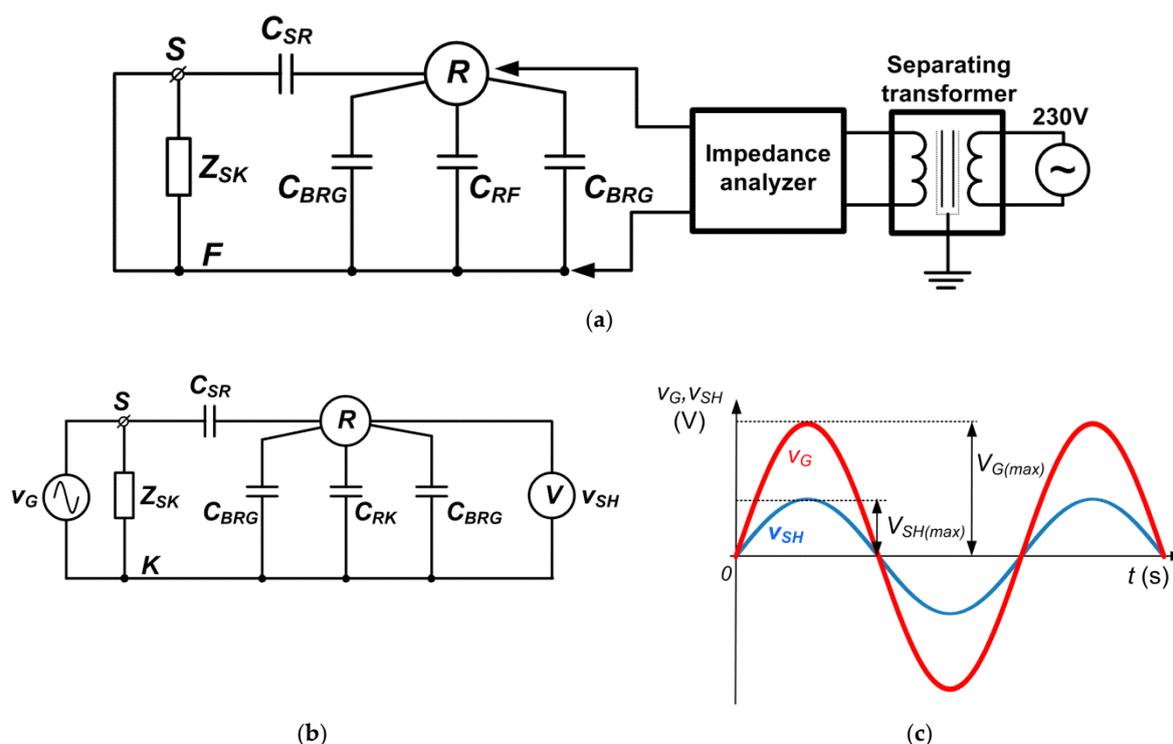
$$C_{m1} = \frac{1}{6\pi f_1 Z_{Cm(1)}} - C_{m2} - C_{m4} \quad (11)$$

where  $Z_{Cm(1)}$  is an absolute value of machine CM impedance at frequency  $f_1$  (Figure 7b). Next, inductor  $L_{ms}$  may be identified:

$$L_{ms} = \frac{1}{4\pi^2 f_{ra}^2 C_{m1}} \quad (12)$$

A set of equations derived from Equations (5)–(7) has to be solved in order to calculate values of resistances  $R_{m1}$  and  $R_{mp}$ . In that case,  $Z_{Cm}$  values measured for frequencies  $f_{ra}$  and  $f_{rb}$  should be taken into account as reference values. Values of winding resistances  $R_{ms}$  are directly measured using measuring devices (impedance analyzer).

Considering IM with hybrid bearings (Figure 6c), impedance  $Z_{RF}$  is determined by capacitances  $C_{BRG}$  and  $C_{RF}$ . Bearing capacitance  $C_{BRG}$  may be directly measured using an LCR meter or impedance analyzer before bearing installation. Then the values of  $C_{SR}$  and  $C_{RF}$  may be extracted. In the first step, an impedance  $Z_{Rm}$  is measured using a test bench presented in Figure 9a.



**Figure 9.** Measurement of  $C_{SR}$  and  $C_{RF}$ : (a) laboratory setup for identification of impedance  $Z_{Rm}$  characteristic; (b) measurement setup for Bearing Voltage Ratio (BVR) evaluation; (c)  $v_G$  and  $v_{SH}$  voltages waveforms.

It should be noted that input terminals of stator windings are short-circuited to the motor frame, hence  $Z_{Rm}$  impedance characteristic results from:

$$|Z_{Rm}(\omega)| = \frac{1}{\omega(2C_{BRG} + C_{RF} + C_{SR})} \quad (13)$$

Next, the Bearing Voltage Ratio (BVR) [6] defined as:

$$BVR = \frac{C_{SR}}{2C_{BRG} + C_{RF} + C_{SR}} \quad (14)$$

may be gained by introducing a sinusoidal voltage source  $v_G$  (e.g., signal generator) between terminals  $S$  and motor frame  $F$  and measuring of shaft voltage  $v_{SH}$  between the motor shaft and frame (Figure 9b). BVR can be calculated as follows:

$$BVR = \frac{V_{G(max)}}{V_{SH(max)}} \quad (15)$$

where  $V_{G(max)}$  and  $V_{SH(max)}$  are amplitudes of voltages  $v_G$  and  $v_{SH}$  (Figure 9c), hence, from Equation (14), the capacitance of  $C_{SR}$  is given by:

$$C_{SR} = \frac{2C_{BRG} + C_{RF}}{1 - BVR} \quad (16)$$

Moreover, substituting Equation (16) to Equation (13) capacitance of  $C_{RF}$  may be obtained:

$$C_{RF} = \frac{1 - BVR}{\omega |Z_{Rm}(\omega)|} - 2C_{BRG}. \quad (17)$$

In the presented model structure (Figure 8), a small resistor ( $R_{SR} = 1 \Omega$ ) was implemented between node  $N$  and capacitor  $C_{SR}$  in order to improve the numerical stability of the model. The proposed model of IM CM impedance may also be applied when another type of bearings are used. If insulated bearings are installed (Figure 6b), parameter identification of capacitors  $C_{SR}$ ,  $C_{BRG}$ ,  $C_{RF}$ ,  $C_{INS}$  is also possible as it is presented in Ref. [9]. In that case, a measurement procedure requires to ensure sufficient speed of motor shaft (more than 300 rpm) to form a thin insulating lubricating grease film within a bearing body [9]. If standard bearings are applied to motor construction (Figure 6a), individual identification of  $C_{BRG}$  and  $C_{RF}$  is impossible. However, the presented measurement method for capacitor  $C_{SR}$  and impedance  $Z_{Rm}$  is still valid if the motor shaft rotates with sufficient speed. A modeling approach of  $Z_{SF}$  impedance may be successfully applied regardless of applied bearing types.

In this paper, a model of 7.5 kW IM with hybrid bearings 6308-2RS (ZCS Ceramit) is being considered. Parameters of the motor model are depicted in Table 1.

**Table 1.** Common-mode impedance model parameters of a 7.5 kW motor with hybrid bearings 6308-2RS (ZCS Ceramit).

| Parameter | Value          |
|-----------|----------------|
| $C_{m1}$  | 1.31 nF        |
| $C_{m2}$  | 64 pF          |
| $C_{m3}$  | 102 pF         |
| $C_{m4}$  | 255 pF         |
| $C_{SR}$  | 105 pF         |
| $C_{RF}$  | 1310 pF        |
| $C_{BRG}$ | 29 pF          |
| $L_{m1}$  | 23.9 $\mu$ H   |
| $L_{ms}$  | 7.53 mH        |
| $R_{m1}$  | 308 $\Omega$   |
| $R_{m2}$  | 2.86k $\Omega$ |
| $R_{mp}$  | 5911 $\Omega$  |
| $R_{ms}$  | 5.31 $\Omega$  |
| $R_{SR}$  | 1 $\Omega$     |

### 2.3. Model of Cable

A model of four-wire cable connecting the inverter and motor is presented in Figure 10. The model is composed of resistances  $R_{cs}$  connected in series with inductors  $L_{cs}$ , mutual inductances  $M_{cs}$ , and ground capacitances  $C_{c1}$  connected in series with resistor  $R_{c1}$ . Terminals  $A, B, C$  are connected to the inverter outputs, and terminals  $A', B', C'$  is linked to the motor. Common-mode cable impedance characteristic  $Z_{Cc}(\omega)$  may be measured within the laboratory setup, as is presented in Figure 11a.

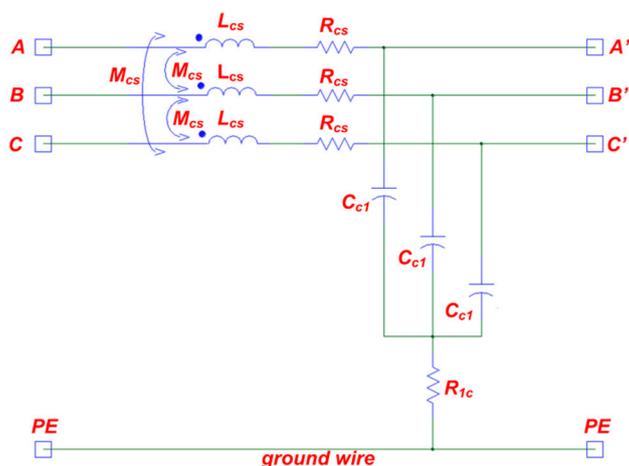


Figure 10. A model of cable.

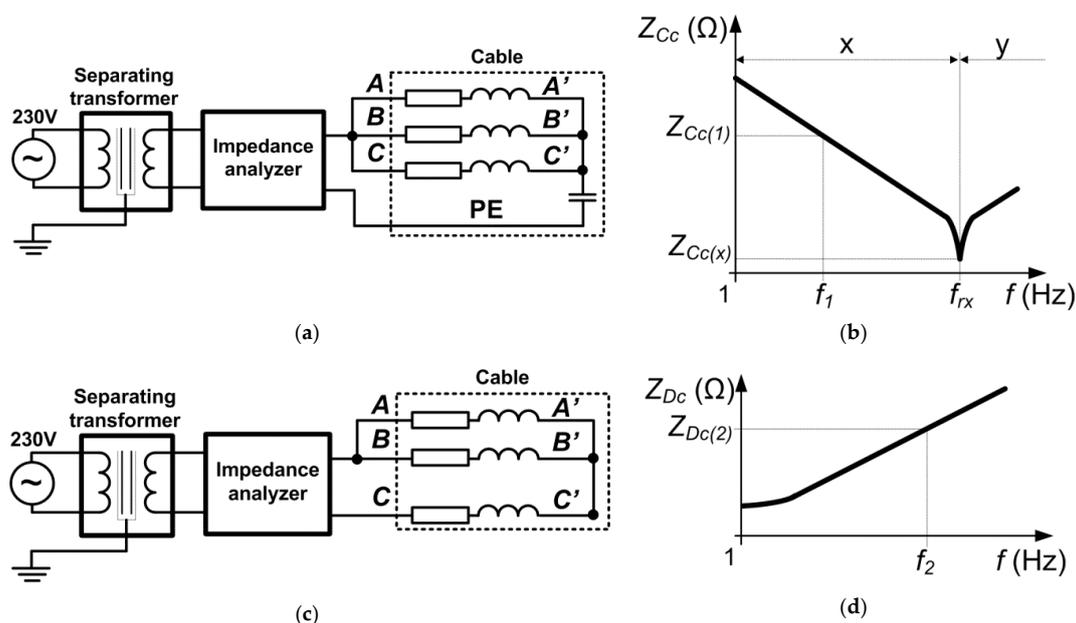


Figure 11. Measurement of cable model parameters: (a) measurement setup for  $Z_{Cc}(\omega)$  characteristic; (b)  $Z_{Cc}(\omega)$  characteristic; (c) measurement setup for  $Z_{Dc}(\omega)$  characteristic; (d)  $Z_{Dc}(\omega)$  characteristic.

Based on  $Z_{Cc}(\omega)$  characteristic (Figure 11b), one resonant frequency ( $f_{rx}$ ) is identified at series resonance, which is formed by cable series inductances and  $C_{c1}$ . If frequency  $f$  is significantly lower than  $f_{rx}$  (band “x”), the characteristic of cable CM impedance is dominated by a ground capacitance  $C_{c1}$ . Hence, the value of  $C_{c1}$  may be extracted as follows:

$$C_{c1} = \frac{1}{6\pi f_1 Z_{Cc(1)}} \tag{18}$$

where  $Z_{Cc(1)}$  is an absolute value of cable CM impedance at frequency  $f_1$  (Figure 11b), next, an equivalent inductance  $L_{Ce}$  may be obtained:

$$L_{Ce} = \frac{1}{4\pi^2 f_{rx}^2 C_{c1}} \tag{19}$$

Based on  $Z_{Dc}(\omega)$  characteristic (Figure 11d), which is obtained by the measurement (Figure 11c), an equivalent differential-mode inductance  $L_{De}$  may be calculated:

$$L_{De} = \frac{Z_{Dc(2)}}{2\pi f_2} \quad (20)$$

where  $Z_{Dc(2)}$  is an absolute value of cable impedance  $Z_{Dc}$  at frequency  $f_2$  (Figure 11d), analyzing schematics presented in Figure 11a, Figure 11c, and considering the proposed cable model structure, the following system of equations may be formulated:

$$\begin{cases} L_{Ce} = L_{cs} + 2M_{cs} \\ L_{De} = (3L_{cs} + M_{cs})/2 \end{cases} \quad (21)$$

Hence, solving a system of Equations (21), values of  $L_{cs}$  and  $M_{cs}$  are described as follows:

$$L_{cs} = (4L_{De} - L_{Ce})/5 \quad (22)$$

and

$$M_{cs} = (3L_{Ce} - 2L_{De})/5 \quad (23)$$

The value of resistor  $R_{cs}$  shall be measured directly using an impedance analyzer or LCR meter. Based on  $Z_{Cc}(\omega)$  characteristic (Figure 11b), it may be assumed that:

$$R_{c1} = Z_{Cc(x)} - R_{cs}/3 \quad (24)$$

where  $Z_{Cc(x)}$  is a value of  $Z_{Cc}$  impedance at resonant frequency  $f_{rx}$ .

In this paper, a cable model of 2.5 m length is considered. The model parameters are presented in Table 2.

**Table 2.** Cable model parameters.

| Parameter | Value         |
|-----------|---------------|
| $C_{c1}$  | 77 pF         |
| $L_{cs}$  | 670 nH        |
| $M_{cs}$  | 310 nH        |
| $R_{cs}$  | 18 m $\Omega$ |
| $R_{c1}$  | 2.52 $\Omega$ |

### 3. Simulation and Experimental Results

Parameters of experimental PQRDCLI are depicted in Table 3. The Control system with the Vector Sigma-Delta modulator [30] is implemented using the STM32F407 microcontroller operating with the sampling frequency of 20 kHz. To simplify control algorithms and calculations, MOSFETs were used as main inverter transistors  $T_F$ . It should be noted that the turn-off process of MOSFET proceeds without tail current observed for IGBT. If IGBT is used, the IGBT turn-off tail current will influence the resonant capacitor current  $i_{CR}$ , which would result in significant changes of  $dv_F/dt$  values during the period  $\langle t_5, t_6 \rangle$ . However, IGBTs were applied in the resonant circuit due to better dynamic parameters of anti-parallel diodes and lower conduction losses compared to MOSFETs.

**Table 3.** PQRDCLI specification.

| Parameter                | Specification                                       |
|--------------------------|---|
| Rated output power       | 2 kW  |
| $V_{DC}$                 | 260 V   |
| $T_1-T_4, D-D_5$         | IGBT-IRG4PC40 (Infineon)                            |
| $T_F-T_{F6}, D_F-D_{F6}$ | N-MOSFET FDA50N50 (Fairchild)                       |
| $C_F, C_1$               | 470 $\mu$ F (electrolytic) + 220 nF (polypropylene) |

Within experimental measurements, the bench power supply unit (EA-PSI 9750-20 3U) provides a stabilized voltage of 260 VDC to fed PQRDCLI. The inverter load was formed by a 7.5 kW induction

motor with hybrid bearings 6308-2RS (ZCS Ceramit, Tłuczań, Poland). Voltage and current waveforms were recorded using the Tektronix (Beaverton, OR, USA) DPO4034 oscilloscope equipped with the high voltage differential probe P5205A (100 MHz) and the current probe TCP2020 (50 MHz). Consequently, the simulation model of proposed PQRDCLI, IM, and cable was built as hierarchical models of Synopsys (Mountain View, CA, USA) Saber@Sketch and were used to model the AC drive system in the configuration, as presented in Figure 12. The star-connected capacitors  $C_d$  ( $3 \times 0.68$  nF) were used to measure the common-mode voltage  $v_{N-PE}$  referred to the Protective Earthing (PE) ground potential.  $V_{DC}$  supply ground capacitances ( $C_{sp1} = C_{sp2} = 72$  nF) were measured using impedance analyzer. Sinusoidal voltage sources  $V_\omega$  were added to simulate an influence of IM rotational electromotive force.

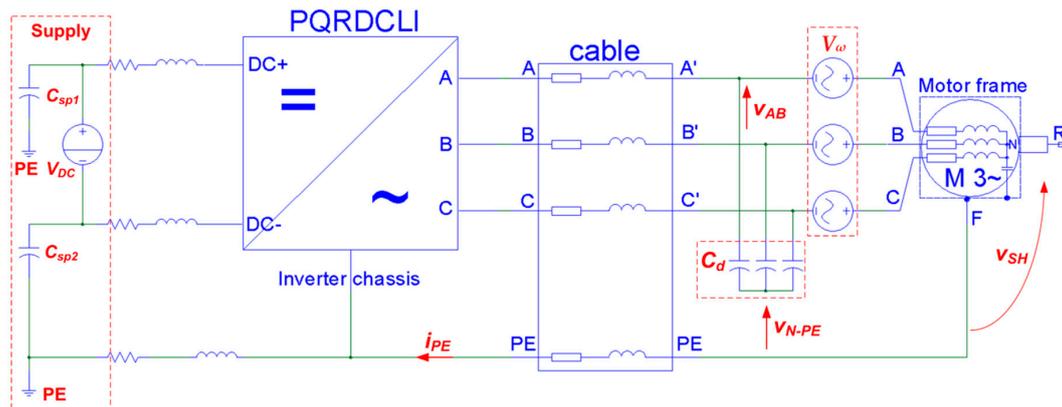


Figure 12. Simulation model of AC drive system with PQRDCLI.

A good coherence between simulated and measured characteristics of CM impedance of motor and cable is depicted in Figure 13. Identified results of measurements confirm resonant frequencies of simulated  $Z_{Cm}(\omega)$  and  $Z_{Cc}(\omega)$  characteristics. Moreover, values of  $Z_{Cm}$  and  $Z_{Cc}$  impedances at resonant frequencies are close to measured values. The proposed method of IM CM impedance modeling may be useful within the range of frequency up to 5 MHz. For higher frequencies, an impact of additional capacitive and inductive couplings, which is not considered within the proposed model, is distinguishable. These additional couplings may cause significant differences of impedance characteristics, even between motors of the same type; hence they should be identified for each machine individually [31].

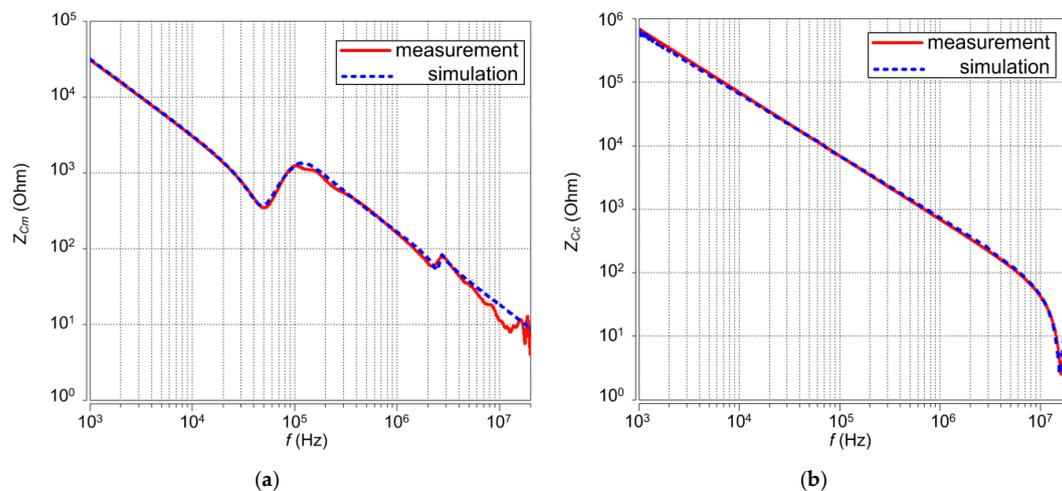
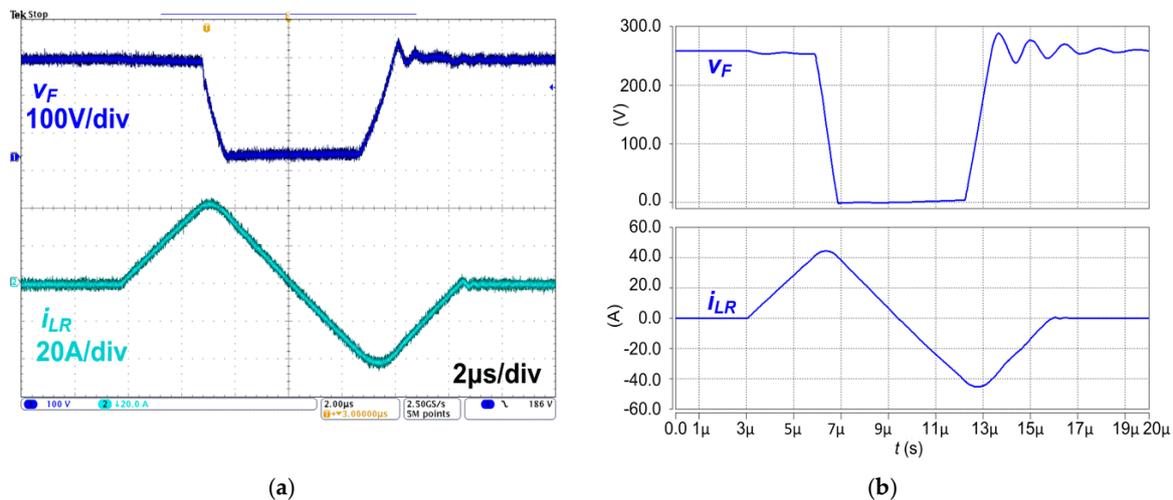


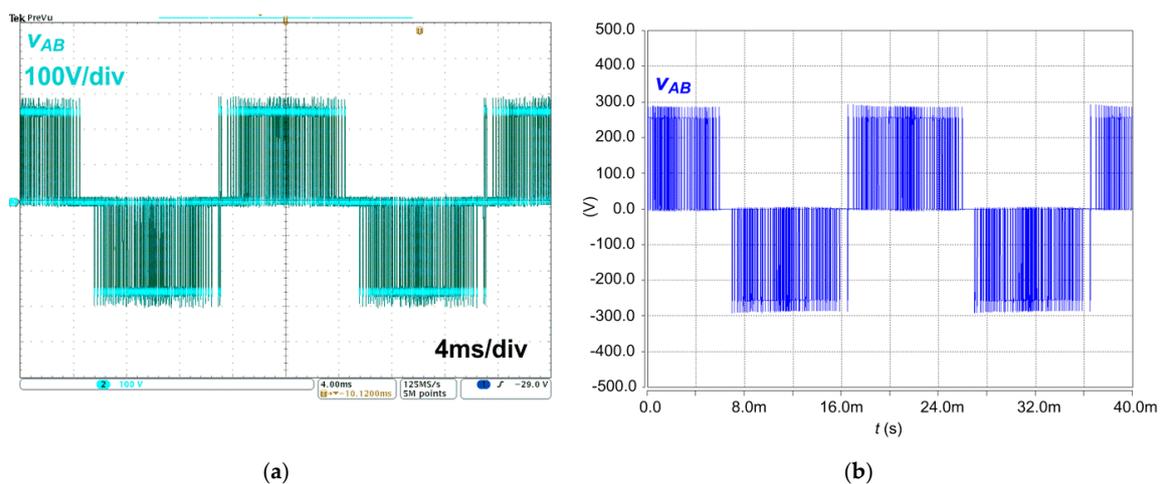
Figure 13. Measured and simulated frequency characteristics of the common-mode impedance: (a) induction machine; (b) cable.

Evaluation of the PQRDCLI model includes a comparison of characteristic operational waveforms with experimental ones. Good accuracy of waveforms between simulation and measurement of voltage  $v_F$  and current  $i_{LR}$  is shown in Figure 14. Special focus is given on comparisons of the maximal and minimal values of  $i_{LR}$  and on the comparisons of the gradient  $dv_F/dt$  during voltage  $v_F$  rise and fall periods ( $dv_F/dt \approx \pm 200$  V/ $\mu$ s). Attenuated voltage  $v_F$  oscillations, which appear after  $v_F$  rise/fall to  $V_{DC}$  value, are visible for both experimental and simulated waveforms.



**Figure 14.** PQRDCLI Quasi-resonant voltage  $v_F$  and current  $i_{LR}$  waveforms at related output power (2 kW): (a) measurement; (b) simulation.

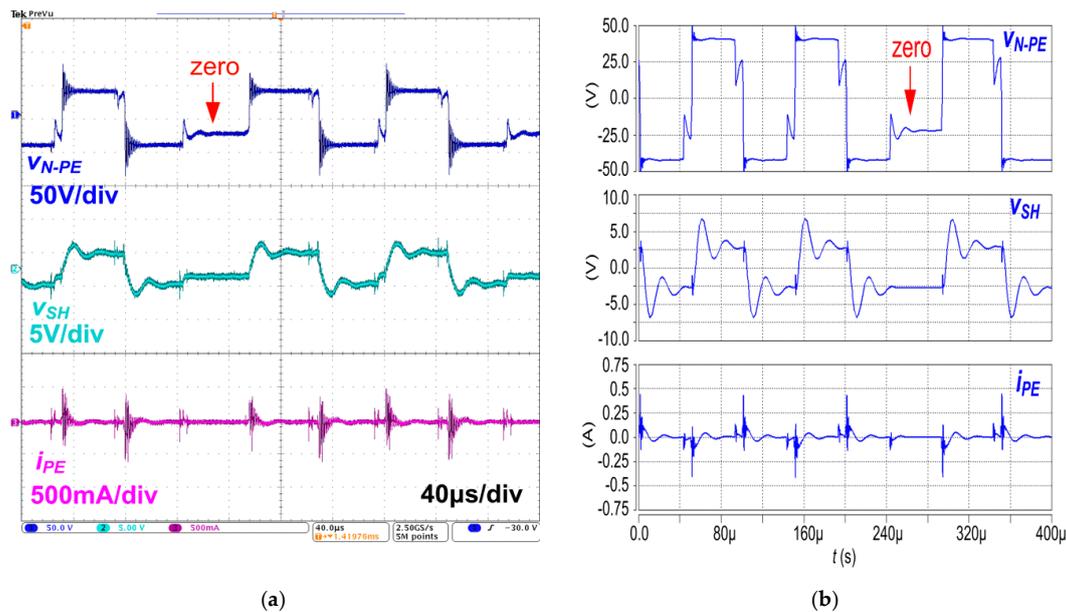
The reduction of gradient  $dv_F/dt$  enables the limitation of  $dv/dt$  values of inverter output voltages. As a result, the overvoltage spikes of the line- to - line voltages acting on motor stator windings do not exceed  $1.16 \times V_{DC}$  (without using any additional filters or snubbers). This statement is confirmed by simulation and measurement as well (Figure 15).



**Figure 15.** Line -to-line voltage  $v_{AB}$ : (a) measurement; (b) simulation.

A satisfied accuracy is recorded for simulated waveforms of common-mode voltage  $v_{N-PE}$ , shaft voltage  $v_{SH}$ , and ground current  $i_{pE}$  (Figure 16). In conventional two-level hard- switched inverter,  $v_{N-PE}$  levels are equal  $\pm V_{DC}/6$  for active vectors and  $\pm V_{DC}/2$  for zero vectors. If PQRDCLI is being used, levels of  $v_{N-PE}$  are limited to  $\pm V_{DC}/6$  due to CM voltage reduction for zero vectors [8]. However, as a result of the interaction between motor ground capacitances and parasitic nonlinear capacitances of transistors  $T_1, T_2$ ,  $v_{N-PE}$  is not fully reduced to zero during the inverter zero vector.

Hence, a bias shift (about 25 V) is observed for both measured and simulated  $v_{N-PE}$  waveforms. Shaft voltage  $v_{SH}$  reflects the envelope of  $v_{N-PE}$  voltage with  $BVR \approx 5\%$ , which is a typical value for off-the-shelf motors [32]. Due to the reduction of CM voltage, levels of shaft voltage  $v_{SH}$  are also limited, decreasing the possibility of EDM bearing's current occurrence. Compared to  $dv_{N-PE}/dt$ , shaft voltage gradients  $dv_{SH}/dt$  are slightly decreased, resulting from the influence of stator windings—motor frame impedance of IM. The waveform of ground leakage current  $i_{PE}$  is determined by  $dv_{N-PE}/dt$  values and CM impedance of propagation path between induction machine and inverter. Simulated  $i_{PE}$  current waveform with peak values close to 500 mA is coherent with measured one, proving correctness of adopted modeling approach.



**Figure 16.** Waveforms of CM voltage  $u_{N-PE}$ , shaft voltage  $u_{SH}$ , and leakage ground current  $i_{PE}$  (a) measurement; (b) simulation.

A satisfying agreement between experimental and simulated spectra of voltage  $v_{N-PE}$  and current  $i_{PE}$  is obtained (Figure 17). Simulated spectra levels and most significant resonant frequencies are recognized in experimental results, especially for frequency range lower than 1 MHz. Nevertheless,  $v_{N-PE}$  spectrum peak at  $f = 1\text{MHz}$  is not as clear as it is observed for measured results. For frequency range lower than 1 MHz, coherence between simulated and measured CM voltage spectra is satisfactory. Comparing simulated and measured spectra of current  $i_{PE}$ , the peak observed at a frequency around 1MHz is recognized for both cases. However, for the simulated spectrum, this peak is shifted about 0.3 MHz relative to the results of the measurement. A significant peak around 3 MHz is noted in the measured  $i_{PE}$  spectrum, but it is not observed for the results of the simulation. It should be noted that an influence of omitted components and phenomena (e.g., skin effects, more detailed modeling of capacitive ground couplings) should be taken into account to increase the accuracy of simulated spectra in a range of frequency higher than 1 MHz. Nevertheless, the complexity of models will increase significantly.

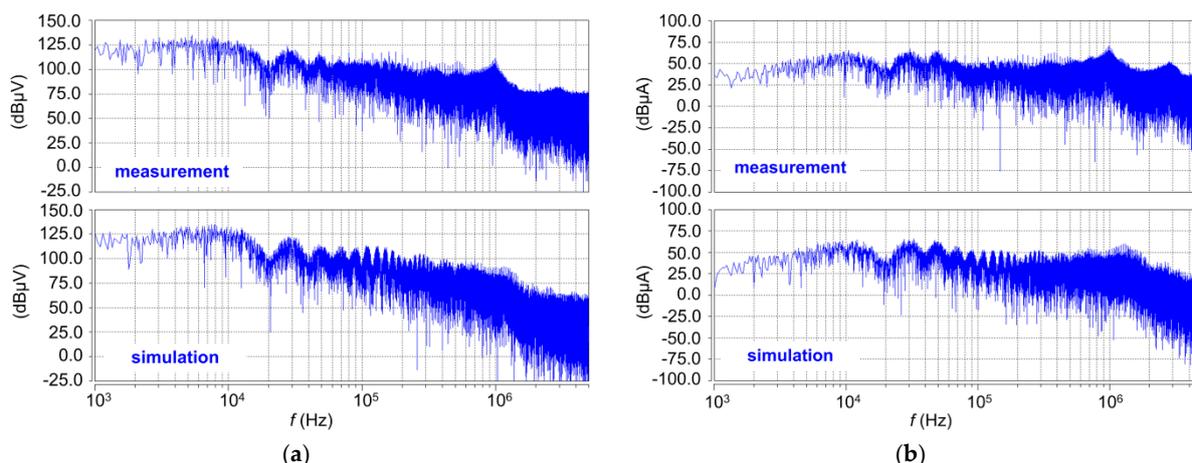


Figure 17. Simulated and measured spectra: (a) voltage  $v_{N-PE}$ ; (b) current  $i_{PE}$ .

The described modeling approach was used to model a system with a hard-switched inverter (considered to be general three-phase, two-level bridge inverter, whose main parameters are the same as reported for PQRDCLI). Such a drive system may be treated as a system without common-mode voltage reduction. The obtained results of simulation have been validated by comparison with the results of experimental measurement. In this case, a good correlation between simulated and measured results was recorded. Characteristic levels (at steady states) of CM voltage are distinguishable for active inverter vectors, when  $v_{N-PE} = \pm V_{DC}/6$  and for zero vectors, when  $v_{N-PE} = \pm V_{DC}/2$  (Figure 18).

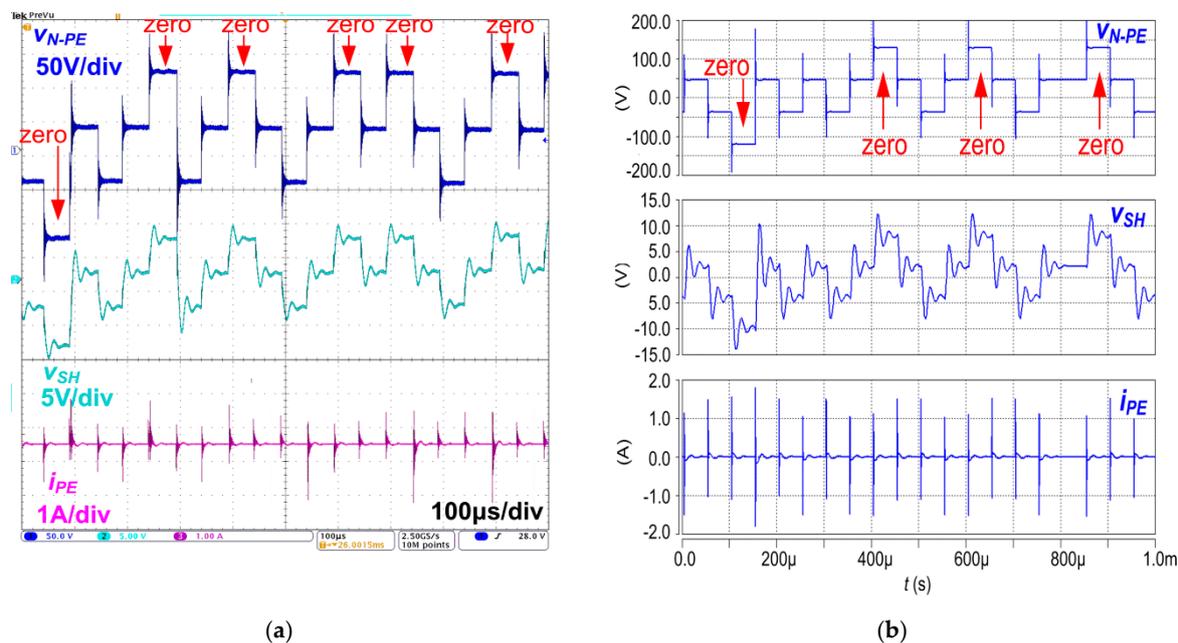
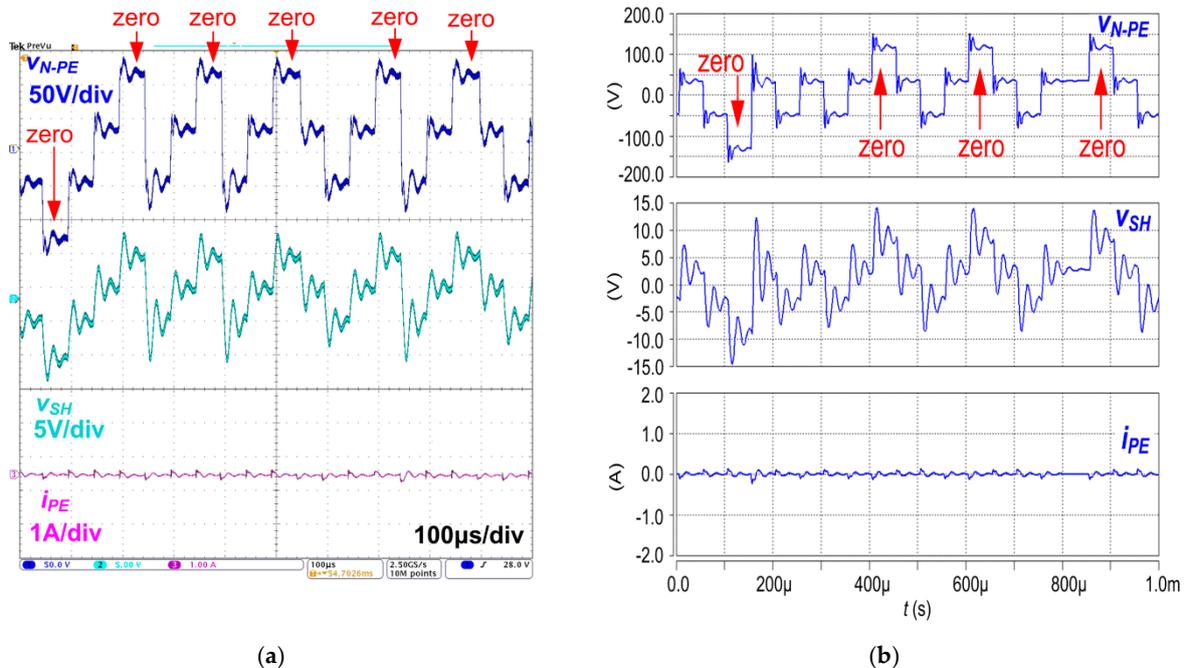


Figure 18. Waveforms of CM voltage  $u_{N-PE}$ , shaft voltage  $u_{SH}$ , and leakage ground current  $i_{PE}$  in a drive system fed by a hard-switched inverter: (a) measurement; (b) simulation.

It should be noted that the maximum values of voltage  $v_{N-PE}$  are significantly higher than recognized for PQRDCLI. It means that maximum levels of shaft voltages in a system fed by a hard-switched inverter are also higher. Hence the probability of EDM bearings currents occurrence is higher. It should also be noted that  $dv_{N-PE}/dt$  gradients are also higher than those for PQRDCLI. Thus, amplitudes of leakage currents current  $i_{PE}$  are also increased.

Additionally, a simulation of a drive system fed by a hard-switched inverter with a 720  $\mu\text{H}$  common-mode choke included between machine and inverter was performed. Despite the significant reduction of  $i_{PE}$  current amplitude, levels of voltages  $v_S$  and  $v_{N-PE}$  at steady states were not limited (Figure 19). However, at transient states, a small reduction of overvoltage spikes in  $v_{N-PE}$  waveform was noticed. Simulated results are confirmed by the results of measurements, which proves the suitability of the presented modeling approach in the evaluation of CM disturbances reduction methods.

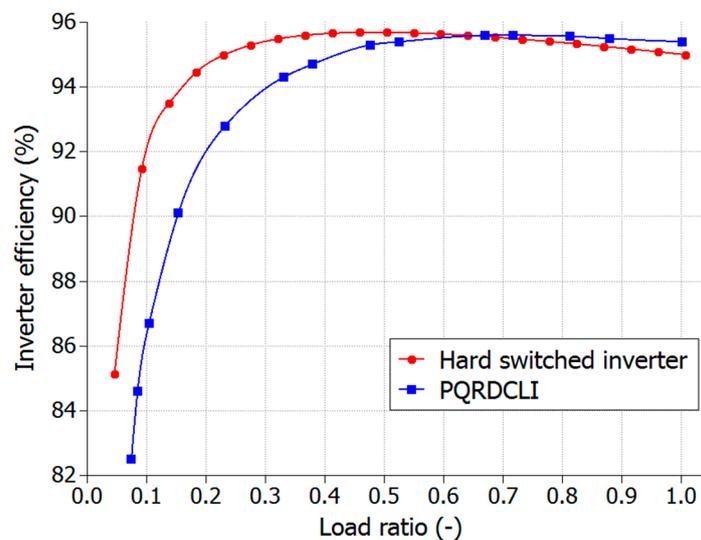


**Figure 19.** Waveforms of CM voltage  $u_{N-PE}$ , shaft voltage  $u_{SH}$ , and leakage ground current  $i_{PE}$  in a drive system fed by a hard-switched inverter with 720  $\mu\text{H}$  common-mode choke: (a) measurement; (b) simulation.

Simulation and experimental results prove that compared to the hard-switched inverter, using PQRDCLI enables a significant reduction of CM voltage amplitudes. Hence shaft voltage levels are also limited, which leads to a decrease of EDM bearing's current occurrence probability. Due to the reduction of common-mode voltage  $dv/dt$  gradients, amplitudes of bearing's currents and ground current  $i_{PE}$  resulting from  $dv_{N-PE}/dt$  are also limited. Moreover, semiconductor devices in PQRDCLI operate under lower dynamic stresses due to reduced  $dv/dt$  and  $di/dt$  gradients. Taking these aspects into account, it can be supposed that the reliability of the drive system fed by PQRDCLI should be higher than obtained for a conventional solution with a hard-switched inverter. However, this conclusion must be confirmed by the results of further research.

Relationship between efficiency and load ratio (load ratio is defined as inverter output power related to the nominal output power) of compared inverters is presented in Figure 20. At nominal output power operation, efficiency of PQRDCLI is about 0.5% higher compared to a hard-switched inverter. However, if load ratio is lower than 0.6, additional losses generated in the PQRDCLI quasi-resonant circuit are higher than switching losses in hard switching conditions. This results in lower PQRDCLI efficiency. In effect, Euro efficiency (calculated as it is presented in Ref. [33]) of PQRDCLI (93.9%) is slightly lower than obtained for compared hard-switched inverter (94.7%). Nevertheless, considering electric motors operational features, recommended operational load range should not be lower than 50% of full-load [34,35]. Hence, taking this aspect into account, PQRDCLI may be an attractive alternative for hard-switched inverters, despite of worse efficiency at light loads. An efficiency analysis proves that loss generated by PQRDCLI quasi-resonant circuit are mainly

dominated by conduction losses of transistors  $T_1, T_4$  [8]. Therefore, further research focused on design and optimization of resonant circuit components is still required.



**Figure 20.** Measured relationship between efficiency and load ratio of hard-switched inverter and PQRDCLI.

#### 4. Comparison with Other Solutions

The presented modeling approach was also compared with other solutions. Five hundred PQRDCLI operation periods using different models of paths, cable, and induction machines were simulated under the same simulation conditions, and the results are presented in Table 4. Tests were performed using Dell Vostro 3560 computer equipped with Intel(R) Core(TM) i7-3632QM CPU 2.20 GHz processor and 8.00 GB RAM. In the first case, models of inverter paths divided into four branches for track and based on Wheeler/Schneider formulas [12,16] were selected as a comparative solution. It should be noted that such an approach introduces a large numerical load of simulator solver, which results in a significant increase in the execution time of the simulation.

**Table 4.** Total execution time of simulation for individual models.

| Model   | Execution Time |
|---|----------------|
| Proposed solution   | 142 s          |
| Model of the drive system with inverter model based on Wheeler/Schneider formulas paths model | 430 s          |
| Model of the drive system with an IM model presented in Ref. [14]                             | 172 s          |
| Model of the drive system with a cable model presented in Ref. [12]                           | 144 s          |

Moreover, a large number of model parameters must be then identified, using complicated methods of parameter extraction, which additionally increases the total time of model development. The presented solution, despite using simplified models of paths (with two parameters describing each path), the accuracy of results is satisfactory with lower numerical complexity. Replacing the proposed model of IM, by solution presented in Ref. [14], results in a slight extension of the total execution time of the simulation. However, it should be noted that the model presented in Ref. [14] is described by more parameters (14 in the proposed solution, 18 in the model [14]). Moreover, the accuracy of the model [14] is lower due to the omission of IM common-mode impedance characteristic changes in-band “e” (Figure 7). It should also be noted that proposals of this paper consider the model of IM with less complicated complexity due elimination of magnetic couplings models (those are implemented in the model [14]). Replacing the described model of the cable by a model shown in Ref. [12] does not result in numerical complexity increase. A similar number of parameters also defines both models.

However, due to the elimination of phase-to-phase capacitors applied in the model [12], a procedure of parameter extraction for the proposed model became less time-consuming. It should be mentioned that the total “cost” of the model is composed of introduced numerical loads, availability of parameters, and the total time required to identify all model parameters. Hence, taking these factors into account, the modeling approach presented in this paper offers a satisfactory accuracy with a moderate “cost” of used models.

## 5. Conclusions

The presented modeling approach with simplified models may be successfully applied in electric drive systems simulation in a range of frequency up to 1 MHz. The procedure of the model’s parameter extraction is easy and it is based on the presented analysis of individual frequency characteristic measurements. Second, it is proposed to use available optimization algorithms. Described models of cable and induction machines may be effectively implemented in the simulation of AC drives fed by other types of inverters, e.g., multilevel DC/AC inverter. Obtained accuracy enables using the proposed modeling approach as a useful tool dedicated to the evaluation of CM disturbances reduction methods. Additionally, usefulness of PQRDCLI in CM voltage and ground leakage current reduction in electric drives was confirmed.

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**Conflicts of Interest:** The authors declare no conflict of interest.

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