**A Line-Commutated Thyristor-Bridge Emulated Rotating
Power Electronic Converter for Brushless Exciter
Applications**

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Dedication

To my family and to my wife for always supporting me throughout this process.

Abstract

Conventional brushless exciter supplies field current to the wound-field synchronous machines through a diode bridge rectifier. Though these diode bridge based exciters are more reliable than the static exciters, which use brushes and slip rings, they can not produce a negative voltage across the field winding, making the rate of decay of the field current slow. In case of an internal fault in the synchronous machine, the field current needs to be removed rapidly.

In this dissertation, a novel thyristor emulated rotating power electronic converter (RPEC) is proposed for brushless exciter (BLE) applications. In this proposed three-phase AC-DC converter, each phase comprises a diode in series with a MOSFET. This converter can produce both positive and negative voltage across the field winding. It also can operate at higher frequencies, and thus exciter will be more compact and energy-efficient. This topology also eliminates the need for a DC-link capacitor, which improves the brushless exciter system's reliability.

Moreover, this converter needs only a single gate drive power supply as the MOSFETs have their source terminals connected together. Furthermore, as the exciter is operated at higher frequencies, there will be a lower torque ripple at the synchronous machine output. Mode change is detected using a higher frequency component superimposed on the rotating transformer's fundamental voltage, due to which the exciter system is more robust and reliable.

The first part of the thesis describes the proposed rotating power electronic converter operation and gate pulse generation during rectification and inversion modes. The second part of the thesis analyzes the field winding current disruption issue while switching from rectification to inversion mode. A new control algorithm is proposed to avoid the problem of mode change from rectification to inversion. The proposed topology, along with the proposed control algorithm, is simulated using Matlab/Simulink. A hardware prototype has been fabricated and tested. The simulation and experimental results verify the operation and advantages of the proposed topology and control.

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Chapter 1

Introduction

1.1 Synchronous machines

1.1.1 Motivation

Synchronous machines can be broadly divided into two types: (a) permanent magnet type and (b) wound-field type. In general, permanent magnet synchronous generators (PMSG) are highly reliable as they do not need any slip rings and brushes. Also, they are very efficient as there are no windings on the rotor. So, they are often used in a wide range of applications like power generation in the wind, hydro, aerospace, electric vehicles, etc. [2]. However, high strength rare-earth permanent magnets used in these generators have some serious drawbacks - they are a single source, imported items subject to price fluctuation, have huge adverse environmental impacts, and imposes temperature limitations. Also, the permanent magnet's flux density cannot exceed 1.2 T, as shown in Fig. 1.1. An additional current component needs to be injected on the stator side during field weakening as the rotor magnets have a fixed flux level. Due to this extra current, the power factor reduces. Also, the stator side converter needs to be rated for higher kVA [3]. Further, there will be an increase in losses in both the stator side converter and the stator.

Unlike PMSG, a wound field synchronous generator (WFSG) uses a rotor winding excitation for creating flux inside the machine. Therefore, it eliminates all the disadvantages mentioned above, which are associated with using rare earth magnets. Also, as

per Equation 1.1, using magnetic materials with higher flux densities than permanent magnets can improve torque/power density and efficiency. Table 1.1 lists some of these magnetic materials. From this table, we can observe that Hyperco 50H requires fewer ampere-turns compared to other soft magnetic materials.

$$T_{em} = 2 \times \int_{\xi=-\pi/2}^{\xi=\pi/2} dT_{em}(\xi) = 2 \frac{N_S}{2} r \ell \hat{B}_r \hat{I}_S \int_{-\pi/2}^{\pi/2} \cos^2 \xi \cdot d\xi = \left(\pi \frac{N_S}{2} r \ell \hat{B}_r \right) \hat{I}_S \quad (1.1)$$

where T_{em} is the electromagnetic torque, $\hat{B}_r$ is the peak of the rotor-produced flux-density, $\hat{I}_s$ is the stator peak current and others are machine parameters.

Therefore, doubling the flux density could lead to doubling torque density, which will be highly advantageous. So, the thesis's primary motivation is to achieve higher efficiencies and higher torque densities using a compact high-frequency brushless exciter to transfer power to the field winding on the rotor of an electrically-excited synchronous machine.

Figure 1.1: B-H Characteristics of different PM materials

Soft Steel Grade	Flux Density (Tesla)	MMF (Oersteds)
416 Stainless Steel	1.614	245.36
M250-35A5 (M15)	1.980	489.23
35ZH135(Oriented)	2.008	122.15
Hiperco 50H	2.005	12.64

Table 1.1: Comparison of mmf required to maximize flux density in soft magnetic materials

1.2 Basic exciters for wound field synchronous machines

The excitation for wound field type synchronous machines can be categorized into two types: (a) static excitation and (b) rotating brushless excitation. Static exciter shown in Fig. 1.2, supplies current to the field winding of the synchronous generator through brushes and slip-rings [4]. These mechanical contacts cause wear and tear, due to which frequent maintenance of synchronous machines is required. This maintenance results in a significant operational cost for the power plant owners [1].

Figure 1.2: Synchronous generator with static exciter

Replacing the static exciter with a rotating brushless exciter will improve the reliability of WFSG. In conventional brushless synchronous machines shown in Fig. 1.3, rotating exciter supplies DC current to the field winding through a diode bridge [5].

In the case of a WFSG, for an internal fault in the stator, the machine's flux needs

Figure 1.3: Synchronous machine with a brushless excitation system

to be reduced immediately by applying a negative voltage across the field winding. However, diode bridge based brushless exciters cannot generate a negative voltage across the field winding. The decay rate of the field current in these brushless exciters purely depends on the time constant of the field winding. The field current i_f shown in Fig. 1.3 takes 5 to 9 s to reach zero in diode-based brushless exciters [6].

The main objective of this thesis is to find a rotating power electronic converter embedded inside the brushless exciter with the following features:

- Should produce both positive and negative voltages across the field winding
- Should operate at a higher frequency so that exciter size would be small
- No DC link capacitors need to be used to improve the reliability of the brushless exciter system
- Number of controlled switches used should be less
- Should detect mode-change without compromising reliability
- Should have the simple control technique to generate the voltages across the field winding

1.3 Current state of the art

1.3.1 Passive rectification based brushless exciter systems

This section describes various brushless exciter topologies that use diode bridge-based rectifiers to supply field current to the main generator.

Conventional aircraft generator uses the configuration as shown in Fig. 1.4. It has the following three stages: (a) pre exciter, (b) main exciter, (c) main generator [7].

Figure 1.4: Conventional aircraft generator

Pre-exciter is a brushless permanent magnet generator that supplies field current to the main exciter through a stationary three-phase diode bridge and a chopper. The main exciter consists of a wound stator and a wound rotor, which provides excitation to the main field winding through the rotating rectifier. One of the advantages of this topology is that it is independent of the grid voltage. Besides, as the chopper circuit is stationary, the active switch is easily controllable. However, this topology requires rare earth magnetic materials and has too many stages.

Fig. 1.5 [8] and Fig. 1.6 depict the topologies without pre-exciter. Instead, they only use a single-phase rotary transformer and a 3 phase ac exciter, respectively.

Figure 1.5: Single phase rotary transformer converter topology

For a given field current, the three-phase exciter per phase current is less compared to the single-phase exciter current [7]. Due to this, as the exciter's power rating increases,

Figure 1.6: Two-stage aircraft generator with the three-phase ac exciter

the single-phase exciter's efficiency goes down. In the topology shown in Fig. 1.6, the main exciter mmf rotation direction is made opposite to the prime mover rotation direction, making field coupling between stator and rotor much stronger. Because of the strong coupling, the stator's fundamental frequency is made much smaller at the higher speeds of the prime mover [7]. In such topologies, it is important because the mmf speed of rotation is comparable to the speed of synchronous speed.

The brushless induction excited synchronous generator shown in Fig. 1.8 is similar to that of the topology given in Fig. 1.6. However, in this topology, the exciter is embedded inside the main generator without magnetically influencing one another [9]. This magnetic decoupling is achieved by having the number of poles different from each other. However, this brushless exciter topology is not practically feasible for high power applications due to winding constraints.

Figure 1.7: Brushless Induction excited Synchronous generator

Figure 1.8: Brushless Induction excited Synchronous generator internal structure

For the brushless exciters discussed so far in this section, the main generator field current freewheels through the diode bridge when the exciter input current is zero. This field current takes several seconds (5-9 seconds) to decay to zero as the time constant of field winding is high [6], which necessitates de-excitation of the field winding before any possible damage in case of any internal fault inside the generator.

1.3.2 Brushless exciters with improved dynamic response

Several converter topologies are proposed for the brushless exciters to improve the field current's transient response during fault conditions in generators. These topologies can be broadly categorized into three types: (a) Uncontrolled rectifier with an additional demagnetization circuit, (b) Thyristor-based converters, and (c) Topologies with controlled switches.

Exciter with de-excitation circuit

External circuits are embedded inside the rotating exciter to apply a negative voltage across the synchronous generator's field winding for de-excitation [6], [10], thereby improving the transient response. The de-excitation circuit, shown in Fig. 1.9, consists of a control circuit and a non-linear resistor parallel with a high voltage IGBT switch. IGBT conduction state depends on the output voltage of the rectifier. During normal operation, as the diode bridge's output voltage is positive, the control circuit applies positive gate voltage to the emitter terminal of the IGBT which will turn it on. When the exciter current is zero, as shown in Fig. 1.10, the diode bridge's output voltage is zero. IGBT remains off during this freewheeling period as there is no positive voltage applied across the switch's gate emitter terminals. Due to this, field current flows

through the non-linear resistor, which causes a negative voltage to be applied across the field winding. As the time constant is reduced due to the resistor's insertion, the field current comes to zero much faster.

Figure 1.9: Brushless exciter with de-excitation circuit during normal operation

Figure 1.10: Brushless exciter with de-excitation circuit during de-excitation

As this de-excitation circuit requires no external signals from outside the exciter, it is easy to implement. During the de-excitation, the field winding energy is dissipated in the form of heat across the resistor. However, this exciter has the following disadvantages: resistor value used needs to be re-evaluated as the exciter rating changes. Also, due to this external resistor, exciter size will not be compact as it needs a heat sink to dissipate the heat generated. Moreover, the voltage rating of the field winding will increase.

Thyristor-based brushless exciters

Direct control over the synchronous machine's field current is achieved by replacing the diodes with thyristors in brushless exciters. A topology of the outer pole thyristor-based brushless exciter is shown in Fig. 1.11. In this topology, the thyristor's firing angle is varied to magnetize and demagnetize the field winding. These firing angles are controlled remotely from outside the exciter using wireless communication [11]-[12].

Figure 1.11: Brushless exciter with rotating thyristor bridge rectifier

A disadvantage of this topology is that it can only be operated at lower frequencies as thyristor switching times are very high. Due to this, there will be a higher ripple in the field current. Correspondingly, torque ripple defined as per equation (1.2) is very high. Also, the exciter size will be huge. As the firing angle increases, the power factor reduces, thereby reducing the exciter's utilization in this converter.

$$TRF = \frac{\Delta T}{T_o} = \frac{T_{\max} - T_{\min}}{T_o} \quad (1.2)$$

Reduction in the torque ripple factor (TRF) in Fig. 1.11 is achieved by six-phase PM excited 12 pulse thyristors connected in series and parallel configuration as shown in Fig. 1.12 and Fig. 1.13, respectively [1].

Figure 1.12: Six phase PM excited series connected thyristor bridges

Figure 1.13: Six phase PM excited parallel connected thyristor bridges

Table 1.2 summarizes these brushless exciter's characteristics given in Fig. 1.12 and Fig. 1.13 at a field current of 46.5 Amps.

The exciter shown in Fig. 1.12 is capable of producing more voltage, whereas the exciter given in Fig. 1.13 can produce less torque ripple [1]. The advantages of these two brushless exciter systems are combined by six-phase PM excited 12 pulse hybrid mode thyristor bridge [1] as shown in Fig. 1.14.

	Series configuration	Parallel configuration
Firing Angle	54.9°	10.5°
Phase current Magnitude	High	Low
Torque ripple	High	Low
Power Factor	Low	High
Phase Current Distortion	Low	High
Output DC Voltage Ripple	High	Low

Table 1.2: Evaluation of the six phase PM excited series and parallel connected rotating thyristors [1]

Figure 1.14: Six phase PM excited hybrid connected thyristor bridges

Since the parallel mode thyristor-bridge has a lower armature current for a given field current, the exciter's thermal stress goes down. Also, it has the benefit of redundancy during thyristor loss or loss of a phase due to failure.

In hybrid mode, interchangeability of the topologies from series to parallel combination is achieved by using one active switch. This brushless exciter is operated in series mode during fault ride-through events, resulting in improved synchronous machine's transient stability. Otherwise, it is generally operated in parallel mode due to the reduced torque ripple. More importantly, in parallel mode, a negative voltage across the field winding is possible as the switch is open. In contrast, it is not possible in the series

mode, as there are freewheeling diodes connected across the field winding.

In these six-phase thyristor-based topologies, the number of semi-controlled switches used is more than the three-phase thyristor topology. Because of this, the reliability of the exciter system goes down.

Embedding fully controlled switches in brushless exciters

This subsection states various brushless exciters with fully- controlled switches for controlling the field current.

The field current's dynamic response is improved by connecting a chopper at the diode bridge rectifier's output, as shown in Fig. 1.15 [13]. In this system, the input voltage to the main exciter is always maintained at the rated voltage. The power electronic topology shown in this system has following advantages: a) field current can be directly controlled by controlling the chopper b) this dual quadrant topology obtains both positive and negative voltage across the field winding.

Figure 1.15: Brushless exciter with diode bridge and chopper

On the other hand, since this brushless exciter is always applied with a rated input excitation current (I_{exc}), there are resistive losses in the exciter that occur even if there is a light load at the output of the synchronous generator. Also, hysteresis and eddy current losses are present due to the rated excitation current at light load.

The following topology shown in Fig. 1.16 has been proposed to improve the efficiency, which is an extension of the exciter shown in Fig. 1.15. In this brushless exciter, an additional single quadrant chopper is connected to the exciter's field winding. In this dual controlled structure, the terminal voltage is controlled by the main chopper, whereas the exciter current is controlled by the exciter chopper leading to lower exciter losses.

Figure 1.16: Brushless exciter with dual control

As shown in Fig. 1.15, there is a similar topology analyzed in [11]. The only exception is that instead of electrical excitation, PM-based excitation is used, as shown in Fig. 1.17. In Fig. 1.15 and 1.17, both active switches are turned on during magnetization. For demagnetization of the field winding, both the switches are turned off to apply negative voltage. However, during this period, in both of these topologies shown in Fig. 1.15, 1.17, all the field energy will be transferred to the capacitor, causing a voltage rise. So the capacitor voltage should be rated large enough to withstand demagnetization.

Figure 1.17: PM exciter with rotating diode bridge and dual quadrant chopper

The voltage rise issue in Fig. 1.17 is eliminated by replacing all the diodes with active switches, as shown in Fig. 1.18 [14]. This brushless exciter consists of a voltage source converter (VSC) based rectifier and H bridge chopper.

Figure 1.18: PM exciter with VSC rectifier and H bridge chopper

Since this brushless exciter has fully controlled switches, the rectifier can be operated at a unity power factor, because of which the size of the exciter goes down. Compared to the thyristor-based brushless exciter given in Fig. 1.11, this exciter has less torque ripple factor. Field current is controlled by changing the duty ratio of the H-bridge.

However, additional sensors are required for the phase-locked loop to generate PWM signals for the rectification.

Converters with fully controlled switches, proposed so far for brushless exciter application, have the following disadvantages: (a) they require a DC-link capacitor due to which the system's reliability goes down (b) the number of switches used is more.

1.4 Proposed solution and outline of the thesis

A line-commutated thyristor-bridge emulated rotating power electronic converter for brushless exciter application is proposed in this dissertation. The overall scheme of the proposed brushless exciter system is shown in Fig. 1.19.

Figure 1.19: Brushless excitation system with thyristor emulated rotating power electronic converter

This exciter consists of an excitation inverter, a high-frequency rotating transformer, and a thyristor emulated rotating power electronic converter connected to the field winding of the wound field synchronous machine (WFSM). Each phase of this rotating power electronic converter has a MOSFET in series with the diode. Due to this configuration, only unidirectional current flow is possible.

The excitation inverter shown in Fig. 1.19 is powered from a battery to be unaffected

by grid side faults. It is used for exciting the stator of the rotating transformer. It also produces a high-frequency voltage of a low magnitude and a fundamental voltage on the transformer's stator side. So, there will be a high-frequency voltage superimposed on the fundamental voltage induced on the rotor side, which acts as a three-phase four-wire AC input source to the rotating power electronic converter. This high-frequency component determines the mode of operation on the rotor side converter. These induced voltages are also used to power the gate drive circuitry for the MOSFETs. As shown in Fig. 1.19, the field winding's negative terminal is connected to the rotating transformer's neutral. An advantage of the proposed half-bridge scheme is that, with the sources of the three MOSFETs connected, only a single power supply is needed in the MOSFET gate-drive circuitry.

This proposed rotating power electronic converter gives both positive and negative voltages that can be applied across the field winding of the wound field synchronous machine. Also, unlike the existing topologies proposed in the literature, which use fully-controlled switches, the proposed line-commuted thyristor bridge emulated converter requires fewer controlled switches. It also eliminates the need for a DC-link capacitor, thus improving the reliability of the system. Another advantage is that this converter can be operated at much higher frequencies, due to which the rotating transformer size goes down. Besides, the ripple in field current will be less due to the high-frequency operation. High-frequency component in the rotor voltage eliminates wireless communication requirements between stationary and rotating components.

The analysis, simulation, and hardware implementation of rotating power electronic converter are done in this dissertation, whereas the design and the FEA analysis of the rotating transformer have been left for future work.

Chapter 2 presents a detailed discussion of the modulation technique for generating the output voltage across the field winding along with the simulation results. The field winding current disruption issue during the mode change is discussed in Chapter 3. This chapter also presents a new algorithm for gate pulse generation, which eliminates the disruption issue. Chapter 4 provides experimental results confirming the analytical predictions and advantages of the proposed control algorithm.

Most of this chapter's contents are reproduced from [15], whereas for chapters 2, it is taken from [16].

Chapter 2

Operation Principles of Proposed RPEC topology and Simulation

A brief introduction of the proposed topology has been presented in the previous chapter. This chapter presents a more in-depth description on some of the operational aspects of the proposed converter.

2.1 Proposed RPEC equivalent circuit

The rotor side equivalent circuit of Fig. 1.19 is shown in Fig. 2.1. Here L_{lr} represents leakage inductance of the rotor side winding, and current source i_d represents the field winding of the wound field synchronous machine. Diode and MOSFET combination in each phase of the converter emulates the thyristor switch. This combination helps to switch at higher frequencies than the available thyristors are capable of, which reduces the excitation system's size. Some of the characteristics of this combination are compared to Thyristor in Table 2.1. One of the similarities between them is that they conduct only in one direction. The main difference is that when there is a positive current, the thyristor can not be turned off by removing the gate pulse. However, the diode and MOSFET combination will turn off if the gate voltage to the MOSFET goes to zero. Therefore we have to be cautious while turning off the gate voltage.

Figure 2.1: Brushless exciter's rotor side equivalent circuit

	Switch type	Current flow direction	Turn-off condition
Thyristor	Semi-controlled	Uni-directional	when current goes to zero
Diode & MOSFET Combination in RPEC	Fully-controlled	Uni-directional	when the gate pulse is off

Table 2.1: Basic characteristics of the switches

Rotor phase and induced voltages are related as given in equations 2.1, 2.2, 2.3, where v_{ra} , v_{rb} , v_{rc} denote rotor side induced fundamental voltages and v_h denotes the high frequency rotor voltage.

$$v_{an} = v_{ra} + v_h \quad (2.1)$$

$$v_{bn} = v_{rb} + v_h \quad (2.2)$$

$$v_{cn} = v_{rc} + v_h \quad (2.3)$$

$$v_{ra} = V_m \sin \omega t \quad (2.4)$$

$$v_{rb} = V_m \sin \left(\omega t - \frac{2\pi}{3} \right) \quad (2.5)$$

$$v_{rc} = V_m \sin \left(\omega t + \frac{2\pi}{3} \right) \quad (2.6)$$

$$v_h = kV_m \sin n\omega t \quad (2.7)$$

In equations 2.4, 2.5, 2.6, 2.7, V_m denotes the peak of the rotor voltage, whereas ω represents the angular frequency. Here k is a real value, and n is an integer. The value of ω is dictated by the magnetic properties of the rotating transformer material. In this thesis, $k=0.1$ and $n=4$ are considered.

2.2 Gate signals generation algorithm for thyristor-bridge emulated RPEC

Fig. 2.2 shows the block diagram for gate signal generation for the rotating power electronic converter. Here, output gate signals depend on the control input signal, as shown in Table 2.2. In Fig. 2.2, Input-1 is connected to Inversion mode logic output, whereas Input-2 is connected to the rectification mode logic output. During the inversion mode, the phase-locked loop (PLL) is required to know the phase angle of the input voltage (v_{an}) in order to generate pulses.

Control Input	1	0
Output Gate Signals	Input-1	Input-2

Table 2.2: Output and input signals relation with respect to mode change control input

The control signal for the mode change value shown in Fig. 2.2 depends on the presence of a high-frequency signal in the rotor induced voltages. If it is present, the control signal value will be 0; else, it is 1. In conclusion, if a high-frequency signal is present on the rotating transformer's secondary side, rotating power electronic converter operates in the rectifier mode; else, it operates in the inversion mode. This way, mode change is detected without using wireless communication between stationary and rotating parts of the exciter system, due to which reliability of the system improves. Gate signals g_1 , g_2 , g_3 for rectification and inversion mode are shown in Fig. 2.3, Fig. 2.4, respectively.

Figure 2.2: Gate signals generation algorithm for thyristor bridge emulated RPEC

Figure 2.3: Rectifier Mode: a) input AC Voltages which includes both fundamental and harmonic components b) a-phase gate voltage c) b-phase gate voltage and d) c-phase gate voltage

Figure 2.4: Inversion Mode: a) input AC Voltages which includes only fundamental component b) a-phase gate voltage c) b-phase gate voltage and d) c-phase gate voltage

Rectification Mode

Under normal conditions, the rotating power electronic converter is operated in rectification mode to supply the field current for field winding of the wound field synchronous machine. In this mode, the exciter inverter on the stator side injects a high-frequency signal along with the fundamental voltage. On the rotating side, the rotating power electronic converter controller detects the high-frequency component in the rotor voltage and turns on all the MOSFET gate signals g_1 , g_2 , g_3 continuously, as shown in Fig. 2.3. Due to this, the converter operates as a three-phase half-wave diode bridge rectifier where DC output voltage is given by Equation 2.8 where the firing angle $\alpha = 0^\circ$. In this equation, the high-frequency component's effect is neglected as its magnitude is less compared to the fundamental voltage.

$$V_d = \frac{3\sqrt{3}V_m}{2\pi} \cos(\alpha) - \frac{3\omega L_{lr}}{2\pi} I_d \quad (2.8)$$

The Rotor side converter shown in Fig. 2.1 is simulated in rectification mode using Matlab/Simulink. The parameters considered for the simulation are given in Table 2.3.

The obtained simulation results are shown in Fig. 2.5. Following observations can be made from Fig. 2.5: (a) Devices in a phase start conducting when the corresponding phase voltage is more than the rest of the voltages. (b) There is a commutation period μ during which the sum of the current through outgoing and incoming device is equal to the load current. This commutation period is given in Equation 2.9. (c) Currents that are flowing through each phase will always be uni-directional. (d) As the gate voltages for MOSFETs are always on in this mode, there will not be any switching losses in these devices. (e) Due to the leakage inductance on the secondary side of the transformer, devices in each phase will have zero current switching at turn on and turn off instants.

$$\mu = \cos^{-1} \left(\cos \alpha - \frac{2\omega L_{lr}}{\sqrt{3}V_m} I_d \right) - \alpha \quad (2.9)$$

peak fundamental input voltage	98 V
frequency of the fundamental voltage	500 Hz
peak of the harmonic component	9.8 V
frequency of the high frequency voltage	2000 Hz
output load current source value	5 A
leakage inductance (L_{lr})	0.34 mH

Table 2.3: Parameters used for the RPEC simulation during rectification mode

Figure 2.5: Rectification Mode: a) input AC voltages and output DC voltage b) gate voltages c) output DC current and d) input phase current

Inversion Mode

When there is an internal fault in the main synchronous machine, the rotating power electronic converter goes to inversion mode, due to which there will be a negative voltage applied across the field winding to make the field current zero. In this mode, the exciter inverter injects only the fundamental voltage on the transformer's stator side. The controller of rotating power electronic converter detects the absence of high-frequency component and generates the gate pulses, as shown in Fig. 2.4. These gate pulses are similar to the pulse pattern given to a three-phase half-wave thyristor bridge converter operating in inversion mode. Fig. 2.6 shows the rotating power electronic converter gate voltages, output voltage, and phase current during the inversion mode.

In Fig. 2.6, the commutation of current from phase-c devices to phase-a devices is shown with a firing angle $\alpha = 120^\circ$. Unlike three-phase half-wave thyristor bridge converters where each thyristor in a phase conducts for 120 degrees, here in rotating power electronic converter, each switch is on for $120 + \delta$ degrees where δ (tolerance angle) should be more than the commutation period μ . This tolerance angle δ makes

sure that the leakage inductor current of the outgoing phase is zero. Care should be taken while considering the value for the tolerance angle δ such that the sum $(\alpha + \delta)$ should be less than 180° to avoid the commutation failure. Commutation from phase a to b and from phase b to c happens similarly. During this mode, the output voltage is given by Equation 2.8 in which α should be more than 90° .

Figure 2.6: RPEC waveforms during Inversion mode

To simulate the rotating power electronic converter in inversion mode, the value of commutation period μ given in Equation 2.9 needs to be evaluated in order to determine the tolerance angle δ . As mentioned earlier, the value of δ should be more than the μ for the converter's safe operation. For the simulation, parameters given in Tables 2.3, 2.4

are used. Fig. 2.7 shows the rotating power electronic converter input voltages, gate voltages, the output voltage across the load, input phase currents during the inversion mode. In this figure, we can observe that the average voltage across the load is negative. When the gate pulse is given to an incoming device, the current through the outgoing device starts decreasing. The tolerance angle that we considered made sure that the outgoing device's phase current goes to zero before the end of that duration.

commutation period μ (evaluated)	4.2°
tolerance angle δ (considered)	10°
firing angle α (considered)	120°

Table 2.4: Additional parameters used for the RPEC simulation during inversion mode

Figure 2.7: Inversion Mode: a) input AC voltages and output DC voltage b) gate voltages c) output DC current and d) input phase current

Chapter 3

Analysis and Simulation of the RPEC during Mode Switching

In the previous chapter, the gate pulse generation algorithm is presented for the rotating power electronic converter to operate it in rectification and inversion mode. This chapter addresses the issue of field current disruption that is caused during the mode change from rectification to inversion. A new mode switching algorithm is proposed to eliminate this issue. In addition, analysis is done on the maximum firing angle for the rotating power electronic converter during the inversion mode to ensure the field current is not disrupted. Finally, simulation results are presented using the Matlab/Simulink platform.

3.1 Phase current disruption issue during the mode change

Fig. 3.1 shows the gate signals for all three phases of the rotating power electronic converter. These gate signals depend on the mode-change control signal input. This figure shows that when the high-frequency component is present in the rotor voltages, the mode-change control signal value is zero. Otherwise, it is one. When the control input is zero, all the gate pulses to MOSFETs are on.

Similarly, when the control input is one, gate pulses are generated according to the inversion mode logic, as shown in Fig. 2.2. Here for the inversion mode, firing angle of 120° and tolerance angle $\delta=30^\circ$ are considered. Even though the gate pulses (g_1 , g_2 , g_3) for all the MOSFETs are on between t_1 to t_2 , shown in Fig. 3.1, only c-phase is

conducting during this period as c-phase devices are forward biased. So, all the load current flows through the c-phase in this duration. The mode-change control signal changing to 1 at t_2 will disrupt c-phase inductor current as c-phase gate pulse is off immediately after t_2 . This c-phase inductor current disruption will cause huge voltage that can damage the devices used in rotating power electronic converter. This phase current disruption can happen in other phases, which depends on the mode switching instant. But here, only one case is shown where mode change is happening at zero degrees with respect to phase-a voltage.

Figure 3.1: Gate signals during the mode change without using the proposed algorithm: a) input AC voltages b) mode-change control signal c-e) a, b, c phase gate voltages

3.2 New mode-switching algorithm and analysis

A new algorithm is proposed to overcome the phase current disruption issue. The mode-change control signal will be delayed to ensure the phase inductor current is not disrupted in any phase during the mode change throughout the time. The implementation of this algorithm is shown in Fig. 3.2. Here inner block takes the PLL output of phase-a as an input and determines the time delay (T_d), assuming the firing angle is 120° during the inversion mode. This evaluation of T_d happens only when the mode-change signal goes from 0 to 1. Furthermore, the outer block shown in Fig. 3.2 takes the mode-change control signal as an input and gives the output as a delayed version of the control signal using the transport delay block.

Figure 3.2: Proposed new algorithm to modify the control signal for mode change

Fig. 3.3 shows the effect of the proposed algorithm on the mode-change control signal. Initially, mode change happens at 0° of phase-a voltage. The algorithm delays the control signal by 30° to make sure the c-phase current is not disrupted. Here we can observe that the c-phase gate pulse is still on after t_2 .

Figure 3.3: Effect of the proposed algorithm on the mode-change control signal during the mode change: a) input AC voltages b) mode-change control signal and modified mode-change control signal c-e) a,b,c phase gate voltages

Analysis of the maximum firing angle during the inversion mode

Fig. 3.4 and Fig. 3.5 show the gate signals for a firing angle of 135° and a tolerance angle of 30° . The only difference between them is that they have a different instance of the mode-change control signal. Fig. 3.4 shows case 1, where mode change occurs at 0 degrees of the a-phase voltage at t_2 . In this case, between t_1 to t_2 , the c-phase carries the load current as its devices are forward biased. After t_2 , we can observe that the c-phase gate voltage is zero. Due to this, there will be a disruption of the c-phase current, which will damage the converter. To avoid this issue, assuming mode-change control signal is delayed by t_d as shown in Fig. 3.5, where the a-phase voltage phase angle is 45 degrees; this will make the c-phase current continues. However, because of

this delay of the control signal, the a-phase devices start carrying load conducting from t_3 as the a-phase devices are forward biased, as shown in Fig. 3.5. After instant t_4 , the a-phase current is interrupted as the a-phase gate voltage is zero after t_4 . So, from this analysis, we can conclude that the firing angle should not be more than 120 degrees during inversion mode. Since the average output voltage goes down as the firing angle reduces, a firing angle less than 120 degrees is not preferred; this is why the firing angle of 120 degrees is considered for the mode-switching algorithm shown in Fig. 3.2.

Figure 3.4: Case 1: mode change occurs at t_2 where the a-phase voltage is at the zero degrees: a) input AC voltages b) mode-change control signal c-e) a, b, c phase gate voltages

Figure 3.5: Case 2: Assuming mode change occurs at t_4 where the a-phase voltage is at 45 degrees: a) input AC voltages b) mode-change control signal and assumed modified mode-change control signal c-e) a,b,c phase gate voltages

3.3 Simulation of the rotating power electronic converter

Matlab/Simulink toolbox is used to simulate the model. The simulation of the model is divided into two major sections, the circuit model and the control. Fig. 3.6 shows the circuit model implementation. As mentioned in Chapter 1, only the power electronic part in the rotating part of the high-frequency brushless exciter is validated in this thesis. The three-phase voltage shown in this figure emulates the rotor voltages of the rotating transformer. This voltage source generates a peak fundamental voltage of 98 V, 500 Hz, and a high-frequency component of 9.8 V peak, 2000 Hz. An inductor of 0.34 mH is series-connected in each phase to emulate the three-phase high frequency

rotating transformer's leakage inductance. We can also observe that the circuit has a diode in series with a MOSFET in each phase. And all these MOSFET's source terminals are connected to the one end of the load terminal. As explained in Chapter 2, connecting the MOSFETs' source terminal requires only one gate drive power supply. Load resistance of $3\ \Omega$ and an inductance of $2.96\ \text{H}$ is considered to represent the field winding of a wound field synchronous machine in the simulation.

Figure 3.6: Rotating power electronic converter model

The implementation of the control section is divided into three major parts. They are PLL implementation, mode-switching algorithm implementation, gate pulse generation implementation.

Fig. 3.7 shows the PLL implementation in which a sawtooth waveform is used to emulate the phase of the rotor fundamental voltages. The actual single phase PLL algorithm is avoided in the simulation, which will significantly improve the simulation time compared to the actual algorithm, which uses proportional and integrator controllers.

Figure 3.7: Phase Locked Loop Emulation

Fig. 3.8 shows the proposed mode-switching algorithm implementation, which is given in Fig. 3.2. The invert command block shown in this Fig. 3.8 mimics the original mode-change control signal. Depending on the instant of the mode switching and the angle of a-phase voltage, the delay time is determined according to the algorithm to make sure the phase current is not disrupted. Finally, invert command is delayed by the evaluated delay time using a transport block.

Fig. 3.9, Fig. 3.10, Fig. 3.11 shows the implementation of gate pulse generation for each phase of the MOSFETs as given in Fig. 2.2. Here, the invert command delayed signal acts as a control input. When this control input is zero, the converter operates in rectification mode; otherwise, it operates in inversion mode.

Figure 3.9: Gate pulse generation implementation for the a-phase MOSFET

Figure 3.10: Gate pulse generation implementation for the b-phase MOSFET

Figure 3.11: Gate pulse generation implementation for the c-phase MOSFET

3.3.1 Simulation results

Fig. 3.12, Fig. 3.14 shows the output waveforms without and with the proposed algorithm. Fig. 3.13 and Fig. 3.15 show the zoomed version of Fig. 3.12, Fig. 3.14, respectively.

Here following observations can be made: 1. When the algorithm is used, output voltage transition is smooth during the mode change. However, when it is not used, we can see a massive spike in the output voltage due to the phase current disruption. 2. During the rectification mode, the steady-state field current 27 A matches the analytical expression given by Equation 3.1 where V_m denotes the peak of the fundamental voltage, and R represents the field winding resistance. 3. For the inversion mode, tolerance angle is evaluated using Equation 3.1 and Equation 2.9, which comes around 26.7 degrees. For this simulation, the firing angle of 120° and the tolerance angle of 30° are considered to ensure the phase current is not disrupted. 4. During the inversion mode, we can observe that field current finally reduces to zero. 5. Output power of this converter is given by the $I_d^2 \cdot R$, which comes around 2.1 kW.

$$I_d = \frac{3\sqrt{3}V_m}{2\pi R} \quad (3.1)$$

Figure 3.12: RPEC waveforms without using proposed algorithm: a) input AC Voltages b) g_1 , g_2 , g_3 gate pulses c) output DC current and DC voltage d) input phase currents

Figure 3.13: RPEC waveforms without using proposed algorithm (zoomed version): a) input AC Voltages b) $g1$, $g2$, $g3$ gate pulses c) output DC current and DC voltage d) input phase currents

Figure 3.14: RPEC waveforms with the proposed algorithm: a) input AC Voltages b) $g1$, $g2$, $g3$ Gate pulses c) output DC current and DC voltage d) input phase currents

Figure 3.15: RPEC waveforms with the proposed algorithm (zoomed version): a) input AC Voltages b) g1, g2, g3 Gate pulses c) output DC current and DC voltage d) input phase currents

Chapter 4

Experimental Results

4.1 Experimental Setup

A hardware prototype of 2 kW rating to validate the proposed converter and the algorithm experimentally is developed. Fig. 4.1 shows the hardware setup's schematic, whereas Fig. 4.2 and Fig. 4.3 show the actual hardware prototype.

Figure 4.1: Experimental setup schematic

Figure 4.2: Hardware prototype

Figure 4.3: Hardware prototype

4.1.1 Voltage source

A three-phase programmable AC power supply is used to emulate the rotating transformer's secondary voltages, as shown in Fig. 4.4. This voltage source is capable of producing the fundamental voltage superimposed on the high-frequency component. As per the safety protocol of the lab, the operating AC voltage should not exceed 30 V. So, a fundamental component of the 30 V AC RMS, 500 Hz, is applied as an input to the converter along with a high-frequency component of 3 V AC RMS, 2000 Hz.

During the mode switching from rectification to inversion, the output power is negative for a small duration. Eventually, this output power goes to zero. However, this three-phase AC chroma power supply does not have the ability to sink the negative power. If precautions are not taken, this negative power will damage the power supply. A resistor bank of $15\ \Omega$ is connected across the AC terminals to protect the supply during the mode change, as shown in Fig. 4.1. So, after connecting only the resistor bank, the power drawn from the chroma power supply will be 180 W. So, we have to make sure the load power does not exceed 180 W during the mode switching so that power drawn from the source is always positive through out the operation.

Figure 4.4: Three phase programmable Chroma AC power supply-61702

The single-phase AC voltage with and without high-frequency components are shown in Fig. 4.5, Fig. 4.6, respectively.

Figure 4.5: Fundamental AC phase voltage of 30 V RMS, 500 Hz

Figure 4.6: High frequency component of 3 V RMS, 2000 Hz superimposed on the fundamental AC phase voltage of 500 Hz

4.1.2 Power semiconductor devices

For the hardware prototype, devices shown in Table 4.1 are used. For driving the SiC MOSFETs, a half-bridge gate driver is chosen to generate +20 V, -5 V that will power the gate driver IC UCC27211DDAR. Because of the SiC devices, this converter can switch at a very high frequency with fewer losses. For a proof of concept, here,

switching is done at only 500 Hz.

Device	Rating
Power SiC MOSFET	1200V, 25A (Littelfuse LSIC1MO120E0080)
Power SiC Diode	1200V, 40A (Littelfuse LSIC2SD120E40CC)

Table 4.1: Power devices and ratings

4.1.3 Load parameters and leakage inductance

Though the designed converter is rated for 2 kW, the system is operated at only 37 W due to the following two constraints. 1. Operating input voltage should not be more than 30 V AC in the lab. 2. Power drawn from the chroma power supply should always be positive. So, a $33\ \Omega$ and an inductance of 117 mH is chosen for the load parameters. To emulate a rotating transformer's leakage inductance, a 340 μ H, 15 A is chosen, which will ensure the sum of the firing angle and commutation angle not exceed 180 degrees.

4.1.4 AC voltage sensing circuit and controller

To run the proposed algorithm, a TI DSP TMS320F28379D is used. It is capable of running at 200 MHz. As shown in Fig. 2.2, the algorithm needs a-phase voltage information to detect the high-frequency component and determine the phase angle of a-phase voltage. So, to sense the a-phase voltage, an op-amp based voltage sensing circuit is used. The output of the sensed voltage is shown in Fig. 4.7. The controller determines the phase of the a-phase voltage based on the sensed voltage, as shown in Fig. 4.8. For the PLL implementation, SPLL_1PH_SOGI_FLL API is used that provides different functions to implement software PLL based on the orthogonal signal generation using second-order generalized integrators and frequency locked loop [17].

Figure 4.7: a-phase voltage and voltage sensor output

Figure 4.8: PLL output of the a-phase voltage

4.2 Hardware Results

4.2.1 Gate signals generation algorithm implementation in DSP

RPEC gate pulse generation algorithm given in Fig. 2.2 and Fig. 3.2 is implemented as shown in Fig. 4.9, 4.10, 4.11, 4.12. Fig. 4.9 shows the main function where clock frequency, EPWM, ADC modules are initialized. After initializing these modules, the trip zone signal is enabled, which will make all the PWM outputs high so that the converter runs in rectification mode at startup. Fig. 4.10 shows the interrupt service routine in which the mode-switch algorithm is executed. This ISR function is executed every $10\mu s$.

As mentioned earlier, SPL11PH.SOG1FLL API is used to detect the angle of the a-phase voltage. This API needs fundamental frequency as an input during the initialization. During ISR execution, this PLL function takes the ADC output of the a-phase voltage as an input and gives the phase angle as output. FFT algorithm is used to detect the high-frequency component in the sensed a-phase voltage. FFT spectrum of the sensed signal is determined after every 512 samples. So, there will always be a delay in the range of milliseconds to detect the high-frequency component. Fig. 4.11 shows the first sub-block of the main ISR flowchart where manipulation of variables being done depends on the high-frequency component. Mode-change detection algorithm implementation is given in Fig. 4.12, where the mode-change signal changes to 1 if there is no high-frequency component in the sensed voltage. It also determines the sector where mode change occurred. Fig. 4.13 shows the algorithm for modifying the mode change transition to ensure phase currents are not disrupted.

Figure 4.9: Algorithm implementation shown in flowchart form

Figure 4.10: Algorithm implementation shown in flowchart form

Figure 4.11: First sub-block of the ISR flow chart

Figure 4.12: Second sub-block of the ISR flow chart

Figure 4.13: Third sub-block of the ISR flow chart

4.2.2 PWM output signals during the mode change

This section shows PWM signals when the mode switch happens between 0° to 30° . In Fig. 4.14, we can observe that when there is no high-frequency component in a-phase

voltage, the mode-change control signal changes to 1. However, there is a delay of 18.2ms detecting the mode change, as shown in Fig. 4.15. This delay is caused due to FFT algorithm, as explained in the previous section. Using a DSP with higher oscillator frequency or using separate analog circuitry to detect the high-frequency component can avoid this. From Fig. 4.16, we can observe that mode-change is detected at 18.3245° of the a-phase voltage. However, as per the proposed algorithm shown in Fig. 3.2, the mode-change signal should be delayed by 12° to ensure the phase current is not interrupted, which can be observed in Fig. 4.16. Here, $100\mu\text{s}$ time difference corresponds to the 18° delay. This additional 6° delay makes sure there is no glitch in the PWM outputs. Final PWM output signals during the mode switching are shown in Fig. 4.17. This gate pulse pattern matches with the simulation results shown in Fig. 3.3. The PWM signals for other cases where mode switch happens other than 0° to 30° range are given in Appendix A.

Figure 4.14: Mode change detection due to the missing of high frequency component in a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure 4.15: Mode change is detected after 18.2ms after missing of high frequency component in a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure 4.16: Mode change is detected at 18.3245° of the a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure 4.17: PWM outputs when the mode-switch occur at 18.3245° of a-phase voltage:
 1) modified mode-change control signal 2-4) phase a,b,c PWM outputs

4.2.3 Rotating power electronic converter output waveforms

Fig. 4.18 shows the rotating power electronic converter output voltage and output current during rectification to inversion mode transition. Here, we can observe that when the modified mode-change control signal is zero, the converter operates in rectification mode. During this period, the output DC voltage is equal to the peak of the instantaneous AC input voltages which is shown in Fig. 4.19. Since there is a delay due to FFT algorithm execution for detecting mode change, we don't see any high-frequency component in the output voltage. Output current shown in Fig. 4.18 matches with the analytical relation given by Equation 3.1. When the modified mode-change control signal changes to one, the output load current reduces to zero within 3.983 ms due to the negative voltage applied across it. Here, we can observe that there is no voltage spike during the mode transition. So, these waveforms validate the converter operation and the algorithm proposed in this dissertation.

Figure 4.18: RPEC waveforms during rectification to inversion mode transition: 1) modified mode-change control signal 2) output DC voltage 2) output DC current

Figure 4.19: Rectified output voltage when there is only a fundamental component 1-3) input AC voltages 4) output DC voltage

Waveforms depicting inversion to rectification mode transition is shown in Fig. 4.20. Here, when the modified mode-change control signal changes to zero, rotating power

electronic converter starts operating in rectification mode. The output voltage during this period is the same as the DC voltage shown in Fig. 4.21. Here we can observe a high-frequency component in the output DC voltage along with the fundamental.

Figure 4.20: RPEC waveforms during inversion to rectification mode transition: 1) modified mode-change control signal 2) output DC voltage 2) output DC current

Figure 4.21: Rectified output voltage when there are both fundamental and high frequency component 1-3) input AC voltages 4) output DC voltage

Chapter 5

Conclusions

The rotating power converter proposed in this dissertation is an ideal solution for brushless exciter applications due to the following advantages: a) Can produce both positive and negative voltage across the field winding, b) Can operate at higher frequencies so that exciter transformer size would be small, c) Eliminates the need of DC-link capacitor due to which system is more reliable, d) Requires only three controlled switches, e) Uses simple control technique to generate the voltages across the field winding, f) Eliminates the use of wireless technology to detect the mode change, g) Requires only one gate drive power supply as source terminals of the MOSFETs are connected and h) Load current ripple is significantly less as it can operate at higher frequency range.

A simple control technique is proposed for rotating power electronic converter to operate it in rectification and inversion mode. Phase current disruption issue is identified while switching from rectification to inversion mode. A mode switching algorithm to eliminate this issue is proposed in this dissertation. Also, analysis has been presented for maximum firing angle that can be used for rotating power electronic converter during the inversion mode.

The proposed topology functionality and the proposed mode switching algorithm have been verified through both simulation and experimental results.

5.1 Future work

The following future work is proposed to improve the viability of the proposed exciter system

- Use a separate analog circuitry to detect the high-frequency component in sensed a-phase voltage rather than using FFT algorithm to reduce the delay time in mode change detection.
- Design a high-frequency transformer along with an excitation inverter and validate the brushless exciter operation with an experimental setup.

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Appendix A

PWM output signals during the mode change for different case scenarios

A.0.1 Mode change between 30° to 150° of a-phase voltage

Figure A.1: Mode change is detected at 91.1491° of the a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure A.2: PWM outputs when the mode-switch occur at 91.1491° of a-phase voltage:
 1) modified mode-change control signal 2-4) phase a,b,c PWM outputs

A.0.2 Mode change between 150° to 270° of a-phase voltage

Figure A.3: Mode change is detected at 246.1187° of the a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure A.4: PWM outputs when the mode-switch occur at 246.1187° of a-phase voltage:
 1) modified mode-change control signal 2-4) phase a,b,c PWM outputs

A.0.3 Mode change between 270° to 360° of a-phase voltage

Figure A.5: Mode change is detected at 345.0935° of the a-phase voltage: 1) input a-phase voltage 2) mode-change control signal 3) modified mode-change control signal

Figure A.6: PWM outputs when the mode-switch occur at 345.0935° of a-phase voltage:
 1) modified mode-change control signal 2-4) phase a,b,c PWM outputs