Enhanced Emulation Techniques for Aircraft Systems

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List of abbreviations

AC Alternating current
ADC Analogue/digital converters
APU Auxiliary power unit
ARMAX Autoregressive moving average with exogenous variables
ARX Autoregressive with exogenous variables
CT Control Techniques
DAC Digital/analogue converters
DC Direct current
DM-LM Drive machine-load machine
ECS Environmental control systems
EMA Electromechanical actuators
EMI Electromagnetic interference
FCS Flight control system
FFT Fast Fourier Transform
FW Field weakening
HIL Hardware-in-the-Loop
HMI Human machine interface
HP High pressure
ICE Internal combustion engine
IEPNEF Intelligent Electrical Power Network Evaluation Facility
IM Induction machine
LP Low pressure
LTI Linear time invariant
MEA More-electric aircraft
MMF Magneto-motive force
OE Output-error
PEM Prediction error method
PI Proportional integral (controller)
PID Proportional integral derivative (controller)
PLA Pilot’s lever angle
PLC Programmable logic controller
<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Definition</th>
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<tbody>
<tr>
<td>PMG</td>
<td>Permanent magnet generator</td>
</tr>
<tr>
<td>PMSM</td>
<td>Permanent magnet synchronous machine</td>
</tr>
<tr>
<td>POA</td>
<td>Power optimised aircraft</td>
</tr>
<tr>
<td>PRBS</td>
<td>Pseudorandom binary sequence</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse width modulation</td>
</tr>
<tr>
<td>RC</td>
<td>Resistor-capacitor</td>
</tr>
<tr>
<td>RTP</td>
<td>Real-time platform</td>
</tr>
<tr>
<td>SCADA</td>
<td>Supervisory control and data acquisition</td>
</tr>
<tr>
<td>S-G</td>
<td>Savitzky-Golay</td>
</tr>
<tr>
<td>SMC</td>
<td>Sliding mode control</td>
</tr>
<tr>
<td>SQIM</td>
<td>Squirrel cage induction machine</td>
</tr>
<tr>
<td>SRSG</td>
<td>Switched reluctance starter/generator</td>
</tr>
<tr>
<td>STATCOM</td>
<td>Static synchronous compensator</td>
</tr>
<tr>
<td>V/F</td>
<td>Voltage/frequency</td>
</tr>
<tr>
<td>VSI</td>
<td>Voltage-source inverter</td>
</tr>
<tr>
<td>WIPS</td>
<td>Wind ice protection systems</td>
</tr>
<tr>
<td>ZOH</td>
<td>Zero-order hold</td>
</tr>
</tbody>
</table>
List of symbols

A  Variable frequency signal amplitude
B  Emulator system damping coefficient
\( \hat{B} \) Estimated damping coefficient
\( B_{em} \) Damping coefficient of emulated mechanical system
\( B(q) \) Numerator in the system identified transfer function
C  Gain of the compensator
D  Viscous damping
d  Ideal duty ratio for each inverter leg
e(k)  System noise
e^{Ts}  Inverse system transport delay model in Laplace transform
e^{-Ts}  System transport delay model in Laplace transform
\( F_{abc} \) Three phasors co-ordinates
\( F_{αβ} \) Two phasors co-ordinates
\( F(q) \) Denominator in the system identified transfer function
f  Frame size
\( f_n \) Resonant frequency
\( f_{rated} \) Rated frequency
\( G(s) \) Mechanical dynamics of the electrical drive system
\( \hat{G}(s) \) Estimated plant model
\( G_{comp}(s) \) Compensator term
\( G_{comp-po}(s) \) Compensator including both speed and load effects
\( G_{comp-o}(s) \) Compensator including only speed dependency
\( G_{em}(s) \) Emulated mechanical system model
\( G_{em}^{-1}(s) \) Inverse emulated load dynamics
\( G_M(s) \) Mechanical equation of motion of the system being emulated
\( G_P(s) \) Plant mechanical dynamics
\( G(q^{-1};\theta) \) Input transfer function
\( G_s(s) \) Speed controller
\( G_f(s) \) Feedback controller
\( G_{TF}(s) \) Emulator system transfer function
\( G_{TF-po}(s) \) System transfer function including both speed and load effects
List of symbols

\( G_{TF-\omega}(s) \)  
System transfer function including only speed dependency

\( G_d(s) \)  
Torque/speed controller

\( H(q^{-1}; \theta) \)  
Noise transfer function

\( I_1 \)  
Phase input current in motor per-phase equivalent circuit

\( I_2 \)  
Load current in motor per-phase equivalent circuit

\( I_m \)  
Magnetising current

\( I_{no-load} \)  
No-load current

\( I_{q-rated} \)  
Rated active current

\( I_{rated} \)  
Rated current

\( i_{rd} \)  
Rotor \( d \) axis current in \( d-q \) co-ordinates

\( i_{rq} \)  
Rotor \( q \) axis current in \( d-q \) co-ordinates

\( i_{sd} \)  
Magnetising current in \( d-q \) co-ordinates

\( i_{sd}^* \)  
Magnetising current reference in \( d-q \) co-ordinates

\( i_{sq} \)  
Torque-producing current in \( d-q \) co-ordinates

\( i_{sq}^* \)  
Torque-producing current reference in \( d-q \) co-ordinates

\( J \)  
Emulator system inertia

\( \hat{J} \)  
Estimated inertia

\( J_E \)  
Engine inertia

\( J_{em} \)  
Inertia of emulated mechanical system

\( J_G \)  
Generator inertia

\( J_M \)  
Inertia of the system being emulated

\( J(\theta) \)  
Cost function

\( K \)  
Driveshaft stiffness

\( K_I \)  
Speed controller integral gain

\( K_{II} \)  
Current controller integral gain

\( K_P \)  
Speed controller proportional gain

\( K_{PI} \)  
Current controller proportional gain

\( K_S \)  
Drive system speed slew rate

\( K_T \)  
Machine torque constant

\( \hat{K}_T \)  
Estimated motor torque constant

\( K'_T \)  
Varying torque constant in field weakening region

\( k \)  
Polynomial order

\( L_{lr} \)  
Rotor self-inductance
List of symbols

$L_{ls}$  Stator self-inductance
$L_m$  Magnetising inductance
$L_r$  Total rotor inductance
$L_s$  Total stator inductance
$P_{\text{rated}}$  Rated power
$P_{F\text{rated}}$  Rated power factor
$pp$  Pole pair
$R(s)$  Emulated speed dynamics
$R_r$  Rotor winding resistance
$R_s$  Stator winding resistance
$s$  Motor slip
$\hat{T}$  Estimated delay
$T_{ADC}$  ADC conversion delay
$T_{Ana}$  Analogue transmission delay
$T_{CAN}$  CAN bus communication delay
$T_D$  Torque resulting from drivetrain-backlash integrated model
$T_{Drive}$  Emulator drive system delay
$T_e(s)$  Electrical driving torque
$T_{em}$  Motor electromagnetic torque
$T_f(s)$  Unmodelled dynamics
$T_f(s)$  Feedforward torque
$T_{Gen}$  SRSG generator torque
$T_GT$  Emulated system torque
$T_l(s)$  Electrical load torque
$T_{plc}$  Local PLC delay
$T_c$  Speed controller delay
$T_{\text{rated}}$  Rated torque
$T_s$  RTP sampling time
$T_{\text{sensor}}$  Sensor circuitry delay
$T_i(s)$  Output torque from the speed controller
$T_1$  Lumped forward delay
$T_2$  Lumped feedback delay
$u(k)$  System input
List of symbols

- $V_{dc}$: DC link voltage
- $V_{no,n=a,b,c}$: AC output voltage of the inverter
- $V_{phase-pk}$: Rated peak phase voltage
- $V_{rated}$: Rated voltage
- $V_{rd}$: Rotor $d$ axis voltage in $d$-$q$ co-ordinates
- $V_{ref}$: Reference voltages ($n=a,b,c$)
- $V_{rq}$: Rotor $q$ axis voltage in $d$-$q$ co-ordinates
- $V_{sd}$: Stator $d$ axis voltage in $d$-$q$ co-ordinates
- $V_{s-max}$: Maximum inverter output voltage
- $V_{sq}$: Stator $q$ axis voltage in $d$-$q$ co-ordinates
- $\bar{V}_{tri}$: Maximum triangular voltage of the PWM
- $V_W$: Wind velocity
- $X_{lr}$: Rotor leakage reactance
- $X_{ls}$: Stator leakage reactance
- $X_m$: Magnetising reactance
- $Y(s)$: Delayed system speed
- $\bar{Y}(s)$: Modified version of the controller output
- $y(k)$: System output
- $\alpha$: Ratio of $\omega_{base}/\omega_r$
- $\beta$: Load scaling factor
- $\delta T_{Gen}$: Small signal change in generator torque
- $\delta \omega^*$: Small signal change in emulator speed reference
- $\delta \omega_r$: Small signal change in measured emulator speed
- $\eta$: Efficiency
- $\theta_b$: Backlash angle
- $\theta_E$: Engine angular position
- $\theta_G$: Generator angular position
- $\zeta$: Damping factor
- $\sigma$: Motor leakage factor
- $\tau_c$: Critical time delay
- $\tau_r$: Rotor time constant
- $\varphi_m$: System phase margin
- $\psi_{rd}$: Rotor $d$ axis flux in $d$-$q$ co-ordinates
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>$\psi_{rq}$</td>
<td>Rotor $q$ axis flux in $d$-$q$ co-ordinates</td>
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<td>$\psi_{sd}$</td>
<td>Stator $d$ axis flux in $d$-$q$ co-ordinates</td>
</tr>
<tr>
<td>$\psi_{sq}$</td>
<td>Stator $q$ axis flux in $d$-$q$ co-ordinates</td>
</tr>
<tr>
<td>$\omega^*_r(s)$</td>
<td>Speed reference</td>
</tr>
<tr>
<td>$\omega_b$</td>
<td>Backlash speed</td>
</tr>
<tr>
<td>$\omega_{base}$</td>
<td>Machine base speed</td>
</tr>
<tr>
<td>$\omega_c$</td>
<td>Crossover frequency</td>
</tr>
<tr>
<td>$\omega_E$</td>
<td>Engine angular speed</td>
</tr>
<tr>
<td>$\omega_{em}(s)$</td>
<td>Emulated speed</td>
</tr>
<tr>
<td>$\omega_f$</td>
<td>Signal frequency</td>
</tr>
<tr>
<td>$\omega_G$</td>
<td>Generator angular speed</td>
</tr>
<tr>
<td>$\omega_n$</td>
<td>Corner frequency</td>
</tr>
<tr>
<td>$\omega_r(s)$</td>
<td>Measured shaft speed</td>
</tr>
<tr>
<td>$\omega_{rated}$</td>
<td>Rated speed</td>
</tr>
<tr>
<td>$\omega_{slip}$</td>
<td>Motor slip speed</td>
</tr>
<tr>
<td>$\omega_{syn}$</td>
<td>Motor synchronous speed</td>
</tr>
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Abstract

Name of University: The University of Manchester

Candidate’s Name: Chengwei Gan

Degree Title: Doctor of Philosophy (PhD)

Thesis Title: Enhanced Emulation Techniques for Aircraft Systems

Date: August 2014

The increasing electrical load demand on on-board (aircraft, marine, automotive, commercial vehicles and rail) power networks, together with the proposed embedded electrical generators introduces new challenges including the potential for engine-generator-electrical system interactions. Accurately emulating such systems is essential to enable the identification and mitigation of such interactions to ensure reliable, robust operation. In this thesis, an advanced emulation method which cancels the natural drive system dynamics using system identification is proposed, designed, simulated and demonstrated experimentally to show its performance in emulating a wide range of aircraft mechanical systems using a commercial electrical drive system.

The thesis firstly examines the performance of the existing emulator system hardware, which consists of an electrical drive, real-time control platform and inline speed and torque sensors, to identify aspects of the drive behaviour which may impose a limit on the range or the accuracy of the systems being emulated. Two features of the experimental data were identified which may compromise the ability of the emulator system to emulate mechanical sources, the first is an 89ms time delay between the reference speed and the measured speed, and the second is the relatively low, 60rad/s, bandwidth of the speed controller with a significant resonant peak at approximately, 40rad/s, which significantly limit the range of frequencies which can successfully be emulated. Existing time delay compensation techniques were analysed and simulated using an experimentally validated simulation model, offering only a modest reduction in time delay.

An advanced natural drive system dynamics compensator method is proposed to cancel the speed control loop characteristics and so extend the emulation bandwidth of the electrical drive-based emulator system without the need for any hardware changes or knowledge of the drive system parameters. This compensator is designed by applying parameter system identification to experimental data from the emulator drive system speed step response, and then modified to incorporate the effects of field weakening and load variations. The effectiveness of the proposed emulation approach is demonstrated by simulation and experimental emulation of a wide range of mechanical systems, with a focus on aerospace applications, including gas engine dynamics and the associated mechanical transmission effects due to a compliant drivetrain and backlash effects. The proposed emulation method is not limited to a specific mechanical system model, and is particularly suitable for the emulation of the high-speed and high-power mechanical systems.
Declaration

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Dedication

This dissertation is lovingly dedicated to my father, Lin Gan, my mother, Huayan Guo and my girlfriend, Bing Han.
Acknowledgements

I would like to express my profound gratitude to Dr Rebecca Todd and Dr Judith Apsley for their excellent supervision throughout the project. Without their continuous enthusiasm, vision and wisdom on this project, my PhD studies in Manchester would not have been so rewarding. Their strong commitments to excellence in research and rigorous research attitude have set an exemplary model for a PhD student and will also inspire me to pursue the extraordinary in my future career.

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Special thanks to my parents and family for their love and continuous support in my life to which I am always indebted. Finally, I wish to thank my girlfriend, Bing Han, for her always encouragement, support and love which is very critical to the completion of this PhD research.
About the author

The author received the B.Eng. degree in 2010 from North China Electric Power University, Baoding, China, and the first-class B.Eng. (Hons.) degree, also in 2010, from The University of Manchester, Manchester, U.K., where he subsequently started his PhD in the Power Conversion Group. His research interests include time delay compensation techniques for use in industrial control platforms, electrical drive-based mechanical system emulation, and the more electric aircraft.

The following papers have been accepted and are awaiting publication, or have been published based on the work presented in this thesis.


Chapter 1

Introduction

Protecting the environment and managing the usage of finite energy reserves are key goals for European Aviation [1]. To achieve these goals, ambitious targets for 2050 of a reduction in CO₂ emissions by 75% per passenger kilometre together with a 90% reduction in NOₓ emissions, both with respect to the emissions of a new aircraft in 2000 are defined. These goals are especially challenging considering the anticipated doubling of air traffic in the next 15 years [2], which will in part be met by 27,000 new 100 plus seat aircrafts by 2031 [2], and the increasing electrical aircraft load largely due to the no-bleed engine design [3] where pneumatic systems are replaced with electrical alternatives.

Diverse technologies are required to achieve the emissions targets including, electrical and hybrid electrical engines [1], fuel additives [4], lighter airframes [4], better aircraft/engine integration [3], optimised flight trajectories for fuel and time efficiency [1], and the replacement of pneumatics [3] and hydraulics [5] with electrical systems. Many of these technologies are directly related to the ‘more-electric’ aircraft concept which is being embraced by the aerospace industry [3].

1.1 More-Electric Aircraft

On traditional aircraft fuel is consumed by the gas engines to produce power, the majority of which is used for propulsion, with the remainder primarily used by pneumatic systems, environmental control systems (ECS) and wind ice protection systems (WIPS), with additional lower power used by hydraulic, mechanical and electrical systems. In the more-electric concept the non-electric powered systems which are heavy, cumbersome, inefficient and prone to leaks, in the case of pneumatics [6], are replaced by electrical systems which are lighter, smaller and more efficient [7]. In addition to the improved efficiency and reduced fuel consumption the more-electric aircraft offers better power management, increased flexibility, reduced aircraft empty weight, lower installation and maintenance costs and faster turn-around times [7] and easier fault-detection which enables localised isolation of faults [8].
The Boeing 787 Dreamliner uses the only commercially available no-bleed engine, where the traditional engine powered pneumatic system are replaced by an electrical equivalent which is expected to extract 35% less power from the engine [3], improving efficiency [6, 9], leading to reduced fuel consumption and emissions. The 787 has engine gearbox driven variable frequency generators [3], which although more efficient, more reliable, with lower spare parts costs [3] than the complex constant speed drive integrated generator, does compromise the engine nacelle design. Removal of the engine gearbox, and so offering a more aerodynamic nacelle, is possible by utilising embedded generators which are directly mounted on the engine shafts as demonstrated in the Rolls-Royce supported power optimised aircraft (POA) Trent 500 demonstrator [10]. The embedded generators must be able to withstand both the high temperatures and also the high rotational speeds of the gas engine spools. Candidate generator topologies include switched reluctance machines [6, 10, 11] and permanent magnet machines [10, 12, 13, 14], owing to their robustness in high speed and high temperature environments.

The use of an embedded generator introduces direct and tight coupling between the gas engine spool(s) and the generators. This, coupled with both an increased electrical load in a more-electric aircraft, which is likely to reduce the ratio of propulsive to non-propulsive power, and more highly dynamic motor drive power converter loads [15], gives rise to the possibility of engine-generator-electrical system interactions. This direct coupling and higher electrical loading on the engine may also result in the gas engine operating on the extremes of the engine performance envelope which may increase the likelihood of engine surge which could have serious implication on aircraft safety. Developing a fundamental understanding of potential engine-generator-electrical system interactions and demonstrating mitigation techniques are critical to future more-electric aircraft developments.

### 1.2 Hardware-in-the-Loop Techniques

In the case of the more-electric-aircraft (MEA) described in Section 1.1, emulation systems are an invaluable tool to system designers and integrators as the capability to accurately emulate future propulsion systems enables potential engine-generator-electrical system interactions to be evaluated and mitigation techniques to be demonstrated prior to integrated ground based testing. Hardware-in-the-Loop techniques form the basis of
modern emulation systems, with an electrical drive system [16] being controlled to replicate the behaviour of the system being emulated. The use of an electrical drive introduces new challenges to emulation systems owing to their fast dynamic response, often less than a few milliseconds, and complexity.

Real-time hardware-in-the-loop (HIL) techniques are an established method of evaluating system performance and design using a blend of hardware, either actual or emulated systems, and a digital real-time simulation platform. Two distinct forms of HIL exist, controller or power HIL systems [17, 18], shown in Figure 1.1.

In the controller HIL systems, Figure 1.1(a), a physical control platform is used to command a model of the system being controlled which is implemented on the real-time platform (RTP). This technique is widely used in power systems applications [19, 20, 21] where large power systems are simulated to evaluate the performance and effectiveness of physical hardware devices, including relays and protection devices. Recent advances in computational methods also enable the real-time emulation of power electronic circuits [22, 23] to enable the evaluation of control algorithms implemented on industrial control platforms before test work on a physical converter. In these applications, the RTP is connected to the controller hardware under-test through the low power interfaces such as analogue/digital (ADC) or digital/analogue (DAC) converters.

Power HIL, Figure 1.1(b), is the opposite approach to the controller HIL in that the system being controlled is the physical hardware with the control being implemented on the RTP.
This technique is extensively used in emulation systems, where physical hardware is commanded to reproduce the behaviour of the system being emulated [24, 25, 26] with the emulated system model being implemented on the RTP. Another common power HIL application is where the RTP is used for rapid controller prototyping [27] to enable advanced controllers to be used to regulate the physical hardware. Unlike controller HIL, power HIL implementation includes representative power hardware components and the RTP is linked to the power device under-test through a much more complex interface, which involves power amplification and conversion apparatus (ADC/DAC), bus communication (CAN bus or Profibus) or analogue transmission, and a sensor.

Emulation techniques based on power HIL are well established to aid the design, development and research of transportation systems [25] and renewable power generation applications [24].

1.2.1 Power HIL Emulation Systems

Power HIL emulation systems are commonly used to emulate mechanical systems, sources or loads, and can be designed based on a range of technologies, including mechanical, hydraulic or electrical drive systems. Electrical drive based solutions, consisting of a power electronic converter, closed-loop vector control and electrical machine, are now extensively used in preference to hydraulic or mechanical systems as the inclusion of a power converter enables a higher controllability and increased flexibility [28] providing that the torque capability of the electrical machine is sufficient for the system being emulated [28].

Electrical drive based emulation systems are inherently complex owing to the use of advanced field oriented control algorithms, multiple measurement sensors, protection systems, and the interconnection with other elements in a test facility including the RTP (shown in Figure 1.1(b)). The emulation system used in this thesis is based on the power HIL techniques which will be introduced in Section 1.3.

1.3 IEPNEF

The Intelligent Electrical Power Network Evaluation Facility (IEPNEF), which is located in the Rolls-Royce University Technology Centre at The University of Manchester, has been established using hardware-in-the-loop (HIL) techniques to examine the electrical
system of a more-electric aircraft in a safe and controlled environment. This 100kW state-of-art facility shown in Figure 1.2 contains the main elements of a future aircraft electrical system with engine embedded generators, enabling aero engine-generator-electrical system interactions to be studied.

IEPNEF enables two gas engine spools to be emulated, with the low pressure (LP) spool emulator being coupled to a 70kW 3000rpm permanent magnet generator (PMG) [29] and the high pressure (HP) spool being coupled to a 30kW 15000rpm switched reluctance starter/generator (SRSG) [30]. Two 115kW variable-speed induction machine drives commanded by a real-time platform (RTP) form the spool emulators, driving the generators to energise the 540V DC power network. The RTP contains a thermodynamic and mechanical aero gas engine model, which receives pre-programmed flight profile data, such as environmental data and throttle (pilot’s lever angle (PLA)) and outputs the corresponding spool speed command to the relevant electrical drive unit. This RTP is labelled ‘flight control system’ (FCS) in Figure 1.2 as the FCS can automatically command the entire IEPNEF [31], enabling full flights to be investigated.

The generated electrical power is distributed through a reconfigurable power distribution network with a 540V DC bus structure to four bi-directional DC/DC converter-based
active load systems [31] and two resistive load banks. The active load systems are used to emulate a number of aircraft loads, include an auxiliary power unit (APU), electromechanical actuators (EMA), and pulse radar loads. In addition, a 1MJ super-capacitor based energy storage system [32] has also been implemented to evaluate advanced power management schemes.

IEPNEF is principally designed as a demonstrator facility to research on-board power generation and management for future aircraft and land based vehicles. Research topics include engine-generator-electrical interactions (the focus of this PhD project), electrical system stability and power quality, energy storage/management, fault tolerance, protection and health monitoring.

1.4 Research Aims and Objectives

The aim of this research project was to devise and demonstrate advanced electrical drive based mechanical system emulation techniques which can accurately reproduce transportation system dynamics, for example an aerospace gas engine. To accommodate future prime mover designs the emulation technique should not be limited to a single engine and should offer the widest bandwidth possible from the existing emulation system hardware. The objectives of this research project are defined below:

- Develop and validate a model of the emulation system to identify the operating envelope and the dominant effects on the system behaviour.
- Investigate limitations in the emulator system and devise/assess mitigation techniques.
- Devise and demonstrate a compensator to cancel the emulation system dynamics and identify any bandwidth limits.
- Develop and emulate experimentally additional mechanical system models with wide speed/power/bandwidth requirements to assess the emulation capability of the drive system dynamics compensator based emulation method.

1.5 Contributions of this Research

This research project advances the current state-of-the-art mechanical system emulation through the introduction and validation of a new approach to drive system dynamics
compensation. The proposed emulation method uses the parametric system identification technique to experimentally capture unmodelled dynamics. This approach offers superior emulation capabilities from the electrical drive system over wider speed range and up to higher bandwidths. The main contributions of this research project are detailed below:

- Methods to mitigate the effects of time delay in modern electrical drive based power hardware-in-the-loop emulation systems have been evaluated with respect to both open-loop and closed-loop scenarios.
- A novel drive system dynamics compensator based mechanical system emulation method has been developed and the technique has been demonstrated to enable enhanced aero engine dynamics emulation over the full operating range.
- The performance limits of the proposed emulation technique, in terms of the range of inertias that can be emulated has also been evaluated, together with the successful emulation of resonant frequencies from mechanical transmission effects such as drivetrain resonances and backlash vibrations. These higher order effects demonstrate the future capability to examine electro-mechanical interactions within the more-electric aircraft.

1.6 Structure of this Thesis

This chapter has provided background information on the more-electric aircraft and a brief overview of the hardware-in-the-loop (HIL) techniques, including typical applications and emulation hardware together with a brief introduction to the experimental facility used in this project. The objectives and contributions of this research are identified.

Chapter 2 gives a comprehensive discussion and evaluation of the Prior art on mechanical system emulation. Additional literature on time delay compensation techniques and system identification is then presented with a focus on power HIL applications.

Chapter 3 presents a detailed simulation model for the emulator system, based on which time delay compensation techniques have been evaluated for the closed-loop scenario while the existing open-loop setting within the emulation system applies to subsequent chapters.

Chapter 4 presents two compensator designs from different fidelity simulation models. The first compensator is determined analytically whereas the second one is designed using the
parametric system identification technique which is applied to small speed step transients from the full simulation model. Both compensators are then evaluated experimentally and by simulation.

In chapter 5 a drive system dynamics compensator based on the measured experimental speed step transients is devised and validated in both the time and frequency domains.

Chapter 6 summarises HIL emulation results for a wide range of mechanical systems, all with a focus on aerospace applications. Furthermore, the main limits on the compensators performance are identified.

Chapter 7 presents the conclusions and suggestions for future work and a speculative study on the open-loop time delay compensation within the emulation system is performed.
Chapter 2

Literature Review

In this chapter existing mechanical system emulation techniques using electrical drive based methods are evaluated. Time delay compensation techniques applicable to the HIL applications are then evaluated. An overview of system identification techniques is then given with a discussion of typical applications.

2.1 Mechanical System Emulation Techniques

Emulating the behaviour of complex systems is a widely used experimental technique to enable the performance of a system over its operating range to be examined in a safe, repeatable, and cost effective way. Often mechanical systems are emulated, for example, the emulation of a renewable power source [24, 33, 34] enables generation techniques to be examined, flight surface actuation systems are emulated to assess control methods, or gas engines are emulated [35, 36, 37, 38, 39] to enable controllers or electrical subsystems to be evaluated and tested. Common features of all systems being emulated are significant safety concerns, high cost, and complex drive/load demands.

2.1.1 Dynamic Emulation of Mechanical Source Systems

Mechanical sources or prime movers are often emulated using electrical drives in both renewable power generation systems, especially wind turbines [24, 33], and transportation applications [35, 37]. Since the drive dynamics are normally significantly faster than the traditional mechanical system they are emulating, the mechanical dynamics of a conventional prime mover can be reproduced by appropriately selecting an electrical machine and drive system.

Wind turbines

Wind energy conversion systems are often emulated [24, 33, 34] to enable advanced generator control or alternative generator designs to be evaluated without the significant time and cost associated with the planning and construction of a wind turbine. There are also benefits associated with the ease of access to an emulation system compared to a wind
turbine nacelle. The wind turbine emulator system shown in Figure 2.1 consists of an electrical drive with an electrical machine coupled to the generator under test and a RTP which contains a model of the wind turbine. The wind turbine model determines a reference torque for the electrical drive system based on the external stimulus such as wind velocity and turbine rotor speed.

![Figure 2.1: Real-time wind turbine emulation system](image)

Many different fidelity wind turbine models are used in the emulator depending on the specific turbine behaviour of interest and also the computation power of the RTP. Simple torque equations [40, 41] can be used to express the steady-state wind turbine behaviour, or a modified inertia model [33, 42] which equates the wind turbine and emulator system torque equations offers a relatively higher fidelity. More detailed wind turbine models are often used to examine specific issues including the effect of varying wind turbine inertia [43], wind shear and tower shadow effects [42] or torsional resonances [43, 44, 45] in the wind turbine drivetrain. The wind shear and tower shadow effects [42] enable the emulator to replicate harmonic torques, and the three times the rotor frequency power pulsations. The torsional resonance [43, 44, 45] was examined using a two/three mass oscillating system to model the dynamic characteristic of the wind turbine which represent the turbine, gearbox and elastic drive shaft.

Induction machines are preferred in wind turbine emulator systems [40, 46], owing to their robustness and cost in comparison to permanent magnet synchronous machines (PMSM) [42] or low maintenance requirements compared to DC machines [47], though this is at the expense of more complex control requirements than DC machines [47].

In the majority of wind turbine emulation systems the model determines the reference torque that would be transmitted on the shaft coupling the blades/gearbox to the generator [24, 34, 48]. Occasionally speed control is employed in the emulator to enable wind turbine pitch control [44, 49, 50] to be examined where the turbine mechanical power is maintained at its rated value [49]. In either case the control must be appropriately designed
to allow good emulation of the wind turbine dynamics of interest. In [43, 51], a DC motor drive emulator is operated under torque control where a PI controller regulates the armature current. A frequency response tuning method in the study [44] was used to design a PI torque controller for a DC machine by specifying the desired control bandwidth and phase margin from the emulated dynamics. [42] employed a PMSM as a vector-controlled motor, which has a torque control bandwidth of approximately 130Hz to enable the dynamics resulting from the tower shadow and wind shear effects to be emulated. [46] proposed a novel control structure for a wind turbine simulator in which two closed-loop controllers, a speed and current control, are applied to the inverter-controlled induction motor.

Much of the wind turbine emulation literature focuses on the dynamic emulation of the wind turbine rotor dynamics, though some researchers propose methods to emulate the mechanical characteristics of a wind turbine drivetrain [43, 44, 50]. The wind turbine model on the RTP is enhanced to contain a mechanical drivetrain model to produce the torque harmonics which then form the torque reference for the electrical drive emulation system. The work in [43] uses a PWM DC-DC converter in the torque control of a DC motor where a hysteresis current control structure was implemented. Some interesting emulation results of the harmonic torque, due to the gradient and tower shadow effects, are presented in [43]. Although the hysteresis controller can enable a high control bandwidth it may result in a poor steady-state performance where the current errors can be up to twice the hysteresis band [52]. Evaluating the emulation accuracy of the torque oscillations is not possible from [43] as only the measured speed variations due to the torque harmonics are presented with no speed reference data to enable a comparison. In [44], a PID speed controller was designed to obtain a wider control bandwidth compared to that of a PI speed control structure. Evaluating the emulation performance of the resonant frequencies is not possible as the PID implementation in the test system was unsuccessful. In [50], successful emulation performance of the third harmonics in the turbine shaft speed due to both the mechanical drivetrain and wind turbulence is presented, though little detail is provided on the controller design or implementation.

Mechanical prime mover emulation in wind energy applications often uses the model-based control method, in which a PI torque controller is designed and employed in the emulator system, taking into account the emulated turbine dynamics. Providing the electrical drive emulation system torque control and power electronic converter have
sufficient bandwidths then accurate steady-state and transient emulation may be achieved. The potential source of interactions in a wind turbine system are limited to the mechanical input to the generator as the electrical network is generally assumed to be a stiff grid, removing network variation as a source of interactions except for during fault conditions. Wind turbine emulation techniques are not appropriate for gas engine emulation which may experience interactions from both the mechanical prime mover and electrical network leading to complete engine-generator-electrical system interactions.

Gas engines

Relatively few technical papers investigate the use of an electrical drive system to emulate the dynamics of gas engines for transportation applications [35, 36, 37, 38, 39]. The general implementation of the gas engine emulation system coupled to an electrical generator is shown Figure 2.2. A speed reference is often provided by the gas engine model to the electrical drive system as the engine speed is principally determined by the required propulsion via the fuel controller with only a slight speed variation resulting from the generator load step. The generator load does however determine the non-propulsive torque extracted from the spool and so in contrast to the wind turbine emulator there is a significantly higher likelihood of engine-generator-electrical system interactions.

In Figure 2.2, if a gas engine model as opposed to a torque equation is used then the inputs to the model are normally environmental data (altitude, Mach number), throttle command (pilot lever angle (PLA)), together with measured generator torque.

In [35], a PI speed controller was designed for the electrical drive system to allow the shaft dynamics of a twin-shaft gas engine to be emulated with a synchronous motor. [36] describes the implementation of an auto-tuned PI speed controller in an induction motor drive system to emulate an internal combustion engine (ICE) system. It was suggested in
[36] that compensation of the electrical drive dynamics are necessary if the drive dynamics have a noticeable effect on the emulation performance. In [37] and [38], some characterisation tests such as ramp, step and frequency response tests were conducted to assess the electrical drive performance on the emulation of a gas engine dynamics, concluding that the electrical drive emulator bandwidth should exceed that of the emulated engine system. [39] presented a model-based control method for emulating an aeroderivative twin-shaft engine which has a 10 times smaller inertia in comparison with the machine of the same power rating. In [39], the PI speed controller, which was designed for slow engine dynamics, is adopted for use in the emulation of the aeroderivative engine and some emulation limitations are evident in [39] resulting from accuracy and stability issues. The paper also demonstrates that the traditional model-based control method for gas engine emulation is limited to a specific system and cannot accommodate applications with faster dynamics. As in the wind turbine emulation, all existing studies on the gas engine emulation are also focused on the model-based control approach where an appropriate speed controller (normally PI type) is designed and implemented.

Gas engine emulation often involves the dynamic emulation of high speed and high power applications such as an aero gas engine. Emulation systems using an AC machine often requires field weakening capability to enable the emulator to achieve a high speed operating range. The field weakening effect, together with the strong coupling between the emulator and the generator, make good emulation performance over the full speed and power range a challenge to achieve.

**Generic mechanical source**

A block diagram that describes the model-based control technique for a generic mechanical source is shown in Figure 2.3. In Figure 2.3, \( G_{em}(s) \) is the emulated mechanical system model and \( G(s) \) represents the mechanical dynamics of the electrical drive system. \( G_c(s) \) is the corresponding torque/speed controller which depends on the system configuration; in Figure 2.3 a speed controller is assumed. The relationship between the actual rotor speed \( \omega_r(s) \) and the driving torque \( T_e(s) \), (2.1) can be obtained by considering Figure 2.3 and is dependent on the rotor dynamics of the electrical drive \( G(s) \) and the emulated (desired) mechanical system \( G_{em}(s) \).

\[
\frac{\omega_r(s)}{T_e(s)} = \frac{G(s)G_c(s)G_{em}(s)}{1 + G(s)G_c(s)} \tag{2.1}
\]
Chapter 2  

LITERATURE REVIEW

Figure 2.3: Overview of the model-based control method

From (2.1), in order to obtain good emulation of the mechanical dynamics, \( G_{em}(s) \), the controller \( G_t(s) \) must be carefully designed, considering both the dynamics of the actual test rig and the emulated mechanical system, to meet the requirement in (2.2).

\[
\frac{\omega_t(s)}{T_e(s)} = G_{em}, \text{ when } G(s)G_t(s) \gg 1
\]  

(2.2)

The controller \( G_t(s) \) often needs to be redesigned for each range of emulated system dynamics in order to allow the conditions in (2.2) to be fulfilled which limits the use of this technique.

2.1.2 Dynamic Emulation of Mechanical Load Systems

Mechanical load emulation using a dynamometer (electrical drive) is very common in many applications including the off-site testing of electrical drives [53], research and development of electrical machines [54], or engine test beds [55, 56]. The dynamometer, shown in Figure 2.4, can be controlled to emulate the specific load characteristics when the load is unavailable. The specific application determines the level of load model required. Simple linear and non-linear loads are often sufficient to load a test electrical drive or machine, using either fixed mechanical parameters, or speed and load dependent values to enable the engine/machine to be tested under steady state or slowly changing load conditions. More detailed drivetrain model are often necessary in engine test beds.

Figure 2.4: Mechanical load emulation system set-up
Inverse mechanical dynamics emulation

Dynamic loads, both linear and non-linear, are very common in the testing of the variable speed and torque drives. The conventional approach for dynamic emulation of mechanical loads is based on the principle of inverse mechanical dynamics [54, 57, 58, 59], in which an inverse dynamic model of the desired load is used to derive the torque reference for the dynamometer from the measured shaft speed. The emulation method based on the inverse load dynamics is shown in Figure 2.5.

\[
G(s) = \frac{1}{Js + B} \quad (2.3)
\]

The relationship between the measured shaft speed \(\omega_r\) and the electrical torque \(T_e\) of the drive machine can be deduced from Figure 2.5 as:

\[
\frac{\omega_r(s)}{T_e(s)} = G_{em}(s) = \frac{1}{J_{em}s + B_{em}} \quad (2.4)
\]

Successful emulation with the inverse load model is not possible in the closed-loop control system as the overall pole-zero structure of the emulated load is violated [60] when (2.4) is inverted. The emulation technique can be used in an open-loop control system, however, filtering is often essential as sampling effects in the RTP introduce noise resulting from the derivative terms of the inverse transfer function, potentially affect the system stability in the open-loop emulation of the load dynamics. Furthermore, non-linear loads or more
detailed loads are complex to be emulated with the inverse load model method as deriving
the inverse transfer function is quite difficult.

**Closed-loop load emulation method**

Since the conventional mechanical load emulation approach is not viable for use in the
closed-loop control systems, a dynamic load emulation method which is effective under
closed-loop conditions has been proposed [60, 61, 62] which is based on feedforward
speed-tracking control of the emulator drive. In this approach the aim is to cancel the
physical drive system dynamics to enable good emulation performance, while at the same
time maintaining the pole-zero structure of the emulated load. Figure 2.6 shows a block
diagram representation of the closed-loop emulation approach using a compensator method.

In this emulation approach, an implicit feedforward of the inverse load dynamics $G_{em}^{-1}(s)$
is implemented for use in the compensation of the closed-loop dynamics.

![Figure 2.6: Closed-loop mechanical load emulation method (adapted from [60])](image)

$T_e(s)$ and $T_l(s)$ are the motor driving torque and dynamometer load torque respectively. $G_f(s)$
is the speed-tracking control. As shown in Figure 2.6, the inverse emulated load dynamics
$G_{em}^{-1}(s)$ are fed forward from the emulated speed $\omega_{em}(s)$ to provide a feedforward torque
$T_f(s)$. A speed-tracking loop which compares the desired speed $\omega_{em}(s)$ with the actual speed
$\omega_f(s)$ outputs the torque $T_f(s)$. From Figure 2.6, the relationship between the drive torque
$T_e(s)$ and machine speed $\omega_r(s)$ can be obtained:

$$\frac{\omega_r(s)}{T_e(s)} = G_{em}(s) \frac{G(s)G_f(s)}{1 + G(s)G_f(s)} G_{comp}(s) \quad (2.5)$$
In order to preserve the mechanical characteristics of the emulated load, the term $G_{\text{comp}}$ should be selected to cancel out the dynamics of the speed-tracking control loop in (2.5). Then the required relationship between drive torque and machine speed can be obtained:

\[
\frac{\omega_r(s)}{T_r(s)} = G_{\text{em}}(s) \frac{G(s)G_t(s)}{1 + G(s)G_t(s)} \frac{1 + G(s)G_t(s)}{G_{\text{comp}}(s)} = G_{\text{em}}(s)
\]  

(2.6)

From (2.6), the physical causality of the emulated load system is preserved, retaining the load’s pole-zero structure, which allows the emulation method to be used in closed-loop control. The improper compensator term (denominator, $G(s)G_t(s)$ in (2.6)) is made proper by introducing a single delay $1/z$ [62], which can itself be compensated by setting $G_{\text{em}}$ as $z \times G_{\text{em}}(z)$.

The method presented in [60, 61, 62] assumes that the test rig has approximately linear dynamics. The dynamics of the PI feedback speed-tracking controller were then integrated in the compensator term in (2.6) to compensate for the speed closed-loop dynamics so that the effective bandwidth of the load machine (dynamometer) is extended. In the emulation approach, both machines are speed-controlled and the torque control loops are not considered as their bandwidths are assumed much higher than that of the speed controllers. This approach preserves the load model dynamics and enables very good closed-loop emulation results, however, test results in [60, 61] only show the compensator’s effectiveness emulating fairly slow and low power dynamics at modest speeds. The bandwidth of the emulated linear loads (less than 0.2 rad/s) in [60, 61] is in fact well within the bandwidth (20 rad/s) of the uncompensated drive system and so the uncompensated system may be capable of successful emulation. In [62], the emulated nonlinear mechanical applications again exhibit relatively slow dynamics (around 25 rad/s).

This closed-loop emulation method in [60] uses an analytical compensator to cancel out the test rig’s speed control loop dynamics. Therefore the emulation capability and flexibility of the test rig in this method is increased compared to the model-based control design method due to the fact that there is no need to design a specific speed-tracking controller $G_t(s)$ taking into account the terms $G_{\text{em}}(s)$ and $G(s)$.

The method in [60] does however assume good knowledge of the system parameters such as the rig inertia $J$, damping coefficient $B$ and machine torque constant $K_T$ and so any
estimation errors would affect the emulation accuracy. In the aero engine emulation system in IEPNEF, the field weakening range is widely used in the drive emulator and so this, together with magnetic and current saturation effects mean parameter variation is likely which would seriously affect the performance of the analytical compensator.

Neglecting the torque control loop response of the load machine in the developed compensator in [60] introduces a torque bandwidth limit on the mechanical load dynamics that can be emulated. The inherent slew rate limit of the electrical drive also affects the emulation performance. In [60], the small inertia of the test rig permits a high slew rate, which enables good emulation performance. However, the emulator for use in the dynamic emulation of aero gas engine system, described in Section 1.3, has a high power rating and so a higher inertia. In this case, the slew rate limit is much lower than that in [60] for safety reasons, which restricts the emulation capability (this issue will be further discussed in Section 5.3).

The test system in [60] utilised a common dc link for the load emulator and the machine under test which simply circulates power. In the aero engine emulation system, the prime mover emulator and its connected generator have independent dc links and the generator is connected to the high power network so is exposed to frequent load transitions. Therefore, the analytical compensator is not suitable for use in the emulation of such a high speed and high power application. Furthermore, embedding the speed control in the analytical compensator means that if a nonlinear control method such as sliding mode control (SMC) is employed in the future, then the compensator will become very complex.

**Emulation using the inverse dynamic model combined with a feedback controller**

A further load emulation method has been proposed [63, 64] which is effectively an enhancement to the technique in [60, 61, 62]. The approach shown in Figure 2.7 considers the unmodelled dynamics of the test rig and uses a feedback PI controller to mitigate the effects, together with a compensator based on the inverse dynamics of the test rig to cancel out the system dynamics.
In Figure 2.7, \( T_e(s) \) is the applied electrical torque of the drive machine and \( T_F(s) \) denotes the unmodelled dynamics of the test rig, \( \omega_r(s) \) is the machine speed and \( \omega_{em}(s) \) is the emulated speed of the mechanical load. \( G(s) \) is the test rig mechanical dynamics and \( G_{em}(s) \), \( G_{comp}(s) \) and \( G_T(s) \) describes the emulated load, the compensator and the feedback controller respectively. From Figure 2.7, the relationship between \( T_e(s) \) and \( \omega_r(s) \) can be obtained:

\[
\frac{\omega_r(s)}{T_e(s)} = \frac{G(s)G_T(s) + G(s)G_{comp}(s)}{G(s)G_T(s) + 1} G_{em}(s) + \frac{G(s)}{1+G(s)G_T(s)} \frac{T_F(s)}{T_e(s)}
\] (2.7)

In the case when the test system parameters are well known and there are no unmodelled dynamics in the system, the compensator \( G_{comp} \) can be derived as the inverse model of the rotor dynamics of the test rig, and then the relationship between the input \( T_e \) and output rig speed \( \omega_r \) is simply the emulated mechanical load dynamics, \( G_{em} \). However, if the unmodelled dynamics cannot be neglected as is often the case, then in [63] it is suggested that the impacts of the disturbance \( T_F \) can be minimised through varying the PI values in the controller \( G_T \). In fact, the test rig considered in [63] is almost identical to that in [60] with only different electrical drive controllers; in [60] the drives are speed controlled whereas in [64] variable torque drives are used. This difference leads to two different compensator implementations, but with the same derivation procedure, which both heavily rely on the estimation accuracy of the system parameters. If \( T_F(s) \) is included into Figure 2.7, a new transfer function representation, considering the unmodelled dynamics \( T_F(s) \), can be derived, which is exactly the same as (2.7). That implies that the disturbance \( T_F(s) \) can also be mitigated by appropriately choosing the parameters of the speed controller \( G_T(s) \) in Figure 2.6. Furthermore, in [63, 64] the mechanisms of the unmodelled dynamics and their resulting effects have not been clearly clarified. From the information
available in literature the approach as explained in the Figure 2.7 [63, 64] may be viewed as just a different implementation of the emulation method in Figure 2.6.

2.2 Time Delay Compensation Methods

The control and power HIL systems in Figure 1.1 both include digital systems (communication and RTP) which function with finite sample rates. These finite sample rates inevitably introduce time delays which may affect the performance of the control or the emulator.

Several papers [65, 66, 67, 68] have been published on the analysis and compensation of time delay effects in controller HIL applications. [65] proposed a compensation method based on the inverse function of the voltage output error for reducing the time delay effects in the current controller of a PWM AC drive. [66] analysed and eliminated the time delay effects in a controller HIL simulation for an automotive drive system. It demonstrated that a total three-step sampling time delay (totalling 125μs) is introduced into the HIL system, which is due to the digital pulse-width-modulation (PWM) logic, ADC and DAC conversion and the encoder. An inverse system transport delay model was utilised to compensate for the small identified time delay. The study in [67] discusses the stability issue of a controller HIL for a power electronics controller, which is related to the time delay resulting from the signal processing, A/D and D/A conversion, and model integration. The Smith Predictor technique is applied in [68] to stabilise the controller HIL simulation of a jet engine fuel controller, which becomes unstable due to the time delay introduced by the transfer system such as the flow meter. All controller HIL time delays were very small in the μs range.

In contrast to controller HIL simulation, the time delay issues for the power HIL applications have not previously been explored in detail in the literature. A key difference between controller HIL and power HIL is the increased system complexity and higher power hardware which result in longer time delays than in a controller HIL. The time delay effects, which relate to system stability and emulation accuracy, have not been well acknowledged or discussed in depth and are only indirectly mentioned in papers. In [17] the limitations of the interface equipment for a STATCOM test system were analysed and a compensation method based on a second-order high pass filter was proposed. However, the time delay itself was not discussed. The work in [18] is focused on the examination of
the effects of different interface algorithms on the emulation accuracy of power HIL; the time delay is not adequately investigated. All recent studies on power HIL mainly deal with the power interface between the simulator and the hardware under test without any detailed examination on the time delay effects. In fact, for the power HIL applications (shown in Figure 1.1(b)), as the systems normally involve a drive system, a sensor circuitry, amplification devices, sometimes even a transmission network such as the CAN bus communication in [16], a relative larger delay could be introduced, which makes the time delay effects in these applications become significant.

Several compensation methods have been published to mitigate time delay effects such as polynomial extrapolation [68], phase lead compensation [69], derivative feedforward compensation [70], augmented Kalman filter [71] and the Smith Predictor [68]. The first three methods devise a new control structure to eliminate the time delay and were proposed specifically for civil engineering systems with a focus on real-time dynamic testing for substructures. Redesigning the control structure in a HIL emulator is complex, as compensating for the time delay is likely to limit the dynamic response of the drive system [72]. The augmented Kalman filter technique is often implemented by increasing the state space representation to accommodate delayed measurements [71]. This approach, which involves solving a partial differential equation and boundary condition equations, is computationally intensive, especially for large time delays and so is often impractical. The Smith Predictor method requires information on the specific time delay in the system so a modified version of the controller output can be subtracted from the input [68]. This compensation method is practical providing that the knowledge of the time delay is known accurately.

2.3 System Identification and its Applications

The system identification technique is generally applied to experimental results to construct a mathematical model of a dynamic system. To develop a system-identified model of a system an input signal transient such as a step, a sinusoid or a random signal is imposed on the dynamic system under test and its output response is recorded. Then a model of the system is designed to fit to the measured input and output responses. The structure of the parametric model must be carefully selected and then the unknown parameters of the identified model are estimated using statistical theory. The model of a
dynamic system is generally obtained by following the procedure as illustrated in Figure 2.8 [73].

Unlike mathematical modelling which derives the system mathematical model from physical laws, system identification does not give much physical information and in most cases, there is not much physical interpretation for the identified parameters of the model. However, the system identified model will produce an output in response to a specific input which will be similar to the original system response. The degree of similarity is the ‘fit’ as explained in detail in Chapter 4. The identified model is often only valid for a finite bandwidth, a limited operating window or a certain kind of input.

2.3.1 System Identification Experiment Design

As shown in Figure 2.8, the first step of the system identification process is to design the experiment according to the planned use of the identified model [73]. The experiment must excite the system to capture the dynamics of interest, for example if a model of a speed controlled electrical drive is required then a reference speed step would be applied to the system and actual speed can be recorded. Therefore, the input to the identified model is reference speed and its output is measured speed. The system identified model should represent the system in the frequency bandwidth of interest.
The next step is to identify the system model from the measured data collected from the identification experiment. In literature, many identification methods have been discussed for both linear time invariant (LTI) and nonlinear models. Assuming the system has approximately linear dynamics over the frequency range of interest as is the case in Chapter 4 and Chapter 5, then the LTI model identification methods are appropriate. For the linear system model identification, two main approaches exist [73, 74]:

- **Nonparametric Methods:** Nonparametric methods are applied to estimate models that are not characterised by a parameter vector. The identified models are usually described by tables, functions and curves such as the step response, impulse response or frequency diagrams (Bode plots). The nonparametric identification is normally performed by means of transient analysis, frequency analysis, correlation analysis and spectral analysis [74].

- **Parametric Methods:** The parametric methods are applied to fit models that are described by a set of parameters derived from both transient and frequency response data. In general, the methods are performed by means of the least squares estimate and its extensions [74].

In this research, the system identified model is implemented in the RTP and so for computation reasons and ease of implementation the parametric identification method is preferred. Parametric models have two common structures, state space and transfer function models, which are both suitable for implementing in the RTP. For parametric identification, the state space model of the system is normally derived from the frequency response, while the transfer function model is normally identified from the transient response [74]. The frequency response method involves time consuming experiments and requires an expensive network analyser [75]. Therefore, in this research, the transient response based system transfer function identification method is preferred.

The transfer function method requires an appropriate model structure and order to be determined which is either set using knowledge of the system or using a trial and error technique.

In the emulation system a transfer function of the compensator is required to cancel the drive system dynamics. So using the transient analysis in the system identification process enables the transfer function to be directly obtained. The input signal used for system
Chapter 2  Literature Review

Identification has a significant impact on the identification performance. Some types of input signals which are often used in the identification experiments include:

- Step function
- Pseudorandom binary sequence (PRBS)
- Autoregressive moving average process
- Sum of sinusoids

In order to obtain an accurate model over the bandwidth of interest, an input signal should be selected to give a persistent excitation to the system [73, 74]. The step signal is particularly well suited to systems with large signal-to-noise ratios as the magnitude of step can be sized to provide full information on the system dynamics such as static gain, settling time and overshoot. Then the step response together with the input step can be used to fit an accurate transfer function model to the system within the frequency of interest. More details regarding on the system identification such as model structure and order selection and fit rate calculation are given in Appendix A and Chapter 4.

2.3.2 Applications

The system identification approach is often used to develop subsystem models for use in large system level simulations [30, 76, 77, 78, 79] to investigate the overall system performance and stability. In [76], a model which reproduces only the average input-output behaviour of solid-state power controllers for use in the power systems was developed using the system identification technique. The study in [77] developed a behavioural model for predicting the electromagnetic interference (EMI) noise in a power converter based on the three-terminal equivalent Norton network. The work in [78] proposed a model for dc-dc converters based on a non-linear static model together with a linear dynamic model, featuring the protections and start-up of the converter. In [30], a behavioural model was developed for a switched reluctance generator for use in aircraft power systems simulations, which significantly reduces the complexity of the whole system and enable a shorter simulation time. Similarly, the authors in [79] proposed a model of three-phase voltage source inverters based on only the input-output responses of the dc-ac converter for use in system level analysis.

System identified models of electrical systems are produced for use in larger system level simulations to examine the performance of energy management controllers or assess and
mitigate system level stability issues. The models do not take any system internal structure into consideration and only replicate the required response.

2.4 Conclusions

In this literature review, the advancements and future challenges of the more-electric aircraft have been fully described in detail. It is expected that the electrical power demand will be significantly increased in future aircraft design. The embedded generation in which the generator is directly employed within the core of the aero gas engine is becoming the main trend of future generation. Due to the highly dynamic on-board electrical power network and the vulnerable mechanical components in the aero engine system, it is possible to produce electro-mechanical interactions, which could have a great impact on both the electrical and mechanical sides.

Little consideration has until now been given to investigate such a two-way interaction within the aero gas engine electro-mechanical system. It is then very important to understand the implications of the interaction in the laboratory environment before a final integrated field test is completed. HIL techniques based mechanical system emulation offers many benefits such as the repeatability, low cost and safety in terms of the lab demonstration and investigation of the aero electro-mechanical interaction. The mechanical source system is normally emulated by following the model-based control design approach, which is limited to a specific system and has a reduced flexibility. For the mechanical load emulation, the main approach is focused on the development of an analytical compensator for use in the compensation of the electrical drive dynamics. This approach heavily depends on the good estimation of the system parameters and does not consider the effects of the system parameter variation and external disturbances such as the electrical load and speed.

In addition, due to the possible CAN bus transmission network and sensor circuitry in power HIL applications, a large time latency is inevitable, which may significantly affect the system stability and emulation accuracy. The consideration of such a delay effect in the power HIL systems has not been explored in the previous work. This thesis investigates the mechanism of the time delay and its resulting effects in IEPNEF to illustrate the importance of the latency effects in the power HIL applications. Potential appropriate
delay compensation techniques are examined using a transient simulation model of HP spool emulation system in IEPNEF.

For the power HIL emulation systems like IEPNEF, the emulator is highly subject to the influences of operating conditions such as the generator electrical power and the operating speed. The model-based control method can allow a good emulation of the existing mechanical engine model, which is implemented in the RTP. However, its flexibility is limited and a redesign of the controller is required for each new application such as a future high power density engine model. A novel compensator design method based on the parametric system identification technique is proposed and its effectiveness has been demonstrated through the emulation of a wide variety of mechanical systems which includes both the different engine dynamics and high order systems such as mechanical drivetrain and backlash vibration effects.
Chapter 3

Emulator System

In this chapter the hardware-in-the-loop (HIL) emulator drive hardware and the coupled starter/generator in the IEPNEF laboratory are described, together with a simulation model of the system. Experimental results are included to validate the emulator drive simulation model. The model is then extended to include a thermodynamic and mechanical model of a Rolls-Royce gas engine; further test results are used to validate the full emulator drive, starter/generator, gas engine simulation model.

The time delay in the emulation system is then quantified and compensation methods are examined, for time delays in the closed-loop path as oppose to the existing open-loop setting of the system, using the validated simulation model. The time delay compensation methods evaluation in Section 3.3 formed the basis of an IEEE Industry Applications Magazine article [80].

3.1 Hardware System

The emulator drive used in this research forms part of the IEPNEF system [25] described in Section 1.3. Only a subset of the full IEPNEF system is used in this research. The remainder of this section describes in detail the equipment used, together with identification of potential time delay sources in the emulator drive. The IEPNEF system was fully commissioned before this research commenced and so only minor modifications to the hardware have been required to enable this research to be undertaken.

3.1.1 Configuration

A single emulator drive from the IEPNEF lab is used in this research to examine electrical drive based emulation of mechanical systems. From Figure 1.2, the HP spool emulator drive is used as this requires a very wide speed range of operation at high power; any emulation techniques developed for the HP spool are likely to be suitable for the lower speed but higher power LP spool. The test hardware used in this research, HP spool emulator drive, real-time platform (RTP), the coupled starter/generator and loads from Figure 1.2 are shown in detail in Figure 3.1.
The emulator drive in Figure 3.1 consists of a commercial 115kW Control Techniques Unidrive bidirectional four quadrant converter (Model: SPMD1402) with an Oswald 115kW, 15,000rpm squirrel cage induction machine (IM). The bidirectional converter is formed by two six-switch converters; an AC/DC rectifier connects to the grid and is controlled to regulate DC-link voltage and ensure unity AC power factor at the grid, and a DC/AC inverter interfaces the DC-link and IM and is controlled to regulate the speed or torque and the flux in the IM. Each converter has an embedded microcontroller which is used to command the switches in the power inverter. The inverter is controlled using a vector control induction machine strategy and space vector PWM with a 3kHz switching frequency [81] and the DC-link voltage has been measured as 697V. Limited control details are given and so this research assumes the use of a standard $d$-$q$ based vector control system which can be configured to regulate speed or torque.

The emulator drive is coupled to a commercial 30kW switched reluctance starter/generator (SRSG) from SR Drives which consists of a three-phase machine with twelve stator poles and eight rotor poles, and a conventional three-phase half-bridge converter. The SRSG can be operated as a motor from zero to 15,000rpm, with the emulator drive imposing a load on the motor, and when the SRSG is operating as a generator the DC bus is regulated to 540V and the machine can supply a maximum of 30kW over a speed range of 8700rpm to 15,000rpm [30].

A Torquetronic inline torque transducer is installed between the emulator machine and the SRSG. The transducer uses the phase shift principle [82] with 100 ‘teeth’ on each signal wheel and measures torque and speed. The Torquetronic is rated to 20,000rpm and 150Nm. The torque sensor is connected to the amplifier circuits which can be configured for
different time constants (filter cut-offs) and signal output scaling for analogue outputs. The amplifier is user-configurable and the two analogue outputs can be set to output any two from torque, speed or mechanical power. In this research two 0-10V analogue outputs are used to transmit torque and speed to the RTP. The analogue outputs are connected via a custom-designed filter board to the RTP.

The custom filter board has six channels and a maximum input voltage of 10V. Standard RC low pass filters are implemented in the board with a cut-off frequency of 500Hz to remove the switching noise from the drive system. The filter board designs are given in Appendix B.1.

The flight control system (FCS) is a dSPACE real-time platform (RTP) which can be autocoded using MATLAB/SIMULINK with the relevant toolbox; details of the hardware specification are outlined in Appendix B.2. The sample rate of the RTP is 1ms and the RTP ADC has a 1.1μs [83] conversion time per channel. The RTP contains a thermodynamic and mechanical gas engine model from Rolls-Royce and flight profile data; commercial confidentiality prevents the exact model being listed. The RTP receives analogue voltage signals equivalent to torque and speed from the Torquetronic measurement equipment, and has bidirectional communication via CAN to the emulator drive and via Profibus to the resistive loads and the master PLC. These communication channels enable the RTP to execute flight profiles on the IEPNEF hardware by coordinating the gas engine, generator and load behaviour. A reference speed value is sent to the emulator drive converter which is operated in the speed control mode. In the RTP simple linear scaling (listed in Appendix B.3) is applied to the analogue voltage signals from the torque meter to compute the speed and torque from the scaled voltage equivalents.

The SRSG generator is loaded using resistive load banks which are connected to the IEPNEF 540V DC power network. The resistive load contains 120Ω, 60Ω, and two 30Ω resistors. Contactors in the resistive load enable the specific load to be selected with load steps of approximately 2.5kW, reaching a maximum of 26.7kW at 540V DC.

### 3.1.2 Parameters

Many parameters from the test system in Figure 3.1 are required to parameterise a simulation model of the emulation system; existing models are used of the SRSG [30] which are introduced later in this chapter. The key parameters required in a simulation
model relate to the emulator drive machine and associated control. The required parameters have been obtained via simple tests or from manufactures datasheets.

**Induction machine parameters**

The rated parameters for the Oswald squirrel cage induction machine model QDI16.3-2FI are listed in Table 3.1. Also shown in Table 3.1 is the pole number for the induction machine.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated voltage, $V_{\text{rated}}$</td>
<td>$400\text{V}_{\text{rms-LL}}$</td>
</tr>
<tr>
<td>Rated current, $I_{\text{rated}}$</td>
<td>173A</td>
</tr>
<tr>
<td>Rated speed, $\omega_{\text{rated}}$</td>
<td>6745rpm</td>
</tr>
<tr>
<td>Rated Torque, $T_{\text{rated}}$</td>
<td>163Nm</td>
</tr>
<tr>
<td>Rated Power, $P_{\text{rated}}$</td>
<td>115kW</td>
</tr>
<tr>
<td>Rated frequency, $f_{\text{rated}}$</td>
<td>113.5Hz</td>
</tr>
<tr>
<td>Pole pair, $pp$</td>
<td>1</td>
</tr>
<tr>
<td>$I_m$ (magnetizing current)</td>
<td>55A</td>
</tr>
<tr>
<td>Power factor, $PF_{\text{rated}}$</td>
<td>0.91</td>
</tr>
<tr>
<td>Efficiency, $\eta$</td>
<td>96%</td>
</tr>
</tbody>
</table>

The electrical parameters of the Oswald induction machine have been found from the manufacturer’s datasheet [84] and are summarised in Table 3.2. Simple impedance measurements were performed on the stator circuit and the stator winding resistance value agrees well with the datasheet. Terminal measurement of motor reactance with the rotor present gives a value of $128\text{m}\Omega$ which is approximately the sum of the stator and rotor leakage reactance. Full IM parameterisation was not possible for safety reasons associated with the high synchronous speed (6810rpm) and difficulties performing the locked rotor test as the system is laser aligned and the SRSG is a relatively low power system (30kW).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Data Sheet Value</th>
<th>Measured Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator winding resistance, $R_s$</td>
<td>10.5mΩ</td>
<td>11mΩ</td>
</tr>
<tr>
<td>Rotor winding resistance, $R_r$</td>
<td>8.6mΩ</td>
<td>N/A</td>
</tr>
<tr>
<td>Stator leakage reactance, $X_{ls}$</td>
<td>78mΩ</td>
<td>$X_{ls}+X_{lr}=128\text{m}\Omega$</td>
</tr>
<tr>
<td>Rotor leakage reactance, $X_{lr}$</td>
<td>47.5mΩ</td>
<td>N/A</td>
</tr>
<tr>
<td>Magnetizing reactance, $X_m$</td>
<td>4.85Ω</td>
<td>N/A</td>
</tr>
</tbody>
</table>
Separately identifying the inertia of each component within the emulator system would require the IM machine to be uncoupled from the torque transducer and SRSG which would be costly as the system must be laser aligned due to the high speed and power requirement. Also, as is discussed later the SRSG model has no mechanical equation, and so no inertia, therefore the total system inertia must be lumped in the IM model in order to enable correct simulation of the emulator with SRSG. The total system inertia is therefore the sum of each subsystem within the emulation facility (induction machine, SRSG, torque transducer and the common shaft). In this study, the inertia of the emulator system is calculated from experimental measurements in [85] and is listed in Table 3.3. An estimation error of ±10% of the total inertia is assumed to represent the difficulties in accurately measuring the system inertia. Due to the relatively small size of the machines the damping is assumed to mainly come from bearing friction and so the damping can be determined by a coast-down test which measures the deceleration rate in machine speed under zero torque. The coast-down test is detailed in Appendix B.4 and the damping value calculated is given in Table 3.3. The damping value is a nominal average over the full operating speed range.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inertia, J</td>
<td>0.11kgm²(±10% accuracy)</td>
</tr>
<tr>
<td>Damping coefficient, B</td>
<td>0.0018Nmsrad⁻¹</td>
</tr>
</tbody>
</table>

A simple mechanical model is used to describe the system in the simulation environment and so the dynamics are governed by the inertia of the system and the damping coefficient as shown in Table 3.3. The shaft stiffness has a very high estimated value of 24698Nmrad⁻¹ [82], giving a natural frequency of approximately 333Hz. Therefore it is assumed to be infinitely stiff and so the effect of shaft stiffness can be neglected.

**Emulator drive control parameters**

The control settings from the emulator drive can be downloaded from the converter drive using the CT net interface and the relevant settings are listed in Table 3.4. The speed controller setting shown in Table 3.4 can be converted from internal units to new SI units (shown in brackets in Table 3.4) by introducing a scaling factor of 180 which is the conversion between the speed controller output and the torque producing current. The controller tuning was done as part of the commissioning process and the autotune function
embedded in the Control Techniques software was used to tune the gain in the proportional and integral (PI) control functions. The settings for both the machine torque constant and the speed slew rate are also included in Table 3.4 as these parameters have a significant effect on the emulator drive’s performance.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Speed controller proportional gain, $K_P$</td>
<td>0.025s/rad (4.5As/rad)</td>
</tr>
<tr>
<td>Speed controller integral gain, $K_I$</td>
<td>1.5rad$^{-1}$ (270Arad$^{-1}$)</td>
</tr>
<tr>
<td>Current controller proportional gain, $K_{PI}$</td>
<td>36V/A</td>
</tr>
<tr>
<td>Current controller integral gain, $K_{II}$</td>
<td>116Vs/A</td>
</tr>
<tr>
<td>Machine torque constant, $K_T$</td>
<td>0.85Nm/A</td>
</tr>
<tr>
<td>Speed slew rate, $K_S$</td>
<td>3759.4rpm/s</td>
</tr>
</tbody>
</table>

### 3.1.3 Time Delay Paths

Figure 3.2 illustrates the distribution of the system time delays within the emulation system, where the closed-loop speed control is on board the emulator drive and the reference speed is determined in the RTP and transmitted to the emulator drive. The latency due to the signal transmission between the RTP and the emulator drive system is the dominant delay and so the delay is in the open-loop reference speed path. In the system configuration in Figure 3.2, the CAN bus communication introduces a delay, $T_{CAN}$ into the forward path while another delay, $T_{Ana}$ is produced by the analogue transmission in the feedback path. Additional small potential sources of time delays are also identified in Figure 3.2, $T_{plc}$, $T_{s}$, $T_{Drive}$, $T_{sensor}$ and $T_{ADC}$ which are linked to the local PLC, speed controller, emulator drive system, sensor circuitry and ADC conversion stages respectively.

The embedded microcontroller in the IM converter is interfaced to CAN and Profibus via a local programmable logic controller (Control Techniques onboard PLC program). The CAN channel is used to exchange signals with the real-time platform (RTP) and the emulator drive system in Figure 3.1 and the Profibus exchanges data with a supervisory master PLC / SCADA HMI (human machine interface) which is not shown in Figure 3.1; further details on the communication channels are given in Appendix B.5. The local PLC is used for safety interlocking of the HP spool emulator drive system and manages the data flow from the different systems. The microcontroller sample rate is 4ms and the local PLC has a 10ms sample rate. The local PLC also enables communication via CT net (a Control
Techniques proprietary communication method) which is used to download/upload code to the local PLC.

![Figure 3.2: Breakdown of the time delays in the open-loop emulation system](image)

In order to obtain an estimation of the signal transmission time, Vector CANalyzer tools [86] have been used to examine the CAN communication latency between the RTP and the emulator drive system. A standard CAN bus transmission, consisting of a payload of 8 bytes, takes 5ms. However, the minimum data length for use in the transmission of the engine speed dynamics is 9 bytes as shown in Figure 3.3, requiring a Multi-Frame format, which takes longer than the standard CAN bus communication. A Vector CANalyzer based test on the forward transmission of the engine dynamics indicates that a minimum 25ms latency exists. Measuring the feedback signal transmission time is difficult as each independent measurement has an inherent sensor and transmission delay.

![Figure 3.3: Data frame of the HP spool engine speed dynamics](image)

Limited knowledge is available on the time delays in Figure 3.2. However, some information is known which enables an estimate to be made. The drive onboard PLC program has a 10ms sampling rate, resulting in an inherent delay $T_{plc}$ for the emulator system. $T_c$, which is due to the speed reference update rate, equals 4ms [81]. The drive system delay $T_{Drive}$, resulting from current controller computation, converter control, and speed measurement, is estimated as 1.25ms [81]. The RTP ADC has a sampling rate of 1.1µs per channel [83], so $T_{ADC}$ is neglected. The RTP itself has a 1ms sample rate; however, as the RTP only uses or displays data in the next sample frame, forward and feedback time delays of 2ms are introduced. In comparison to the signal transmission
delays, $T_{CAN}$, $T_{Ana}$, and $T_{sensor}$, the delays, $T_c$, $T_{Drive}$, and $T_{ADC}$, are quite small. The total system time delay, $T_{CAN}$, $T_{plc}$, $T_c$, $T_{Drive}$ and $T_{RTP}$ is estimated to be a minimum of 44ms without taking $T_{Ana}$ and $T_{sensor}$ into consideration.

3.2 Simulation Model

A companion simulation model of the test set-up in Figure 3.1 has been developed to provide an environment to evaluate the emulation capability of the emulator drive system under a range of conditions. This section presents an overview of the separate subsystem simulation models, with experimental results for validation, and then the integration of the models is discussed with further experimental results validating the full model.

3.2.1 Overview

An overview of the simulation model is shown in Figure 3.4 consisting of the emulator drive model, SRSG model [30], resistive loads and a mechanical system model (an aero gas engine) which is to be emulated by the drive; the modelling environment only includes models of the key hardware components and effects such as the time delays identified in Figure 3.2, the microcontroller and the RTP are omitted.

![Figure 3.4: The overall diagram of the full simulation model for the emulation system](image)

3.2.2 Power Hardware

In Figure 3.4 there are two main pieces of power hardware, the emulator drive and the SRSG generator with resistive loads.

Emulator drive

The emulator drive principally consists of a squirrel cage induction machine (SQIM), power converter and control. A fifth order $d$-$q$ based SQIM model is used in the simulation model as it enables machine behaviour during transient changes to be observed, dynamic
and steady state comparisons to be made with the test rig and the effects of controller tuning to be analysed. Such machine models are widely published and the key equations are summarised in Appendix C.1-C.3. Non-linear effects such as temperature effects, saturation, iron losses, eddy currents, skin effects and space harmonics are neglected. If a reduced order machine model is used, such as those widely used in power systems, then the accuracy of the system response will be reduced [87] as the stator transients are neglected. Higher order models, such as those used in harmonic analysis [88], are increasingly complex, requiring higher computational requirements which would slow the run time of the model. The fifth-order SQIM model is considered the most appropriate for the aim of this work which is to investigate the emulation capability of the hardware system in the IEPNEF lab. The SQIM parameters in Table 3.1 and Table 3.2 are used to parameterise the SQIM simulation model.

In the simulation model only the inverter of the bidirectional converter in the test system, Figure 3.1, is modelled. The AC/DC rectifier in the test system regulates the DC-link voltage which is then synthesised by the inverter, and so the inverter DC-link is assumed to be stiff and so is modelled as a simple controlled voltage source. An averaged-value model [89] has been used for the inverter which provides information about the local averaged-value behaviour of circuit voltages and currents, however, the detail in the waveforms, such as peak value and switching transients are lost. The averaged-value models are frequently used in platform level simulations where the detail of a fully switched inverter model is not required and would simply cause significantly higher simulation run times. The averaged-value model equations which map the reference voltage output of the control algorithm to the fundamental phase voltage using the DC-link voltage are listed in Appendix C.4. The DC-link controlled voltage source is set to 697V, as measured in Section 3.1.1.

Only an overview is given in the Control Techniques manual [81] for the inverter control in the test system. From this overview, and the control parameters downloaded from the drive system listed in Table 3.4, a standard rotor flux oriented $d$-$q$ vector control method has been implemented as shown in Figure 3.5.
Figure 3.5: The system block diagram of vector-controlled induction motor drive

The $d$ axis control is formed by an inner $d$ axis current control with an outer field weakening scaling function. The field weakening is required in the emulator drive as the maximum speed of the induction machine, 15,000rpm, is approximately double the 6745rpm rated speed of the machine. Beyond the machine base speed, the voltage limit of the drive converter is reached and so the rotor flux must be weakened in order to maintain a constant output power. In Simulink, the field weakening function is implemented using the inverse proportional relationship between the magnetising current ($i_{sd}$) and the machine mechanical speed as shown in Figure 3.6.

Figure 3.6: The relationship between the machine speed and the magnetising current

Based on Figure 3.6, (3.1) is used to field weaken the flux in the machine above base speed where $i_{sd}$ varies with the measured speed, $\omega_r$.

$$i_{sd}^* = \frac{\omega_{base}}{\omega_r} i_{sd \_base} \quad (3.1)$$
The $q$ axis loop in Figure 3.5 has series connected PI controllers, the inner loop regulates the $q$ axis current (torque-producing component) and the outer loop regulates the machine speed. If the drive is torque-controlled then the outer speed loop is removed [81], leaving the drive in the open-loop mode. The rotor flux space angle in Figure 3.5 is estimated in Simulink as the sum of the rotor electrical angle and the calculated slip angle corresponding to the slip frequency based on the motor stator currents and machine speed as shown in Appendix C.3. Two separate decoupling terms ($\omega_{\text{syn}} L_s i_{sd}$ and $\omega_{\text{syn}} L_s i_{sq}$) are used to facilitate the rotor flux-oriented vector control, allowing the stator currents ($i_{sd}$ and $i_{sq}$) to be controlled independently.

The vector control PI gain values in the simulation model are set to those downloaded from the emulator drive as listed in Table 3.4, giving the speed control loop bandwidth of around 10Hz and a much higher current control loop bandwidth: approximately 1100Hz [81]. The transfer functions based on the simplified diagrams for speed/current control loops in Appendix C.5 can be used to analyse the corresponding control bandwidth of the emulator drive system and the Bode plots for (C.30) and (C.31) are shown in Figure 3.7. From Figure 3.7, the control bandwidths can be identified as 68rad/s (10.8Hz) for the speed loop and 5500rad/s (875Hz) for the current control loop, corresponding well to the those stated in Control Techniques Manual [81].

![Bode plots for speed and current control loops](image)

**Figure 3.7: Bandwidth analysis for speed and current control loops**

**Starter / generator and loads**

A system-level Simulink simulation model of the IEPNEF switched reluctance starter generator (SRSG) [30] has been developed as part of another related research project and is
available for use in this research. The SRSG generator model was developed based on a
behavioural modelling technique, which only reproduces the average behaviour of the
input-output responses. The model is particularly suitable for system-level analysis, for
example, examining the interactions with other subsystems in an aircraft power distribution
system.

The SRSG in generator mode is modelled as a controlled current source, supplying current
to the DC bus. The SRSG is controlled to regulate the DC bus voltage to a nominal value
of 540V. The control stage is realised by implementing a proportional-integral (PI) type
voltage regulator with clamping and anti-windup functions as used in the hardware system
[30]. In this model, the slow mechanical speed response is neglected as it is assumed the
speed varies much more slowly than the fast electrical transients. The SRSG generator
behavioural model [30] has been validated against a wide range of experimental results for
different operating conditions.

The SRSG generator model [30] is loaded using a switchable resistive load in the test
system, described in Section 3.1.1. In the simulation model the resistive loads are assumed
to be purely resistive and are connected directly to the SRSG generator output terminals
and so the cable impedance in the test system between the SRSG and the resistive load is
neglected.

**Experimental validation**

The integrated vector controlled IM drive and SRSG simulation model has been
experimentally validated for a range of speed and load steps over the full operating range.
A sample result comparing the performance of the simulation model, red traces, to the test
data, blue traces, is shown in Figure 3.8 for the speed and load conditions in Table 3.5. In
Figure 3.8(a) and (b), speed and torque, and SRSG DC voltage and current respectively,
the data is presented for three time windows with portions of steady-state data removed
between 5s-6s and 9s-10s to enable the transients to be more clearly viewed. Figure 3.8(c)
shows IM phase current during short time periods of the third, fourth and seventh time
steps in Table 3.5 for clarity. The expansion of the corresponding subplots in Figure 3.8 is
shown in Appendix D; the magnified figures provide an explicit view on different system
dynamics.
This time delay is investigated further in Section 3.3. A delay of approximately 0.1s is evident between the reference speed and measured speed. The black trace in Figure 3.8 is the reference speed sent from the RTP to the IM drive and a slight time delay of approximately 0.1s is evident between the reference speed and measured speed. This time delay is investigated further in Section 3.3.

The correlation between the simulation model and test data is good in Figure 3.8 for all variables during both the steady-state periods and the multiple transients. The black trace in Figure 3.8 is the reference speed sent from the RTP to the IM drive and a slight time delay of approximately 0.1s is evident between the reference speed and measured speed. This time delay is investigated further in Section 3.3.

### Table 3.5: Operating conditions in Figure 3.8

<table>
<thead>
<tr>
<th>Time step, s</th>
<th>Speed, rpm</th>
<th>Load power, kW</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>9000</td>
<td>0</td>
</tr>
<tr>
<td>0.5</td>
<td>9000</td>
<td>10</td>
</tr>
<tr>
<td>1.4</td>
<td>10000</td>
<td>10</td>
</tr>
<tr>
<td>3.4</td>
<td>9000</td>
<td>10</td>
</tr>
<tr>
<td>6.5</td>
<td>9000</td>
<td>15</td>
</tr>
<tr>
<td>8.6</td>
<td>9000</td>
<td>10</td>
</tr>
<tr>
<td>10.6</td>
<td>9000</td>
<td>0</td>
</tr>
</tbody>
</table>

The correlation between the simulation model and test data is good in Figure 3.8 for all variables during both the steady-state periods and the multiple transients. The black trace in Figure 3.8 is the reference speed sent from the RTP to the IM drive and a slight time delay of approximately 0.1s is evident between the reference speed and measured speed. This time delay is investigated further in Section 3.3.

**Figure 3.8: Experimental validation of the integrated IM and SRSG model**

(a) Speed and torque

(b) SRSG DC voltage and current

(c) IM phase current
3.2.3 Model Integration

As shown in Figure 3.4, the emulator drive system model receives the speed command from the aero gas engine model as the reference speed input and is integrated with the SRSG behavioural model by coupling the emulator machine speed output to the generator speed input. The generator model also receives data on load current and the reference DC bus voltage and calculates DC bus voltage which is subsequently used by the resistive load bank model to determine load current. In the full integrated emulation system model, the system inertia is lumped into the emulator motor model as the SRSG behavioural model itself has no mechanical equation.

Gas engine model

Rolls-Royce has an extensive library of civil and defence gas engine models. The models range in fidelity depending on the purpose of the model and the life cycle stage of the specific gas engine.

In the IEPNEF lab and this research a gas engine model is required to define the emulator operation to enable the gas engine dynamics to be reproduced by the emulator system. The gas engine model must be capable of running in real-time with a sample rate which allows the engine dynamics to be identified. As IEPNEF is used to assess future electric architectures with engine embedded generators the gas engine model needs electrical power off-take on both engine spools. Also, for IEPNEF the gas engine model needs a sub idle model to enable the electric engine start using the SRSG as a motor to be tested.

A Rolls-Royce engine used in medium sized aircraft has been identified as being suitable for IEPNEF. The model contains thermodynamic and mechanical subsystems together with a simple fuel controller and is able to reproduce sub idle and normal engine behaviour. The model runs in real-time with the gas engine having a 1ms sample rate and the fuel controller has a 10ms sample rate.

An overview of the aero gas engine model is shown in Figure 3.9, where three main components are identified: environmental conditions, fuel controller and thermodynamic gas engine model. The environmental conditions block receives the altitude and Mach number and determines the corresponding environmental conditions such as pressure and temperature to the engine model. The fuel controller uses the thrust command (PLA), turbine temperature and pressure, and both spool speeds to determine the fuel command
and the relevant mode command to the engine model. The engine model takes those inputs
along with the corresponding spool electrical power off-take and determines the spool
speed commands which are then mapped for use in the relevant emulator system.

![Figure 3.9: Overview of the aero gas engine model used in IEPNEF](image)

This model originally only had electrical power off-take on the HP spool and so a small
modification was made by Rolls-Royce to the model to enable electrical power off-take
from HP and LP spools. The electrical off-take power, speed, and inertia for the engine
model and IEPNEF emulator drive systems are shown in Table 3.6. The gas engine model
computes spool speed, using environmental data (temperature, altitude, mach number and
pressure), electrical power off-take and pilots lever angle (PLA) demand, as a percentage
of full spool speed. The percentage speed is then mapped to the IEPNEF emulator drive
system speed with 100% engine speed being equal to the maximum emulator spool speed.
The actual engine maximum speeds are not listed in Table 3.6 due to commercial
restrictions. The actual emulator drive inertias are less than the spool inertias in Table 3.6
and methods of matching the inertias using the emulation technique are presented in
Section 6.1. The full engine model has previously been fully validated against the
experimental results by Rolls-Royce.

<table>
<thead>
<tr>
<th>Table 3.6: Mechanical parameters for the aero gas engine model and the IEPNEF emulator systems</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Parameter</strong></td>
</tr>
<tr>
<td>----------------</td>
</tr>
<tr>
<td>HP Electrical Power Off-take</td>
</tr>
<tr>
<td>LP Electrical Power Off-take</td>
</tr>
<tr>
<td>HP Speed Range</td>
</tr>
<tr>
<td>LP Speed Range</td>
</tr>
<tr>
<td>HP Shaft Inertia</td>
</tr>
<tr>
<td>LP Shaft Inertia</td>
</tr>
</tbody>
</table>
Experimental validation

(a) Speed and torque

(b) SRSG DC voltage and current

(c) Engine fuel demand

(d) IM phase current

Figure 3.10: Experimental validation of the emulation system simulation model

The full simulation model, IM drive, SRSG and gas engine model has been validated using experimental data from the IEPNEF system. A selection of tests have been performed for different pilots lever angles (PLAs) and loads, and a sample result is shown in Figure 3.10 for the operating conditions listed in Table 3.7. The corresponding magnified views are shown in Appendix D. Speed and torque are shown in Figure 3.10(a), SRSG DC voltage and current in Figure 3.10(b), gas engine fuel demand in Figure 3.10(c) and IM phase
current in Figure 3.10(d); the experimental data is in blue and the simulation data in red. Again only time windows of data are shown in Figure 3.10.

Table 3.7: Operating conditions for Figure 3.10

<table>
<thead>
<tr>
<th>Time step, s</th>
<th>PLA, deg</th>
<th>Load power, kW</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>38</td>
<td>0</td>
</tr>
<tr>
<td>0.4</td>
<td>38</td>
<td>27</td>
</tr>
<tr>
<td>4.3</td>
<td>42</td>
<td>27</td>
</tr>
<tr>
<td>5.3</td>
<td>38</td>
<td>27</td>
</tr>
<tr>
<td>6.4</td>
<td>38</td>
<td>0</td>
</tr>
</tbody>
</table>

The overall correlation between the simulation model data and test data in Figure 3.10 is good for all variables. During the electrical power off-take steps at 0.4s and 6.4s the engine speed dynamics shown in Figure 3.10(a) exhibit an approximately first order response with a 0.66s time constant which are correctly reproduced by the emulator drive system. The fuel demand changes slightly with electrical power off-take as shown in the magnified plots in Figure 3.10(c). When the PLA step change is applied to the engine model at 4.3s and 5.3s a significant transient occurs in the fuel demand in Figure 3.10(c) and the engine speed in Figure 3.10(a) increases from 12125rpm to 12500rpm; a slight time delay is again visible between the measured speed and the reference speed (black trace).

Both the measured and simulated responses in Figure 3.10(a) show a higher overshoot and a longer settling time in comparison with the engine reference speed which indicates that the emulator drive system is not able to correctly reproduce the PLA engine dynamics. The emulation performance limit which is evident in Figure 3.10 results from the 40rad/s resonant peak of the speed control loop in Figure 3.7 which limits the maximum frequency that can be correctly reproduced. The engine response to an electrical load change is approximately 12rad/s, and so is correctly emulated, whereas the PLA step response is approximately 30rad/s which is around the resonant peak of the speed control loop and so is poorly reproduced. The next chapter addresses the design of a compensator to cancel the natural drive system dynamics to enable a higher frequency range from the emulator drive system to accurately reproduce the engine electrical load step and PLA step dynamics.
3.3 Time Delay Effects in Emulator Hardware

Multiple potential sources of time delays in the emulator hardware system were identified in Section 3.1.3 and data, from measurements or datasheets, was included where possible to estimate the minimum time delay of 44ms. Results included in Section 3.2.2, Figure 3.8, showed evidence of a time delay between the change in reference speed from the RTP and the measured speed of the emulator drive, however, the long dataset time length made a measurement of the delay inaccurate. Reference speed steps have been imposed on the emulator drive in Figure 3.1 using the RTP over the full speed and power range of the emulator system to quantify the actual time delay between a change in RTP reference speed and the corresponding change in measured speed. A sample plot from a 0 to 2000rpm step with a 1ms rise time with the SRG at no-load is shown in Figure 3.11 and an 89ms time delay is apparent between the reference and measured speed response. Table 3.8 contains values for the measured time delay between the change in reference speed and resulting change in measured speed for the full system operating envelope.

![Figure 3.11: Sample HP spool response due to speed step](image)

<table>
<thead>
<tr>
<th>Speed step, rpm</th>
<th>Load power, kW</th>
<th>Time delay, ms</th>
</tr>
</thead>
<tbody>
<tr>
<td>0-2000</td>
<td>0</td>
<td>89</td>
</tr>
<tr>
<td>0-5000</td>
<td>0</td>
<td>86</td>
</tr>
<tr>
<td>8000-9000</td>
<td>15</td>
<td>92</td>
</tr>
<tr>
<td>12000-14000</td>
<td>27</td>
<td>88</td>
</tr>
</tbody>
</table>

The data in Table 3.8 shows the measured time delay is approximately double the 44ms estimated value from the identified potential time delay sources in Section 3.1.3. The error is attributed to unmodelled time delays which occur in the test system but are either
difficult to measure, as in the case of the speed sensor feedback, which includes $T_{\text{sensor}}$ and $T_{\text{Ana}}$ in Figure 3.2, or are not identifiable and so are not considered in the analytical estimation.

The value of 89ms was measured in the emulator hardware when the speed controller is embedded in the emulator drive, whereas the reference and measured speed are both sent or recorded by the RTP; in this case the significant time delays occur external to the speed control loop, shown in Figure 3.2, and so the time delay occurs in the open-loop path. This configuration is used in the emulator system in Figure 3.1 to minimise the computational demands on the RTP. If the emulator drive speed control was disabled (with the current control still active) and the speed control was coded in the RTP (and so a reference current value was sent to the emulator drive) then the time delays would be internal to the speed control and so the time delay would occur in the closed-loop path. Although this configuration is possible with the system in Figure 3.1 significant software modification would be required and so no tests have ever been performed with the drive in this mode.

As a result, the effect of the 89ms time delay is considered in a simulation model (from Section 3.2) for evaluation of the closed-loop path time delay compensation techniques. It is acknowledged that a closed-loop path time delay would usually be much less than 89ms in a practical system.

### 3.3.1 Closed-loop Path

Industrial drive systems and some HIL emulation systems [35, 68] are often configured with a single microcontroller which performs protection, control and measurement functions. In this case the time delays which occur throughout the system are internal to the outer control loop and so the time delays occur in the closed-loop path. The outer control loop is often regulating the speed of the system, however, other variables could be used including torque. Several compensation techniques for closed-loop path time delays were reviewed in Section 2.2, including the use of a Smith Predictor to minimise or eliminate time delays.

The Smith Predictor [68] is a well-established feedback control scheme which may be used for compensating the closed-loop path time delay effects without affecting the original system controller. In terms of practical implementation, the Smith Predictor is simple, has low computation requirements and is easy to implement in the RTP. The remainder of this
section discusses the design of the Smith Predictor for the emulator system and evaluates
the performance of the method over a range of time delay durations.

**Overview of Smith Predictor**

Figure 3.12 shows a simplified emulation system model with two lumped delays, $T_1$ and $T_2$. The closed-loop RTP control and electrical drive emulator are described by $G_x(s)$ and $G(s)$ respectively. $R(s)$ is the emulated speed dynamics and $Y(s)$ is the delayed system speed.

![Diagram of Closed-loop time delayed system](image)

In the forward path of the control loop as shown in Figure 3.2, the multiple delays can be reduced to a single delay $T_1$ as given by (3.2). In the same manner, $T_2$ can also be derived from Figure 3.2 for the feedback path, (3.3).

$$T_c + T_{CAN} + T_{plc} + T_{Drive} = T_1 \quad \text{(3.2)}$$

$$T_{sensor} + T_{Analog} + T_{ADC} = T_2 \quad \text{(3.3)}$$

The closed-loop transfer function for the system in Figure 3.12 is:

$$\frac{Y(s)}{R(s)} = \frac{G_x(s)G(s)e^{-T_1s}}{1 + G_x(s)G(s)e^{-T_1s}} \quad \text{(3.4)}$$

![Diagram of Practical implementation for the Smith Predictor in the delayed emulation system](image)
The time delays in Figure 3.12, $T_1$ and $T_2$, can be compensated by an appropriately designed Smith Predictor [68]. Figure 3.13 shows a practical implementation of the Smith Predictor for the system in Figure 3.12. The Smith Predictor uses information on the specific time delays, $T_1$ and $T_2$, to subtract a modified version of the controller output from the controller input.

Provided that the estimated delay $\hat{T}$ equals to $T_1+T_2$ and the estimated plant model $\hat{G}(s)$ is known accurately, the closed-loop transfer function for the system in Figure 3.13 can be derived as:

$$\frac{Y(s)}{R(s)} = \frac{G(s)e^{-(T_1+T_2)s}}{1 + G(s)G(s)e^{-(T_1+T_2)s}}$$ (3.5)

The system in Figure 3.13 can then be reduced to Figure 3.14 using (3.5), where a closed-loop delay-free system is obtained with only a pure transportation delay $T_1+T_2$ which occurs external to the closed-loop control.

![Figure 3.14: Equivalent reduced system for the delayed system including the Smith Predictor](image)

**Analysis of Closed-loop Path Time Delay in the Emulation System**

Figure 3.15 shows a simplified block diagram of the emulator system shown in Figure 3.1 with the closed-loop path time delays. In Figure 3.15, $G_M(s)$ is the simplified mechanical equation of motion for the system being emulated and is given by (3.6), where $J_M$ is the inertia of the system being emulated.

$$G_M(s) = \frac{1}{J_M s}$$ (3.6)

$G(s)$ is the rotor dynamics of the emulator system which are assumed to be approximately linear and are given by (3.7). The parameters $J$ and $B$ are the inertia and damping coefficient of the lab hardware (motor and generator). The emulator drive control structure has two series controllers, an inner current loop with an outer speed loop. The bandwidth
of the current loop is at least an order of magnitude larger than the speed loop bandwidth as illustrated in Figure 3.7 and so in response to changes in speed, the current loop can be assumed to be unity. Equation (3.8) defines the PI speed controller for the emulator drive, \( G_\delta(s) \) in Figure 3.15.

\[
G(s) = \frac{1}{Js + B} \quad \text{(3.7)}
\]

\[
G_\delta(s) = K_p + \frac{K_i}{s} \quad \text{(3.8)}
\]

Figure 3.15: Emulator system with closed-loop path time delay

\( K_T \) is the torque constant of the emulator drive and \( T_{GT} \) is emulated system torque. \( T_{Gen} \) is the SRSG generator torque, which varies according to the changes of electrical load. \( \omega_r \) is the motor speed, which in the RTP is delayed by two lumped delays, \( T_1 \) and \( T_2 \).

For the emulator system shown in Figure 3.15, the closed-loop transfer function between gas engine torque and motor speed is:

\[
\omega_r = \frac{G_M(s)G(s)G_\delta(s)K_T e^{-(T_1+T_2)s}}{1+G(s)G_\delta(s)K_T e^{-(T_1+T_2)s}} \quad \text{(3.9)}
\]

Figure 3.16 shows the root locus of the closed-loop transfer function in (3.9), where the total system delay, \( T=T_1+T_2 \), is varied from zero to 89ms. The parameters in (3.9) are from nominal values of inertia \( J=0.156 \text{kgm}^2 \) and damping coefficient \( B=0.0018 \text{Nmsrad}^{-1} \) and Table 3.4; \( K_P \) and \( K_I \) are 4.5 and 270 respectively, and \( K_T \) is 0.85Nm/A. \( J_M \) is set to 1.87kgm\(^2\) to match the gas engine inertia in Table 3.6. The time delay in (3.9) is approximated by a rational transfer function using the first order Pade approximation [90] so:

\[
e^{-Ts} \approx \frac{2-Ts}{2+Ts} \quad \text{(3.10)}
\]
The accuracy of first order Pade approximation is shown in Figure E.1 in Appendix E.1 where phase response for a sample delay time of 89ms show a maximum 15° error at 20rad/s.

![Root locus of the emulation system in response to time delay variation](image)

**Figure 3.16: Root locus of the emulation system in response to time delay variation**

In Figure 3.16, the delay-free system using (3.9) has two negative complex-conjugate poles close to the zero imaginary axis and a single negative zero at approximately -60seconds⁻¹. As the time delay increases, the two dominant negative poles move to the right half plane. When the time delay $T_1+T_2=24$ms, the two poles cross over to the unstable region and the emulation system will become unstable. It can also be seen that a separate single positive zero and single negative pole are produced due to the Pade approximation of the time delay which both move towards the origin as the time delay increases.

The 24ms delay time in Figure 3.16 is known as the critical time delay $\tau_c$ as defined in (3.11), where $\phi_m$ is the system phase margin and $\omega_c$ is the crossover frequency. The critical time delay is the maximum time delay the system can experience while maintaining its stability.

$$\tau_c = \frac{\phi_m}{\omega_c}$$ (3.11)

The open-loop Bode plot of the delay-free system and 89ms delayed system are shown in Figure 3.17. The phase margin of the delay-free system is 38.7° at the crossover frequency.
of 27.6rad/s which gives a critical time delay from (3.11) of 24.5ms. The calculated critical
time delay confirms the root locus analysis in Figure 3.16. Furthermore, the system critical
time delay would increase as the speed controller proportional gain $K_P$ increases or the
corresponding integral gain $K_I$ decreases. When the emulation system inertia in Table 3.3
is used in the open-loop Bode plot as in Figure 3.17, the calculated critical time delay
would decrease to 14ms.

![Figure 3.17: Bode plot for open-loop delayed system](image)

**Time Domain Analysis of the Time Delay Effects in the Emulation System**

The simulation model of the emulator system described in Section 3.2 has been used to
confirm the critical closed-loop path time delay value and to evaluate the effectiveness of
the time delay compensation methods. As the Smith Predictor is a model-based approach,
an accurate model of the hardware system (plant $G(s)$) is required to determine $\hat{G}(s)$ which
is the estimated plant model. The plant behaviour in the emulator system with closed-loop
path time delay is dominated by two elements, the IM rotor dynamics, (3.7) and the
converter dynamics which can simply be represented by the drive torque constant due to
the high bandwidth (5500rad/s) of the current control dynamics. $\hat{G}(s)$ for the emulator
drive system is given by:

$$\hat{G}(s) = \frac{\hat{K}_T}{\hat{J}s + \hat{B}}$$  \hspace{1cm} (3.12)

$\hat{J}$ and $\hat{B}$ are the estimated inertia and damping coefficient of the emulator system
respectively, $\hat{K}_T$ is the estimated motor torque constant. In order to examine the
effectiveness of the Smith Predictor in the closed-loop path time delayed emulator system, the full simulation model of the emulator drive, SRSG and gas engine from Section 3.2.3 is used with the parameters listed in Table 3.1 to Table 3.4. The gas engine experiences a 16% to 26% PLA step increase at t=20s and decrease at t=25s while the system has closed-loop time delays of 24ms and 89ms; the results are shown in Figure 3.18. In the PLA step test, the DC link voltage for the IM drive is 697V and the SRSG load is 0kW.

![Figure 3.18(a) shows the gas engine speed reference and emulator speed responses to a 10% step change in PLA for the time delay of 24ms. The 24ms time delay is selected as this is close to the critical time delay identified in Figure 3.17. The emulator speed response without the Smith Predictor contains a 5Hz oscillation with a settling time of approximately 3s as $T_1 + T_2$ is around the critical time delay. When the Smith Predictor is enabled, the correlation between the emulated speed response and reference speed in Figure 3.18(a) is very good. When $T_1 + T_2$ is 89ms, as in the aircraft emulator facility, the emulator speed in Figure 3.18(b) without the Smith Predictor enabled is unstable (exhibits an increasing oscillation which causes the solver to terminate) and so is not shown in Figure 3.18(b). Enabling the Smith Predictor stabilises the emulator speed response in Figure 3.18(b) and enables the emulator to track the reference speed, however, the time
delay $T_1+T_2$ still exists. The Smith Predictor is only effective in mitigating the time delay effects i.e. stabilising the unstable system due to the delay. The time delay $T_1+T_2$ cannot be removed by the Smith Predictor.

Figure 3.19 shows the emulator performance for a 0 to 30kW increase in SRSG load (and therefore power off-take from the gas engine) at $t=20s$ and then a 30kW reduction in load at $t=25s$ for 39% PLA. The Smith Predictor improves the emulator performance by eliminating the speed oscillation for both time delays considered in Figure 3.19.

Analysis of the Estimation Errors on the Smith Predictor Performance

The Smith Predictor is heavily dependent on the model accuracy and may be affected by parameter errors associated with estimating the parameters in (3.12) and the time delay, $T_1+T_2$. Figure 3.20 shows the effect of estimation errors in $\hat{G}(s)$ and $\hat{T}(s)$. As the accuracy of the gain of the estimated system model $\hat{G}(s)$ reduces by varying $\hat{K}_T$ in (3.12) as would happen during the field weakening range, the system will become more under damped as in Figure 3.20(a), or more over damped if $\hat{G}(s)$ is overestimated which implies that the effectiveness of the Smith Predictor would be impaired. As the estimated system time
delay $\hat{T}(s)$ deviates from the base value (89ms), some gain oscillations can be seen in the magnitude plot in Figure 3.20(b) which indicates system stability is being adversely affected.

![Figure 3.20: Sensitivity analysis to estimation errors in the Smith Predictor](image)

The system stability due to estimation errors in the Smith Predictor can also be assessed using a root locus plot. Equations (3.13) and (3.14) can be determined from Figure 3.13, (3.13) describes the system closed-loop transfer function with the estimated plant model $\hat{G}(s)$ and (3.14) is the closed-loop transfer function with an estimated time delay $\hat{T}(s)$.

\[
\frac{\omega}{T_{GT}} = \frac{G_M(s)G(s)G_p(s)K_T e^{-\hat{T}_T}}{1 + G(s)G_p(s)K_T e^{-\hat{T}_T} + G(s)G_G(s)K_T (1 - e^{-\hat{T}_T})}
\]  

\[
\frac{\omega}{T_{GT}} = \frac{G_M(s)G(s)G_p(s)K_T e^{-\hat{T}_T}}{1 + G(s)G_p(s)K_T + G(s)G_G(s)K_T (e^{-\hat{T}_T} - e^{-\hat{T}_T})}
\]  

(3.13)  

(3.14)
The root locus of the denominator of (3.13) with estimation errors in the plant model, $\hat{G}(s)$ is shown in Figure 3.21. When $\hat{G}(s)$ is overestimated in (3.13), the two negative complex poles in Figure 3.21 move to the left which indicates that the system remains stable. However, when the plant model is underestimated, the two poles move to the right half plane, where the transition into the unstable region occurs at $\hat{G}(s)=0.7G$.

![Figure 3.21: Root locus of the emulation system with varying $\hat{G}(s)$](image)

Figure 3.22 shows the root locus of the denominator of (3.14) with variable time delay estimation.

![Figure 3.22: Root locus of the emulation system with varying $\hat{T}(s)$](image)
Figure 3.22 shows that the system stability is also significantly affected by estimation errors in the time delay $\hat{T}$. When the time delay is estimated to be less than the actual delay, both complex poles move to the right, where the transition to the right half plane occurs at $\hat{T}=0.6T$. A similar result applies if the time delay is overestimated as the system loses stability at $\hat{T}=1.3T$.

Some additional poles and zeros are shown in Figure 3.21 and Figure 3.22 resulting from the first order Pade approximation of the system time delay in (3.13) and (3.14). However, as these poles remain well within the left half plane they have little effect on stability and so their effects can be neglected.

### 3.4 Conclusions

The aircraft demonstrator system, IEPNEF, has been described in this chapter with a focus on the high pressure spool emulation system hardware and its coupled generator system. The emulation system consists of a real-time platform (RTP), which contains the model of the mechanical system being emulated (in this chapter a gas engine) and an induction machine based electrical drive. A transient simulation model of the emulation system has been developed and integrated with existing models of the coupled generator system and the gas engine being emulated. The full integrated simulation model has been validated experimentally and there is a good correlation between the model and test system. The validation results have highlighted two aspects of the emulator system which may compromise the emulation performance; the first is a noticeable time delay between the change in reference speed and the corresponding change in actual speed; the second is the bandwidth of the speed control which limits the maximum frequency which can be correctly emulated.

Experimental data has been inspected over the full speed and load range of the emulation system to estimate the time delay to be 89ms. Methods to mitigate time delay effect in the closed-loop path, where the time delay is internal to the emulator drive speed control, have been evaluated using the full integrated simulation model. For the closed-loop path time delay root locus analysis using a simplified-transfer-function based system model was used to determine the unmodified emulator drive would become unstable if the time delay exceeds 24ms. Time domain results from the full integrated simulation model with the Smith Predictor during gas engine PLA and generator load steps showed that the Smith
Predictor was able to stabilise the system and also confirmed the critical time delay permissible for an unmodified stable system. The effect of estimation errors on the effectiveness of the Smith Predictor demonstrated that system stability is significantly compromised by estimation errors of ±40%.

The bandwidth limit of the speed controller results in unwanted speed variations in response to a PLA step change. The following chapter presents two model based compensator designs, an analytically derived compensator and a system identified compensator, to improve the dynamic response of the emulator drive system.
Chapter 4

Model Based Compensator Design

A proposed solution to meet the higher power requirements of future aircraft is to embed electrical generators within the engine core. Examining embedded generators in a real engine has high cost and safety implications. A solution, introduced in Section 1.2 is to use hardware-in-the-loop (HIL) techniques, where an electrical drive is commanded by a real-time platform (RTP) which contains an engine model to emulate the mechanical behavior of the engine. Section 1.2 reviewed the mechanical source emulation; two main methods were identified, a model-based control technique, in which a speed/torque controller is designed for the emulator system taking into account the emulated mechanical model, and a compensator technique which effectively modifies the control reference to cancel the natural dynamics of the emulator drive system. The compensator approach is independent of the system being emulated, providing greater flexibility on systems which may be emulated compared to the model-based control method which is limited to a narrow range of systems being emulated. The control method also has the disadvantage that the commercial drive control would need to be modified, potentially affecting the system stability.

This chapter assesses the mechanical source emulation capability of a commercial induction machine drive system which is required to emulate the high pressure (HP) spool behaviour of the aero engine in IEPNEF. Two model-based compensators for the drive system dynamics are devised to improve the dynamic emulation performance. The first compensator is designed based on a simple analytical model of the emulator drive system, whereas the second compensator uses the transient model from Section 3.2. Both methods require knowledge of the emulator drive control and so parameter sensitivity analysis is used to demonstrate the effect of estimation errors.

4.1 Simplified Mechanical Source Emulation Model

The emulator system, described in Section 3.1, principally consists of a real-time platform (RTP) and an electrical drive system, which themselves consist of multiple systems. The
emulator system can be modelled using first-order transfer functions as shown in Figure 4.1, where the drive dynamics associated with the power converter and inner current loop are neglected, since both operate much faster than the outer speed control loop and so are assumed not to influence the slower dynamics; this assumption is supported by Figure 3.7 in Section 3.2.2.

\[
G_p(s) = \frac{K_t}{Js + B}
\]

\[
G_{em}(s) = 1
\]

\[
T_{Gen}(s)
\]

\[
\omega^*(s) = \frac{1}{Js + B} = G_{comp}(s)
\]

\[
G_{em}(s) = \frac{1}{Js + B}
\]

\[
G_{comp}(s) = 1
\]

Figure 4.1: Generic mechanical source emulation using the electrical drive

In Figure 4.1 the model of the system being emulated is implemented on the RTP, it receives external stimuli which affect the behaviour of the system being emulated, and produces the speed output reference \( \omega^*(s) \) which is sent to the emulator drive and is used to command the power electronic converter. The emulated speed \( \omega^*(s) \) is compared to the actual rotor speed \( \omega_r(s) \) of the electrical machine and the speed controller \( G_s \) then uses the speed error to determine the torque-producing current demand for the current control loop. The transfer function \( G_p \) accounts for the total rotor dynamics of the emulator machine drive and its coupled load. \( K_T \) is the drive torque constant.

### 4.2 Performance of Uncompensated Emulator System

Assuming \( G_{comp} = 1 \) in Figure 4.1 and \( T_{Gen} = 0 \), then the relationship between the emulated mechanical system model output speed dynamics \( \dot{\omega}(s) \) and the drive system output \( \omega_r(s) \) can be derived as:

\[
\frac{\dot{\omega}(s)}{\dot{\omega}^*(s)} = G_{TF}(s) = \frac{G_s(s)K_T(s)G_p(s)}{1 + G_s(s)K_T(s)G_p(s)}
\]

In order to obtain good emulation performance, the condition in (4.2) must be fulfilled over the bandwidth of the emulated mechanical system model, \( G_{em} \).
When an electrical drive system is employed to emulate a mechanical system, this requires the speed regulator $G_s$ to be designed taking into account the dynamics of both $G_p$ and $G_{em}$ to ensure the emulator system bandwidth is sufficient for the system being emulated.

The emulation system in IEPNEF consists of a Control Techniques 115kW power electronic converter based induction machine drive. The control loops (speed, torque and current) which command the switches in the power converter have been tuned using an industrial auto-tune function with the parameters summarised in Section 3.1.2. Figure 4.2 shows the Bode plot of the gain $G_s K_T G_p$ for the emulation system in IEPNEF using the estimated parameters listed in Table 3.3 and Table 3.4 in Section 3.1.2, which are summarised as $K_p=4.5$, $K_I=270$, $K_T=0.85\text{Nm/A}$, $J=0.11\text{kgm}^2$, and $B=0.0018\text{Nmsrad}^{-1}$.

The crossover frequency of the open-loop transfer function $G_s(s) K_T(s) G_p(s)$ is identified from Figure 4.2 as 54rad/s, which is assumed to be the bandwidth of the closed-loop system shown in Figure 4.1. Comparing Figure 4.2 with the condition for perfect emulation performance in (4.2), then for frequencies below 10 rad/s the emulation system should enable good emulation performance. As the frequency approaches 54 rad/s, the emulation performance will be affected by the bandwidth of the emulator and as frequency increases beyond 54 rad/s the emulation performance will be poor.
Figure 3.10 in Section 3.2.3 shows the HP spool dynamics of the engine have an approximate first order response with a time constant of 0.66s during electrical power off-take, which is fairly slow, approximately 12rad/s and so the uncompensated emulator drive with a bandwidth of 54rad/s can cope with such slow dynamics. Results in Section 3.2.3 (Figure 3.10) did however identify a condition where there is a mismatch between the engine and the emulator response for step changes in throttle demand (pilot lever angle (PLA)). When a PLA step is applied to the thrust command, the corresponding engine speed dynamics include a component at approximately 30rad/s where the gain $G_s(s)K_T(s)G_p(s)$ is close to 0dB as shown in Figure 4.2, and so good emulation performance from the emulator drive is not obtained. To improve the emulation performance for PLA step responses, a compensator will be designed to cancel the emulator drive system dynamics and so extend the bandwidth of the emulator drive. Additionally, the compensator is flexible and so will enable good emulation of a wide range of mechanical systems, including those which have faster dynamics.

### 4.3 Model-Based Analytical Compensator

The compensator is designed to cancel the natural emulator drive system dynamics so that the physical causality of the emulated system is preserved. Figure 4.1 shows how the compensator is integrated in the emulation system, where the feedforward compensator $G_{comp}$ is implemented in the RTP between the emulated mechanical system model $G_{em}$ and the emulator drive system. The reference speed sent to the emulator drive system is modified by the compensator to cancel the system dynamics. Provided that detailed information is available for the emulator system, including machine drive parameters, control structure and tuning values, an inverse analytical transfer function based compensator can be designed [60].

Considering the compensator $G_{comp}$ design, then $T_{Gen}=0$ and so the system input-output relation from Figure 4.1 is:

$$\frac{\omega_r(s)}{\omega'(s)} = G_{comp}(s) \frac{G_s(s)K_T(s)G_p(s)}{1 + G_s(s)K_T(s)G_p(s)} = 1$$  \hspace{1cm} (4.3)
where

\[
G_{\text{comp}}(s) = G_T^{-1}(s) = \frac{1 + G_r(s)K_T(s)G_p(s)}{G_r(s)K_T(s)G_p(s)} \tag{4.4}
\]

$G_{TF}$ in (4.4) is given by (4.1) and using the PI controller and rotor dynamics transfer functions shown in Figure 4.1, the closed-loop transfer function $G_{TF}$ relating $\dot{\omega}_r(s)$ to $\omega^*(s)$ with $G_{\text{comp}}=1$ can be expressed as:

\[
G_{TF} = \frac{\omega^*(s)}{\dot{\omega}_r(s)} = \frac{K_p\hat{J}_r K_T + K_i\hat{B}_s}{J s^2 + B s + K_p\hat{J}_r K_T + K_i\hat{B}_s} \tag{4.5}
\]

where $\hat{J}$ and $\hat{B}$ are the estimated total system inertia and friction coefficient respectively. $K_P$ and $K_I$ are the proportional and integral values respectively for the speed controller $G_s(s)$ in the emulator system. $\hat{K}_T$ is the estimated torque constant.

To derive the analytical system transfer function as defined by (4.5), control tuning values $(K_P, K_I)$ are required together with the estimated inertia, $\hat{J}$, friction coefficient, $\hat{B}$ and IM drive torque constant, $\hat{K}_T$. As discussed in Section 3.1, the emulator induction machine is coupled to a switched reluctance starter/generator and so for successful compensation of the emulator drive system dynamics, the total inertia must be used in the analytical compensator design. The total emulator inertia was estimated at 0.11kgm$^2$ with an estimation accuracy of ±10% in Table 3.3 in Section 3.1.2. Similarly the total emulator friction coefficient was estimated at 0.0018Nmsrad$^{-1}$ from a machine coast-down test in Section 3.1.2, which represents only the average value over the speed range.

The drive manual [81] provides an expression (4.6) for torque constant $K_T$ calculation.

\[
K_T = \frac{\sqrt{3}V_{\text{rated}} I_{\text{rated}} PF_{\text{rated}} \eta}{\omega_{\text{rated}} I_{q-\text{rated}}} \tag{4.6}
\]

where $V_{\text{rated}}$, $I_{\text{rated}}$, $\omega_{\text{rated}}$, $PF_{\text{rated}}$ and $I_{q-\text{rated}}$ are the rated values of motor voltage, current, speed, power factor and active current (torque-producing current) respectively and $\eta$ is the motor efficiency.

Rated $I_q$ is the squared difference between the motor rated current $I_{\text{rated}}$ and the no-load (magnetising) current, $I_{\text{no-load}}$, (4.7).

\[
I_q = \sqrt{I_{\text{rated}}^2 - I_{\text{no-load}}^2} \tag{4.7}
\]
In the emulator drive, the parameters used in (4.6) are shown in Table 3.1 in Section 3.1.2. The torque constant $K_T$ is then estimated from (4.6) to be 0.99Nm/A. The drive microcontroller parameter list shown in Table 3.4 gives a value for $K_T=0.85$Nm/A using a fixed motor efficiency of 90% [81], and so an estimation error will occur in $K_T$ if the $\eta$ varies. The difference between the drive’s $K_T$ value and that calculated using (4.6) indicates that the torque constant calculated by the drive is quite conservative.

Figure 4.3 shows the frequency responses measured in the experimental system at 1000rpm and 13000rpm respectively together with the response of (4.5) based on the estimated parameters in Table 3.3 and Table 3.4 in Section 3.1.2. The experimental results were obtained by superimposing a 10rpm variable of sinusoidal component on the reference speed value; either 1000rpm or 13000rpm. The frequency of the sine wave was varied from 1 to 120rad/s; a 10rad/s resolution was used up to 30rad/s and above 30rad/s, and a 2rad/s interval was used between 30rad/s and 50rad/s to capture the detail around the resonant peak. The sample MATLAB code used to obtain the corresponding drive system frequency dynamics is in Appendix F.

![Figure 4.3: Experimental and analytical estimated frequency response comparison](image-url)

Figure 4.3 shows a mismatch between the measured and estimated frequency response above 10rad/s, especially apparent for the high speed case. The estimated resonant peak is 3dB lower in magnitude and 4rad/s higher in frequency than the measured resonant peak of 8dB at 40rad/s for the speed of 1000rpm. After reaching the resonant frequency, the roll-off rate also differs with a higher gradient in the measured response. As the speed increases beyond the rated speed, the discrepancy between the analytical and experimental measured response becomes more significant.
Below 10 rad/s, the system is capable of reasonable system emulation without the need of a compensator. Above 10 rad/s, a compensator must be used, however, the analytical model is not accurate enough for the compensator design. Errors in the estimates of $J$ and $B$ may be responsible for the difference between the measured and theoretical response in Figure 4.3, as these parameters are measured values. The torque constant $K_T$ for use in (4.5) is calculated in the drive, which is conservative and will also affect the performance of the analytical compensator in a similar manner. Other effects not included in $G_{TF}$ such as magnetic and current saturation may also affect the accuracy. These effects, together with parameter errors are considered as unmodelled system dynamics.

### 4.3.1 Parameter Sensitivity Analysis

To further evaluate the source of the error between the measured and the theoretical responses in Figure 4.3, sensitivity analysis is used to demonstrate the effect of parameter estimation errors. If the analytical compensator is used in Figure 4.1 and $T_{Gen}=0$, then the transfer function relating emulated mechanical system model output $\omega^*(s)$ and emulator drive system speed $\omega_r(s)$ can be expanded out by substituting the inverse of (4.5) and the PI controller and rotor dynamics transfer functions from Figure 4.1 into (4.3):

$$\begin{align*}
\omega_r(s) &= \frac{\hat{J}s^2 + B_s + K_p\hat{K}_T s + K_i\hat{K}_T}{K_p\hat{K}_T s + K_i\hat{K}_T} \frac{K_pK_T s + K_iK_T}{s^2 + B_s + K_pK_T s + K_iK_T} \omega^*(s)
\end{align*}
(4.8)
$$

For perfect compensation, the estimated plant model $\hat{G}=G$, giving $\hat{J} = J$, $\hat{B}=B$ along with $\hat{K}_T=K_T$ and then the emulator speed $\omega_r(s)$ in (4.8) is exactly the same as emulated speed $\omega^*(s)$. When perfect compensation of the system dynamics is achieved, the Bode plot of (4.8) will have a 0dB gain, so unity gain, and a phase of 0 degree. However, it is likely that estimation errors occur in $J$ and $B$ due to the limitation on measurement accuracy and $K_T$ is likely to vary with speed under the field weakening range.

Figure 4.4 shows the impact of both larger and smaller values of the base $J$ of 0.11kgm$^2$, base $B$ of 0.0018Nmsrad$^{-1}$ and base $K_T$ of 0.85Nm/A on the compensation performance of (4.8) for the emulation system. Errors in $J$ and $K_T$ have a significantly larger effect than an error in $B$ on the mechanical system emulation performance. Both underestimation and overestimation on $J$ and $K_T$ would compromise the analytical compensator’s performance, significantly affecting the emulation performance over the bandwidth range from 10 rad/s
to 1000 rad/s. Therefore, in order to obtain good emulation, the value of inertia $J$ and drive torque constant $K_T$ must be accurately known.

Figure 4.4: Sensitivity analysis of estimation errors in $\hat{J}$, $\hat{B}$ and $\hat{K}_T$ on (4.8)
As previously mentioned, both $J$ and $K_T$ in the emulator drive system are subject to relatively large estimation errors, potentially affecting the performance of the analytical compensator. Furthermore, in practice, it is not always possible to identify or have access to all required parameters to derive the compensator, if for example a commercial motor drive system is used, as is the case in this study. Therefore, the analytical drive system dynamics compensator is not appropriate for use in this emulation system. The remainder of this section includes further analysis using the transfer function from Figure 4.1 to demonstrate the impact of speed and load variation on the emulation system dynamics.

### 4.3.2 The Effects of the Operating Speed on the Emulation System Dynamics

Induction motor drives using variable voltage and frequency (V/F) or closed-loop vector control often operate at a fixed flux below the machine base speed. Above the base speed, the induction machine flux is reduced to prevent the inverter from over-modulating as the motor enters into the constant power region. The induction machine in this emulator drive system can operate as a motor up to 15000rpm. As shown in Table 3.1 in Section 3.1.2, the emulator induction motor has two poles and a base frequency of 113.5Hz, giving an estimated base speed of 6745rpm, so for approximately 60% of the motoring speed range (6745rpm to 15000rpm), the induction motor is field weakened and $K_T$ may reduce to $0.45K_T$ from (4.6).

$G_{TF}$ (4.5) can be used to illustrate the impact of $K_T$ variation as the emulator system speed increases. Figure 4.5 illustrates the system dynamics in (4.5) for variable $K_T$ based on the estimated parameters in Table 3.3 and Table 3.4 in Section 3.1.2. The variation in $K_T$ has a noticeable effect on the phase plot only around the resonant peak frequency shown in Figure 4.5. At low frequency (<10rad/s) the gain of all plots converge to zero and at higher frequency (>60rad/s) the plots converge to $-90^\circ$. The low frequency response (below 10rad/s) of the frequency response in Figure 4.5 are relatively unaffected by the value of $K_T$. As $K_T$ increases in Figure 4.5, the magnitude of the resonant peak reduces, and the frequency of the resonance increases.
If the analytical compensator (4.4) is considered and field weakening effect is neglected in the compensator design, then the estimated torque constant in the compensator will be at least $2.22 \left( \frac{K_T}{0.45K_T} \right)$ times the actual $K_T$ in the emulator drive from (4.8). The sensitivity analysis in Figure 4.4(c) demonstrates that if field weakening is neglected in the compensator design then the emulation performance will be significantly affected over the frequency range from 10 rad/s to 1000 rad/s. Therefore the dynamics of the field weakening algorithm should also be considered for the drive system dynamics compensator design.

4.3.3 The Effect of the Generating Load on the Emulation System Dynamics

The effect of varying the load on the emulator can be represented as an external disturbance as shown in Figure 4.1. The torque $T_{Gen}$, which varies with electrical load, is applied to the emulator drive system and may affect the system dynamics.

In order to examine the effect of electrical generating load on the system output speed $\omega_r$ due to changes in load torque disturbance, $T_{Gen}$, separate small-signal transfer functions can be determined from Figure 4.1. From Figure 4.1, the transfer function relating emulator drive speed, $\delta \omega_r$, to the emulated speed dynamics, $\delta \omega^*$ and electrical generating load torque, $\delta T_{Gen}$ is:

$$
\delta \omega_r = \frac{G_p(s)G_p(s)K_T}{1+G_p(s)G_p(s)K_T} \delta \omega^* - \frac{G_p(s)}{1+G_p(s)G_p(s)K_T} \delta T_{Gen}
$$

(4.9)

where the $\delta$ prefix denotes the small signal change in a variable.
The relationship between $\delta \omega_r$ and $\delta T_{Gen}$, (4.10), can be used to perform load torque sensitivity analysis, where the speed input reference, $\omega^*$ is assumed to be constant.

$$\delta \omega_r = \frac{G(s)}{1 + C(s)G(s)K_T} \delta T_{Gen} \quad (4.10)$$

The sensitivity analysis is shown in Figure 4.6 for the existing speed controller $G_s(s)$ ($K_P=4.5$, $K_I=270$), in comparison with other PI values where some extreme $K_P$ and $K_I$ variations are used to demonstrate the sensitivity analysis. In Figure 4.6(a), it can be seen that the effect of the load torque on the motor speed is quite minor in the low frequency range (less than 20rad/s), however, the influence becomes significant between 20rad/s and 100rad/s. The resonant peak in the amplitude-frequency plot of Figure 4.6(a), where the effect of the load torque is significant, can be reduced by increasing the proportional gain, $K_P$, in the drive speed controller as illustrated in Figure 4.6(a). Figure 4.6(b) shows the sensitivity of the motor speed to the load torque in response to the variation in the controller’s integral gain, $K_I$. The resonant frequency in Figure 4.6(b) gradually increases with the increase in $K_I$ value, demonstrating the $K_I$ value has a very minor influence on the effect of load torque on speed. For the existing emulator drive speed controller settings (green traces in Figure 4.6), the effects of the load disturbance on the drive system speed response is quite minor.

It is possible to move the resonant peak above the emulation bandwidth of interest by varying the $K_I$ gain in the drive speed controller. In order to do so, from Figure 4.6(b), a very high $K_I$ value is required, which may cause large overshoot, oscillation and/or instability issues to the drive response. This further justifies the choice of a compensator based method of improving emulator performance and not a controller design based method.
Another option to develop a model based compensator is to use the full emulator drive system model described in Section 3.2 to develop the required $\omega_r/\omega^*$ transfer function. This full emulator model includes the inner current control loop, speed slew rate limit, an average value technique based inverter model and a fifth-order induction machine model.

Figure 4.6: Sensitivity analysis of the emulated speed dynamics to the electrical load disturbance

**4.4 Model-based System Identified Compensator**

Another option to develop a model based compensator is to use the full emulator drive system model described in Section 3.2 to develop the required $\omega_r/\omega^*$ transfer function. This full emulator model includes the inner current control loop, speed slew rate limit, an average value technique based inverter model and a fifth-order induction machine model.
which were all either simplified or neglected in Section 4.3. Analytically identifying the required transfer function from the full model would be complex and so a system identification based method is employed. The system identification method can be applied to complex detailed models to identify the relationship between specific input(s) and output(s), which can be used to develop a parametric transfer function based model [74]. The system identified transfer function then captures the behaviour of the drive dynamics, \( \omega_r/\omega^* \), with the speed, current controller, and the mechanical dynamics effects embedded in the single transfer function.

An overview of the emulator system with the system identified transfer function is shown in Figure 4.7. In Figure 4.7, the emulator drive system is represented by a system identified transfer function, which is developed based on the simulated input and output speeds captured from the full emulator system model in section 3.2 and a slew rate limiter. The slew-rate limiter in Figure 4.7 is set at 393.7rad/s\(^2\) (3759.4rpm/s), which is the acceleration/deceleration limit in the hardware drive system. In Figure 4.7, the transfer function between reference speed and actual speed is determined by a system identification method, to identify the transfer function, \( \omega_r/\omega^* \), a speed reference step is performed with the resulting actual speed response being captured.

Figure 4.7: Block diagram of the emulator arrangement with the system identified drive model

Figure 4.8 shows reference speed, slew-rate limited reference speed and the simulated output speed from the full emulator system model in Section 3.2 when a 500rpm speed step from stationary is applied while the generator load is zero.

Figure 4.8: Simulated speed step test from 0rpm to 500rpm, \( T_{\text{load}}=0\text{Nm} \)
Before performing the identification algorithm, the model structure and order must be determined. Appendix A discusses the system identification method and explains the model structure and order for the emulator drive system transfer function. The output error model structure is selected as this model maps well to the reference speed/output speed application, the noise model is decoupled, and the structure will yield a transfer function which can be inverted to obtain the corresponding compensator. The model is selected to have the same number of parameters as the theoretical model, (4.5), and so has a second order denominator and a first order numerator.

The slew-rate limited input data and the simulated output speed from Figure 4.8 are used with the MATLAB ‘oe’ function [91], which uses a prediction error method (PEM), to determine the drive system transfer function as:

$$G_{TF} = \frac{0.0388z - 0.03655}{z^2 - 1.96z + 0.9619}$$

(4.11)

The PEM method determines the drive system transfer function, (4.11), in the discrete domain, which is not suitable for implementation in the RTP. The corresponding model in the continuous time domain (4.12) is obtained by using the “d2c” function of MATLAB. The sampling time for the conversion was 1ms, which is sufficiently small so that the model accuracy is maintained during the model transformation over the frequency range of interest up to 120rad/s.

$$G(s) = \frac{38.43s + 2310}{s^2 + 38.88s + 2310}$$

(4.12)

The estimated model, (4.12), has to be validated to check if the “best” model is good enough in terms of reproducing the real system behaviour. A popular validation method is to compare the identified transfer function response with the full simulation response to the same input. The fitting rate, which can give a good feel on whether the essential system dynamics have been replicated, can be obtained using the “compare” function of MATLAB and the fit rate is derived as (4.13).

$$Fit\ rate = 100 \times \left( 1 - \frac{\text{norm}(y_{simulated} - y_{identified})}{\text{norm}(y_{simulated} - \text{mean}(y_{simulated}))} \right) \%$$

(4.13)

Figure 4.9 shows comparison results from the identified transfer function model, (4.12) and full simulation model for four sets of data. Experimental data fit is also included in
Figure 4.9. There is a very good correlation between the full simulation model and system identified transfer function model responses for all case tests in Figure 4.9 with the fit rate being in excess of 98.5% in all cases. A poor correspondence is evident between the experimental measurements and the system identified model with the highest fit rate being only 90%, indicating that the identified model is not a good representative of the actual test system.

![Graphs showing speed responses](image)

**Figure 4.9: Sample full model and identified transfer function speed responses**

Figure 4.10 compares the frequency response of the identified transfer function model with that of the full simulation model and also the corresponding experimental measured data. In the full simulation model, an AC frequency sweep with the amplitude of 1rpm was performed on the reference speed of 1000rpm from 1rad/s to 120rad/s, within which 70 frequencies are spaced logarithmically. The output speed was recorded and the gain and phase characteristics are then computed using Linear Analysis Tool in MATLAB. There is a very good correlation between the identified transfer function response and the full simulation model dynamics, suggesting that the transfer function model can reproduce the essential features of the full simulation system dynamics. However, a consistent mismatch between the identified model and experimental measured response is identified in Figure 4.10, further demonstrating that the estimated model in (4.12) is not capable of replicating the actual test system dynamics.
Figure 4.10: Frequency response comparison between the transfer function model and the estimated frequency response of the full simulation model

Figure 4.9 and Figure 4.10 demonstrate that the system identification method can estimate an accurate linear time invariant transfer function for the detailed emulator simulation model, which is valid for a wide speed and load range. However, the identified model cannot represent the real test rig and so is not appropriate for use in the experimental emulation system.

4.4.1 Extension to Include the Field Weakening Effect

As discussed in Section 4.3.2, the field weakening effect should be considered when developing the system transfer function in order to allow it to be valid over the full operating speed range. Figure 4.11 shows four speed steps above base speed (and so in the field weakening region) with the fitting performance for both no-load and full load scenarios, with data captured from the full simulation model and the identified transfer function, \(K_T\). The full load of 20Nm is calculated based on the generation capability of the switched reluctance generator (SRG), which is 30kW in the emulation system. In Figure 4.11, a difference is apparent between the full simulation response, which varies the torque constant, \(K_T\), in according to the operating speed and the identified transfer function response (as illustrated in Figure 4.11 ‘without FW fit’), which has a fixed \(K_T\) with the discrepancy being more noticeable for higher motor speeds.
Chapter 4  Model Based Compensator Design

(a) 500rpm step up response from 8000rpm at no-load

(b) 500rpm step up response from 10000rpm at no-load

(c) 500rpm step up response from 14000rpm at no-load

(d) 500rpm step up response from 14000rpm at 20Nm load

Figure 4.11: Sample full model and identified transfer function speed responses under field weakening operation

When the induction machine is in the field weakening region, the torque constant, \( K_T \), is inversely proportional to the motor speed \( \omega_r \) [92] and so the relationship between the torque constant and motor speed can be described by (4.14).

\[
K_T' = \frac{\omega_{\text{base}}}{\omega_r} K_T
\]  

(4.14)

where \( K_T' \) is the varying torque constant in the field weakening region and \( K_T \) is the constant value below base speed. Assuming that the system model is represented by the theoretical transfer function, (4.5), the dynamics of field weakening effect can be incorporated into the system model by substituting (4.14) into (4.5), resulting in (4.15) which is valid over the full operating speed envelop.

\[
G_{TF} = \frac{(K_p K_T s + K_T' s) \frac{\omega_{\text{base}}}{\omega_r}}{J s^2 + B s + (K_p K_T s + K_T' s) \frac{\omega_{\text{base}}}{\omega_r}}
\]  

(4.15)

The effect of field weakening can also be introduced into the identified transfer function model, (4.12) using the same approach to take the full operating speed range into consideration. The resulting transfer function, (4.16) can reproduce the simulated model
response over the whole field weakening range quite well as illustrated in Figure 4.11 ‘with FW fit’, where a much better fit performance (all>99%) between the model, (4.16) and the simulated responses is achieved.

\[ G(s) = \frac{(38.43s + 2310) \frac{\omega_{base}}{\omega_r}}{s^2 + (38.88s + 2310) \frac{\omega_{base}}{\omega_r}} \]  

(4.16)

The transfer function, (4.16) can then be inverted to obtain the corresponding compensator model, (4.17) which is suitable for use with the full simulation model of the emulation system.

\[ G_{comp} = \begin{cases} 
1 + \frac{s + 0.45}{38.43s + 2310}s, & \text{when } \omega_r \leq 6745 \\
1 + \frac{s + 0.5}{38.43s + 2310}s \cdot \frac{\omega_r}{6745}, & \text{when } \omega_r > 6745 
\end{cases} \]  

(4.17)

4.5 Conclusions

Two compensator design methods have been presented. The first was an analytical compensator design technique that was shown to be highly dependent on the estimation accuracy of the drive system parameters such as system inertia, \( J \) and torque constant, \( K_T \). In the HIL emulation systems such as the emulation system in IEPNEF, the use of the analytical compensator method is restricted due to the system unmodelled dynamics, including the estimation errors in the rig inertia and torque constant. A second design was formed using a parametric system identification method applied to the full simulation model of the emulation system. The compensator was derived using a simple system transient speed step test and was enhanced to account for the field weakening effects of the machine drive to cover the full system operating speed range. The fit performances in both time and frequency domains have yielded good correlation between the full simulation model and identified transfer function responses. However, the developed system identified compensator in this chapter is only valid for the full simulation model and has been shown to have a poor correlation with the experimental test rig.
Chapter 5

Experimental Data Based Compensator Design and Validation

To provide an improved compensator design to cancel the natural dynamics of the experimental test system described in Chapter 3, this chapter describes the design of a compensator which is based on applying parametric system identification techniques to experimental measured data. The influence of the specific operating point, speed and power, on the compensator performance is considered.

The implementation of the compensator in the real-time platform is then described. The compensator and the transfer function which is used to determine the compensator are validated using time and frequency domain data from the simulation model and experimental system.

The contents of this chapter formed the basis of an IEEE Transactions on Industrial Electronics paper [93].

5.1 Experimental Data Based Compensator Design

The process for developing a compensator to cancel the natural drive system dynamics using the system identification technique was explained in Chapter 4 where the technique was applied to the experimentally validated simulation model from Section 3.2. The same system identification approach is applied to experimental measured data in this chapter to develop a compensator which performs correctly in the test system. Some additional initial steps are required when the system identification method uses test data to ensure that the derived compensator describes only the desired relationship. Then the effects of various features of the test system, including speed, load, and the RTP discrete sample time, on the compensator are explored.

5.1.1 Signal Pre-Filtering

In the emulator drive system the mechanical systems being emulated often require a feedback variable such as speed and/or torque. Both of these variables are measured using
an inline sensor between the induction machine and switched reluctance starter/generator (SRSG). The speed and torque are output from the inline sensor amplifier as unfiltered analogue signals which are connected to the RTP via a custom designed filter board (described in Section 3.1.1). Simple first order low pass filters are built on the PCB to ensure a robust design. The cut-off frequency is set to 500Hz to remove the dominant noise components at 2kHz and 9kHz without interfering with the low frequency dynamics of interest. Further details and validation results for the filter board are given in Appendix B.1.

Figure 5.1(a) shows the measured speed from the RTP in red which has been scaled from the voltage output of the filter board to speed using the scaling in Appendix B.3. The 89ms time delay as identified in Section 3.3 has been removed by post-processing to enable the identification of the desired system transient characteristics. If this time delay is not removed then the system identified transfer function would include a redundant time delay term which could make the developed compensator model more complex. At 0.12s in Figure 5.1(a) the speed reference, black trace, was stepped from 1000rpm to 1500rpm with a 1ms rise time, with the SRSG on no-load. The overall emulator drive speed response to the reference speed step is classic second order; with the initial speed response being restricted by the drive’s inherent 3759.4rpm/s (393.7rad/s²) speed slew rate, set for safety considerations. The measured data in Figure 5.1(a) also exhibits some low frequency noise effects which should not be reproduced by the identified transfer function model and so some pre-filtering is required before the system transfer function can be identified.

![Image](attachment:image1.png)

(a) Time domain

![Image](attachment:image2.png)

(b) Fast Fourier transform (FFT) analysis

Figure 5.1: Emulator drive system speed response to 1000rpm to 1500rpm step at no-load condition
Figure 5.1(b) shows the Fast Fourier transform (FFT) analysis of the 500rpm step response for the time window between 0.24s and 1.6s shown in Figure 5.1(a). As FFT analysis is usually applied to a periodical signal it is only used in this case to identify the key frequencies dominating the initial overshoot response. The most dominant component is the 500rpm peak at zero frequency which corresponds to the speed step value. The magnified view in Figure 5.1(b) shows several peaks below 10Hz (2.2Hz, 4.4Hz, and 6.5Hz) which are associated with the initial overshoot oscillation of the step response which must be preserved. Unwanted noise in the step response is scattered throughout the frequency spectrum from 10Hz up to around 60Hz. These low frequency noise components are attenuated using a software filter to ensure the overshoot dynamics of the step response are preserved as it is compensating for these features that determines the success of the system identified compensator.

**Software filter design**

The software filter is a Savitzky-Golay (S-G) digital smoothing polynomial filter, which performs a local least-squares polynomial approximation on the data series [94]. The S-G method can preserve the main dynamics of the signal response, including the peak, minima and width, which are often attenuated by other filtering techniques (such as the moving-average filter). In addition, the S-G smoothing filter has zero phase distortion for its pass band.

The S-G filter is defined by two parameters, the polynomial order $k$ and frame size $f$ [94]. The filter pass band increases as the polynomial order $k$ increases. When the frame size, $f$ (which must be odd [94]) increases, the filter will have a higher roll-off rate in its stop band, potentially giving a better noise rejection performance. The S-G filter is implemented in MATLAB using the “sgolayfilt” function. Table 5.1 contains the performance of several S-G filter designs with different $k$ and $f$ settings together with the unfiltered case for the same 1000rpm to 1500rpm step at no-load used in Figure 5.1(a). The performance of the S-G filter is evaluated by how well the speed transient after the speed step is preserved, and so Table 5.1 lists settling time and overshoot to aid the comparison.

Table 5.1 in general shows better preservation of the speed transients for higher polynomial order $k$ or lower frame size $f$. The S-G with $k=5, f=255$, highlighted in bold in Table 5.1, offers the best performance in Table 5.1 having a 3.3% error in settling time and 10% overshoot error compared to the unfiltered case.
Table 5.1: S-G filter performance with different parameter settings

<table>
<thead>
<tr>
<th>S-G filter settings (k and f)</th>
<th>Key features of the step response</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Settling time (s)</td>
</tr>
<tr>
<td>Unfiltered</td>
<td>0.62</td>
</tr>
<tr>
<td>k=5, f=201</td>
<td>0.59</td>
</tr>
<tr>
<td>k=4, f=255</td>
<td>0.58</td>
</tr>
<tr>
<td>k=5, f=255</td>
<td>0.6</td>
</tr>
<tr>
<td>k=6, f=255</td>
<td>0.58</td>
</tr>
<tr>
<td>k=5, f=301</td>
<td>0.57</td>
</tr>
</tbody>
</table>

Figure 5.2 shows step response data and the FFT for two cases; unfiltered and filtered using the S-G with k=5 and f=255. In Figure 5.2(a), the filtered step response captures the essential dynamics of the original measurement while removing the noise components. Figure 5.2(b), the FFT data performed on the same time window between 0.24s and 1.6s as in Figure 5.1(a), further confirms that the key overshoot dynamics in the raw data below 10Hz have well been preserved and the noise components above 10Hz are all significantly attenuated. The S-G filter with k=5 and f=255 is therefore used to pre-process the measured system response before the system identification is performed.
5.1.2 Transfer Function Identification

The parametric system identification approach, as used in Section 4.4 and described in Appendix A, is applied to experimental data from the emulator system in order to obtain an accurate emulator system transfer function. Either frequency or time domain data can be used [74] to determine the transfer function, with the time domain data being preferred as the frequency response method involves time-consuming experiments and requires an expensive network analyser [75]. The remainder of this section discusses the choice of time domain data used to determine the emulator drive system transfer function, and evaluates alternative transfer function model designs.

(1) Model structure selection

The structure of the identified transfer function model depends on how the relationship between the input, output and noise is described. The Output-Error (OE) model structure, described in Appendix A.2, is:

\[ y(k) = \frac{B(q)}{F(q)} u(k) + e(k) \]  \hspace{1cm} (5.1)

where \( y(k) \) is the system output, \( u(k) \) is the system input, \( e(k) \) is the system noise and \( B(q)/F(q) \) is the transfer function to be estimated.

The Output-Error model was selected as it allows the system transfer function \( B(q)/F(q) \) to be parameterised independent of the noise model. This model structure also simplifies the identification process and enables good identification performance [95].

(2) Transient response selection

Measured and reference speed, time domain data is required to enable the transfer function to be determined which expresses the relationship between these variables. The measured speed data is time shifted by 89ms to remove the time delay which was identified in Section 3.3, as such an open-loop path delay has no effect on the system’s transient characteristic and should not be reproduced by the identified transfer function.

The measured and reference speed response is best captured during a step in reference speed as this yields the fastest system response. If a speed ramp was used this would slow down the system response and so limit the frequency bandwidth of the identified model.
Figure 5.3 shows the step response for four different step magnitudes, 100rpm, 500rpm, 1000rpm and 2000rpm, from an initial value of 1000rpm at no-load. In all cases the test data is filtered using the S-G filter from Section 5.1.1. A simple transfer function model $B(q)/F(q)$ with a 2nd order $F(q)$ and 1st order $B(q)$ has been determined for each step to evaluate the fit performance. The fit performance in Figure 5.3 improves as the step magnitude increases with a dramatic improvement in fit of from 72.35% to 96.72%, when the step is increased from 100rpm (Figure 5.3(a)) to 500rpm (Figure 5.3(b)) due to the higher signal-to-noise ratio for the 500rpm step. For speed steps larger than 500rpm, Figure 5.3(c) and (d), the fit increases only slightly. The speed steps in Figure 5.3(c) and (d) also exhibit a slight reduction from the initial gradient as the measured speed increases beyond 1500rpm; the reason for this is not clear. Based on Figure 5.3 a 500 rpm step is used in the system identification process as this response exhibits a high signal-to-noise ratio, enables a good fit performance of the simple transfer function, can give a proper persistent excitation to the emulator drive system within the bandwidth of interest and also avoids the dual gradient region.

![Graphs showing step response model fit performance for different speeds](image)

Figure 5.3: Step response model (2nd order $F(q)$ and 1st order $B(q)$) fit performance for different speed step magnitudes

(3) **Model order selection**

Next, the order of both $B(q)$ and $F(q)$ for the OE structure is considered. Figure 5.4 shows the fit performance for four different model orders for the 1000rpm to 1500rpm step at no-
load as used in Figure 5.3(b). Good fit accuracy is obtained for both the 3rd and 2nd order model (95.64%) and 4th and 4th order model (96.84%) cases shown in Figure 5.4(b) and (d) respectively and the 2nd and 1st order model (96.72%) shown in Figure 5.3(b). Since the compensator will be implemented in the real-time platform, computational efficiency is important and so the lower order model shown in Figure 5.3(b) is preferred at the expense of a 0.12% reduction in model fit. Therefore, a 2nd \( F(q) \) and 1st \( B(q) \) has been selected for the transfer function which also matches the analytical system inner structure model order shown in Section 4.3.

![Figure 5.4: Step response model fit performances for different model orders](image)

(4) Transfer function design

As the appropriate step magnitude and model order have been identified, the system transfer function can be obtained using the MATLAB function “oe” from the filtered 500rpm step test data set shown in Figure 5.2(a). The identified drive system model in the discrete time domain, (5.2) is then converted to the corresponding transfer function in the continuous domain, (5.3) using the “zero-pole matching” method with 1ms sampling rate, matching the RTP sample rate. The fit between (5.3) and the measured response is 96.7% as shown previously in Figure 5.3(b).

\[
G_{pe}(z) = \frac{0.01664z - 0.01509}{z^2 - 1.982z + 0.9831}
\]  
(5.2)
Chapter 5  Experimental Data Based Compensator Design and Validation

\[ G_{tr}(s) = \frac{16s + 1569}{s^2 + 17s + 1569} \]  \hspace{1cm} (5.3)

A frequency chirp study was performed, sweeping the speed reference to the emulator drive system, from 10 rad/s to 120 rad/s in 10 rad/s steps; a 2 rad/s interval is used between 30 rad/s and 40 rad/s to capture the detail around the resonant peak. The chirp test amplitude was chosen to be 10 rpm (less than 0.1% of the full measurement range of 15000 rpm), which is the lowest value that can be differentiated from the noise in the signal frequency spectrum analysis as shown in Appendix G.1.

The frequency chirp reference speed, together with the measured motor speed is then used to calculate the frequency characteristics of the emulator drive system using MATLAB and sample MATLAB code used to obtain the corresponding drive system frequency dynamics is in Appendix F. Figure 5.5 shows the frequency response comparison between the identified models, (5.2), (5.3) and experimental measured frequency response of the drive system. The general correlation is excellent, with a slight 0.5 dB error between 10 rad/s and 30 rad/s and a phase error of approximate 4° between 20 rad/s to 60 rad/s.

![Figure 5.5: Fitting performance of the parametric identified drive system model](image)

5.1.3  Drive System Dynamics Compensator Design and Development

The emulator system compensator is determined by inverting the identified system model which in the continuous domain is given by (5.3). The compensator determined from (5.3) is improper, though it can be made proper by introducing an extra term \( ks^2 \), which is
associated with a certain phase shift, into the numerator of (5.3) and so the new $G_{TF}$ becomes:

$$G_{TF} = \frac{k s^2 + 16s + 1569}{s^2 + 17s + 1569}$$

(5.4)

When $k$ is chosen to be small, the inclusion of the term $ks^2$ into the numerator of (5.3) will result in a slight phase advance over a specific frequency range. Considering the emulation bandwidth of interest (120rad/s) which is limited by the emulator drive speed slew rate limit, $k$ is selected to be 0.01, which has a negligible effect on the compensator performance over the bandwidth of interest as shown in Figure 5.6, where a phase advance of around $7^\circ$ occurs at 120rad/s which is the maximum frequency of interest.

![Bode Diagram](image)

**Figure 5.6: Frequency response of original identified system model, (5.3) and modified model, (5.4)**

If $k$ is set to be a smaller value than 0.01, the phase advance due to the term $ks^2$ will occur at a higher frequency, which implies that the effect of $ks^2$ can be neglected over a wider frequency range. However, when the coefficient $k$ is further reduced, the additional zero introduced by $ks^2$ in the pole-zero plot of (5.4) in the discrete time domain, as explained in Section 5.1.4, will move outside the unit circle, as shown in Figure 5.7. In Figure 5.7, the extra zero on the left half plane of the unit circle, which is produced by the term $ks^2$, moves to the perimeter of the unit circle when $k=0.008$. When $k$ is smaller than 0.008, the system in (5.4) will become a non-minimum phase system, which may introduce an unstable pole into the corresponding developed compensator model and so should be avoided. Therefore, the limit of the $k$ value is identified as 0.008. For $k=0.01$, the corresponding compensator model is:
Chapter 5 Experimental Data Based Compensator Design and Validation

\[ G_{\text{comp}} = \frac{s^2 + 17s + 1569}{0.01s^2 + 16s + 1569} \]  
(5.5)

Figure 5.7: Pole-zero analysis of (5.4) in the discrete time domain

5.1.4 Discretisation Time Delay Effect

In the test system the developed compensator (5.5) will be implemented in a real-time platform (RTP) and so the sampling effects such as the time delay due to the digital implementation need to be considered. Figure 5.8 shows the sampled-data representation for the emulator system, where \( T_s \) (1ms) is the RTP sample time and the analogue/digital (A/D) and digital/analogue (D/A) conversions are modelled by a sampler and a zero-order hold (ZOH) respectively. The ZOH introduces a time delay which is equal to half the sampling time \( (T_s/2) \) into the discretised mechanical system dynamics [96].

Figure 5.8: Discrete time representation of the emulator system
Figure 5.9: Compensator frequency response comparison between continuous and discrete time domains

Figure 5.9 compares the frequency response of the developed compensator, (5.5) in continuous domain with the discretised equivalent (5.6) which is obtained using the zero-pole matching method [96] and the same sampling time (1ms) as used in the RTP. From Figure 5.9, it can be seen that the delay effect due to the discretisation is quite minor within the emulation bandwidth of interest (120 rad/s) and so the discretisation effect can be neglected in the compensator design.

\[ G_{comp}(z) = \frac{49.69z^2 - 98.46z + 48.85}{z^2 - 1.125z + 0.2019} \]  

(5.6)

5.1.5 Influences of Operating Conditions

The emulator system operates over a wide speed and load range and so the drive system dynamics variation with either variable needs to be carefully considered for the design of the experimental data based compensator. In parametric system identification the polytopic structure, which comprises a set of local “small-signal” linear time invariant (LTI) models [97], is often used to account for a system’s dependency on operating conditions. However, defining the weighting of separate LTI models is complex and cannot easily be done analytically. In addition, as the compensator is obtained by inverting the identified system model the polytopic model is not appropriate for this study. In this section, the system dynamics dependencies on both the speed and load conditions will be evaluated and any noticeable effects are incorporated into the compensator model by modifying the numerator and/or denominator of the compensator.
Effects of Speed Dependency

The operating speed was demonstrated analytically in Section 4.3.2 to have a significant effect on the emulator system dynamics in the induction machine’s (IM) field weakening region. In order to experimentally examine the drive system speed dependency, a series of frequency chirp tests, as used in Section 5.1.2, were performed. The test results at no-load condition are shown in Figure 5.10, with additional results for 27kW and 15kW load power shown in Appendix G.2. Figure 5.10 illustrates the dynamic dependency on the machine speed by showing the responses at 9000rpm, 11000rpm and 13000rpm at no-load. As shown in Figure 5.10, the drive system resonant frequency decreases slightly and the system damping increases as the machine speed increases. The variation of the frequency response in Figure 5.10 with speed is again attributed to the field weakening effect as all speeds shown in Figure 5.10 are above base speed and so are within IM field weakening range. The experimental results in Figure 5.10 match the analytical responses in Section 4.3.

As in Section 4.4, the speed dependency due to the field weakening effect can be incorporated into the compensator in (5.5) by considering the theoretical transfer function model of the drive system, (5.7).

$$G_{TF} = \frac{K_p K_T s + K_f K_T}{J s^2 + B s + K_p K_T s + K_f K_T}$$

(5.7)
The field weakening effect will cause the torque constant, $K_T$, to reduce when the machine is operating above base speed. The torque constant is inversely proportional to the speed, $\omega$, so:

$$K'_T = \frac{\omega_{\text{base}}}{\omega_r} K_T = \alpha K_T$$  \hspace{1cm} (5.8)

where $\omega_r$ is machine speed, $K'_T$ is the varying torque constant in the field weakening region, $K_T$ is the constant value below the machine base speed, $\omega_{\text{base}}$, and $\alpha$ is the ratio of $\omega_{\text{base}} / \omega_r$. The effect of field weakening can be considered by multiplying $K_T$ in (5.7) by $\alpha$ to give (5.9), which is valid over the full speed range.

$$G_{TF} = \frac{(K_p s + K_i)K_T \alpha}{J s^2 + B s + (K_p s + K_i)K_T \alpha}$$  \hspace{1cm} (5.9)

where

$$\alpha = 1 \quad \text{when} \quad \omega_t \leq \omega_{\text{base}}$$  \hspace{1cm} (5.10)

$$\alpha = \frac{\omega_{\text{base}}}{\omega_t} \quad \text{when} \quad \omega_t > \omega_{\text{base}}$$  \hspace{1cm} (5.11)

Since $B$, (0.0018Nmsrad$^{-1}$ from Table 3.3 in the emulation system), is usually at least an order of magnitude less than $K_p$ which is 4.5 (from Table 3.4) in the drive system, it can be neglected. Then, the $\alpha$ scaling to represent the effect of field weakening can also be used with the system identified transfer function (5.4) to give (5.12), which enables the full model response to account for the effect of the specific operating speed, subject to the clauses in (5.10) and (5.11).

$$G_{TF-\alpha} = \frac{0.01 s^2 + (16 s + 1569)\alpha}{s^3 + (17 s + 1569)\alpha}$$  \hspace{1cm} (5.12)

The numerator of (5.12) is second order and so can be compared to the classic second order differential equation to determine expressions for the corner frequency and damping factor. The corner frequency, $\omega_n$, of the emulation system, (5.13) and the damping factor, $\xi$, both decrease with the machine speed increasing beyond the base speed, $\omega_{\text{base}}$, which is consistent with Figure 5.10.
\[\omega_n = \sqrt{1569 \cdot \frac{\omega_{\text{base}}}{\omega_r}} \quad (5.13)\]

A new speed dependent compensator which takes the speed dependency into consideration can be derived from (5.12). When the machine speed, \(\omega_r\), is smaller than the base speed, \(\omega_{\text{base}}\), the inverse of (5.12) with \(\alpha = 1\) is:

\[G_{\text{comp}} = 1 + \frac{0.99s^2 + s}{0.01s^2 + 16s + 1569} \quad (5.14)\]

When \(\omega_r\) is higher than the base speed, \(\omega_{\text{base}}\), the corresponding compensator model can be obtained from (5.12) as shown in (5.15).

\[G_{\text{comp-\omega}} = 1 + \frac{0.99s^2 + s\alpha}{0.01s^2 + (16s + 1569)\alpha} \quad (5.15)\]

Since the extra term \(0.01s^2\) has a negligible effect over the emulation bandwidth of interest, as shown in Figure 5.6, and \(\alpha\) is less than one in the field weakening region, (5.15) can be modified by deliberately introducing the \(\alpha\) scaling into the \(0.01s^2\) term to give:

\[G_{\text{comp-\omega}} = 1 + \frac{0.99s^2 + s\alpha}{0.01s^2\alpha + (16s + 1569)\alpha} \quad (5.16)\]

Decoupling the \(\alpha\) scaling from the denominator in (5.16) results in (5.17).

\[G_{\text{comp-\omega}} = 1 + \left[\begin{array}{c} 0.99s^2 \\ 0.01s^2 + (16s + 1569) \end{array}\right] \frac{1}{\alpha} \quad (5.17)\]

Figure 5.11 shows the variation of the \(\alpha\) term in the field weakening range has little effect on the frequency response of part b in (5.17), and so \(G_{\text{comp-\omega}}\) can then be further simplified by setting \(\alpha\) to unity in part b in (5.17) to give (5.18) which is easier to implement in the RTP.

\[G_{\text{comp-\omega}} = 1 + \left[\begin{array}{c} 0.99s^2 + s \\ 0.01s^2 + (16s + 1569) \end{array}\right] \frac{1}{\alpha} \quad (5.18)\]
Therefore the compensator model, which accounts for the full speed range of the emulator system, is:

\[
G_{\text{comp-\omega}} = \begin{cases} 
1 + \frac{0.99s^2 + s}{0.01s^2 + 16s + 1569}, & \omega \leq \omega_{\text{base}} \\
1 + \frac{0.99s^2 + s}{0.01s^2 + (16s + 1569) \frac{\omega_\text{base}}{\omega}}, & \omega > \omega_{\text{base}} 
\end{cases} \quad (5.19)
\]

In wide speed range, high power closed-loop controlled induction machine drives the rated speed of the machine is often lower than the base speed at which the voltage limit of the converter is reached and so field weakening must occur, as in this emulator drive system, to enable the high load demands to be fulfilled.

The base speed of the drive is not known, however it is determined by the maximum possible output voltage of the inverter, which is associated with the available dc link voltage, and the rated motor parameters [99]. An approximate analytical estimate of the machine base speed (5.20) is given by equating the ratio of base speed to rated speed, to the ratio of the maximum inverter output voltage \(V_{s-\text{max}}\) to rated peak phase voltage \(V_{\text{phase-pk}}\).

\[
\omega_{\text{base}} = \frac{V_{s-\text{max}}}{V_{\text{phase-pk}}} \cdot \omega_{\text{rated}} \quad (5.20)
\]

The maximum voltage \(V_{s-\text{max}}\) that the inverter can apply to the machine is limited by the available dc-link voltage \(V_{dc}\) and the pulse width modulation (PWM) strategy used. In the emulator drive system, a voltage space vector PWM strategy is used and \(V_{dc}=697V\) [81].
and so \( V_{s\text{-max}} \) is limited to \( V_{dc}/\sqrt{3}=697/\sqrt{3} \) [92]. The nominal voltage of the star connected induction motor from Table 3.1 is \( V_{\text{rated}}=400\,\text{V (line to line rms)} \). The machine base speed from (5.20) for \( V_{s\text{-max}}=402.4\,\text{V} \) and \( V_{\text{phase-pk}}=326.6\,\text{V} \) is 8310rpm.

However, the motor rated voltage \( V_{\text{rated}} \) may vary with the temperature rise in the motor windings. For cold winding resistance values (room temperature, 20\(^\circ\)C), \( V_{\text{rated}} \) is calculated as 330V (line to line rms) at rated load and rated slip (details are given in Appendix G.3) which gives a machine base speed of 10,073rpm from (5.20). If the hot resistance values are considered \( (R_{\text{hot}}=1.4R_{\text{cold}}) \), then \( V_{\text{rated}} \) is calculated as 385V at rated load and rated slip and so the machine base speed is 8634rpm from (5.20). This demonstrates significant variation in estimated machine base speed if slight temperature effects are considered, and so accurately calculating \( \omega_{\text{base}} \) is difficult.

The estimated base speed range from (5.20) can be validated using the test data from Figure 5.10. The corner frequencies for the three speeds shown in Figure 5.10 are plotted in Figure 5.12 (grey stars) together with (5.13), for base speeds of 8300rpm (400V\( _{\text{LL}} \)), red trace and 10073rpm (330V\( _{\text{LL}} \)), blue trace. The test data points in Figure 5.12 are bounded by the two analytical estimates of base speed, with the test data being approximately half way between the two estimates. Selecting a base speed of 9000rpm in (5.13) yields a curve (black trace) which correlates very well with the test data points. Therefore the base speed of 9000rpm is used in the compensator model, (5.19).

![Figure 5.12: The relationship between actual speed and frequency ratio](image)

**Effects of Load Dependency**

In this section, the drive system dynamics dependency on the load operating points is evaluated by applying the frequency chirp tests, as used in Section 5.1.2, to the emulator system under different load and speed conditions. The SRSG generating range is limited to
between 8700rpm and 15000rpm [30]. Figure 5.13 shows 27kW, 15kW and no-load frequency responses at 9000rpm and 13000rpm when a 10rpm amplitude variable sine wave is added (0-120rad/s) to the speed. From Figure 5.13(a), it can be seen that the resonant frequency in the on-load cases at 9000rpm is 4rad/s lower than the no-load peak at 40rad/s with a slightly higher damping. 9000rpm is the lowest speed tested for the generator to ensure the minimum SRSG speed is not exceeded during large load steps. For the higher speed of 13000rpm, Figure 5.13(b), the different load traces are very similar; indicating that the generator load has little effect on the drive system dynamics at higher speeds.

![Figure 5.13: Experimental frequency response at 9000rpm and 13000rpm with variable load](image)

The results in Figure 5.13 show that the load influence on the drive system dynamics is more noticeable at low speed. For the speed of 9000rpm, the on-load frequency responses have a lower resonant frequency with a higher damping, which exhibits a similar effect as that of the speed dependency field weakening effect. Therefore, within the emulation bandwidth of interest, the load dependency at 9000rpm can be incorporated into the compensator model by introducing an additional $\beta$ scaling factor. The $\beta$ factor value, (5.21) at 9000rpm can be experimentally derived as the square of the ratio of the on-load corner frequency $\omega_{n1} = 36$rad/s in Figure 5.13(a) to the no-load corner frequency $\omega_{n2} = 40$rad/s from Figure 5.13(a):

$$\beta = \frac{\omega_{n1}^2}{\omega_{n2}^2} = \frac{36^2}{40^2} = 0.8$$

(5.21)

As the machine speed increases, the load influence become less significant as shown in Figure 5.13(b), where the $\beta$ scaling factor is approaching unity at 13000rpm. In order to account for the load dependency over the whole generating speed range of 9000rpm to
15000rpm, a weighting function on the $\beta$ scaling factor needs to be defined. Analytically identifying the weighting function is very difficult as the load effect is very minor within the high speed range. In this case, based on the assumption that the load influence varies linearly within the generating range, the $\beta$ scaling factor is weighted as a linear function of the speed operating points as illustrated in Figure 5.14, where the $\beta$ scaling factor is assumed to be unity at the maximum speed of 15000rpm.

![Figure 5.14: The relationship between the $\beta$ scaling factor and the speed operating point](image)

The relationship between the $\beta$ scaling factor and the operating speed from Figure 5.14 is given by:

$$\beta = 0.8 + \frac{0.2}{9000}(\omega_r - \omega_{base}) \text{ when } \omega_r > \omega_{base}$$  \hspace{1cm} (5.22)$$

Incorporating (5.22) into (5.12) results in a system transfer function model (5.23) for the generator speed region $\omega_r > 9000$rpm, which accounts for both the speed and load effects.

$$G_{TF-\omega_r} = \frac{0.01s^2 + (16s + 1569)\alpha\beta}{s^2 + (17s + 1569)\alpha\beta}$$  \hspace{1cm} (5.23)$$

Therefore the compensator model for generator operation, which accounts for speed and load dependencies, can be obtained from (5.23) as:

$$G_{comp-\omega_r} = 1 + \frac{0.99s^2 + s}{0.01s^2 + (16s + 1569)} \cdot \frac{1}{\alpha \cdot \beta}, \omega_r > 9000$$(5.24)$$

As the machine speed increases from 9000rpm, the system load dependency which is associated with the $\beta$ scaling factor reduces. The compensator in (5.24) is equal to that in (5.19) when the system speed is the maximum value of 15000rpm.
5.1.6 Effect of the Current Loop Response in the Drive System

The compensator in Section 5.1.5 is designed for use in eliminating the speed control loop characteristics of the emulator drive system. The details of the speed control and any inner current control loops are not relevant to the compensator design and so can be neglected providing that the inner current control is sufficiently fast compared to the outer speed control. For the emulator drive system, from Section 3.2.2, the current loop bandwidth is 1100Hz and the speed control loop bandwidth is 10Hz [81], and so neglecting the current loop dynamics will have a negligible effect on the compensator performance. If this assumption is not valid or if only the inner current control is used, as the machine speed increases into the field weakening region the available voltage for controlling the current may be limited and so applying a step reference to the current loop could potentially result in a large overshoot and long settling time in the current response. In this case the proposed compensator method might offer a limited emulation performance and the current loop dynamics could be considered to be incorporated into the model which is the subject of future work.

5.2 Experimental Validation

Experimental validation tests have been undertaken to evaluate the performance of the speed-varying, and combined speed- and load-varying transfer functions, $G_{TF-\omega}$ (5.12) and $G_{TF-p_o\omega}$ (5.23) respectively, as it is these transfer functions that are used to determine the compensators. Further experimental tests are then presented for the compensators $G_{comp-\omega}$ (5.19) and $G_{comp-p_o\omega}$ (5.24) which were both implemented in the RTP to show their performance and enable any performance limits to be identified.

5.2.1 Time Domain

In Section 5.1.5, two system transfer function models were derived, $G_{TF-\omega}$ (5.12) which included only the speed dependency ($\alpha$ scaling), or $G_{TF-p_o\omega}$ (5.23) which included both the speed and load effects ($\alpha$ and $\beta$ scaling).

Sample results from the experimental validation tests for $G_{TF-\omega}$ and $G_{TF-p_o\omega}$ are shown in Figure 5.15 to Figure 5.17 for a range of speeds and loads over the allowable operating range. Time domain results are used in this section to enable the transient performance to be evaluated. Figure 5.15 shows the response of $G_{TF-\omega}$ and the measured drive response.
(after signal pre-filtering) for no-load condition (which is why $G_{TF-p0}$ is not considered).

Step responses at two different speed conditions (3250rpm, so below the base speed and 11950rpm which is above base speed, so in the field weakening range) are chosen to show cross validation over the full speed range, while avoiding the conditions used for the system model identification. The correlation between the measured and identified model response is excellent in both validation tests, which demonstrates that $G_{TF-\omega}$ (5.12) is able to accurately reproduce the no-load system dynamics.

![Graph](image1)

(a) 700rpm step up response from 3250rpm at no-load condition

![Graph](image2)

(b) 700rpm step down response from 11950rpm at no-load condition

Figure 5.15: Time domain validation test for no-load condition

Figure 5.16 shows step responses at four different speeds when the generator is on load and so $G_{TF-\omega}$ and $G_{TF-p0}$ are shown together with experimental data. For the validation tests at 7.29kW, Figure 5.16(a)-(b), both system models exhibit a response which correlates very well with the measured drive speed dynamics. As the system speed increases further to approach the maximum operating point of 15000rpm, the results in Figure 5.16(c)-(d) show a slight time shift between the model responses and the actual drive system dynamics with the effect being more noticeable for the higher speed test in Figure 5.16(d). Since the system is approaching the maximum power and speed, the time shift is likely to be associated with the saturation of the machine speed controller which is attributed partly to the converter overcurrent protection features designed to avoid the saturation of the leakage inductance path and prevent the motor overheating [100]. In Section 5.1.5 $G_{TF-\omega}$ and $G_{TF-p0}$ were both designed assuming a linear operating range of the emulator system.
and so they are not able to reproduce any nonlinear dynamics of the system such as the controller saturation effects as in Figure 5.16(c)-(d). However, the correlation between the data sets is still good and the system model responses are stable when the saturation region is reached. In all tests in Figure 5.16, the responses of $G_{TF-co}$ and $G_{TF-pco}$ are overlapped, indicating that the additional load scaling factor makes little difference to the transfer function responses for speeds above 11000rpm.

Figure 5.16: Time domain validation test for on-load condition

Figure 5.17 shows the measured emulator system speed response in red together with $G_{TF-co}$ (grey trace) and $G_{TF-pco}$ (blue trace) for a speed step from 9000rpm to 9500rpm with a 15kW load power.

Figure 5.17: Time domain validation test at 9000rpm with 15kW power

In Figure 5.17, the measured response and $G_{TF-pco}$ response have a 94% fit, with the $G_{TF-co}$ having an 87% fit to the measured data. This shows the load effect is more noticeable at
lower speeds and this effect is better captured by $G_{TF-p\omega}$ than $G_{TF-\omega}$. The $G_{TF-p\omega}$ function is however more complex and requires measured power information in addition to the speed.

### 5.2.2 Frequency Domain

The results in this section directly show the performance of the natural drive system dynamics compensators. As in Section 5.2.1, two compensators are examined, $G_{comp-\omega}$ (5.19) which only has speed dependency and is determined from $G_{TF-\omega}$ (5.12), and $G_{comp-p\omega}$ (5.24) which has speed and load dependencies and is derived from (5.23). Both $G_{comp}$ functions were derived assuming that the system dynamics are approximately linear over the frequency range of interest and so the system input-output behaviour can be described by the frequency response. The frequency response tests demonstrate the compensator bandwidth and enable the performance of the compensators for a wide range of operating conditions to be shown on a single graph.

Figure 5.18 shows the test set-up for the frequency response validation tests where the variable frequency speed reference and $G_{comp}$ functions are both implemented on the RTP. The compensator models are implemented in the RTP, receiving the pre-programmed variable frequency reference speed signal as input. The variable frequency speed reference is created by superimposing a 10rpm amplitude variable frequency chirp signal on the reference speed as in Section 5.1.2. The chirp signal is swept from 10rad/s to 120rad/s in 10rad/s steps, with additional 2rad/s steps used between 30rad/s and 40rad/s to capture the detail of the resonant peak.

![Figure 5.18: Block diagram of the test system for validating the compensator](image)

The variable frequency reference speed, together with the measured motor speed are used to calculate the frequency characteristics of the drive system with and without the compensator in place, using the MATLAB example code shown in Appendix F.

The results in Figure 5.19 and Figure 5.20 show the frequency responses of the emulator drive system with (red and black traces) and without (blue and grey traces) the
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compensator $G_{\text{comp-\omega}}$ to demonstrate the effectiveness of the compensator over the whole system operating range. The simulation results in Figure 5.19 and Figure 5.20 are obtained using the full simulation model of the emulator system described in Section 3.2 with the model-based system identified compensator in (4.17) from Section 4.4. The base speed used in (4.17) is set to 9000rpm to match the base speed identified in Section 5.1.5. The same variable frequency reference speed is used in both the simulation model and test system.

Figure 5.19 shows the compensation performance of the compensator $G_{\text{comp-\omega}}$ for no-load at 3250rpm where the compensator very effectively cancels the uncompensated system resonant peak and so the system bandwidth is doubled from 60rad/s to 120rad/s. The phase difference is also well compensated for frequencies up to 120rad/s. The compensated measured and simulated data correlate well in Figure 5.19. A slight difference is noticeable in the measured and simulated uncompensated results at frequencies over 20rad/s. The small variation in the measured compensated phase response is due to slight measurement noise. The phase values of the frequency of interest from the FFT of the measurement and reference speed are subtracted to get the phase difference. The relatively low signal-to-noise ratio in the measured speed may affect the phase measurement accuracy.

![Figure 5.19: Frequency domain validation results for $G_{\text{comp-\omega}}$ at no-load](image)

Figure 5.20 shows the validation results for $G_{\text{comp-\omega}}$ for the following on-load operating conditions: 9000rpm at 15kW and 27kW, and 13000rpm at 15kW and 27kW. As in Figure 5.19, the simulation results are included for comparison purposes and to verify the effectiveness of the compensator in the simulation environment.
In Figure 5.20, good compensation performance on both magnitude and phase aspects is obtained for all test cases, enabling the system bandwidth to be significantly extended. The extension of system bandwidth reduces slightly with operating speed due to the drive’s inherent speed slew-rate safety feature. This is discussed in more detail in Section 5.3. From Figure 5.20(a)-(c), it can be seen that the compensated phase plots experience a consistent small ripple which is due to measurement noise. The compensated magnitude response has a peak of around 5dB for the 27kW test at 13000rpm in Figure 5.20(d) with the uncompensated case peak being almost 14dB. In Appendix G.4, additional on-load frequency validation tests are shown, further verifying the effectiveness of the compensator \(G_{comp-oa}\) over the system’s full operating envelope.

In order to evaluate the significance of the load effect, compensators \(G_{comp-oa}\) and \(G_{comp-poa}\) have been tested and compared for several on-load operating conditions; 9000rpm at 15kW and 27kW, and 11250rpm at 7.29kW and 13000rpm at 27kW. The compensator comparison results are shown in Figure 5.21, where the red trace is the response of \(G_{comp-oa}\), and the blue trace is the response of \(G_{comp-poa}\). Only measured data with the compensators enabled is shown in Figure 5.21 to enable the responses to be clearly viewed.
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Figure 5.21: Performance comparison between compensator $G_{comp-pw}$ and $G_{comp-pw}$

Figure 5.21(a)-(b) show the load dependency evaluation results for the speed of 9000rpm at 15kW and 27kW respectively. For both case tests, the compensator $G_{comp-pw}$ including both the speed and load dependencies, has a slightly better compensation performance with a slightly lower gain gradient between 80rad/s and 120rad/s. Within the frequency range between 1rad/s and 80rad/s, both compensators have similar performance, allowing both the magnitude and phase angle to be maintained at approximately zero.

Figure 5.21(c)-(d) show the performance of both compensators for two higher speed conditions: 11250rpm at 7.29kW and 13000rpm at 27kW. In Figure 5.21(c)-(d), $G_{comp-pw}$ response performs almost identically to $G_{comp-pw}$, with both compensated system responses having a magnitude and phase angle of close to zero up to around 100rad/s.

The experimental comparison results in the frequency domain are consistent with the time domain validations for the developed system models in Section 5.2.1, showing that the load dependency effect is only noticeable in the high frequency range (above 80rad/s) for the low generating speed condition (around 9000rpm). Incorporating the load effect in the compensator increases the model complexity and only results in a limited improvement on the compensator performance over a small portion of the allowable operating range.
Therefore, for ease of implementation on the RTP, the load influence can be neglected and so $G_{\text{comp-}\omega}$ (5.19) is used in the emulator drive system.

### 5.3 Effect of Speed Slew Rate Limit

Often emulator systems are based on a low power, low speed drive system for example [60] (0.55kW and 1400rpm) where the speed slew rate is quite high and so its effect is negligible within the emulation system bandwidth of interest. The IEPNEF HP spool emulator drive system has a much higher power rating, 115kW, and a maximum speed of 15000rpm and so is often required to operate in the field weakening region. In the field weakening region the motor output peak torque decreases in inverse proportion to the speed which significantly limits the drive slew rate, which is determined considering both speed and power operating conditions. The maximum theoretical speed slew rate can be calculated using (5.25) assuming that the machine is operating at full power (30kW) and full speed (15000rpm). The motor friction loss (friction coefficient $B$) is neglected, the maximum electrical output torque $T_e = 70\text{Nm}$ shown in the machine datasheet [84], estimated system inertia $J=0.11\text{kgm}^2$ from Table 3.3 and the corresponding load torque $T_{\text{Gen}}=19.1\text{Nm}$. From (5.25), the estimated maximum speed slew rate, $d\omega/dt$, is calculated as $462.7\text{rad/s}^2$ (4418.7rpm/s).

\[
T_e = T_{\text{Gen}} + J \frac{d\omega}{dt} + B\omega
\]  

(5.25)

The calculated slew rate represents the analytical estimated value. In the emulator drive hardware, a more conservative slew rate limit of $393.7\text{rad/s}^2$ (3759.4rpm/s) is set for safety reasons. For the compensator to be effective, the compensator output must be within the drive’s slew rate limit, $K_S$, so:

\[
Y(\omega_f) = \frac{K_S}{A \cdot C(\omega_f) \cdot \omega_f} \geq 1
\]  

(5.26)

where $A$ is the variable frequency signal amplitude, $\omega_f$ is the signal frequency and $C(\omega_f)$ is the gain of the compensator which is dependent on the signal frequency $\omega_f$. Therefore, for a specific emulated mechanical dynamics with a bandwidth $\omega_f$ and amplitude $A$, the compensator’s gain $C(\omega_f)$ is limited to $K_S/(A \cdot \omega_f)$. Beyond this gain limit, the effectiveness of the compensator will be impaired. In addition, as the machine speed increases, the
corresponding effective frequency range of the compensator, which is restricted by the compensator gain limit, decreases. In this emulation system, $K_S$ is 3759.4rpm/s as used in the emulator drive, the frequency range of interest has a maximum value of 120rad/s and $A$ is set at 10rpm, so the limit of the compensator output gain $C(\omega_f)$ is calculated to be 3.13, which corresponds to 9.9dB in this chirp test. The effective compensation bandwidth of the compensator due to the slew rate limit can be analytically identified using (5.26), as shown in Figure 5.22, where only $G_{comp,\omega}$ is considered. In Figure 5.22, the condition in (5.26) is met above 0dB and the corresponding frequency limits are identified as 113rad/s, 104rad/s and 89rad/s for 3250rpm, 9000rpm and 13000rpm respectively.

![Figure 5.22: Condition for identification of the compensator bandwidth due to the slew rate limit for different speeds](image)

In Figure 5.22, as the machine speed increases, the compensator’s effective frequency range slightly decreases. For the 3250rpm case test in Figure 5.19, the compensated system response has a magnitude and phase angle of approximately zero up to 120rad/s. Zero magnitude and phase angle can be maintained until the frequency reaches approximately 110rad/s for the 9000rpm case in Figure 5.20(a)-(b) and approximately 100rad/s for the 13000rpm case test in Figure 5.20(c)-(d). These experimental results confirm the validity of the bandwidth limit analysis in Figure 5.22 for the developed compensator.

The current loop dynamics may also impose a limit on the emulated mechanical dynamics, however, as the current loop bandwidth is 1100Hz [81]; the drive speed slew rate limit dominates the overall bandwidth of the emulated mechanical system. The slew rate limit could be increased by using a converter with a higher current rating, up to the thermal and saturation limit of the induction motor, but this may increase the system cost. Increasing the machine rating would increase both torque and inertia, giving at most marginal gains on slew rate and hence compensator performance.
5.4 Conclusion

A parametric system identification technique based method for the industrial drive system dynamics compensator has been presented in detail and applied to an emulation system in IEPNEF. The compensator is designed using experimental step response data, considering the full operating range of the test system. The effects of field weakening and load dependency have been incorporated into the experimental identified system model, yielding two compensator designs, which account for the speed dependency, and both the speed and load dependencies respectively.

The time domain validation for the developed drive system models has been performed by comparing the responses of the transfer function models with the measured system step responses. Both compensators perform well over the full speed and power range, with the system model including both the speed and load effects having a slightly better fit for the on-load test at 9000rpm. The frequency chirp results again demonstrate good performance from both compensators, $G_{\text{comp-o}}$ and $G_{\text{comp-po}}$, over the full speed and load range. The compensator $G_{\text{comp-o}}$ can accurately eliminate the drive system dynamics up to 100rad/s, enabling a good compensation performance for the full system operating range. The load effect is again only significant at around 9000rpm, with $G_{\text{comp-po}}$ having a slightly better compensation performance in the high frequency range between 80rad/s and 120rad/s. The $G_{\text{comp-po}}$ offers slightly better compensation performance of the natural drive system dynamics compared to $G_{\text{comp-o}}$, though this is at the expense of a more complex implementation and the requirement for measured torque in addition to measured speed. In this emulator drive system $G_{\text{comp-o}}$ is preferred as the performance is similar to $G_{\text{comp-po}}$ and the implementation in the RTP is much simpler. The slew rate limit of the drive system has been evaluated analytically and is shown to limit the effective bandwidth of the compensator to approximately 120rad/s, which is double the bandwidth of the uncompensated system.
Chapter 6

Hardware-in-the-Loop Emulation for Aerospace Propulsion Systems

In this chapter different mechanical systems are emulated to further demonstrate the performance of the compensators developed in Sections 4.4 and 5.1. A Rolls-Royce aircraft gas engine model is used to evaluate the emulator system’s ability to accurately replicate throttle (PLA) and electrical power off-take transients. A selection of separate mechanical systems, containing linear, drivetrain or backlash vibration effects, are used to demonstrate the higher frequency emulation capability which may be required in future gas engines designed for more-electric aircraft.

The mechanical system emulation is also examined using the Simulink model described in Section 3.2 which is validated using the hardware-in-the-loop (HIL) emulator system, presented in Section 3.1. The base speed in the emulator drive system model is set to 9000rpm to match the base speed identified in Section 5.1.5.

The contents of this chapter formed the basis of a paper presented at the 2014 IET Power Electronics, Machines and Drives (PEMD) conference [101].

6.1 Emulator System

The emulator hardware system and the controller bandwidth are briefly reviewed in this section as both have a significant influence on the achievable bandwidth of the emulation system.

6.1.1 Configuration

An overview of the emulator system has been shown in Figure 3.1. The emulator consists of a dSPACE real-time platform (RTP) containing a model of the mechanical system of interest, and a 115kW commercial vector controlled induction machine (IM) drive. The difference in emulator drive power and generator power provides the available transient acceleration/deceleration power. The acceleration/deceleration power is effectively used by
the compensator to achieve good speed tracking beyond the natural dynamics of the emulator drive system.

The hardware test system (Section 3.1) and a matching transient simulation model (Section 3.2.3) are used to evaluate the emulator system’s performance with the drive system dynamics compensator. The test system uses the experimental data based compensators from Section 5.1 \( G_{\text{comp-\omega}} (5.19) \) and \( G_{\text{comp-p\omega}} (5.24) \)) and the full simulation model uses the compensator (4.17) from Section 4.4 with the base speed of 9000rpm.

The emulator system enables different mechanical systems to be emulated by implementing a Simulink model of the system of interest on the RTP; no hardware changes are necessary. The RTP has a sample rate of 1ms (1kHz) which is sufficient to enable dynamics up to 120rad/s (19.1Hz) or even higher to be adequately reproduced. It is acknowledged that the emulation bandwidth is limited by the drive system speed slew rate and the software filter characteristics used in the system identification.

### 6.1.2 Controller Bandwidth Regions

The emulator drive in Figure 3.1 uses an induction machine which is regulated by a PI based standard vector controlled algorithm consisting of inner current loops with outer speed/torque and flux regulation loops as described in Section 3.2.2. The emulator system bandwidth is dominated by the 10Hz outer speed control loop bandwidth identified in Section 3.2.2, which is approximately two orders of magnitude smaller than the inner current loop bandwidth, and so the inner current loops can be neglected. Two sample emulator drive hardware frequency responses within the field weakening range \( \geq 9000 \text{rpm} \) are shown in Figure 6.1 for 9000rpm and 13000rpm at 0kW load. Only no-load data is shown in Figure 6.1 as the variation in the frequency response with load was shown to be relatively minor in Figure 5.13. The data is determined using a frequency chirp test with a 10rpm peak-peak variable frequency ripple from Section 5.2.2. Three distinct regions are identified in Figure 6.1 for the 9000rpm test which exhibits significantly different emulator system performance if no compensator is used.

If the emulated system dynamics are relatively low frequency \( \leq 10 \text{rad/s} \) and so within region I of the drive frequency characteristic in Figure 6.1, then the drive emulation performance is good with or without compensator in place as both have a 0dB magnitude and 0° phase characteristics in this range. The upper frequency range of region I does not...
vary with machine speed as confirmed by the frequency response from the 13000rpm test in Figure 6.1.

Figure 6.1: Frequency response division for HP spool emulation system

For emulated mechanical system dynamics within region II in Figure 6.1 (between 10rad/s to 60rad/s for 9000rpm case), with the compensator disabled, the emulated speed dynamics will be amplified by the drive system due to the resonant peak at approximately 40rad/s. When the machine is operating at high load and speed, the current controllers are likely to be approaching their saturation limit and so the amplified response in region II may cause the drive system to hit the torque producing limit, which could result in a large overshoot and long settling time in the speed response. With the compensator enabled the resonant peak in region II will be cancelled out (compensated), enabling good tracking performance for the emulated dynamics. As the system operating speed increases further, for example to 13000rpm as shown in Figure 6.1, the upper frequency limit of region II will be reduced to 50rad/s.

In region III of Figure 6.1 ($f > 60$rad/s for 9000rpm case test, or $f > 50$rad/s for 13000rpm), the emulated dynamics will be significantly attenuated by the drive system when the compensator is disabled. Enabling the compensator will improve the speed tracking performance by compensating the drive system attenuation effect to the input dynamics. However, the compensation performance in region III is limited by the inherent speed slew rate of the drive system as discussed in Section 5.3, resulting in a limited effective compensation bandwidth. In Figure 6.1, region III is expanded as the machine speed increases, as demonstrated by the 13000rpm case test.
6.2 Aero Gas Engine Dynamics Emulation

The aircraft electrical system demonstrator, described in Section 3.1, is currently set up to examine the impact of generators embedded on the gas engine spools. The gas engine model used is a two spool Rolls-Royce defence engine model (presented in Section 3.2.3) which was originally used for performance validation of their engine design. The model has been modified by Rolls-Royce to incorporate electrical power off-take from the high pressure (HP) and low pressure (LP) spools and so interactions between the generator and engine spools can be examined. Only the HP spool is considered in this thesis; the LP spool is out of scope, but does form a potential area of future work.

6.2.1 Experimental Validation

The Rolls-Royce gas engine model was loaded into the RTP shown in Figure 3.1 together with the compensator, which can be enabled in the speed output of the engine model or bypassed to disable the compensator.

The emulation performance is evaluated using both the hardware emulation system in Figure 3.1 and the transient model of the emulation system described in Section 3.2.3 with the parameters listed in Table 3.1 to Table 3.4. The estimated emulation system inertia is 0.11kgm\(^2\) while the emulated gas engine HP spool inertia is 1.87kgm\(^2\). The gas engine being tested has relatively slow dynamics (< 50rad/s) and so both experimental system compensators \(G_{\text{comp-\omega}}\) (5.19) and \(G_{\text{comp-p\omega}}\) (5.24) shown in Section 5.1 offer similar performance within this frequency range (shown in Figure 5.21) and so the speed only dependent compensator \(G_{\text{comp-\omega}}\) in (5.19) is used for simplicity.

Two different gas engine dynamics are examined; a step change in electrical power off-take on the HP spool at constant throttle (pilot lever angle (PLA)) and a step change in pilots lever angle (PLA) at constant electrical power off-take on the HP spool. In [37] the PLA step was identified as a particularly onerous operating point for gas engine emulation systems. In both tests the default environmental data is 15000 feet with a Mach number of 0.5 which represent a normal engine operating condition. The performance of the emulation system is evaluated by comparing both the steady-state and dynamic speed responses. To enable the transient characteristics to be visually compared, the 89ms time delay from Section 3.3 is removed from the test data in Figure 6.2 to Figure 6.5. This is preferred to adding the 89ms time delay to the simulation model as the exact split of the
time delay to the feedforward and feedback path is unknown and cannot easily be measured.

**Electrical power off-take**

Figure 6.2 shows the speed tracking performance of the emulator motor drive during a 15kW electrical load step on the SR generator at a PLA of 18%; the power increases from 0kW at $t=0.5s$ and decreases back to 0kW at $t=4.5s$. A 65rpm decrease in speed from 9995rpm is apparent in Figure 6.2(a) and (b) when the electrical load is applied to the generator. The HP spool speed variation is due to the gas engine model fuel controller directly regulating the LP spool speed (which is not shown in this chapter) with the HP spool speed being determined by the thermodynamics of the engine model and the electrical off-take power. In the test system the 115kW emulator drive can hold the machine speed during a generator load change, preserving the engine dynamics, until the engine derives the reference speed changes in response to the load change. The time delay only affects the specific time of the speed transition in response to the load change and does not affect any feature of the actual transient.

![Figure 6.2: Compensator performance for engine dynamics during an 15kW electrical power off-take transient at 18% PLA](image)

In Figure 6.2, the settling time for the engine to the load step is 3s (less than 10rad/s) which is well within region I in Figure 6.1. Therefore, the emulator speed, both from the test
system and simulation, has an excellent correlation with the reference speed from the engine model with or without the compensator for both power step up and step down scenarios.

**Pilot lever angle (Throttle)**

![Graph](image)

(a) Compensator disabled

(b) Compensator enabled

(c) Experimental system torque responses

Figure 6.3: Compensator performance for engine dynamics during a 58%-60% PLA step change at no-load

In Figure 6.3, a PLA step change between 58° and 60° (the maximum angle) is applied to the engine model when the generator load is zero. The PLA steps have a 0.23s rise time (determined by the gas engine model) and the PLA increase occurs at t=0.13s and decreases at t=4.13s. The uncompensated emulation performance is shown in Figure 6.3(a) and the compensated response is in Figure 6.3(b). The corresponding torque responses for both tests are shown in Figure 6.3(c). The measured and simulated responses show close correspondence in Figure 6.3(a) and (b). When the PLA step change is applied, the engine dynamics have an approximately 30rad/s oscillation during the initial transient. Figure 6.1
shows this oscillation frequency is just below the resonant peak in region II of the uncompensated system. The high gain at the resonant peak frequency means the uncompensated system speed input will be significantly amplified, causing a high overshoot and long settling time in comparison with the engine model reference, as shown in Figure 6.3(a). Enabling the compensator means the system resonant peak is accurately compensated, therefore, the emulator speed response is significantly improved, enabling the excellent dynamic speed tracking shown in Figure 6.3(b). The 89ms time delay is removed from the experimental data to enable the transient to be easily compared as the time delay affects only the specific start time of the transient and no other aspect.

![Speed Response Diagram](image)

**Figure 6.4: Compensator performance for engine dynamics during a 38%-40% PLA step change at 27kW load power**

Figure 6.4 and Figure 6.5 show the emulation performances for engine dynamics when subject to a 38% to 40% and 58% to 59% PLA step changes at full generator load of 27kW. The PLA increases at t=0.25s and decreases at t=4.25s; both steps have a 0.23s rise time.
When the generator is highly loaded, the uncompensated system responses exhibit a more oscillatory under-damped response than that seen in the no-load PLA test (Figure 6.3), especially for the high PLA test (Figure 6.5). The initial transient oscillation frequency in Figure 6.4(a) with the compensator disabled is around 32rad/s which is in region II in Figure 6.1, causing the lightly damped response. When the emulator drive system is operating at high torque with a large speed oscillation, the motor needs to produce torque to supply the high generator load and significant acceleration/deceleration torque for the much amplified speed dynamics.

![Graphs showing speed and torque responses with and without compensator](image)

Figure 6.5: Compensator performance for engine dynamics during a 58%-59% PLA step change at 27kW load power

At the high speed of 14050rpm the torque requirement of the uncompensated emulation drive system, Figure 6.5(c), approaches the maximum torque capability of the machine shown in the machine datasheet [84] which may cause the current controller to saturate. Internal measurements within the drive system were not available so this could not be
verified. However, a similar saturation phenomenon is identified in [39] and attributed to the motor torque-producing limitation. When the compensator is enabled, the drive system dynamics are well compensated, enabling excellent speed reference tracking, Figure 6.6(b), significantly reducing the acceleration/deceleration torque requirement in Figure 6.4(c) and Figure 6.5(c). The validation results in this section, along with Figure 6.2, demonstrate the effectiveness of the proposed compensator in regions I and II in Figure 6.1 and confirm the validity of the frequency analysis in Section 5.2.2.

6.3 Emulation of a Linear System of Variable Inertia

Examining the emulator system’s capability to follow the dynamics of emulated systems accurately with larger and smaller inertias than the hardware system inertia will identify the hardware (emulator motor drive) emulation limit for variable inertia system emulation.

Hardware-in-the-Loop (HIL) emulation of a mechanical source during power transients should present no problem when the emulation system inertia is smaller than that of the emulated power source, as demonstrated in Figure 6.2, providing that the torque rating of the emulator system is sufficient. However, if the prime mover has a high power density and so a much smaller inertia, then the electrical drive system emulation performance may be unsatisfactory as discussed in [39]. The study in [39] presented emulation results from an aeroderivative twin-shaft engine which had an inertia 10 times smaller than the emulation system inertia; some emulation limitations were evident resulting from accuracy and stability issues.

6.3.1 Linear System Model

A generic aero gas engine model accommodating different inertias is complex to develop and so in this section, a heuristic investigation of the emulation performance of gas engines of different inertias is performed through the emulation of a generic mechanical equation of motion, (6.1).

\[ T_e = J_{em} \frac{d\omega_{em}}{dt} + B_{em}\omega_{em} + T_{Gen} \]  

(6.1)

where \( T_e \) and \( T_{Gen} \) represents the electrical input torque and load torque to the mechanical equation, \( J_{em} \) and \( B_{em} \) corresponds to the inertia and damping coefficient for the emulated mechanical system respectively. The emulated mechanical dynamics for this case in the Laplace domain can be derived as:
\[ G_{em}(s) = \frac{\omega_{em}(s)}{T_e - T_{Gen}} = \frac{1}{J_{em} s + B_{em}} \] (6.2)

Although this generic linear model (6.2) does not contain any inherent engine dynamics, it does allow the emulation capability of the test rig to be examined. The model is implemented in the simulation environment (RTP) shown in Figure 3.1, taking the pre-defined electrical torque \( T_e \); and experimental measured load torque \( T_{Gen} \) from the inline transducer, as inputs. The speed output \( \omega_{em} \) from (6.1) is then sent to the emulator drive system as the reference. The electrical torque \( T_e \) and damping coefficient \( B_{em} \) are calculated as 700Nm and 0.6Nmsrad\(^{-1}\) respectively based on the corresponding mechanical rotor equation in the gas engine model used in Section 6.2.

### 6.3.2 Experimental Validation

The emulation capability of (6.2) with different inertias is examined for both steady state and transient conditions using the hardware system in Figure 3.1, described in Section 3.1, and the full simulation model from Section 3.2.3. No emulator drive parameters are changed in these tests, only the inertia value \( J_{em} \) in model \( G_{em} \) varies. As in Section 6.2.1 the time delay in the test system has been removed to enable the transient characteristics to be easily compared.

Two linear systems, both with inertias less than the estimated emulator system actual inertia of 0.11kgm\(^2\), have been tested with and without the compensator in place. The emulated inertias are 0.0187kgm\(^2\) (low \( J_{em} \) case) and 0.011kgm\(^2\) (minimum \( J_{em} \) case); \( B_{em} \) was 0.6Nmsrad\(^{-1}\) in both cases. The emulation system speed was set to 11140rpm using \( T_e \) of 700Nm at \( T_{Gen} = 0 \)Nm and the emulation performance was tested by applying a 15kW step increase and step decrease in load power from 0kW. In this emulation performance study, the compensator \( G_{comp-\omega} \) (5.19), which incorporates only the speed dependency, is employed in the real-time platform (RTP) as the dynamics resulting from the inertia model (6.1) are within regions I and II of Figure 6.1.
Figure 6.6: Emulation performance for the low inertia system (0.0187kgm$^2$)

Figure 6.6 shows the emulation performance for $J_{em}$ of 0.0187kgm$^2$ which is the low inertia case from both simulation and experiment studies and $J_{em}$ is approximately 6 times smaller than the actual inertia of the emulator system. Figure 6.6(a) and (b) show the emulation performance with the compensator disabled and Figure 6.6(c) and (d) shows the corresponding emulation performance for the same condition when the compensator is active. The measured and simulated responses show close correspondence in Figure 6.6(a) and Figure 6.6(c). Poor speed tracking performance is evident in Figure 6.6(a) when the
Chapter 6 Hardware-in-the-Loop Emulation for Aerospace Propulsion Systems

compensator is disabled during both power step up and step down transients. When the actual emulation system inertia is larger than the emulated inertia the emulated dynamics are faster than the drive system and so the motor torque-producing requirement may increase. For the low inertia test (Figure 6.6), the bandwidth of the model speed dynamics is around 32rad/s which is within region II of the uncompensated drive system frequency response in Figure 6.1. The high gain at the resonant frequency in region II and the high torque requirement mean the motor current controller may saturate which contributes to the oscillations in both measured and simulated speed and torque signals in Figure 6.6(a) and Figure 6.6(b). When the compensator is enabled, the resonant peak is eliminated and the emulation bandwidth of the test system is extended which enables excellent dynamic speed tracking, Figure 6.6(c) and Figure 6.6(d).

Figure 6.7(a) shows the compensator performance for a 0.011kgm$^2$ inertia value, which is smaller than that in Figure 6.6. In this case, a small speed oscillation is visible in both simulated and measured responses, which results from the compensator’s output exceeding the drive’s speed slew rate (393.68rads$^2$) over part of the rise process as shown in Figure 6.8(b). When the emulated inertia is smaller than 0.011kgm$^2$, the amplitude of the observed oscillation is larger, which indicates that the minimum inertia that can be emulated with the system is around 0.011kgm$^2$.

Figure 6.7: Emulation performance for the limiting inertia system (0.011kgm$^2$) with the compensator enabled

(a) Speed with compensator enabled

(b) Compensated output speed reference
Figure 6.8 shows the speed tracking performances of the emulator system for the high inertia case, 1.87\(\text{kgm}^2\) which is 16 times the inertia of the actual emulation system with and without the compensator. During the same test condition as in the low inertia case, it takes 9s for the high inertia model speed to reach steady state after the 15kW load is applied or removed. Such slow dynamics are well within region I of the uncompensated system frequency response in Figure 6.1. Therefore, the compensator is not required to improve the emulation performance, as the uncompensated system is already capable of emulating the high inertia dynamics on its own, although importantly the compensator does not degrade the emulation performance. This is consistent with the result in Figure 6.2.

The emulation performance of the test rig for the basic inertia system is highly related to the drive system inherent speed slew rate, limiting the minimum inertia to around 0.011\(\text{kgm}^2\) for a load power of 15kW at the speed of 9000rpm. For load power steps higher than 15kW or system operating speeds above 9000rpm, the test system emulation performance will be reduced for the same inertia case test due to a potential higher speed change rate during the load transient as can be seen from (6.1) and a lower emulation bandwidth as discussed in Section 5.2.2.

![Figure 6.8: Emulation performance for the large inertia system (1.87\(\text{kgm}^2\)](image)
6.4 Aero Engine Drivetrain Emulation

A mechanical drivetrain is very common in transportation applications when mounting the electrical generator or driven load directly on the prime mover is not practical. Other researchers [102, 103, 104] have studied electro-mechanical interactions in traditional aircraft gas engines with externally mounted generators or loads which are coupled to the engine spool through a mechanical drivetrain. Drivetrain resonant frequencies of between 12Hz to 30Hz were identified [103, 104] which if emulated on the test system in Figure 3.1 would be in region III. This section examines the emulator performance to reproduce these higher frequency dynamics.

6.4.1 Drivetrain Model

A simple aero mechanical drivetrain model is used in this section which is developed from a full gas engine model by combining the lumped inertias of auxiliary loads and the torsional stiffness of all driveshafts, leaving a two inertia system, which is dominated by the inertias of the engine spool and the generator, and the stiffness of a common driveshaft [103]. The two-inertia mechanical drivetrain model maintains the main resonant frequencies of the complete system [103] with significantly lower computational demands. Figure 6.9 shows the simplified electro-mechanical system with an externally mounted generator connected to the spool via a driveshaft. In Figure 6.9 the SRSG is assumed to represent the externally mounted generator and so no hardware changes are required.

![Figure 6.9: Overview of aero engine drivetrain generator test system](https://via.placeholder.com/150)

In Figure 6.9, \( J_E \), \( \theta_E \) and \( \omega_E \) are the engine inertia, angular position and angular speed respectively and \( J_G \), \( \theta_G \) and \( \omega_G \) are the generator inertia, angular position and angular speed respectively. The engine inertia, \( J_E \), in the drivetrain model is 1.87kgm\(^2\) to match the engine inertia tested in Section 6.2. The generator inertia, \( J_G \), driveshaft stiffness, \( K \), and the viscous damping, \( D \), are varied to give the required resonant frequency from the classical mechanical resonance equation for a free-free system [105] as described in (6.3). This drivetrain model can be tuned for different natural resonant frequencies which if
excited, may lead to fatigue in mechanical components and instability in the electrical
network [103].

$$f_n = \sqrt{K \left( \frac{1}{J_E} + \frac{1}{J_G} \right)}$$  \hspace{1cm} (6.3)

This section considers a single tunable mode model, shown in Figure 6.10, which can be
determined from Figure 6.9 to demonstrate the emulation capability of the test system.

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**6.4.2 Experimental Validation**

Three different resonant frequencies (12Hz, 18Hz and 27Hz), from the range identified in
[103, 104] are emulated with and without the compensator enabled. Since the resonant
frequencies to be emulated are all within region III in Figure 6.1, both the speed dependent
compensator $G_{comp-\omega}$ (5.19) and the speed and load dependent compensator $G_{comp-\rho\omega}$ (5.24)
from Section 5.1 are tested to compare the resulting emulation performance. Again the
hardware system in Figure 3.1 and the full simulation model from Section 3.2.3 are used to
evaluate the drivetrain emulation performance. In the case of the hardware, the model in
Figure 6.10 is loaded into the RTP. The speed controller in the emulator drive system has a
10Hz bandwidth, as identified in Section 3.2.2. As in Section 6.3.2 the time delay in the
test system has been removed to enable the transients to be easily compared.

**12Hz resonant mode at 9000rpm with 15kW load step up**

In order to excite the drivetrain resonances, an external 15kW power step is applied at
t=0.08s when the emulator drive is running at a steady-state speed of 9000rpm. Figure 6.11
and Figure 6.12 compare the compensation performances of both compensators for the
emulation of a low frequency resonant mode (12Hz). Uncompensated results have been
omitted for the clarity of the other traces in Figure 6.11(a). In this case study, the generator
inertia $J_G$ in Figure 6.10 is 0.05kgm$^2$, the driveshaft damping $D$ is set to 0.2Nmsrad$^{-1}$ and the shaft stiffness $K$ is selected as 263Nmrad$^{-1}$ to achieve the 12Hz resonance.

Figure 6.11: Time domain waveforms for the resonant mode of 12Hz in the mechanical drivetrain

Figure 6.11(a) shows the measured speed transients during the power step up in comparison with the speed reference from the drivetrain model and Figure 6.11(b) is the corresponding measured torque transients for both uncompensated and compensated scenarios. In Figure 6.11(a), it is quite difficult to differentiate the measured motor speed due to a wide range of low frequency noise. However, a periodic (approximately 12Hz) variation at the resonant frequency can be identified in the magnified torque plot in Figure 6.12(b) between 0.2s to 0.6s for both $G_{comp-o}$ and $G_{comp-po}$ cases. The expansion of this torque plot is shown in Appendix H. The corresponding simulation results for the 12Hz resonance emulation in Figure 6.11 and Figure 6.12 are shown in Appendix G.5; the simulation results show a good emulation performance for this resonant mode.

In order to demonstrate the effectiveness of the compensators, a frequency analysis using Fast Fourier transform (FFT) is performed on both the speed and torque transients. Figure 6.13(a) compares the measured motor speed transients with the compensators $G_{comp-o}$ and $G_{comp-po}$ enabled, together with the model reference output in the frequency domain for a 0.7s time window after the electrical load step settles at t=0.2s. In Figure 6.13(a), a very strong resonant frequency at 12Hz is clear in the speed reference, which suggests a correct torsional speed oscillation was triggered by the power step. The
12Hz resonant mode is also present in the measured motor speed FFT shown in Figure 6.13(a) with similar 12Hz amplitudes for both compensators. Such a speed variation is attenuated very quickly by the system damping, as shown in Figure 6.11(a). A second peak at 17Hz is apparent in Figure 6.12(a), which is not present in the reference speed. This frequency is not understood and could be analysed further in the future work.

![FFT of the measured speed transients](image)

Figure 6.12: FFT of 12Hz resonance transients for the 0.7s time window after the electrical power step

Figure 6.12(b) allows the torque transient response of the compensated and uncompensated 12Hz resonant mode model to be compared for both $G_{\text{comp-\omega}}$ and $G_{\text{comp-p\omega}}$. The 0.7s time window between 0.2s and 0.9s is chosen to be the same as that in the speed FFT analysis for a fair comparison. With the compensators enabled, a peak at 12Hz is identified for both tests, which is the expected excited drivetrain resonance. The peak with $G_{\text{comp-\omega}}$ enabled is slightly more prominent than for the $G_{\text{comp-p\omega}}$ test, however, such difference is less obvious in the corresponding time domain responses in Figure 6.11(b). When the compensator is disabled, the emulator is unable to follow the 12Hz resonance which is as expected, since the uncompensated system is tuned for a 10Hz bandwidth. The corresponding FFT analysis based on the full simulation model is shown in Appendix G.5, giving similar compensation performance. Based on the presented results so far the HIL emulation with either $G_{\text{comp-\omega}}$ or $G_{\text{comp-p\omega}}$ enabled respectively is able to introduce the correct frequency of resonance exhibited by a high order system such as a mechanical drivetrain, but with some slight
variation in magnitude. Since the resonance magnitude depends on damping, it is hard to obtain accurately.

**18Hz resonant mode at 9000rpm with 15kW load step up**

Figure 6.13 and Figure 6.14 show the compensator performance for the emulation of a high frequency resonance (18Hz) when excited by the same 15kW load step up at t=0.07s at a steady-state speed of 9000rpm. In this case, the driveshift damping $D$ is 0.2Nmsrad$^{-1}$ and the stiffness $K$ is increased to 634Nmrad$^{-1}$ in order to get the desired 18Hz resonant frequency. As in Figure 6.11(a), the measured speed in Figure 6.13(a) contains a variety of low frequency noise components, so the 18Hz frequency is not obvious in the speed measurements. However, a low frequency component at approximately 5Hz can be identified in the measured speeds (Figure 6.13(a)) and also for the compensated torque transients shown in Figure 6.13(b) where the torque dynamics are expanded in Appendix H. In order to show the emulation performance of the compensators $G_{comp-o}$ and $G_{comp-p+o}$, the frequency spectrum analysis is performed over the same 0.7s time window immediately after the load settles at t=0.2s as the low frequency resonance test for both speed and torque transients and is shown in Figure 6.14. The corresponding simulation studies on this resonant mode emulation can be seen in Appendix G.5.

![Figure 6.13: Time domain waveforms for the resonant mode of 18Hz in the mechanical drivetrain](image-url)
The FFT analysis of the speed transients in Figure 6.14(a) with $G_{\text{comp-}\omega}$ enabled shows two main peaks at 6Hz and 18Hz respectively. The peak at 18Hz is the desired resonant frequency to be emulated which is triggered by the load step. This measured frequency has a peak of 18rpm that is slightly lower than the reference of 26rpm in Figure 6.14(a), which is as expected as the resonant mode is beyond the 15.9Hz cut-off frequency of the compensator $G_{\text{comp-}\omega}$ shown in Figure 5.20. $G_{\text{comp-}\omega}$ at 18Hz has a slightly higher peak magnitude than that of the compensator $G_{\text{comp-p\omega}}$ (12rpm). The other frequency at approximately 6Hz as seen in both time domain and frequency domain results is attributed to the speed slew rate limit effect where the compensator output exceeds the drive’s inherent slew rate. The effect of the emulator drive speed slew rate limit is investigated at the end of this section using the full simulation model.

The torque FFT in Figure 6.14(b) which is applied over the 0.7s time window between 0.2s and 0.9s for the compensated systems also shows peaks at 6Hz and 18Hz for both $G_{\text{comp-}\omega}$ and $G_{\text{comp-p\omega}}$ with $G_{\text{comp-}\omega}$ having a more noticeable peak as in speed FFT which is, however, less defined in the corresponding time domain results shown in Figure 6.13. In comparison to the uncompensated test, both compensators still allow the emulated resonant frequency to occur although the emulated resonance peak is attenuated and an unexpected low frequency component is produced.

![FFT of measured speed transients](image)

(a) FFT of the measured speed transients

![FFT of measured torque transients](image)

(b) FFT of the measured torque transients

Figure 6.14: FFT of 18Hz resonance transients for the 0.7s time window after the electrical power step
**27Hz resonant mode at 9000rpm with 15kW load step down**

Figure 6.15 and Figure 6.16 describe the emulation performance for a much higher frequency resonant mode of 27Hz when excited by a load step down from 15kW at t=0.08s at the same steady-state operating speed of 9000rpm. In this test, the driveshaft damping $D$ is still 0.2Nmsrad$^{-1}$ and the stiffness $K$ is set to 1344Nmrad$^{-1}$, allowing the emulated resonant frequency to occur. In Figure 6.15, the desired resonant frequency of 27Hz is not observed, however, a strong low frequency component at approximately 5Hz is evident in both the speed and torque waveforms for both $G_{\text{comp-}\omega}$ and $G_{\text{comp-p} \omega}$ tests and the magnified view of the torque inset figure is shown in Appendix H. The corresponding simulation studies on this resonant mode emulation can be seen in Appendix G.5 which shows a similar time domain response.

The FFT analysis in Figure 6.16 has the same 0.7s time window between 0.2s and 0.9s as in the previous tests. In Figure 6.16(a), the desired resonant frequency cannot be seen in the speed FFT analysis with only the unexpected 5Hz low frequency noise occurring. Since the emulated resonance (27Hz) is much higher than the compensator effective bandwidth (15.9Hz), the desired frequency component has been significantly attenuated as shown in the time domain speed simulation result in Appendix G.5 which in the test system cannot be differentiated from the measurement noise.

![Time domain waveforms for the resonant mode of 27Hz in the mechanical drivetrain](image-url)
Figure 6.16 FFT of 27Hz resonance transients for the 0.7s time window after the electrical power step

The emulated resonance (27Hz) is apparent in the torque FFT for both compensator tests shown in Figure 6.16(b) which is also significantly attenuated in comparison to other unexpected frequencies at 6Hz and 16Hz. In this case study, the peak at 27Hz in Figure 6.17(b) with compensator $G_{\text{comp-w}}$ enabled is almost the same as that in the compensator $G_{\text{comp-pw}}$ test. Two additional peaks at 6Hz and 16Hz respectively are also apparent in Figure 6.16(b) which is related to the speed slew rate limit effect in the drive system as will be discussed at the end of this section.

**13Hz resonant mode at 13000rpm with 10kW load step down**

Previous resonant frequency emulation results were for a steady-state speed of 9000rpm and a 15kW load step. To demonstrate the compensator performance at higher power and speed, a 13Hz resonance frequency at a steady-state speed of 13000rpm with a 27kW (approximately full SRSG generating power) to 17kW load step is tested to excite the drivetrain model.
The measured torque is shown in Figure 6.17(a) and an FFT of the torque transient from Figure 6.17(a) is shown in Figure 6.17(b) for a 0.7s time window after the load torque just settles at t=0.2s. The load reduction results in a 10Nm step-down load torque excitation to the drivetrain model, Figure 6.17(a). A short duration oscillation is visible between 0.2s and 0.7s in the magnified view of Figure 6.17(a) which is further expanded in Appendix H for both compensator tests which reduces due to the damping term in the model. A frequency mode at 13Hz is present in Figure 6.18(b) after the load step for both $G_{\text{comp-co}}$ and $G_{\text{comp-po}}$ tests, suggesting a correct torsional oscillation has been triggered by the power step down. The peak value at 13Hz is almost unchanged when either compensator is enabled, indicating very similar effectiveness of both compensators at the high speed which is consistent with the analysis in Section 5.2.2.

**Simulation study on the effect of the speed slew rate**

The results in this section demonstrate a wider emulation capability through the successful emulation of drivetrain resonant frequencies at frequencies beyond the emulator drive speed control bandwidth of 10Hz and the compensator effective bandwidth of 15.9Hz. As the emulated resonant frequency increased significantly beyond the control or compensator
bandwidth some unexpected frequency components were identified in the frequency spectrums as shown in Figure 6.14 and Figure 6.16. These additional frequency components are attributed to the pre-set speed slew rate in the emulator drive system of 3759.4rpm/s (393.7rad/s²). The full simulation model from Section 3.2.3 with the emulator drive parameters listed in Section 3.1.2 is used to investigate the effect of the speed slew rate and its association with the unexpected low frequency components when higher resonant frequencies are emulated.

![Time domain response](image1.png)

(a) Time domain response

![FFT analysis](image2.png)

(b) FFT analysis

**Figure 6.18: Example simulation results showing the slew rate effect on resonance emulation**

The simulation model is initially at a steady-state speed of 9000rpm and a 0kW load. A 31.8Hz (200rad/s) 200rpm peak-peak sinusoidal component is added to the 9000rpm speed reference. The resulting simulated emulator drive speed is shown in Figure 6.18; time domain results shown in (a) and FFT results shown in (b). As the reference speed and so the compensator output is well beyond the drive’s slew rate limit, the simulated speed dynamics are significantly attenuated by the drive’s inherent slew rate. In Figure 6.18(a), a low frequency mode, approximately 5Hz, is noticeable in the magnified simulated machine speed plot, along with the emulated frequency mode: 31.8Hz. The FFT analysis of the simulated speed, Figure 6.18(b), confirms that three additional frequency modes at 2.5Hz, 5.5Hz and 13.5Hz are produced due to the slew rate effect, with the most prominent peak
being at 5.5Hz. It has been found that the unexpected low frequency modes are related to the desired frequency of the emulated mechanical dynamics.

The simulation model has been used to demonstrate the range of additional unexpected frequency components which are introduced by the 393.7rad/s² slew rate limit as the reference speed frequency component varies. The emulator steady-state speed is 9000rpm and the load is 0kW and a variable frequency sinusoid from 0 to 200rad/s with a 200rpm peak to peak value is added to the reference speed. The simulated emulator drive speed was processed using an FFT and Table 6.1 summarises the identified unexpected frequency components; in all cases the desired frequency component was also identified correctly but is omitted from the table for clarity.

Table 6.1: Simulation results for the location of unexpected frequencies

<table>
<thead>
<tr>
<th>Frequency range of reference speed sinusoid</th>
<th>Unexpected frequency mode ranges</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>First</td>
</tr>
<tr>
<td>0-13Hz</td>
<td>N/A</td>
</tr>
<tr>
<td>13Hz-20Hz</td>
<td>1Hz-3Hz</td>
</tr>
<tr>
<td>20Hz-32Hz</td>
<td>1Hz-3Hz, 5Hz-10Hz</td>
</tr>
</tbody>
</table>

As illustrated in Table 6.1, when the input frequency is below 13Hz (approximately 80rad/s), only the frequency of the emulated dynamics are identified in the FFT analysis of the simulation speed. As the frequency increases beyond 13Hz (up to around 20Hz), the slew rate effect begins to emerge, presenting two unexpected frequency modes and as the frequency increases further from 20Hz up to 32Hz a third unexpected frequency mode is visible. In all cases the unexpected frequency mode between 5Hz to 10Hz has the highest peak in comparison to the other unexpected frequency modes, though the reason for this is not understood and could be analysed further in the future work.

The simulation study data in Table 6.1 is supported by the test data in Figure 6.14 where one unexpected frequency mode was apparent at around 6Hz, however no unexpected first low frequency mode as suggested by Table 6.1 was evident which is possibly due to the short measurement data window and its associated low frequency noise potentially affecting the FFT accuracy. The emulator drive speed for 27Hz resonant frequency emulation test should have three unexpected frequency components from Table 6.1 whereas the test data in Figure 6.16 only shows unexpected frequencies at around 6Hz and 16Hz; again the unexpected first low frequency component is missing.
6.5 Aero Engine Drivetrain Including Backlash Emulation

A mismatch in prime mover speed and generator or load speed often necessitates the use of a gearbox as part of the drivetrain system. The clearance between gear pairs at the engagement operating point is referred to as the backlash, which is defined to account for lubrication space and machining tolerances.

For systems with a constant or slowly varying load on the gearbox the backlash has a negligible steady-state effect and so is neglected. The effect of backlash is more apparent in dynamic systems where the load changes frequently. The load disturbances introduced by the backlash can induce vibrations between the engaged gears which may result in speed fluctuations [106].

In conventional aero engine drivetrains, backlash can be a significant effect, due to the multiple stages in the transmission system mapping spool speed to the generator speed. A modified version of Figure 6.9 is shown in Figure 6.19 where backlash is added to a simple two-spool aero gas engine electro-mechanical system. The transient loading on the generator in Figure 6.19 combined with gear backlash may lead to excessive mechanical wear and hence a reduced lifespan [107].

In the hardware system used for the experimental testing, in Figure 3.1, the emulator drive is directly coupled to the SR generator and so the mapping between gas engine spool speed and generator speed is performed in the RTP, as discussed in Section 3.1. The performance of the emulator drive to reproduce the effect of backlash with a 1:1 gear ratio will be examined to demonstrate the emulation capability of the emulator drive.

6.5.1 Backlash Model

The backlash effect can be incorporated into the drivetrain model in Figure 6.10 by adding a backlash sub-model [108] shown in Figure I.1 in Appendix I to the drivetrain model so
both the backlash effect and drivetrain resonance can be modelled. The backlash submodel is developed based on the concept presented in [106] and implemented in Simulink according to the layout in [108]. When the backlash is considered, the torque resulting from the drivetrain system, $T_D$, is given by:

$$T_D = K(\theta_E - \theta_G) + D(\omega_E - \omega_G) - (K\theta_b + D\omega_b)$$  \hspace{1cm} (6.4)

where $\theta_E$ and $\theta_G$ are the engine and generator angular positions respectively, $\omega_E$ and $\omega_G$ are the engine and generator angular speeds respectively. $\theta_b$ is the backlash angle which can vary between $\alpha$ and $-\alpha$, and $\omega_b$ is the backlash speed and is obtained using (6.5) [108].

$$\omega_b = \begin{cases} 
\min(0, \omega_E - \omega_G + \frac{K}{D}(\theta_E - \theta_G - \theta_b)) & \text{for } \theta_b = \alpha \\
\max(0, \omega_E - \omega_G + \frac{K}{D}(\theta_E - \theta_G - \theta_b)) & \text{for } \theta_b = -\alpha \\
\omega_E - \omega_G + \frac{K}{D}(\theta_E - \theta_G - \theta_b) & \text{for } -\alpha < \theta_b < \alpha 
\end{cases}$$  \hspace{1cm} (6.5)

The inclusion of the backlash model with the original drivetrain will enable the emulator drive to reproduce the effect of both subsystems if the full integrated model is excited by generator load steps. The excited emulator speed response will be initially in the backlash mode, during which the vibration frequency increases and the amplitude of the oscillation decays. The speed oscillation will then be purely due to the drivetrain model; and this effect eventually decays to zero, depending on the damping value.

### 6.5.2 Experimental Validation

In this section, the engine inertia, $J_E$, generator inertia, $J_G$, driveshaft stiffness, $K$, and the shaft damping, $D$, are selected based on the scaled civil gas engine in [103] and all emulation tests are focused on the backlash vibration period. $J_E$ and $K$ are set to 1.87kgm$^2$ and 2630Nmrad$^{-1}$ respectively, and $J_G$ is varied to give the desired drivetrain resonance from (6.3). The viscous damping, $D$ is set to 0.3Nmsrad$^{-1}$ for all tests. Due to the high machining tolerances in aero engine drivetrains, most of the backlash angles in the drivetrain system are less than 0.1° (0.0018rad) [109]. To demonstrate the concept, three slightly larger backlash angles, 0.1rad, 0.007rad and 0.004rad, are selected from [103] and are emulated with and without the appropriate compensator enabled and the corresponding emulation results showing measured speed and torque along with the speed FFT are shown in Figure 6.20 to Figure 6.22. The hardware in Figure 3.1 is used with the combined
drivetrain-backlash model in the RTP commanding the emulator. The full simulation model from Section 3.2.3 has also been tested with the backlash and drivetrain model and the corresponding simulation results are shown in Appendix G.6. In Figure 6.20 and Figure 6.22, only $G_{\text{comp-}\omega}$ has been tested as the vibration dynamics resulting from the integrated backlash model are within region II of the uncompensated emulator drive frequency response in Figure 6.1. For the last emulation test, both $G_{\text{comp-}\omega}$ and $G_{\text{comp-p}\omega}$ have been evaluated and the corresponding comparison results are shown in Figure 6.22. As in Section 6.3.2 the time delay in the test system has been removed to enable the transients to be easily compared.

Figure 6.20 and Figure 6.21 show the emulation performance for the backlash angles of 0.1rad and 0.007rad when the steady-state emulator speed is 9000rpm. A 0 to 15kW SR generator load transient is applied to excite the combined drivetrain-backlash model and $J_G$ is set to 0.5kgm$^2$ to give a 13Hz drivetrain resonance. The measured speed transients with and without $G_{\text{comp-}\omega}$ enabled are shown in Figure 6.20(a) and Figure 6.21(a) together with the speed reference from the backlash model. Figure 6.20(b) shows the corresponding measured torque transients and the torque inset plot is expanded in Appendix H while the FFT analysis of the speed dynamics is illustrated in Figure 6.20(c) and Figure 6.21(c).

In Figure 6.20(a), a backlash vibration with a first cycle frequency of 5Hz is triggered by the load step and the frequency of the saw-tooth oscillation increases with time until the drivetrain resonant mode of 13Hz is reached at $t=1.64s$ as shown in Appendix G.6, indicating the system enters into the pure drivetrain mode and the backlash effect has fully decayed. Due to the good compensation performance, the measured compensated speed correlates reasonably well with the reference speed in Figure 6.20(a) between 0.1s to 1.1s, after which it is quite difficult to differentiate the actual machine speed from the low frequency noise. Amplification and phase lag effects are apparent in the uncompensated speed response in Figure 6.20(a). In comparison to the uncompensated result, the compensator allows a better emulation of the backlash effect with respect to both amplitude and phase. The FFT analysis in Figure 6.20(c), which applies to the 1s time window between 0.1s and 1.1s of the speed transients in Figure 6.20(a), shows a much higher peak (38rpm) at 5.8Hz for the uncompensated system in comparison to the reference peak of 13.5rpm and the peak with the compensator $G_{\text{comp-}\omega}$ of 18.5rpm, further demonstrating the effectiveness of the $G_{\text{comp-}\omega}$. The frequency of 5.8Hz is the average frequency of the emulated backlash vibration over the 1s time interval.
Good tracking from the emulator drive system is obtained for the emulation test with the angle of 0.007rad in Figure 6.21. A backlash vibration of approximately 9.1Hz at the first cycle is excited with the 0.007rad backlash angle which is higher than the frequencies excited by the larger backlash angles. The correspondence between the reference and the measured compensated speed is good over the whole time period while the uncompensated speed also correlates properly with the reference speed in Figure 6.21(a). Due to measurement noise, the emulator drive system speed waveform after 0.7s are dominated by noise and so are not shown here. The FFT analysis (Figure 6.21(b)) performed on the time window between 0.1s and 0.7s in Figure 6.21(a), shows the compensated speed has a peak of around 9rpm at the emulated frequency of 10Hz which is almost the same as the reference peak. In contrast, the uncompensated speed response is attenuated by the drive
system with only a small peak of less than 5rpm at 10Hz. The frequency of 10Hz is the average frequency of the emulated backlash vibration over the 0.6s time interval.

![Graph](image)

In Figure 6.22, the emulator drive is operated at a higher steady-state speed of 13000rpm before the backlash model (with 0.004rad angle) is excited by applying a 10kW power step reduction from the high load (27kW) condition. The drivetrain resonant frequency at this operating condition is 16Hz ($J_G$ is set to 0.3kgm$^2$ in this case). The triggered backlash vibration has a frequency of around 11Hz; higher than in Figure 6.20 and Figure 6.21 since the smallest backlash angle of 0.004rad is emulated in this case. The measured speed with the $G_{comp-o}$ and $G_{comp-po}$ enabled is corrupted by some low frequency noise after 0.45s in Figure 6.22(a) and so evaluating the emulation performance of the backlash oscillation using speed responses is not possible. The effectiveness of the developed compensators $G_{comp-o}$ and $G_{comp-po}$ can be evaluated using the frequency content of the speed transients in Figure 6.22(a) and is shown in Figure 6.22(b). The FFT analysis performed on the time window between 0.1s and 0.8s in Figure 6.22(a) shows two main peaks at 6Hz and 12.5Hz respectively, where 12.5Hz is the average frequency of the emulated backlash vibration over the time duration of 0.1s to 0.8s. The 6Hz frequency component is attributed to the effect of the emulator drive’s speed slew rate as discussed at the end of Section 6.4.2. With $G_{comp-po}$ enabled, the peak at 12.5Hz is the same magnitude as that for the $G_{comp-o}$ test,
showing the same effectiveness of both compensators for such small backlash angle emulation.

![Graph](image)

(a) Speed

(b) FFT of the speed transients

Figure 6.22: Emulation results for the backlash angle of 0.004rad (0.23°)

### 6.6 Conclusions

In this chapter, the effectiveness of the proposed natural drive system dynamics compensators from Chapter 4 and Chapter 5 has been demonstrated by simulation and experimentally with the successful emulation of a range of mechanical systems relevant to aerospace applications. The working range of the developed compensators has been divided into three regions, each with separate mechanical system models demonstrating its compensation performance.

The emulated mechanical dynamics within region I (<10rad/s) were demonstrated by the gas engine embedded generator electrical power off-take and the high inertia linear system tests. The compensator has no effect on the emulation performance in either case as the emulator drive speed control bandwidth is sufficient to enable good emulation.

In region II, due to the resonant peak within this range, good emulation is a challenge as the resonant peak amplifies the transient response and may cause the drive speed response
to saturate. The compensators accurately compensate for the resonant peak, enabling very
good emulation performance for the region II tests; spool speed dynamics during the PLA
step and also low inertia (less than the emulator drive system inertia) linear mechanical
systems. In region II $G_{\text{comp}-\omega}$ and $G_{\text{comp-p}}$ are equally effective.

The emulation capability of the compensators in region III is restricted by the drive’s
inherent speed slew rate, which is set conservatively for safety reasons. The developed
compensators still offer improved emulation performance by compensating for the drive’s
attenuation effect over this region. To demonstrate the effectiveness of the $G_{\text{comp}-\omega}$ and
$G_{\text{comp-p}}$ in region III, a wide range of drivetrain resonant modes and backlash angles have
been adequately emulated, with either compensator having an improved response
compared to the uncompensated case in the high bandwidth region. Some unexpected low
frequency modes are produced by the slew rate limit effect. In region III, $G_{\text{comp-p}}$ enables
similar emulation performance as $G_{\text{comp-\omega}}$.

Considering all three regions, then overall compensator $G_{\text{comp-\omega}}$ is preferred as it is easy to
implement on the RTP and only dependent on speed which can be accounted for
analytically using the field weakening algorithm. Compensator $G_{\text{comp-p}}$ is more complex to
implement than $G_{\text{comp-\omega}}$ as it is dependent on both speed and load (with the load scaling
being only approximate, see Section 5.1.5). $G_{\text{comp-p}}$ only offers limited improved
emulation performance over a small speed range in the third region only.
Chapter 7

Conclusions and Future Work

This chapter summarises the work contained in this thesis, highlighting the key contributions to the research area. Potential areas of future work are then outlined where a speculative study on the open-loop time delay compensation is performed which applies to the existing emulation system.

7.1 Summary of the Thesis

This project has focused on the development and examination of enhanced emulation techniques for transportation systems, such as aero gas engine dynamics through the use of a power HIL emulator system. Research areas covered include the emulation of mechanical power sources, time delay compensation methods and natural drive system dynamics compensator development where the theoretical analysis has been validated using a 115kW emulator drive system in the Intelligent Electrical Power Network Evaluation Facility (IEPNEF).

7.1.1 Investigation of the Time delay Compensation Techniques

Test results from the emulation system hardware showed a 89ms time delay in the open-loop path which may affect the emulation performance and lead to incorrect assessment of an aero generator if a high power density gas engine is emulated. Two types of time delay effects were considered, closed-loop path, and open-loop path, depending on where the emulator outer torque/speed control loop is implemented. For closed-loop time delays, the analysis is validated by full transient simulation model results which show the emulator system is unstable for time delays exceeding 24ms. However, the system can be stabilised by a Smith Predictor based on estimated emulator drive parameters and system delay. Root locus analysis has indicated that Smith Predictor parameter estimation errors of 40% for time delay $\hat{T}(s)$ and 30% for system model $\hat{G}(s)$ may compromise the stability of the emulation system.
The gas engine model used in this research has a relatively slow speed variation (between 0.5s to 2.5s) in response to PLA step changes or generator load steps and so the identified 89ms time delay is quite small by comparison. The small significance of the time delay on the gas engine model under test, together with the derivative noise from the inverse transportation delay compensation method implementation, both contributed to the time delay compensation not being experimentally tested.

7.1.2 Advanced Drive System Dynamics Compensator Design

A detailed analysis was performed of the analytical derived compensator method based on the machine parameters and control structure, showing that it is vulnerable to unmodelled system dynamics resulting from parameter estimation errors and so not suitable for the dynamic emulation of high speed and high power applications such as an aero gas engine. This thesis then introduced a novel approach to the design of the natural drive system dynamics compensator based on the parametric system identification technique, which is applied to the measured speed step transients from the test rig. System speed and load dependencies have both been considered, resulting in two different compensators: \( G_{comp-\omega} \), which considers only the speed dependency and \( G_{comp-\omega p} \), which takes both speed and load effects into consideration. The field weakening algorithm has been used to analytically incorporate the system speed dependency into the basic compensator model, whereas the load effect is included in the model using an approximate experimentally derived weighting function.

Extensive time and frequency domain validation tests have been performed for the developed compensators, showing excellent compensation performance for the drive system dynamics in comparison to the analytical identified compensator. The compensators can accurately eliminate the resonant peak of the drive system dynamics, extending the uncompensated system bandwidth from 60rad/s to 120rad/s for the full system operating range. The bandwidth extension from the compensator is limited by the speed slew rate, which is set conservatively to 393.7rad/s\(^2\) (3759.4rpm/s) in the drive system for safety reasons. The load effect is most noticeable around the system base speed of 9000rpm where the compensator \( G_{comp-\omega p} \) offers slightly better performance than \( G_{comp-\omega} \) at higher frequencies between 80rad/s to 120rad/s. Of the two developed compensators, \( G_{comp-\omega} \) is preferred as it is simpler to implement in the RTP, requires only a single
measurement value (speed) and provides similar compensation performance to $G_{\text{comp-pao}}$ over the majority of the power/speed operating range.

### 7.1.3 HIL Emulation for Aerospace Prime Mover Systems

A wide range of mechanical systems with a focus on aerospace applications have been emulated to demonstrate the effectiveness of the proposed system identified compensator based emulation technique. Analysis of the uncompensated drive system frequency response indicates three distinct frequency regions with different compensator requirements for correct emulation: region I, no compensator is required, region II, the compensator eliminates the resonant peak and region III, the compensator cancels the drive system’s attenuation. The electrical power transient and high inertia tests on the gas engine embedded generator demonstrated emulator operation in region I as good emulation performance is obtained with or without the compensators enabled. The gas engine PLA step test and the low inertia tests are in region II (10rad/s to 60rad/s) and $G_{\text{comp-ω}}$ enables an 8dB reduction in resonant peak to 0dB and so good tracking performance. Region III relates to higher order system emulation, mechanical drivetrain resonances and backlash vibration effects, which have higher frequency components and so $G_{\text{comp-ω}}$ and $G_{\text{comp-pω}}$ were tested. The emulation system test results within this region have shown similar performance for both compensators as both can enable good tracking to the desired resonance within their effective bandwidth. When the emulated resonant frequency dynamics are beyond the drive system speed slew rate, some unexpected low frequency components may be introduced, and were examined in detail using the transient simulation model.

### 7.2 Contributions of this Research

This research project advances the current state-of-art mechanical system emulation through the design and validation of a new approach to natural drive system dynamics compensation. The proposed emulation method uses the parametric system identification technique to experimentally capture unmodelled dynamics and offers superior emulation capabilities from the electrical drive system over a wider speed range and up to higher bandwidths.
7.2.1 Time Delay Compensation

While a number of time delay compensation methods focus on controller HIL applications, fewer researchers have investigated the time delay effects and examined the appropriate compensation techniques for power HIL emulation systems. In this thesis the impact of time delays on the power HIL emulation system with respect to the closed-loop scenario has been thoroughly examined through the full simulation model. Furthermore appropriate time delay compensation techniques for use in the emulation system have also been evaluated for time delays in the closed-loop path, which provide useful reference and guidance for other researchers in this field for the compensation of such time delay effects in future power HIL emulation systems.

7.2.2 Enhanced Mechanical Emulation Technique

A model-based controller design and analytical compensator design methods for electrical drive based mechanical system emulation have been investigated, identifying the main limitations for each emulation scheme. An enhanced emulation approach based on deriving a more accurate drive system dynamics compensator is then described, which is particularly suitable for the emulation of the high speed and high power mechanical systems, for example an aero gas engine.

The parametric system identification technique is applied to the simple experimental system speed step transients to estimate a system transfer function which has an accurate representation of the test system over the emulation bandwidth of interest (120rad/s). The compensator dependency on the system speed and load operating conditions has been fully analysed and included in the model. The proposed emulation scheme is generic and not fixed to a specific mechanical system, providing greater flexibility without redesigning the emulation technique.

7.2.3 HIL Mechanical System Emulation

The compensator based emulation technique has been demonstrated to enable enhanced aero engine dynamics emulation over the full operating range. The performance limits of the proposed emulation technique, in terms of the range of inertias that can be emulated and the unexpected low frequency components due to the slew rate effect have also been evaluated, together with the successful emulation of resonant frequencies from mechanical transmission effects such as drivetrain resonances and backlash vibrations, these higher
order effects demonstrate the future capability to examine electro-mechanical interactions within the more-electric aircraft.

7.3 Future Work

The work presented in this thesis has met the project objectives defined in the introduction. In this section a speculative investigation, focusing on the open-loop time delay compensation in the emulation system, is undertaken together with the suggestions given for future research topics, either to further develop the mechanical emulation method, or using this mechanical emulation method.

7.3.1 Speculative Investigation of Open-loop Time Delay Compensation

The emulator system in the aircraft demonstrator facility, described in Section 3.1, has a microcontroller in the emulator drive, which is responsible for protection and closed-loop current and speed control functions, with a separate RTP which is used to determine the reference speed for the emulator, either manually or determined by a mechanical system model, and also has a measurement capture capability. This means $T_1$ and $T_2$ in Figure 3.2 are both in the open-loop path which is external to the control loop and so can be combined into a single delay $T$. Therefore, the open-loop time delay does not affect the system stability as the time delay element is not present in the denominator of the resulting system transfer function.

As stability is not a concern in the open-loop configuration, closed-loop time delay compensation techniques such as the Smith Predictor are not appropriate as they are only effective in stabilising time delayed systems. Open-loop time delay compensation methods are rarely discussed in literature. In this section, the inverse transportation delay method as described in [66] has been designed to eliminate or minimise the time delay effect. This section presents an inverse transportation delay function for the emulation system and evaluates the performance.

Overview of Inverse Transportation Time Delay Compensation Method

The inverse transportation time delay compensation method requires a time delay compensation term, the truncated inverse system transport delay model $e^{z_T}$, to be added in the RTP open-loop path to form a new set point for the emulator drive as shown in
Figure 7.1. The delay compensation term $e^{Ts}$ can be expanded using the Taylor series as shown in (7.1), where $T=89$ms for the aircraft emulation facility.

$$e^{Ts} = 1 + \frac{T}{1!}s + \frac{T^2}{2!}s^2 + \frac{T^3}{3!}s^3 + \cdots$$ (7.1)

When the time delay is small enough (in the magnitude of $\mu$s), the second order and higher terms in (7.1) are close to zero and can be omitted, so $e^{Ts}$ can be approximated by $1+Ts$. When the time delay value increases, the higher order terms cannot be ignored and must be included in the delay compensation term to get satisfactory compensation performance. The higher order Taylor series implementation of the compensator contains derivative terms, which are very sensitive to the high frequency signals.

**Frequency Analysis of the Inverse Delay Function in the Emulation System**

The order of Taylor series required is dependent on the specific system time delay. First and third order representations of the delay compensation term using (7.1) have been evaluated for three different time delays, 10ms, 45ms and 89ms to determine the compensator bandwidth (3dB point on a Bode plot) and Table 7.1 summarises the results. The compensation bandwidth increases as the higher order terms are introduced for all time delays in Table 7.1, with the improvement being more significant for larger time delays. Conversely, for the same delay compensation order, the compensator is effective over a wider frequency range for a smaller time delay.

**Table 7.1: Open-loop time delay compensator bandwidth**

<table>
<thead>
<tr>
<th>Time delay (ms)</th>
<th>First order</th>
<th>Third order</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>11</td>
<td>16</td>
</tr>
<tr>
<td>45</td>
<td>8</td>
<td>14</td>
</tr>
<tr>
<td>89</td>
<td>4</td>
<td>12</td>
</tr>
</tbody>
</table>
Figure 7.2 shows a sample frequency response for $\frac{\omega_r}{\omega_o}$ in Figure 7.1 with equations shown in Appendix E.2 for the first and third order compensation representations and a 89ms time delay. The inverse transportation delay method can eliminate the system time delay up to the bandwidth stated in Table 7.1 where the gain is 3dB and the phase is $0^\circ$, above this bandwidth, the output response of the system deviates from the reference in both amplitude and phase as shown in Figure 7.2.

![System frequency responses for 89ms open-loop path time delay](image)

**Figure 7.2: System frequency responses for 89ms open-loop path time delay**

**Time domain Analysis of the Inverse Delay Function in the HIL Aircraft Emulator**

Figure 7.3 shows the simulated compensation performances of first and third order compensators for a 89ms open-loop time delay from the full simulation model of the emulator system described in Section 3.2.3. The simulation conditions and model parameters are the same as in Section 3.3.1. A 16% to 26% PLA step reference increase at $t=20s$ and decrease at $t=25s$ is examined for 0kW SRSG load. The first order delay compensator is satisfactory in removing the system transportation delay during the step decrease in PLA shown at $t=25s$ in Figure 7.3(a), however, it has a negligible effect during the step up in PLA reference at $t=20s$. When a step increase is applied to PLA, the emulator speed with the first order delay compensator representation has a 89ms delay with respect to the reference speed and a small, approximately 16rad/s, oscillation during the settling time. Figure 7.2 indicates the first order representation (green trace) at 16rad/s will have a gain of 1.58 which matches the gain appearing in Figure 7.3(a) at approximately $t=20.3s$.  

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When the third order time delay compensator is implemented, Figure 7.3(b), the time delay compensator performance is similar to the first order delay compensator response in Figure 7.3(a) during the step decrease in PLA at t=25s. Figure 7.3 shows that neither the first or third order time delay compensators can cope with the step increase in PLA as the effective frequency (around 30rad/s) of the PLA step increase exceeds the bandwidth of either compensator listed in Table 7.1.

Figure 7.4 illustrates the performance of the first and third order compensators during a 0-30kW step up in SRSG load at t=20s and a 30kW load decrease at t=25s with an open-loop path time delay of 89ms. The gas engine responds relatively slowly (approximately 2.5s) to the SRSG load change and so both compensators provide an acceptable response to compensate for the time delay. The effective frequency of the load step response is approximately 0.5Hz which is well within the bandwidth of the compensators listed in Table 7.1.
This inverse transportation delay compensation technique may also be effective in removing the closed-loop time delay in Section 3.3.1 providing the time delay is very small (in the magnitude of µs) [110]. For such a small delay, a much wider bandwidth is enabled for the first order compensator, which is beyond the closed-loop system natural frequency. However, for the emulator system, delays in the order of a few µs will not affect the system stability as illustrated in Figure 3.16. For time delays over the critical value of 24ms, the bandwidth of the compensator is significantly compromised and hence its use is limited.

**Discrete-time Domain Analysis of the Inverse Delay Function**

Figure 7.2 showed the compensator performance using the continuous-time open-loop transfer function of Figure 7.1. However, in practice, the compensator will be implemented in RTP which is a digital system and so the discretisation effects should be considered where the corresponding system transfer functions in discrete-time domain are shown in Appendix E.3. Figure 7.5 shows magnified views of the root locus about the unit circle for
the emulation system in the discrete-time domain for different order time delay compensators. In Figure 7.5(a), both poles and the zero are within the unit circle, which indicates that the open-loop delayed system with no open-loop delay compensator is stable. If the open-loop delay compensator is active then as the compensator order increases some zeros are introduced which eventually fall outside the unit circle as shown in Figure 7.5(d) for the third order compensator. The zeros which exist outside the unit circle make the compensated emulator system behave like a non-minimum phase system. The response of a non-minimum phase system to a step input will first have a negative deviation before changing direction to its positive steady state value which is evident in Figure 7.3(b) after the PLA increase at t=20s.

Figure 7.5 illustrates that a trade-off is required between the open-loop time delay compensator bandwidth and the system response which can suffer high frequency noise.
Chapter 7 Conclusions and Future Work

and non-minimum phase behaviour due to the open-loop delay compensator. If the time delay in a system is comparable to the transients of interest, then the first order compensator should be considered as it maintains the system minimum phase characteristic. Furthermore it is least sensitive to the high frequency noise from the derivative components and offers good compensation performance over a reasonable bandwidth for the load transient response, which is essential when examining gas engine-generator-electrical system interactions.

The gas engine model used in this research has a relatively slow speed variation (between 0.5s to 2.5s) in response to PLA step changes or generator load steps and so the identified 89ms time delay is quite small by comparison. The small significance of the time delay on the gas engine model under test, together with the derivative noise from the Taylor series implementation, both contributed to the time delay compensation not being experimentally tested. This work may be revisited if the emulation of a high power density gas engine is required, as this engine would have faster dynamics [39] and so emulation errors from the open-loop time delay would become more prominent and may result in incorrect assessment of the aero generator [37].

7.3.2 Enhanced Time Delay Compensation Techniques

Techniques to mitigate the closed-loop and open-loop path time delays for power HIL applications have been examined only by simulation study. An experimental demonstration of the identified methods on the test system may be undertaken in the future to show the practical implementation of the methods. Further development and demonstration of advanced open-loop path time delay compensation methods may be required if future systems contain fast dynamics. Higher bandwidth sensors, amplifiers, RTP and communications could also be added to the test rig to reduce the time delay.

7.3.3 Advanced Compensator

It has been demonstrated that the key dominating factor on the proposed compensator working bandwidth is the drive system speed slew rate limit.

To increase the effective bandwidth of the compensator and allow a wider emulation range, the emulator drive PI speed controller may be modified to give a slower roll-off gradient of the uncompensated system closed-loop frequency response, which would result in a smaller compensator output at the same frequency in comparison to that of the existing PI
setting. However, the change to the controller setting may affect the emulation system stability and so, in this case, a closed-loop path configuration of the speed controller which is implemented in the RTP may be preferred.

Another possible way is to either increase the converter current rating up to the thermal and saturation limit of the emulator machine or simply increase the machine rating, which however, would result in an increase to the system cost and inertia and give at most marginal gains on slew rate and hence compensator performance.

In order to extend the compensator bandwidth, potential work may include selecting proper PI controller settings, investigating the system nonlinear effect and using a higher current converter or larger emulator machine. Further work includes the design and development of a more advanced compensator which may incorporate the current loop dynamics and to further extend the emulation bandwidth. Furthermore, since the speed- and load-varying compensator $G_{\text{comp-p}}$ was derived simply based on the experimental measurements, future work also includes the analytical modelling of the load effects in $G_{\text{comp-p}}$ in the case where the gas engine embedded generator load has a significant impact on the emulation system dynamics.

### 7.3.4 Engine-generator-electrical System Interactions

The work presented in this thesis will enable the future examination of aero engine-generator-electrical system interactions in IEPNEF. The emulation work would allow the system to emulate various electro-mechanical system behaviours in the next generation of aircraft, which would necessitate the future validation of the corresponding mitigation techniques, including alternative mechanical configurations, energy management and generator control schemes. Future work also includes the mechanical load emulation using the existing hardware with reversed power flow for the examination and validation of the gas engine starting process and in-flight electrical boost of the gas engine.
References


References


References


References


References


Appendices

A. Model Structure and Order Selection for System Identification

In Figure 2.9, after the identification experiment is designed and the relevant measurements are collected, the model structure needs to be selected properly. The general structure of a linear time-invariant model which can be estimated using parametric methods is shown in Figure A.1 and is described by (A.1) [73], where $y(k)$ is the measured system output, $u(k)$ is the measured system input, $e(k)$ is a sequence of independent and identically distributed random variables with zero mean which is the white noise, and $q^{-1}$ denotes the backward shift operator meaning: $q^{-1}u(k) = u(k-1)$.

$$y(k) = G(q^{-1}; \theta)u(k) + H(q^{-1}; \theta)e(k) \quad (A.1)$$

$G(q^{-1}; \theta)$ and $H(q^{-1}; \theta)$ are the input transfer function and noise transfer function respectively. $\theta$ is the unknown parameter vector of the relevant transfer functions.

![Figure A.1: Block diagram of the general model structure](image)

Based on the different characteristics of $G$ and $H$, the model structure can be divided into three categories, which are equation error models, output error models and Box-Jenkins models respectively [79].

A.1 Equation Error Model

The equation error models [73] have two common structures, ARX and ARMAX, which can be described by (A.2) and (A.3) respectively.

$$A(q^{-1})y(k) = B(q^{-1})u(k) + e(k) \quad (A.2)$$
The ARX structure (A.2) is the simple model, which cannot fully describe the characteristic of the disturbance signal, \( e(k) \) (white noise). ARMAX [73] model structure (A.3) adds flexibility on how the disturbance term is modelled by introducing a moving average term \( C(q^{-1}) \) to the white noise signal. In the equation error models, the white noise is assumed to influence the system dynamics before being exported to the output. ARX and ARMAX models can be identified simply by a linear regression and the signal flow of both structures is illustrated in Figure A.2.

![Equation error model structures](image)

**Figure A.2: Equation error model structures**

### A.2 Output Error Model

In equation error model structures, the transfer functions \( G(q^{-1}; \theta) \) and \( H(q^{-1}; \theta) \) share a common denominator \( A(q^{-1}) \) as shown in Figure A.2 and so it is complex to describe and parameterise these transfer functions separately. However, sometimes it is meaningful and natural to have two transfer functions \( G \) and \( H \) characterised independently if the disturbance noise is only applied at the output. An output error model structure is shown in Figure A.3 and is given by (A.4).

\[
y(k) = \frac{B(q^{-1})}{F(q^{-1})} u(k) + e(k)
\]  

(A.4)

![The output error model structure](image)

**Figure A.3: The output error model structure**
Compared to equation error model structure, output error models have a key advantage, allowing $G(q^{-1}; \theta)$ to be parameterised independent of the noise model for which the white noise characteristics are assumed.

The Box-Jenkins model structure, (A.5), is developed from the output error model and is the most complex in terms of model identification. The signal flow of the Box-Jenkins model structure is illustrated in Figure A.4.

$$y(k) = \frac{B(q^{-1})}{F(q^{-1})} u(k) + \frac{C(q^{-1})}{D(q^{-1})} e(k)$$ \hspace{1cm} (A.5)

Figure A.4: Box-Jenkins model structure

Selecting an appropriate model structure is a very important aspect of the system identification procedure. Equally important is the selection for the order of the model, which denotes the number of parameters in the identified transfer functions. In the emulation system in IEPNEF, some high frequency noise, such as the switching ripple of the drive system, and low frequency background disturbance are introduced into the output speed of the emulator drive system, which cannot be replicated by the identified system model. The high switching frequency noise can be removed by a purpose-built low pass filter shown in Appendix B.1. The background noise, which has a relatively constant energy level over the frequency range of interest, is assumed to have a negligible effect on the emulator system speed dynamics. The output error model structure is proposed for the emulation system identification for the ease of the identification process due to the fact that the white noise is simply taken as the disturbance. Furthermore, from a practical point of view, the model order selection is also simpler without the need to identify the order for the noise model. Only the order of $B(q^{-1})$ and $F(q^{-1})$ in (A.4) needs to be selected. In
addition, \( \frac{B(q^{-1})}{F(q^{-1})} \) polynomials can still be identified properly even if the noise model is not correct.

In terms of the model order selection for the emulation system identification, only the input to output dynamics, which is \( \frac{B(q^{-1})}{F(q^{-1})} \), have to be considered, as the output error model has been selected in this work. Normally, a proper order selection is obtained by testing different orders and comparing the relevant model validation results. However, the physical insight of the system and the intended use of the identified model can provide a lot of useful information for selecting the model order. For the emulation system, the physical process can be described by a closed-loop speed control motor system comprising both the speed controller and the rotor dynamics over a limited frequency range assuming the current loop is neglected. The theoretical model of the emulation system as given by (A.6) is derived by means of mathematical modelling.

\[
TF = \frac{K_p K_T s + K_i K_T}{J s^2 + (B + K_p K_T)s + K_i K_T}
\]

(A.6) has two poles and one zero. In addition, the frequency response shown in Figure 4.3 in Section 4.3 has a resonant peak at around 40 rad/s, which implies the identified model at least has two complex poles. In other words, the order of system denominator needs to be at least two. As for the intended model application, the identified model will be used to develop a compensator, which will be placed in the real-time simulation platform (RTP). The requirement to optimise the computational resources of the RTP where large systems like the mechanical and thermodynamic aero gas engine model are simulated makes a low and simple model order a preference.

Moreover, the system identification algorithm usually needs an initial condition to start the identification. A random starting point cannot assure the best model to be identified due to the presence of local minima [111]. In the so-called black box identification, if the initial condition is not specified, the starting values are forced to a null vector. Therefore it is quite beneficial if an initial condition close to the general solution can be given.

As the identified model order is the same as that of the theoretical system transfer function, the initial parameter vector \( \theta_T \) from the estimated model can be used to help the system
identification to converge to the best solution in a short time. Although the estimated model parameters that are taken as the initial values are not a final solution, they can direct the parametric method to obtain a relative best model.

Preprocessing of measurements is usually a prerequisite for the final model estimation. The pre-filtering of the measured experimental data by means of a low-pass or high-pass filter is always taken to remove undesired disturbance features in the data. The input-output behaviour of a LTI system will not be modified when both the input and output signals are pre-filtered through $L(q)$ as illustrated in (A.7).

$$y(k) = G(q^{-1}; \theta)u(k) + H(q^{-1}; \theta)e(k)$$

$$\Downarrow$$

$$L(q^{-1})y(k) = L(q^{-1})G(q^{-1}; \theta)u(k) + L(q^{-1})H(q^{-1}; \theta)e(k)$$

The system identification for the model structure described in (A.1) is performed by the prediction-error method (PEM), which minimizes the cost function $J(\theta)$: the squared predicted errors as given by (A.8) [74]. The PEM approach normally has two special cases which are least-squares method and the maximum-likelihood method respectively [74].

$$J(\theta) = \frac{1}{N} \sum_{k=1}^{N} \left[ \frac{1}{H(q^{-1}; \theta)} (y(k) - G(q^{-1}; \theta)u(k)) \right]^2$$

where prediction error: $e(k) = \frac{1}{H(q^{-1}; \theta)} (y(k) - G(q^{-1}; \theta)u(k))$

In (A.8), the prediction errors are weighted by the input signal spectrum [74], which implies that the identification performance is better at the frequencies where the input signal has high power. As slew-rate limited step signals are used as the input in this identification study, the identified system model should reproduce the system’s low frequencies dynamics better than the high frequency dynamics. There are quite a few commercial software tools which are available for system identification using PEM. In this work the MATLAB System Identification Toolbox [91] is used for this research.
Appendices

B. Hardware System

B.1 Filter Board Designs

The filter circuit was developed based on the standard RC low pass filter shown in Figure B.1 where the circuit has five components: attenuator, RC filter, isolation amplifier, differential circuit and inverting amplifier.

The attenuator (a step down inverting amplifier) firstly attenuates the input signal to the working range of the isolation amplifier after which the attenuator output is conditioned by the RC low pass filter. Then the filtered signal is fed into an isolation amplifier, producing a differential voltage at the other side that is converted to a single ended signal by a differential circuit. The signal is then recovered to the 98% of the original input signal using a step up inverting amplifier. The isolation amplifier (HCPL-7800) rejects very high common-mode transient slew rates and monitors the input voltage signals in the high noise motor control environment.

Figure B.2(a) shows the FFT analysis of the input signal to the filter board which identifies the dominant switching noise at around 2kHz and 9kHz respectively. The cut-off frequency $f_c$ (B.1) of the RC filter is then selected as 500Hz in order to remove the switching noise from the drive system and Figure B.2(b) shows the FFT analysis of the output signal from the filter board which gives a very good filter performance.

\[
R = 33k\Omega \\
C = 0.01\mu F \\
f_c = \frac{1}{2\pi RC} \approx 482.3Hz
\] (B.1)
Figure B.2: FFT of the input and output signals of the filter board

(a) Input signal
(b) Output signal
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B.2 The Specifications of the dSPACE Real-Time Platform (RTP)

The dSPACE real-time platform (RTP) in the emulation system uses DS1006 as the processor board, and send/receive data to/from the emulator drive system using DS4302 CAN interface and DS2202 HIL I/O boards respectively. The specifications for the dSPACE boards are shown in Table B.1 to Table B.3.

Table B.1: The specifications for the dSPACE processor board

<table>
<thead>
<tr>
<th>Board Name</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>DS1006</td>
<td>2.13 GHz Variant</td>
</tr>
<tr>
<td></td>
<td>Intel (R) Core (TM)2 CPU processor running at 2.13GHz</td>
</tr>
<tr>
<td></td>
<td>1. Design for laboratory use with PX10 and PX20 Expansion Boxes</td>
</tr>
<tr>
<td></td>
<td>2. Building multiprocessor systems with further processor boards:</td>
</tr>
</tbody>
</table>

Table B.2: The specifications for the interface board DS2202

<table>
<thead>
<tr>
<th>Board Name</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>DS2202</td>
<td>Analogue inputs</td>
</tr>
<tr>
<td></td>
<td>1. 16 14-bit differential A/D channels (multiplexed)</td>
</tr>
<tr>
<td></td>
<td>2. No sample-and-hold functionality</td>
</tr>
<tr>
<td></td>
<td>Analogue outputs</td>
</tr>
<tr>
<td></td>
<td>20 12-bit differential D/A channels</td>
</tr>
<tr>
<td></td>
<td>Digital I/O</td>
</tr>
<tr>
<td></td>
<td>1. PWM in (24 shared channels, up to 50ns resolution)</td>
</tr>
<tr>
<td></td>
<td>2. PWM out (9 channels, up to 50ns resolution)</td>
</tr>
<tr>
<td></td>
<td>3. PWM in/out channels can also be used for F2D and D2F</td>
</tr>
<tr>
<td></td>
<td>4. Digital in (14+24 shared channels)</td>
</tr>
<tr>
<td></td>
<td>5. Digital out (16 channels)</td>
</tr>
<tr>
<td></td>
<td>Communication interfaces</td>
</tr>
<tr>
<td></td>
<td>1. Serial interface (RS232, RS422 mode)</td>
</tr>
<tr>
<td></td>
<td>2. CAN bus interfaces</td>
</tr>
</tbody>
</table>
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Table B.3: The specifications for the interface board DS4302

<table>
<thead>
<tr>
<th>Board Name</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>DS4302</td>
<td>1. 4 independent CAN channels</td>
</tr>
<tr>
<td></td>
<td>2. 3 different on-board CAN transceivers per channel</td>
</tr>
<tr>
<td></td>
<td>3. Piggyback module site for up to 4 customer-specific CAN transceivers</td>
</tr>
<tr>
<td></td>
<td>4. Interrupt controller</td>
</tr>
</tbody>
</table>

B.3 Linear Scaling in the RTP

The outputs from the torque meter are the scaled voltage signals which are then scaled down by the board DS2202. The linear scaling in the RTP is shown in Table B.4.

Table B.4: Linear scaling factor in the RTP

<table>
<thead>
<tr>
<th>Signal Type</th>
<th>Analogue output scaling</th>
</tr>
</thead>
<tbody>
<tr>
<td>Speed</td>
<td>60×2000 rpm/V</td>
</tr>
<tr>
<td>Torque</td>
<td>60×15 Nm/V</td>
</tr>
</tbody>
</table>

B.4 Coast-down Test for Emulator Drive System

Two spin-down tests were taken, which are for 7000rpm and 15000rpm starting speeds respectively. For the starting speed of 7000rpm, the test results are shown in Figure B.3.

Figure B.3: Coast-down speed test-upper graph and the characteristic curve-lower graph

During the period of the coast-down test, there is no electromagnetic torque, $T_e$ in the emulator motor system and only the friction torque, $T_f$ acts on the common shaft of the motor-generator system. For constant load torque $T_{load} = 0$, the motor mechanical equation of motion (B.2) can be rewritten in (B.3).

$$T_e = J \frac{d\omega}{dt} + B\omega + T_{load}$$  \hspace{1cm} (B.2)
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\[ T_f = B \omega = - J \frac{d\omega}{dt} \]  \hspace{1cm} (B.3)

A section of the dataset in Figure B.3 which starts from 20.792s to 275.49s was taken and divided into 125 partitions. For each partition, all \( \frac{d\omega}{dt} \) and \( \omega \) were calculated and the corresponding mean values for \( \frac{d\omega}{dt} \) and \( \omega \) were then obtained. With all partitions’ mean values for \( \frac{d\omega}{dt} \) and \( \omega \), the characteristic graph which shows the relationship between \( \frac{d\omega}{dt} \) and \( \omega \) was plotted in Figure B.4-upper graph using MATLAB from Figure B.3.

It can be seen that the relationship between \( \frac{d\omega}{dt} \) and \( \omega \) is approximately linear which confirms the form of the friction torque modelling in (B.2). A best first order polynomial fit to the characteristic curve in Figure B.4-upper graph, has been computed using MATLAB and also shown in Figure B.4-upper graph where the slope of the fitted line is calculated to be approximately 0.015s\(^{-1}\).

Figure B.4: The best first order polynomial fits to the characteristic curves for the coast-down tests

The damping coefficient \( B \) can be calculated from (B.3) to be approximately 0.0016Nmsrad\(^{-1}\). For the coast down test at the start speed of 15000rpm/min, the curve showing the characteristic between \( \frac{d\omega}{dt} \) and \( \omega \), together with the corresponding best first
order polynomial fit is shown in Figure B.4-lower graph where a slope value of approximate 0.018 was obtained for the fitted line. Then the damping coefficient $B$ in the high speed coast-down test can be calculated to be $0.0020\text{Nmsrad}^{-1}$ using (B.3). By taking the average of the above two calculated damping coefficient values, $B$ of $0.0018\text{Nmsrad}^{-1}$ can be obtained as the damping coefficient for the emulation system in IEPNEF.

### B.5 Communication Channels

Further details on the communication channels interfacing between the RTP and the emulator drive hardware system is illustrated in Table B.5.

<table>
<thead>
<tr>
<th>Channel Type</th>
<th>Board Type</th>
<th>Baudrate (kbit/s)</th>
<th>Sample Mode</th>
<th>Transceiver Type</th>
<th>Data Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>CAN Bus</td>
<td>DS4302</td>
<td>1000</td>
<td>3 samples per bit</td>
<td>ISO11898</td>
<td>Double</td>
</tr>
</tbody>
</table>
Appendices

C. Key Equations for the SQIM Drive Model

A mathematical model of the emulator system in IEPNEF is developed based on the dynamic, $d$-and $q$-axis equivalent circuit model of the squirrel cage induction machine (SQIM). Dynamic modelling of the induction machine using $d$-$q$ windings is selected, which enables the machines to be controlled properly, for example using the vector control principles [112, 113].

C.1 The Clark and Park Transform

The Clarke ($abc - \alpha\beta$) and Park transforms ($\alpha\beta - dq$) are commonly used in vector control of high performance drive systems, which are related to many synchronous and asynchronous machines (for example induction machine). For the induction machine, the three electrical phase currents produce a magnetic field (mmf) which creates the interaction within all three phases in both the stator and rotor. Although a well-balanced sinusoidal steady-state condition can be assumed for the machine, the development of a full three phase machine model is still complicated due to the complex phase interactions. Then, in terms of dynamic modelling and control of the machines, a two phase system, which is orthogonal, is required so that the machine torque and flux can be controlled independently.

The Clarke transform is a power invariant transform that convert three phase winding system ($abc$) into the equivalent two phase winding system of a fictitious two phase machine ($\alpha\beta$), depicted in Figure C.1. The three phasors ($F_a, F_b, F_c$), which can be either flux linkage, voltage or current, can be transformed into two phasors ($F_\alpha, F_\beta$) using the Clarke transform. The resulting two phasors are decoupled because of their 90 degrees separation and the zero sequence term ($\gamma$) is neglected due to the isolated star-connection and well-balanced windings in the machine. The complexity of the dynamic machine model is simplified greatly by the Clarke transform and a $d$-$q$ induction model can be developed. The Clarke transform can be derived from Figure C.1 and depicted in (C.1).
The scale factor of $\sqrt{\frac{2}{3}}$ is selected to ensure ‘power invariance’ in the transformation. The inverse Clarke transform can also be used to map $\alpha\beta$ to $abc$ phasors, as shown in equation (C.2).

$$
\begin{bmatrix}
F_a \\
F_b \\
F_c
\end{bmatrix}
= \sqrt{\frac{2}{3}}
\begin{bmatrix}
1 & -\frac{1}{2} & -\frac{1}{2} \\
0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\
1 & 1 & 1
\end{bmatrix}
\begin{bmatrix}
F_α \\
F_β \\
F_γ
\end{bmatrix}
$$

(C.1)

The Park transform, as stated in equation (C.3), is one that provides a reference frame conversion on the Clarke transformed machine model. It transforms variables in the $\alpha\beta$ 1stationary two axis system to another $d-q$ two axis system rotating initially at a reference frame frequency, $\omega_0$, shown in Figure C.1 where $\omega_0 = \frac{d\theta}{dt}$. As with the Clarke transform, the Park transform is also power invariant.
\[
\begin{bmatrix}
F_d \\
F_q \\
F_\gamma
\end{bmatrix}
= \begin{bmatrix}
\cos \theta & -\sin \theta & 0 \\
\sin \theta & \cos \theta & 0 \\
0 & 0 & 1
\end{bmatrix}
\begin{bmatrix}
F_x \\
F_y \\
F_\gamma
\end{bmatrix}
\] (C.3)

The frequency of \( \omega_s \) can be selected arbitrarily, however, three common reference frames are always chosen to facilitate the vector control algorithms for fast dynamic control of ac machines, Table C.1 [113].

<table>
<thead>
<tr>
<th>Reference frame</th>
<th>Reference frame frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stationary</td>
<td>0</td>
</tr>
<tr>
<td>Rotor</td>
<td>( \omega_r )</td>
</tr>
<tr>
<td>Synchronous</td>
<td>( \omega_{syn} )</td>
</tr>
</tbody>
</table>

The inverse Park transform shown in (C.4) is used to convert the \( d-q \) two-axis system to \( \alpha\beta \) two-axis system.

\[
\begin{bmatrix}
F_x \\
F_y \\
F_\gamma
\end{bmatrix}
= \begin{bmatrix}
\cos \theta & \sin \theta & 0 \\
-\sin \theta & \cos \theta & 0 \\
0 & 0 & 1
\end{bmatrix}
\begin{bmatrix}
F_d \\
F_q \\
F_\gamma
\end{bmatrix}
\] (C.4)

C.2 SQIM Modelling

The induction machine is modelled based on [112] and the dynamic \( d-q \) machine circuit diagrams are shown in Figure C.2. In terms of the reference frame selection for the machine modelling, the synchronous speed \( \omega_{syn} \) is selected for the \( d-q \) winding speed. Under a balanced sinusoidal steady state, the selection of synchronous reference frame results in the transformed \( dq \)-windings rotating at the same speed as both rotor and stator field distributions. Therefore all motor variables which are related to the stator and rotor \( dq \) windings become constant (dc). With dc quantities, the rotor flux oriented vector control scheme can be easily developed for the dynamic \( d-q \) machine model.
Based on Figure C.2, a mathematical model of the induction machine can be derived which is shown in (C.5) to (C.14).

\[ V_{sd} = R_s i_{sd} + \frac{d}{dt} \psi_{sd} - \omega_{syn} \psi_{sq} \]  
(C.5)

\[ V_{sq} = R_s i_{sq} + \frac{d}{dt} \psi_{sq} + \omega_{syn} \psi_{sd} \]  
(C.6)

\[ V_{rd} = R_s i_{rd} + \frac{d}{dt} \psi_{rd} - \omega_{slip} \psi_{rq} \]  
(C.7)

\[ V_{rq} = R_s i_{rq} + \frac{d}{dt} \psi_{rq} + \omega_{slip} \psi_{rd} \]  
(C.8)

\[ \psi_{sd} = L_s i_{sd} + L_m i_{rd} \]  
(C.9)

\[ \psi_{sq} = L_s i_{sq} + L_m i_{rq} \]  
(C.10)

\[ \psi_{rd} = L_s i_{rd} + L_m i_{sd} \]  
(C.11)

\[ \psi_{rq} = L_s i_{rq} + L_m i_{sq} \]  
(C.12)

\[ L_s = L_{rs} + L_m \]  
(C.13)

\[ L_r = L_{rtr} + L_m \]  
(C.14)
The torque equation and mechanical equation of motion are described in (C.15) and (C.16).

\[
T_{em} = \frac{p L_m}{2 L_r} (\psi_{rd} i_{sq} - \psi_{rq} i_{sd}) \quad \text{(C.15)}
\]

\[
\frac{d}{dt} \omega_{mech} = \frac{T_{em} - T_L - B\omega_{mech}}{J} \quad \text{(C.16)}
\]

Rearranging equations (C.5) to (C.8) results in (C.17) and (C.18). Combining equations (C.9) to (C.12) can provide the relationship between fluxes and currents shown in (C.19).

\[
\begin{align*}
\begin{bmatrix} V_{sd} \\ V_{sq} \end{bmatrix} &= R_s \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \psi_{sd} \\ \psi_{sq} \end{bmatrix} + \omega_{\text{syn}} \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \begin{bmatrix} \psi_{sd} \\ \psi_{sq} \end{bmatrix} \\
\begin{bmatrix} V_{rd} \\ V_{rq} \end{bmatrix} &= R_r \begin{bmatrix} i_{rd} \\ i_{rq} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \psi_{rd} \\ \psi_{rq} \end{bmatrix} + \omega_{\text{slip}} \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \begin{bmatrix} \psi_{rd} \\ \psi_{rq} \end{bmatrix}
\end{align*}
\]

\[
\begin{bmatrix} \psi_{sd} \\ \psi_{sq} \\ \psi_{rd} \\ \psi_{rq} \end{bmatrix} =
\begin{bmatrix} L_r & 0 & L_m & 0 \\ 0 & L_{sq} & 0 & L_m \\ L_{sd} & 0 & L_r & 0 \\ 0 & L_{sq} & 0 & L_r \end{bmatrix}
\begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \end{bmatrix} \quad \text{(C.19)}
\]

The machine equations (C.17) to (C.19), which are implemented in Simulink, give the dynamic induction machine model. With input voltages applied to the machine model, combining the machine equations with the torque equation (C.15) and the mechanical equation of motion (C.16) can result in the full induction machine \(d-q\) model in Simulink.

In terms of the model implementation in Simulink environment, the machine equations (C.15) to (C.19) are incorporated with the Clark and Park’s transforms to create a three phase machine model in the synchronous reference frame. The overall block diagram for model implementation is shown in Figure C.3. Since it is assumed that the induction machine has a squirrel cage rotor, the rotor voltage input is zero. The five machine equations result in a \(dq\)-windings dynamic machine model. Both stator and rotor parts are modelled in the synchronous reference frame.
Appendices

Figure C.3: Equivalent dq-winding circuit model for the induction machine

C.3. Vector Control of the SQIM

Existing ac machine control strategies such as rotor flux oriented control, stator flux orientation, magnetising flux orientation and rotor physical orientation are well documented in literature [114, 115, 116]. In this study, the rotor flux oriented control is implemented in Simulink for the emulator motor drive system in IEPNEF.

Rotor flux oriented control keeps the rotor flux constant and aligns the $d$ axis in the direction of the rotor flux vector as illustrated in Figure C.4.

Figure C.4: Rotor flux orientation

As shown in Figure C.4, the rotor flux rotates at synchronous speed. The quadrature rotor flux must remain zero so that the direct rotor flux is equal to the total rotor flux. The direct and quadrature rotor flux components and machine torque equation under the field orientation conditions are described in (C.20) to (C.22) respectively.
Appendices

\[ \psi_{sd} = L_r i_{rd} + L_m i_{sd} = |\psi_r| = L_m i_{sd} \]  
(C.20)

\[ \psi_{sq} = L_r i_{rq} + L_m i_{sq} = 0 \]  
(C.21)

\[ T_{em} = \frac{p}{2} \frac{L_m}{L_r} (\psi_{rd} i_{sq} - \psi_{rq} i_{sd}) = \frac{p}{2} L_m \psi_{rd} i_{sq} \]  
(C.22)

In Figure C.4, the rotor flux space angle is determined by (C.23) where the synchronous speed \( \omega_{syn} \) equals to the sum of the rotor electrical speed \( \omega_r \) and the slip speed \( \omega_{slip} \). The slip speed can be obtained using (C.24) where \( \tau_r \) is the rotor time constant. The rotor flux angle estimation is shown in Figure C.5.

\[ \theta = \int \omega_{syn} \; dt = \int (\omega_{slip} + \omega_r) \; dt \]  
(C.23)

\[ \omega_{slip} = \frac{i_{sq}}{\tau_r j_{sd}}, \quad \tau_r = \frac{L_r}{R_r} \]  
(C.24)

![Figure C.5: Block diagram for rotor flux space angle calculation](image)

As with a separately excited DC machine, the SQIM torque and flux control are decoupled. The field orientation simplifies the IM control with \( i_{sd} \) setting for rotor flux magnitude and \( i_{sq} \) for machine torque.

Two separate PI controllers are employed in the current loops to maintain the quadrature and direct stator current, which set the machine torque and rotor flux respectively. An additional single-stage PI controller, which provides the torque producing current (\( i_{sq} \)) setpoint, is used to regulate the machine speed.

The stator voltages, which are applied to the SQIM under vector-controlled conditions, are calculated as described by (C.25) and (C.26), where \( \sigma \) represents the leakage factor of the
SQIM. In order to make the stator currents equal to the references, the compensation part for both $V_{sd}$ and $V_{sq}$ must be introduced.

\[ V_{sd} = R_{sd}i_{sd} + \sigma L_s \frac{d}{dt} i_{sd} - \omega_{syn} \sigma L_{sd} \frac{i_{sq}}{V_{sd-comp}} \]  
(C.25)

\[ V_{sq} = R_{sq}i_{sq} + \sigma L_s \frac{d}{dt} i_{sq} + \omega_{syn} \sigma L_{sd} \frac{i_{sd}}{V_{sq-comp}} \]  
(C.26)

\[ \sigma = 1 - \frac{L_{so}^2}{L_s L_r} \]  
(C.27)

### C.4 Drive Inverter Modelling

In the emulator drive system, an industrial voltage-source inverter (VSI) is used and the VSI modelling should properly be addressed. Various modelling techniques for the VSI have been presented in the literature with the aim to analyse and optimise the performance of the inverters [117, 118, 119]. In this study, the developed induction drive model is mainly used for system level analysis within the full HP spool emulation system. Adopting a full and complex inverter model would significantly increase the simulation run time, giving no difference to the machine mechanical dynamics.

A simplified inverter model is implemented in Simulink, using the average value modelling technique [89]. The AC output voltage of the inverter can be defined in (C.28), assuming bipolar pulse width modulation (PWM) and $n$ in (C.28) represents the phase and can be either $a$, $b$ or $c$ [120].

\[ V_{no} = (2 \cdot d - 1) \frac{V_{dc}}{2} \]  
(C.28)

where $d$ is the ideal duty ratio for each inverter leg and $V_{dc}$ is the dc bus voltage. $d$ can be written as:

\[ d = \frac{1}{2} \left(1 + \frac{V_{ref}}{V_{tri}}\right) \]  
(C.29)

where $V_{ref}$ are the reference voltages and $V_{tri}$ is the maximum triangular voltage of the PWM.
C.5 Simplified Diagrams for Speed/Current Control Loop in the SQIM

Assuming that the current loop dynamics in SQIM are neglected, the speed control loop can be reduced as illustrated in Figure C.6(a) where \( K_P \) and \( K_I \) are the proportional and integral values respectively for the speed controller in the emulator system, \( K_T \) is the torque constant and \( J \) and \( B \) are the system inertia and damping ratio respectively. The transfer function relating \( \omega^* \) to \( \omega_r \) can be derived from Figure C.6(a) and is shown in (C.30).

\[
\frac{\omega^*}{\omega_r} = \frac{K_P K_T s + K_I K_T}{J s^2 + (B + K_P K_T) s + K_I K_T}
\]  

(C.30)

Based on (C.25) and (C.26), the current control loops can be simplified as shown in Figure C.6(b), where \( K_{PI} \) and \( K_{II} \) are the proportional and integral values respectively for the current controller in the emulator system, \( R_s \) and \( L_s \) are the stator resistance and inductance respectively and \( \sigma \) is the leakage factor of the SQIM. The transfer function relating \( i_{sd}^*/i_{sq}^* \) to \( i_{sd}/i_{sq} \) can be derived from Figure C.6(b) and is shown in (C.31).

\[
\frac{i_{sd} / i_{sq}}{i_{sd}^*/i_{sq}^*} = \frac{K_{PI} s + K_{II}}{\sigma L_s s^2 + (R_s + K_P) s + K_I}
\]  

(C.31)

Figure C.6: Simplified block diagrams for speed and current control loops
Appendices

D. Expansion of Experimental Validation Results in Chapter 3

Figure D.1: Expansion of torque and speed subplots in Figure 3.8
Figure D.2: Expansion of SRSG DC voltage and current subplots in Figure 3.8
Figure D.3: Expansion of IM phase current subplots in Figure 3.8
Figure D.4: Expansion of Speed and torque subplots in Figure 3.10
Appendices

Figure D.5: Expansion of SRSG DC voltage and current subplots in Figure 3.10
Figure D.6: Expansion of Engine fuel demand subplots in Figure 3.10
Figure D.7: Expansion of IM phase current in Figure 3.10
Appendices

E. Time Delay Effects in Emulator Hardware

E.1 Pade Approximation of the Time Delay

The first order Pade approximation of a sample delay time of 89ms has been compared to the true delay and the results are shown in Figure E.1 which yields good delay estimation accuracy, for frequencies up to 20rad/s.

![Figure E.1: Comparison between the true delay and its Pade approximation in the time and frequency domains](image)

E.2 Transfer Functions for Continuous-Time Domain Analysis of Inverse Delay Compensation Method

The continuous-time transfer functions for $\frac{\omega_j}{\omega_r}$ in Figure 3.23 with or without the open-loop time delay compensator in place is derived as:

Without inverse delay compensator in place:

$$\frac{\omega_j}{\omega_r} = \frac{K_p K_f s + K_i K_f}{J s^2 + (B + K_p K_f) s + K_i K_f} \frac{2 - Ts}{2 + Ts} \quad (E.1)$$

With first order inverse delay compensator in place:

$$\frac{\omega_j}{\omega_r} = (1 + Ts) \frac{K_p K_f s + K_i K_f}{J s^2 + (B + K_p K_f) s + K_i K_f} \frac{2 - Ts}{2 + Ts} \quad (E.2)$$
With third order inverse delay compensator in place:

\[
\frac{\omega_i(z)}{\omega_f(z)} = \left(\frac{T^3s^3 + 3T^2s^2 + 6Ts + 6}{6} \right) \frac{K_pK_fs + K_fK_T}{Js^3 + (B + K_pK_T)s + K_fK_T} \frac{2 - Ts}{2 + Ts}
\]  

(E.3)

**E.3 Transfer Functions for Discrete-Time Domain Analysis of Inverse Delay Compensation Method**

The discrete-time transfer functions are derived from their continuous domain equivalents (see (E.1) to (E.3)) using a continuous-to-discrete time domain transformation (zero-pole matching method) in MATLAB. The corresponding equations are shown as:

Without inverse delay compensator in place:

\[
\frac{\omega_i(z)}{\omega_f(z)} = \frac{-0.01726z^2 + 0.0339z - 0.01662}{z^3 - 2.96z^2 + 2.921z - 0.961}
\]  

(E.4)

With first order inverse delay compensator in place:

\[
\frac{\omega_i(z)}{\omega_f(z)} = \frac{-69.08z^6 + 410.5z^5 - 1016z^4 + 1341z^3 - 996z^2 + 394.4z - 65.06}{z^5 - 4.96z^4 + 9.84z^3 - 9.762z^2 + 4.843z - 0.961}
\]  

(E.5)

With second order inverse delay compensator in place:

\[
\frac{\omega_i(z)}{\omega_f(z)} = \frac{-1.544z^4 + 6.106z^3 - 9.049z^2 + 5.958z - 1.471}{z^4 - 3.96z^3 + 5.88z^2 - 3.882z + 0.961}
\]  

(E.6)

With third order inverse delay compensator in place:

\[
\frac{\omega_i(z)}{\omega_f(z)} = \frac{-2060z^8 + 16340z^7 - 56680z^6 + 112400z^5 - 139200z^4}{z^6 - 5.96z^5 + 14.8z^4 - 19.6z^3 + 14.6z^2 - 5.804z + 0.961}
\]

\[
+ \frac{110400z^3 - 54690z^2 + 15490z - 1918}{z^6 - 5.96z^5 + 14.8z^4 - 19.6z^3 + 14.6z^2 - 5.804z + 0.961}
\]  

(E.7)
F. MATLAB Code for the Drive System Frequency Dynamics Calculation

% FFT calculation for frequencies of interest in a frequency chirp test

N=8   %Signal index for measured drive speed
Fs=1000   %Sampling rate
L=length(data(21051:26051,N))   %Length of signal

%FFT calculation for measured drive speed
Y1=fft(data(21051:26051,N), L)/ L
Y2=fft(data(31051:36051,N), L)/L
Y3=fft(data(41051:46051,N), L)/L
Y4=fft(data(51051:56051,N), L)/L
Y5=fft(data(61051:66051,N), L)/L
Y6=fft(data(71051:76051,N), L)/L
Y7=fft(data(81051:86051,N), L)/L
Y8=fft(data(91051:96051,N), L)/L
Y9=fft(data(101051:106051,N), L)/L
Y10=fft(data(111051:116051,N), L)/L
Y11=fft(data(121051:126051,N), L)/L
Y12=fft(data(131051:136051,N), L)/L
Y13=fft(data(141051:146051,N), L)/L
Y14=fft(data(151051:156051,N), L)/L
Y15=fft(data(161051:166051,N), L)/L
Y16=fft(data(171051:176051,N), L)/L
Y17=fft(data(181051:186051,N), L)/L
Y18=fft(data(191051:196051,N), L)/L
Y19=fft(data(201051:206051,N), L)/L
Y20=fft(data(211051:216051,N), L)/L

Y1(1)=Y1(1)/2
Y2(1)=Y2(1)/2
Y3(1)=Y3(1)/2
Y4(1)=Y4(1)/2
Y5(1)=Y5(1)/2
Y6(1)=Y6(1)/2
Y7(1)=Y7(1)/2
Y8(1)=Y8(1)/2
Y9(1)=Y9(1)/2
Y10(1)=Y10(1)/2
Y11(1)=Y11(1)/2

N=9   %Signal index for speed reference
Fs=1000   %Sampling rate
L=length(data(21051:26051,N))   %Length of signal

%FFT calculation for frequency chirp speed reference
YR1=fft(data(21051:26051,N), L)/L
YR2=fft(data(31051:36051,N), L)/L
YR3=fft(data(41051:46051,N), L)/L
YR4=fft(data(51051:56051,N), L)/L
YR5=fft(data(61051:66051,N), L)/L
YR6=fft(data(71051:76051,N), L)/L
YR7=fft(data(81051:86051,N), L)/L
YR8=fft(data(91051:96051,N), L)/L
YR9=fft(data(101051:106051,N), L)/L
YR10=fft(data(111051:116051,N), L)/L
YR11=fft(data(121051:126051,N), L)/L
YR12=fft(data(131051:136051,N), L)/L
YR13=fft(data(141051:146051,N), L)/L
YR14=fft(data(151051:156051,N), L)/L
YR15=fft(data(161051:166051,N), L)/L
YR16=fft(data(171051:176051,N), L)/L
YR17=fft(data(178051:183051,N), L)/L
YR18=fft(data(187051:192051,N), L)/L
YR19=fft(data(201051:206051,N), L)/L
YR20=fft(data(209051:214051,N), L)/L

YR1(1)=YR1(1)/2
YR2(1)=YR2(1)/2
YR3(1)=YR3(1)/2
YR4(1)=YR4(1)/2
YR5(1)=YR5(1)/2
YR6(1)=YR6(1)/2
YR7(1)=YR7(1)/2
YR8(1)=YR8(1)/2
YR9(1)=YR9(1)/2
YR10(1)=YR10(1)/2
YR11(1)=YR11(1)/2

h=figure

%Calculation of the ratio of measured speed to reference speed at frequencies of interest
A=[1, Y1(9)/YR1(9), Y2(17)/YR2(17), Y3(25)/YR3(25), Y4(26)/YR4(26),
  Y5(28)/YR5(28), Y6(30)/YR6(30), Y7(31)/YR7(31), Y8(33)/YR8(33),
  Y9(34)/YR9(34), Y10(36)/YR10(36), Y11(38)/YR11(38), Y12(39)/YR12(39),
  Y13(41)/YR13(41), Y14(49)/YR14(49), Y15(57)/YR15(57), Y16(65)/YR16(65),
  Y17(73)/YR17(73), Y18(81)/YR18(81), Y19(89)/YR19(89), Y20(97)/YR20(97)]

B=20*log10(A)
f=[1, 10,20,30,32,34,36,38,40,42,44,46,48,50,60,70,80,90,100,110,120]

%Amplitude frequency response plot
subplot(2, 1, 1)
semilogx(f, B, '-kx', 'lineWidth', 2)

%Calculation of the phase differences at frequencies of interest
I1=unwrap(atan2(imag(Y1(9)), real(Y1(9)))*180/pi-atan2(imag(YR1(9)),
real(YR1(9)))*180/pi)
I2=unwrap(atan2(imag(Y2(17)), real(Y2(17)))*180/pi-atan2(imag(YR2(17)), real(YR2(17)))*180/pi)
I3=unwrap(atan2(imag(Y3(25)), real(Y3(25)))*180/pi-atan2(imag(YR3(25)), real(YR3(25)))*180/pi)
I4=unwrap(atan2(imag(Y4(26)), real(Y4(26)))*180/pi-atan2(imag(YR4(26)), real(YR4(26)))*180/pi)
I5=unwrap(atan2(imag(Y5(28)), real(Y5(28)))*180/pi-atan2(imag(YR5(28)), real(YR5(28)))*180/pi)
I6=unwrap(atan2(imag(Y6(30)), real(Y6(30)))*180/pi-atan2(imag(YR6(30)), real(YR6(30)))*180/pi)
I7=unwrap(atan2(imag(Y7(31)), real(Y7(31)))*180/pi-atan2(imag(YR7(31)), real(YR7(31)))*180/pi)
I8=unwrap(atan2(imag(Y8(33)), real(Y8(33)))*180/pi-atan2(imag(YR8(33)), real(YR8(33)))*180/pi)
I9=unwrap(atan2(imag(Y9(34)), real(Y9(34)))*180/pi-atan2(imag(YR9(34)), real(YR9(34)))*180/pi)
I10=unwrap(atan2(imag(Y10(36)), real(Y10(36)))*180/pi-atan2(imag(YR10(36)), real(YR10(36)))*180/pi)
I11=unwrap(atan2(imag(Y11(38)), real(Y11(38)))*180/pi-atan2(imag(YR11(38)), real(YR11(38)))*180/pi)
I12=unwrap(atan2(imag(Y12(39)), real(Y12(39)))*180/pi-atan2(imag(YR12(39)), real(YR12(39)))*180/pi)
I13=unwrap(atan2(imag(Y13(41)), real(Y13(41)))*180/pi-atan2(imag(YR13(41)), real(YR13(41)))*180/pi)
I14=unwrap(atan2(imag(Y14(49)), real(Y14(49)))*180/pi-atan2(imag(YR14(49)), real(YR14(49)))*180/pi)
I15=unwrap(atan2(imag(Y15(57)), real(Y15(57)))*180/pi-atan2(imag(YR15(57)), real(YR15(57)))*180/pi)
I16=unwrap(atan2(imag(Y16(65)), real(Y16(65)))*180/pi-atan2(imag(YR16(65)), real(YR16(65)))*180/pi)
I17=unwrap(atan2(imag(Y17(73)), real(Y17(73)))*180/pi-atan2(imag(YR17(73)), real(YR17(73)))*180/pi)
I18=unwrap(atan2(imag(Y18(81)), real(Y18(81)))*180/pi-atan2(imag(YR18(81)), real(YR18(81)))*180/pi)
I19 = unwrap(atan2(imag(Y19(89)), real(Y19(89)))*180/pi - atan2(imag(YR19(89)), real(YR19(89)))*180/pi)

I20 = unwrap(atan2(imag(Y20(97)), real(Y20(97)))*180/pi - atan2(imag(YR20(97)), real(YR20(97)))*180/pi)

T = 0.089  % Time delay value

phase = [0, 11+10*T*360/2/pi, 12+20*T*360/2/pi, 13+30*T*360/2/pi, 14+32*T*360/2/pi, 15+34*T*360/2/pi - 360, 16+36*T*360/2/pi - 360, 17+38*T*360/2/pi - 360, 18+40*T*360/2/pi + I8-360, 42*T*360/2/pi + I9-360, 44*T*360/2/pi + I10-360, 46*T*360/2/pi + I11-360, 48*T*360/2/pi + I12-360, 50*T*360/2/pi + I13-360, 60*T*360/2/pi + I14-360, 70*T*360/2/pi + I15-360, 80*T*360/2/pi + I16-360, 90*T*360/2/pi + I17-360, 100*T*360/2/pi + I18-360, 110*T*360/2/pi + I19-360, 120*T*360/2/pi + I20-360]

f = [1, 10, 20, 30, 32, 34, 36, 38, 40, 42, 44, 46, 48, 50, 60, 70, 80, 90, 100, 110, 120]  % Phase frequency response plot

subplot(2, 1, 2)
semilogx(f, phase, '-kx', 'LineWidth', 2)
G. Experiment and Simulation Results and Supporting Analysis

G.1 Frequency Spectrum Analysis

(a) Speed Time domain Data

(b) FFT Analysis of the Drive Speed Measurement

Figure G.1: Key waveforms for the frequency chirp test with the frequency of 10 rad/s
G.2 Drive System Speed Dependency Evaluation at On-load Conditions

Figure G. 2: Experimental frequency response at on-load conditions with variable speed
G.3 Motor Rated Voltage Calculation in Response to Resistance Variation

Figure G.3 shows the equivalent circuit of the induction motor, where the resistances $R_1$ and $R_2'$ may change due to the winding temperature variation.

![Figure G.3: Per phase equivalent circuit of the induction motor](image)

For the cold winding resistances (room temperature, 20°C) shown in Section 3.1.2, the load current $I_2$ at rated load and rated slip can be calculated as:

$$I_2 = \sqrt{3 \cdot \frac{P_{\text{rated}}}{R_2'}} = \sqrt{\frac{T_{\text{rated}} \times \omega_{\text{rated}}}{3 \times \frac{R_2'}{s}}} = 206.5 \text{A} \quad (G.1)$$

Where the motor rated slip is obtained as:

$$s = \frac{\omega - \omega_{\text{rated}}}{\omega_{\text{rated}}} = \frac{6810 - 6745}{6810} = 0.00954 \quad (G.2)$$

The phase input current $I_1$ can be found as:

$$I_1 = \sqrt{I_m^2 + I_2^2} = \sqrt{55^2 + 206.5^2} = 213.7 \text{A} \quad (G.3)$$

To find phase input voltage $V_{rate}$:

$$V_{\text{rate}} = \sqrt{3 \times \left((I_1 \times (R_1 + jX_{1s}))^2 + (I_2 \times \left(\frac{R_2'}{s} + jX_{2s}'\right))^2\right)} = 330 \text{V} \quad (G.4)$$

When the resistances are converted to operating temperature ($R_{\text{hot}} = 1.4R_{\text{cold}}$), the load current $I_2$ becomes:

$$I_2 = \sqrt{3 \times \frac{P_{\text{rated}}}{R_2 \times 1.4}} = \sqrt{\frac{T_{\text{rated}} \times \omega_{\text{rated}}}{3 \times \frac{R_2}{s} \times 1.4}} = 174 \text{A} \quad (G.5)$$
Appendices

The phase input current $I_1$ can be calculated as:

$$I_1 = \sqrt{I_m^2 + I_n^2} = \sqrt{55^2 + 174^2} = 182A$$  \hspace{1cm} (G.6)

In the same manner, the phase voltage $V_{rate}$ can be found as 385V.

G.4 Frequency Domain Experimental Validation Results

![Figure G.4: Frequency domain validation results for the developed compensator $G_{comp-es}$ at 11250rpm with 7.29kW](image)

Figure G.4: Frequency domain validation results for the developed compensator $G_{comp-es}$ at 11250rpm with 7.29kW
Appendices

G.5 Simulation Results for Drivetrain Resonances Emulation

12Hz resonant mode at 9000rpm with 15kW load step up

Figure G.5: Simulated time domain waveforms for the resonant mode of 12Hz in the mechanical drivetrain
Appendices

(a) FFT of the simulated speed transients

(b) FFT of the simulated torque transients

Figure G.6: FFT of simulated 12Hz resonance transients for the 0.7s time window after the electrical power step

18Hz resonant mode at 9000rpm with 15kW load step up

(a) Simulated speed transients in the time domain

(b) Simulated torque transients in the time domain

Figure G.7: Simulated time domain waveforms for the resonant mode of 18Hz in the mechanical drivetrain
Appendices

Figure G.8: FFT of simulated 18Hz resonance transients for the 0.7s time window after the electrical power step

27Hz resonant mode at 9000rpm with 15kW load step down

Figure G.9: Simulated time domain waveforms for the resonant mode of 18Hz in the mechanical drivetrain
Appendices

(a) FFT of the simulated speed transients

(b) FFT of the simulated torque transients

Figure G.10: FFT of simulated 18Hz resonance transients for the 0.7s time window after the electrical power step

13Hz resonant mode at 13000rpm with 10kW load step down

(a) Simulated torque transients in the time domain

(b) FFT of the simulated torque transients

Figure G.11: Simulation results for the resonance of 13Hz at 13000rpm with 10Nm torque step down
Appendices

G.6 Simulation Results for Backlash Emulation

Simulation results for backlash angle of 0.1rad (5.7°)

![Graphs showing simulated speed and torque transients](image)

**Figure G.12:** Simulation results for the backlash angle of 0.1rad (5.7°)
Appendices

Simulation results for backlash angle of 0.007rad (0.4°)

![Graph showing speed transients in the time domain](image)

(a) Simulated speed transients in the time domain

![Graph showing FFT of speed transients](image)

(b) FFT of the simulated speed transients

Figure G.13: Simulation results for the backlash angle of 0.007rad (0.4°)

Simulation results for backlash angle of 0.004rad (0.23°)

![Graph showing speed transients in the time domain](image)

(a) Simulated speed transients in the time domain

![Graph showing FFT of speed transients](image)

(b) FFT of the simulated speed transients

Figure G.14: Simulation results for the backlash angle of 0.004rad (0.23°)
Appendices

H. Expansion of Inset Figures in Chapter 6

(a) Figure 6.11

(b) Figure 6.13

(c) Figure 6.15

(d) Figure 6.17

(e) Figure 6.20

Figure H.1: Magnified view of inset plots in Chapter 6
Appendices

I. Drivetrain and Backlash Simulation Models

(a) Integrated drivetrain-backlash model

(b) Backlash sub-model

Figure I.1: Full drivetrain-backlash simulink model diagram