Switched-Doubly-Fed-Machine Drive for High Power Applications
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Abstract

Converting electricity to mechanical motion is a foundation of modern civilization. A controllable “knob” is often necessary in these electromechanical energy conversion systems to achieve adjustable motion or a process control. An energy-efficient approach to realize this “knob” is through variable-speed drives (VSD), which are power-electronic based converters with associated control operating as an interface between the electrical machine and the electrical source. These drives are not only critical in a wide range of applications including industrial processes, electric propulsion systems, and power generation plants but also becoming increasingly relevant for optimizing energy consumption. For example, a motor without a VSD running at fixed speed can potentially waste 30% to 80% of energy in mechanical throttles located upstream from a compressor or downstream of a pump. In addition to being a controllable knob for energy conversion, these VSDs are configurable to support the electrical source, e.g., the electric grid, through appropriate reactive power support and controllable power factor – a vital feature required for the future electric grid comprising more complex electrical networks.

However, merely 13% of global loads in mega-watt class high-power applications are driven by VSDs. At these higher power levels, the VSD design is significantly challenging due to the limited available power-electronic device ratings and allowable switching frequency leading to design trade-offs among size, efficiency, performance, reliability, and cost. This dissertation proposes a switched-doubly-fed machine (switched-DFM) drive that uses a parallel architecture for electromechanical energy conversion to reduce the required power processing capability of the power-electronic converter by two-thirds while operating seamlessly over a wide speed range. Additionally, the proposed architecture provides exciting opportunities for supporting the electric grid with reactive power not only through the VSD but also using the electrical machine. The approach confronts the challenges of high power electromechanical energy conversion from the perspective of electromagnetics, power electronics, circuit designs, embedded computing, and control to push the trade-off boundary for the VSD to be physically small, efficient, reliable, flexible, inexpensive, and electric-grid friendly.

The thesis contributions include a design procedure for the proposed switched-DFM drive based on a required drive-torque-speed capability, a control architecture that can achieve seamless performance across the entire speed range from the perspectives of the electrical grid and the mechanical load, multiple transfer-switch circuit topologies enabling uninterrupted on-the-fly reconfiguration of the DFM, steady-state and dynamic performance comparison between different switched-DFM drive topologies, and an exploration of DFM electromagnetic design considerations that suit the proposed architecture. A lab-scale experimental setup that emulates an entire power system from generation to consumption is designed and built to demonstrate seamless, wide-speed range, and four-quadrant operation of the proposed switched-DFM drive. The proposed methodologies open up opportunities to create efficient, cost-effective, and sustainable solutions for high power electromechanical energy conversion systems.
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2.20 Experimental result: Speed response of the switched-DFM drive for a step command in speed in the LSI topology.

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Nomenclature

\( \bar{T}_s, \bar{T}_r \)  Stator and rotor current space vectors (A)

\( J, B \)  Total inertia (kgm\(^2\)) & frictional coefficient (Nm-s/rad)

\( T_1, T_2, T_3 \)  Filter time constants (s)

\( P \)  Number of poles of the DFM

\( R_s, R_r \)  Stator and rotor resistances (Ohms)

\( L_s, L_r, M \)  Stator, rotor, and mutual inductances (H)

\( S_w \)  Switching signal for mode changeover

\( T, T_L \)  Electromagnetic and load torques (N-m)

\( \bar{V}_{grid}, \bar{V}_{dc} \)  AC and DC source voltage space vectors (V)

\( \bar{V}_s, \bar{V}_r \)  Stator and rotor voltage space vector (V)

\( \Omega_e \)  Rotor speed in electrical-radian/s

\( \Omega_s, \Omega_{ac} \)  Stator flux frequency & ac supply frequency (electrical-radian/s)

\( \delta \)  Angle between the stator voltage and the stator flux vectors

\( \bar{\Psi}_s, \bar{\Psi}_r \)  Stator and rotor flux space vectors (V-s)

\( \psi_{sa}, \psi_{sa} \)  Stator flux components in \( \alpha\beta \) ref. frame (V-s)

\( X_{ds}, X_{d} \)  d-axis and q-axis components of any space vector \( \bar{X} \)

\( X^* \)  Reference value of variable \( X \)

\( r_s, r_r \)  Stator and rotor resistances in normalized form (p.u.)

\( x_s, x_r, x_m \)  Stator, rotor, and mutual inductances in normalized form (p.u.)

\( v_{dc}, v_{PE} \)  DC supply voltage & power electronics voltage rating (p.u.)

\( i_s, i_r \)  Stator and rotor currents (referred to stator) (p.u.)

\( v_s, v_r \)  Stator and rotor voltages (referred to stator) (p.u.)

\( \psi_s \)  Normalized stator flux magnitude (p.u.)

\( I_r \)  Rated rotor current (referred to stator) (p.u.)
\(\omega_s, \omega_e\)  Stator flux frequency and rotor speed (p.u.)

\(\omega_T, \omega_{\text{max}}\)  Mode transition speed and maximum rotor speed (p.u.)

\(\tau, \tau_L\)  Electromagnetic and load torques (p.u.)
Chapter 1

Introduction

Electromechanical energy conversion is a foundation of modern civilization. A controllable valve is often necessary to achieve adjustable motion in these electromechanical energy conversion systems. A real-world approach for realizing this valve is through variable-speed drives (VSD). VSDs are power-electronic based converters with associated control that interface the electrical machine with the electrical source. These drives are critical in a wide range of applications including industrial processes, electric propulsion systems, and power generation plants. They are becoming increasingly relevant for optimizing energy consumption, e.g., in fans, pumps, compressors, grinding mills, metal rolling, mine hoists, refineries, heating, ventilation, and air conditioning (HVAC) units. For example, a motor without a VSD running at fixed speed can potentially waste 30% to 80% of energy in mechanical throttles located upstream from a compressor or downstream of a pump [1]. In addition to being a controllable knob for energy conversion, these VSDs are configurable to support the electrical source, e.g., the electric grid, through appropriate reactive power support and controllable power factor - a vital feature required for the future electric grid comprising more complex electrical networks. The demand for efficient, cost-effective, and sustainable solutions for these different electromechanical energy conversion systems has led to larger market penetrations of VSDs [2].

Conventionally, power flow is controlled by the VSD to different electrical machine types, including permanent magnet [3], squirrel-cage induction [4], wound-rotor induction [5], and synchronous types [6]. Of these four machine types, permanent magnet, squirrel-cage induction, and synchronous machines are inherently configured with a single electrical port and a single mechanical port for electromechanical energy conversion. For these motors, a VSD connected in series with the electrical port of the machine must be rated at the shaft power to control the entire energy conversion. At higher power levels, for mega-watt (MW) class (i.e., > 1000 HP) motors, the power converter design is limited by the available semiconduc-
tor device ratings and allowable switching frequency. For example, currently the voltage rating of typical Insulated-Gate Bipolar Transistors (IGBTs) used in medium voltage VSDs ranges between 3.3 kV and 6.5 kV with an allowable switching frequency of a few hundred Hz [7]. Often times, multiple devices are stacked in series/parallel as in multilevel converters [8, 9, 10, 11, 12], or the electrical machine is designed with more than three phases as in reference [13]. Reduction in allowable device switching frequency leads to large passive filters that are used at each power conversion interface to ensure allowable distortion in the voltage/current waveforms. Considerable effort from research to product development focuses on different topologies and combinations of multilevel converters and multiphase machines [14, 15, 16] to create high power VSD architectures. However, in reality, only 13% of MW class motors for industrial applications use VSDs despite 48% of these large motors being suitable for VSD integration. Even a 90% deployment of VSDs to these MW class motors has the potential for energy savings of up to 0.7% to 1.8% of total U.S. electricity consumption [17]. The low adoption of VSDs in the MW class applications is primarily due to the high capital cost, large footprint, unfavorable harmonic injections to the electric grid, and high losses in the VSDs. Besides land-based applications, the need for high power VSDs is seeing a considerable traction in transportation electrification where the current trend promises to electrify the propulsion systems, e.g., in a ship or aircraft, that handle powers of several megawatts. Merely transferring the available technology from the land-based, stationary applications to these space-constrained transportation vehicles leads to significant challenges in terms of accommodating real “payloads”.

1.1 Switched-doubly-fed-machine drive architecture

Unlike single-electrical-port machines, a doubly-fed machine (DFM) (also known as wound-rotor induction machine) has two accessible electrical ports – a stationary winding (stator) and a rotating winding (rotor) – that add flexibility in the drive design. This flexibility has been widely used in limited-speed range applications, such as in wind power generation [18] and fly-wheel energy storage [19, 20], where the stator of the DFM is directly connected to the ac source while a power converter interfaces between the rotor and the ac source. Based on the power converter rating, the DFM can be controlled across a speed range around the ac source synchronous speed [5]. For example, in a typical wind power generation system, the power converter rating is one-third compared to that of the electrical machine size while operating on a speed range of ±33% around the synchronous speed. However, a wide operating speed range is required in many industrial and propulsion applications, including operation at zero speed. The advantage of the conventional DFM drive is lost because in this case the power electronic converter rating must be equal to that of the motor rating. Other approaches like feeding the DFM with controlled double inverters on both
the windings increases the speed range with rated torque capability at all speeds [21, 22]. Although this has the advantage of having complete control on both the stator and the rotor excitation, the total power electronic converter rating matches the rated mechanical shaft power, leading back to the same challenge of massive power electronic converters.

Operating DFM drives in different configurations using external switches opens up opportunities in achieving a wider operating speed range with reduced power electronics. Denoting this group of DFM-based drives as “Switched DFM”, these drives typically have a controlled back-to-back converter on one of their electrical ports while the other electrical port is connected to one or many sources on-the-fly depending on the operating speed. For example, reference [23] reports a drive configuration in which during low speed operation the stator is shorted while during high speed operation, the stator is connected to an ac source. The rotor is fed with a back-to-back converter connected to the ac source for the entire speed range. Similar approaches have been used for developing and conceptualizing variable speed drives for hydro-electric power stations [24], ship propulsion [25], and centrifugal loads [26]. Unlike the configuration of shorting the stator, the switched-DFM drive in [25] proposes to use a dc source excitation during the low speed operating mode.

Based on these different architectures reported in the literature, switched-DFM drives can be broadly classified into two categories. In the first category, denoted as “Low-speed induction” (LSI) topology in this thesis, the DFM stator is shorted at lower drive speed to emulate a cage-induction machine. The topology is shown in Fig. 1.1(a). In the second category, termed as “Low-speed synchronous” topology (LSS) in this thesis, the DFM stator is connected to a dc source at lower drive speed thereby emulating a wound-field synchronous machine. The topology is depicted in Fig. 1.1(b). The functionality of the dc source is similar to an exciter of a synchronous generator used in power generation plants. In both the topologies, the DFM stator is connected to the ac source/grid at higher drive speed making the DFM operate as a grid-connected doubly-fed induction machine. For simplicity, the low speed operating configuration of the drive is denoted as low-speed mode while the high speed operating configuration is denoted as high-speed mode. In low-speed mode, the power to the shaft is entirely provided from the rotor while in high-speed mode, the bulk of the power to the shaft is directly provided by the stator port (without being processed by the rotor variable speed drive) and the rotor controls only the differential power. Switching between the two regimes at an appropriate transition speed leads to a reduction in the rotor power electronics requirement as well as a wide speed range operation.
Switched-DFM drive
Shorted
AC source/Grid (Constant voltage, constant frequency)

Transformer-Rectifier (AC-DC) (constant voltage) DFM
AC source/Grid (Constant voltage, constant frequency) Shorted
Stator
Back-to-back converter (AC-DC-AC)

(a)

Figure 1.1: Architectures of switched-doubly-fed machine drives (a) Low-speed induction topology (LSI): At lower shaft speed, the stator is shorted (b) Low-speed synchronous topology (LSS): At lower shaft speed, the stator is connected to a dc source. At higher shaft speed, the stator is connected to the ac source for both the topologies. The transfer switch allows the on-the-fly reconfiguration of the DFM stator connection to the external sources/short depending on the operating mode.

1.2 Challenges for switched-doubly-fed-machine drive

There are multiple challenges in realizing a switched-DFM drive in practice. These challenges have led to a very restricted utilization of the switched-DFM drive instead of the full-power-converter-based-electromechanical drives. Some challenges are listed below:

1. Restricted drive-design procedure: While reference [23] apparently uses an iterative method to choose a transition speed to minimize the rotor converter size, references [25, 26] use steady-state load torque-speed characteristic to determine the size of the controlled rotor converter. However, there are multiple design variables that need to be considered in order to design a switched-DFM drive holistically based on
the required drive-torque speed characteristics and for given DFM parameters. The transient behavior of the drive at the mode transition instant needs to be considered to ensure that the minimum drive size is capable to ride-through the stator transients.

2. Extensive use of mechanical switches for on-the-fly reconfiguration of the DFM electrical port: Sluggish mechanical switches controlling the stator connection to the sources significantly impact the shaft behavior during the mode transition. This leads to poorly controlled speed/torque variations during mode transition [23]. In fact, the switched-DFM drive has been made to operate in two isolated modes—starting and operational—to ensure that the drive does not operate near the transition speed under normal conditions [26].

3. Absence of seamless control: The switched-DFM drive has significantly different challenges from a control perspective compared to a standard grid-connected-doubly-fed induction machine [27, 28, 29, 30] or a wound-field synchronous machine [31]. The rotor power electronics control has to adapt to the change in the machine configuration on-the-fly and should be able to control the individual modes (low-speed or high-speed) of the drive operation. Moreover, the control has to ensure that the DFM undergoes stable mode transition at the transition instances.

4. Unknown grid interaction in a weak-ac grid system: A switched-DFM drive presents a distinct discontinuity in the operation mechanism to the ac source. For example, the connection of the stator to the ac grid during the mode transition instant can be conceived as a step change in power demand to the ac source, if not appropriately controlled. For power systems, as in a ship, where the switched-DFM propulsion drive will consume a considerable share of the generated power, proper control must be ensured so that the ac grid is stable under all operating conditions of the drive, including the instances of the mode transition.

5. Lack of machine design considerations for selection of drive topology: In a switched-DFM drive, the same doubly-fed motor is operated as different conventional electrical machines (either induction/doubly-fed induction or synchronous/doubly-fed induction) depending on the operating speed range. Special considerations must be taken into account from a machine design perspective such that the motor design is suitable for getting the maximum advantage in terms of reducing the required power electronics. Selection of a particular switched-DFM drive topology is driven by the machine parameters for a pre-existing DFM machine design. Alternatively, a particular choice of a switched-DFM drive topology based on additional requirements leads to specific design guidelines for the DFM in “blank-slate” applications.
1.3 Thesis contributions

This thesis addresses the above challenges for a switched-DFM drive to ensure that the drive is operated as any other standard full-power-converter drive but with reduced power electronics requirement and with enhanced reactive power support to the grid for wide-speed-range high-power applications. While the intended application is mainly for marine propulsion, the drive might also be of great interest in other transportation systems such as locomotives, off-highway vehicles, and for large land-based applications like chillers, HVAC fans, compressors, and pumps where an ac source/electric grid is available and adjustable speed is required.

In Chapter 2, a design methodology for a switched-DFM drive is presented that is driven by the required drive torque-speed capability at the mechanical port. The drive power architecture analysis explains the operational benefits of the switched-DFM drive and outlines the key design parameters influencing the drive design. Considering the parameters of a practical doubly-fed machine, an exemplary drive design is presented that explores the design space and draws connections between the different design parameters to achieve minimum rotor power electronics requirement for the entire drive operation. Experimental results validate the proposed design approach for the two topologies of the switched-DFM drive.

A critical challenge for the seamless performance of a switched-DFM drive is the need for an appropriate transfer switch. In Chapter 3, several solid-state transfer switch topologies are presented that enable reconfiguration of the DFM terminals on-the-fly without affecting the sources or disrupting the DFM operation. Thyristors are used to create these transfer switch topologies to ensure a balance between the affordable complexity and the required performance. Thyristors, being non-gate-turn-off devices, act as a constraint to the rest of the drive operation and a careful selection of the mode transition instant enables seamless performance of the transfer switch without requiring any additional supporting circuitry.

In Chapter 4, a control architecture is presented that adheres to the constraints set by the thyristor-based transfer switch while simultaneously achieving a “bumpless” drive performance relative to the mechanical shaft or the electrical source, especially at the mode transition instances. In-built mechanisms in the control scheme allow a seamless control of the DFM from the rotor-side converter, irrespective of the excitation in the stator while remaining within the designed drive torque-speed specifications. A stator flux estimation method presented in this chapter is crucial in making the control scheme feasible in practice.

The switched-DFM drive must also interact with the ac source seamlessly irrespective of the operating modes. This is critical for the stability of the ac grid as well as for the intermediate dc-link between the grid-side and the rotor-side converters. In Chapter 5, a control scheme of the grid-side converter is presented that enables the seamless grid interactions of the switched-DFM drive and provides reactive power support to the grid without adding extra power electronics. Constraints based on the thyristor-based transfer switch
topologies are analyzed to bound the feasible region of operation in terms of the reactive power support.

As there are multiple topologies of the switched-DFM drive, a comparison is inevitable in terms of selecting the correct topology that maximizes the reduction in the rotor power electronics for a given DFM design. In Chapter 6, performance comparison of different switched-DFM drive topologies is presented that leads to machine design guidelines. The comparison is useful from two different perspectives. First, the comparison can be used to select the topology that satisfies the drive requirement for a given DFM design. Second, the comparison can also be used to specify design guidelines for a DFM that suit the chosen topology for specific applications based on additional requirements. An exemplary 30 MW, 4160 V, 60 Hz, doubly-fed machine dimensional design shows the feasibility of the proposed switched-DFM drive for ship propulsion applications.
Chapter 2

Power Architecture and Rotor-side Converter Design

This chapter presents a design methodology for the power architecture of a switched-DFM drive based on a desired drive torque-speed capability. The design objective is to minimize the rotor power electronics requirement while the DFM is operated within rated conditions over a wide speed range. The design process is illustrated using an example DFM with practical machine parameters. The drive design includes the selection of different design variables such as the transition speed between the operating modes, dc source voltage, rotor power electronics voltage and current ratings, maximum speed, and reactive power capability in high-speed mode.

In this chapter, a DFM machine model is described first, which is followed by the analysis of the switched-DFM drive under ideal conditions. The analysis shows the benefit of the proposed drive in reducing the size of the required power electronics and emphasizes other critical design considerations. Next, the non-idealities in the DFM are included to illustrate the entire drive design process using practical DFM parameters. Experimental results validate the proposed drive design methodology in reducing the required power electronics while operating over a wide speed range. The practical DFM is operated in different switched-DFM drive topologies using an ac source of 146 V (line-to-line, rms), 40 Hz which drives a quadratic load torque profile that is similar to a ship-propeller load.
Table 2.1: Defined base quantities for normalization

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<thead>
<tr>
<th>Quantity</th>
<th>Nomenclature</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>$V_B$</td>
<td>AC source voltage magnitude (peak phase)</td>
</tr>
<tr>
<td>Frequency</td>
<td>$\omega_B$</td>
<td>AC source frequency</td>
</tr>
<tr>
<td>Current</td>
<td>$I_B$</td>
<td>Per-phase peak DFM stator current rating</td>
</tr>
<tr>
<td>Flux</td>
<td>$\psi_B$</td>
<td>$V_B/\omega_B$</td>
</tr>
<tr>
<td>Power</td>
<td>$P_B$</td>
<td>$3V_B I_B/2$</td>
</tr>
<tr>
<td>Torque</td>
<td>$\tau_B$</td>
<td>$P_B P/2\omega_B$</td>
</tr>
<tr>
<td>Impedance</td>
<td>$Z_B$</td>
<td>$V_B/I_B$</td>
</tr>
</tbody>
</table>

2.1 Doubly-fed-machine model in stator-flux orientation

A DFM in the stator-flux reference frame is described using standard electrical machine equations [32]. The convention for the two-axis representation of the analytical model is chosen such that the orientation of the stator flux coincides with the d-axis while the q-axis leads the stator flux by 90°. Using normalized variables and parameters with respect to corresponding base quantities as given in Table 2.1, the dynamics of the stator flux magnitude is governed by:

$$
\frac{1}{\omega_B} \frac{d\psi_s}{dt} = v_{sd} - r_s i_{sd}.
$$

The dynamics of the d-axis rotor flux is given by

$$
\frac{1}{\omega_B} \frac{d\psi_{rd}}{dt} = v_{rd} - r_r i_{rd} + (\omega_s - \omega_e)\psi_{rq},
$$

while the dynamics of the q-axis rotor flux is given by

$$
\frac{1}{\omega_B} \frac{d\psi_{rq}}{dt} = v_{rq} - r_r i_{rq} - (\omega_s - \omega_e)\psi_{rd}.
$$

The normalized stator-flux frequency, $\omega_s$, is an algebraic function of the q-axis component of the stator voltage, stator current, and the stator flux magnitude and is given by

$$
\omega_s = \frac{1}{\omega_B} \frac{d\theta_s}{dt} = \frac{v_{sq} - r_s i_{sq}}{\psi_s}.
$$

The electromagnetic torque is governed by:

$$
\tau = -\frac{x_m}{x_s} \psi_s i_{rq}.
$$

The stator flux magnitude is related to the d-axis components of both the stator and the rotor currents and
is given by

$$\psi_s = x_s i_{sd} + x_m i_{rd}. \quad (2.6)$$

However, by the choice of reference frame, the q-axis stator and rotor current components are directly proportional to each other and is represented as

$$0 = x_s i_{sq} + x_m i_{rq}. \quad (2.7)$$

The d-axis component of the rotor flux is influenced by the d-axis components of both the stator and the rotor currents such that

$$\psi_{rd} = x_r i_{rd} + x_m i_{sd}. \quad (2.8)$$

Similarly, the q-axis component of the rotor flux is influenced by the q-axis components of both the stator and the rotor currents such that

$$\psi_{rq} = x_r i_{rq} + x_m i_{sq}. \quad (2.9)$$

A simplified mechanical model for the DFM is represented as:

$$\frac{2J}{P} \frac{d\Omega_e}{dt} + \frac{2B}{P} \Omega_e = T - T_l \quad (2.10)$$

where $P$ corresponds to the number of poles in the DFM, $J$ corresponds to the combined inertia at the mechanical shaft, and $B$ corresponds to the viscous frictional coefficient at the shaft. For example, the load torque can be represented in normalized-form as

$$T_l = k\omega_e^2$$

for a quadratic load torque-speed characteristic of a ship propeller, where $k$ is the constant of proportionality [33].

Eliminating the time-derivative term in (2.1), the d-axis stator voltage is directly proportional to the d-axis stator current in steady-state operation such that

$$v_{sd} = r_s i_{sd}. \quad (2.11)$$
Rearranging (2.4), leads to

\[ v_{sq} = \omega_s \psi_s + r_s i_{sq}. \] (2.12)

Elimination of the time-derivative term in (2.2), results in

\[ v_{rd} = r_r t_{rd} - (\omega_s - \omega_c) \psi_{rq}, \] (2.13)

while elimination of the time-derivative term in (2.3) yields

\[ v_{rq} = r_r t_{rq} + (\omega_s - \omega_c) \psi_{rd}. \] (2.14)

These four equations (2.11)-(2.14) represent the DFM behavior in steady-state operating condition and will be used to design the drive with the proposed architecture.

### 2.2 Operation of an ideal switched-doubly-fed machine drive

Based on the DFM machine model, an analysis for an ideal DFM is considered next to illustrate the benefit of the proposed drive architecture in reducing the power electronic requirement for wide speed range applications. The impact of the drive torque requirement on minimizing rotor power electronics is evaluated that directly translates into the design choice of the mode transition speed and the maximum operating drive speed. Assuming zero resistances, leakages and negligible magnetizing current, the normalized rotor current rating (referred to the stator) is equal to the stator current rating of unity. Using (2.6)-(2.9), (2.13), and (2.14), the required rotor voltage magnitude is given by

\[ v_r = \sqrt{v_{rd}^2 + v_{rq}^2} = |\omega_s - \omega_c| \psi_s. \] (2.15)

In steady-state high-speed mode, with the DFM stator connected to the ac source, the normalized stator flux frequency \( \omega_s \) is identical to the normalized ac source frequency of unity. The stator flux magnitude \( \psi_s \big|_{\text{high}} \) is directly determined by the ac supply voltage and frequency and, hence is also unity in the normalized form. Consequently, the required rotor voltage is proportional to the slip speed and is given by

\[ v_r \big|_{\text{high}} = |1 - \omega_c|. \] (2.16)

The required rotor voltage as a function of rotor speed is shown by the V-curve AEB in Fig. 2.1.
Figure 2.1: Ideal DFM: Rotor power electronics voltage rating of 0.5 p.u. (instead of 1 p.u. as required in standard full-power converter drives) enables operating speed range from 0 to 1.5 p.u. (OAEB) with rated drive torque capability when the rotor speed is normalized to the ac source synchronous speed. For a lower (75%) drive torque requirement in low-speed mode, the required rotor power electronics voltage rating can be further reduced to 0.43 p.u. enabling an operating speed range of 0-1.43 p.u. (OCED).

For maximum utilization of the DFM, the rotor current rating sets the current rating of the rotor power electronics to unity. Unity rotor current and stator flux ensures that the rated torque capability of the DFM is utilized in high-speed mode.

For the ideal DFM in low-speed mode, the normalized stator flux frequency $\omega_s$ is zero irrespective of the drive topology. The stator flux magnitude is a design choice that depends on the desired drive torque requirement. With a unity rotor current rating, the stator flux magnitude $\psi_s|_{low}$ must be equal to the maximum drive torque requirement in low-speed mode $\tau|_{low}$ as governed by (2.5). Therefore, the required rotor voltage is proportional to the product of the rotor speed and the chosen stator flux magnitude and is given by

$$v_r|_{low} = |\omega_c| \psi_s|_{low}.$$  \hspace{1cm} (2.17)

For example, a unity drive torque requirement in low-speed mode implies a required rotor voltage as represented by the line segment $OA$ in Fig. 2.1. Consequently, a transition at $A$ from low speed to high-speed mode ensures drive operation from zero to 1.5 p.u. rotor speed while requiring a maximum rotor voltage of 0.5 p.u. (operating on $OAEB$ path in Fig. 2.1).

With unity drive torque capability across the entire operating speed range, the mechanical power output
Figure 2.2: Ideal DFM: Rotor power electronics power rating of 0.5 p.u. enables operation of the electromechanical drive to provide a maximum shaft power of 1.5 p.u. while operating over a speed range from 0 to 1.5 p.u.

for the ideal DFM is proportional to the rotor speed as shown in Fig. 2.2. The maximum mechanical power output is achieved at 1.5 p.u. rotor speed. The electrical power input from the rotor side of the DFM that is processed by the rotor power electronics is given by

\[ P_r = v_r i_r. \]  

(2.18)

With 1 p.u. rotor current required to obtain the drive torque of 1 p.u., (2.16) and (2.17) are used to calculate the required rotor power across the entire speed range and is shown in Fig. 2.2. Finally, the stator power is obtained by the product of the stator voltage and stator current and is given by

\[ P_s = v_s i_s. \]  

(2.19)

Obviously for the ideal DFM with no losses, zero stator power is required in low-speed mode. With the stator being connected to unity normalized ac source voltage and drawing unity stator current in high-speed mode, the electrical power input from the stator is 1 p.u. as shown in Fig. 2.2. Thus, the rotor power electronics processes only one-third of the maximum shaft power while operating over the entire speed range from 0 to 1.5 p.u.

If a lower drive torque capability is acceptable at low speed, for example 75% of that at high speed, the stator flux magnitude in low-speed mode can be reduced to 0.75 p.u. while maintaining the rotor current
at unity. The reduction in stator flux magnitude directly impacts the required rotor voltage as evident from 
(2.17) (graphically represented by the line segment OC in Fig. 2.1). In this case, transitioning at \( C \) enables 
the drive to operate on a speed range from zero to 1.43 p.u. while requiring a maximum rotor voltage of 
0.43 p.u. \( (OCED \) path). In general, the transition speed that minimizes the required rotor voltage is given 
by equating (2.16) and (2.17) computed at the mode transition speed \( \omega_T \) i.e.;

\[
v_r \big|_{low@\omega_T} = |\omega_T| \psi_s \big|_{low} = |1 - \omega_T| = v_r \big|_{high@\omega_T} = v_{PE}.
\]

(2.20)

With the stator flux magnitude \( \psi_s \big|_{low} \) identical to the maximum drive torque \( \tau \big|_{low} \) in low-speed mode, 
(2.20) simplifies to

\[
\omega_T = \frac{1}{1 + \tau \big|_{low}}.
\]

(2.21)

Using (2.16) and (2.21), the minimum required rotor power electronics voltage rating \( v_{PE} \) is given by

\[
v_{PE} = \frac{\tau \big|_{low}}{1 + \tau \big|_{low}}.
\]

(2.22)

Using (2.15) in high-speed mode and (2.22), the associated maximum achievable rotor speed \( \omega_{max} \) in high-
speed mode with the minimum rotor power electronics voltage rating is

\[
\omega_{max} = \frac{1 + 2\tau \big|_{low}}{1 + \tau \big|_{low}}.
\]

(2.23)

As the maximum mechanical power output is achieved at the maximum speed, (2.23) implicitly represents 
the maximum mechanical power output of the DFM. The minimum rotor power electronics power rating that 
enables the drive operation for the entire speed range is identical to (2.22) since the rotor power electronics 
current rating is designed to be at unity. Therefore, (2.22) and (2.23) can be used to compute the ratio of 
\( v_{PE}/\omega_{max} \), which is also the ratio of minimum rotor power electronics rating to maximum mechanical power 
output. For the example, in which the low-speed mode drive torque requirement is 75% of the high speed 
torque, the minimum rotor power electronics power rating can be as low as 30% of maximum mechanical 
power while operating on a speed range of 0-1.43 p.u.

Figure 2.3 shows the benefit and the design trade-offs for the proposed drive architecture using (2.22) and 
(2.23). The maximum speed requirement for the drive can be achieved in three ways. First, by choosing the 
ac supply frequency precisely in cases where system level design is possible, e.g., a “blank slate” design for a

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ship or locomotive. Second, choosing and designing a DFM with a specific number of poles. Finally, if either of the above two is already constrained, the maximum speed will be set by the minimum voltage rating of the rotor power electronics. An increase in the maximum speed requirement will increase the voltage rating of the rotor power electronics.

To summarize, in high-speed mode the stator flux being strictly determined by the ac supply voltage and frequency, the choice of high speed torque requirement strongly influences the rotor power electronics current rating. On the contrary, the stator flux magnitude in low-speed mode is a design choice, which is driven by the drive torque capability requirement in low-speed mode. The designed stator flux magnitude in low-speed mode impacts not only the rotor power electronics voltage rating but also the transition speed, the maximum speed and, of course, the required rotor power electronics power rating.

2.3 Extended negative speed range operation

The modulus of rotor speed in (2.17) signifies that the required rotor voltage magnitude is independent of the direction of rotation in low-speed mode. Therefore, with no further modification in the power architecture, the proposed drive can operate over the speed range between -0.5 p.u. and 1.5 p.u. with the rotor power electronics sized at one-third of the maximum mechanical power output. However, many applications such as ship propulsion require full four-quadrant operation including a wider negative speed range. A wider
negative speed range is achievable by augmenting the switched-DFM drive with an additional switching mechanism that can change the phase sequence of the ac source relative to the DFM stator connection.

In high-speed mode, the direction of rotation of the stator flux is determined by the phase sequence of the DFM stator connection. Considering ABC phase sequence of the DFM stator connection to the ac source as positive rotation implies that, in steady state,

\[ \omega_s = 1 \]  \hspace{1cm} (2.24)

while changing the phase sequence of the DFM stator to ACB implies

\[ \omega_s = -1. \]  \hspace{1cm} (2.25)

The change in polarity of \( \omega_s \) in high-speed mode depending on the DFM stator phase sequence directly impacts the required rotor voltage as computed using (2.15). The required rotor voltage for the entire speed range is plotted in Fig. 2.4 with respect to normalized rotor speed. For positive rotor speed, the DFM stator is connected to the ac source in ABC phase sequence, represented by PQR segment, while for negative rotor speed the phase sequence is reversed to ACB, represented by segment XYZ. A mode transition between the operating modes at P and X in Fig. 2.4 enables an operating speed range of \( \pm 1.5 \) p.u. of the propulsion drive while requiring a maximum rotor voltage of 0.5 p.u. and a rotor power electronics rated at one-third...
of the maximum mechanical power.

2.4 Advantages and benefits

The reduction in the required power converter rating for the switched-DFM drive pushes the design trade-off curve in which often size, efficiency, and performance are negotiated to remain within the allowable design specifications. Reduction in required power converter rating directly translates into multiple benefits including higher efficiency, reduced output/EMI filter rating, reduced dc-bus capacitor rating, lower fundamental drive frequency, improved machine-side and source-side harmonic injection, and reduced converter cooling requirements. A conventional DFM drive for a wind turbine is known for the better efficiency of the power converter, as only one-third of the total power is processed by the converter compared to a full converter drive [5]. Similar benefits in efficiency are expected from the proposed switched-DFM drive. Reduction of the required dc-bus voltage to control the drive over the complete speed range directly translates into reduction of conduction and switching losses in the converter. In high power applications, where available device rating limits the power handling capacity of the converter, the proposed drive architecture extends the capability limit of the drive without sacrificing the speed range [7]. Additionally, the switched-DFM drive offers seamless interface, controllable power factor, and reactive power support to the ac grid.

2.5 Rotor power converter sizing for practical switched-doubly-fed machine drives

In practice, the presence of non-zero resistances, leakages, and magnetizing current must be considered during rotor power converter design in a switched-DFM drive. The magnetizing current in a DFM can be shared between the stator and rotor windings and is determined at the design stage of the machine. For DFMs designed for grid-connected motoring operations, the stator typically supplies the complete magnetizing current while the speed/torque control is achieved using the rotor-side converter [34]. Alternatively, for DFMs designed for wind power generation, the rotor-side converter is designed not only to provide the entire magnetizing current but also additional stator reactive power [35]. Due to such wide possibilities of providing magnetizing current, the stator and rotor current ratings may not be equal for a practical DFM. Assuming the rated stator current as the base for normalization, the normalized rated rotor current (referred to the stator) is given by,

$$I_r = \frac{N_M}{L_{\text{rms}}|_{\text{rotor}}}$$

$$I_{\text{rms}}|_{\text{stator}}$$

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Table 2.2: Parameters of a 1 HP, 220 V/150 V, 60 Hz, 4 Pole, 1800 rpm off-the-shelf DFM, which are used to design the exemplary switched-DFM drive.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Actual value</th>
<th>Normalized (p.u.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator resistance</td>
<td>3.57 Ω</td>
<td>0.1013</td>
</tr>
<tr>
<td>Rotor resistance (ref. to stator)</td>
<td>4.23 Ω</td>
<td>0.1199</td>
</tr>
<tr>
<td>Stator leakage inductance</td>
<td>9.6 mH</td>
<td>0.1024</td>
</tr>
<tr>
<td>Rotor leakage inductance (ref. to stator)</td>
<td>9.6 mH</td>
<td>0.1024</td>
</tr>
<tr>
<td>Mutual inductance (ref. to stator)</td>
<td>165 mH</td>
<td>1.7630</td>
</tr>
<tr>
<td>Stator current rating</td>
<td>5.09 A (peak)</td>
<td>1.0000</td>
</tr>
<tr>
<td>Rotor current rating (ref. to stator)</td>
<td>3.857 A (peak)</td>
<td>0.7576</td>
</tr>
<tr>
<td>Turns ratio (rotor to stator)</td>
<td></td>
<td>0.682</td>
</tr>
</tbody>
</table>

where \( N_M \) is the rotor-to-stator turns ratio and \( I_{rms} \) is the rms current rating of the individual windings. \( I_r \) quantifies the sharing of the magnetizing current for a specific DFM design.

In this section, the rotor power converter design procedure for a practical switched-DFM drive will be illustrated using parameters from an off-the-shelf DFM (1 HP, 220 V/150 V, 60 Hz, 4 Pole, 1800 rpm) as given in Table 2.2. The ac source is considered to be 220 V, 60 Hz for this analysis. The mechanical load for the drive is assumed to be a ship propeller implying that the load torque is proportional to the square of the rotor speed. Therefore, the rated torque of the DFM is fully utilized at the maximum rotor speed.

The goal is to minimize the rotor power electronics requirement while the machine is operated within rated conditions along with a desired drive torque-speed capability. The sizing of the rotor power electronics starts with selecting the current rating of the rotor converter such that the candidate DFM is utilized to its maximum capability in high-speed mode. Once the current rating is resolved, the stator flux in low-speed mode is determined by the low-speed drive-torque requirement. Minimizing the stator flux in low-speed mode directly impacts the rotor power electronics voltage rating by affecting the slope of the required rotor voltage as highlighted in Fig. 2.1. For the LSS topology, a choice of the stator-flux magnitude in low-speed mode also sets the required dc source voltage. Finally, the transition speed and the maximum speed are chosen to minimize the voltage rating and, therefore, the power rating of rotor power converter. Additionally, the limits on the d-axis and q-axis rotor currents are computed for both the operating modes to ensure that the DFM operates within the rated limits of stator and rotor across the entire speed range.

2.5.1 High-speed mode: Choice of rotor-side converter current rating

Due to the possibility of different current ratings in the stator and rotor in a practical DFM, the rotor power electronics current rating must be chosen such that the machine stator and rotor current limits are not exceeded. The objective is to utilize the candidate DFM to its maximum torque capability in high-speed mode considering a quadratic load torque-speed characteristic. Assuming that the stator is connected to a
stiff ac voltage source implies

\[ v_{sd}^2 + v_{sq}^2 = 1. \] (2.27)

As \( r_s \) in a typical machine is small, (2.27) is simplified using (2.7), (2.11), and (2.12) to compute the stator flux in high-speed mode and is given by

\[ \psi_s \bigg|_{\text{high}} = 1 + \frac{r_s x_m}{x_s} i_{rq}. \] (2.28)

The rotor current limit requires that

\[ i_{rd}^2 + i_{rq}^2 \leq I_r^2. \] (2.29)

Similarly, the stator current limit necessitates

\[ i_{sd}^2 + i_{sq}^2 \leq 1. \] (2.30)

Using (2.6) and (2.7), the stator currents can be substituted with rotor current components in (2.30) resulting in

\[ \left( \frac{\psi_s \bigg|_{\text{high}} - x_m i_{rd}}{x_s} \right)^2 + \left( \frac{-x_m i_{rq}}{x_s} \right)^2 \leq 1. \] (2.31)

Constraints (2.29) and (2.31) are plotted in rotor d-q current plane as shown in Fig. 2.5. The intersection area of the two constraints sets the safe operating current limits in high-speed mode. The maximum torque capability of the DFM is achieved when maximum negative q-axis current is provided to the rotor of the DFM within the safe operating area. In the example DFM, the maximum torque capability is achieved when the rotor q-axis current is identical to rotor current rating while the rotor d-axis current is zero. The stator d-axis current supports the entire stator flux. Therefore, the rotor power electronics current rating must be equal to the current rating of the DFM rotor winding.

### 2.5.2 High-speed mode: Maximum torque capability of the candidate DFM

With the maximum q-axis rotor current being \( I_r \), the maximum torque capability of the candidate DFM is computed using (2.5) and (2.28) and is given by,
Figure 2.5: High-speed mode: Intersection of the stator and rotor current limits for the practical DFM sets the safe operating area in high-speed mode. For the candidate DFM, current limit of the rotor winding specifies the rotor-side power converter current rating.

\[
\tau_{\text{max}} = \frac{x_m}{x_s} \left( 1 - \frac{r_x x_m}{x_s} I_r \right) I_r.
\] (2.32)

If the drive torque requirement at the maximum rotor speed is lower than \(\tau_{\text{max}}\), the candidate DFM will be an oversize machine, left under-utilized from the torque capability perspective. The "extra" capability may be utilized for reactive power control. If the torque requirement is higher than \(\tau_{\text{max}}\), an alternate DFM design must be selected. For the example DFM, \(\tau_{\text{max}}\) is computed to be 0.663 p.u. Figure 2.6 shows the drooping behavior of the stator flux magnitude relative to the drive torque demand due to the non-zero resistances and leakages as computed using (2.5) and (2.28).

### 2.5.3 High-speed mode: Choice of rotor d-axis current limit

In high-speed mode, a possible secondary objective for the drive is to achieve controllable power factor on the stator. For example, any "extra" or spare rotor inverter and DFM capability may be used to provide reactive power support to the ac source of the ship micro-grid. This is an exciting opportunity to allow the drive to assist with overall power system stability, and an opportunity that may make the switched-DFM drive attractive in many other land-based applications besides propulsion. The limit of the rotor d-axis current rating, which affects stator power factor, can be obtained from the stator and the rotor current rating. From (2.28), (2.29), and (2.31) for a particular q-axis rotor current, the upper limit for the d-axis rotor current is
Figure 2.6: High-speed mode: Droop characteristic of the stator flux magnitude relative to the drive torque due to the presence of stator resistance and leakages for the example DFM.

\[ i_{rd, \text{max}} = \min \left[ \sqrt{I_r^2 - i_{r q}^2} \left( \frac{1}{x_m} \frac{r_s}{x_s} + \frac{x_s}{x_m} \sqrt{1 - \left( \frac{x_m i_{r q}}{x_s} \right)^2} \right) \right] \]  

while the lower limit is given by

\[ i_{rd, \text{min}} = \max \left[ \sqrt{I_r^2 - i_{r q}^2} \left( \frac{1}{x_m} \frac{r_s}{x_s} - \frac{x_s}{x_m} \sqrt{1 - \left( \frac{x_m i_{r q}}{x_s} \right)^2} \right) \right]. \]  

For the parameters of the example DFM, the stator current limit imposes a minimum d-axis rotor current of negative value, which does not provide any benefit in high-speed mode unless the DFM has to operate as a reactive power sink. The minimum d-axis rotor current is, therefore, chosen to be zero. The safe operating area for the rotor current is shown by the filled-half-circle in Fig. 2.5.

2.5.4 Low-speed mode in LSS topology: Choice of stator flux magnitude and dc source voltage

The design process continues with selection of stator flux magnitude in low-speed mode based on the required drive torque capability. This directly influences the to-be-selected rotor power electronics voltage rating by affecting the slope of the required rotor voltage curve relative to the operating speed in low-speed mode. As an example, the required drive torque in low-speed mode is chosen as 75% of the maximum torque capability in high-speed mode (equivalent to 0.498 p.u. for the example DFM). Assuming that the drive is configured in LSS topology, the positive terminal of the dc source is connected to the stator A-phase while the negative terminal is connected to the B and C phases. Figure 2.7 shows the relative orientation of the stator flux, voltage, and current space vectors in the stator ABC winding reference frame.

Denoting the angle between the stator voltage and the stator flux vector as \( \delta \), the drive torque in low-speed
Figure 2.7: Low-speed mode in LSS topology: Steady-state orientation of the stator flux, voltage, and current space vectors in the stator ABC winding reference frame.

mode can be expressed using (2.5) and (2.7) as,

\[ \tau_{low} = i_s \psi_{low} \sin \delta \]  

(2.35)

where \( i_s = \frac{v_{dc}}{r_s} \).

The required torque in low-speed mode should be achieved at minimum stator flux magnitude, which implies that the stator current and the angle \( \delta \) needs to be maximized within allowable limits. The upper bound on the angle \( \delta \) for steady operation is 90° and the upper bound on the stator current is determined by two constraints. First, the stator current rating of the machine must not be exceeded. Second, the reflected rotor current must not exceed the power electronics and the DFM current ratings both in the steady-state and during transients. These bounds can be simultaneously mapped on a \( \delta - i_s \) plane as shown in Fig. 2.8.

To keep the machine stator current within its r.m.s rating (since the stator current is dc),

\[ i_s \leq \frac{1}{\sqrt{2}} \]  

(2.36)

Using (2.7) and (2.35), the q-axis rotor current is expressed as

\[ i_{rq} = -\frac{x_s}{x_m} i_s \sin \delta \]  

(2.37)
Similarly (2.6) is rearranged in low-speed mode resulting in

\[ i_{rd} = \frac{\psi_s}{x_m} - \frac{x_s i_s \cos \delta}{x_m}, \]  

(2.38)

At steady-state, the stator voltage vector, current vector, and the A-phase axis coincides as shown in Fig. 2.7. Therefore, to keep the rotor current within its rated limit at steady state (2.35), (2.37), and (2.38) are combined to obtain

\[ i_r = \sqrt{\left(\frac{x_s}{x_m} i_s \sin \delta\right)^2 + \left(\frac{\psi_s}{x_m} - \frac{x_s i_s \cos \delta}{x_m}\right)^2} \leq I_r. \]  

(2.39)

The rotor current must also be within its rated limit during transient torque demand. A step change in torque demand of \( \tau \) from zero can be considered as the worst case scenario in this regard. Rearranging (2.4),

\[ v_{sq} = r_s v_s + \frac{1}{x_m} \left(\frac{1}{\omega_B} \frac{d\delta}{dt}\right). \]  

(2.40)

With zero torque demand initially, the angle \( \delta \) is zero implying the stator flux and dc source voltage vectors are aligned along the stator A-phase axis. With the stator current vector aligned to the dc source voltage vector implies

\[ i_{sd} = \frac{v_{dc}}{r_s} = i_s. \]  

(2.41)

Using (2.6) and (2.41), the required rotor d-axis current to maintain the stator flux constant is obtained as

\[ i_{rd} = \frac{\psi_s}{x_m} - \frac{x_s i_s}{x_m}. \]  

(2.42)

Equation (2.40) also suggests that the angle \( \delta \) does not change instantaneously with change in torque demand. A step demand in torque of \( \tau \) will require a q-axis rotor current from (2.5) as,

\[ i_{rq} = -\frac{x_s \tau}{x_m \psi_s}. \]  

(2.43)

Therefore, using (2.35) in (2.42) and (2.43), the dynamic limit on the d-axis and q-axis rotor currents is formulated as,
Equations (2.36), (2.39), and (2.44) set the allowable stator current limit for the required low-speed mode torque as shown in Fig. 2.8.

The minimum low-speed mode stator flux can be computed from (2.35) by framing as a minimization problem given by

$$
\text{minimize } \psi_s \bigg|_{\text{low}} (i_s, \delta)
$$

subject to:

$$
\tau \bigg|_{\text{low}} = i_s \psi_s \bigg|_{\text{low}} \sin \delta,
$$
$$
\delta < 90^\circ,
$$
$$
i_s(\delta) \text{ bounded by (2.36), (2.39), and (2.44)}.
$$

The minimum stator flux magnitude is obtained at the angle $\delta_{\text{max}}$ as shown in Fig. 2.9. The minimum stator flux obtained for the example DFM is nearly 0.75 p.u. The angle $\delta_{\text{max}}$ and the obtained minimum stator flux magnitude is used to compute the required dc source voltage using (2.35). Figure 2.10 shows that the DFM stator and rotor currents that are within the machine ratings for every operating point within the allowable $\delta_{\text{max}}$. The rotor d-axis and q-axis current limits in low-speed mode are different than in high-speed mode. This must be appropriately accounted in the drive controller.
Figure 2.9: The maximum operating $\delta_{\text{max}}$ is chosen such that stator flux in low-speed mode is minimized to reduce the rotor power electronics voltage rating.

### 2.5.5 Low-speed mode in LSI topology: Choice of stator flux magnitude

In the LSI topology, the stator flux is entirely created by the rotor magnetizing current in low-speed mode.

The rotor $d$-axis current required to establish the stator flux $\psi_s|_{\text{low}}$ is given by

$$i_{rd} = \frac{\psi_s|_{\text{low}}}{x_m}.$$  \hfill (2.45)

As in the LSS topology, the DFM must operate within the constraints of the allowable stator and the rotor currents, which can be formulated from (2.29), (2.30), and (2.7) as:

$$\left(\frac{z_s\tau_{\text{max}}}{x_m} i_{rq}\right)^2 + i_{rq}^2 \leq I_r^2,$$

and

$$i_{rq}^2 \leq \left(\frac{x_m}{x_m}\right)^2.$$  \hfill (2.46)

The minimization function for the operating stator flux magnitude in LSI topology is, therefore, obtained as

$$\text{minimize } \psi_s|_{\text{low}}(i_{rq})$$

subject to:

$$\tau|_{\text{low}} = -\frac{z_m}{x_s} i_{rq} \psi_s|_{\text{low}}, \text{ and}$$

$i_{rq}$ is bounded by (2.46).

### 2.5.6 Choice of rotor-side converter voltage rating: Computation of transition speed and maximum speed

With the minimized stator flux in low-speed mode and the rotor current in high-speed mode, (2.11)-(2.14) and (2.6)-(2.9) are used to compute the rotor voltage requirement for the entire operating speed range. For
the LSS topology, this is plotted in Fig. 2.11. Comparing with the rotor voltage requirement of the ideal DFM in Fig. 2.1, there are notable differences due to presence of non-idealities in the DFM. Not only that non-zero rotor voltages are required at zero and unity speed, but also the magnitude depends on the polarity of the demanded electromagnetic/drive torque. This implies a different rotor voltage requirement during acceleration and deceleration for the drive load. To minimize required rotor voltage to operate the drive under all operating conditions, a mode transition speed is chosen such that required rotor voltage in low-speed mode for maximum positive drive torque equals that in high-speed mode for maximum negative drive torque. The maximum operating speed of the drive can be obtained based on the chosen minimum rotor voltage and voltage requirement in high-speed mode for the maximum positive drive torque. For the example DFM, the required rotor power electronics voltage rating is 0.52 p.u. while operating on a speed range of 0-1.49 p.u. The sharing of total active power between the stator and rotor for the non-ideal DFM is shown in Fig. 2.12. The maximum power being handled by the rotor power electronics is 0.39 p.u. while the maximum total active power fed to the DFM is 1.13 p.u. Therefore, a rotor power electronics of one-third power rating can be used to control the example DFM over the complete speed range. Depending on the efficiency of the DFM and the mechanical drive train, the net output shaft power is determined. Figure 2.13 shows the steady-state torque-speed capability of the example DFM with the proposed drive architecture. A typical propulsion load torque is also shown that is proportional to the square of rotor speed and matches the maximum torque capability of the DFM at the maximum speed. In general, the procedure can be used for designing the switched-DFM drive with different drive torque capability requirement in low speed. Figure
2.14 shows the effect of different low-speed mode torque capability requirement on the rotor converter size and operating drive speed for the example DFM. The effect is similar to an ideal DFM as shown in Fig. 2.4.

Assuming the mechanical load for the switched-DFM drive to be ship propeller, the steady-state load torque requirement is proportional to the square of the shaft speed. Based on mechanical requirements, the maximum commandable shaft torque required in low-speed mode can be either the same as or lower than the rated drive torque in high-speed mode.

2.6 Experimental results

Experiments have been performed on the example DFM to illustrate the sizing benefit of the rotor power electronics for the proposed switched-DFM drive. The laboratory setup comprises an ac grid of 146 V (line-to-line, rms), 40 Hz that is created using two synchronous generators operating in parallel as described in details in Appendix A. The choice of the ac grid frequency ensures that the off-the-shelf example DFM remains within the rated operating speed of 1800 rpm even when running at super-synchronous speed. Correspondingly, the ac source voltage magnitude is adjusted to ensure a rated stator flux operation of the DFM in high-speed mode.

The experimental switched-DFM drive is designed to achieve a low-speed torque capability of 70% of that of high-speed torque capability. This torque requirement in low-speed mode translates to a selection of stator flux magnitude of 0.36 V-s at low speed in the LSI topology while a stator flux magnitude of 0.32 V-s
Figure 2.12: Example DFM in LSS topology: One-third rotor power electronics power rating controls the net active power flowing to the DFM while operating on a speed range of 0-1.49 p.u.

Figure 2.13: Designed drive torque-speed capability of the example DFM using the proposed architecture with a typical propulsion load torque.
at low speed in the LSS topology. The mode-transition speed is designed to be 775 rpm for the LSS topology and 615 rpm for the LSI topology each with a hysteresis band of ±18 rpm around the mode transition speed to ensure a chatter-free mode transition. The dc electronic load bank is programmed to achieve a quadratic load torque profile with respect to the rotor speed. Individual experimental results are shown for the two topologies in the following sections.

### 2.6.1 Switched-DFM drive operating in low-speed synchronous topology

With the DFM initially running at 1800 rpm, the drive is subjected to a step-command in reference speed of −1800 rpm (at A in Fig. 2.15) and back to 1800 rpm (at B) to evaluate the full-speed range operation and the requirement on the rotor power converter. A solid-state transfer switch, described in Chapter 3, ensures an on-the-fly reconfiguration of the stator winding of the DFM while a unique controller, described in Chapter 4, ensures a smooth operation of the proposed drive in and through the two modes. For this entire operating speed range, the DFM initially operates in high-speed mode with stator connected to the ac grid through ABC phase sequence. As the rotor speed goes below 775 rpm, the DFM operates in low-speed mode with the DFM stator connected to the dc source of 20 V. Finally, the DFM operates back in high-speed mode but with the stator connected to the ac grid in ACB phase sequence when the rotor speed goes below −775 rpm. From A to B in Fig. 2.15, the DFM operates with a negative drive torque command while from B to C, the DFM operates with positive drive torque command.
Figure 2.15: Experimental result: Speed response of the switched-DFM drive for a step command in speed in the LSS topology.

The torque-speed capability of the switched-DFM drive during the complete four-quadrant operation is shown in Fig. 2.16. The load torque is also shown in the figure, which is programmed as proportional to the quadratic rotor speed. The presence of static frictional torque in the drive shaft at zero speed results in appearance of the discontinuity in the load torque-speed characteristic. As designed, the low-speed mode torque limit is observed as 70% to that of high speed torque mode torque limit.

Figure 2.17 shows the required rotor voltages in the LSS topology during the DFM operation with the positive drive torque (i.e. from B to C in Fig. 2.15). The operating rotor voltage profile is a mirror image along y-axis for negative drive torque requirement. Changing the operating modes of the DFM at correct transition speeds ensures that the required rotor voltage is always less than 58 V while the DFM drive operates over the entire speed range.

The sharing of active power between the stator and rotor of the DFM in the LSS topology with the positive drive torque (i.e. from B to C in Fig. 2.15) is shown in Fig. 2.18. A power of 60 W is provided by the separate dc source to supply the losses in the stator winding during low-speed operation. The rotor power electronics processes 416 W of power while the total active power fed both by the stator and the rotor ports of the DFM is 1100 W at the maximum operating speed. Assuming negligible losses, the rotor power electronics rating is, therefore, 38% of the total active power available to the mechanical port while operating on a speed range of ±1800 rpm.

Finally, Fig. 2.19 shows the stator and rotor flux frequencies during the full speed range operation. The stator flux frequency is zero in low-speed mode while the rotor flux frequency is equal and negative of the
Figure 2.16: Experimental result: Four-quadrant drive torque-speed capability of the designed switched-DFM drive for a quadratic load torque-speed characteristic. Low-speed mode has 70% torque capability to that of high-speed mode by design.

Figure 2.17: Experimental result: Measured rotor power electronics voltage command during acceleration (full positive torque) relative to rotor speed.
rotor speed. In high-speed modes, the stator flux frequency is either 40 Hz or -40 Hz depending on whether the stator is connected to the ac grid in $ABC$ phase sequence or $ACB$ phase sequence. Again, the difference between the stator flux speed and the rotor flux speed in high-speed mode equals the rotor speed suggesting continuous steady torque creation for the entire operating speed range.

2.6.2 Switched-DFM drive operation in low-speed induction topology

The switched-DFM drive operation in the low-speed induction topology is evaluated next. Figure 2.20 shows the speed performance in the LSI topology, which is identical to the LSS topology because of an equal drive torque capability design. The mode transition speed is designed as 615 rpm. The drive torque capability in the LSI topology during this full-speed range operation is shown in Fig. 2.21 which is almost similar to that in LSS topology with minor differences due to the designed mode transition speeds. For the example DFM, the required rotor voltage in the LSI topology is 18% higher compare to the LSS topology due to higher operating stator flux level as shown in Fig. 2.22. The maximum power supplied by the rotor power electronics is 5% higher in the LSI topology as compare to the LSS topology that accounts for the losses in the stator of the DFM as shown in Fig. 2.23. Unlike in the LSS topology, the stator flux frequency in the LSI topology is dependent on the drive torque demand. For example, for the maximum positive drive torque demand in low-speed mode for the example DFM, the corresponding stator flux frequency is $-5$ Hz.
Figure 2.19: Experimental result: The difference between the stator and the rotor flux frequencies must be equal to the rotor speed for steady torque creation in the DFM irrespective of the operating mode. In low-speed mode, the stator flux frequency is zero while in high-speed mode the stator flux frequency is either 40 Hz or -40 Hz based on stator phase sequence connection. The rotor flux frequency is appropriately commanded by the rotor power electronics to ensure seamless rotor speed irrespective of the operating mode.

as shown in Fig. 2.24. This represents the slip frequency of the DFM that mimics an inside-out induction machine in this particular mode.

2.7 Summary

This chapter discussed a design process for different topologies of switched-DFM drives. Two mode operation permits full speed range VSD when an ac utility is available. While high-speed mode operation impacts the current rating, low-speed mode operation impacts the voltage rating of the rotor-side power converter. Given a DFM and drive torque requirements, the rotor power electronics can be minimized with proper choice of transition speed between modes and maximum desirable speed. Although an exemplary design has been shown with a propulsion load torque profile, the design can be iterated for other desirable load torque profiles like traction, which requires higher torque in low speed and lower torque in high speed. The drive architecture also provides flexibility of introducing a third mode—“high speed dc” mode—when the ac source connection to the stator is switched back to the dc source or shorted configuration but operated with a lower effective stator flux magnitude similar to a field-weakening operation in a standard drive.
Figure 2.20: Experimental result: Speed response of the switched-DFM drive for a step command in speed in the LSI topology.

Figure 2.21: Experimental result: Four-quadrant drive torque-speed capability of the designed switched-DFM drive for a quadratic load torque-speed characteristic. Low-speed mode has 70% torque capability to that of high-speed mode by design.
Figure 2.22: Experimental result: Measured rotor power electronics voltage command during acceleration (full positive torque) relative to rotor speed in the LSI topology.

Figure 2.23: Experimental result: The rotor power electronics processes a maximum of 437 W of active power while the total active power fed both by the stator and the rotor ports of the DFM is 1100 W at the maximum operating speed. Neglecting losses, the rotor power electronics rating is, therefore, 40% of the total active power available to the mechanical port while operating on a speed range of ±1800 rpm.
Figure 2.24: Experimental result: The difference between the stator and rotor flux frequencies must be equal to the rotor speed for steady torque creation in the DFM irrespective of the operating mode. In low-speed mode, the stator flux frequency is dependent on the drive torque demand (negative in this case for a maximum positive drive torque) while in high-speed mode the stator flux frequency is either 40 Hz or -40 Hz based on stator phase sequence connection. The rotor flux frequency is appropriately commanded by the rotor power electronics to ensure seamless rotor speed irrespective of the operating mode.
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Chapter 3

Transfer Switch Design

A critical challenge for practically realizing a switched-DFM drive is in developing an effective switching circuit that can connect the stator to either of the sources, dc or ac, or short all the windings together on-the-fly. A complex switching circuit in the DFM stator made with self-commutating devices like IGCTs, GTOs, IGBTs counters the benefit of reduced power electronics requirement achieved in the rotor-side converter [36]. Alternatively, sluggish mechanical switches [37] controlling the stator connection can significantly impact the shaft behavior during the mode transition, leading to poorly controlled speed/torque variations [23], and even forcing the switched-DFM drive to operate in two isolated stages - the starting stage and the operating stage [26]. Using thyristors (SCRs) for the stator winding reconfiguration can provide a balance between the affordable complexity and the required performance. SCRs are still some of the most capable high-power devices available [10]. A single device can handle large currents and block large voltages. In fact, SCR-based static-transfer-switches (STS) have been used extensively to supply uninterrupted power to critical loads [38, 39, 40, 41, 42]. A traditional operational scheme in these STS is detecting a disturbance in the primary ac source, followed by a transitioning of the load to an alternate ac source, both of the sources being of identical voltage and frequency.

However, for the proposed switched-DFM drive, the transfer occurs between an ac and a dc source or a shorted condition with widely different voltage magnitudes, frequencies, variable operating load power factor, and load regeneration capability. This is a challenge because the two sources cannot be “synchronized” [43] for transfer. Additionally, the load, in this case the DFM, cannot withstand partial phase transfer. For example, a partial phase transfer occurs when the three phases of the DFM stator are not connected to one specific source at any given instant of time [44]. Moreover, the transfer operation between the low-speed mode and high-speed mode is triggered based on the operating speed of the DFM and the designed mode-
transition speed, making the source-transfer mechanism integral to the switched-DFM drive operation rather than just as a means to provide occasional back up power.

This chapter presents multiple SCR-based static-transfer-switch topologies that connect the stator of the proposed DFM drive to either the ac or the dc source or short the stator winding together on-the-fly with well-controlled speed/torque variations at the mechanical shaft output. The analysis begins with a review of the ideal requirements for a transfer of the stator of the DFM between two sources. Conditions are identified in an ac cycle such that all the three phases can undergo simultaneous natural commutation during the source transition because SCRs require special considerations to ensure device turn-off. Two sets of SCRs are used to connect the individual sources to the stator of the DFM. Only one set is turned on during each operating mode of the DFM drive to connect the corresponding source to the stator. In the LSS topology, the dc-source-side SCRs are continuously on when the DFM drive is operating in low-speed mode, while in the LSI topology the shorted-side SCRs operate at the slip frequency. In both topologies, the ac-source-side SCRs switch at line frequency during the high-speed mode of operation of the DFM drive. This results in ideally zero switching losses in the overall transfer switch during steady-state operation without increasing the harmonic distortions of the stator waveforms.

The development of an STS suitable for the proposed DFM drive is focused on two different aspects. First, appropriate conditions based on the DFM stator current and source voltages are derived that ensure natural commutation of all outgoing SCRs simultaneously during both the low-speed-to-high-speed and high-speed-to-low-speed mode transitions. Next, these conditions are superimposed with the DFM requirements to ensure bumpless mechanical behavior at the shaft as will be shown in Chapter 4.

3.1 Requirements for an ideal transfer switch

A schematic of an ideal transfer switch connected to the DFM is shown in Fig. 3.1. The DFM stator is connected to different sources based on the topology and operating mode by appropriate configuration of switch $S_1$ and $S_2$. It is assumed that the dc and the ac sources share a common reference. In practice, the dc source may be created from the ac source using a transformer-rectifier. In low-speed mode, the dc source only supplies the resistive losses in the stator of the DFM. Consequently, the power rating of the transformer-rectifier is a fraction of the DFM power rating.

Several characteristics of the transfer switch operation are essential:

- The ac and dc sources should not be shorted; i.e., $S_1$ and $S_2$ cannot be ON simultaneously.

- The DFM stator current should not be interrupted instantaneously; i.e., $S_1$ and $S_2$ cannot be both
Figure 3.1: Ideal transfer switch connected to the DFM stator winding to alter the connection among ac source, dc source (LSS), or short (LSI) as necessary, based on operating speed.

*OFF* abruptly with a non-zero DFM stator current.

- The three phases should always switch together; i.e., the three phases of the stator are connected to a single source, dc, ac or short, at any instant. This ensures that the DFM does not experience severe unbalanced stator voltages that can lead to unacceptable disturbances in the stator flux.

A few other characteristics are also necessary:

- The DFM should experience minimal perturbation in the drive torque and rotor speed during the source transition.

- Additional "supporting" circuitry, e.g., forced commutation circuit, should be minimized to limit cost and complexity [45].

### 3.2 Single cell analysis for SCR-based transfer switch commutation requirement

As a precursor, a simplified SCR-based transfer switch is evaluated for its commutation requirements. These familiar results for a constant current load are essential for extending the analysis to three phase sources of different form, ac and dc, or even under shorted condition with the DFM as the load. Figure 3.2 shows a single cell of an SCR-based transfer switch connected to a constant current load. The load can be connected
to arbitrary voltage sources $V_X$ and $V_Y$ reference to a common ground. Assuming initially the load is connected to source $V_X$, a transition to source $V_Y$ is desired. Four well-known commutation scenarios arise based on the load current polarity and the relative magnitudes of $V_X$ and $V_Y$.

1. $V_Y > V_X$ and $I_L > 0$: As the load current is positive, SCRs $T_{XR}$ and $T_{YR}$ are non-functional and their gate control is permanently disabled. Load current initially flowing in $T_{XF}$ naturally commutates to $T_{YF}$ when the gate control for $T_{YF}$ is enabled. $T_{XF}$ is automatically reverse biased, as $V_L$ is now pinned to $V_Y$, and is turned ‘OFF’. No additional commutation circuit is required for this transition.

2. $V_Y < V_X$ and $I_L > 0$: Switching in this case would require a forced commutation circuit. $T_{YF}$ is reverse biased and will not “naturally” pick up the load current even if its gate signal is enabled. This case corresponds to an undesirable region for attempting a stator transition in the DFM.

3. $V_Y < V_X$ and $I_L < 0$: As the load current is negative, SCRs $T_{XF}$ and $T_{YF}$ are non-functional and their gate control is permanently disabled. Similar to Case 1, enabling gate signal $T_{YR}$ can naturally commutate the load current from $T_{XR}$ to $T_{YR}$. No additional commutation circuit is required for this transition.

4. $V_Y > V_X$ and $I_L < 0$: Similar to Case 2, additional forced commutation circuit is required to turn $T_{XR}$ OFF and forms an undesirable region for source transition for the DFM.

### 3.3 Twelve-thyristor-based transfer switch

The single cell analysis is extended for a DFM load with three phase source connection. As switch $S_1$ in Fig. 3.1 should block bidirectional voltage and carry bidirectional current, an anti-parallel configuration of SCRs for each phase is required similar to a standard STS. Switch $S_2$ appears to carry only unidirectional current and to block bidirectional voltage, so a single SCR for each phase is sufficient in the LSS topology. However, the DFM stator-current polarity can reverse during load transients in low-speed mode, and also during the
Figure 3.3: Twelve thyristor-based transfer switch (TTB) - The neutral of the ac source voltage is the reference for the transfer switch operation.

The ac-to-dc source transition. Therefore, an anti-parallel configuration of the SCRs for each phase is required to replace the ideal switch $S_2$. These considerations give rise to a twelve-thyristor-based (TTB) transfer switch as shown in Fig. 3.3. One of the sources is either a dc source of magnitude $V_{dc}$ for the LSS topology or a short for the LSI topology, and the other is a three-phase ac source represented by phase voltages $V_a$, $V_b$, and $V_c$.

In high-speed mode, SCRs $T^A_{high,R}$, $T^A_{high,F}$, $T^B_{high,F}$, $T^B_{high,R}$, $T^C_{high,F}$, and $T^C_{high,R}$ are turned ON thereby connecting the DFM stator to the ac source. These six SCRs form the high-speed SCR-bank. Each of these SCRs conduct for half of the time period corresponding to the ac source fundamental frequency. Alternatively, in low-speed mode, SCRs $T^A_{low,F}$, $T^A_{low,R}$, $T^B_{low,F}$, $T^B_{low,R}$, $T^C_{low,F}$, and $T^C_{low,R}$ are turned ON either to connect the stator to the dc source in the LSS topology or to short them together in the LSI topology. These six SCRs are collectively termed as the low-speed SCR-bank.

### 3.3.1 LSS topology: Commutation requirement during low-speed-to-high-speed mode transition

Assuming initially the stator of the DFM is connected to the dc source, at steady state, the stator current is positive in the $A$ phase, and negative in the $B$ and $C$ phases. This implies that SCRs $T^A_{low,F}$, $T^B_{low,R}$, and $T^C_{low,R}$ are conducting. These three SCRs are classified as the “conducting bank.” The complementary SCRs $T^A_{low,R}$, $T^B_{low,F}$, and $T^C_{low,F}$, classified as the “non-conducting bank”, do not take part during this transition and the gate signals to these SCRs are disabled at the onset of the transition. A commutation diagram can be established for all the three phases using the stator $ABC$-phase winding axis as reference, as shown in Fig.
3.4. The dc-source voltage magnitude, shown as a dashed line, is relatively small compared to the ac-source voltage. Before the transition, the stator voltage vector $\overline{V}_{dc}$ (where $|\overline{V}_{dc}| = \frac{2}{3} V_{dc}$) and the current vector $\overline{I}$ are both stationary, aligned, and oriented towards $A$-phase axis, while the incoming ac source voltage vector $\overline{V}_{grid}$ (where $|\overline{V}_{grid}| = \text{peak-phase voltage of the ac source}$) is rotating at the ac-source frequency, $\omega_{ac}$.

For $A$-phase, turning off the gate signal to SCR $T_{low,F}^{A}$ and firing SCR $T_{high,F}^{A}$ when the ac-source $A$-phase voltage is greater than the dc-source voltage naturally steers the stator current from the dc to the ac source. This implies that the ac-source voltage vector must be within the arc $PP'$ in Fig. 3.4 to ensure natural commutation of SCR $T_{low,F}^{A}$. Similarly, for the $B$ and $C$ phases, SCRs $T_{low,R}^{B}$ and $T_{low,R}^{C}$ are turned off naturally by deactivating the gate signal to these SCRs and firing gate pulses to SCRs $T_{high,R}^{B}$ and $T_{high,R}^{C}$ during the negative half cycle of the $B$ and $C$ phase ac-source voltages, respectively. These conditions are met when the ac-source voltage vector is within the arc $QQ'$ and $RR'$ respectively, in the commutation diagram of Fig. 3.4. Different sectors evolve in the commutation diagram, which demarcates possible commutation for the outgoing three-phase SCRs. The sector of $\pm 30^\circ$ where all the three arcs overlap indicates the special region for the ac-source voltage vector where natural commutation can occur for the dc-source-side conducting bank.

After the source transition, SCRs $T_{high,F}^{A}$, $T_{high,R}^{B}$, and $T_{high,R}^{C}$ continue to supply the DFM stator current. These SCRs form the “succeeding bank.” With a sufficient dead time, depending on the turn-off time of the SCRs, the gate signals to the remaining SCRs on the ac-source side $T_{high,R}^{A}$, $T_{high,F}^{B}$, and $T_{high,F}^{C}$, collectively named as “concluding bank,” are enabled. This completes the transfer of the stator of the DFM to the ac source.

### 3.3.2 LSS topology: Commutation requirement during high-speed-to-low-speed mode transition

Conversely, when the stator of the DFM is initially connected to the ac source, all ac-source-side SCRs $T_{high,F}^{A}$, $T_{high,R}^{A}$, $T_{high,F}^{B}$, $T_{high,R}^{B}$, $T_{high,F}^{C}$, and $T_{high,R}^{C}$ operate on $180^\circ$ conduction mode depending on the stator voltage and current. This implies that three of the six SCRs conduct at any instant including the instant of mode transition, which constitutes the conducting bank. The remaining three complementary SCRs, forming the non-conducting bank, do not influence the transition and the gate signals to this bank can be turned off. On the dc-source side, three SCRs (succeeding bank) turn on at the instant of transition for each phase commutating the current from the ac source to the dc source. The remaining three dc-source side SCRs (concluding bank) are turned on after a sufficient dead time to avoid shoot-through between the ac and dc sources. Unlike in the dc-to-ac source transition, the SCRs that constitute these four banks
depend on the stator-current polarity at the instant of transition. Consequently, the conditions of the source voltages required for natural commutation of the conducting bank during ac-to-dc source transition need to be determined based on the stator-current polarity.

For example, when the stator-current vector $\overrightarrow{I}$ is in sector of $\pm 30^\circ$ in Fig. 3.5, SCRs $T^A_{high,F}$, $T^B_{high,R}$, and $T^C_{high,R}$ constitute the conducting bank while SCRs $T^A_{high,R}$, $T^B_{high,F}$, and $T^C_{high,F}$ form the non-conducting bank. During the period when the ac source A-phase voltage is less than the dc source voltage, turning off the gate signal to SCR $T^A_{high,F}$ and firing SCR $T^A_{low,F}$ will naturally commutate the stator A-phase current from the ac source to dc source. This condition is met when the ac source voltage vector is within the arc $PP'$. Similarly for the B and C phases, SCRs $T^B_{high,R}$, and $T^C_{high,R}$ can naturally commutate the stator current to SCRs $T^B_{low,R}$, and $T^C_{low,R}$ at the instant when the ac source B and C phase voltages are positive shown by arcs $QQ'$ and $RR'$ respectively. In this case, SCRs $T^A_{low,F}$, $T^B_{low,R}$, and $T^C_{low,R}$ forms the succeeding bank. To ensure all the three phases undergo natural commutation at the same instant, the ac-source voltage vector must be within $150^\circ$ and $210^\circ$ where the arc $PP'$, $QQ'$, and $RR'$ overlap as shown in Fig. 3.5. Finally, the complementary dc-source side SCRs $T^A_{low,F}$, $T^B_{low,R}$, and $T^C_{low,R}$ in the concluding bank are provided with gate pulses after a sufficient dead time to complete the source transition.

This analysis can be extended for other possible polarities of the three-phase stator current such that natural commutation of the outgoing ac-source side SCRs are ensured. The sectors of natural and forced commutation for each phase also rotate along with the stator-current vector location. Table 3.1 summarizes
the SCRIs that form the conducting and succeeding banks at the instant of transition depending on the stator-current polarity. Also, this table highlights the requirement on the ac source voltage vector $V_{grid}$ to make simultaneous natural commutation of the conduction bank SCRs. The angle $\varepsilon$ in the Table 3.1 depends upon the magnitudes of the dc and the ac source voltage vector and can be calculated as,

$$\varepsilon = \arccos\left(\frac{V_{dc}}{|V_{grid}|}\right)$$  \hspace{1cm} (3.1)

Based on Table 3.1, the common requirement for simultaneous natural commutation of the ac-source side SCRs is that the stator power factor angle $\theta$ must be between 120° and 240° irrespective of the location of the stator current vector. The complementary region of natural commutation of the conduction bank SCRs between the two mode transitions, dc-to-ac in Fig. 3.4 and ac-to-dc in Fig. 3.5, is due to the relative magnitude of the source voltages, as expected.

### 3.3.3 LSI topology: Commutation requirement during mode transitions

In the LSI topology, the dc source of the LSS topology is replaced by a short as shown in Fig. 3.3. The stator winding carries a bidirectional current at slip frequency during low-speed mode implying that all SCRs in the low-speed-SCR bank $T_{low,F}^A$, $T_{low,R}^A$, $T_{low,F}^B$, $T_{low,R}^B$, $T_{low,F}^C$, and $T_{low,R}^C$ operate in 180° conduction mode.
Table 3.1: Classification of outgoing ac-source side and incoming dc-source side SCRs for a high-speed-to-
low-speed mode transition in the LSS topology based on the stator current vector location.

<table>
<thead>
<tr>
<th>Sector</th>
<th>Location of stator current vector</th>
<th>Conducting bank (ac-source side)</th>
<th>Succeeding bank (dc-source side)</th>
<th>Required ac-source voltage vector location</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>$-30^\circ &lt; \angle \bar{I}_s &lt; 30^\circ$</td>
<td>$T^A_{high,F}$, $T^B_{high,R}$, $T^C_{high,F}$</td>
<td>$T^A_{low,F}$, $T^B_{low,R}$, $T^C_{low,F}$</td>
<td>$150^\circ &lt; \angle V_{grid} &lt; 210^\circ$</td>
</tr>
<tr>
<td>II</td>
<td>$30^\circ &lt; \angle \bar{I}_s &lt; 90^\circ$</td>
<td>$T^A_{high,F}$, $T^B_{high,F}$, $T^C_{high,R}$</td>
<td>$T^A_{low,F}$, $T^B_{low,R}$, $T^C_{low,F}$</td>
<td>$210^\circ &lt; \angle V_{grid} &lt; 360^\circ - \varepsilon$</td>
</tr>
<tr>
<td>III</td>
<td>$90^\circ &lt; \angle \bar{I}_s &lt; 150^\circ$</td>
<td>$T^A_{high,R}$, $T^B_{high,F}$, $T^C_{high,F}$</td>
<td>$T^A_{low,R}$, $T^B_{low,F}$, $T^C_{low,F}$</td>
<td>$360^\circ - \varepsilon &lt; \angle V_{grid} &lt; 330^\circ$</td>
</tr>
<tr>
<td>IV</td>
<td>$150^\circ &lt; \angle \bar{I}_s &lt; 210^\circ$</td>
<td>$T^A_{high,R}$, $T^B_{high,F}$, $T^C_{high,F}$</td>
<td>$T^A_{low,R}$, $T^B_{low,F}$, $T^C_{low,F}$</td>
<td>$-30^\circ &lt; \angle V_{grid} &lt; 30^\circ$</td>
</tr>
<tr>
<td>V</td>
<td>$210^\circ &lt; \angle \bar{I}_s &lt; 270^\circ$</td>
<td>$T^A_{high,R}$, $T^B_{high,F}$, $T^C_{high,F}$</td>
<td>$T^A_{low,R}$, $T^B_{low,F}$, $T^C_{low,F}$</td>
<td>$30^\circ &lt; \angle V_{grid} &lt; \varepsilon$</td>
</tr>
<tr>
<td>VI</td>
<td>$270^\circ &lt; \angle \bar{I}_s &lt; 330^\circ$</td>
<td>$T^A_{high,R}$, $T^B_{high,F}$, $T^C_{high,F}$</td>
<td>$T^A_{low,R}$, $T^B_{low,F}$, $T^C_{low,F}$</td>
<td>$\varepsilon &lt; \angle V_{grid} &lt; 150^\circ$</td>
</tr>
</tbody>
</table>

The alternating current polarity opens up possibility of the SCR current commutation in the other sectors as well during a low-speed-to-high-speed mode transition. Extending the commutation analysis from the LSS topology as detailed in Sec. 3.3.1, the three outgoing SCRs (conducting bank) will have simultaneous natural commutation provided that the incoming ac source voltage vector is in the same sector as the stator current vector at the instant of mode transition. For example, with the stator current vector $\bar{I}_s$ in sector IV as shown in Fig. 3.6, the stator current from the SCRs $T^A_{low,R}$, $T^B_{low,F}$, and $T^C_{low,F}$ can be commutated to the SCRs $T^A_{high,R}$, $T^B_{high,F}$, and $T^C_{high,F}$ when the incoming ac source voltage vector $V_{grid}$ is also in sector IV. The direction of rotation the stator current vector, which rotates at the slip frequency $\omega_s$, is dependent on the polarity of the drive torque demand while the direction of rotation of the ac source voltage vector depends on the intended phase sequence connection relative to the $ABC$ stator winding axis. The three SCRs on the shorted side and the three SCRs on the ac-source side that participate in the current commutation are determined by the location of the stator current vector.

The high-speed-to-low-speed mode transition in the LSI topology has identical constraints on the ac-source voltage vector as outlined for the LSS topology in Sec.3.3.2. However, the sector boundaries are required to be adjusted to incorporate the presence of the short instead of the dc source voltage implying $\varepsilon = 0$, evident from (3.1).
stationary -- rotating at slip frequency

\[ m \sim \text{rotating at ac source frequency} \]

Figure 3.6: LSI topology - Commutation diagram highlighting the commutation requirement during low-speed-to-high-speed mode transition. The incoming ac source voltage vector \( \vec{V}_{\text{grid}} \) must be in the same sector as the stator current vector \( \bar{I}_S \) for simultaneous natural commutation for the relevant SCRs in the low-speed-SCR bank while the current is steered to the high-speed SCR-bank.

3.4 Eight-thyristor-based transfer switch

If one of the phases of the three-phase ac source is chosen as the reference potential, instead of the neutral of the three-phase ac source as in the twelve-thyristor-based transfer switch, the number of thyristors required to perform the on-the-fly reconfiguration of the DFM reduces to eight. The proposed eight-thyristor-based (ETB) transfer switch is shown in Fig. 3.7. In this switch configuration, the A-phase of the DFM is permanently connected to the A-phase of the ac source voltage. In high-speed mode, SCRs \( T_{\text{high},B}^R, T_{\text{high},B}^L, T_{\text{high},C}^C, \) and \( T_{\text{high},C}^R \) are turned ON thereby connecting the remaining two phases of the DFM stator to the ac source. These four SCRs form the high-speed SCR-bank. Each of these SCRs conduct for half of the time period corresponding to the ac source fundamental frequency. Alternatively, in low-speed mode, SCRs \( T_{\text{low},B}^R, T_{\text{low},B}^L, T_{\text{low},C}^C, \) and \( T_{\text{low},C}^R \) are turned ON either to connect the stator to the dc source in the LSS topology or to short them together in the LSI topology. These four SCRs are collectively termed as the low-speed SCR-bank. In the LSS topology, the positive polarity of the dc source is permanently connected to the A-phase of the ac source. As the dc source is created using a transformer-rectifier from the ac source, there is inherent isolation between the sources. In the LSI topology, the three phase stator windings are all connected to the A-phase of the ac source voltage during low-speed operation. The SCRs used in this topology are rated to withstand ac source line-to-line voltages unlike in the TTB transfer switch in which the SCRs must be rated to withstand ac source line-to-neutral voltage. Also, SCR \( T_{\text{low},B}^A \) in the TTB transfer switch incurs the maximum conduction loss when the switched-DFM drive operates in low-speed mode in the LSS topology. Removal of this SCR leads to significant reduction in the conduction loss and, therefore,
cooling requirement for the ETB transfer switch. In the following section, the commutation requirement of the ETB transfer is analyzed in light of the drive topology and operating mode transition. Constraints on the ac source voltage are derived based on the DFM stator current that ensures simultaneous commutations of all the outgoing SCRs without shorting the sources or requiring any additional auxiliary circuit for SCR commutation.

3.4.1 LSI topology: Commutation requirement during low-speed-to-high-speed mode transition

With the DFM stator initially shorted, the low-speed SCR-bank is ON and the stator current alternates at the slip frequency. Depending on the stator current polarity in the B and C phases, one of the two SCRs per phase conducts at any instant of time. This gives rise to four possible combinations for the SCR commutation requirement based on the polarity of the current in the B and C phases. The four possible combinations are demarcated in the commutation diagram by the filled-blue sectors relative to the stator ABC winding axis as shown in Fig. 3.8. For example, a negative polarity of stator currents in the B and C phases corresponds to the condition when the stator current vector is in sector I as shown in the Fig. 3.8(a).

The polarities of the incoming ac source line voltages $V_{ba}$ and $V_{ca}$ must be same as the polarity of the respective phase stator currents in the B and C phases to naturally steer the stator current to the ac source at the instant of mode transition. These constraints are satisfied when the incoming ac source voltage vector is within the arc $PP'$ and $QQ'$ respectively. The intersection of the two arcs form the hatched sector that

Figure 3.7: Eight thyristor-based transfer switch (ETB) - A-phase of the ac source is the reference for the transfer switch operation.
Table 3.2: Commutation constraints on allowable ac source voltage vector location based on the location of the stator current vector for simultaneous natural commutation of all outgoing SCRs in the LSI topology.

<table>
<thead>
<tr>
<th>Stator current location</th>
<th>Low-speed-to-high-speed mode transition</th>
<th>High-speed-to-low-speed mode transition</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Required ac-source voltage vector location</td>
<td>Outgoing SCRs</td>
</tr>
<tr>
<td>Sector I</td>
<td>$-60^\circ &lt; \angle V_{grid} &lt; 60^\circ$</td>
<td>$T_{low,R}^B$, $T_{low,R}^C$</td>
</tr>
<tr>
<td>Sectors II, III</td>
<td>$60^\circ &lt; \angle V_{grid} &lt; 120^\circ$</td>
<td>$T_{low,F}^B$, $T_{low,F}^C$</td>
</tr>
<tr>
<td>Sector IV</td>
<td>$120^\circ &lt; \angle V_{grid} &lt; 240^\circ$</td>
<td>$T_{low,F}^B$, $T_{low,F}^C$</td>
</tr>
<tr>
<td>Sectors V, VI</td>
<td>$240^\circ &lt; \angle V_{grid} &lt; 300^\circ$</td>
<td>$T_{low,R}^B$, $T_{low,F}^C$</td>
</tr>
</tbody>
</table>

represents the allowable region for the ac source voltage vector at the instant of mode transition for a natural commutation of the outgoing SCRs in the low-speed SCR-bank. The required constraints on the ac source voltage vector and the participating SCRs for all possible stator current vector locations have been detailed in Table. 3.2.
Figure 3.8: Commutation diagram during low-speed-to-high-speed mode transition in the LSI topology. Red arc (PP') : allowable line-to-line voltage between B and A phases for natural commutation of the B-phase SCRs. Green arc (QQ') : allowable line-to-line voltage between C and A phases for natural commutation of the C-phase SCRs. Blue-filled sector: stator current vector location. Hatched sector: allowable ac source voltage vector for simultaneous natural commutation of all the outgoing SCRs. (a) Current polarities in the B and C phases are both negative (b) Current polarity in the B-phase is positive while in the C-phase is negative (c) Current polarities in the B and C phases are both positive (d) Current polarity in the B-phase is negative while in the C-phase is positive.
3.4.2 LSI topology: Commutation requirement during high-speed-to-low-speed mode transition

With the DFM stator initially connected to the ac source, the high-speed SCR-bank is ON. The stator current and the ac source voltage alternate at the ac source frequency. Figure 3.9 depicts the constraints on the allowable ac source voltage vector location (hatched sector) based on the stator current vector location (filled-blue sector) for simultaneous natural commutation of all outgoing SCRs. For example, a negative stator current amplitudes in the B and C phases require that the ac source voltage vector must be within the sector $POQ'$ in Fig. 3.9(a).

As evident from Fig. 3.9, ideally the stator power factor angle must be between $90^\circ$ and $270^\circ$ at the instant of high-speed-to-low-speed mode transition independent of the stator current polarity for a natural commutation of the outgoing SCRs in the high-speed SCR-bank. The allowable range of power factor angles for the ETB transfer switch is wider by $60^\circ$ compared to the TTB transfer switch. The wider range of allowable stator power factor angles add more flexibility in terms of allowable reactive power and transition condition from the DFM perspective as will be shown in Chapter 5. Table 3.2 also includes the required ac source voltage vector constraint and the participating SCRs during a high-speed-to-low-speed transition for all the possible stator current vector locations.
Figure 3.9: Commutation diagram during high-speed-to-low-speed mode transition in the LSI topology. Red arc ($PP'$): allowable line-to-line voltage between $B$ and $A$ phases for natural commutation of the $B$-phase SCRs. Green arc ($QQ'$): allowable line-to-line voltage between $C$ and $A$ phases for natural commutation of the $C$-phase SCRs. Blue-filled sector: stator current vector location. Hatched sector: allowable ac source voltage vector for simultaneous natural commutation all outgoing SCRs. (a) Current polarities in the $B$ and $C$ phases are both negative (b) Current polarities in the $B$-phase is positive while in the $C$ phase is negative (c) Current polarities in the $B$ and $C$ phases are both positive (d) Current polarities in the $B$-phase is negative while in the $C$-phase is positive.
3.4.3 LSS topology: Effect of the dc source on the commutation requirement for the ETB transfer switch

During steady-state low-speed operation, the stator current is dc in the LSS topology. The stator $A$-phase is connected to the positive polarity of the isolated dc source of magnitude $V_{dc}$ and the $B$ and $C$ phases are connected to the negative polarity as shown in Fig. 3.7. The connection enforces a stationary stator current vector $\overrightarrow{T_s}$ directed along the $A$-phase axis as shown in Fig. 3.10. The negative current polarities in the stator $B$ and $C$ phases resembles the condition of Fig. 3.8(a) in the LSI topology. Accounting for the dc source voltage magnitude, the required condition on the incoming ac source voltage for a natural commutation of the outgoing SCR $T_{low,R}^B$ in the $B$-phase is given by

$$V_a - V_{dc} > V_b.$$  \hspace{1cm} (3.2)

Similarly, the required condition for the outgoing SCR $T_{low,R}^C$ in the $C$-phase is given by

$$V_a - V_{dc} > V_c.$$  \hspace{1cm} (3.3)

Rearranging (3.2) in terms of the ac source line-to-line voltage $V_{ba}$ results in

$$V_{ba} < -V_{dc}.$$  \hspace{1cm} (3.4)

Similarly, rearranging (3.3) in terms of the ac source line-to-line voltages $V_{ca}$ yields

$$V_{ca} < -V_{dc}.$$  \hspace{1cm} (3.5)

The above constraints are shown by the lines $XX'$ and $YY'$ in Fig. 3.10. Intersection of the two constraints forms the hatched sector $P'OQ$ that sets the bound on the allowable location of the incoming ac source voltage vector for a successful natural steering of the stator current from the dc source to the ac source.

The angle $\epsilon$ of the sector $P'OQ$ in Fig. 3.10 is dependent on the voltage magnitude of the dc source and the line-to-line voltage of the ac source and is given by,

$$\epsilon = 60^\circ - \arcsin \frac{V_{dc}}{V_{ab}}.$$  \hspace{1cm} (3.6)

For example, for a dc-source of 20 V magnitude and an ac source of 146 V (rms, line-to-line), the incoming ac voltage vector must be within $\pm 55.5^\circ$ relative to the stator $A$-phase winding axis for a simultaneous natural commutation of SCRs $T_{low,R}^B$ and $T_{low,R}^C$ during low-speed-to-high-speed mode transition. In the
Figure 3.10: Allowable ac source voltage vector location for a low-speed-to-high-speed mode transition in the LSS topology of the switched-DFM drive. Prior to mode transition, the stator current vector is stationary and aligned with the stator A-phase axis. The allowable sector boundary incorporates the necessary constraints due to the presence of the dc source voltage.

TTB transfer switch, the equivalent allowable sector angle is $\pm 30^\circ$. The extension in the allowable stator voltage vector location helps in reducing the stator flux transients and shaft torque perturbation particularly during mode transition with a low-drive torque demand as will be shown in Chapter 4.

The high-speed-to-low-speed mode transition in the LSS topology has identical constraints on the ac source voltage vector as outlined for the LSI topology in Sec. 3.4.2. However, the sector boundaries are required to be adjusted to incorporate the presence of the dc source voltage.

3.5 Alternative eight-thyristor-based transfer switch configurations

The transfer switch configuration proposed in Fig. 3.7 assumes that the A-phase of the ac source is the reference phase for the transfer switch. Alternatively, the B and C phases of the ac source can also be chosen as the reference phase. The choice of reference phase does not influence the transition requirement and constraints for the LSI topology due to symmetricity of the shorted-stator configuration. However, it is important to consider the implications of the alternative possibilities of the transfer switch configuration for the LSS topology.

The eight SCRs can be configured between the ac and the dc sources in the LSS topology by choosing either B or C phase as the reference as shown in Fig. 3.11. With the DFM initially operating in low-speed mode, the stationary stator current vector $I_s$ is aligned with the stator A-phase axis irrespective of the choice of reference phase as shown in Fig. 3.12. Based on the stator current vector and the reference phase,
the allowable region for the ac source voltage vector for a natural commutation of the low-speed SCR-bank is shown by the hatched sector. The location of the hatched sector with the B-phase as the reference, as shown in Fig. 3.12(a), can simultaneously satisfy the DFM constraints for a seamless transition only with a positive drive torque demand and when the incoming ac source voltage has ABC phase sequence. However, the location of the hatched sector with the C-phase as reference, as shown in Fig. 3.12(b), is suitable for a seamless transition from the perspective of the DFM only under a negative drive torque demand and with the incoming ac source voltage having ACB phase sequence.

Therefore, the A-phase reference configuration is preferable over the B-phase reference or C-phase reference configurations to ensure a seamless transition from the DFM perspective for all operating conditions. Moreover, in low-speed mode, the stator current in A-phase is twice that of the stator current in B- and C-phases for a balanced stator winding of the DFM. Therefore, placing a SCR on the A-phase not only needs a higher current rating but also incurs higher conduction loss. In conclusion, the transfer switch topology with the A-phase as reference has better performance in terms of losses and seamless mode transition compared to the other two choices in the LSS topology.
Figure 3.11: Alternative configuration for the eight-thyristor-based transfer switch for the LSS topology. (a) B-phase of the ac source is the reference for the transfer switch. (b) C-phase of the ac source is the reference for the transfer switch.

Figure 3.12: Allowable ac source voltage vector location (hatched regions) in the LSS topology for alternative configuration of eight-thyristor-based transfer switch. (a) B-phase of the ac source is the reference for the transfer switch. (b) C-phase of the ac source is the reference for the transfer switch.
3.6 Six-thyristor-based transfer switch configuration for LSS topology

Alternatively in the LSS topology, one more of pair of SCRs can be eliminated by operating the DFM only with two phases in the low-speed mode as shown in Fig. 3.13. In this transfer switch configuration, two of the stator windings (for example, $A$ and $B$ phases) conduct equal and opposite currents as set by dc source while the current in the $C$-phase is zero in low-speed mode. In steady-state condition, the stator current vector $\vec{i}_s$ in low-speed mode is given by

$$\vec{i}_s = \left( \frac{3}{2} - j \frac{\sqrt{3}}{2} \right) \frac{v_{dc}'}{2r_s} = \frac{1}{\sqrt{3}} \frac{v_{dc}'}{r_s} \angle -30^\circ$$

where $v_{dc}'$ is the new dc source voltage. On the contrary, the stator current vector in low-speed mode for the TTB or the ETB transfer switch is given by

$$\vec{i}_s = \frac{v_{dc}}{r_s}.$$  \hspace{1cm} (3.8)

This requires that the dc source voltage $v_{dc}'$ has to be marginally increased such that the magnitude of the stator current vector remains identical for all the transfer switch topologies achieving identical stator flux magnitude. Therefore,

$$v_{dc}' = \frac{2}{\sqrt{3}} v_{dc}.$$  \hspace{1cm} (3.9)

The operation of the DFM in low-speed mode remains same as the TTB or the ETB transfer switch except that the stator current vector $\vec{i}_s'$ is oriented along $-30^\circ$ relative to the stator $A$-phase winding axis as shown in Fig. 3.14. With the current in the $B$-phase stator winding being negative, for a successful commutation of the $T_{low,RT}$ thyristor during low-speed-to-high-speed mode transition, the required constraint on the ac source voltage is given by

$$v_{ba} < -v_{dc}'.$$  \hspace{1cm} (3.10)

The possible commutation region for the incoming ac source voltage vector $\overline{V_{grid}}$ is shown by the yellow sector in Fig. 3.14 with the $\epsilon$ being calculated as

$$\epsilon = 90^\circ - \arcsin \left( \frac{v_{dc}'}{\sqrt{3}} \right).$$  \hspace{1cm} (3.11)
Equation (3.11) is similar to (3.6) but with a wider range of allowable stator voltage vector location during low-speed-to-high-speed mode transition. For the example DFM, $\epsilon$ is calculated as $87^\circ$ implying that the dc-to-ac source transition can be initiated for any positive drive torque as will be detailed in Chapter 4.

In the absence of the SCR $T_{low,F}^C$ in Fig. 3.13, the ac-to-dc source transition can only be performed when the stator current in the $C$-phase is zero. Presence of the SCR $T_{low,F}^C$ assists during the ac-to-dc source transition without affecting the low-speed mode operation.
Figure 3.13: An alternative transfer switch topology for a switched-DFM drive in the LSS topology that achieves wider possible commutation region when compared to the ETB transfer switch. SCR $T_{low,F}^{C}$ does not conduct in low-speed mode but aids in current commutation only during transients in particular during the ac-to-de source transition.

Figure 3.14: Commutation diagram during low-speed-to-high-speed mode transition for the alternative proposed six-thyristor-based transfer switch. A wide range of allowable ac source voltage vector location is feasible during a dc-to-ac source transition as shown by the yellow sector to meet the natural commutation requirement of the SCR $T_{low,F}^{B}$. The dc source voltage vector and the steady-state stator current vector is at an angle $-30^\circ$ relative to the stator A-phase winding axis.

3.7 Experimental results

Two thyristor-based transfer switches are built for the switched-DFM drive - one with twelve thyristors as shown in Fig. 3.15(a) and the other with eight thyristors as shown in Fig. 3.15(b). Diodes are placed in series with each SCR to enable faster turn off during mode transition. An RC snubber of $330 \, \Omega$ and $6.8 \, nF$ is
Figure 3.15: Prototype thyristor-based transfer switches used for the experimental demonstration. (a) Twelve-thyristor-based transfer switch (b) Eight-thyristor-based transfer switch

connected in parallel with the SCRs to limit the dv/dt stress. In practice, the analytically-derived windows for natural commutation of outgoing SCRs are reduced taking into account the dead-time of the SCRs. For example, considering a turn-off time of 250 µs for the chosen SCR results in the allowable window to be ±24° (for a 60-Hz ac source) for safe commutation of the outgoing SCRs where the analytically-derived window corresponds to ±30°. This safety margin also ensures correct current commutations in presence of noise and harmonics in source voltages and DFM stator currents. Augmenting the design of the thyristor-based-transfer switch with fast-recovery high-power diodes can enable the reduction in the required dead-time in high-power applications.

3.7.1 Twelve-thyristor-based transfer switch

3.7.1.1 LSS topology: Mode transition between ac and dc source connections of the DFM stator

Figure 3.16 shows the performance of the TTB transfer switch during the instant of dc-to-ac source transition. The mode transition occurs at $t = 0$ when the ac source $A$-phase voltage is greater than dc-source voltage and the ac-source $B$ and $C$ phase voltages are negative, shown in Fig. 3.16(a). Initially, the dc-source side SCRs $T_{low,F}^A$, $T_{low,R}^B$, and $T_{low,R}^C$ conduct as shown in Fig. 3.16(b). After the transition, ac-source side SCRs start conducting as shown in Fig. 3.16(c) without any discontinuity in the DFM stator current. At the instant of mode transition, both the incoming ac source voltage vector and the stator current vector are in sector I, as evident from the polarity of the signals, which is required for the stator current to seamlessly transfer from the dc source to the ac source. The ac source for this test is created using an inverter (Texas Instruments High Voltage Motor Control & PFC Developer’s Kit) that is programmed to operate open loop PWM, generating 110 V (phase peak) and, 40 Hz (fundamental). An output LC filter is used to filter the
harmonics of Inverter-I and generate sinusoidal ac voltages emulating an ac source.

Similarly, Fig. 3.17 shows the performance of the transfer switch during the ac-to-dc source transition. At the instant of the transition ($t = 0$), the rotor power converter ensures that the phase currents are of opposite polarity to that of the ac source phase voltages for simultaneous natural commutation of all the three phases as can be seen in Fig. 3.17(a) and (b). The DFM stator current transitions seamlessly from the ac source to the dc source as shown in Fig. 3.17(c). After the DFM stator is transitioned to the dc source, the stator A-phase current is initially negative implying momentarily power flows back to the dc source.

In practical scenario, where the dc source is created using a transformer-rectifier with an output capacitor, a mechanism must be installed to ensure that the rectified output voltage across the output capacitor is controlled during this reverse power flow. In this experimental setup, a switch is turned on particularly during ac-source-to-dc-source transition to connect a resistance across the output capacitor that protects the dc source by dissipating power.

Figure 3.18 shows the current commutation during mode transition between two SCRs in micro-second time scale. In this small time scale, the DFM stator current behaves like a constant current load. The pre-transition SCR commutates and the load current is transitioned to post-transition SCR without interruption or short. The presence of fast-recovery diodes in series with the SCRs aids in the faster current commutation.
Figure 3.16: Performance of the twelve-thyristor-based transfer switch during the dc-to-ac source transition. The source transition occurs at \( t = 0 \) (a) AC and DC source voltages during the transition. (b) Low-speed SCR-bank currents go to zero at the instant of transition. (c) High-speed SCR-bank picks up the DFM stator current seamlessly after the transition.

Figure 3.17: Performance of the twelve-thyristor-based transfer switch during the ac-to-dc source transition. The source transition occurs at \( t = 0 \) (a) AC and DC source voltages during the transition. (b) High-speed SCR-bank currents go to zero at the instant of transition. (c) Low-speed SCR-bank picks up the DFM stator current seamlessly after the transition.

Figure 3.18: With the DFM stator current remaining unperturbed during the transition, the pre-transition SCR current commutates to post-transition SCR, shown in \( \mu s \) time scale.
3.7.1.2 LSI topology: Mode transition between ac source connection and shorted configuration of the DFM stator

In the LSI topology, the stator currents and the ac source voltages during mode transitions are shown in Fig. 3.19(a) and (b) respectively. The ac source voltage follows a $ABC$ sequence. At the low-speed-to-high-speed mode transition instant in Fig. 3.19(a), the polarity of the stator currents match the polarity of the respective ac source phase voltages implying a natural commutation of the outgoing SCRs. In this particular mode transition instant, both the stator current vector and the incoming ac source voltage vector are in sector V. The phase sequence of the stator current changes from $ACB$ to $ABC$ at the transition instant implying that the stator current vector changes direction of rotation at the instant of mode transition. At the high-speed-to-low-speed mode transition instant in Fig. 3.19(b), the polarity of the stator current is opposite to the polarity of the ac source phase voltage due to the reverse active power flow and a stator reactive power control using the rotor converter. In this particular mode transition instant, the stator current vector is in sector II while the ac source voltage vector is in sector V. The phase sequence of the stator current after and before the transition remains the same in this case. The ac source in this experiment is created using two parallel synchronous generators operating at 146 V (line-to-line, rms) and 40 Hz.

Figure 3.19: Experimental results: Ac source phase voltages and the stator currents during the mode transition instances in the LSI topology. (a) The stator current is at slip frequency when shorted and transitions to grid frequency at the low-speed-to-high-speed mode transition instant. (b) The stator current is initially at grid frequency and changes over to slip frequency during high-speed-to-low-speed mode transition.
3.7.2 Eight-thyristor-based transfer switch

3.7.2.1 LSS topology: Mode transition between ac and dc source connections of the DFM stator

Figure 3.20 and Fig. 3.21 show the performance of the ETB transfer switch during different mode transitions in the LSS topology. A low-speed-to-high-speed mode transition is shown in Fig. 3.20 (a)-(c). In this case, the low-speed SCR-bank carries dc current prior to transition as evident from Fig. 3.20 (b). The B and C phase current polarities being negative before transition, the outgoing SCRs in each of the phases undergo natural commutation when the ac source line-to-line voltages are lesser than the negative dc source voltage as shown in 3.20 (a).

A high-speed-to-low-speed mode transition is shown in Fig. 3.21 (a)-(c). Initially, the high-speed SCR-bank conducts the stator current at the ac source frequency, shown in Fig. 3.21 (b). After the transition, the stator current is commutated to the low-speed SCR-bank where the frequency changes to dc eventually, after initial transients similar to that described for the case of twelve-thyristor-based transfer switch. At the instant of transition, current polarities in the B and C phases are both positive. Therefore, the transition is activated when the ac source line-to-line voltages $V_{ba}$ and $V_{ca}$ are both negative. The stator condition is precisely obtained using proper current commands to the rotor d-axis and q-axis currents.
3.7.2.2 LSI topology: Mode transition between ac source connection and shorted configuration of the DFM stator

Figure 3.22 and Fig. 3.23 show the performance of the ETB transfer switch during different mode transitions in the LSI topology. A low-speed-to-high-speed mode transition in the LSI topology is shown in Fig. 3.22 (a)-(c). Prior to transition, the low speed SCR-bank carries slip frequency stator current as shown in Fig. 3.22 (b). After the transition at the zero second, the high-speed SCR-bank carries ac source frequency stator current shown in Fig. 3.22 (c). The B and C phase current polarities are both positive and, therefore, a positive polarity of ac source line-to-line voltages $V_{ba}$ and $V_{ca}$ make the commutation of the outgoing SCRs feasible without any additional supporting circuitry as shown in Fig. 3.22(a). This transition condition corresponds to Fig. 3.8(c).

A high-speed-to-low-speed mode transition in the LSI topology is shown in Fig. 3.23 (a)-(c). Prior to
transition, the high-speed SCR-bank carries the stator current at the ac source frequency as shown in Fig. 3.22(b). After the transition at the zero second, the low-speed SCR-bank carries slip frequency stator current shown in Fig. 3.23(c). The B and C phase current polarities are both positive and, therefore, a negative polarity of ac source line-to-line voltages \( V_{ba} \) and \( V_{ca} \) make the commutation of the outgoing SCRs feasible without any additional supporting circuitry as shown in Fig. 3.22(a). This transition condition corresponds to Fig. 3.9(a).

![Figure 3.22: Performance of the eight-thyristor-based transfer switch during low-speed-to-high-speed mode transition in the LSI topology. The source transition occurs at (t = 0) (a) Ac source line-to-line voltage (b) low-speed SCR-bank current and the A-phase current on the shorted side (c) high-speed SCR-bank current and the A-phase current on the ac-source side.](image)

![Figure 3.23: Performance of the eight-thyristor-based transfer switch during high-speed-to-low-speed mode transition in the LSI topology. The source transition occurs at (t = 0) (a) Ac source line-to-line voltage (b) high-speed SCR-bank current and the A-phase current on the ac-source side (c) low-speed SCR-bank current and the A-phase current on the shorted side.](image)

### 3.8 Summary

This chapter presented and demonstrated multiple thyristor-based transfer switches, which can change the stator supply between an ac or a dc source/shorted configuration on-the-fly for a switched-DFM drive. Using SCRs to construct the transfer switch, while allowing instantaneous transition without shorting the
sources or impairing the load currents, enables a significant reduction of the required power electronics and associated passive-auxiliary components. With no additional circuitry required for the commutation of the SCRs and a reduced rotor converter size, the proposed transfer-switch based drive architecture is a cost effective solution suitable for high power applications. The proposed transfer switch opens up a mechanism for operation of the switched-DFM drive that can be adjusted based on the specific drive design and transient performance requirement. With the availability of high-power SCRs with large short-time current ratings, a hybrid transfer switch with parallel mechanical relays can be devised to reduce conduction losses in the transfer switch to an even greater extent. Table 3.3 summarizes the key constraints set by the different transfer switch topologies to the incoming voltage vectors and stator current vectors. In the next chapter, rotor-side current control will be used not only to adhere to these constraints but also to achieve bumpless mode transitions.
Table 3.3: Summary of commutation requirement and feasibility during mode transition for different drive and transfer switch topologies.

<table>
<thead>
<tr>
<th>Switched-DFM drive topology</th>
<th>Transfer switch topology</th>
<th>Low-speed-to-high-speed mode transition</th>
<th>Low-speed-to-high-speed mode transition</th>
</tr>
</thead>
<tbody>
<tr>
<td>LSI</td>
<td>TTB</td>
<td>Feasible in all sector provided incoming ac source voltage vector is in the same sector as the stator current vector</td>
<td>Feasible in all sector provided the stator power factor angle is between 120° and 240°</td>
</tr>
<tr>
<td>ETB</td>
<td>Feasible in four sub-sectors -30° to 60°, 120° to 150°, 210° to 240°, and 300° to 330° - provided incoming ac source voltage vector is in the same sector as the stator current vector</td>
<td>Feasible in all sector provided the stator power factor angle is between 90° and 270°</td>
<td></td>
</tr>
<tr>
<td>LSS</td>
<td>TTB</td>
<td>Feasible only when the incoming ac voltage vector is within ±30°</td>
<td>Feasible in all sector provided the stator power factor angle is between 120° and 240°</td>
</tr>
<tr>
<td>ETB</td>
<td>Feasible only when the incoming ac voltage vector is within ~±60°</td>
<td>Feasible in all sector provided the stator power factor angle is between 90° and 270°</td>
<td></td>
</tr>
<tr>
<td>STB</td>
<td>Feasible only when the incoming ac voltage vector is within ~±90°</td>
<td>Feasible in all sector provided the stator power factor angle is between 90° and 270°</td>
<td></td>
</tr>
</tbody>
</table>
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Chapter 4

Drive control architecture

The proposed switched-DFM drive has significantly different challenges from a control perspective compared to a standard grid-connected doubly-fed induction machine [27, 28, 29, 30, 46] or a wound-field synchronous machine [31]. The rotor power electronics control has to adapt to on-the-fly machine reconfiguration and should be able to control the individual modes (dc, short, or ac) of the drive operation. As the excitation in the stator undergoes a step change during mode changeover, the control scheme must ensure a stable and “bumpless” transition such that the coupled mechanical load and electrical sources are subjected to minimal disturbances. During the low-speed mode, the proposed drive either has to operate with a dc stator flux in the LSS topology or with a very-low-frequency (slip) stator flux in the LSI topology that makes stator flux estimation challenging in the presence of measurement offset and drift. Moreover, the stator-to-source connection change-over must be within the allowable constraints of the thyristor-based transfer switch as described in Chapter 3. In addition, all of this must work in the context of a wide variety of load torques that can experience sudden disturbances. This chapter addresses these above challenges and proposes an architecture that can ensure a seamless control for the DFM drive during all operating modes, including the instance of mode transition. In addition, the DFM can provide a controllable power factor to an ac grid or ship micro-grid which is also explored in this chapter.

The proposed drive control architecture is based on the indirect field-oriented control. A doubly-fed induction machine for wind application has been traditionally controlled in the stator-flux orientation [46] or in the natural-flux/grid-flux orientation [29]. In a motoring application, e.g. on-board a ship with a fixed voltage and frequency ac service, ignoring winding resistance, the stator flux magnitude is fixed during the high-speed mode by the ship’s ac service. The rotor inverter is used to strictly control the shaft torque. Based on the remaining drive capability, the rotor inverter can also share the magnetizing current and control
the stator power factor. During the low-speed mode, the stator flux magnitude is either set by both the stator and rotor magnetizing currents in the LSS topology or entirely by the rotor magnetizing current in the LSI topology. As the stator is either connected to an uncontrolled dc source or shorted, depending on the topology, the rotor inverter must ensure a desired stator flux level before commanding the necessary torque-producing current component. Therefore, to control both low-speed and high-speed modes seamlessly, it makes sense to develop the control architecture in a reference frame oriented to the stator flux.

### 4.1 Control architecture block diagram for rotor-side converter

The control architecture for the proposed drive comprises two sections:

1. The stator flux magnitude $\psi_s$, stator flux angle $\theta_s$ (with respect to the stator A-phase axis), and the stator flux frequency $\Omega_s$ are estimated based on the measurement of the source voltages connected to the stator, rotor current, and rotor mechanical speed, as shown in Fig. 4.1. The estimated stator flux angle is used to transform relevant variables to the stator flux reference frame. In steady state, the stator flux frequency is zero in low-speed mode in the LSS topology, slip frequency in low-speed mode in the LSI topology, and equal to the ac source frequency in high-speed mode. However, at mode changeover, the stator flux frequency goes through a transient. Estimating the stator flux frequency enables smooth torque production under all operating conditions by providing appropriate feed-forward terms for the rotor current controllers.

2. The commands to the rotor-side converter and the stator source transition $S_W$ are generated based on the estimated internal variables, terminal measurements, and the reference set points, as shown in Fig. 4.2. Full mechanical control of the DFM is achieved from the rotor.

![Figure 4.1: Estimation of machine internal variables based on terminal measurements required for the proposed control architecture.](image)
Figure 4.2: Control architecture of the proposed DFM drive using the estimated machine internal variables and terminal measurements to command the rotor-side inverter and the stator-side thyristor-based transfer switch based on the reference set point.

The rotor $q$-axis current reference is generated depending on the desired torque. Based on the required drive torque-speed characteristic, the drive may be designed to have different torque capability in low and high speed region. This results in different saturation limits at the output of the speed controller $T_{low}$ or $T_{high}$ based on the operating mode as shown in Fig. 4.2. The torque limits are transitioned with a low pass filter of appropriate cut-off frequency during mode transition. This avoids a step change in the demanded $q$-axis current when the drive is operating at its peak torque capability while undergoing a mode transition.

For example, in a drive designed with lower torque capability in low-speed mode compared to high-speed mode, the time-constant $T_1$, relevant during acceleration, is chosen to be equal to or slower than the DFM stator time constant. This allows the stator flux to settle to a steady value in high-speed mode before a higher torque is demanded. However, the time-constant $T_2$, relevant during braking, is chosen to be significantly faster than the DFM stator time constant.

The rotor $d$-axis current $I_{rd}$ performs different functions during different operating modes. In low-speed mode, $I_{rd}$ is used to maintain the stator flux magnitude irrespective of the switched-DFM drive topology. In high-speed mode, $I_{rd}$ is used to control the stator reactive power within the ratings of the rotor power electronic converter. Thus, based on the operating mode, $I_{rd}$ is generated from either the stator reactive power controller or the stator flux controller. The rotor $d$-axis current limit is based on the commanded rotor $q$-axis current such that the DFM/rotor power electronics converter current rating $I_{rated}$ is not exceeded during any of the operating modes as shown in Fig. 4.2.
During mode transition, for example, at the instant of low-speed-to-high-speed mode transition, the stator flux vector changes from being stationary in the stator reference frame (in the LSS topology) or rotating at slip frequency (in the LSI topology) to a rotating one at the ac source frequency. The operating stator flux magnitude may also change based on the drive design as was highlighted in Chapter 2. This directly impacts the shaft torque and the power transients seen by the ac source during the mode transition. Two degrees of freedom can be utilized to ensure that the stator flux is minimally perturbed. The first degree of freedom is the switching instant at which the stator connection of the DFM is changed between the sources. This is appropriately chosen by the synchronizer, shown in Fig. 4.2, that commands binary signal $S_w$ to the thyristor-based transfer switch. Based on the relative magnitude of the designed transition speed and the actual mechanical speed of the rotor, a transition command $S_w^*$ is generated. $S_w^*$ is fed to the synchronizer along with the estimated stator flux angle and the measured ac and dc source voltages that computes $S_w$.

The remaining degree of freedom to suppress the stator flux oscillation during the mode transition is by actively commanding the d-axis rotor current. This is achieved by appropriate design of the stator flux transition controller. Similarly, the same two degrees of freedom are utilized to suppress perturbations on the shaft torque and to the ac source during high-speed-to-low-speed mode transition instant. The following sections describe in details the individual functionality of the different blocks in the control architecture that together achieves seamless control of the switched-DFM drive across the entire speed range.

### 4.1.1 Stator flux estimator

The most important aspect of stator-flux oriented control is the estimation of stator flux magnitude and angle under all operating conditions. In a DFM, the stator flux is generally estimated either through a voltage model based on the differential equations or a current model based on the flux linkage equations [47]. Alternatively, the stator flux can be estimated from stator voltage vector by assuming that it lags the voltage by 90° [46]. As the proposed drive configuration undergoes voltage and flux transients, the assumption of constant angle between the stator voltage and the stator flux vector need not be always true. The voltage model incorporates integrators in the stationary $\alpha\beta$ reference frame given by,

$$
\Psi_{s\alpha} = \int (V_{s\alpha} - R_s I_{s\alpha}) \, dt
$$

and

$$
\Psi_{s\beta} = \int (V_{s\beta} - R_s I_{s\beta}) \, dt.
$$

In high-speed mode, with the stator connected to the ac source, the integrators can be approximated with low pass filters in practical implementations [32], avoiding problems with measurement offset and drift in the integral computations [47]. The cutoff frequency of these low-pass filters must be well below the signal...
frequency. However this approximation of the integral with a low pass filter remains impractical in low-speed mode – more specifically for the configuration where the stator is connected to a dc source with the signal frequency being zero.

The estimation problem in both the modes can be greatly simplified by computing the stator flux with a hybrid estimator that uses flux linkage equations (a current model) augmented with the voltage model (4.1). Using stator-flux equation (current model) in (4.1) to replace the stator current yields,

$$\bar{V}_s = \frac{R_s}{L_s} (\bar{\Psi}_s - MT_r e^{j\epsilon}) + \frac{d}{dt} \bar{\Psi}_s.$$  \hspace{1cm} (4.2)

The rotor current vector in the stationary reference frame is given by:

$$I_{r\alpha} + jI_{r\beta} = \bar{T}_r e^{j\epsilon},$$  \hspace{1cm} (4.3)

where $\epsilon$ is the angle between the rotor $ABC$ winding axis and the stator $ABC$ winding reference frame.

Using (4.3) in (4.2) and rearranging real and imaginary parts gives,

$$\frac{d}{dt} \bar{\Psi}_{s\alpha} + \frac{R_s}{L_s} \bar{\Psi}_{s\alpha} = V_{s\alpha} + \frac{R_s M}{L_r} I_{r\alpha},$$

and

$$\frac{d}{dt} \bar{\Psi}_{s\beta} + \frac{R_s}{L_s} \bar{\Psi}_{s\beta} = V_{s\beta} + \frac{R_s M}{L_r} I_{r\beta}.$$  \hspace{1cm} (4.4)

These expressions for the components of stator flux are governed by first order differential equations that correspond to “low-pass” transfer functions relating the “input” voltages and currents to the “output” fluxes. This flux estimator is attractive, as its low pass quality can minimize the effect of voltage and current measurement noise.

By avoiding the direct use of stator current in the flux estimator, a mechanism for potentially detecting parameter variations as the machine operates is preserved. Based on the estimated stator flux and the measured rotor currents, the stator current components can be estimated using

$$I_{sd} = \frac{\bar{\Psi}_s - M I_{rd}}{L_s}$$  \hspace{1cm} (4.5)

and

$$I_{sq} = -\frac{M}{L_s} I_{rq}.$$  \hspace{1cm} (4.6)

A direct measurement of the stator currents can potentially be compared against the estimated values to verify the stator flux estimation. Disagreement between the measured and the estimated stator current can be used to trigger the need for in-situ revision of the machine parameters.
Apart from the stator flux magnitude and angle, other necessary estimated variables for proper control of the proposed drive include the stator flux frequency $\Omega_s$ and the angle $\delta$, defined as the angle between the stator flux vector and the stator voltage vector. While $\delta$ can be easily calculated based on the estimated stator flux angle and the measured stator voltage angle, the stator flux frequency can be estimated from the transformed stator voltage, rotor current, and the DFM parameters using

$$\Omega_s = \frac{d\theta_s}{dt} = \frac{V_{sq} + \frac{R_r M}{L_s} I_{rq}}{\Psi_s}. \quad (4.7)$$

### 4.1.2 Design of the rotor currents and speed controller

Irrespective of the operating mode and the switched-DFM drive topology, the rotor-side inverter control requires a feedback loop to control the speed and the $d$ and $q$ axis rotor currents. The $q$-axis current controller can be designed based on the $q$-axis rotor current dynamics, which is given by

$$V_{rq} = R_r I_{rq} + \left(L_r - \frac{M^2}{L_s}\right) \frac{d}{dt} I_{rq} + \left[ (\Omega_s - \Omega_e) \Psi_{rd} \right]. \quad (4.8)$$

The drive $V_{rq}$ is provided by the rotor inverter. The desired $V_{rq}^*$ is computed as the output of a proportional-integral (PI) compensator added to a speed voltage term that cancels the cross-coupling effect of $\Psi_{rd}$ in (4.8). $\Psi_{rd}$ is indirectly estimated using the DFM machine model as:

$$\Psi_{rd} = \left(L_r - \frac{M^2}{L_s}\right) I_{rd} + \frac{M}{L_s} \Psi_s. \quad (4.9)$$

The overall $q$-axis current controller along with the involved machine dynamics based on (4.8) is shown in Fig. 4.3 where

$$R_E = R_r, \quad L_E = \left(L_r - \frac{M^2}{L_s}\right), \quad FF = - (\Omega_s - \Omega_e) \left[ \left(L_r - \frac{M^2}{L_s}\right) I_{rd} + \frac{M}{L_s} \Psi_s \right]. \quad (4.10)$$

$R_E$, $L_E$, and $FF$ represents the equivalent resistance, equivalent inductance, and the computed feed-forward quantity that is used to design the controller for the first-order plant of Fig. 4.3. Proportional and integral gains ($K_p$ and $K_i$) of the controller are chosen to ensure loop stability, desired transient response, and acceptable steady-state error.

The design of the $d$-axis current controller is similar. The $d$-axis rotor current dynamics is given by
$V_{rd} = \left( R_r + R_s \frac{M^2}{L_s} \right) I_{rd} + \left( L_r - \frac{M^2}{L_s} \right) \frac{d}{dt} I_{rd} + \left[ \frac{M}{L_s} V_{sd} - \frac{R_s M}{L_s^2} \Psi_s - (\Omega_s - \Omega_c) \left( L_r - \frac{M^2}{L_s} \right) I_{rq} \right]. \quad (4.11)$

This equation is used to design a d-axis current controller with a PI compensator. The equivalent inductance and resistance of the machine dynamics are appropriately changed in Fig. 4.3 using (4.11) to design $K_p$ and $K_i$ for the PI compensator by making

$$R_E = \left( R_r + R_s \frac{M^2}{L_s} \right)$$

$$L_E = \left( L_r - \frac{M^2}{L_s} \right) \quad (4.12)$$

$$FF = - \left[ \frac{M}{L_s} V_{sd} - \frac{R_s M}{L_s^2} \Psi_s - (\Omega_s - \Omega_c) \left( L_r - \frac{M^2}{L_s} \right) I_{rq} \right].$$

Finally, the shaft speed control is designed using

$$\frac{2J}{P} \frac{d\Omega_s}{dt} + \frac{2}{P} B \Omega_s = T - T_l, \quad (4.13)$$

and by noting that the drive torque is related to the q-axis rotor current through

$$T = \frac{3P M}{2L_s} \Psi_s I_{rq}. \quad (4.14)$$

The bandwidth of the speed controller is chosen such that it is much lower than the q-axis current controller bandwidth. This is a design choice that will determine the final performance of the propulsion drive. A PI compensator stabilizes the outer speed loop control.
4.1.3 Design of the stator flux controller in low-speed mode

In low-speed mode, the stator flux magnitude for the DFM needs to be either controlled and/or created through the rotor-side inverter for both of the switched-DFM drive topology. In the LSS topology, the dc source connected to the stator, which is created using a transformer-rectifier is uncontrolled while in the LSI topology, there is no excitation from the stator. Therefore, a stator flux controller is essential in low-speed mode that can command the d-axis rotor current for appropriate stator flux magnitude. The stator flux magnitude dynamics is given by

\[ I_{rd} = \frac{1}{M} \Psi_s + \left( \frac{L_s}{M R_s} \right) \frac{d}{dt} \Psi_s - \left[ \frac{L_s}{M R_s} V_{sd} \right]. \]  \hspace{1cm} (4.15)

A PI compensator responsive to the estimated flux error is used to drive the d-axis rotor current and, therefore, the stator flux magnitude. The equivalent inductance and resistance of the machine dynamics are appropriately changed in Fig. 4.3 using (4.11) to design \( K_p \) and \( K_i \) for the PI compensator by making

\[ R_F = \frac{1}{M} \]
\[ L_F = \left( \frac{L_s}{M R_s} \right) \]
\[ FF = \left[ \frac{L_s}{M R_s} V_{sd} \right] \hspace{1cm} (4.16)\]

The previously discussed rotor d-axis current compensator is a minor loop within the stator flux controller, as shown in Fig. 4.2. Therefore, the bandwidth of the flux controller is chosen such that it is significantly lower than the d-axis current controller. In the LSS topology, \( V_{sd} \) is calculated by projecting the dc source voltage onto the stator flux reference frame while \( V_{sd} \) is equated to zero in the LSI topology.

4.1.4 Design of the reactive power controller in high-speed mode

In high-speed mode, the stator flux magnitude is predominantly set by the ac source voltage and frequency. Assuming a negligible stator resistance in (4.7)

\[ \Psi_s = \frac{V_{sq} + \frac{K_p M I_{sq}}{\Omega_s}}{\Omega_s} \approx \frac{V_{sq}}{\Omega_s} \approx \frac{|V_{grid}|}{\Omega_{ac}}. \]  \hspace{1cm} (4.17)

The d-axis rotor current in this mode is commanded to control the flow of stator reactive power to and from the ac mains as in a standard wind turbine application [28]. Any desired stator reactive power, within the machine and inverter limits, can be provided according to:

\[ Q_s = \frac{3}{2} (V_{sq} I_{sd} - V_{sd} I_{sq}). \]  \hspace{1cm} (4.18)
Substituting stator current components in terms of rotor currents using (4.5) and (4.6) in (4.18) and rearranging, the required rotor d-axis current necessary to command the steady-state reactive power flow on the stator is computed as:

\[
I_{sd}^* = \frac{1}{M} \psi_s + \frac{V_{sd}}{V_{eq}} I_{rq} - \frac{2}{3} \frac{L_s}{M V_{eq}} Q_s^*.
\] (4.19)

4.2 Control of the switched-DFM drive during mode transitions

The doubly-fed machine used in the drive can be considered to be either a synchronous machine or an induction machine during low-speed mode depending on the topology or as an ac-source-connected doubly-fed induction machine in the high-speed mode. In individual modes, the control architecture described using indirect field-oriented control enables smooth operation of the drive. The major challenge appears at the mode transition instant when the DFM stator and the rotor power electronic control must be reconfigured “on-the-fly” i.e., while the shaft is turning and producing torque. In this section, first, a generalized stator-flux transition model with appropriate assumptions is presented. This model can universally represent the mode transition behavior of the DFM irrespective of the stator connection. The model is then used to create an analytical framework to evaluate and compare the performances of different thyristor-based transfer switch topologies outlined in Chapter 3 during all possible mode transition instances.

4.2.1 Generalized stator-flux transition model

Assuming a high current-controller bandwidth and a large mechanical time constant of the drive, a simplified non-linear plant model of the DFM can be developed to investigate the stator flux transition during source change-over. Rearranging (4.15),

\[
\frac{d}{dt} \psi_s = -\frac{R_s}{L_s} \psi_s + V_{sd} + \left( \frac{M R_s}{L_s} \right) I_{rd}.
\] (4.20)

Replacing the q-axis rotor current in (4.7) using (4.14) yields,

\[
\Omega_s = \frac{V_{eq}}{\psi_s} - \frac{1}{\psi_s^2} \frac{4R_s}{3P} T.
\] (4.21)

The angle between the stator flux vector and the stator voltage vector \( \delta \) can be computed as:

\[
\frac{d}{dt} \delta = \Omega_{source} - \Omega_s.
\] (4.22)
and of course, the total stator voltage vector magnitude equals the resultant from the d-axis and q-axis components.

\[ V_{sd}^2 + V_{sq}^2 = V_{source}^2; \quad V_{sq} = V_{source} \sin \delta; \quad V_{sq} = V_{source} \cos \delta. \]  

(4.23)

Using (4.20)-(4.23), the non-linear DFM plant model can be described in state-space form as,

\[
\begin{bmatrix}
\frac{d}{dt} \Psi_s \\
\frac{d}{dt} \delta
\end{bmatrix} = \begin{bmatrix}
\frac{-R_s}{L_s} \Psi_s + V_{source} \cos \delta \\
\Omega_{source} - \frac{V_{source} \sin \delta}{\Psi_s} + \frac{1}{\omega_s^2} \frac{4R_s}{3T} I_d
\end{bmatrix} + \begin{bmatrix}
\frac{M}{L_s} \\
0
\end{bmatrix} I_{rd}. 
\]  

(4.24)

The model is equivalent to the well-known model in [48] that has been used in grid-connected doubly-fed induction generator stability studies except that the q-axis current is replaced with the drive torque demand. Normalizing the model relative to the respective base quantities results in

\[
\begin{bmatrix}
\frac{1}{\omega_s} \frac{d}{dt} \Psi_s \\
\frac{d}{dt} \delta
\end{bmatrix} = \begin{bmatrix}
\frac{-R_s}{x_s} \Psi_s + V_{source} \cos \delta \\
\omega_{source} - \frac{1}{\Psi_s} v_{source} \sin \delta + \frac{1}{\omega_s^2} r_s r
\end{bmatrix} + \begin{bmatrix}
\frac{r_s x_s}{\omega_s} \\
0
\end{bmatrix} I_{rd}. 
\]  

(4.25)

The parameters of the normalized state-space model are the stator resistance \( r_s \), the stator inductance \( x_s \), and the mutual inductance \( x_m \). The steady-state condition of the stator flux for the individual stator connection is obtained by equating (4.25) to zero with an appropriate stator source voltage \( v_{source} \) and the frequency \( \omega_{source} \) based on the connection. For example, \((v_{source}, \omega_{source})\) changes in (4.25) as \((1, 1)\), \((v_{dc}, 0)\), and \((0, 0)\) depending on the stator connection to the ac source, to the dc source \(v_{dc}\), or whether the stator windings are shorted together respectively. In the shorted configuration, the stator flux angle \( \delta \) is relative to the A-phase stator axis as the stator voltage is zero. Assuming a positive drive torque \( \tau \), the steady-state condition is graphically represented by \( SS \) in the \( \psi_s - \delta \) phase plane as shown in Fig. 4.4 depending on the stator winding connection. When the operating point is perturbed from the steady-state condition, the vector field drives the stator flux trajectory in the phase plane.

When the stator is connected:

1. to the dc source (low-speed mode, LSS topology) - The steady-state is represented by a stationary point in the \( \psi_s - \delta \) plane with the \( \psi_s \) corresponding to the designed stator flux magnitude (for example, 0.72 p.u. in this case). The sine of the angle \( \delta \) depends on the drive torque demand. The operating point \( SS \) is inherently a saddle point as observed based on the vector field direction. The operating point is stabilized by proper feed-forward terms in the stator flux magnitude controller.

2. as a short (low-speed mode, LSI topology) - The DFM operation is represented by a line in the \( \psi_s - \delta \)
Figure 4.4: Comparison of the steady-state operating condition $SS$ and the vector field for the individual stator connections. The stator is (a) connected to the dc source, (b) shorted together, or (c) connected to the ac source. A mode transition maps the steady-state condition in the pre-transition vector field as a perturbation in the post-transition vector field.
plane with the $\psi_s$ corresponding to the designed stator flux magnitude (for example, 0.9 p.u. in this case). In this case, the time derivative of $\delta$ depends on the drive torque demand. The time derivative of $\delta$ corresponds to the slip speed.

3. to the ac source (high-speed mode) - The steady-state is once again represented by a stationary point in the $\psi_s - \delta$ with $\psi_s \simeq 1$ and $\delta \simeq 90^\circ$. The operating point is fairly independent of the drive torque demand for a negligible stator resistance.

A source change-over instantly maps the operating point from one vector field to the other (depending on the pre- and post-transition stator connection) with the identical stator flux magnitude and with initial $\delta$ decided by the transition instant. The state trajectory to the final steady-state condition is governed by the vector field and the d-axis current controller command $i^*_d$. At the source transition instant, the different constraints on the DFM and on the SCR-based transfer switch are:

1. seamless and controllable drive torque,

2. required rotor voltage within the designed maximum limit based on the steady-state drive design to prevent uncontrolled rotor current,

3. minimal stator flux magnitude swing to prevent saturation in the DFM magnetic circuit and power sloshing between the DFM and the source connected to the stator,

4. natural commutation of the SCRs in the transfer switch by remaining withing the commutation constraints of the selected thyristor-based transfer switch topology, and

5. stator and rotor currents within its respective ratings.

### 4.2.2 Low-speed-to-high-speed mode transition in LSS topology

With the DFM stator initially connected to the dc source, the low-speed SCR-bank is ON. The stator current vector $i_s$ and the stator voltage are stationary relative to the $ABC$ stator winding axis as shown in Fig. 4.5. Connecting the A-phase of the stator to the positive polarity and B and C-phases to the negative polarity of the dc source enforces the stator voltage vector orientation along the A-phase axis. Under steady-state condition, the stator current vector is also directed towards the A-phase axis. The incoming ac source voltage vector $\tilde{v}_{grid}$ rotates in $ABC$-phase sequence at the grid frequency $\omega_{ac}$ relative to the stationary reference frame. The blue sector represents the feasible region for the incoming ac source voltage vector that ensures natural commutation for all outgoing SCRs of the low-speed SCR-bank in the TTB transfer switch.

Reducing the number of thyristors to eight, as in ETB transfer switch, extends the feasible region to the
Figure 4.5: Space vector diagram superimposed on the commutation diagram during low-speed-to-high-speed mode transition that ensures minimum flux perturbation and natural commutation of the conducting SCR bank under rated drive torque demand. Turning ON the high-speed SCR-bank of the ETB transfer switch when the ac source voltage vector is within the green sector naturally turns OFF the low-speed SCR-bank. The corresponding region for the TTB transfer switch is shown in blue.

Prior to the mode transition, the orientation of the stator flux vector $\vec{\psi}_s$ is dependent on the operating drive torque $\tau$, the normalized dc source voltage $v_{dc}$ and the designed stator flux magnitude in low-speed mode $\psi_s$ and is given by

$$\delta_{dc} = \arcsin \frac{\tau \psi_s}{v_{dc} \psi_s},$$

(4.26)

where $\delta_{dc}$ represents the angle between the stator flux vector and the stator voltage vector in low-speed mode.

Assuming that the stator flux magnitude before and after the mode transition is identical, for minimal perturbation of the stator flux magnitude, (4.20) must be equated to zero during mode transition. This implies that the d-axis component of the incoming ac voltage must be equal to the existing d-axis voltage, ignoring the effect of the stator resistance. Additionally, the stator flux frequency must transition from zero to the ac supply frequency, which is governed by (4.7). A positive q-axis component of ac voltage drives the stator flux frequency towards the ac supply frequency. Thus, minimum perturbation in the stator flux during mode transition is achieved by connecting the ac source to the stator at the instant when the incoming ac voltage vector is at an angle $\delta_{sw}$ relative to the stator flux vector $\vec{\psi}_s$ as shown in Fig. 4.5, where

$$\delta_{sw} = \arccos \left( v_{dc} \cos \delta_{dc} \right).$$

(4.27)
1.2 - Nullcline setting magnitude dynamics to zero.

Nullcline setting angle dynamics to zero

S1.1

0.9

0.8

0

0.7

-1.3 -1.4 1.5

1.6 1.7

6 (rad)

Figure 4.6: Phase plane analysis for finding optimum switching instant during stator source change-over from dc to ac: The stator flux magnitude swing is minimum when the initial condition in (4.28) is at C as compared to A or B. The stator flux transition controller is disabled in this analysis. XY: stator flux magnitude level in low-speed mode.

A generalized approach utilizing the generalized stator-flux transition model can also be used to determine the correct transition instant in cases where the stator flux magnitudes in the two modes are unequal. For example, assuming the stator flux in low-speed mode to be 0.75 p.u. instead of nearly 1 p.u. in the high-speed mode stator flux, the switching instant can place the initial condition for the state trajectory governed by

\[
\frac{1}{\omega_0} \frac{d}{dt} \begin{bmatrix} \psi_s \\ \delta \end{bmatrix} = \begin{bmatrix} -\frac{L_s \psi_s}{x_s} + \cos \delta \\ 1 - \frac{1}{\psi_s} \sin \delta + \frac{1}{\psi_s} \frac{r_d}{x_s} i_{rd} \end{bmatrix} + \begin{bmatrix} \frac{v_d}{x_s} i_{rd} \\ 0 \end{bmatrix},
\]

(4.28)

anywhere on the line XY in Fig. 4.6. Without the stator flux transition controller, i.e. by setting \(i_{rd}\) to be zero, if the transition instant is set at A or B, the state trajectory makes a larger magnitude swing of the stator flux when compared to the transition instant C. Essentially, C is the intersection of the first null-cline with the stator flux magnitude in low-speed mode. This is the identical condition of equating (4.20) to zero. The second condition of positive q-axis voltage is satisfied by the choice of positive \(\delta\).

Therefore, the desired angle for the incoming ac voltage vector relative to the A-phase axis at the instant of mode transition is obtained from Fig. 4.5 and is given by

\[\delta_{opt} = \delta_{sw} - \delta_{dc}.\]

(4.29)

Based on the chosen transfer switch topology, the feasible region for \(\delta_{opt}\) is bounded by the requirement of the natural commutation of the outgoing SCRs in the low-speed SCR-bank, which is represented as
\[ \delta_{\text{opt}} \in [-\epsilon, \epsilon]. \] (4.30)

For the TTB transfer switch, \( \epsilon = 30^\circ \) as represented by the blue sector while for the ETB transfer switch

\[ \epsilon = 60^\circ - \arcsin \left( \frac{v_{dc}}{\sqrt{3}} \right), \] (4.31)

as represented by the green sector where the peak phase voltage of the ac source is chosen as the base voltage for normalization.

During the mode transition, the DFM must be operated within the allowable stator and rotor currents ratings, which implies

\[ i_{sd}^2 + i_{sq}^2 \leq 1 \] (4.32)

and

\[ i_{rd}^2 + i_{rq}^2 \leq I_r^2, \] (4.33)

where \( I_r \) represents the ratio of the rotor current rating (referred to the stator) to the stator current rating. The rotor q-axis current component is represented by \( i_{rq} \) while \( i_{sd} \) and \( i_{sq} \) represent the d and q-axis current components for the stator. Additionally, the maximum voltage that can be impressed on the rotor is limited by the designed voltage rating of the rotor power electronic converter based on the steady-state drive design detailed in Chapter 2. Considering \( v_{rd} \) and \( v_{rq} \) as the d and q-axis components of the rotor voltage, the limit on the allowable rotor voltage \( V_r \) enforces

\[ v_{rd}^2 + v_{rq}^2 \leq V_r^2. \] (4.34)

Constraint (4.32) is mapped as a function \( f \) in the \( \delta - \psi_s \) plane using the DFM machine model in the stator-flux orientation as given in Chapter 2 by

\[ f (\psi_s, \tau, i_{rd}) = \left( \frac{\psi_s - x_m i_{rd}}{x_s} \right)^2 + \left( \frac{\tau}{\psi_s} \right)^2 - 1 \leq 0. \] (4.35)

This constraint is dependent on the stator flux magnitude, drive torque, and the rotor d-axis current input.

Similarly, constraint (4.33) is mapped as a function \( g \) in the \( \delta - \psi_s \) plane by
Finally, the individual components of the rotor voltage is calculated using

$$v_{rd} = r_e i_{rd} + \frac{x_m}{x_s} \cos \delta - \frac{r_s x_m}{x_s} \psi_s - x_e (\omega_s - \omega_T) \left( - \frac{x_s \tau}{x_m \psi_s} \right)$$

and

$$v_{rq} = r_r \left( - \frac{x_s \tau}{x_m \psi_s} \right) + (\omega_s - \omega_T) \left[ x_e i_{rd} + \frac{x_m \psi_s}{x_s} \right]$$

where $\omega_T$ represents the low-speed-to-high-speed mode transition speed. The stator flux frequency is represented by $\omega_s$ and is calculated by

$$\omega_s (\psi_s, \delta, \tau) = \frac{\sin \delta}{\psi_s} r_e - \frac{r_s \tau}{\psi_s^2}.$$

$r_e$ and $x_e$ represents the equivalent resistance and inductance and is given by

$$r_e = r_r + r_s \frac{x_m^2}{x_s^2}, \quad x_e = x_r - \frac{x_m^2}{x_s}.$$  

Using (4.37) and (4.38), the rotor voltage constraint (4.34) is mapped as a function $h$ in the $\delta - \psi_s$ plane and is represented by

$$h (\psi_s, \delta, \tau, i_{rd}) = v_{rd}^2 + v_{rq}^2 - V_r^2 \leq 0.$$  

In the following sections, the developed model along with the constraints projected on the $\delta - \psi_s$ plane are used to evaluate the performance of the ETB and the TTB transfer switch during the mode transition under different operating drive torque conditions.

4.2.2.1 Low-speed-to-high-speed mode transition at rated drive torque condition

At the rated drive torque condition during the desired mode transition, $\tau$ equals 0.498 p.u. for the example DFM as derived in Chapter 2. With a dc source voltage of 0.068 p.u. and a designed stator flux magnitude of 0.75 p.u., full drive torque demand sets the operating point of the DFM in low-speed mode at $(80', 0.75)$ denoted by A in the $\delta - \psi_s$ plane as shown in Fig. 4.7. The optimum transition instant $\delta_{opt}$ for minimizing the stator flux perturbation is calculated as 10' using (4.29). The optimum transition instant lies well within the constraint $\epsilon$ of the TTB transfer switch to enable a natural commutation of the outgoing SCRs in the
Figure 4.7: Stator flux trajectory during dc-to-ac source transition with zero d-axis rotor current command and under rated drive torque condition lies within the constraint boundaries set by the stator current, rotor current, and the rotor power electronics voltage rating. A: Initial operating point in the $\delta - \psi_s$ plane while the DFM is operated in low-speed mode. B: Operating point right after the mode transition. C. Steady-state operating point in the high-speed mode.

low-speed SCR-bank. For the ETB transfer switch, $\epsilon$ is calculated as $\sim 55^\circ$ using (4.31) implying that the dc to ac source transition is equally feasible at the optimum transition instant. After the source transition, the operating point is placed instantly at $(90^\circ, 0.75)$ denoted by B in the $\delta - \psi_s$ plane. The stator flux transition dynamics follows (4.28) to reach the steady state in the ac source connected mode denoted by C. The stator flux transition trajectory shown in Fig. 4.7 is with the d-axis rotor current input set to zero.

The constraints from the stator current rating, rotor current rating, and the voltage rating of the rotor power electronic drive, using (4.35), (4.36), and (4.41) respectively, are mapped to the $\delta - \psi_s$ plane. The entire state trajectory remains within the constraint boundaries during the mode transition ensuring that the low-speed-to-high-speed mode transition is achievable seamlessly for the example DFM under rated drive torque condition using either of the ETB or the TTB transfer switches. Additionally, the rotor d-axis current can be commanded to damp the oscillations in the stator flux magnitude using a full-state feedback controller as described next.

4.2.2.2 Stator-flux transition controller

A stator flux transition controller is designed to affect the trajectory of the states of the simplified model during the mode transition, and to damp the stator flux oscillations. This is a full-state feedback controller with estimated state variables setting the d-axis rotor current command. The non-linear plant model, described by (4.24), is linearized across the steady-state operating point in high-speed mode to design the controller gains.
Figure 4.8: Non-linear simplified plant model valid during transition from dc source to ac source change-over with the stator flux transition controller designed using pole placement method.

Using the pole placement method, the poles of the closed loop plant can be appropriately placed to ensure stability, sufficient damping, and lower bandwidth compared to the d-axis current controller. The complete controller is shown in Fig. 4.8, where gain $K_1$ and $K_2$ correspond to the gains of the stator flux transition controller. High pass filters are used with time period $T_3$, equal to the stator time constant, such that the controller is effective only during the mode transition and does not affect the steady-state rotor d-axis current command. This eliminates the effects of inaccuracy in estimation of the steady-state operating point in high-speed mode. With the designed full state feedback controller, the transition from B to C is without any oscillation as can be seen in Fig. 4.9. As the control input vector in (4.24) has a single non-zero element, the control vectors are effective only along the stator flux magnitude axis in the phase plane, as shown by arrows in Fig. 4.9. The rotor d-axis current limiter has a large impact on the allowable damping and, therefore, on the stator flux trajectory.

4.2.2.3 Low-speed-to-high-speed mode transition at low drive torque condition

Under lightly loaded condition, the demanded drive torque is lower than the rated value for the DFM. For example, assuming a drive torque demand $\tau$ to be 0.174 p.u. (35% of the rated torque in low-speed mode) for the example DFM, the operating point of the DFM in low-speed mode is at $(20^\circ, 0.75)$ denoted by D in the $\delta - \psi_s$ plane as shown in Fig. 4.10. The optimum transition instant $\delta_{opt}$ for minimizing the stator flux perturbation is calculated as $66^\circ$ using (4.29). The optimum transition instant is not feasible to be achieved using either of the TTB or the ETB transfer switches as this would violate the condition required to satisfy the natural commutation for the outgoing SCRs. The condition is depicted in the vector diagram shown in Fig. 4.11 where the optimum transition instant from the perspective of minimizing stator flux transients is achieved when the ac source voltage vector is at $\underline{V}_{grid}$ while lying outside the feasible sectors for natural commutation of the outgoing SCRs. To ensure natural commutation of the SCRs, the transition
nullcline setting magnitude dynamics to zero
nullcline setting angle dynamics to zero
autonomous response during dc to ac source transition
state trajectory with stator flux transition controller
controller effort and direction

Figure 4.9: Stator flux trajectory during dc-to-ac source transition with d-axis rotor current commanded using the stator flux transition controller and under rated drive torque condition. B: Operating point right after the mode transition. C. Steady-state operating point in the high-speed mode.

must be initiated at $\overline{v}_{grid}$ for the ETB transfer switch and at $\overline{v}_{grid}$ for the TTB transfer switch. The natural commutation of the SCR constraint enforces the operating point of the DFM after the source transition to be at $(50^\circ, 0.75)$ denoted by E for the TTB transfer switch and at $(75^\circ, 0.75)$ denoted by F for the ETB transfer switch as shown in Fig. 4.10(a) and (b) respectively.

Using (4.28), the stator flux trajectory with zero d-axis rotor current input for these two initial conditions are also shown in Fig. 4.10. The stator flux magnitude swing is 20% higher for the TTB transfer switch in comparison to the ETB transfer switch. A higher flux swing results in saturation of the magnetic circuit of the DFM affecting the overall performance of the DFM. Moreover, the TTB transfer switch will not be able to provide a seamless torque at the shaft during the mode transition. This can be observed by superimposing the rotor converter voltage constraint of $V_r = 0.52$ p.u. in the $\delta - \psi_s$ plane of Fig. 4.10(a). Multiple points in the state trajectory including the initial condition of E lies outside the rotor voltage constraint boundary implying that the rotor voltage will be limited during the mode transition. The limited rotor voltage leads to limited q-axis rotor current and consequently an uncontrolled drive torque. The constraints for the stator current rating and the rotor current rating are much relaxed in the $\delta - \psi_s$ plane and do not directly affect the drive performance under low drive torque condition. For the ETB transfer switch, all the points in the state trajectory including the initial condition of F lies within the rotor voltage constraint boundary implying that a seamless transition can be achieved with zero d-axis rotor current command or an optimally damped control law as will be discussed next.
Figure 4.10: Stator flux trajectory during the dc-to-ac source transition under low drive torque condition (35% of the rated drive torque). D: Initial operating point in the \( \delta - \psi_s \) plane while the DFM is operated in low-speed mode. G: Steady-state operating point in the high-speed mode. (a) TTB transfer switch: E is the operating point right after the mode transition \((DE = 30^\circ)\). (b) ETB transfer switch: F is the operating point right after the mode transition \((DF = 55^\circ)\).

### 4.2.2.4 Optimum rotor d-axis current input to minimize stator flux transients

The analysis shown in the previous sections explicitly sets the rotor d-axis current to zero to emphasize on the state trajectories during the mode transition with zero control input. However, appropriate rotor d-axis current damps the oscillations in the stator flux within the allowable bounds of the stator current, rotor current and the rotor converter voltage rating. Multiple approaches can be used to optimize the d-axis rotor current input based on the system-specific requirement. For example, one approach is to maximize the damping of the stator flux during a low-speed-to-high-speed mode transition.

Two steps are involved to compute the optimum rotor d-axis current that maximizes the stator flux damping. First, identification of the correct polarity of \( i_{rd} \) that can maintain stability of the DFM during the mode transition and second, maximizing \( i_{rd} \) that can be impressed along the state trajectory to damp the stator flux within the allowable limits. As the rotor d-axis current affects the state trajectory only along the \( \psi_s \) axis, as evident from (4.28), the stabilizing polarity of \( i_{rd} \) at each instant is such that the state trajectory is driven inwards towards the steady-state operating point G in Fig. 4.10. The null-cline obtained by equating \( d\delta/dt \) in (4.28) to zero is shown as by the line segment PQ in Fig. 4.12(a). For the operating points below the null-cline PQ, a non-negative rotor d-axis current results in forcing the state flux to inner trajectories in the phase plane. Conversely for the operating points above the null-cline PQ, a non-positive rotor d-axis current ensures that the stator flux is drawn towards the inner trajectories in the phase plane.

The maximum rotor d-axis current that can be impressed to damp the stator flux transients is governed
Figure 4.11: Commutation diagram during low-speed-to-high-speed mode transition under low drive torque condition. From the DFM perspective, a dc-to-ac source transition instant when the ac source voltage vector is at $\vec{v}_{\text{grid}}$ incurs least perturbation in the stator flux. However, constraints for the natural commutation of the outgoing SCRs in the low-speed SCR-bank enforces the transition instant to be at $\vec{v}_{\text{grid}}$ for the ETB transfer switch and at $\vec{v}_{\text{grid}}$ for the TTB transfer switch respectively.

by the constraints described by (4.35), (4.36), and (4.41). For the example DFM with a 35% drive torque demand during the mode transition, the optimum d-axis rotor current is shown in Fig. 4.12(b). During the sections FX and YG in the state trajectory, a positive d-axis rotor current is commanded while in the section XY, a negative rotor d-axis current is used to damp the stator flux oscillation. The magnitude of the rotor d-axis current is governed by the rotor voltage constraint in the section FX as shown in Fig. 4.12(c).

In the section XY, the magnitude of the rotor d-axis current is set by the stator current constraint while in the section YG, the limit on the rotor current plays the dominant role in determining the magnitude of the rotor d-axis current. The above outlined procedure can be repeated to find the optimal d-axis rotor current that provides maximum stator flux damping for other drive torque conditions.
Figure 4.12: Maximizing rotor d-axis current input to damp stator flux oscillation during dc-to-ac source transition. (a) Comparison of the stator flux trajectories with and without d-axis rotor current damping under light drive torque condition. (b) Optimal rotor d-axis current input ensuring that the constraints of the stator current, rotor current, and rotor converter voltage rating are satisfied at all operating points of the damped trajectory. (c) The constraint functions $f$, $g$, and $h$ during the dc-to-ac source transition that remain within the respective limits for the example DFM.
4.2.2.5 Boundary of low-drive torque demand

The boundary that defines a low-drive-torque demand for the thyristor-based transfer switch is obtained when the synchronizer requirement and the commutation constraints are marginally satisfied and is given by,

\[ V_{dc} \cos \delta_{dc,\text{min}} = \cos \delta_{sw} \]

and \( \delta_{sw} - \delta_{dc,\text{min}} = \epsilon \),

where \( \delta_{dc,\text{min}} \) corresponds to the angle between the stator A-phase and the stator flux vector. Solving the two equations in (4.42) for \( \delta_{dc,\text{min}} \) gives,

\[ \delta_{dc,\text{min}} = \arctan \left( \frac{1}{\sin \epsilon} (\cos \epsilon - v_{dc}) \right). \]

(4.43)

For example, with a dc source voltage 20 V and an ac source voltage 146 V (rms, line-line), the boundary of the low-drive torque for the ETB transfer switch is achieved when \( \delta_{dc,\text{min}} = 28^\circ \). For the same dc and ac source voltage magnitudes, the equivalent boundary in the TTB transfer switch corresponds to \( \delta_{dc,\text{min}} = 56^\circ \). This implies that the ETB transfer switch enables optimum transition from the DFM perspective along with the natural commutations of the SCRs for even lighter load compared to that by the TTB transfer switch.

The minimum torque below which the demanded drive torque is termed “low” can be written in terms of the stator flux magnitude in low-speed mode \( |\psi_s| \), stator current vector magnitude \( |i_s| \), and the number of pole of the DFM \( P \) as,

\[ \tau_{low,\text{min}} = |\psi_s|i_s \sin \delta_{dc,\text{min}}. \]

(4.44)

4.2.3 Low-speed-to-high-speed mode transition in LSI topology

There are two criteria to be satisfied for a bump-less low-speed-to-high-speed mode transition from the DFM perspective as explained in Sec. 4.2.2. First, the d-axis component of the incoming ac voltage vector must be aligned with the existing d-axis stator voltage prior to the mode transition. Second, the q-axis component of the incoming ac source voltage must be a positive quantity. Prior to mode transition, the stator voltage is zero due to the shorted configuration in the LSI topology. With no source of magnetizing current from the stator side, the stator current vector is always orthogonal to the stator flux vector. A positive drive torque implies that the stator current vector leads the stator flux vector by 90° as shown in Fig. 4.13.

For the DFM rotor rotating in the counter-clockwise direction relative to the stator ABC winding reference frame for a positive drive torque, the stator current vector and the stator flux vector rotate in the
clockwise direction at slip frequency. The incoming ac voltage vector, however, rotates in the counterclockwise direction at ac source frequency. The in-phase requirement can be achieved at least once every fundamental ac source cycle and automatically satisfies the constraint for natural commutation of the outgoing SCRs for any positive drive torque for the TTB transfer switch. In general, the low-speed-to-high-speed mode transition inherently involves a positive drive torque requirement at the shaft.

For the ETB transfer switch, the mode transition requirement from the DFM perspective is achievable along with simultaneously satisfying the commutation constraints for the outgoing SCRs for all the conditions where the filled-blue sectors overlap with the hatched sectors in the commutation diagrams of Fig. 3.8. While the overlapping of the two sectors forms the major share of the fundamental ac cycle, there exists, however, four sub-sectors 30° to 60°, 120° to 150°, 210° to 240°, and 300° to 330° in the commutation diagram when both the conditions of minimal stator flux perturbation and natural commutation of the outgoing SCR cannot be simultaneously satisfied. If the stator current vector is located within these four forbidden sectors at the instant a mode transition is demanded, either the transition has to be delayed such that the stator current vector is outside of these sectors or a non-optimum transition instant has to be chosen to satisfy the natural commutation requirement.

4.2.4 High-speed-to-low-speed mode transition in LSS topology

The two major differences in the high-speed mode of DFM operation compared to that of low-speed mode operation in the LSS topology are, first, the angle between the stator voltage and the stator flux (δ) is
relatively fixed and is close to 90° independent of the mechanical load (assuming small stator resistance). Second, the stator voltage $V_{\text{grid}}$, the stator current $I_s$, and the stator flux $\Psi_s$ vectors rotate at ac-source frequency relative to the stationary dc-source voltage vector $V_{\text{dc}}$ and the stator-winding axis. For the TTB transfer switch, the stator voltage and current must be at least 120° and at most 240° out of phase for the ac-to-dc source transition to occur with the high-speed SCR-bank undergoing natural commutation while for the ETB transfer switch the stator power factor angle must be between 90° and 270°. This implies that the active power flowing to the stator must be negative during the transition which is inherently achieved by the drive operation. The mode transition of ac-to-dc is necessary when the drive speed goes below the designed transition speed which requires a braking action and, therefore, a reverse power flow in the stator. The braking torque is commanded using the rotor $q$-axis current. The remaining requirement on maintaining the stator power factor can be achieved using the reactive power controller on the rotor power electronics.

The reactive-power controller commands the rotor $d$-axis current to create the required phase between the stator voltage and the current for natural commutation of the outgoing ac-side SCRs (conducting bank).

With properly controlled rotor currents, the ac-to-dc source transition can be initiated during any part of the ac cycle. This flexibility is due to the presence of anti-parallel SCR-bank on the dc source side that can potentially shorten the ac-to-dc source-transition time. However, a careful choice of transition instant is necessary to ensure smooth transition of the stator flux (from ac to dc) of the DFM and, therefore, a controlled torque output at the shaft. To illustrate the correct transition region for ac-to-dc source transition from a DFM perspective, the state-plane analysis is used. The non-linear DFM plant during ac-to-dc source transition can rewritten using (4.24) as,

$$
\frac{d}{dt} \begin{bmatrix} \Psi_s \\ \delta \end{bmatrix} = \begin{bmatrix} \frac{R_s}{L_s} \Psi_s + V_{\text{dc}} \cos \delta \\ \frac{V_{\text{dc}} \sin \delta}{\Psi_s} + \frac{1}{\Psi_s^2} \frac{M R_s}{L_s} \Psi_s \theta \end{bmatrix} + \begin{bmatrix} \frac{M R_s}{L_s} \\ 0 \end{bmatrix} I_{rd}. 
$$

(4.45)

The initial condition of the stator flux magnitude $\Psi_s$ equals to the high-speed mode stator flux magnitude prior to the transition while the target stator flux magnitude is $\Psi_{s,\text{low}}$ based on the DFM drive design. Similarly, the initial condition for the angle $\delta$ is determined by the switching instant from the ac source to the dc source. After the transition to the dc source, the rotor $d$-axis current is commanded based on the stator flux controller with a output limiter that follows a control law given by,

$$
I_{rd}^* = (K_p + \frac{K_i}{s}) [\Psi_{s,\text{low}} - \Psi_s] - \frac{L_s}{M R_s} V_{\text{dc}} \cos \delta; \quad |I_{rd}^*| < I_{rd,\text{limit}}. \quad (4.46)
$$

The assumption of high rotor-current controller bandwidth allows to couple (4.45) and (4.46) by,
Figure 4.14: Phase plane analysis: Determination of the desired switching instant during the ac-to-dc source transition. Improper switching instant can lead to a collapse of the stator flux magnitude, leading to an uncontrolled shaft torque.

\[ I_{rd} = I_{rd}^* \in [-I_{rd,limit}, I_{rd,limit}] \]  

(4.47)

The non-linear dynamics for the state trajectories are shown in Fig. 4.14. It is assumed that the initial ac-mode stator-flux magnitude is unity, while the final stator flux magnitude in low-speed mode is 0.72 p.u. based on a drive design. Based on the instant of transition, governed by the initial \( \delta \), the states for the non-linear system described by (4.45), (4.46), and (4.47), follows different trajectories during the high-speed-to-low-speed mode transition. The steady-state condition for the angle \( \delta \) is determined by the demanded braking torque as given by (4.30). As can be seen in Fig. 4.14, the least perturbation in the stator flux during the transition occurs when the initial \( \delta \) is within the stable zone marked in Fig. 4.14. For other switching instances, the stator flux magnitude completely collapses leading to a significant uncontrolled shaft torque behavior. Therefore, for a guaranteed stable ac-to-dc transition, the desirable range of the stator flux angle at the instant of transition must be within zero and 90°. With the stator flux lagging the stator voltage nearly by 90°, the corresponding suitable region is when the ac source voltage vector is within 0° and 90° with respect to the stator A-phase axis.

The requirement of reverse active-power flow necessary for the natural commutation of the ac-side SCRs might not be satisfied under certain drive-torque requirements. For example, a gradual ramp-down command for the drive reference speed, or a mode transition initiated because of load disturbances does not necessarily need a negative drive torque at the transition speed. In these cases, either a forced commutation circuit should be added on to the ac source side SCRs, or a secondary transition speed can be designed depending on the applications. If the rotor speed goes below the designed secondary-transition speed, the speed controller
is disabled momentarily with deliberate, sufficient braking torque commanded to the drive. This ensures proper conditions necessary for the natural commutation of the ac source side conducting bank. The speed controller comes back into the control loop after the mode transition. Fluctuations to drive speed are insignificantly perturbed during this short maneuver because of the presence of the mechanical inertia of the drive.

4.2.5 **High-speed-to-low-speed mode transition in LSI topology**

With the DFM initially operating in the high-speed mode, the constraint of the stator power factor angle for the natural commutation of the outgoing SCRs can be satisfied by a reverse active power flow and with an appropriate stator reactive power control using the rotor converter command. The stator flux trajectory at the instant of the high-speed-to-low-speed mode transition in the LSI topology is shown in Fig. 4.15. Assuming the initial normalized stator flux magnitude is unity and the final stator flux magnitude for low-speed mode is 0.72 p.u., the stator flux trajectory follows a different path based on the transition instant set by the initial angle \( \delta \). Unlike in the LSS configuration, in which the initial \( \delta \) has to be within zero and 90° for a stable high-speed-to-low-speed mode transition, ac-source-to-shorted-stator configuration change-over is always stable in the LSI topology, irrespective of the initial \( \delta \) set by the mode transition instant as shown in Fig. 4.15. This implies that the ac source voltage vector can be anywhere relative to the stator winding during high-speed-to-low-speed mode transition. The direction of the state trajectory is opposite to that of the vector field shown in Fig. 4.4 due to the negative drive torque demand during braking. After the transition, the steady-state condition corresponds to the line \( XY \) in Fig. 4.15.
4.2.6 Operation of the transfer switch and phase change over relay

Based on the above analysis, a flowchart of the transfer-switch operation in the LSS topology is shown in Fig. 4.16. A hysteresis band is introduced around the designed mode-transition speed to prevent chattering between the operating modes. The absolute rotor speed $|\omega_r|$ must be greater than $\omega_{T1}$ for a low-speed to high-speed mode transition. An ac-to-dc source transition is initiated when $|\omega_r|$ is below $\omega_{T2}$, with $\omega_{T1} > \omega_{T2}$. For cases where the braking torque is insufficient to cause reverse power flow in the stator even with the drive speed lower than $\omega_{T2}$, a secondary transition speed $\omega_{T3}$ is selected, where $\omega_{T2} > \omega_{T3}$. As the rotor speed reaches $\omega_{T3}$ while decelerating, being in high-speed mode, a braking torque is commanded at the shaft with the speed controller disabled. The delay $T_d$ introduced between enabling gate signals to the succeeding SCR-bank and the concluding SCR-bank depends on the turn-off time of the chosen SCRs.

The phase-sequence change-over relay, connected between the ac source and the SCR-based transfer switch to extend the DFM drive operation range to all four quadrants, operates only during the low-speed-mode operation of the drive with a hysteresis band around zero speed. The ac source phase sequence is commanded to be $ABC$ for rotor speed greater than $\omega_{T0}$ and $ACB$ for drive speed less than $-\omega_{T0}$ where $\omega_{T0} < \omega_{T3}$. This prevents on-load operation of the relay. As an alternative, the relay can be replaced with two additional anti-parallel SCRs cross-connecting the phases of the DFM to the ac source. This also increases operational reliability for applications that need frequent four-quadrant operation, replacing the mechanical operation of relay contact surfaces.
4.3 Experimental results

The performance of the switched-DFM drive along with the proposed control architecture is evaluated using the experimental setup described in Appendix A. In the following sections, different experimental results are highlighted, which expose the behavior and response of the switched-DFM drive for a wide variety of operating conditions.
4.3.1 Drive response for different reference speed command and with different load torque profile

The switched-DFM drive is subjected to different speed command and load torque profile to evaluate the performance and stability of the drive under different operating conditions. A electronic-load bank connected to a permanent-magnet synchronous generator, which is mounted on the same shaft as the DFM, effectively loads the mechanical port of the DFM. The load bank is programmed to draw current proportional to the square of the rotor speed, which emulates a quadratic load-torque curve similar to a typical ship propulsion load. The step response of the DFM drive for the full speed range, four-quadrant operation is shown in Fig. 4.17(a). As evident, the speed of the DFM is controlled seamlessly across the complete speed range. In a ship, this kind of requirement can arise during crash-back situations. The braking torque in the high-speed mode is commanded such that the drive can operate at unity power factor as required for the natural commutation of the high-speed SCR-bank.

To evaluate the performance of the drive under reduced-drive-torque requirement at the transition speeds, particularly for the LSS topology, a ramp is commanded in the reference speed with a slope of 360 rpm/s. The speed response of the drive is shown in Fig. 4.17(b). During acceleration the demanded drive torque at the low-speed-to-high-speed mode transition instant is 2.2 Nm while the “low drive torque” limit for the prototype drive calculated using (4.33) is 2.6 Nm. During deceleration at instant B, the rotor speed goes below the chosen secondary transition speed of 648 rpm. The commanded drive torque required, even at this rotor speed, is still positive such that the ramp down in speed can be maintained. In this case, the requirement of reverse power flow for the natural commutation of the ac side SCRs cannot be met. A short pulse of a controlled negative torque ensures the correct condition for the natural commutation of the outgoing ac-source-side SCRs and a source transition to dc source is achieved. The speed response of the DFM is unperturbed and the ramp speed response is achieved without distortion.

Next, the drive is subjected to an alternating reference speed for the LSS topology across the complete speed range as shown in Fig. 4.17(c). The reference speed oscillates with a step size of 400 rpm at 0.5 Hz across the entire speed range. The speed response profiles A and B are completely in high-speed mode with positive ac-source frequency. The profiles D, E, and F are completely in low-speed mode, and the profiles H and I are again in high-speed mode with stator ac-source phase sequence reversed. The speed profiles C and G correspond to the case when the DFM drive alternates between the low-speed and high-speed modes across the designed transition speed. As can be seen, the mode transitions are seamless even with an alternating reference speed. The settling time of the speed response increases with the operating speed because of the increase in the load torque (as programmed using the electronic load bank).
Figure 4.17: Performance of the proposed switched-DFM drive subjected to different operating conditions. (a) Speed response of the DFM for step command in the reference speed. (b) Speed response of the DFM for a ramp command in the reference speed. (c) Speed response of the DFM for alternating reference speed command with a step size of 400 rpm. (d) Speed response of the DFM under load torque disturbance, ranging between no-load and 70 percent of the rated load.

Finally, the drive is subjected to a load disturbance of zero to 70 percent of the rated load at different operating speeds for the LSS topology as shown in Fig. 4.17(d). The speed response profile A remains completely in low-speed mode for the load disturbance, and the profiles C, D, E, and F are completely in high-speed mode. For profile B, the reference speed is set at 684 rpm, which is midway between the dc-to-ac source transition speed (720 rpm) and secondary dc-to-ac source transition speed (648 rpm). As the load goes to zero, the rotor speed overshoots beyond 720 rpm, leading to a dc-to-ac source transition. Eventually, the speed settles to 648 rpm as the speed controller pulls back the operating drive speed to the reference speed but now operating in high-speed mode. In the following instant, when the drive is subjected to the 70 percent rated load, the rotor speed falls below the secondary transition speed of 648 rpm, leading to an
ac-to-dc source transition. In this case, the stator-flux magnitude oscillates between the designed dc-mode level and the ac-mode level, as set by the ac source. The drive performance is stable and equivalent without any notable differences under all operating regimes.

4.3.2 Drive torque performance during low-speed-to-high-speed mode transition in LSS topology using an emulated ac source

The drive torque performance during the low-speed-to-high-speed mode transition in LSS topology is shown in Fig. 4.18(a). The ac source in this test is emulated by an inverter (Texas Instruments High Voltage Motor Control & PFC Developer's Kit) that is programmed to create either an ac source of 134 V, 40 Hz or a dc source of 20 V through open-loop duty ratio command such that the control architecture for the switched-DFM drive can be evaluated without any second-order effects of source harmonics. The torque limit is changed at the DFM stator time constant ensuring a smooth transfer from the dc-to-ac source transition. Additionally, the acceleration test is repeated with identical low-speed mode and high-speed mode torque limits as shown again in Fig. 4.18(a). In both the cases, the electromagnetic torque undergoes a near “bumpless” transition as the mode changes from the low-speed to high-speed mode. Figure 4.18(b) illustrates the transition of the estimated stator flux frequency from zero to 251 rad/s. The transient in the estimated stator flux frequency is critical for providing the feed-forward terms of the current controller to ensure proper reference tracking of the individual axis currents as shown in Fig. 4.18(c). The q-axis reference current is tracked properly by the q-axis feedback current even during mode transition. Perturbations in the stator flux magnitude results in perturbation of the q-axis current reference for a commanded torque. The frequency of these perturbation is in the order of ac source frequency.
4.3.3 Drive torque performance during low-speed-to-high-speed mode transition in LSS topology using realistic ac source

The performance of the overall switched-DFM drive under different drive torque conditions in the LSS topology is evaluated next by using two synchronous generators operating in parallel as the ac source. The drive torque command is kept constant during the low-speed-to-high-speed mode transition at six different torque magnitude levels $a - f$ as shown in Fig. 4.19(a). The estimated drive torque has different harmonic content in low-speed mode compared to in high-speed mode, due to the presence of harmonics in the ac
source voltage that is produced by the off-the-shelf synchronous generators. In low-speed mode, the DFM is designed to operate with the stator flux magnitude of 0.32 V-s while in high-speed mode is 0.42 V-s. The operating stator flux magnitude in high-speed mode is governed predominantly by the ac source voltage and frequency. The stator flux magnitude and the angle between the stator flux vector and the stator voltage vector, \( \delta \), undergoes transition at the instant of source change-over as shown in Fig. 4.19(b) and (c). As pointed out earlier, angle \( \delta \) depends on the drive torque demand in low-speed mode of the LSS topology while is fairly constant near 90° in the high-speed mode. The stator-flux magnitude perturbation is higher for a lower-drive torque demand as expected. This is due to the non-optimum transition enforced by the allowable region of the ac source voltage vector location due to the SCR commutation constraint. Figure 4.19(d) shows the orientation of the normalized stator flux vector and the ac source voltage vector at the instant of low-speed-to-high-speed mode transition for the six demanded drive torque cases. The stator flux vector is nearly orthogonal to the ac source voltage vector for an optimum transition for the drive torque cases a – c. The optimum transition instant for the ac source voltage vector falls well within the allowable constraints for the SCR commutations. However, for the remaining drive torque-cases d – f, the ac source voltage vectors are staggered together to ensure that the SCR commutation constraints are satisfied even if the transitions happen at non-optimum instances from the DFM perspective. A safety margin of 7° is provided to the allowable computed theoretical region of ±55.5° to ensure that the noises in the measurement and the harmonics in the source voltage do not lead to uncertainty during mode transition leading to the shorting of the sources.
Figure 4.19: Drive torque performance during the low-speed-to-high-speed mode transition in LSS topology using two synchronous generators operating in parallel as the ac source: (a) estimated drive torque, (b) stator flux magnitude transition, (c) transition of the angle between the stator flux vector and the stator voltage vector, δ, (d) stator flux and ac source voltage vectors location relative to the stator ABC winding axis at the instant of mode transition.

4.3.4 Stator flux trajectory during mode transitions

The phase-plane analysis used to ascertain the stator flux trajectory during mode transitions is validated experimentally. Figure 4.20(a) shows the stator flux transition in the δ - ψs plane during dc-to-ac source change over in the LSS topology. A1, A2, and A3 represent three initial operating points in low-speed mode under three different drive torque demands, which are 56%, 74%, and 94% of the low-speed drive torque capability respectively. Using the ETB transfer switch, the operating points are moved to B1, B2, and B3 respectively after the stator of the DFM is connected to the ac grid to ensure that the outgoing low-speed SCR-bank undergoes natural commutations and minimum perturbation is incurred by the stator flux swing.
Figure 4.20: Stator flux behavior during mode transitions: (a) Stator flux transition during low-speed-to-high-speed mode transition in the LSS topology under different drive torque condition where $A_1$, $A_2$, and $A_3$ are the initial operating points with the DFM stator connected to the dc source, $B_1$, $B_2$, and $B_3$ are the operating points post transition to the ac source, and $C_1$, $C_2$, and $C_3$ are the steady-state operating points in high-speed mode respectively. (b) Stator flux trajectory during high-speed-to-low-speed mode transition in the LSS topology.

Finally, the stator flux settles down to steady-state operating points $C_1$, $C_2$, and $C_3$ respectively. As expected from the analytical framework, experimental results show similar behavior in the stator flux transition during low-speed-to-high-speed mode transition. Under lower drive torque demand, $A_1$ in this case, the transition instance is set by the SCR commutation requirement rather than optimum instance for least perturbation of the stator flux post-transition.

Alternatively, Fig. 4.20(b) shows the stator flux transition in the $\delta - \psi_1$ plane during ac-to-dc source change over in the LSS topology. Different initial conditions of $\delta$, depending on the switching instant, lead to the different trajectories of the stator flux magnitude from the ac-mode level to the dc-mode level. As expected, for the Case $A$, the stator flux collapses during the ac-to-dc transition, leading to uncontrolled shaft torque.

### 4.3.5 Transient performance comparison of different switched-DFM topologies

The two switched-DFM drive topologies dynamic performances are identical in individual operating modes. However, the response of the switched-DFM drive during transients in particular during the mode transition instant depends on the drive topology. As predicted by the analytical framework, the LSS topology of the switched-DFM drive has to operate under mode stringent constraints compared to the LSI topology to ensure natural commutations of the outgoing SCR-bank along with satisfying the requirement from the DFM perspective for a seamless transition. The following experiments in which the two topologies are compared
Figure 4.21: Transient performance comparison of different switched-DFM topologies (a) Comparison of the switched-DFM drive topologies based on the speed response to an alternating reference speed with a constant load torque around the mode transition speed. The responses of the two topologies are identical. (b) Comparison of the switched-DFM drive speed response to a constant reference speed with a load torque variation between 18% to 55% of full-load torque. The speed perturbation enforces the drive to operate in the two modes alternatively. The LSI topology exhibits a smaller swing in the rotor speed during the mode transition due to the relaxed transition requirements.

for their transient performances near the mode transition speed validates this argument. The drive response for an alternating reference speed command encompassing the mode transition speed is compared in Fig. 4.21(a). With an identical constant load torque at the shaft, both the topologies speed response is identical. A higher overshoot during high-speed-to-low-speed mode transition in comparison to low-speed-to-high-speed mode transition is due to the requirement of the reverse power flow to achieve the natural commutation of the outgoing SCRs.

In the next experiment, the reference speed is set midway between the lower and higher threshold of the hysteresis comparator that commands the mode transition. An oscillating load between 18% and 55% of the full load makes the perturbation in the rotor speed such that the DFM drive operates in the low-speed and high-speed modes alternatively. The LSS topology exhibits a larger swing in the rotor speed due to the load disturbance as it has more constraints to satisfy for proper mode transition as shown in Fig. 4.21(b).

4.4 Summary

This chapter discussed and demonstrated a control architecture for a switched-DFM drive. The seamless transition of the DFM along with the natural commutation of the SCR-based transfer switch enables a smooth or “bumpless” mechanical operation of the switched-DFM drive over a wide range of drive torque-
speed requirements. The results show that the proposed controller works equivalently under the tested operating conditions (low-speed mode, high-speed mode, and during mode transitions). The results also validate the operation of the stator flux estimator, mode-transition controller, reactive power controller, and stator flux transition controller. A more robust controller may include an on-line parameter estimation scheme to not only update the DFM parameters, but also tune the controller for variable load friction and inertia. Finally, proper switching between the low-speed and high-speed modes within the constraints set by the thyristor-based transfer switch and a controlled flux transition with reactive power control (during the high-speed mode) opens up opportunities for system-level design choices to be made, which can minimize weight and cost of the propulsion drive.
Chapter 5

Grid Interactions

The switched-DFM drive has so far been shown to operate with seamless shaft behavior in and through low-speed and high-speed modes with appropriate control on the rotor-side converter and mode transition instant of the stator transfer switch. However, the drive must also interact with the ac source seamlessly irrespective of the operating modes. This is critical for the stability of the ac grid as well as for the intermediate dc-link between the grid-side and the rotor-side converters. Assuming an ideal DFM (with no resistive losses and leakages) producing a rated drive torque of 1 p.u., the active power required in the stator during low-speed operation is zero and the shaft power equals the rotor power as shown in Fig. 5.1. At the transition instant (denoted by $P$ in Fig. 5.1) from low-speed to high-speed mode, the active power in the stator of the DFM changes from zero to 1 p.u. Instantaneously, the rotor back-to-back converter withdraws power from the rotor to ensure a mechanically "bumpless" transition at the shaft. The power withdrawn from the rotor must be seamlessly transferred to the ac source at the point of common-coupling (PCC) between the transfer switch and the rotor power electronics such that the active power drawn from the ac source is "bumpless" as well. A converse scenario appears with respect to the power consumption from the ac source during the braking of the drive with a torque of $-1$ p.u. as shown in Fig. 5.1.

A stiff ac source with a sufficient active power reserve can sustain a sudden change in power demand if the active power balance at the PCC is not met during mode transition. The seamless grid interaction at the PCC becomes more critical in applications where the switched-DFM drive may consume a major share of the generated ac power or in cases where a large inductive line exists between the PCC and the ac source. For example, in an islanded ship microgrid, the propulsion drive may consume as much as ninety percent of the ship ac generator capability [49]. In this case, a large swing in power requirement can destabilize the ac grid. Similar arguments are valid for the reactive power requirement of the switched-DFM drive in the two
The switched-DFM drive interacts with the ac grid through the rotor under all operating modes and through the stator only during high-speed mode. The two degrees of freedom of the grid-side converter are used to maintain the intermediate dc-link voltage and provide additional reactive power support to the grid, if
necessary. This section highlights the sizing and control of the grid-side converter to ensure a seamless grid interaction of the overall drive under different operating conditions.

### 5.1.1 Grid-side converter sizing analysis

Denoting $V_b$ (peak per-phase) to be the required maximum rotor voltage referred to the stator, $N_M$ to be the rotor-to-stator winding turns-ratio of the DFM, and assuming that the rotor-side converter operates as a two-level voltage source converter operating with conventional space-vector modulation, the required nominal dc link voltage, $V_{dc,nom}$ to operate the drive over the full speed range is given by,

$$V_{dc,nom} = \sqrt{3}V_b N_M G_{dc}, \quad (5.1)$$

where $G_{dc}$ is an extra margin provided to the dc-link voltage and is typically in the order of $1.04 - 1.1$. More often than not, a grid-side transformer interfaces between the grid-side converter and the ac source. Denoting the nominal duty ratio of the grid-side converter to be $D_{FE,nom}$, the nominal grid voltage magnitude to be $V_{g,nom}$ (peak per-phase), and assuming that the grid-side converter is also a two-level voltage source converter operating with conventional space-vector modulation, the required turns-ratio of the grid-side transformer ($N_T$) for proper grid interfacing is given by,

$$N_T = \frac{D_{FE,nom} V_b N_M G_{dc}}{V_{g,nom}}. \quad (5.2)$$

For an appropriate design of the DFM with a carefully-selected $N_M$, the required turns-ratio $N_T$ can be reduced to unity, implying possible elimination of the grid-side transformer as will be illustrated in Chapter 6. For an active power $P_r$ to be supplied to the DFM rotor from the rotor-side converter, the required dc link current is:

$$I_{dc} = \frac{P_r}{\eta_{de} V_{dc,nom}} \quad \text{for } P_r > 0 \quad (5.3)$$

and

$$I_{dc} = \frac{\eta_{de} P_r}{V_{dc,nom}} \quad \text{for } P_r < 0,$$

and the required current to be drawn from the ac grid by the grid-side converter is:

$$I_f = \frac{2}{3} \frac{\eta_{fe} P_r}{\eta_{de} V_{g,nom} N_r V_{g,nom}} \quad \text{for } P_r > 0 \quad (5.4)$$

and

$$I_f = \frac{2}{3} \frac{\eta_{de} P_r}{N_r V_{g,nom}} \quad \text{for } P_r < 0,$$

where $\eta_{de}$ and $\eta_{fe}$ are the efficiencies of the rotor-side and the grid-side converters respectively. Additionally, the grid-side converter may be designed to handle a part of the DFM magnetizing current or provide
additional grid-side reactive power support based on the DFM design or the application requirement.

Assuming constant dc-link and ac grid voltages, the dc-link current and the grid-side converter current injected to the ac grid are a scaled version of the rotor active power. In conclusion, the grid-side converter voltage rating is governed by the required dc link voltage for the drive to operate over the full speed range. In addition, the current rating is governed by the required rotor power and the additional reactive power support, if necessary.

**5.1.2 Grid-side converter control**

The grid-side converter interfaces the ac grid through a filter and the grid-side transformer as shown in Fig. 5.2. The filter and the grid-side transformer can be modeled as a lumped series resistance \( R_f \) and an inductance \( L_f \) for control purposes. In the d-q reference frame, with the d-axis aligned with the ac source voltage vector, the system equations governing the power flow between the ac-grid and the grid-side converter are given by:

\[
V_{fd} = L_f \frac{dI_{fd}}{dt} + R_f I_{fd} - \omega_g L_f I_{fq} - V_{grid}, \tag{5.5}
\]

and

\[
V_{fq} = L_f \frac{dI_{fq}}{dt} + R_f I_{fq} + \omega_g L_f I_{fd}, \tag{5.6}
\]

where \( V_{grid} \) is the ac-grid voltage magnitude (peak per-phase), \( \omega_g \) is the ac-grid frequency, \((V_{fd}, V_{fq})\) are the d-axis and q-axis voltages, respectively, that are commanded by the grid-side converter, and \((I_{fd}, I_{fq})\) are the d-axis and q-axis currents, respectively, fed by the grid-side converter to the grid. The required reference
frame angle $\theta_g$ for the d-q transformation of the measured grid phase voltages ($V_{ag}$, $V_{bg}$, and $V_{cg}$) and the inverter currents ($I_{ai}$, $I_{bi}$, and $I_{ci}$) is obtained from a three-phase phase-locked loop (PLL) designed to track the grid voltage and frequency [52] as shown in the control architecture for the grid-side converter in Fig. 5.3.

The inner-loop proportional-integral (PI) current controllers are designed based on (5.5) and (5.6) with appropriate feed-forward terms to eliminate the effect of the cross-coupling terms due to $L_f$ and the grid-voltage magnitude. The reference command to the inner loop d-axis current controller is set by the outer loop dc-link voltage controller. The q-axis current controller is commanded based on the required reactive power support from the grid-side converter. The coordination between the grid-side and the rotor-side converters is taken into consideration during the design of the outer loop controllers to ensure a seamless mode transition of the switched-DFM drive relative to the ac-grid.

The dc-link voltage $V_{dc}$ is set by the net charging current to the dc-link capacitor $C_{dc}$, which is the difference between the current drawn from the grid-side converter $I_{dcf}$, the current delivered to the rotor-side converter $I_{dcr}$, and the current in the braking resistor $I_{brake}$ as shown in Fig. 5.2. The dynamical equation governing the dc-link voltage is given by,

$$C_{dc} \frac{dV_{dc}}{dt} = I_{dcf} - I_{dcr} - I_{brake}. \quad (5.7)$$
Neglecting ripple currents at switching frequency, the active power $P_r$ drawn by the rotor of the DFM from the dc link can be related to the rotor-side converter current $I_{rdc}$ by,

$$P_r = \eta_{de} V_{dc} I_{rdc} \quad \text{for } P_r > 0$$

and

$$P_r = \frac{1}{\eta_{de}} V_{dc} I_{rdc} \quad \text{for } P_r < 0.$$

(5.8)

Defining $S_{brake}$ as the switching function of the braking resistor $R_{brake}$ connected to the dc link, the current in the braking resistor is given by,

$$I_{brake} = S_{brake} \frac{V_{dc}}{R_{brake}}.$$

(5.9)

Therefore, the current $I_{rdc}$ and $I_{brake}$ in (5.7) can be substituted using (5.8) and (5.9) as:

$$I_{dfc} = C_{dc} \frac{dV_{dc}}{dt} + \left( \frac{P_r}{\eta_{de} V_{dc}} \right) + S_{brake} \frac{V_{dc}}{R_{brake}} \quad \text{for } P_r > 0$$

and

$$I_{dfc} = C_{dc} \frac{dV_{dc}}{dt} + \left( \frac{P_r}{\eta_{de} V_{dc}} \right) + S_{brake} \frac{V_{dc}}{R_{brake}} \quad \text{for } P_r < 0.$$

(5.10)

The dc-link voltage controller is designed using (5.10) with appropriate feed-forward terms that command the required the grid-side converter current $I_{dfc}^*$. The feed-forward terms include the active power supplied to the rotor of the DFM by the rotor-side converter and the switching function of the braking resistor. The feed-forward terms help in minimizing the delay in dc-link power balance in case a step perturbation in active power requirement is demanded by the switched-DFM drive during the mode transition. Similar control strategies have been widely used in the literature for generic ac-to-dc converter control [53, 54, 55].

The relationship between the output of the dc-link voltage controller $I_{dfc}^*$ and the reference d-axis converter current $I_d^*$ can be established through the active power balance across the grid-side converter and is given by,

$$V_{dc} I_{dfc}^* = -\frac{3}{2} \eta_{fe} V_{grid} I_d^* \quad \text{for } I_{dfc}^* > 0$$

and

$$V_{dc} I_{dfc}^* = -\frac{3}{2} \eta_{fe} V_{grid} I_d^* \quad \text{for } I_{dfc}^* < 0.$$

(5.11)

The negative sign appears due to the assumed direction of the grid-side converter current. For simplicity, the conditional gains and feed-forward terms that depend on the direction of the power flow are not explicitly shown in the control architecture in Fig. 5.3.

In a ship or a locomotive micro-grid, with the propulsion load power taking the major share of the total power generation capability, a complete regenerative braking using the grid-side converter may destabilize the ac grid. It is necessary to share the power recovered from the mechanical drive train between passive dissipation in the braking resistor and the active regeneration to the ac grid during the braking operation.
of the drive. The braking resistors are sized such that the power regenerated to the ac grid is within the allowable grid-frequency stability limits.

The braking resistors can be turned ON ($S_{\text{brake}} = 1$) or OFF ($S_{\text{brake}} = 0$) based on the relative magnitude of the measured dc link voltage and a maximum allowable dc-link voltage threshold. However, this may lead to unnecessary turn-on of the braking resistor during low-speed to high-speed mode transition as the power flow to the dc link reverses instantaneously. Using an alternative approach, the braking resistor is operated based on the active power balance in the DFM. The total active power delivered to the DFM is the sum of the stator and the rotor active powers under all operating conditions. A low-speed to high-speed mode transition can be differentiated from a braking operation by the sign of the total active power delivered to the DFM. This is used to command the switching signal of the braking resistor as

$$S_{\text{brake}} = 1 \quad \text{for} \quad P_s + P_r < P_L$$
$$\text{and} \quad S_{\text{brake}} = 0 \quad \text{for} \quad P_s + P_r > P_H,$$

(5.12)

where $P_L$ and $P_H$ are the negative active power bounds for a hysteresis comparator. The active power bounds are designed based on the allowable regenerative braking to the ac grid and the size of the braking resistor.

### 5.2 Reactive power support using DFM

Apart from reducing the required power electronics for controlling the electromechanical energy conversion, a switched-DFM drive offers a unique opportunity of providing reactive power support to the ac grid not only using the grid-side converter but also using the DFM stator connections in high-speed mode. The rotor d-axis current when specifically controlled in high-speed mode, a controllable power factor is achievable at the DFM stator port, which opens up a wide range of possibilities to use the electrical machine to support the utility grid. This is particularly very advantageous for power systems where the electromechanical load dominate the power consumption such as in ship micro-grid or in a weak utility grid.

While the steady-state reactive power capability of the DFM has been shown in Chapter 2, this section analyzes the influence of the transfer switch topology on allowable range of reactive power support with a simultaneous opportunity to transition to low-speed mode, if required. During high-speed-to-low-speed mode transition, for an ideal DFM with a sinusoidal stator current and ac source voltage, any negative drive torque command can achieve a power factor angle between $90^\circ$ and $270^\circ$ required for the ETB transfer switch to satisfy the commutation requirement of high-speed SCR-bank. Alternatively, the corresponding power factor angle must be between $120^\circ$ and $240^\circ$ for the TTB transfer switch. In practice, a margin is required to
accommodate the non-idealities of the DFM and the effect of noise and harmonics of the stator voltage and
current waveforms. Discrepancies on the source voltages and currents from an ideal sinusoidal waveform,
particularly near the zero crossings, can lead to a shorting between the sources. To enforce the margin, it is
ensured that the voltage across the incoming thyristor is greater than a threshold voltage \( v_{th} \) and the current
through the outgoing thyristor is greater than a threshold current \( i_{th} \) at the instant of mode transition. The
threshold voltage is represented in normalized form relative to the peak ac source phase voltage while the
threshold current is specified as a normalized quantity with respect to the peak rated stator current.

As an example, a high-speed-to-low-speed mode transition is considered with the stator current polarities
in the \( B \)-phase and \( C \)-phase being negative. This is represented in Fig. 5.4. The stator current vector \( \overline{T_s} \)
and the ac voltage vector \( \overline{V_{grid}} \) are placed to satisfy the commutation constraint required for the transition.
The threshold voltage constraint for the thyristors is incorporated by specifying the bounds on the allowable
ac source voltage vector as,

\[
120^\circ - \lambda < \theta + \gamma < 240^\circ + \lambda, \quad (5.13)
\]

where

\[
\lambda = \arcsin \left( \frac{v_{th}}{\sqrt{3}} \right)
\]

and \( \gamma \) is the angle between the stator current vector and the stator \( A \)-phase axis, and \( \theta \) is the stator power
factor angle. The stator current vector angle is bounded by,

\[
-30^\circ < \gamma < 30^\circ. \quad (5.14)
\]

The threshold current constraint is incorporated by specifying the bounds on the allowable stator \( B \)- and
\( C \)-phase currents at the instant of mode transition by,

\[
|i_b| = |i_s \cos (120^\circ - \gamma)| > i_{th} \quad \text{and} \quad |i_c| = |i_s \cos (240^\circ - \gamma)| > i_{th}, \quad (5.15)
\]

where \( i_b \) and \( i_c \) are the normalized stator currents through the thyristors \( T_{high,R}^B \) and \( T_{high,R}^C \) respectively.
\( i_s \) is the normalized stator current magnitude.

With the d-q reference frame oriented to the stator flux vector, the relationship between the stator
voltage components \( (v_{sd}, v_{sq}) \), stator current components \( (i_{sd}, i_{sq}) \), and the stator flux magnitude \( \psi \) in the
normalized form is given by,
Figure 5.4: High-speed-to-low-speed mode transition: Incorporating safety margins on the allowable location of the ac source voltage vector location and current vector location for guaranteed mode transition in the presence of non-idealities of the DFM and noise and harmonics in the source waveforms.

\[
v_{sd} = r_s i_{sd} \\
v_{sq} = \psi_s + r_s i_{sq} \\
v_{sd}^2 + v_{sq}^2 = 1 \\
i_{sd}^2 + i_{sq}^2 = i_s^2
\]  
(5.16)

where \( r_s \) is the normalized stator resistance. The stator current components are related to the rotor current components \((i_{rd}, i_{rq})\) according to,

\[
\psi_s = x_s i_{sd} + x_m i_{rd} \\
0 = x_s i_{sq} + x_m i_{rq}
\]  
(5.17)

where \( x_s \) and \( x_m \) are the normalized stator and mutual inductances respectively. The drive torque \( \tau \) is given by,

\[
\tau = -\frac{x_m}{x_s} \psi_s i_{rq}
\]  
(5.18)

The stator and the rotor currents of the DFM must be within the rated values, which in normalized form can be represented as,

\[
i_{sd}^2 \leq 1 \\
i_{rd}^2 + i_{rq}^2 \leq i_{rpu}^2
\]  
(5.19)
where \( I_r \) is the normalized rated rotor current reflected to the DFM stator. For specific values of rotor current components that satisfy the constraints of (5.19), the stator flux magnitude can be computed by solving (5.16) and (5.17) as,

\[
\left( \frac{r_s \psi_s - x_m i_{rd}}{x_s} \right)^2 + \left( \frac{\psi_s - r_s x_m i_{rq}}{x_s} \right)^2 - 1 = 0. \tag{5.20}
\]

Knowledge of the stator flux magnitude is used to calculate the components of the ac source voltage and stator current using (5.16) and (5.17) respectively. Finally, the stator power factor angle is computed by,

\[
\theta = \arctan \frac{v_{sq}}{v_{sd}} - \arctan \frac{i_{sq}}{i_{sd}}. \tag{5.21}
\]

Therefore, for a specific rotor q-axis current \( i_{rq} \), the range of rotor d-axis current \( i_{rd} \) that satisfy the constraints in (5.13), (5.14), (5.15), and (5.19) can be computed using (5.16), (5.17), (5.20), and (5.21).

For example, setting a current threshold magnitude of 0.2 p.u. and a voltage threshold magnitude of 0.1 p.u. for the example DFM, the allowable rotor current components that ensure a natural commutation of the high-speed SCR-bank is shown by area \( ABCD \) in the \( i_{rd}-i_{rq} \) plane in the top trace of Fig. 5.5. The allowable range of \( i_{rq} \) being all positive implies that a braking torque must be commanded to the drive during the high-speed-to-low-speed mode transition, as given by (5.18). The corresponding magnitude is plotted on the secondary axis. The segment \( BC \) represents the non-negative bound on \( i_{rd} \) that prevents unnecessary circulating reactive power in the DFM. The segment \( CD \) represents the bound on the rotor current magnitude based on the rating of the rotor winding. As for the example DFM, the rotor current rating is lower than the stator current rating. Satisfying the rotor current constraint naturally meets the constraint of the stator current rating. The segments \( AB \) and \( DA \) represent the boundary of satisfying the threshold voltage and current constraint of the thyristors.

Alternatively, the commutation constraint can also be satisfied with the stator current in the B-phase being positive and that of the C-phase being negative as represented in Fig. 3.9(b). The above procedure can be repeated to obtain the allowable operating condition of the rotor current components by suitably modifying the bounds on \( \lambda \) and \( \gamma \) as,

\[
240^\circ - \lambda < \theta + \gamma < 300^\circ + \lambda
\]

\[
30^\circ < \gamma < 150^\circ
\]  \tag{5.22}

These constraints extend the allowable rotor current components by additional area \( ADEF \). The remaining two possible polarity combinations of the B-phase and C-phase currents are within the area of \( ABCDEF \).

The bottom plot of Fig. 5.5 shows the allowable \((i_{rd}, i_{rq})\) under identical voltage and current thresholds.
Figure 5.5: Allowable rotor current components \((i_{rd}, i_{rq})\) that simultaneously satisfy the SCR commutation requirement and maintain a voltage higher than a threshold of 0.1 p.u. across the incoming SCRs and a current higher than a threshold of 0.2 p.u. through the outgoing SCRs at the instant of mode transition. These constraints ensure a safety margin during high-speed-to-low-speed mode transition in the presence of DFM non-idealities and noise and harmonics in the source waveforms. (Top) Eight-thyristor-based transfer switch. (Bottom) Twelve-thyristor-based transfer switch.

Comparing the allowable regions in the \(i_{rq}-i_{rd}\) plane between the ETB transfer switch and the TTB transfer switch, it is evident that the ETB transfer switch has more flexibility in terms of operating region during the high-speed-to-low-speed mode transition. The ETB transfer switch can operate at a lower braking torque during mode transition compared to the TTB transfer switch. For the chosen voltage and current threshold, the minimum braking torque required during the high-speed-to-low-speed mode transition is 33% of the full braking torque in the ETB transfer switch. Alternatively, in the TTB transfer switch, the minimum braking torque required is 30% of the full braking torque. The required minimum braking torque influences the drive torque and the shaft speed transients during a gradual deceleration of the drive. Moreover, a wider allowable \(i_{rd}\) implies that the drive can provide an extensive reactive power support to the ac grid with an opportunity to transition to low-speed mode, if required. Finally, in the ETB transfer switch, the rotor \(d\)-axis current command can be set to zero for any braking torque higher than 40% of the maximum braking torque without violating the commutation constraint of the high-speed SCR-bank.
5.3 Experimental Results

5.3.1 Seamless grid behavior of switched-DFM drive for the entire operating speed range

The reference speed of the DFM is commanded with a step function as shown in Fig. 5.6(a). The seamless operation of the DFM drive across the complete speed range with bumpless torque at the shaft is shown in Fig. 5.6(a). The active power shared between the stator and the rotor is shown in Fig. 5.6(b). At the instant of low-speed to high-speed mode transition, the stator active power undergoes a step increase while instantaneously the rotor active power polarity reverses to ensure a minimum perturbation at the shaft torque/speed. At this instant, the dc-link current polarity also reverses. The dc-link voltage is maintained within 3% of the nominal value as shown in Fig. 5.6(c). The ac-grid voltage and frequency decrease as the loading on the generator increases and vice versa due to the programmed droop in the generator controller as shown in Fig. 5.6(d). The ac-grid voltage and frequency remain within ±6% of the nominal values as the drive goes through the different operating regimes during acceleration and braking. The sharing of active power between the two synchronous generators is achieved under all loading conditions as seen in Fig. 5.6(e). The coordinated control of the back-to-back converters ensures that the ac generators do not perceive any sudden disturbance in active power at the mode transition instants. The generators are loaded with a base load of 350 W using a resistive load to prevent reverse power flow during the braking operation of the drive. This load may be considered as equivalent to the "hotel load" in a typical ship microgrid. The voltage command to the dc motor driving one of the synchronous generators along with the voltage command to the generator field winding is plotted in Fig. 5.6(f). Both of these commands have a much lower bandwidth but operate seamlessly irrespective of the DFM operating mode. Similar behavior is observed in all the values during the negative speed range operation of the drive.
Figure 5.6: Experimental results for a step command in reference speed with the switched-DFM drive operating within the full speed range. (a) DFM rotor speed and the drive torque (b) Sharing of the active power between the stator and the rotor of the DFM (c) Dc-link voltage and the dc-link current (d) Ac-grid voltage magnitude and frequency (e) Sharing of the generated active power between the two synchronous generators (f) Voltage command to the dc motor (that mimics the prime-mover input command) and the field winding of one of the synchronous generators.
Figure 5.7: Experimental results: Variation in the dc-link voltage, ac-grid voltage, and the ac-grid frequency with alternate reference speed for the DFM around the mode transition speed. A: High-speed mode (the stator is connected to the ac grid) B: Low-speed mode (the stator is connected to the dc source)

The DFM drive is next subjected to an alternating speed reference command between 680 rpm and 860 rpm with a frequency of 1 Hz. The drive operates alternatively in low-speed mode (marked by B) and high-speed mode (marked by A) as shown in Fig. 5.7. The coordinated control of the switched-DFM drive ensures that the dc-link voltage, ac-grid voltage, and the ac-grid frequency are not only stable but also maintained within small variations.

The DFM drive is next commanded with a constant reference speed of 775 rpm. The dc electronic load bank is programmed to alternate the loading on the shaft between 20% and 60% of the full load torque with a frequency of 1 Hz. As the load torque is turned on and off, the speed perturbation at the shaft causes the DFM drive to alternate between low-speed mode (marked by B) and high-speed mode (marked by A). The dc-link voltage, ac grid voltage, and ac grid frequency are once again stable and maintained within small variations during the load disturbances as shown in Fig. 5.8.

### 5.3.2 Allowable reactive power support based on transfer switch topology

The range of rotor d-axis and q-axis currents that can be supported by the proposed transfer switches while undergoing a high-speed-to-low-speed mode transition is assessed next. The DFM is operated at different rotor d-axis and q-axis current combinations within the allowable rotor current limits while decelerating the
Figure 5.8: Experimental results: Variation in the dc-link voltage, ac-grid voltage, and the ac-grid frequency with an oscillating load disturbance at the DFM shaft. The DFM operates in low-speed and high-speed mode alternatively. A: High-speed mode (the stator is connected to the ac grid) B: Low-speed mode (the stator is connected to the dc source).

drive speed as shown in Fig. 5.9. Each operating point has an uncertainty due to the measurement noise that includes the effect of switching ripple. To illustrate the uncertainty in the actual operating condition, each operating point, in reality, is a circle similar to the one represent by C. The thresholds for the voltage across the incoming SCRs and the current through the outgoing SCRs are both kept at 0.1 p.u. For each of the combinations of the rotor d-axis and q-axis currents, the feasibility of the natural commutation of the outgoing SCRs is evaluated such that the commutation constraints are satisfied. The test result obtained is superimposed to the analytically estimated regions with the desired thresholds. As seen in Fig. 5.9(a), the ETB transfer switch has a wide range of operating conditions in terms of the rotor d-axis and q-axis currents that can simultaneously satisfy the commutation requirements of the high-speed SCR-bank. Comparatively, Fig. 5.9(b) shows a much smaller region of feasible natural commutation of the outgoing SCRs for the TTB transfer switch for identical threshold magnitudes.

5.4 Summary

This chapter presented a coordinated control of the front-end converter and the drive-end converter for a switched-DFM drive. This is necessary to ensure a bumpless operation of the switched-DFM drive not only
relative to the machine's shaft torque/speed, but also relative to the ac source. The seamless performance of the switched-DFM drive can result in a compact and efficient drive with reduced power converter size, thus enabling operation over a wide speed range. The dependency of the choice of transfer switch topology on the reactive power support capability of the switched-DFM drive to the ac source was also explored.

Figure 5.9: Evaluation of SCR commutation criteria during high-speed-to-low-speed mode for different combinations of rotor current components \( (i_{rd}, i_{rq}) \) with identical threshold constraints on voltage and current: (a) eight-thyristor-based transfer switch (b) twelve-thyristor-based transfer switch.
Chapter 6

Drive Topology Selection and Machine Design Considerations

The LSI topology of the switched-DFM drive offers a significant reduction in the required components compared to the LSS topology that requires an additional dc source for low-speed operation. This chapter compares the performance and requirements of the two topologies on the basis of the steady-state design considerations such as torque capability and utilization of the DFM, rotor power converter sizing, component count, transfer switch requirement, and structural noise. The comparison is useful from two different perspectives. First, the comparison can be used to select the topology that satisfies the drive requirement for a given DFM design. Second, the comparison can also be used to specify design guidelines for a DFM that suit the chosen topology for specific applications based on additional requirements. The chapter concludes with a sizing analysis of a 30 MW doubly-fed machine intended for a ship propulsion application.

6.1 Steady-state design and performance comparison

The per-unit d-q model of a DFM in the stator-flux reference frame, as given in the Chapter 2, is used to illustrate the commonalities and the differences between the two topologies from the perspective of the steady-state operation. The parameters of the example DFM, as given in Table. 2.2, are used to compare the two topologies. The constraints for each of the operating modes in steady state are:

1. For the LSS topology: At low speed,

\[ v_{sd}^2 + v_{sq}^2 = v_{dc}^2; \omega_s = 0 \]  

(6.1)
where \( V_{dc} \) is the dc-source voltage.

2. For the LSI topology: At low speed,

\[
v_{sd} = v_{sq} = 0. \tag{6.2}
\]

3. For both topologies: At high speed,

\[
v_{sd}^2 + v_{sq}^2 = 1; \quad \omega_s = 1. \tag{6.3}
\]

### 6.1.1 Torque capability

The torque capability of the DFM depends on the operating flux magnitude and the allowable currents in the stator and the rotor. As seen from (2.7), the torque producing q-axis currents in the stator \( i_{sq} \) and the rotor \( i_{rq} \) are directly proportional. However, the magnetizing d-axis currents in the stator \( i_{sd} \) and the rotor \( i_{rd} \) both contribute to the stator flux \( \psi_s \) at low-speed mode given by (2.6). The objective is to produce the desired drive torque \( \tau \big|_{low} \) in the DFM at low speed with the minimum stator flux magnitude and within the bounds of the allowable stator and the rotor winding currents. Minimizing the stator flux magnitude to achieve the required drive torque directly reduces the required rotor voltage, which impacts the drive size of the rotor converter as was shown in Chapter 2.

An important parameter for the torque capability analysis is the designed rotor-to-stator current rating for the DFM \( I_r \), as given in (2.26) reproduced here for convenience,

\[
I_r = \frac{N_M I_{rms|rotor}}{I_{rms|stator}} \tag{6.4}
\]

where \( N_M \) is the rotor-to-stator turns ratio and \( I_{rms} \) is the rms current rating of the winding. \( I_r \) governs the sharing of the magnetizing current of the DFM and is determined at the design stage of the machine.

For the LSI topology, the rotor d-axis current required to establish the stator flux \( \psi_s \) is given by

\[
i_{rd} = \frac{\psi_s}{x_m}, \tag{6.5}
\]

The torque that can be produced at the DFM shaft is simplified from (6.5) and (2.5) to get

\[
\tau = -\frac{x_m^2}{x_s} i_{rd} i_{rq}. \tag{6.6}
\]

However, the DFM must operate within the constraints of the allowable stator and the rotor currents and can be formulated as

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\[ i_{rd}^2 + i_{rq}^2 \leq I_r^2 \]
\[ i_{rq}^2 \leq \left( \frac{x_m}{x_m} \right)^2 \]
(6.7)
as derived in Chapter 2.

On the contrary, for the LSS topology, the stator flux is set by the stator current, while the d-axis rotor current is controlled to ensure a constant stator flux magnitude under different drive-torque demands. The torque that can be achieved from the DFM in this topology is given by,

\[ \tau = \psi_s i_s \sin \delta \]
(6.8)

where \( i_s \) is the normalized stator current magnitude (dc) and \( \delta \) represents the angle between the dc stator voltage and the stator flux vectors. Therefore, for a chosen stator flux magnitude \( \psi_s \), the maximum torque that can be achieved from the DFM in the LSS topology is obtained by maximizing the product of the stator current magnitude and the sine of the angle \( \delta \) subject to the following constraints below:

\[ i_s \leq 1/\sqrt{2}, \]
\[ \left( -\frac{x_m}{x_m} i_s \sin \delta \right)^2 + \left( \frac{\psi_s}{x_m} - \frac{x_m}{x_m} i_s \cos \delta \right)^2 \leq I_r^2, \]
\[ \left( -\frac{x_m}{x_m} i_s \sin \delta \right)^2 + \left( \frac{\psi_s}{x_m} - \frac{x_m}{x_m} i_s \right)^2 \leq I_r^2, \]
\[ \text{and } \delta < \pi/2. \]
(6.9)

The derivations of the constraints in (6.9) were detailed in Chapter 2. For the example DFM, a comparison between the stator flux-torque characteristic achieved in the LSS topology, using (6.8) and (6.9), and the LSI topology, computed using (6.6) and (6.7), is shown in Fig. 6.1. The off-the-shelf example DFM is designed with \( I_r \) equals to 0.76 p.u. The LSS topology can extract higher torque from the machine compared to the LSI topology for a wide range of stator flux magnitude while remaining within the bounds of the stator and rotor current ratings. Therefore, for a desired drive torque it is more advantageous to use the example DFM in LSS topology as it allows for reduced operating stator flux magnitude. The reduction in operating stator flux magnitude directly translates into reduction in the required rotor converter voltage rating as will be shown in Section 6.1.2.

However, the advantage on the torque capability for the individual topologies depend on the DFM design and more specifically on the \( I_r \) parameter defined in (6.4). The per-unit leakage and the magnetizing inductances of the example DFM are comparable to a typical DFM designed for multi-megawatt power rating [34]. The torque capabilities of the LSI and the LSS topologies are next evaluated using the same per-unit leakage and the magnetizing inductances of the example DFM, but as a function of \( I_r \), ranging
6.1.2 Rotor-side power converter rating

The rating of the rotor-side power converter depends on the voltage, current, and the maximum fundamental frequency rating required for the drive to operate over the full-speed range. The current rating is governed by the drive torque requirement and by the DFM design for sharing of the magnetizing current. For proper utilization of the DFM, the current rating of the rotor converter is predominantly set by the required rotor current to achieve the rated torque at full-speed operation with the stator connected to the ac-source. Therefore, the rotor converter current rating does not depend on the topology of the switched-DFM drive.
However, the voltage rating is dependent on the low-speed mode operation and, therefore, depends on the choice of topology. Two case studies using different drive torque requirements at low-speed for the example DFM highlight the impact of the choice of stator flux magnitude and DFM parameters on the required rotor converter voltage rating for the individual topologies.

6.1.2.1 Identical drive torque achieved at different stator flux magnitude

For example, assuming the required drive torque at low speed corresponds to 0.48 p.u. (72% of the high-speed torque capability of 0.66 p.u.) for the example DFM implies that the required operating stator flux has to be 0.94 p.u. in the LSI topology and 0.725 p.u. in the LSS topology. This corresponds to the operating points $A$ and $B$ respectively in Fig. 6.1. Based on the chosen operating stator flux magnitude, the required rotor voltage for the operation of the DFM in the individual topologies for the maximum positive and negative drive torque is calculated using (2.6)-(2.9) and (2.11)-(2.14) and is shown in Fig. 6.3(a). Not surprisingly, the required rotor voltage is identical in the high-speed ac-source-connected operation of the DFM for both the topologies. This corresponds to the segments $BCD$ and $FGH$. However, the higher operating stator flux in the LSI topology increases the slope of the required rotor voltage versus rotor speed curve at low-speed operation (corresponding to the segment $AB$ and $EF$). This results in a 16% increase in the required rotor voltage for the LSI topology compared to the LSS topology. The transition speed for low-speed to high-speed configuration is adjusted to minimize the required rotor voltage for operation across the complete speed-torque range. An increase in rotor voltage margin also influences the maximum operating speed range in the high-speed grid-connected mode. In the example DFM, the maximum speed that can be achieved by the LSI topology is 6% higher compared to that of LSS topology for the same ac source synchronous speed.
Figure 6.3: Comparison of the LSI and the LSS topologies based on (a) the required rotor converter voltage rating and, (b) active power sharing between the stator and the rotor for the example DFM.

The active power sharing between the stator and the rotor of the DFM at the maximum positive drive torque for the two topologies is compared in Fig. 6.3(b). In the low-speed operating regime in the LSS configuration, the stator supplies the resistive losses in the stator winding. This is represented by non-zero segment of $PQ$. The maximum active power handled by the rotor converter is $34\%$ of the maximum total active power delivered to the DFM in the LSS configuration for the example DFM (corresponding to the operating points $R$ and $S$ in Fig. 6.3(b)). During low-speed operation in the LSI topology, all the active power to the DFM is supplied through the rotor including the stator resistive losses. The maximum active power handled by the rotor converter is $38\%$ of the maximum total active power delivered to the DFM, corresponding to the operating points $Y$ and $Z$ respectively in Fig. 6.3(b).

6.1.2.2 Identical drive torque achieved at identical stator flux magnitude

The next case-study illustrates the effect of the DFM parameters on the required rotor voltage operating with an identical stator flux magnitude. For example, a drive torque requirement of 0.38 p.u. ($58\%$ of the high-speed mode torque capability) for the example DFM during low-speed operation results in identical stator flux requirement in both the topologies as can be seen in Fig. 6.1 by operation point $C$. The goal is to investigate whether the DFM parameters (excluding resistance) have any significance on the choice of topology based on the required rotor voltage under identical stator flux operation. Assuming negligible resistances, the required rotor voltage magnitude can be written using (2.6) - (2.9), (2.11) - (2.14), and (2.5) as,
\[ v_r = (\omega_s - \omega_n) \sqrt{\left(\frac{\sigma}{\psi_s}\right)^2 + \left(\frac{x_r \psi_s}{x_m} - \sigma i_{sd}\right)^2} \] 

(6.10)

where \( \sigma = \left(\frac{x_s x_r - x_m^2}{x_m}\right) \). Under same drive torque requirement and stator flux magnitude, the d-axis component of the required rotor voltage is independent of the topology. The q-axis component of the rotor voltage is dependent on the d-axis stator current. For the LSI topology, \( i_{sd} \) is inherently zero by configuration. For the LSS topology, \( i_{sd} = i_s \cos \delta \) tends towards zero at the maximum drive torque as the angle \( \delta \) approaches 90°. The operating condition at full drive torque implies that the stator current vector is nearly orthogonal to the stator flux vector and the stator flux magnitude is completely supported by the d-axis rotor current. Therefore, the q-axis rotor voltage is not significantly different in either of the LSS or the LSI topologies at an identical drive torque and stator flux conditions. At steady state, the stator flux frequency \( \omega_s \) is zero in the LSS topology. In the LSI topology, \( \omega_s \) corresponds to the slip frequency, which tends to be very small for high-power machines. Hence, for negligible resistances, the required rotor voltage is independent of the choice of topology for a given DFM for identical drive torque requirement and operating stator flux magnitude.

In conclusion, the required rotor voltage is dependent predominantly on the choice of the stator flux magnitude to achieve the desired drive torque at low-speed operation. The preferable topology in this regard greatly depends on the topology that allows one to achieve the desired torque at minimum stator flux, which indirectly depends on the DFM design based on the sharing of the magnetizing current between the stator and the rotor.

6.1.3 Requirement of a separate dc source

The absence of a separate dc source makes the LSI topology attractive due to reduced component counts and complexity of additional source requirement. The dc source needed for the LSS topology can be made using a transformer-diode bridge rectifier. Typically, a transformer is connected between the rotor-side converter and the grid. An auxiliary winding on this transformer may be used to supply a three phase bridge rectifier to create the dc source. The dc source power rating depends on the stator copper losses, which are typically a fraction of the rated drive power. For the example DFM with stator resistance of 0.1 p.u., the required dc source voltage is \( \sim 0.1 \) p.u. and the power rating is of the order of 0.045 p.u. for the different low-speed-to-high-speed drive torque design requirement as shown in Fig. 6.4. The p.u. resistance of a high-power DFM is in the order of 0.01 – 0.02 p.u. This calls for an even smaller dc source with a lower voltage and power rating requirement. The required power rating of the dc source is similar to the additional power rating of the rotor-side converter required in the LSI topology.
6.1.4 SCR-ratings for the transfer switch

The same thyristor-based transfer switches (TTB as well as ETB) can be used for both the LSS and the LSI topology. The six SCRs connected between the ac source and the stator of the DFM behave identically irrespective of the low-speed topology. In the LSI topology, six of the SCRs $T_{\text{low,F}}^A$, $T_{\text{low,R}}^A$ (for TTB transfer switch only), $T_{\text{low,F}}^B$, $T_{\text{low,R}}^B$, $T_{\text{low,F}}^C$, and $T_{\text{low,R}}^C$ that short the stator carry half cycle slip-frequency stator current and, therefore, need to be rated with identical current carrying capability. In the LSS topology, the SCRs $T_{\text{low,F}}^A$ (for TTB transfer switch only), $T_{\text{low,R}}^B$, and $T_{\text{low,R}}^C$ carry the dc current to the stator winding under steady-state operation. The complementary SCRs on the dc-source side, $T_{\text{low,F}}^A$ (for TTB transfer switch only), $T_{\text{low,F}}^B$, and $T_{\text{low,F}}^C$ provide a path for the stator current only during the transients and high-speed-to-low-speed source transition operation. Therefore, these complementary SCRs can be selected based only on the transient current requirements.

6.1.5 Structural vibration and noise

Structural vibration and noise in an electrical machine may arise due to mechanical forces, electromagnetic forces, and aerodynamic forces [56]. The most important source of structural vibration is the electromagnetic force that arise due to the time-varying magnetic fields. Magnetic flux densities in the stator and the rotor interact with each other to produce magnetic stress waves. The magnitude of the stress wave is proportional to the amplitude of the flux densities and acts in the radial direction on the stator and rotor active surfaces. The radial force produce deformations in the magnetic circuit both in the rotor and the stator structures. The stator frame forms the primary source of the machine noise as it is typically connected to a mounting platform, which can act as a transmitting medium. Additional inconsistencies in the magnetic path due to...
the effect of slotting, winding distribution, rotor eccentricity, magnetic saturation, and saliency add on to the variation in radial magnetic force for generation of the noise and vibration in the machine.

The switched-DFM drive in LSS topology produces a stationary magnetic field in the stator and the rotor. Although the radial force does exist, the lack of the time-varying feature of the force on the stator ensures static deformation of the stator structure. This implies that the LSS topology is beneficial in applications as in naval propulsion, where low structural noise is of prime importance during low-speed stealthy maneuvers. The switched-DFM drive in the LSI topology produces a stator magnetic field that rotates at the slip frequency. The low frequency time varying magnetic field produces different modes of magnetic stress waves that form the source of vibration and noise for the electrical machine. As low frequency sound can travel very far underwater, the LSI topology can result in higher levels of detectability when used in naval applications [57]. Additional consideration during the mechanical design of the DFM must be taken into account to ensure that these modes do not overlap with the mechanical resonant frequencies of the stator structure. Proper structural mounting and damping can reduce the propagation of the noise through the stator structure, and can be easily adopted for many civilian applications.

6.2 Exemplary machine design of a 30 MW, 4160 V doubly-fed machine for ship propulsion application

An exemplary doubly-fed machine dimensions are presented next that is designed using analytical methods for a typical ship propulsion application. The design specification includes that the available ship-board ac source is a 4160 V, 60 Hz, 3-phase ac bus while the maximum required rotor speed for the propeller is 200 rpm with a rated shaft power of 30 MW. As the maximum speed in a switched-DFM drive is typically one-and-half times the ac source synchronous speed, the ac source synchronous speed for the DFM must be 133.33 rpm. The required ac source synchronous speed along with the available ac source frequency drives the selection of the number of poles for the DFM to be 54.

Using analytical methods [35, 58], the electromagnetic design of the DFM is performed to meet the required specifications. Table 6.1 shows the terminal specifications of the designed DFM. The choice of stator-to-rotor turns ratio for the DFM being two is driven by the consideration that no additional transformer would be necessary to interface the grid-side converter to the ship-board ac source. Also, the rotor winding is designed to provide the entire magnetizing current for this DFM, which makes the parameter $I_\tau$ as 1.04 when computed using (6.4).

Assuming the windings of the stator and the rotor are water-cooled, the actual volume current density is
Table 6.1: Doubly-fed machine terminal specifications designed for a typical ship-propulsion application.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Line-to-line stator voltage</td>
<td>4160 V</td>
<td>rms</td>
</tr>
<tr>
<td>Ac source frequency</td>
<td>60</td>
<td>Hz</td>
</tr>
<tr>
<td>Rated speed</td>
<td>200</td>
<td>rpm</td>
</tr>
<tr>
<td>Rated shaft power</td>
<td>30</td>
<td>MW</td>
</tr>
<tr>
<td>Rated shaft torque</td>
<td>1492</td>
<td>KNm</td>
</tr>
<tr>
<td>Ac source synchronous speed</td>
<td>133.33</td>
<td>rpm</td>
</tr>
<tr>
<td>No. of pole</td>
<td>54</td>
<td></td>
</tr>
<tr>
<td>Motor winding connection</td>
<td>star</td>
<td></td>
</tr>
<tr>
<td>Stator rated current per phase</td>
<td>2775 A</td>
<td>rms</td>
</tr>
<tr>
<td>Rotor rated current per phase</td>
<td>1450</td>
<td>A (rms)</td>
</tr>
<tr>
<td>Stator-to-rotor turns ratio</td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>Rotor magnetizing current</td>
<td>421 A</td>
<td>(rms)</td>
</tr>
</tbody>
</table>

Table 6.2: Dimensional details of the designed doubly-fed machine

<table>
<thead>
<tr>
<th>Stator</th>
<th>Rotor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Design parameters</td>
<td>Value</td>
</tr>
<tr>
<td>Air-gap diameter</td>
<td>2.572 m</td>
</tr>
<tr>
<td>Stack length</td>
<td>2.572 m</td>
</tr>
<tr>
<td>Slot width</td>
<td>8.3 mm</td>
</tr>
<tr>
<td>Tooth width (at air gap)</td>
<td>8.3 mm</td>
</tr>
<tr>
<td>Slot height</td>
<td>58.8 mm</td>
</tr>
<tr>
<td>Yoke height (minimum)</td>
<td>24.2 mm</td>
</tr>
<tr>
<td>Total number of slots</td>
<td>486</td>
</tr>
<tr>
<td>Number of slots per pole per phase</td>
<td>3</td>
</tr>
<tr>
<td>Number of turns per current path</td>
<td>51</td>
</tr>
<tr>
<td>Number of parallel path</td>
<td>54</td>
</tr>
<tr>
<td>Slot-fill factor</td>
<td>0.55</td>
</tr>
</tbody>
</table>

chosen as 6.5 A (rms) per square mm while the peak tooth flux density and the peak fundamental airgap flux density is chosen as 1.55 T and 0.75 T respectively. Based on these assumptions, the different dimensions of the designed DFM are given in Table 6.2.

6.3 Summary

This chapter compared two topologies of the switched DFM drive under different steady-state considerations. An exemplary machine design utilizes analytical methods for sizing a DFM that can be used for ship propulsion applications. The comparison can be used not only to select the preferable topology for a given DFM design and specific drive requirements based on the end-application but also to specify design guidelines for the DFM for a particular switched-DFM topology. Relative sharing of the magnetizing current to the DFM dictates the preferable choice of topology to achieve the torque capability of the DFM and minimize the required electrical power processing capability. Although the LSI topology has a lower component count, the
LSS topology can have lower structural noises and vibration. In both topologies, the switched DFM drives have the unique potential of reducing the required power converter size while operating over a wide speed range with bipolar torque capability, making them suitable for many high power applications.
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Chapter 7

Conclusion and Future Work

7.1 Conclusion

The central theme of this dissertation is to create a variable speed drive (VSD) architecture with the performance of a conventional full-power-converter drive but with reduced power electronics requirement and with enhanced reactive power support to the grid for wide-speed-range high-power applications. The work in this thesis succeeded in reducing the power electronics requirement by two-thirds as compared to that required by a conventional full-power-converter drive while the entire electromechanical energy conversion system operated seamlessly over a wide speed range. In addition to reducing the power electronics requirement, the thesis explored opportunities through which a controllable reactive power support is provided to the ac utility not only through the power-electronic-based converter but also using the electrical machine itself.

The main motivation for this research is to find solutions for achieving energy-efficient high-power electromechanical energy conversion systems that are severely constrained by the available power electronics devices with limited ratings. Typically, connecting a VSD is the most energy-efficient way of controlling electromechanical energy conversion in electrical machines. However, when these electrical machines handle mega-watt scale powers, creating a VSD becomes significantly challenging. These challenges have led to a very poor adoption rate of 13% for VSD-based technology in real-world applications [17]. The thesis used a doubly-fed machine in a parallel architecture for electromechanical energy processing along with an appropriate mode transition to make variable speed drives that not only provided full shaft speed control but also avoided the need for a dc electrical bus capable of providing full mechanical shaft power. The drive architecture used in this thesis is shown in Fig. 7.1. The switched-DFM drive is attractive for any VSD application where the primary power source is ac. Even though conventional DFMs are well-known
for limited speed range applications, such as in wind power generation, due to their reduced requirement in power electronic rating, switched-DFM drives offer the same benefit on power electronic requirement while operating on a wider speed range. This opens up tremendous opportunities of using switched-DFM drives for a variety of applications requiring a wider speed range, including mega-watt class pumps, compressors, fans, industrial material processing and handling drives, propulsion systems, and HVAC systems.

In order to achieve the goals of this thesis, several underlying research questions addressing the functionality of the entire drive architecture had to be solved. Considering the entire electromechanical energy conversion system represented by black-box "A" in Fig. 7.1, the drive achieved seamless performance both at the mechanical port and the electrical port while simultaneously reducing the size of the required power electronics and providing controllable power factor at the electrical port. In Chapter 2, the design methodology for the rotor-side converter of the drive was presented based on a required drive torque-speed capability at the mechanical shaft. This was crucial to explain the operational benefits of the switched-DFM drive while outlining the key design parameters influencing the drive design. An exemplary drive design presented a systematic approach to explore the design space.

In Chapter 3, several solid-state transfer switch topologies were presented that are the key to the seamless operation of the switched-DFM drive. These transfer switches enable reconfiguration of the DFM terminals on-the-fly without affecting the sources or disrupting the DFM operation. Using thyristors to create the switch topologies allowed a balance between the affordable complexity and the required performance. Con-
strains were derived on the source voltages and the DFM stator currents such that no additional supporting circuitry is required to turn-off the thyristors - an essential requirement when designing mega-watt class power electronics circuits.

In Chapter 4, a control architecture was presented that adheres to the constraints set by the thyristor-based transfer switch while simultaneously achieving a "bumpless" drive performance relative to the mechanical shaft or the electrical source, in particular at the mode transition instances. The control scheme had in-built mechanism to achieve a seamless control of the DFM from the rotor-side converter irrespective of the excitation in the stator while remaining within the designed drive torque-speed specifications. A stator flux estimation method presented in this chapter was crucial in making the control scheme feasible in practice.

As the switched-DFM drive must interact with the ac source seamlessly irrespective of the operating modes to ensure the stability of the ac grid and the intermediate dc-link between the grid-side and the rotor-side converters, the sizing and control scheme of the grid-side converter was introduced in Chapter 5. The control architecture of the grid-side converter enabled seamless grid interactions with the switched-DFM drive and provided reactive power support to the grid without adding extra power electronics. Constraints based on the thyristor-based transfer switch topologies were also analyzed to find the feasible region of operation in terms of the reactive power support.

Finally, the two topologies of the switched-DFM drive were compared in Chapter 6. This chapter explored the selection procedure of a particular topology of a switched-DFM drive that maximizes the reduction in the size of the rotor power electronics for a given DFM design. The comparison is useful from two different perspectives. First, the comparison is useful to select the topology that satisfies the drive requirement for a given DFM design. Second, the comparison is advantageous to specify design guidelines for a DFM that suit the chosen topology for specific applications based on additional requirements presented in this chapter. An exemplary 30 MW, 4160 V, 60 Hz, doubly-fed machine dimensional design showed the feasibility of the proposed switched-DFM drive for ship propulsion applications. The entire drive architecture presented in this thesis along with each individual blocks outlined in Fig. 7.1 were designed, built, and evaluated using an experimental testbed as detailed in Appendix A. The relevant experimental results obtained from the setup were presented and highlighted at the end of each chapters, thereby validating the expected performance and the projected benefits of the switched-DFM drive.

7.2 Future work

The thesis showed the huge potential of using a switched-DFM drive for high-power applications by pushing the trade-off boundary for VSD design to be physically small, efficient, reliable, flexible, inexpensive, and
electric-grid friendly. Going forward, there are multiple opportunities to make the proposed switched-DFM drive, which was demonstrated in a 1 kW laboratory setup, realizable in mega-watt scale commercial applications. Some of the future work that would contribute to make the switched-DFM drive the future of variable speed drives include:

- **Removal of brushes** – Doubly-fed machines inherently require brush and slip-ring arrangements for accessing the rotor port. Although there has been tremendous improvements in brush and slip-ring technologies due to their wide-spread usage in wind turbine systems, removal of brushes and slip rings can always prove beneficial in terms of saving the cost of maintenance and improving the reliability of switched-DFM drives.

- **Evaluation of switched-DFM drive in mega-watt scale systems** – The experimental test platform used in this thesis is a 1 kW laboratory-scaled off-the-shelf DFM. An interesting proposition would be to build the switched-DFM drive in a mega-watt scale and evaluate the performance of the proposed methodologies in the thesis in the mega-watt scale system.

- **Robust control using parameter estimation of DFMs** – The control architecture presented in Chapter 4 is designed assuming constant parameters of the DFM. As in any standard variable speed drive, the performance of the switched-DFM drive can be made progressively more robust by making provisions for online parameter estimations of the DFM.

- **DFM electromagnetic design optimization** – While a conceptual analytical-based sizing analysis of the DFM was presented in Chapter 6, it would be worthwhile to explore the complete design space of the doubly-fed machine including the electromagnetic, thermal, and mechanical designs that are optimized for performance given a switched-DFM drive topology.

### 7.3 Publications

The relevant publications originating from this research are:


• Banerjee, A; Leeb, S.B.; Kirtley, J.L., “Switched doubly-fed machine for ship propulsion”, Electrical Machines Technology Symposium, American Society of Naval Engineers, May 2014


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Appendix A

Experimental Setup

The schematic of the laboratory setup used to evaluate the overall operation of a switched DFM drive is shown in Fig. A.1. Two dc motor-driven synchronous generators operating in parallel make the ac grid of 146 V (line-to-line, rms), 40 Hz. The choice of the ac grid frequency ensures that the off-the-shelf DFM remains within the rated operating speed of 1800 rpm. Correspondingly, the ac source voltage is adjusted to ensure that the DFM operates with a rated stator flux in ac grid-connected mode. The total active power generation capability by the two generators is 1.4 kW. The switched-DFM drive consumes 80% of the total generated power at full-load condition. The generators are loaded with a 20% base resistive load that emulates the auxiliary load of a ship. The rotor of the DFM is connected to the back-to-back converters that can produce a variable voltage variable frequency ac waveform at the rotor terminal from the fixed frequency ac grid. Two Texas Instruments High Voltage Motor Control & PFC Developer’s Kits make the front-end and the drive-end converters. A grid-side auto-transformer and a three-phase LCL filter interface between the ac grid and the front-end converter.

The stator of the DFM is connected to the SCR-based transfer switch that can toggle the stator connection based on the operating drive speed. The phase sequence change over relay ensures that the drive can operate over ±1800 rpm with the same rotor power converter rating. At low-speed mode, the stator can be connected to a separate 20 V dc source in the LSS topology or shorted together in the LSI topology using the switch $S_{ind}$. The shaft of the DFM is loaded by a permanent magnet synchronous generator (PMSG), which is connected to a dc-electronic load bank. The diode bridge placed between the load bank and the PMSG ensures unidirectional current flow irrespective of the direction of shaft rotation. The following sections describe the control for each subsystem of the setup.
Figure A.1: Experimental setup to validate the overall performance and the grid interaction of a switched-DFM drive. The setup emulates a ship power system with 80% share for the propulsion load and the remaining 20% share for the auxiliary loads of the total generated power.
A.1 Control of the ac source: Parallel operation of the synchronous generators

The salient-pole wound-field synchronous generators are driven by two separate dc motors that are individually powered from two controllable dc power supplies. The output voltage of the controllable dc power supply is equivalent to the throttle input for a mechanical prime-mover. The field windings of the synchronous generators are individually excited from two additional controllable dc power supplies. The generator speed, terminal voltage, and the armature currents are measured and used as feedback signals by the generator controller. The control algorithm runs at 960 Hz. Two proportional-integral (PI) controllers, with droop for active power sharing, command the voltage of the dc motors emulating a typical prime-mover control for a synchronous generator. Two additional PI controllers, with feedback for reactive power sharing, command the dc power supplies that energize the field windings of the generators based on the measured terminal voltage.

A.2 Control of the switched-DFM drive: Coordinated control of the drive-end converter, front-end converter and the SCR-based transfer switch

The control of the switched DFM drive is achieved by programming the control architectures of the grid-side converter and the rotor-side converter in two Piccolo microcontrollers - TMS320F28035 that individually have a 32-Bit CPU, 12-bit ADCs, and a 60 MHz clock. The two controllers communicate based on the requirement of the feedforward terms by the grid-side converter. The SCR-based transfer switch is commanded at an appropriate transition instant from the drive-end converter through a programmable system-on-chip (PSoC).

A.3 Control of the load: Current control of the dc electronic load bank

The PMSG coupled to the DFM is connected to the dc electronic load bank (BK Precision 8512), which is operated in constant current mode. The load bank is commanded with a current value every 120 ms by a controller programmed in NI-Labview through serial communication. The controller measures the speed of the DFM rotor and can generate various load torque-speed characteristic. Apart from loading the DFM with a quadratic load torque profile, typical for a ship propulsion, other load perturbations can also be enforced.
at the DFM shaft.

Table A.1: Specifications and make of the equipments used to create the laboratory setup.

<table>
<thead>
<tr>
<th>Equipment</th>
<th>Ratings</th>
<th>Manufacturer and part number</th>
<th>Purpose</th>
<th>Quantity</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC power supply</td>
<td>0-150 V, 0-18 A</td>
<td>Sorensen</td>
<td>Supplies power to the DC motors that drives the synchronous generators</td>
<td>2</td>
</tr>
<tr>
<td>DC power supply</td>
<td>0-150 V, 0-7 A</td>
<td>Xantrex</td>
<td>Supplies power to the field-winding of the synchronous generators</td>
<td>2</td>
</tr>
<tr>
<td>DC permanent magnet motors</td>
<td>120 V, 2 H.P., 14 A, 4800 rpm</td>
<td>Leeson</td>
<td>Drives the synchronous generator through a geared-belt arrangement</td>
<td>4</td>
</tr>
<tr>
<td>Synchronous generators</td>
<td>3 phase, 277/480 V, 5 kW, 7.5 A, 60 Hz, 1800 rpm, 0.8 pf, excitation voltage - 90 V, excitation current - 3.7 A</td>
<td>Mindong - 08041183</td>
<td>Creates the 146 V, 40 Hz ac grid</td>
<td>2</td>
</tr>
<tr>
<td>Equipment Type</td>
<td>Description</td>
<td>Manufacturer</td>
<td>Notes</td>
<td>Quantity</td>
</tr>
<tr>
<td>--------------------------------</td>
<td>-----------------------------------------------------------------------------</td>
<td>--------------------</td>
<td>----------------------------------------------------------------------</td>
<td>----------</td>
</tr>
<tr>
<td>Autotransformer</td>
<td>Three phase, 208-240 V, 50-60 Hz, 20 A, 10.8 kVA</td>
<td>Variac - W20G3M</td>
<td>Interfaces between the ac grid to the front-end converter</td>
<td>1</td>
</tr>
<tr>
<td>HV Motor Control Kit</td>
<td>3-phase motor inverter stage rated for 1KW loads</td>
<td>Texas Instruments - TMD - SHVMTRIN- SPIN</td>
<td>Front-end convert and drive-end converter</td>
<td>2</td>
</tr>
<tr>
<td>Doubly-fed motor</td>
<td>1 H.P., 60 Hz, 4 pole, Frame - 203 stator - 220 V, 3.6 A rotor - 150 V, 4 A</td>
<td>Reuland Electric - 155137</td>
<td>Example DFM</td>
<td>1</td>
</tr>
<tr>
<td>Permanent magnet synchronous generator</td>
<td>3.0 H.P. 2500 rpm, Frame - 66S open 36 VDC, 82 A</td>
<td>Tennant 621-354-000</td>
<td>Load machine that is mechanically coupled to the doubly-fed motor</td>
<td>1</td>
</tr>
<tr>
<td>Electronic Load Bank</td>
<td>600 W, 30 A, programmable</td>
<td>BK Precision - 8512</td>
<td>Connected at the output of the PMSG for loading the DFM</td>
<td>1</td>
</tr>
<tr>
<td>DC power supply</td>
<td>24 V, 100 W ac/dc converter, adjustable output</td>
<td>CUI - VGS-100-24</td>
<td>Separate source that creates 20 V power supply for the LSS topology</td>
<td>1</td>
</tr>
</tbody>
</table>
Appendix B

PCB Designs

B.1 Riser board

This board is designed to interface Piccolo TMS320F28035 Isolated controlCARD to the TI High-voltage motor control kit. The riser board allows access to necessary input/output pins of the controlCARD for appropriate implementation of the control architecture described in Chapter 4.
Figure B.1: Board layout of the riser board
Table B.1: Bill of material for the riser board

<table>
<thead>
<tr>
<th>Qty</th>
<th>Value</th>
<th>Device</th>
<th>Package</th>
<th>Parts</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.1u</td>
<td>C-USC0805</td>
<td>C0805</td>
<td>C11, C12, C39</td>
</tr>
<tr>
<td>3</td>
<td>10n</td>
<td>C-USC0805</td>
<td>C0805</td>
<td>C14, C15, C36</td>
</tr>
<tr>
<td>1</td>
<td>12k</td>
<td>8R-N</td>
<td>DIL16</td>
<td>RN1</td>
</tr>
<tr>
<td>1</td>
<td>1u</td>
<td>C-USC0805</td>
<td>C0805</td>
<td>C35</td>
</tr>
<tr>
<td>8</td>
<td>2.2k</td>
<td>R-US_M0805</td>
<td>M0805</td>
<td>R1, R2, R3, R4, R5, R6, R15, R16</td>
</tr>
<tr>
<td>1</td>
<td>2.2u</td>
<td>C-USC0805</td>
<td>C0805</td>
<td>C16</td>
</tr>
<tr>
<td>8</td>
<td>220n</td>
<td>C-EUC0805</td>
<td>C0805</td>
<td>C1, C2, C3, C4, C5, C6, C25, C26</td>
</tr>
<tr>
<td>10</td>
<td>47n</td>
<td>C-EUC0402</td>
<td>C0402</td>
<td>C7, C8, C27, C28, C29, C30, C31, C32, C33, C34</td>
</tr>
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<td>3</td>
<td>47u</td>
<td>CPOL-USE2.5-7</td>
<td>E2,5-7</td>
<td>C9, C10, C38</td>
</tr>
<tr>
<td>8</td>
<td>5n</td>
<td>C-EUC0805</td>
<td>C0805</td>
<td>C17, C18, C19, C20, C21, C22, C23, C24</td>
</tr>
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<td>680</td>
<td>R-US_M0805</td>
<td>M0805</td>
<td>R7</td>
</tr>
<tr>
<td>5</td>
<td>ADUM2400</td>
<td>ADUM1301</td>
<td>SO-16W</td>
<td>ADUM2400-1, ADUM2400-2, ADUM2400-3, ADUM2400-4, ADUM2400-5</td>
</tr>
<tr>
<td>1</td>
<td>DIMM 100</td>
<td>-</td>
<td>-</td>
<td>U$1</td>
</tr>
<tr>
<td>1</td>
<td>Dsub-15pin</td>
<td>F15HP</td>
<td>F15HP</td>
<td>ADC</td>
</tr>
<tr>
<td>3</td>
<td>Dsub-9pin</td>
<td>F09HPS</td>
<td>F09HP</td>
<td>DAC1, DAC2, GPIO1</td>
</tr>
<tr>
<td>2</td>
<td>Jumper-2pin</td>
<td>JP1E</td>
<td>JP1</td>
<td>0, 3V3</td>
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<tr>
<td>12</td>
<td>Jumper-3pin</td>
<td>JP2E</td>
<td>JP2</td>
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<tr>
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<td>22-23-2031</td>
<td>INPUT, OUTPUT</td>
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<td>MOLEX87630</td>
<td>-</td>
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<td>U$2</td>
</tr>
<tr>
<td>1</td>
<td>TPS77033</td>
<td>TPS77033</td>
<td>SOT23-DBV</td>
<td>IC2</td>
</tr>
</tbody>
</table>

B.2 Thyristor-based transfer switch board

This board is designed to realized thyristor-based transfer switch topologies described in Chapter 3. The board connects the stator port of the DFM to different sources.
Figure B.2: Schematic of the voltage/current measurement board
Figure B.3: Board layout of the thyristor-based transfer switch
<table>
<thead>
<tr>
<th>Qty</th>
<th>Value</th>
<th>Device</th>
<th>Package</th>
<th>Parts</th>
<th>Parts</th>
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<tbody>
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<td>MPT2</td>
<td>2POL254</td>
<td></td>
<td></td>
</tr>
<tr>
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<td>C225-113X268</td>
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<td>C8</td>
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<tr>
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<td>1u</td>
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<td></td>
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<tr>
<td>13</td>
<td>294-1024</td>
<td>SK104</td>
<td>SK104</td>
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<td></td>
</tr>
<tr>
<td>12</td>
<td>2k</td>
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<td>M0805</td>
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<td></td>
</tr>
<tr>
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</tr>
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<td>6</td>
<td>330 (OA331KE-ND)</td>
<td>R-US_0617/3V</td>
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<tr>
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<td>TP</td>
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B.3 Voltage/current measurement board

This board is designed to measure three-phase ac source voltages, dc source voltage, and six currents (three stator and three rotor of the DFM), which are converted to appropriate signals that can be read by Piccolo TMS320F28035 Isolated controlCARD used in the rotor-side converter. The board is also used to measure three-phase ac source voltages, and front-end converter currents for appropriate control of the front-end converter.
Figure B.4: Schematic of the voltage/current measurement board
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### B.4 Programmable System-On-Chip (PSOC) Interface board

This board is designed to interface the switching command issued by Piccolo TMS320F28035 Isolated control-CARD placed in the rotor-side converter to the thyristor-based transfer switch. This board then generates the necessary commands to the individual SCRs placed in the thyristor-based transfer switch board.
Figure B.6: Schematic of the PSoC Interface board.
Figure B.7: Board layout of the PSoc interface board
Table B.4: Bill of material for the PSoC interface board

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B.5 LCL-filter board

This board is designed to interface the front-end converter to the ac source created by using the two synchronous generators. The board has in-built starting resistors for pre-charging the dc link.
Figure B.8: Schematic of the LCL-filter board
Figure B.9: Board layout of the LCL-filter board
Table B.5: Bill of material for the LCL-filter board

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B.6 Motherboard

This board is designed to interface the measured signals from the voltage/current measurement board to the data acquisition system (LabJack).
Figure B.10: Schematic of the Motherboard
Figure B.11: Board layout of the Motherboard
### Table B.6: Bill of material for motherboard

<table>
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Appendix C

Setup Pictures

Figure C.1: Doubly-fed machine coupled to the permanent-magnet synchronous generator.
Figure C.2: Rotor-side converter: TI box modified with the riser board to ensure all the required variables are measured and communicated to the Control Card.

Figure C.3: SCR-Based transfer switch with the phase-change over relays.
Figure C.4: LCL-filter that interfaces the grid-side converter (TI Box), voltage/current measurement board, and labjack measurement system.
Bibliography


