

CRANFIELD UNIVERSITY

KINGSLEY IBRAHIM

DESIGN FEASIBILITY OF THE ELECTRICAL NETWORK FOR
TURBOELECTRIC AIRCRAFT PROPULSION

SCHOOL OF AEROSPACE, TRANSPORT AND
MANUFACTURING

Aerospace Propulsion

PhD

Academic Year: 2016 - 2019

Supervisor: Dr. Suresh Sampath
Associate Supervisor: Dr. Devaiah Nalianda
January 2020

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This thesis is submitted in partial fulfilment of the requirements for
the degree of Doctor of Philosophy

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ABSTRACT

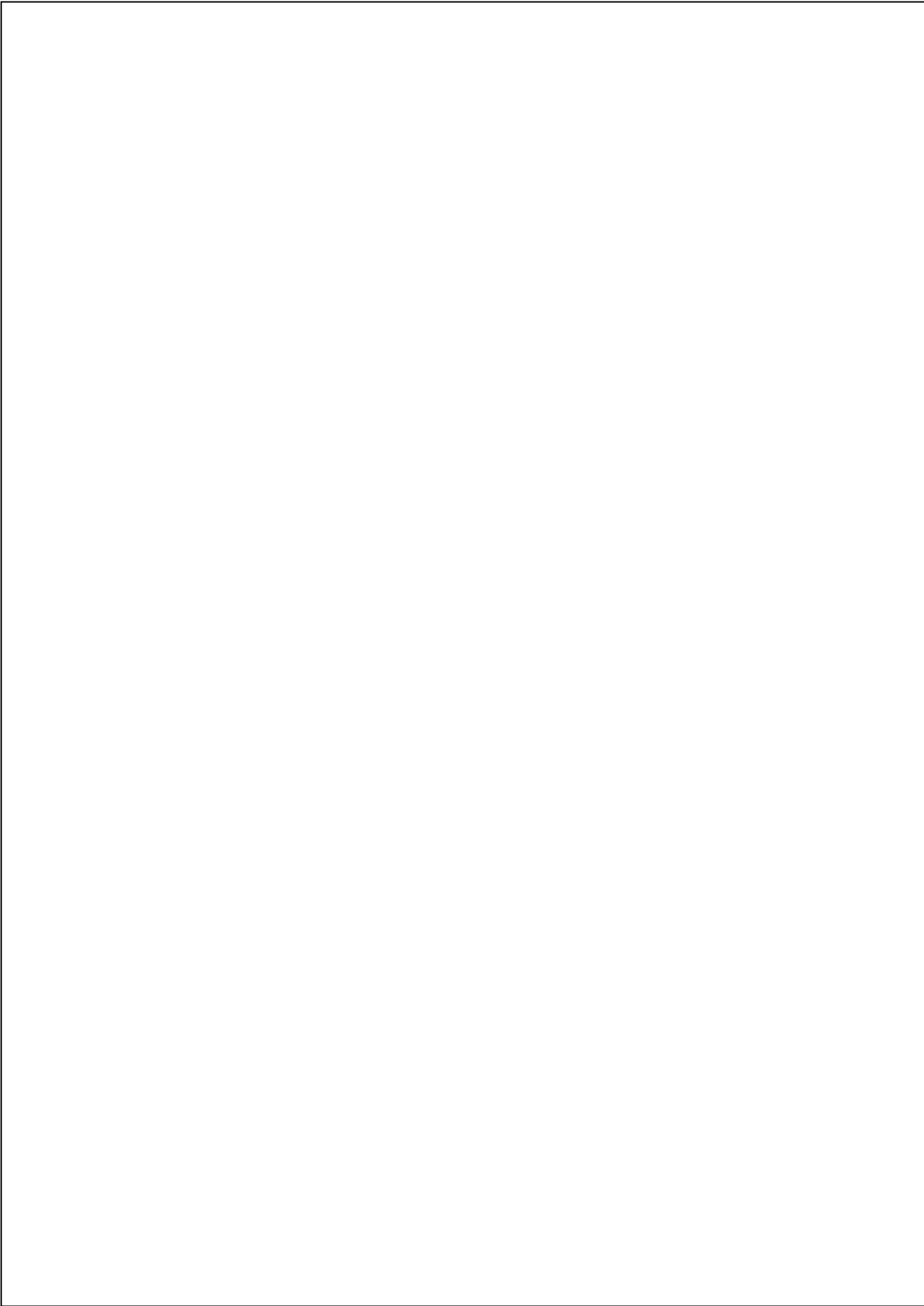
The motivation for this research is the need for safer and more environmentally friendly air transport system. Electrical propulsion systems have been identified as a potential method for improving aircraft performance going forward. The implementation of electrical drive trains for future aircraft propulsion comes with many challenges, due to the novelty and scale of the intended deployments. Major technological advancements and research are ongoing at system and component level to meet this ambition. However, the feasibility aspects of these studies have focused more on the engine side than on the electrical aspects, especially with regards to system reliability and stability. These have been considered in the earlier proposed sizing methods, using assumed fault and transient current magnitudes. Such assumption implies that the control and protection systems, may not properly handle abnormal operational scenarios.

The aim of this research is to establish a procedure for sizing components of an electric propulsion system considering reliability and stability. The major objective is to properly quantify the operating parameters in non-steady state operations, like transients and fault scenarios, and establish that components are operating within their thermal limits at all operational stages.

The contribution of this work is the development of a method that incorporates stability and reliability in the sizing process of electrical propulsion networks. The practicality of the proposed methods has also been validated experimentally, using a test facility set up for this study. The impact of this work is the reduction of the design uncertainties, resulting from assumed fault and transient characteristics of an electrical propulsion system. The results show that the assumptions in earlier researches do not suffice for the investigated architectures. A considerable mass penalty is incurred, with the power electronic devices having to be sized for slightly higher than the maximum transient currents.

Keywords:

Fault, Temperature, Voltage, Motor, Generator.



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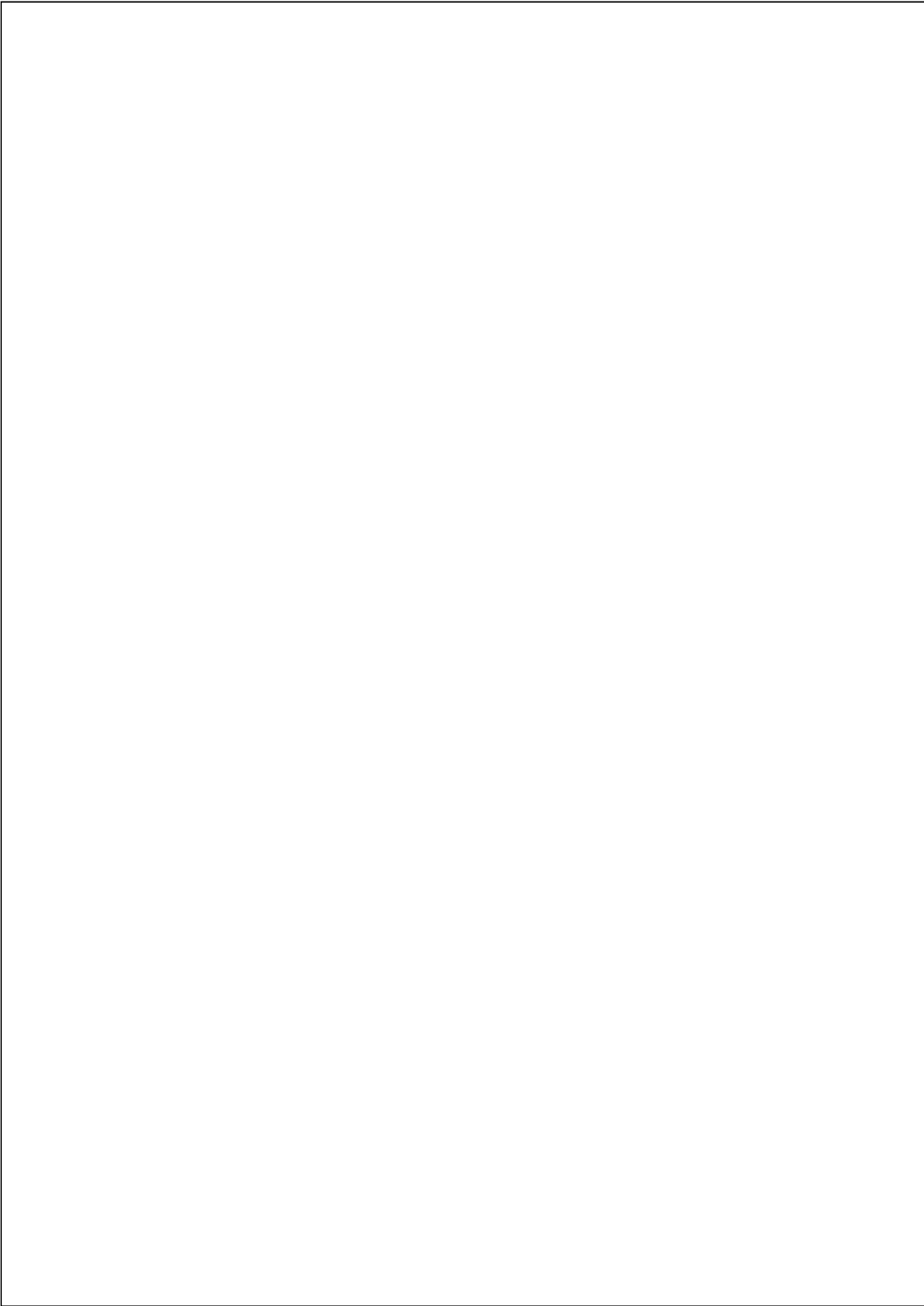


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LIST OF ABBREVIATIONS AND SYMBOLS

AC	Alternating Current
ADP	Aerodynamic Design Point
AEA	All Electric Aircraft
BLI	boundary layer ingestion
BSCCO	Bismuth Strontium Calcium Copper Oxide
BWB	blended wing-body airframe
CESTOL	Cruise Efficient Short Take-off and Landing
DC	Direct current
DEAP	Distributed Electrical Aerospace Propulsion
ESAero	Empirical Systems Aerospace
FAA	Federal Aviation Administration
FPR	Fan Pressure Ratio
HWB	Hybrid wing body
IGBT	Insulated-Gate Bipolar Transistor
MEA	Electric Aircraft
MgB ₂	Magnesium Diboride
NASA	National Aeronautics Space Administration
PAI	propulsion airframe integration
PMSM	Permanent Magnet Synchronous Machine
PWM	Pulse width modulation
SCEPTOR	Scalable Convergent Electric Propulsion Technology and Operations Research
SFC	Specific Fuel Consumption
SiC MOSFET	Silicon Carbide Metal–Oxide–Semiconductor Field-Effect Transistor
SMES	Superconducting Magnetic Energy Storage
SSCB	Solid State Circuit Breaker
TeDP	Turbo-electric distributed propulsion
TRL	Technology Readiness Level
TSEC	Thrust specific energy consumption
YBCO	Yttrium Barium Copper Oxide
ϕ	Flux
λ	Flux Linkage
ρ	Density
τ	Time Constant
$\emptyset$	Phase Angle
α	Voltage Angle



1 INTRODUCTION

1.1 Introduction

Increasing demand for commercial air travel is expected to have additional environmental impact[1], as more flights will be required to meet this rising demand. Along with improvement in other sectors of aviation, far reaching improvement of deployed aircrafts has become inevitable. Precisely, a need to reduce current propulsion system's contribution to emissions.

Significant efforts have been put in, to develop a framework for the reduction of aviation related emissions and other environmental impacts like noise. The NASA N+3 goals [2,3] set in 2008 and the European Flightpath 2050[4] set in 2011, are examples. These frameworks highlight a holistic approach to solving these problems, by viewing aviation activities as a single integrated system. This implied that a worthwhile improvement, can only be achieved on a single aviation component with due consideration of the other components. A broad division into aircrafts, airports, logistics and passengers is made[5], and these require individual and synergistic optimization.

On the aircrafts side of things, much of the possible optimization have been shown to rely on disruptive technologies[6]. Several concepts have been proposed, to improve the performance of aircrafts, in terms of efficiency (fuel burn per mission) and noise pollution[3,7]. The key improvement dependent parameters are; the fan pressure ratio (FPR) and airframe configuration to increase propulsive efficiency, the turbine operational speed to improve the thermal efficiency, and the propulsor exit velocity to minimize noise[8].

Optimum thermal and propulsive efficiencies can be achieved, if the power generation and propulsion components of turbofans are decoupled[8]. Limits of fan diameter and tip speed are a constraint to improving propulsion and thermal efficiencies. This is because, the present configuration in most long-range passenger aircrafts, couples the turbine and fan in a single engine.

Two major approach for uncoupling the fan and turbine are either mechanical or electrical transmission. The mechanical approach would require a gear box

between the turbine spool and the fan shaft. The electrical approach on the other hand, involves the deployment of electromechanical power conversion devices (generators and motors), along with cables, to transmit the power from the turbine to the fan shaft. Such an arrangement offers not only the benefits of running individual propulsors at optimized speed as with gear boxes, but also a flexibility in positioning them around the aircraft for optimized thrust production and less noise. Other reasons for exploring the electric propulsion option, is the superior efficiency anticipated. While traditional propulsion systems offer lower thermal efficiency of about 60%, the electrical propulsion systems are proposed to go as high as 90%[9]. Hence, electrical incorporated systems would result in lower emissions through reduced fuel burn and reduced running costs for aircraft operators.

Despite these promises, the implementation of such a propulsion system still poses challenges, due to the novelty required in implementing such magnitude of power for aircraft propulsion. Unlike with land-based deployments, aerospace deployments do not have the luxury of space, less concern for incurred mass, ease of power management choices and servicing. The implementation of such an electric network for aircraft propulsion follows two major approaches; The more electric aircraft (MEA) and all electric aircraft (AEA). These propose to partly or totally replace the mechanical propulsion system with electrical systems. There are several possible configurations, with overlapping attributes of these two major approaches. All these are with the view to achieve aircraft propulsion at the safest, most economic and power dense cost. It is expected that the more feasible configurations are those compatible with the existing system and are unconstrained by the challenge of energy storage. An example is the turboelectric distributed propulsion (TeDP), which is proposed to be integrated on a number of airframes[10,11].

The power requirement for such an electrical propulsion system reaches an unconventional magnitude, compared with current electrical deployment on aircrafts. Existing electrical networks in the area of aviation have been deployed to meet needs other than the active propulsion of the aircraft. This involved less

power, which is easier to generate, transmit, distribute and manage. With the imminent change, several aspects of electrical network systems have been researched, for both conventional and superconducting electrical network systems. These were in a bid to set the stage for the required change, by addressing the possible challenges.

This work presents a design feasibility of the propulsion system for turboelectric propulsion systems. Focus is given to impact of systems stability and reliability consideration on the sizing and viability of such a propulsion system.

1.2 Aims and Objectives

The aims and objectives of this research are drawn from the research questions intended to be answered, which are;

- What are the electrical stability and reliability challenges associated with electrical propulsion systems?
- What is the incurred mass penalty for providing the contingency?
- How does the magnitude of these challenges impact the design space?

1.2.1 Aims

The aim of this research is to investigate the impact of electrical transients, faults and thermal characteristics, on the overall mass and reliability of electrical propulsion systems.

1.2.2 Objectives

The Objectives through which the aims would be achieved are as follows;

1. Establish the electrical power capacity of the propulsion system.

This is the power required to meet the design thrust at the set altitude and Mach number. The fan rotational speed is also derived, since power for thrust must be delivered as a product of speed and torque.

2. Selection of optimal voltage and current for the electrical network.

Electrical networks usually operate at predefined voltage levels. This requirement is often determined by component size, nature of utility demands and existing regulations. These form the design boundaries for the selection

of an optimal combination of voltage and current for an electrical propulsion network.

3. *Devices identification and sizing.*

The functional devices of such an electric network for aircraft propulsion needs to be identified and sized. The sizing approach for these components are necessarily done along with the voltage and current selection, to arrive at an optimal point.

4. *Transients and Fault Estimation.*

A study of the network branches to derive the susceptibility and reliability of each is required. The magnitude of the electrical network instability and failure is obtained for various points on the network. This influences the position and sizing of protection devices for the network.

5. *Thermal Characterization.*

The instantaneous heat dissipation of the components and sub-components, in reaction to the passage of electrical current needs to be quantified. This would aid the proper sizing of the cooling systems, to keep them running within their temperature limits both at normal and abnormal conditions.

1.3 Thesis Structure

The chapters and summary of the aspects covered in each chapter are outlined below;

Chapter 2 reviews the previous researches carried out for future electric aircraft propulsion systems. A few candidate configurations are covered, these reveal the knowledge gap intended to be filled.

Chapter 3 covers the transient, fault and thermal characterization of a notional electrical network. A novel cooling strategy is also presented, in pursuit of optimal system sizing.

Chapter 4 presents two approaches for optimal voltage and current selection for electric propulsion networks. The presented approaches have an ordered target of feasibility, optimal mass and efficiency.

Chapter 5 presents a validation of the presented design principles, using experimental data and existing literature. This would give credence to the simulated values and aid reliance for scaled up application of the methods presented in this work.

Chapter 6 is a case study application of the derived principles. Here, the effects of considering transient, fault currents and thermal behaviour in the design process is made evident. This is done by investigating a few electrical configurations.

Chapter 7 outlines the conclusions that are derived from the research conducted and presents an outlook for future investigation.

The expected contribution of this thesis is to present clearly defined procedure, to size components of a turboelectric propulsion system. This is done with deliberate focus on their ability to withstand abnormal and transient scenarios, while operating within their thermal limits.

Figure 1 shows the procedure of the conducted investigation.

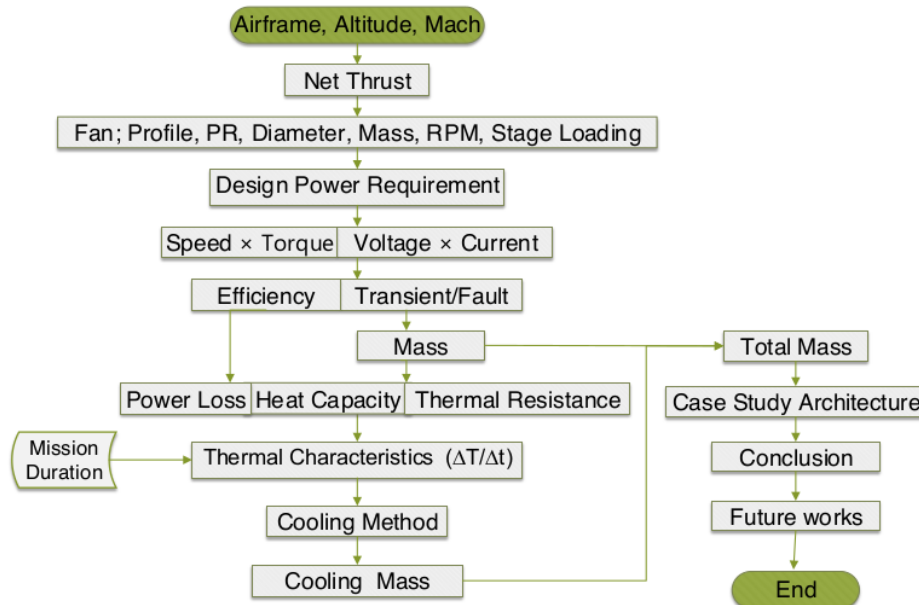


Figure 1 – Flow Chart of Performed Research

2 LITERATURE REVIEW

Electric propulsion refers to a method of aircraft propulsion, where an electric motor or an electrically powered device, is utilized in the generation of thrust. Several possible implementations of electrical propulsion abound, and are greatly dependent on the specific needs, feasibility and availability of the enabling technologies. Figure 2 depicts the broad categories of electrical propulsion configurations.

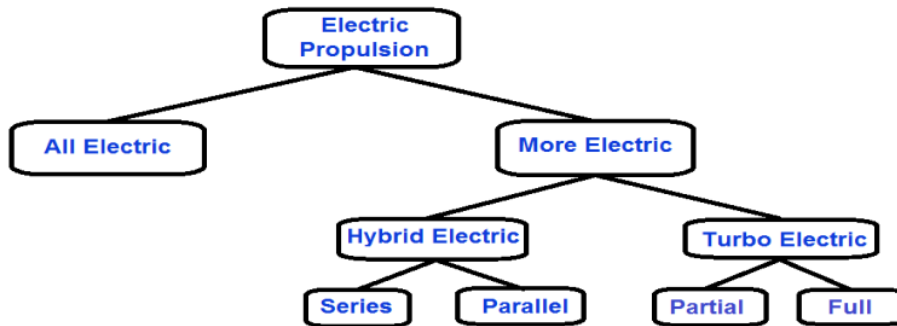


Figure 2 – Overview of Electrical Propulsion Configurations

All electric propulsion requires that the primary source of energy for the propulsion systems, be entirely derived from non-combustible processes. This is typically from electrical charge banks like batteries. More electric propulsion systems on the other hand, refer to deployments where electrical propulsion systems either serves to augment combustion engines, or delivers the entire propulsion power, using energy converted from combusted fuel. The following sections presents a review of the previous works done on the core aspects of electrical propulsion systems and technologies. This is with the view to identify the knowledge gap for which this thesis aims to fill.

2.1 Main Features of Electrical Propulsion

Electrical propulsion systems are proposed to exhibit significant deviation from the existing deployments. These relate to the overall conceptual through implementation and operation stages. These peculiarities are discussed next.

2.1.1 Distributed Propulsion

Early concepts in the area of distributed propulsion have been reviewed in[12]. These employ jet flaps, crossflow fans, gas driven multi-fans, or multiple discrete engines. The main advantage of these concepts, is the emphasis on propulsion airframe integration (PAI), which helps achieve maximum aerodynamic efficiency, reduction in noise and optimal propulsion energy consumption[10]. Another advantage is the feasibility of deploying multiple lower power rating machines for propulsion, as against the currently deployed large turbine engines. These advantages are friendlier with electrical propulsion as the power source and propulsor can be optimally positioned on the aircraft and interconnected using cables.

2.1.2 Electric Powered Fans

The implementation of electrical propulsion utilizes electric powered fans to produce thrust. These fans are driven by electric motors which source their power from an electrical system. The speed and diameter of the fans is optimally selected without the typical restrictions imposed by the turbine shaft speed limits[8] and ground clearance. This increases the propulsive efficiency and enables thrust to be produced by devices other than a turbine.

2.1.3 Electric Machines

Electric machines are the main enablers of electric propulsion systems. They include generators and motors, which perform the mechanical-electrical-mechanical power conversion. Although these devices have existed for several decades, they do not possess a similar power density as the gas turbines. Therefore, a straightforward substitution would rather defeat the aim of electrical propulsion, due to higher mass penalty. Researches performed so far, have identified the potential to surpass this challenge if the machines are made with superconductors. Superconductivity refers to the quality of some materials to conduct electricity with zero resistance[13], if cooled to a certain temperature. Common superconductors in this category are bismuth strontium calcium copper oxide (BSCCO), magnesium diboride (MgB₂) and yttrium barium copper oxide (YBCO), which have been evaluated by NASA. These have shown viability for

electrical propulsion systems, with expected efficiencies in the region of 99.9%[14].

Superconducting machines are characterised by higher power densities and minimal losses[15,16]. They are maintained at their operational temperatures by the deployment of cooling technologies. For BSCCO and YBCO based machines, cryocoolers using liquid nitrogen as the cryogenic fluid are deployed[17,18]. While in the case of MgB₂, liquid hydrogen is the proposed cryogenic fluid[19]. The ability of these cooling systems to deliver cooling at an aerospace adaptable mass is still a subject of ongoing research.

Meanwhile, advancements in the development of conventional electric machines have yielded an aerospace friendly device, rated at 1MW with a specific power of >13kW/kg[20]. This offers great promise for the actualization of electrical propulsion in the near term.

2.1.4 Turbine Engines

Turbine engines still play an important role in electrical propulsion architectures, due to the high power density offered. Therefore, they are relied upon as one of the primary power sources. Depending on the design targets, they are deployed either purely for power generation or a combination of power and thrust generation. This is expressed using the term thrust split, which ranges from 0% to 100%. The later percentage represents an arrangement where the turbine only produces shaft power and no thrust directly. The implementation of thrust split between the electrical fans and a turbine propulsion is sensitive to the overall system efficiency[21]. An assessment of TeDP configuration utilizing turbopumps, reveals possible fuel savings when thrust is split between electrical propulsors and core-engines, especially when the propulsors produce 60% to 90% of the thrust. However, this was highly dependent on the intake pressure losses of the propulsor and the TeDP system bypass ratio[21].

2.1.5 Electrical Energy Storage

Besides the turbine engines, electrical power banks serve as the other power sources relied upon by electrical propulsion systems. These are batteries,

supercapacitors and superconducting magnetic energy storage (SMES). These offer various advantages and suffer a few implementation and feasibility challenges. The measure of the viability of energy storage devices are their specific energy and specific power. Lithium batteries have high specific energy and have been identified as the key enablers of electric flight[22]. However, their specific energy is still limited to about 300Wh/kg, compared to the aerospace target of 1000Wh/kg[23]. Also in [24] it was shown that the performance of batteries are strongly affected by their state of charge, with significant degradation in delivered voltage over the operation time. Supercapacitors offer better specific power and are noted to have better performance than batteries[25], but suffers poor specific energy due to very low voltage capability. SMES offer promising value for the two evaluation criteria but incurs high mass penalties in cooling and charging accessories[26].

2.1.6 Electrical Transmission

Electrical transmission refers to the segment of the electrical network between the power generators and electric motors. The use of superconducting transmission cables has also been proposed, with the main challenge being the mass penalty and complexity of the associated cooling systems. Due consideration is also given to conventional copper and aluminium cables. Previous work in[23] shows that a transmission system utilizing aluminium can be achieved, with endurable mass penalty.

2.2 Electric Propulsion Aircraft Concepts

Several aircraft concepts have been proposed for the implementation of electrical propulsion systems. These are broadly based on passenger capacity and mission targets. Figure 3 is a chart of data obtained by the United States of America's federal aviation administration (FAA) in 2012, for fuel consumption of different aircraft sizes. This identifies the aircraft types, in which conventional propulsion systems need to be replaced or augmented with electrical propulsion systems.

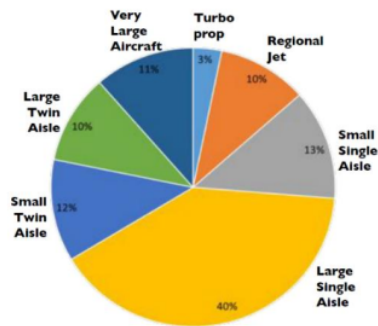


Figure 3 – 2012 Fuel Consumption, FAA US Operations Data[27]

A progression of technology development, passenger targets and required power capacity have been utilized, to propose candidate aircraft concepts to fulfil the electrical propulsion ambition. Some of these are presented in Figure 4.



Figure 4 – Proposed Aircraft Concepts for Electrified Airplanes[27]

Table 1 shows some of the proposed electrical aircraft concepts and the expected entry into service (EIS) timelines.

Table 1 – Proposed Electrical Aircraft Concepts

Entry into Service	More Electric Aircraft	All Electric Aircraft (AEA)
SHORT TERM (2020)		Small Planes (4 - 9 PAX)
		NASA SCEPTOR (X-57)
MEDIUM TERM (2035)	ESAero ECO-150	Medium Planes(18-21PAX)
	Boeing SUGAR Volt	
	Propulsive Fuselage (STARC-ABL)	
LONG TERM (2050)	BWB(N3-X)	

Five of these electrical aircraft concepts appear to have been researched more and are further discussed in the next sub-sections.

2.2.1 N3-X

The NASA N3-X is a derivative from the cruise efficient short take-off and landing (CESTOL) study in 2008[2]. It is proposed as an environmentally friendlier substitute for aircrafts similar in specification to the Boeing B777-200 aircraft. Up to 70%–72% reduction in energy usage is proposed to be achievable, with a significant reduction in exhaust gas emissions[14].

This concept involves the use of turbo-generators to generate and convert mechanical to electrical power for driving the propulsors. This is intended for a hybrid-wing-body (HWB) airframe configuration. The HWB airframe is designed specifically to improve aerodynamics of aircrafts and reduction of propulsion noise from engines[10,28]. Smaller size engines can be deployed, meaning their position can be changed without disturbing the propulsors' location and their performance. Also, the propulsors which are another noise source, are located such that the airframe serves as a noise barrier during take-off and landing[29]. Figure 5 is a representation of such a configuration.



Figure 5 – NASA N3-X Aircraft Concept[14]

The configuration consists of 2 to 4 wingtip mounted generators, with 14 to 16 propulsors [8,14,30]. The generated power is transmitted electrically to the electric propulsors, which are distributed across the top rear of the airframe. This concept is proposed for the 2050 EIS.

Three electrical power transmission architectures were reviewed for this concept in [31], with the view to obtain the optimal configuration. These were; alternating current (AC)[32], direct current (DC)[33,34] and a hybrid of AC-DC[31]. They comprised 4 major segments with sometimes overlapping components including, source, transmission, distribution and utilization[35]. Other architectures exist as variants of the DC architecture, with changes in number of propulsors and bus tie patterns for redundancy purposes[26,36,37]. Figure 6 shows a typical arrangement of segments common to all three architectures.

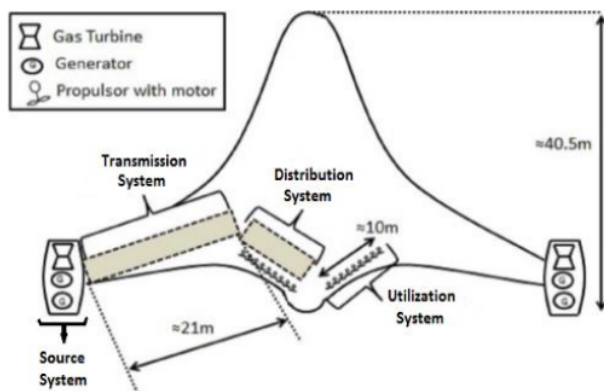


Figure 6 – Example of TeDP Power Segments[31]

An assessment of the 3 architectures was based on components, mass and efficiency. The conclusion reached was that a suitable choice would be dependent on the balance of mass penalty of DC converters and the cable losses in AC transmission.

The implementation of the concept relied greatly on advanced technologies such as superconductivity, cryocooling, high power density machines and energy storage devices. The main promises offered are, near-perfect efficiency profiles and an electric propulsion system, capable of similar thrust as currently deployed turbofans, with same or less mass penalty incurred. However, the technology readiness level (TRL) of these enabling devices, are still immature for the anticipated promises[30,38]. Specific challenges include materials and sustained cooling.

2.2.2 ECO-150R

The ECO-150R is an aircraft concept developed by Empirical Systems Aerospace (ESAero) in 2015. The main philosophy of the concept is the split-wing distributed propulsion system, which offered powered-lift by spreading the thrusting mass flow span wise[39]. This concept employs turboelectric configuration in a tube and wing airframe arrangement. It incorporates 16 electric propulsors, embedded in the wing. These propulsors are to be powered by a turbogenerator comprising of a turboshaft engine and an electric generator. Figure 7 shows the airframe configuration of the ECO-150



Figure 7 – ECO-150R Airframe Configuration[40]

A recent research on the concept was to optimize and improve on the challenges encountered in the development of its predecessors. All in pursuit of a projected

2035 EIS timeline. These challenges were mainly with the feasibility of superconducting technologies in the near term[41], propulsion-airframe integration[42] and electrical thermal management[43]. Figure 8 shows the developmental stages from concept to the currently discussed model.

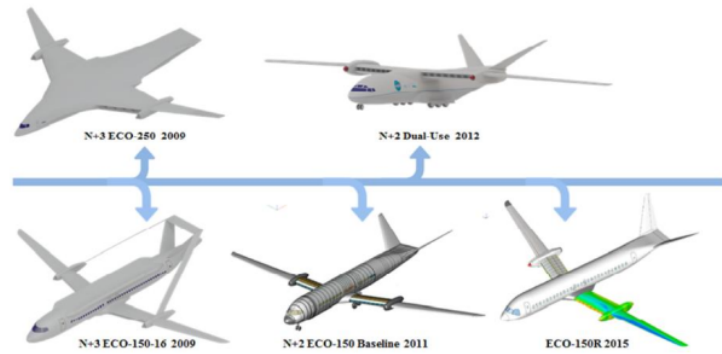


Figure 8 – Genealogical Evolution of the ECO-150R[39]

Table 2 shows the maximum values of design conditions and performance indicators for the ECO-150R as presented in [39]. The two key contributions from [39] include; the proposal of an inner bus tie architecture to improve the redundancy of the electrical system. The proposal of a thermal management system, which was a 10% improvement in weight ratio of the cooled device, arriving at a 20% weight ratio.

Table 2 – ECO-150R Design Conditions

Performance/Design Indicator	Value(SI Units)	Unit
Maximum Range	1646	nm
Cruise Mach	0.7	Mach
Altitude Limit	30000(9144)	ft
Maximum Thrust	12703(56505.8)	lbf
Total On-board power	10487(7820154.6)	hp

Despite these contributions, detailed complexity investigation of the associated drag and coolant piping network was still required. Also, it was recommended that a physics-based electric machine design and analysis algorithm be proposed. This is to pave way for the possible elimination of the gearboxes, since their efficiency and thermal characteristics were not considered.

2.2.3 NASA SCEPTOR X-57

The scalable convergent electric propulsion technology and operations research (SCEPTOR) project, which started in 2014 proposed to achieve flight testing in phases beginning in 2018[44]. Also known through the development stages as SCEPTOR and Maxwell, this 2 seat passenger capacity aircraft concept, was developed by NASA to achieve a minimum of 3.95 lift coefficient' for a stall speed of 58 knots[45]. This was intended to be achieved using distributed fans on the wings. Figure 9 shows the NASA SCEPTOR X-57 aircraft. The aircraft is designed for a cruise speed of 150 knots at 8,000ft, requiring 60kW per motor[46].



Figure 9 – NASA SCEPTOR X-57 Aircraft[46]

In [47] a study of nacelle cooling for the motors was performed. Candidate airflow paths investigated are shown in Figure 10.

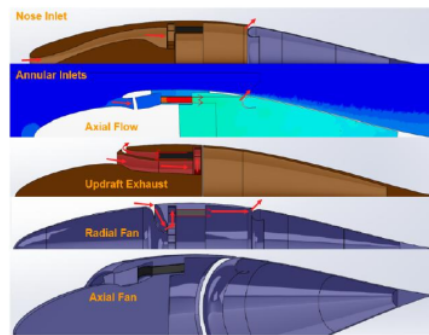


Figure 10 – Candidate Cooling Air Flow Paths for the NASA SCEPTOR Motor[47]

It was found that the annular inlets offered the best performance with least complexity. This was enough for an efficiency of 95%, keeping the temperature below 120°C despite ambient temperatures as high as 45°C. It was also found that the prop wash turbulence had just little contribution to the heat removal. The

motors were further equipped with fins and inter coil air gaps through which air was forced, to enhance heat removal. These successful inclusions led to the conclusion that increasing convective area yields more effective motor cooling than increased pressure, flow velocity or mass flow rate.

Further in [48], this ambient cooling technique was investigated for the cooling of the motor inverters. However, the air used was the exhaust air from the cooled motor, estimated to be 60°C. This air was channelled through a heat sink optimally designed to fit the inverter heat dissipating components. However, practical validation of the approach was recommended, since only a high-fidelity software was utilized for the thermal analysis. The performed analysis was for steady state operation, with a few assumptions. The results were promising, as the temperature was within 97°C, based on the assumptions made.

A transient thermal analysis was later performed in [49]. This was extended beyond the motor and inverter to the battery and transmission cables. A low fidelity modelling approach was employed this time, with some test data. The aim was to establish that the temperatures of non-air-cooled devices are within acceptable limits for the intended range of operation. The cable and battery temperatures were found to be within 68°C and 70.3°C respectively. The limitation of the work was the non-inclusion of component sizing in the performed optimization.

2.2.4 Boeing SUGAR Volt

The Boeing SUGAR comprise several electrical aircraft concepts, developed in a study conducted as part of the NASA subsonic ultra-green aircraft research (SUGAR) [50]. 10 aircraft models have been proposed as seen in Figure 11. An important outcome of the initial phase of the research was the conceptual design of the truss-braced wing (TBW) aircraft model[51]. The current stage of research has seen a detailed investigation up to the SUGAR Volt.

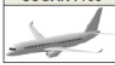
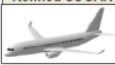
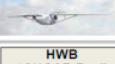
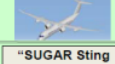
	N	N+3	N+3 Hybrid Electric	N+4
Conventional Tube & Wing	Baseline "SUGAR Free" 	Reference "Refined SUGAR" 		N+4 Reference "Super Refined SUGAR" 
Unconstrained* Conventional Tube & Wing	* Optimum span	"Super Refined SUGAR" 	"SUGAR Electric Eel" 	
High Span Truss Braced Tube & Wing		TBW High L/D "SUGAR High" 	"SUGAR Volt " 	Advanced N+4 Technology "SUGAR Freeze" 
HWB		HWB "SUGAR Ray" 	"SUGAR Sting Ray" 	

Figure 11 – SUGAR Hybrid Electric Aircraft Concepts[50]

The SUGAR Volt (Figure 12) is proposed to use two motor assisted turbofan engines in a tube and truss brace wing configuration. The concept is proposed to achieve a range of 900nm, with the capability of lifting 154 passengers. Two versions of the power utilization sequence were developed, the balanced and core shutdown. The balanced version utilizes the battery stored energy evenly throughout the mission while the core shut down utilizes full electrical power at take-off and climb, then cruises on fuel.



Figure 12 – Boeing SUGAR Volt Aircraft

The balanced version was seen to have an overall superior performance, with 60% fuel burn and 54% energy use compared to the baseline SUGAR Free, which is a representation of current conventional single aisle aircrafts.

Despite the benefits offered, integration concerns have been raised, especially for the batteries. Also, the aircraft concept is not able to meet the set-out noise reduction target.

2.2.5 Propulsive Fuselage (STARC-ABL)

For a near term EIS of 2035, the propulsive fuselage concept has been proposed. This concept intends to incorporate an additional propulsor at the tail of a conventional tube and wing aircraft configuration. This is with the view to optimally cancel out the airframe drag using the boundary layer of air on the airframe to generate thrust. This is proven in [52,53] to be a more efficient method to obtain net thrust.

As the free stream air collides with the fuselage in motion, effects of viscosity reduce the velocity of the air just adjacent the fuselage surface. Hence, a boundary layer of air with a different profile from the freestream air is created. Utilizing this reduced velocity air in the production of thrust, results in more net thrust for same input power.

Characterization of boundary layer air for aircraft propulsion has been done extensively in previous works[54][55], using the Navier-Stokes equations for viscous flow. Besides a reduced velocity compared to the free stream air, reduction in mass, momentum and pressure are also observed. These reduction results in about 1% distortion efficiency penalty[56]. Even then, BLI propulsors offers significantly better propulsive efficiency, estimated to reach 96.7% compared with 89% for a freestream ingesting engine at the same aerodynamic design point (ADP)[56].

Other research contributions on utilizing this BLI phenomenon to achieve better efficiency have been made by Bauhaus Luftfahrt [57] and NASA[11]. These utilize, mechanical and electrical power sources respectively. Both approaches offered promising expectations for the EIS of 2035, based on a correlation between TRL and a clearly defined roadmap. Figure 13 shows the earlier proposed power arrangements for the concept.

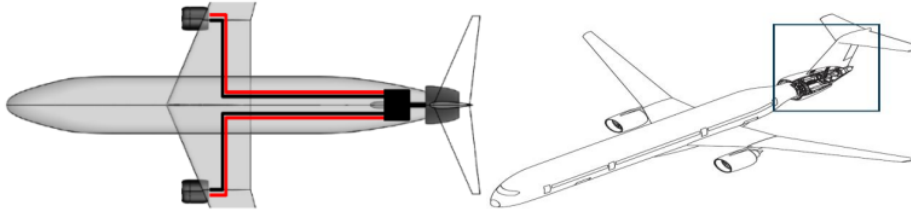


Figure 13 – Power Supply Approaches for the Propulsive fuselage Concept [58,59]

In [60], a side by side comparison of these two power supply sources shows that the mechanical approach had better block fuel reduction (11% vs 7.3% respectively) compared to a conventionally propelled aircraft of similar size. This was due to higher mass penalty of an electrical deployment, compared to the mass of a third power plant. However, in a more promising work by NASA[11], a 12% less mission fuel was estimated for single aisle 150PAX aircraft for a total range of 3500nm. A total additional mass of 3,480kg compared with a conventional tube and wing aircraft was estimated. This was intended to be electrically driven. It is worth mentioning, that the mechanical approach placed certain integration challenges like the position of the third power plant due to aft loading limits. Also, in [52], a comparison is made of these power supply methods, where the structural and flexibility advantage of an electrically driven propulsive fuselage is highlighted. It could be deduced therefore, that an electrical approach becomes more viable in terms of mission fuel reduction, if the electrical systems mass is optimized further. Figure 14 shows a notional power system comprising the electrical components for driving the fuselage propulsor.

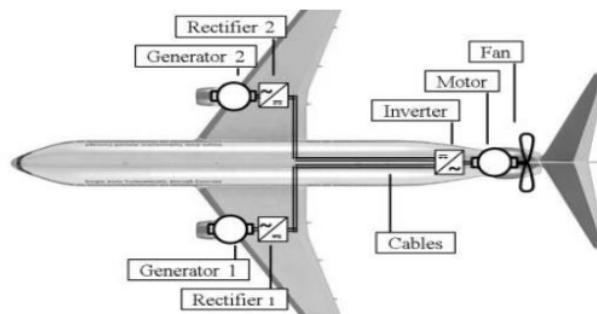


Figure 14 – Notional Electrical Power System of a Propulsive fuselage Aircraft Configuration[61]

In the electrically driven configuration, the main source of power are the two wing mounted turbofans. It is expected that more shaft power is extracted from one of the spools to drive an electric generator, installed coaxially with the turbine's fan. The electrical power output from the driven generators are transmitted to the motor at the aft of the aircraft through a rectifier and inverter. The generator and motor implement the decoupling ambition of turboelectric systems, while the rectification and inversion stages actualize the speed control flexibility. These are interconnected with high voltage capacity cables.

Preliminary sizing of these components have been performed in [61]. This was done based on specific power for a 2.6MW electrical system, designed to meet the thrust requirements presented in [62]. The both investigations showed that the proposed concept was promising enough to meet the set mission goals. However, further feasibility investigation was required, regarding the structural and electrical systems integration. The operational details of the electrical system need to be quantified and investigated to ensure the continued health of the system. This includes the power flow and optimal parameters selection.

In [63], the control of the entire propulsion system was modelled to investigate the ability of the proposed concept to meet the FAA transient response requirement. This included the electrical dynamics of the system. However, the performed investigation was reliant on the assumption that the two turbine spools were always at equal speed. This appears to be a scarce reality in practice, since prime mover speed droops are usually experienced with changing electrical loads. The consequence of a shaft speed disparity between the two turbines is that the generators would generate different output voltages. Where these generators are intended to power one device, it becomes a power management requirement to synchronize their output voltages before tying them up. Otherwise, the generator of lower voltage becomes an additional electrical load to the one of higher voltage.

In [11] it was stated that a thermal modelling approach was not available to demonstrate the heat rejection process for this concept. Although a rough

estimate approach based on SOA revealed that the loss of efficiency was significant.

2.3 Summary of Technical Challenges

There exist promising benefits associated with the successful deployment of the reviewed distributed propulsion systems, especially on civil aircrafts. The use of electrical transmission between the power sources and the propulsors, enables the physical decoupling, thereby enhancing the feasibility of many PAI concepts. However, certain technical challenges have been identified and have been grouped in the following subsection, highlighting the efforts made at proffering solutions.

2.3.1 Design and Integration

In [64], a vehicle trade study is performed for a hybrid-electric propulsion systems. The aim was to derive an optimum engine cycle for both cryogenic and conventionally driven motor system. The outcome was a revelation that the electrical system efficiency and mass were not impacted by the thermal engine cycle, and that this trend had minimal difference in both a cryogenic and conventional electrical system. A recommendation is made for such a study to be performed on varying motor sizes, since larger electrical systems would significantly impact vehicle weight. Also, a study for different mission lengths was recommended, since the amount of electric power on board the aircraft is a function of mission duration. For this, a wholly integrated sizing approach was required.

Further work was performed in [23] for a hybrid-electric system, where it is shown that for minimal CO₂ emission, the best hybridization strategy was one which lets the gas turbines operate at their optimum specific fuel consumption (SFC). However, the presented mass and efficiency trends were only declared valid for a hybrid-electric propulsion system, with a recommendation for similar studies to be performed for turbo-electric systems. More so, that the results obtained did show that a pure hybridization would not fulfil the set environmental target.

The importance of incorporating the electrical systems sizing in the conceptual aircraft design has been highlighted in [65]. A method was hence proposed to meet this requirement, to prevent the trend of oversized electrical systems for future aircraft propulsion systems. The presented approach was based on mathematical techniques, with focus on the logical operational sequence of components.

In [66], a retrofit study was performed, to evaluate the impact of changing subsystem performance characteristics at system and mission levels. An iterative approach was used, along with physics-based models. However, this was done for a hybrid electrical system with the hybridization occurring only at cruise. The possibility of generalizing the presented approach to non-hybrid systems was not mentioned.

2.3.2 Systems Mass and Efficiency Estimation

A study of mass and efficiency of the turboelectric propulsion components was conducted by [38], in which a configuration consisting 2 superconducting generators, 14 superconducting motors, superconducting transmission lines, inverters and cryocoolers was adopted. A total electrical system mass and efficiency values were presented as 20.7klb and 98.55% respectively for BSCCO and 25.7klb and 98.03% for MgB_2 . It was further revealed that there is a high dependency of the overall system mass and efficiency on mass per input power of the cryocooler and the percentage loading[38]. Another study by [67] has presented a total electrical component mass of 16klb and an overall efficiency of 98%. These values were for components in a DC based superconducting transmission for a rated total power of 50MW. Presented mass and efficiency estimates were for both electrical components and the turboshaft engine.

A methodology for the estimation of the mass of fully superconducting machines have been proposed by [68], in which a relationship between the machine torque per unit rotor volume is exploited, to determine the entire machine mass. This was achieved by the product of the machine volume and an active density. Active density values between 2000kg/m^3 and 8000kg/m^3 was recommended. This method proved to be very useful and was validated, as values obtained were

within acceptable range of previously designed similar machines and was also used in the mass estimation of the machines in the distributed electrical aerospace propulsion (DEAP) project.

A tool for the evaluation of the sensitivity of designs to mass and efficiency has been developed in [34]. The tool helps in the appraisal of electric power designs for TeDP by comparing the impacts of the architecture features, redundancy configuration and protection strategies on the overall system mass and efficiency. System components and their various sensitivity and control variables are highlighted. The study further revealed that the reduction of electrical losses is key to achieving a viable superconducting electrical system for TeDP. Most of the degradation in mass and efficiency are traceable to the solid-state switching components. However, a trade-off between components seems inevitable since the overall system sensitivity was found to tilt after a reduction of the impact of one element on the system. A formal contingency analysis performed by [37] also highlights the impact of transients and redundancy considerations on mass and efficiency sensitivity of systems.

The sizing of power components for a non-superconducting electrical system was performed in [61]. This was done using specific power values obtained for SOA components for 2 airframe configurations. An overall electrical system mass of 1394kg, was estimated for a 2.8MW turboelectric propulsion system, operating at 89.1% efficiency. Values obtained for a series hybrid configuration were 2250kg for a 381kW system operating at 82.1%.

A mass versus efficiency optimized sizing approach was investigated in [23]. It revealed that mass was more sensitive to little efficiency differences than efficiency was, to little mass differences. Therefore, it was recommended to focus components design and sizing on optimized mass.

2.3.3 Electrical Thermal Management

The management of heat dissipated by electrical components of a propulsion train are expected to be significant, since they are a result of efficiency losses.

In [9], five cooling techniques were reviewed and compared for their viability towards aircraft propulsion applications. The measure of viability used was the ability to evacuate large amounts of heat. The highlighted challenges included the mass of the coolant reservoirs and pumps, potential leakages, slow response, changes in coolant characteristics with altitude and the requirement for additional power to drive the cooling systems. Despite these revelations, the actual sizing of the thermal systems was not performed.

In [39], a recirculating liquid cooling system was developed for the ECO-150R electric transmission. This utilized ducted radiators for the rejection of heat with the aid of ram air. A liquid coolant was required to transport the heat from the electrical devices to the ducted heat exchanger. The proposed thermal management system was estimated to have a 20% weight ratio of the cooled device. Despite this, the approach could not successfully cool the electrical system as proposed. Recommendations were made for the optimization of the duct characteristics and the advancement of the method. Other concerns include the added complexity and weight of the coolant pumps and overhead accessories.

A thermal investigation have also been performed in [48,49] for a 2 seater aircraft, with a recommendation for such an investigation to be done on a scalable basis. An overall electrical system sizing was performed in [23] for a hybrid propulsion configuration. It was shown that the cooling systems contribute about 50% of total architecture losses and about 4% to 8% the total system mass.

An integrated systematic approach was utilized in [69] for the investigation of aircraft systems thermal management. It was concluded after a sensitivity study, that a 150-seater TeDP aircraft was infeasible due to thermal handling constraints. This was despite conservative application of SOA technology estimates. The identified challenges include the additional power required, incurred mass and ram drag penalties. An estimated 11,000lb mass is incurred for a 0.8MW heat.

Thermal analysis of potted Litz wire was performed in [70]. Litz wire has become an enabler for high power density machines, due to its adaptability to high

frequency AC currents with low losses. The performed investigation revealed that the machines built with Litz wire could operate at temperatures up to 193°C, with the aid of a fast-flowing coolant. However, the mass of the cooling system was not presented.

2.3.4 High Voltage Selection and Impact

The selection of optimal voltage is an important part of the electrical systems design, since all devices should be rated for a certain range of voltage, within which they would function normally.

In [71], a method for the calculation of safe voltage rating was presented. This was only for the insulated cables in an aircraft propulsion system and did not consider the impact of the selected voltage on the sizing and performance of the other architecture components. Similar studies aimed at the selection of the optimal voltage at the optimal size were performed in [36] and [26]. The two studies covered parameter selection and design procedure, based on superconducting technologies. However, non-overlapping voltage range was proposed for the same aircraft from both studies. In [68], the study of electrical components, sizing of electrical machines and investigation of superconducting turboelectric aircrafts was performed. The voltage selection was done for only the electrical machines at the design point. In [72] Presented an approach for modelling superconducting motors using a modified conventional machine dynamic model. An assumed voltage of 28V was used, detailed propeller to motor matching is performed but other electrical components were not considered. In [73] the power flow technique is used to correctly predict buses voltages and currents at different torque-speed combination. This was obtained for steady state and transient state scenario involving a sudden change (increase) in torque demand. The approach only covers power circulation within a system with known system voltage. No initial architecture voltage derivation, sizing process, start-up and fault scenarios was considered. In [23], a fixed motor voltage of 1000V is used for the investigation of a hybrid electric propulsion system. It was shown that the system mass was optimized if the upstream components were adapted to variable voltages, to eliminate the voltage regulation devices. A trend of

increasing mass and decreasing efficiency was observed for increasing ratio of voltage to power.

2.3.5 Safety, Stability and Reliability

The implementation of an optimized electrical propulsion system appears feasible on the long run and would help fulfil the set environmental goals. However, issues of operational safety, stability and reliability still occupy an important position in the scheme of things. For such large magnitude power as would constitute future aerospace propulsion trains, an effective protection from faults and ride through transients are required.

In [74] it was stated that redundancy and reliability levels were part of the crucial functional requirements for a turboelectric aircraft power system, together with independent speed and power control. In [75], stability, transient response, control, and safety studies was performed for a superconducting DC architecture. The recommendations put forward were for further studies on protection strategies, power rerouting and vehicle intersystem interactions. In [76] a stability study of electric power systems is performed, based on converters for different aircraft architectures. It was concluded that a pure DC source and bus system offered the best stability for a MEA. However, system dynamics (especially of motors), overall system efficiency and thermal management issues were not considered. A recommendation is made for a more comprehensive research.

2.4 Identified Knowledge Gaps

The reviewed literature does reflect significant research progress towards achieving better aircraft performance for the future. However, there are identifiable knowledge gaps which needs to be filled, to enable the implementation of these very promising electrical propulsion concepts.

- **Design and Integration** - Many of the design and integration approaches [23] [64] [65][66] appear to be tailored towards a select aircraft concept. It is still unclear what design parameter has the most influence on the viability of turboelectric propulsion systems. This leaves the designer with too much

uncertainty in the evaluation of the viability of the immediate electrical propulsion concept.

- **Voltage and Current Selection** - Despite works done in [71], [36] [26] and [68], the impact of voltage and current selection on the fidelity of the propulsion system design, based on conventional conducting devices is still not clear. More so, is the absence of a clear procedure for deriving the optimal voltage for a specific electrical propulsion configuration.
- **Stability and Reliability** - There also appears to be a diminished attention to the systems susceptibility to faults and the handling strategies thereof. Attempt is made in [23], using simple multiple of the propulsion system parameters, since it was not the focus of the study. The absence of a study showing the impact of stability and reliability considerations, on turboelectric propulsion systems sizing and integration, does constitute a knowledge gap.
- **Devices Thermal Limits** - An extension of system instability concerns is the requirement to know the thermal behaviour of the propulsion systems both at normal and abnormal operation. This stems from the reality of temperature rise with such increased current flow, as is associated with peak loading and fault times. A knowledge gap is identified in the absence of a clear procedure to establish that the devices stay within their temperature capability at all operating conditions. Also is the absence of a clear roadmap to guide the necessary improvements that are required for high magnitude thermal handling requirements.

2.5 Summary

From the literatures reviewed, systems optimization is the key driver for future aircraft design. Most of the challenges are power density related, manifesting in form of mass penalties as an attempt is made to scale up currently existing technologies. The introduction of superconducting devices appeared to be a way out, but the incurred mass and complexity of the associated cooling devices puts their use to the far term. This is due to a low technology readiness level. The limits of energy storage devices also appear to be a major challenge for the implementation of hybrid or all electric aircraft concepts. Therefore, the



The intended contribution of this work is the characterization of the fault and thermal behaviour of such a system, since a knowledge gap exists. Several approaches exist for electrical networks deployed for terrestrial, train and ship utilities. However, these are not necessarily transferable to aircraft propulsion due to power density challenges. The interrelationship between these two aspects will be demonstrated, since there is typically an instantaneous temperature change tracking the current changes. The need to keep these temperatures below the thermal capabilities of the on-board devices, becomes the motivation for this work. Their correlative impact on cooling systems and the overall sizing of the electrical network would also be demonstrated.

3 FAULT, TRANSIENT AND THERMAL CHARACTERIZATION

The need for fault and thermal characterization has been enumerated in the previous chapters of this thesis, owing to their contribution to the stability and reliability of the electrical propulsion system. This chapter presents the fault and thermal characterization procedure. This is done considering all operation critical equipment. The aim is to identify parameters that influence faults and thermal characteristics of electrical network, in the context of aerospace propulsion. This will provide a designer with some insight, in pursuit of reliability and optimization.

3.1 Initial Electrical Network Modelling

In order to put the fault and thermal characterization in perspective, a notional electrical network dedicated for aircraft propulsion is required. Figure 15 shows a combination of devices typical of a turboelectric propulsion system.

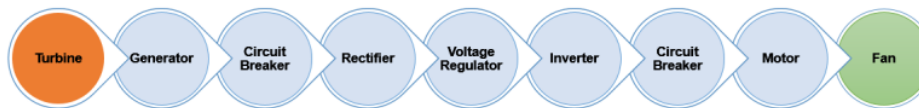


Figure 15 – Components in a Notional Turboelectric Propulsion Train

The blue background represents the electrical devices which are considered in this chapter. They are interconnected using cable transmission lines which are the assumed fault locations. The thermal behaviours of these devices both in normal and abnormal operation will also be studied. A detailed electrical network modelling would be aircraft specific, because for different aircraft configurations, some devices may occur in multiples or be omitted. Chapter 6 of this thesis presents an example.

3.2 Fault Characterization

Electrical network faults imply all that constitute an abnormal functioning of part or whole of electrical networks. Depending on the magnitude, their impact could range from harmless short periods of abnormal power delivery, to total interruption of the power supply and sometimes fatality. Technically, they can be



to the fault point. In such a scenario, the continued spinning of the motor due to reserved fan inertia, causes a current to flow back towards the fault point.

Hence, the need for fault characterization for electrical propulsion systems is to properly quantify the location and magnitude of possible fault hazards, considering the peculiarities and limitations of an aircraft environment. This would make it possible to consider appropriate contingency arrangements, early in the design stages.

Previous work[75] describing the fault characteristics in aerospace propulsion systems was based on a superconducting network system, which utilized unconventional transmission cables. However, the sizing of the protection devices was based on a scaled approach of existing systems, using voltage and current trends at normal operation. Also, the sizing of propulsion systems components, including protection devices was done in [61]. Specific power scaling was used, without regard to the voltage and current combination of the considered system, compared to the referenced protection device upon which the scaling was performed. The work in [23] went a step further, by using a fault current multiplier of two times the rated current. However, it was not established if this multiplying factor would be true for all cases. This poses concerns as the current rise time in a short circuit scenario should differ for various networks and the multiplier may be exceeded earlier than expected, leading to a hazard.

These early attempts at protection device sizing did not establish the magnitude of the fault currents for the considered propulsion system. This information is important for its influence on protection systems and overall architecture sizing, as well as transient response durations that govern fault interruption settings. The presented approaches also do not reflect the influence of faults and transient magnitude on the thermal behaviour of the electrical network. It needs to be established that the components of the propulsion system must be able to withstand the faults for the duration of the fault transient, without exceeding their maximum temperature capability.

The next sections present a procedure for the derivation of fault current values at a given point on an electrical propulsion network. A combination of RLC circuit analysis and simulation are utilized to achieve this. As is typical, only short circuit faults are considered, since they have the greatest magnitude[78]. The high magnitude is due to them being initially fed by the power source. Other fault scenarios are not covered also, because the intension of this research is to enable the adequate sizing of the protection devices and not the proposal of circuit protection strategies and control schemes.

3.2.1 Fault Current Estimation

Fault currents comprises the steady state and transient values. The DC fault current at a given point between two interconnected DC devices or network branches, can be obtained using equation (3-1)[79]

$$I_{DC_Fault} = \frac{V_{DC}}{R} \left(\frac{1 - e^{-\left(\frac{t}{\tau}\right)}}{1 - e^{-\left(\frac{t_p}{\tau}\right)}} \right) \quad (3-1)$$

Where V_{DC} is the DC system voltage, t is the instantaneous time after the fault occurrence and t_p is the time it takes the fault to rise to approximately its maximum value. The rising time constant τ for a series RLC circuit is given by equation (3-2).

$$\tau = \frac{2L}{R} \quad (3-2)$$

R and L are the fault network resistance and inductance respectively and $t_p \approx 5\tau$.

Equation (3-3) from [78], gives the AC fault current at a given point between two interconnected AC devices or network branches.

$$I_{AC_Fault} = \left(\frac{\sqrt{2}V_{rms}}{\sqrt{R^2 + X^2}} \sin(\omega t + \alpha - \phi) \right) + \left(\frac{\sqrt{2}V_{rms}}{\sqrt{R^2 + X^2}} \sin(\alpha - \phi) e^{-\left(\frac{R}{L}\right)t} \right) \quad (3-3)$$

Where V_{rms} is the AC system voltage, X is the reactance of the fault network, given by equation (3-4). ϕ is the AC network phase angle given in equation (3-5), while α is the voltage angle. These are shown on the graphical presentation of an AC voltage in Figure 16.

$$X = 2\pi fL \quad (3-4)$$

$$\phi = \tan^{-1} \frac{X}{R} \quad (3-5)$$

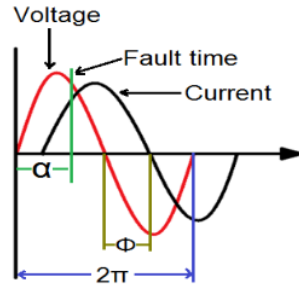


Figure 16 – AC Voltage Graphical Representation

The maximum AC fault current (available short-circuit current) is obtained when the sin terms $\sin(x) = \pm 1$.

In order to compute the fault currents using equations (3-4) and (3-5), it is required to know, not only the power rating of devices in terms of voltage and currents, but also the components' impedances. In [80] a single supply single load fault scenario was considered, using only a simulation approach with previously obtained RLC values. Also in [81], fault analysis involving multiple DC sources were performed, using a matrix based analysis and known impedance values. The aim was to show that numerical calculations using an RLC circuit

analysis, could yield correct values and save more time and computation resources, compared to a complete network simulation. These presented approaches did not cover effects of DC link faults on load sides. Also, these fault analysis were not for the bus topology, which is more peculiar to most proposed aerospace concepts. Therefore, a procedure for obtaining RLC estimates for components of a propulsion system, in a bus topology configuration is presented in the next section.

3.2.2 RLC Network Computation

An RLC network is the basic building block of electrical systems. It relies on the tendency of components to either dissipate heat, build a magnetic field or store charges when electric current is passed through them. These phenomena are respectively described as resistive (R), inductive (L) and capacitive (C) behaviours. They represent an impedance to the flow of electric current, leading to a reduction in power delivered and hence the system's efficiency. However, more capacitive materials foster flow of current due to their inherent ability to store charges.

The formation of RLC circuits and computations for an aerospace propulsion system, begin with the assumption that the generators have their voltages rectified to DC before any interconnections. This is also advisable, as such early interconnection are usually avoided where practicable, to avoid the challenges and complexity of the required synchronization. This assumption simplifies the stability studies, as no multi-branch analysis is required on the AC supply side and only the back effects of the upstream faults in the DC side is considered. Figure 17 depicts the RLC representation of the interconnected components in Figure 15. Block A to H represent the generator, AC transmission, rectifier, DC link filter, DC-DC converter (optional), DC transmission line, inverter and motor respectively.

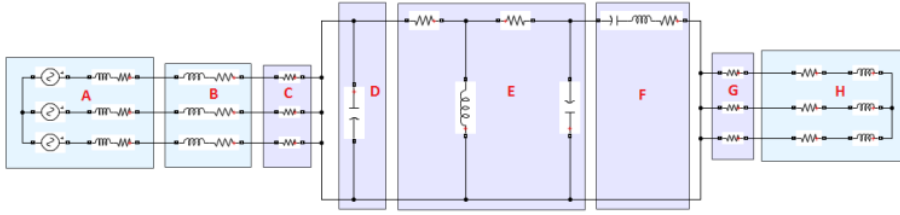


Figure 17 – RLC Representation of a notional Electrical Propulsion Network

The equations for the derivation of the impedance for the propulsion network, are presented as a function of the motor variables. This is because the entire network is set up, purposely to drive the motor which turns the propeller. Equation (3-6) gives the impedance of the n^{th} device in a serial propulsion system configuration.

$$Z_n = \frac{\left(\frac{1}{\eta_n} - 1\right) V_0^2}{P_0 \prod_{k=0}^{n-1} \eta_k} \quad (3-6)$$

Where P_0 and V_0 denote the power and voltage of the motor or the ultimate electrical load while η is the efficiency. The derived impedance comprises the resistive, inductive and capacitive components, which are obtained using equations (3-7), (3-9) and (3-10) respectively.

$$R_n = Z_n \times pf \quad (3-7)$$

$$X_n = \sqrt{Z_n^2 - R_n^2} \quad (3-8)$$

$$L_n = \frac{X_n}{2\pi f_n} \quad (3-9)$$

$$C_n = \frac{1}{2\pi f_n X_n} \quad (3-10)$$

The computation of equations (3-7) to (3-10) requires known values for the power factor (pf), operating frequency (f_n) and each components efficiency. These are

obtained during the propulsion system sizing. However, for preliminary studies, iterative processes could be utilized. Recommended minimum values of 0.95 and 50 Hz [26] could be used for pf and f_n respectively. Also, the efficiency values for SOA devices could be obtained from[61].

It is important to state also, that for purely DC network branches, $pf = 1$ and $f_n = 0$. However, at the instance of fault occurrence the change in current as it rises to fault level, meets an inductance in addition to the very minimal ripples in the DC supply. This inductance needs to be derived, as it forms part of the time constant in equations (3-1) and (3-3).

Figure 18 is a flow chart of the algorithm utilized for the derivation of the RLC and fault currents.

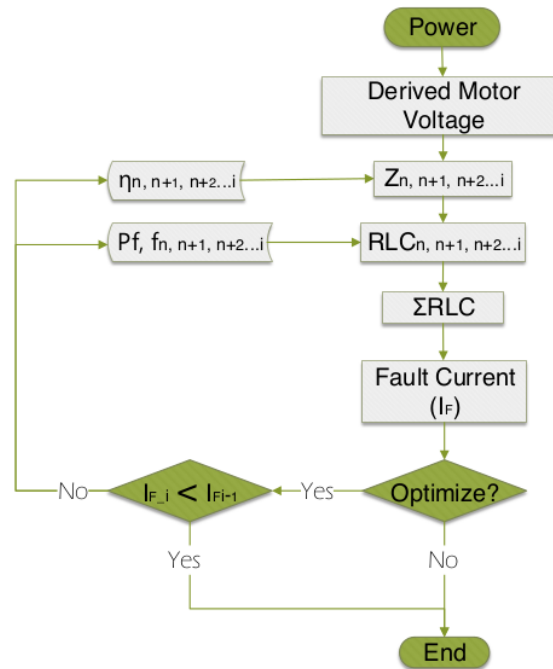


Figure 18 – Flow Chart for Fault Current Estimation

The estimated fault current refers to a fault occurring after a device. It is a function of the total current supplied to that point, the impedance of the current sources, of the transmission medium up to that point and of all components in-between

source and fault, which retain their connection during the fault. These are represented by the series in the flow chart, where the n^{th} term corresponds to the device interfacing with the fault, while the i^{th} term corresponds to the power source. The optimization check is only required for conceptual studies in which final component sizing and operating conditions have not been set. The voltage and current derivation procedure for the propulsion system, would be presented in chapter 4.

3.3 Motor Transient Behaviour

The need to study motor transient responses is informed by the peculiarity of aerospace applications. For instance, the changing loads on the fan blades due to continuously changing flow characteristics. Also is the requirement to move from idle to 95% thrust in 5 seconds[63]. These constitute sources of electrical transients.

In practice, motor start-up and transient responses involves a current inrush, which depending on design fidelity could be likened in magnitude to a short circuit current. So, it is important to estimate the inrush current along with the fault current magnitudes, to ensure appropriate switch loading limits and establish a set point, at which the control system distinguishes between an inrush current and a fault current. This enables the protection device to discriminate and handle such inrush current without interrupting services. Earliest interruption of fault current is desirable, to eliminate arc formation[82] and minimize the associated temperature rise.

It is shown in [78] that there exists a relationship between electrical network stability, faults and machines inertia. This informs the need to investigate the influence of the motor's inertia on the network stability characteristics and sizing.

Common methods for the estimation of motor inrush current, is the use of standard values for different classes of motors or the values provided on device nameplates, as multiples of the rated current. However, these do not reflect the actual influence of optimized motor impedances and soft starting methods on the

inrush value. Hence these values may be either too optimistic or pessimistic and could not be used for an optimized propulsion systems design.

A simulation approach is adopted in this work, based on the motor dynamic equations. The minimum instantaneous current drawn by a permanent magnet electric motor is given by equation (3-11).

$$I = \frac{4 \left(\frac{d\omega_R}{dt} J + T_m + B\omega_R \right)}{3N_p \lambda} + I_R \quad (3-11)$$

This represents the best inrush scenario, where the angle between the stator and rotor field flux are at 90° [83]. This is a feature peculiar to non-salient pole machines. In this state, the motor torque responds fastest to change in stator currents. Therefore, the time rate of speed change $\frac{d\omega_R}{dt}$ can be utilized to estimate the required current, counting from initial time to a time required for full thrust delivery. I_R is the rated motor current, J represents the total inertia load on the motor shaft, T_m is the mechanical torque delivered by the motor, B is the coefficient of friction of the motor bearings, N_p is the number of stator poles while λ is the flux linkage between stator and rotor. A validation of this approach is done experimentally and presented in chapter 5.

3.4 Thermal Characterization

In order to model the thermal behaviour of the propulsion system components at fault times, a full thermal characterization for normal operation needs to be done, since the devices are usually in operation prior to a fault. Despite being related to the fault investigation, the thermal characterization of the entire electrical network also forms a major contribution of this work, since it enables a propulsion system designer to arrive at models, which operate within temperature limits all through the mission. The RLC models need to be advanced into physical models, to help assess the correlation between size, heat generation and removal rate.

Thermal characterization for electric propulsion systems has been performed earlier in [48] and [49]. This was for an all-electric system rated at 14kW. The

limitation of the work was the non-incorporation of component sizing in the thermal characterization procedures. This limits the extent of scalability of the presented methods, with respect to temperature limits, mission range and type of aircrafts. This section presents a modelling of electrical propulsion components and equipment, proposed for large, long range aircrafts.

The principle of heat flow as depicted in Figure 19 is adopted, which represents heat flow from source to ambient through interfaces i , $i + 1$ up till the n th interface.

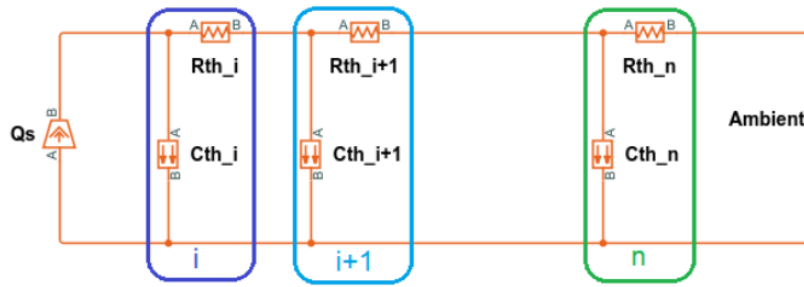


Figure 19 – Typical Heat Dissipation Path

Q_s is the heat flow from the heat source, R_{th_i} , $R_{th_{i+1}}$, and R_n represent the thermal resistances of the interface materials while C_{th_i} , $C_{th_{i+1}}$, and C_{th_n} represent the thermal capacitances of the interface materials. The instantaneous temperature of a device (or interface) is a function of the heat flow into it, its thermal resistance and capacitance, and those of the interfaces downstream towards the ambient air. This can be expressed mathematically as in equation (3-12)[84].

$$T_s = T_{Amb} + Q_s R_{th_i} (1 - e^{\frac{-t}{RC}}) \quad (3-12)$$

Where T_{Amb} is the temperature of air around the device, t is the instantaneous time, while R and C represent the thermal resistances and capacitances respectively of the entire system interacting with the heat. The thermal resistance and capacitance of an interface material can basically be derived using equation (3-13)[85] and (3-14) respectively.

$$R_{th} = \frac{L}{kA} \quad (3-13)$$

$$C_{th} = mc_p \quad (3-14)$$

Where L is the thickness of the interface or the perpendicular distance between the two temperature regions in which it is situated, A is the total surface area in contact with the colder region, m is the mass of the material, k is the thermal conductivity of the material, while c_p is the specific heat capacity of the material.

These principles are next applied to obtain the instantaneous temperatures of the propulsion system devices, considering the geometric peculiarities of each.

3.4.1 Electrical Machines Thermal Modelling

Electrical machines refer to motors which convert electrical to mechanical power and generators which convert mechanical power to electrical power. These machines are the main enablers for electrical propulsion. Since the shape and constituent materials affect the rate of heat emission, it would be required to select a type of electrical machine to be considered. Electrical machines are broadly divided into family groups as shown in Figure 20.

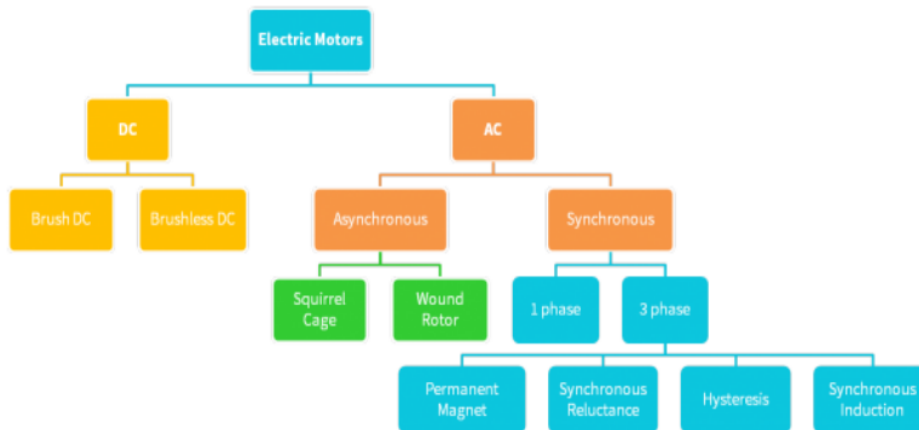


Figure 20 – Electric Machines Family Tree[86]

AC machines are preferable to DC machines for high power applications, because they eliminate commutation brushes which often require replacement. Synchronous machines track the load in both rotations per minute (RPM) and torque, whereas asynchronous machines only track the torque (i.e. drops speed to maintain torque in the event of changing load). This may be undesirable in such application where the fan must be kept at a certain RPM to produce the required thrust. Hence synchronous machines are considered for aerospace propulsion applications. Investigation in [87] showed that permanent magnet machines are better than electro-magnet rotor machines, on the basis of better mass. Furthermore, electric machines currently deployed in small aircrafts propulsion are permanent magnet synchronous machines (PMSM)[47,88]. These have also been proposed in [89][90] for future medium and large aircraft, due to their high specific power and minimal complexity. Thus, enabling them to be seamlessly deployed as motors or generators, by the application of voltage or torque respectively. Following this trend, a PMSM is considered for the thermal modelling approach presented in this work.

Previous works on electrical machines thermal modelling[91–93] were based on the thermal resistance and conductivity of the machine sub components. However, detailed procedure for the calculation of these thermal resistance and capacitance were not presented. This section presents a procedure for these derivations. The presented approach has the added benefit of capturing in same instance, the effect of machine's mass build-up on the thermal behavior.

Given the interrelationship between efficiency and device temperatures, the machine efficiency is first computed, as a function of the winding losses. It is shown in [94] that the winding and lamination conductivity had the most influence on the motor temperature. Equation (3-15)[95] gives the machine efficiency η_M .

$$\eta_M = \frac{VI - (I^2 R_w + VI_0)}{VI} \quad (3-15)$$

Where V is the machine rated voltage, I_0 is the no load current, I is the instantaneous current and R_w is the winding resistance. The heat flow can be obtained using equation (3-16).

$$Q_M = P_{Rated} (1 - \eta_M) \quad (3-16)$$

The thermal resistances R_{th_M} and capacitances C_{th_M} are computed based on the cross-sectional representation of the machine as depicted in Figure 21.

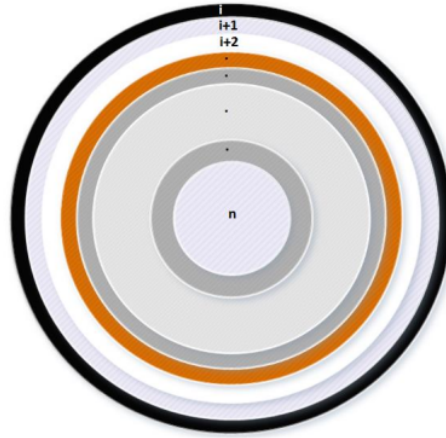


Figure 21 – Cross-Sectional Representation of a Notional PMSM

The i th annulus represent the machine casing while the n th represents the machine shaft, with a few annuli in-between, depending on individual design peculiarities. Equation (3-17) and (3-18) respectively, give the total thermal resistance and capacitance of the machine.

$$R_{th_M} = \frac{1}{2\pi L k_n} + \sum_{i=1}^{n-1} \frac{\ln\left(\frac{r_{i+1}}{r_i}\right)}{2\pi L k_i} \quad (3-17)$$

$$C_{th_M} = \pi r_n^2 L \rho_n c_{p_n} + \sum_{i=1}^{n-1} \pi (r_i^2 - r_{i+1}^2) L \rho_i c_{p_i} \quad (3-18)$$

Where, the r terms represent the outer radius of the referred annulus, the ρ , k and c_p terms respectively represent the density, thermal conductivity and specific heat capacity of the material constituting the referred annulus, while L is the length of the machine. The instantaneous temperature of the machine winding T_M is of most concern, because it constitutes the heat source due to its resistance to the flow of current. Operating beyond its thermal limit leads to a breakdown of the insulation. This temperature can be estimated using equation (3-19).

$$T_M = T_{Amb} + Q_M R_{th_W} \left(1 - e^{\left(\frac{-t}{R_{th_M} C_{th_M}} \right)} \right) \quad (3-19)$$

Where R_{th_W} is the thermal resistance of the winding and can be computed from one of the series components of equation (3-17).

3.4.2 Power Electronics Thermal Modelling

Power electronics refer to electrical elements, which do not exhibit translational or rotational motion, in order to perform their power conversion or control functions. They include semiconductor-based switches, integrated circuits, inductors and capacitors. A functional combination of two or more power electronics yields an electronic device. Power electronics devices proposed for aerospace propulsion include rectifiers, inverters, voltage regulators, and circuit protection devices. The following subsections present the sizing, efficiency derivation and thermal modelling of these power electronics components. Device level details would be presented in chapter 6, since different aircraft configuration and power requirements affect the constituents, number and arrangement of these power electronics components in the realization of the device.

3.4.2.1 Switch

Switches proposed for use in aerospace are insulated-gate bipolar transistor (IGBT) and silicon carbide metal–oxide–semiconductor field-effect transistor (SiC MOSFET). Their sizing is reliant mostly on the current and voltage handling limits. Previous works have approached the sizing of switches in aerospace application, by referencing details of already existing switches[23] and by use of

regression[26]. The sizing adopted in this work aims a balance between the two previous approaches, by the consideration of the SOA. This is to reflect the improved power density of the currently available switches. Also, SiC MOSFET are considered in this work, in contrast to previous works which were based on IGBTs. SiC MOSFETs boasts superior efficiency and less thermal dissipation[96], and hence has arrived at a TRL considerable for aerospace applications.

The instantaneous temperature T_S of the switching device can be obtained with equation (3-20).

$$T_S = T_{Amb} + P_S R_{th_S} \left(1 - e^{\left(\frac{-t}{R_{th_S} C_{th_{CG}}} \right)} \right) \quad (3-20)$$

Where P_S is the instantaneous power loss of the switch, R_{th_S} is the thermal resistance of the switching device, $C_{th_{CG}}$ is the thermal capacitance of the cooling group. The cooling group here refers to the attached heat sinks and other heat evacuation mechanisms. The procedure for the derivation of $C_{th_{CG}}$ will be presented in section 3.5, along with the introduction of a novel cooling strategy. Only switches with incorporated diodes have been considered, since they can handle a reverse flow of current. Equation (3-21) can be used to derive the switch power loss P_S .

$$P_S = P_T + P_D \quad (3-21)$$

Where, P_T and P_D are the switching transistor and diode losses respectively. These are given by equation (3-22) and (3-23) respectively from [97].

$$P_T = I_{RMS}^2 R_{DS_{ON}} + f_{sw} E_{sw} \quad (3-22)$$

$$P_D = I_{F_{av}} V_{F0} + I_{RMS}^2 R_F + f_{sw} E_{Rec} \quad (3-23)$$

Where, I_{RMS} is the RMS value of the input current to the device, R_{DS_ON} is the drain-source on-state resistance of the switch, V_F is the conduction voltage drop in the diode, I_F is the maximum forward current capability of the diode, I_{F_av} is the average value of the current through the diode and V_{F0} is the diode on-state zero-current voltage. f_{sw} is the switching frequency while E_{sw} and E_{Rec} are the switching and recovery energies in the transistor and diode respectively. A variation of R_{DS_ON} with instantaneous temperature can be obtained using equation (3-27).

$$R_{DS_ON} = R_{DS_ON_25} \times (1 + \alpha(T_S - 25)) \quad (3-24)$$

Where the temperature coefficient α is given by equation (3-28)

$$\alpha = \frac{R_{DS_ON_Max} - R_{DS_ON_25}}{T_{S_Max} - 25} \quad (3-25)$$

Where $R_{DS_ON_25}$ is the switch resistance at 25°C, T_{S_Max} is the maximum switch temperature, and $R_{DS_ON_Max}$ is the switch resistance at the maximum switch temperature. The same procedure for obtaining the temperature dependent value of R_{DS_ON} is utilized for R_F . It is required to obtain E_{Rec} , this can be done using equation (3-27) [97].

$$E_{Rec} = \frac{(Q_{rr} \times V_D)}{4} \quad (3-26)$$

Where Q_{rr} is the reverse recovery charge and V_D is the supply voltage. The parameters for the switch power loss calculations could be obtained from a published switch datasheet. Alternatively, they could be derived using regression from a collection of sample switches. The approach adopted in this work is to use data of the SiC MOSFET most suitable for the present research. This choice has been made on the combined balance between efficiency, thermal endurance, maximum current capability and specific power. Appendix A presents details of SiC MOSFETs reviewed. While Table 4 shows the design parameters of the switch adopted in this work.

Table 4 – Selected Switch Specifications

Voltage(V)	1200	$R_{F_{25}}$ (Ω)	0.0045
Max Voltage(V)	4000	$R_{F_{Max}}$ (Ω)	0.01
Current I_D (A)	523	E_{sw} (J)	0.01663
I_{DM} (A)	1280	Q_{rr} (nC)	330
I_{SC} (A)	4000	$R_{th_{JS}}$ (K/W)	0.075
$R_{DS_{ON_{25}}}$ (Ω)	0.0056	Case temperature($^{\circ}$ C) T_c	125
$R_{DS_{ON_{Max}}}$ (Ω)	0.0095	Maximum temperature($^{\circ}$ C) T_j	150
V_F (V)	1.40	Length(mm)	106.4
I_F (V)	100	Width(mm)	64.1
V_{F0} (V)	0.95	Mass(g)	325

3.4.2.2 Inductors

Inductors are electronic components that stores and discharges magnetic energy following the passage of electrical current. This charge and discharge action may be done in a filter or converter mode, depending on the deployment. For the present research, the latter is of more consequence, since the inductor is intended for a DC-DC converter rather than as an AC current ripple filter. The DC-DC converter considered is the inverting buck-boost converter, which can step-up and step-down voltages. The inductance L_{Des} of an inductor for such a converter can be derived using equation (3-27) [98].

$$L_{Des} = \frac{V_{in} D}{\Delta I_{pp} f} \quad (3-27)$$

Where V_{in} is the converter input voltage and f is the switching frequency, ΔI_{pp} is the peak-to-peak ripple current. D is the switching duty cycle, obtained with equation (3-28).

$$D = \frac{V_o}{V_o + V_{in}} \quad (3-28)$$

Where V_o is the converter output voltage.

After deriving the required inductance for the inductor, the physical dimensions and mass need to be obtained. Previous work[23] on inductor sizing for

aerospace devices, performed the sizing based on energy capacity of the inductor. A mass density of 3kg/J was used to estimate the mass, while a value of 0.2mΩ was assumed for the direct current resistance (DCR) in derivation of the power loss. These did not reflect the changing size of the inductor. Also, it is a requirement of the current study to obtain a physical dimension of the inductor, in order to investigate the thermal dissipation to the environment. Hence, an analytical derivation of the dimension, mass and DCR is done, following a correlation of actual inductors and the constituent elements.

Inductors come in a wide variety of types. However, for high power application, a toroidal inductor with a ferromagnetic core shows the best performance[99]. The ferromagnetic core helps to increase the inductance density.

Figure 22 depicts the dimensional representation of the toroidal inductor considered.

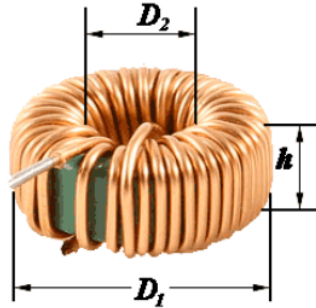


Figure 22 – Inductor Dimensional Representation[100]

Equation (3-29) is used to compute the inductance L_{ind} of an inductor of given dimension, comprising a ferromagnetic core and coils.

$$L_{ind} = \frac{N^2 A_{cr} \mu_{cr}}{l_{cr}} \quad (3-29)$$

Where N is the number of coil turns, A_{cr} is the cross sectional area of the magnetic core, μ_{cr} is the permeability of the core and l_{cr} is the magnetic path length, obtained as the mid-point circumference of the toroid core. The diameter

of the inductor coil D_{cl} , the number of turns N and the core cross sectional area A_{cr} can be obtained using equation (3-30), (3-31) and (3-32) respectively.

$$D_{cl} = \frac{D_1}{20} \quad (3-30)$$

$$N = \frac{\pi D_2}{D_{cl}} \times wf \quad (3-31)$$

$$A_{cr} = \frac{D_1 - D_2}{h} \quad (3-32)$$

Where D_2 and D_1 are the inner and outer diameter of the magnetic core respectively, while h represents the height of the core. wf is the winding factor representing the concentration of the coil turns on the magnetic core. The 20 in equation (3-30) represents an assumption that the core outer diameter is 20 times the coil diameter. This is done to maintain proportionality between the inductor core and coils sizes. A range of 15 to 20 has been obtained empirically from deployed inductors. The mass m_{ind} of the inductor is obtained with equation (3-33).

$$m_{ind} = m_{cr} + m_{cl} \quad (3-33)$$

Where m_{cr} is the mass of the magnetic core while m_{cl} is the mass of the inductor coil, given by equation (3-34).

$$m_{cl} = \pi \left(\frac{D_{cl}}{2} \right)^2 N A_{cr} cf \rho_{cl} \quad (3-34)$$

cf is the connection factor while ρ_{cl} is the density of the coil material in kg/m^3 .

The efficiency η_{Ind} , power loss and direct current resistance R_{DC} of the inductor are estimated using equations (3-35), (3-36) and (3-37) respectively.

$$\eta_{Ind} = 1 - \frac{P_{L_{Ind}}}{P_{in}} \quad (3-35)$$

$$P_{L_{Ind}} = I_{out}^2 R_{DC} + P_{L_{cr}} \quad (3-36)$$

$$R_{DC} = \frac{\rho_{cl} l_{cl}}{A_{cl}} \quad (3-37)$$

Where P_{in} is the power input of the inductor, $P_{L_{cr}}$ is the power loss of the core, I_{out} is the current flowing out of the inductor, while A_{cl} , l_{cl} and ρ_{cl} are the cross sectional area, length and resistivity of the inductor coil respectively. The voltage drop used for the estimation of the inductor efficiency, considers only the DC resistance of the inductor coil. This is because it is currently deployed in a DC circuit, where it serves as a power source, supplying the load for a portion of the switching period, after being pre-charged in the alternate portion. This is governed by the duty cycle D . However, when the inductor is deployed in an AC circuit, the AC resistance and reactance needs to be considered in estimating the efficiency.

The thermal characteristics of the inductor can be expressed in terms of the instantaneous temperature while delivering power. This is given by equation (3-38).

$$T_{Ind} = T_{Amb} + P_{L_{Ind}} R_{th_{ind}} \left(1 - e^{\left(\frac{-t}{R_{th_{ind}} C_{th_{ind}}} \right)} \right) \quad (3-38)$$

Where, $R_{th_{cl}}$ is the thermal resistance of the inductor coil and $C_{th_{ind}}$ is the thermal capacitance of the inductor. These are given in equation (3-39) and (3-40) respectively.

$$R_{th_ind} = \frac{1}{2\pi k_{cl} N A_{cr} c f} + \frac{\ln\left(\frac{D_1}{D_2}\right)}{2\pi h k_{cr}} \quad (3-39)$$

$$C_{th_Ind} = m_{cl} C_{cl} + m_{cr} C_{cr} \quad (3-40)$$

Where k_{cl} and k_{cr} are the thermal conductivity of the coil and core materials, while C_{cl} and C_{cr} are the specific heat capacity of the coil and core materials. The $m_{cr} C_{cr}$ term is computed as the sum of the mass and specific heat of the constituent elements, according to their percentage contribution to the total inductor core mass.

Since the aim is to arrive at an optimal design in terms of mass, thermal capability and efficiency, it is logical to aim for high per-unit current handling capabilities, as this will reduce or eliminate the required parallel interconnections. This is based on RLC circuit principles. By examination of equations (3-27) and (3-29), it is observable that the current capability can be improved through the permeability reduction. An investigation is thus performed, for different combinations of inductor core dimensions and permeability. For each combination, a current handling capability of the inductor is derived. Molypermalloy powder (MPP) and high flux cores with a copper coil are considered, based on their minimal losses compared to other materials[101]. A tabular list of the considered inductor core samples is presented in Appendix B. Figure 23 shows a variation in mass and temperature of all 24 samples, which have been designated 0 – 23. The input data for the investigation were constant design voltage, current and inductance of 1000V, 2000A and 5mH respectively. These have been held constant to enable a tailored investigation.

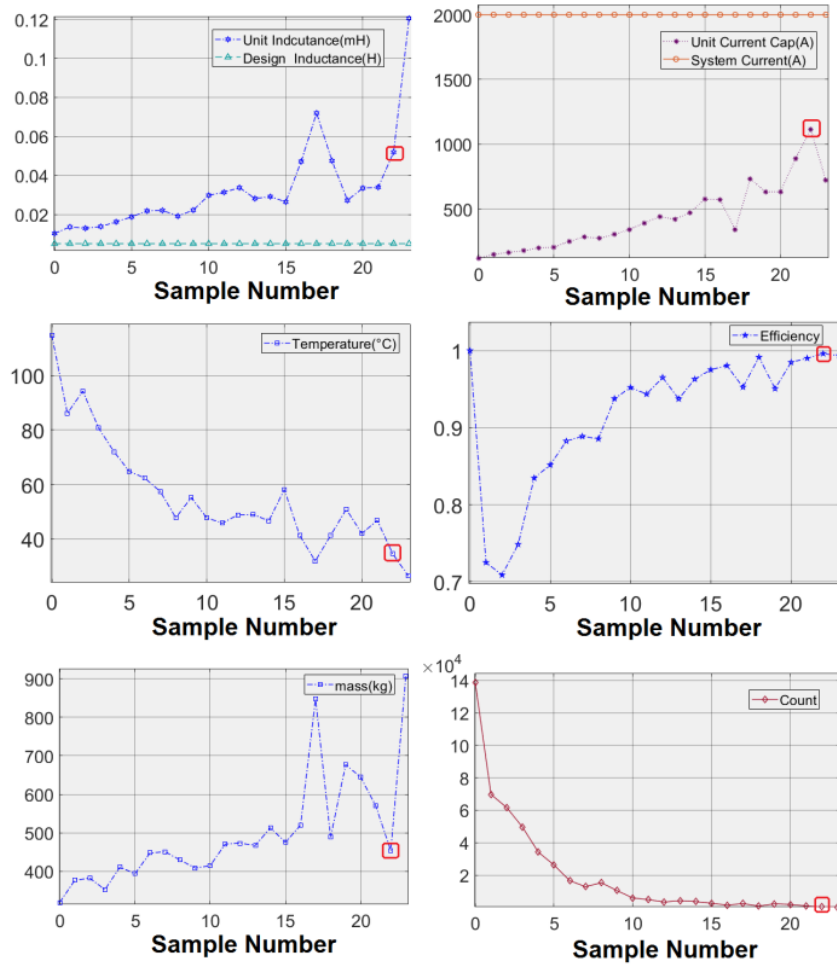


Figure 23 – Inductor Current Capability and Mass variation with Permeability

From the figure, it could be seen that *sample 22* (red boxes) exhibited the best per-unit current handling capability. Although it doesn't offer the best total mass value, it does offer a low count compared to *sample 0*, this means easier installation and lesser interconnection losses. Figure 24 presents the variation in the performance of *sample 22* with changing design current rating.

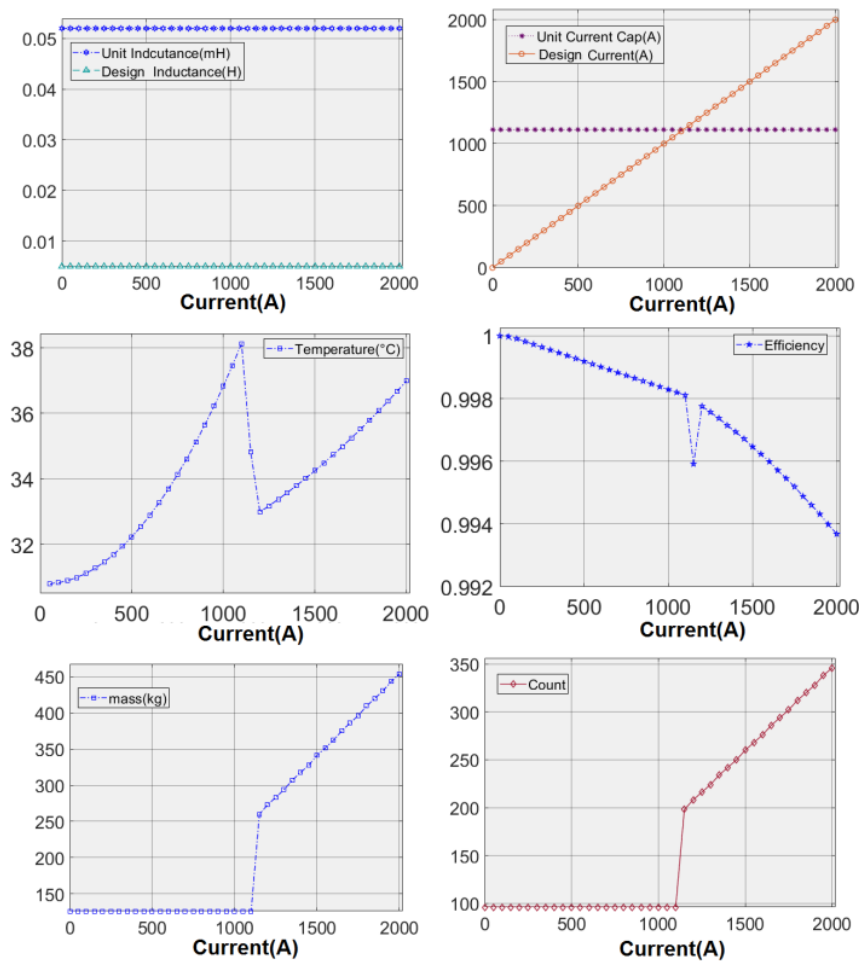


Figure 24 – Inductor Mass Variation with Current

There is a sharp increase in total inductor mass when the per-unit inductor current capability is less than the rated design value. This is a result of reduction in inductance as several inductors are paralleled to accommodate the excess current. The performed investigation shows that unit inductor cores with least permeability and highest current capability offer more mass optimization for high current application. Also worthy of note is that the investigated inductor cores' voltage handling capability is limited to 2kV. Table 5 shows the design values for the high flux inductor core adopted for this research. It is typically made of 50% Nickel and 50% Iron[102,103].

Table 5 – List of Optimal Inductor Core Parameters

Core type	High Flux
Permeability (μ_r)	14
Watt Loss @ 100 kHz, 50mT max (mW/cm ³)	965
Voltage Breakdown wire to wire min (VAC)	2000
Mass (g)	1300
Outer diameter D_1 (mm)	134
Inner diameter D_2 (mm)	77.2
Height h (mm)	26.8
Volume V_e (mm ³)	220000
Magnetic path length l_{cr} (mm)	324
Cross sectional area A_{cr} (mm ²)	678

3.4.2.3 Capacitors

Capacitors perform voltage filtering for electrical circuits by storing and releasing charge in a continuous manner. This section presents an approach to size capacitors and derive their instantaneous temperature while in operation. Similar procedure to that used for the previous components is used.

The capacitance required to filter a given voltage ripple value is derived using equation (3-41) [26].

$$C = \frac{I_{DC}}{2f_{AC}\Delta V_{DC}} \quad (3-41)$$

Where I_{DC} is the DC current, ΔV_{DC} is the DC voltage ripple and f_{AC} is the supply frequency. The capacitor's temperature is given by equation (3-42).

$$T_C = T_{Amb} + P_{L_C} R_{th_C} \left(1 - e^{\left(\frac{-t}{R_{th_C} C_{th_C}} \right)} \right) \quad (3-42)$$

The power loss P_{L_C} is computed as a function of the thermal power loss P_{th} and dielectric power loss P_d given in equations (3-43) and (3-44) [104] respectively.

$$P_{th} = I^2 \times ESR \quad (3-43)$$

$$P_d = \left(\frac{CV_{peak}^2 f}{2} \right) \times \left(\frac{ESR}{2\pi fC + 2\pi fL} \right) \quad (3-44)$$

The total power loss is given by equation (3-45).

$$P_{L_c} = P_d + P_{th} \quad (3-45)$$

The thermal resistance R_{th_c} and equivalent series resistance ESR used in this work have been arrived at following an investigation of 385 capacitors. The choice was based on power capacity, specific power, efficiency, thermal response and physical dimension. However, a major challenge is in the derivation of the thermal capacitance. This is because of the difficulty in deriving the proportions of the constituent materials. Therefore, the thermal resistance is used, since they could be obtained from published datasheets, along with equal proportions of the constituent materials. Table 6 shows the operational parameters of the selected capacitor at its rated power.

Table 6 – Selected Capacitor's Specifications

Type	Film Capacitor
Capacitance(μ F)	560
Voltage DC(V)	700
Current(A)	80
Maximum peak current(A)	8000
ESR($m\Omega$)	1.1
Rth($^{\circ}$ C/W)	2.9
Mass(g)	880
Max Temperature ($^{\circ}$ C)	85
Operating Life(hrs)	200000
Diameter(mm)	116
Height(mm)	70

3.4.3 Transmission Thermal Modelling

The design and sizing of transmission systems is performed with two factors as boundaries: maximum current and temperature handling capability. Sizing is performed based on maximum temperature because a limit could be set as target. Using the short circuit current handling capability as boundary is tricky,

since the actual size of the cable determines the magnitude of the short circuit current, through its inductance. Equation (3-46) has been used to obtain the minimum radius r_c of the conductor to meet this requirement.

$$r_c = \sqrt{\frac{P_{Rated}(1 - \eta_T)}{(T_{max} - T_{Amb})\pi\rho_C C_{p_C}}} \quad (3-46)$$

Where P_{Rated} is the system rated power, T_{max} is the maximum allowable transmission cable temperature, η_T is the efficiency of the transmission cable, ρ_C and C_{p_C} are the density and specific heat capacity of the transmission core material. With known core radius, the mass of the transmission cable of a given length, can be derived using equation (3-47).

$$m_T = v_C\rho_C + v_{ins}\rho_{ins} + v_{sh}\rho_{sh} \quad (3-47)$$

Where v_C and ρ_C are the volume and density of the conductor respectively, v_{ins} and ρ_{ins} are those of the insulation material, while v_{sh} and ρ_{sh} are those of the sheath material. A cross-linked polyethylene (XLPE) insulation and a polyvinyl chloride (PVC) sheath is considered. Also, no consideration is given to armor protection since the considered application does not involve underground installation or similar deployments that expose the cables to cuts. Figure 25 shows the composition and concentric layers of cable considered in this research.

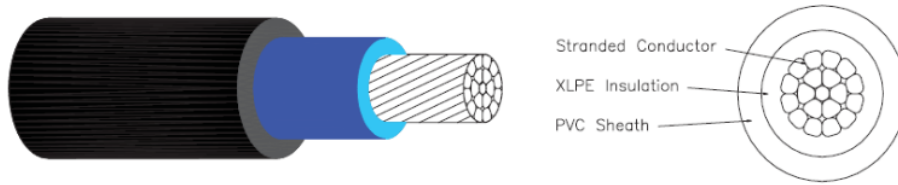


Figure 25 – Cross Sectional View of Considered Cable [105]

A single core configuration has been selected to enable the separation distance to be selected in such a way that voltage breakdown is avoided. Hence, for the three phase AC segment, three of such cable would be installed while two will be installed for a DC transmission segment.

The temperature of the conductor T_T is given as a function of its total mass and the transmission power loss as shown in equation (3-48) according to [106].

$$T_T = T_{Amb} + \frac{P_{Loss_T}(1 - e^{\frac{-t}{RC}})}{m_T C_T} \quad (3-48)$$

Where C_T is the specific heat of the transmission cable and P_{Loss_T} is the transmission power loss, given by equation (3-49).

$$P_{Loss_T} = I_T V_T - I_T (V_R - I_T R_T - X_T I_T) \quad (3-49)$$

Where V_T is the transmission voltage, R_T the resistance of the transmission line and X_T is the inductive reactance of the transmission line. The resistance and reactance of the transmission line are obtained using equations (3-50) and (3-54) according to [107].

$$R_T = R_{DC}(1 + \delta_s + \delta_p) \quad (3-50)$$

Where R_{DC} given by equation (3-51), is the DC resistance, δ_s is the skin effect and δ_p is the proximity effect.

$$R_{DC} = \left(\frac{L_T \times \rho_{20}}{A_c} \right) [1 + \alpha_{20}(T_T - 20)] \quad (3-51)$$

ρ_{20} and α_{20} are the resistivity and temperature coefficient of the conductor at 20°C respectively. The skin effect is given by equation (3-52).

$$\delta_s = \left(\frac{\left(\frac{8\pi f}{R_{DC}} \times 10^{-7} \right)^2}{192 + 0.8 \left(\frac{8\pi f}{R_{DC}} \times 10^{-7} \right)^2} \right) \quad (3-52)$$

While the proximity effect is given by equation (3-53).

$$\delta_p = \delta_s \times \left(\frac{d_c}{s}\right)^2 \left[0.312 \left(\frac{d_c}{s}\right)^2 + \left(\frac{1.18}{\delta_s + 0.27}\right) \right] \quad (3-53)$$

Where, f is the transmission frequency, d_c is the conductor diameter and s is the distance between the conductor centers.

The inductive reactance X_T of the transmission line is given according to [108] by equation (3-54).

$$X_T = 2\pi f \left(0.2L_T \times \left[\ln\left(\frac{2L_T}{r_c}\right) - 1 \right] \right) \quad (3-54)$$

Where the bracket term is the inductance of the cable and r_c is the conductor radius.

It is important to state that for a DC transmission line, the frequency related terms δ_s , δ_p and X_T results to zero and are omitted, since DC voltage does not have an alternating component. Hence R_T will be equal to R_{DC} .

R and C in equation (3-48) can be derived using equations (3-17) and (3-18) respectively. However, only three layers are involved here, r_n , r_{i+1} and r_i . These would be mapped respectively to the conductor, insulation and sheath radii.

The insulation and sheath thickness can be obtained using the trendline equation in Figure 26, obtained from [105].

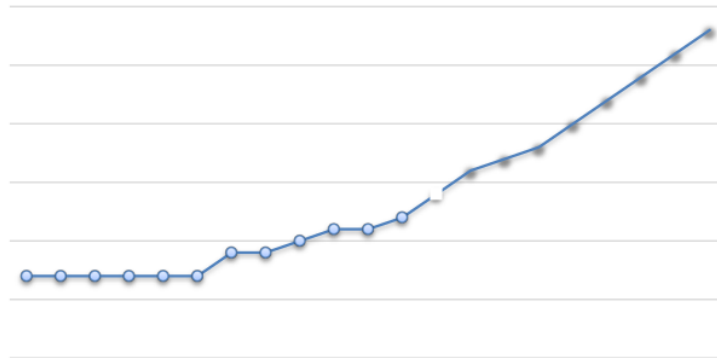


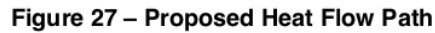
Figure 26 – Relationship between Conductor Diameter and Insulation Thickness

It has been suggested in previous work [23] that a propulsion system utilizing aluminium transmission is better on mass optimization, this was also highlighted in [61]. Following these, an aluminium transmission core material is adopted. Also, a three wire AC transmission is adopted.

3.5 Novel Power Electronics Thermal Management Technique

The thermal management technique presented in this section, is anticipated to greatly reduce the mass and complexity penalty, incurred by the inclusion of the thermal management system for electric propulsion systems. Focus is on the power electronics, since the SOA electrical machines for electric propulsion are proposed to be self-cooled. Also, the deployment of conventional rather than superconducting transmission lines, imply that no special cooling is required for the transmission lines. Various thermal management systems have been investigated and compared in [9], with the conclusion that these all require auxiliary power and accessory for which little mass estimation was performed. Also, the cooling sizing presented in [23], was shown to have a significant mass

This section presents a proposal to minimize the mass of the cooling system by the elimination of accessory components like refrigeration devices, coolants and coolant tank. This is intended to be achieved by an optimal sizing of heatsinks and a strategic positioning of the power electronic components, thereby implementing a passive cooling arrangement using ambient air. Hence, the heat flow path would be as shown in Figure 27.



3.5.1 Feasibility Analysis

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Figure 28 – Assumed Power Electronics Layout

An extension of this layout is intended to a larger power system, for the heat exchange between power electronics and the ambient. This would be achieved by the introduction of fins, protruding from beneath the heat sink into the propulsor duct. This will enable the airflow to aid the convection of heat from the electrical system. Figure 29 is an adaptation of a ducted fan from [11] to the proposed concept.

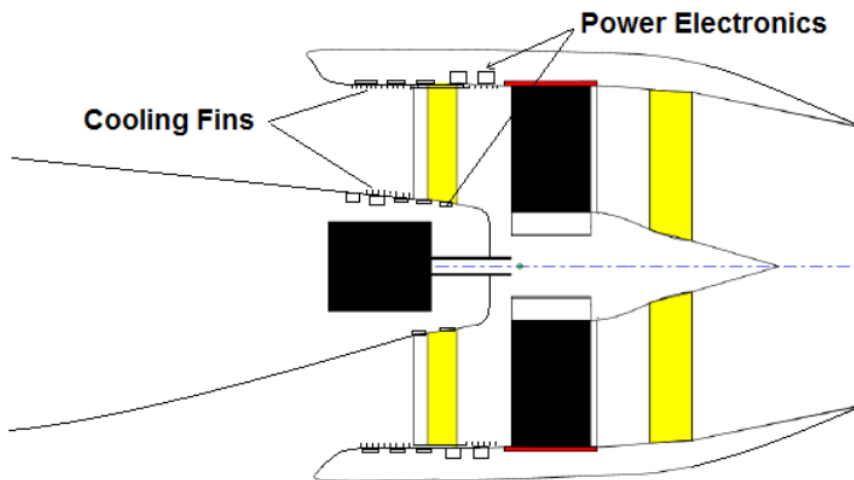


Figure 29 – Ducted fan equipped with the propose layout

The same mass flow of air through the duct is proposed to be utilized for thrust generation and cooling. This appears viable, since the machines requiring the power electronics would be deployed axially with the air flow direction. The power electronics are best positioned closest to these machines to minimize AC losses. Also, the ram air is usually cold and at forced velocity, requiring little or no additional power to effect heat exchange. Part of the nacelle mass can also be

substituted for the heat exchangers (heatsinks), thereby taking advantage of the duct body for heat spreading. This is an improvement of the nacelle cooling method presented in [47], with the main difference here being that the cooling air is not ingested into the aircraft or nacelle body, but only flows through the duct from inlet to exhaust. The mass penalty of the cooling system was also not presented. Another heat exchanger based thermal management system was presented in [39] which resulted in a 20% weight ratio of the cooled device.

3.5.2 Design Assumptions

The following are the assumptions made for the heat sink design:

1. The airflow particles are homogenous.
2. The heat flow is unidimensional from source to sink to air.
3. Heat flows equally from all parts of a fin, implying a constant heat transfer coefficient across the heat sink.
4. The heatsinks are assumed to be pin finned.

These represent common assumptions made for a preliminary design of a forced-air cooling system as stated in [84]. Pin finned heatsinks have been selected, since they have lower thermal resistance than plate finned ones, in applications where the air pumping power per unit area is expected to be high[112]. They also possess better heat transfer coefficient[113]. A study conducted in [114] for seven fin geometries revealed that of all pin finned geometries, the circular geometry had the best performance for applications with high airflow pumping power. Also in [115], it was concluded that a staggered arrangement of the fins has better air performance than an inline arrangement, since it offered better intermixing of the cooling fluid.

3.5.3 Heat Sink Sizing

The sizing aim is to achieve heat removal, at a rate that keeps the components running within their temperature limits, with minimal mass and volume incurred. Two factors are important to consider in heat sink sizing: rate of heat generation and the presence or absence and performance of auxiliary heat removal devices. It is a requirement that the rate of heat evacuation should match the rate of heat

generation. For the intended deployment, the mass flow rate of the ducted fan performs the auxiliary heat removal. Therefore, some characteristics of the flow are incorporated in the heat sink sizing. The measure of the design fidelity is an optimal mass.

As an initial step, two parameters are imposed for the heat sink: A maximum power loss from the cooled device and a minimum length, to ensure the sized heat sink can accommodate the device to be cooled. By using the thermal resistance and power capacity data obtained for a given device, a minimum efficiency can be selected, through which a maximum power loss can be obtained. Equation (3-55) gives the instantaneous temperature of a device resulting from a maximum power loss P_L .

$$T_J = T_{Amb} + P_L R_{thJA} \left(1 - e^{\frac{-t}{RC}} \right) \quad (3-55)$$

It is intended that the temperature obtained from equation (3-55) is never exceeded. The evolution of heat in a device happens thus: The electrical resistance of the device impedes flow of electricity leading to power loss P_L which manifests as heat. This heat is dissipated to the environment through a *junction-to-ambient thermal resistance* R_{thJA} impeding its flow. This raises the device's temperature from the ambient temperature T_{Amb} to T_J . The heat capacity of the device governs the rate of temperature rise expressed by $1 - e^{\frac{-t}{RC}}$. The initially generated heat increases the electrical resistance leading to more heat. In order to keep the device temperature T_J lower and within safe limits, R_{thJA} could be replaced with a much lower value specified for a device. This is the *junction-to-sink thermal resistance* R_{thJS} and is more often the published value. However, in order to utilize R_{thJS} , a heat sink needs to be directly attached to the device. This introduces a thermal resistance R_{HS} which when combined with R_{thJS} should still be significantly lesser than R_{thJA} . Equation (3-56) expresses the new instantaneous temperature resulting from the addition of a heat sink.

$$T_J = T_{Amb} + P_L(R_{thJS} + R_{HS}) \left(1 - e^{\frac{-t}{RC}}\right) \quad (3-56)$$

Heat sinks can maintain a lower temperature at the contact with the cooled device, by exploring the possibility to lower their thermal resistance using larger convection area and high velocity air assisted evacuation. In order to determine the thermal resistance R_{HS} , which when put in equation (3-56) yields an acceptable value of T_J , the maximum allowable temperature of the heat sink T_{HS} needs to be known. This can be obtained by considering the maximum allowable values of the device case temperature T_C relative to that of the internal junction T_J . These are presented for the switch on Table 4 as 125°C and 150°C respectively. It means that at the instance when the device case is 125°C, the internal temperature will be 150°C. Therefore, it is desirable to keep T_{HS} at 125°C, to enable it to hold the device from exceeding its limit of 150°C. These values may vary for different devices and attention needs to be paid to these variations.

By substituting T_{HS} in equation (3-57), the value of R_{HS} can be found

$$T_{HS} = T_{Amb} + P_L R_{HS} \quad (3-57)$$

The obtained thermal resistance R_{HS} corresponds to a heat sink which based on the assumptions in 3.5.2 can be notionally represented as in Figure 30.

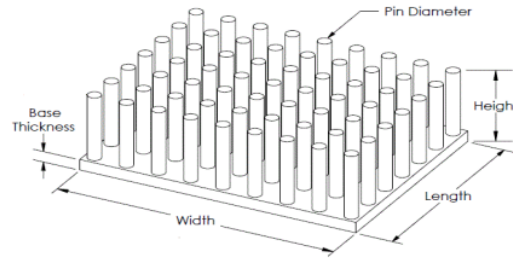


Figure 30 – Sample Heatsink Structure and Dimension[116]

The thermal resistance of such a heat sink can be derived with equation(3-58) [112].

$$R_{HS} = \left(\frac{1}{h(A_b + \eta A_f)} \right) + \left(\frac{1}{\rho_A C_{p,A} \dot{V}} \right) \quad (3-58)$$

Where h is the heat transfer coefficient, A_b is the exposed surface area of the heat sink base, which is not occupied by fins. η is the efficiency of the fins, A_f is the total surface area of the fins, ρ_A is the density of the ambient air, $C_{p,A}$ is the specific heat capacity of the air and $\dot{V}$ is the volumetric flow rate. Equation (3-59)[112] gives the heat transfer coefficient h .

$$h = \frac{k_A Nu}{D_f} \quad (3-59)$$

Where k_A is the thermal conductivity of air, Nu is the Nusselt number of the flow and D_f is the diameter of the heatsink fin, assumed to be the hydraulic diameter as in [112]. The Nusselt number is obtained by equation (3-60) [112].

$$Nu = \frac{0.142 Re^{0.561}}{0.35 + \left(\frac{w_f}{D_f} \right)^{-0.65}} \quad (3-60)$$

Where w_f is the fin pitch and Re is the Reynolds number given by equation (3-61).

$$Re = \frac{\rho_A v D_f}{\mu_A} \quad (3-61)$$

Where v and μ_A are the velocity and viscosity of the cooling air respectively. The efficiency of the heat sink fins can be derived using equation (3-62)[85].

$$\eta = \frac{\tanh m H_f}{m H_f} \quad (3-62)$$

Where H_f is the height of the heat sink fins and m is given by equation (3-63).

$$m = \sqrt{\frac{hP_f}{k_{HS}D_f}} \quad (3-63)$$

Where P_f is the perimeter of the fin and k_{HS} is the thermal conductivity of the heat sink material. The volumetric flow rate $\dot{V}$ is given by equation (3-64).

$$\dot{V} = H_f L v \quad (3-64)$$

Where L is the length of the heat sink.

An iterative procedure is followed to obtain optimized values of A_b , A_f , D_f , w_f , H_f and L . This is shown by the flow chart in Figure 31. The clearance factor represents a measure of the distance between the component edges and heat sink edges. The iteration is repeated for the full mission duration, until the point where best mass keeps the components' temperature below their maximum limit.

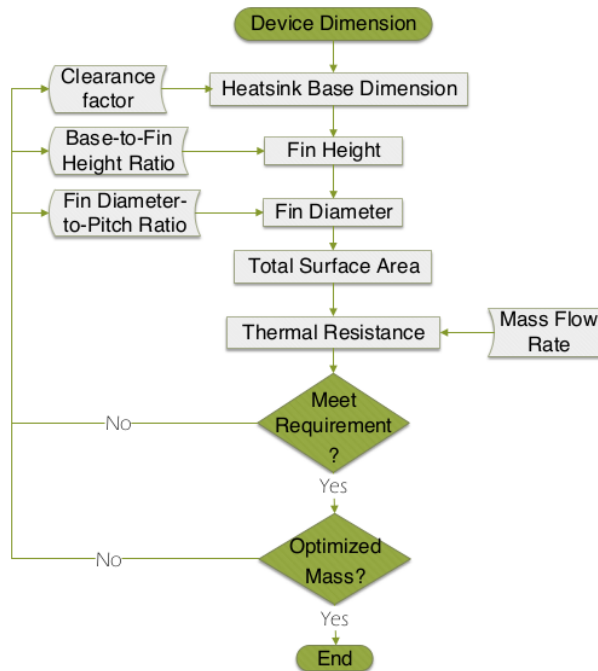


Figure 31 – Flow Chart for Heat sink Sizing

The time constant RC in equation (3-56) can be obtained using equation (3-65).

$$RC = (R_{th_{JS}} + R_{th_{HS}}) \times C_{th_{CG}} \quad (3-65)$$

As mentioned in 3.4.2.1, $C_{th_{CG}}$ represents the thermal capacitance of the cooling group, which in this case consists of the heat sink and mass flow of air. This can be obtained using equation (3-66).

$$C_{th_{CG}} = m_{HS}C_{p_{HS}} + m_A C_{p_A} \quad (3-66)$$

m_{HS} and m_A are the masses of the heat sink and air given by equation (3-67) and (3-68) respectively.

$$m_{HS} = \left(LWt_b + LWH_f \left(1 - \frac{w_f - D_f}{w_f} \right) \right) \rho_{HS} C_{p_{HS}} \quad (3-67)$$

$$m_A = LWH_f \left(\frac{w_f - D_f}{w_f} \right) \rho_A C_{p_A} \quad (3-68)$$

Where W is the width of the heat sink, t_b is the thickness of the heatsink base, ρ_{HS} is the density of the heatsink material and $C_{p_{HS}}$ is the specific heat capacity of the heat sink material.

3.6 Summary

A procedure to estimate the fault current for a turboelectric propulsion system was presented in this chapter. The presented approach focused on both the estimation of the maximum current values and the derivation of the elementary circuit features which were a prerequisite for the estimation of these faults. By examination of equations (3-1) and (3-3), it could be seen that higher line impedances would yield lower fault currents. Thermal characteristics of typical propulsion system devices have also been presented. The temperature of the

device is greatly dependent on the thermal resistance of the device hot spot and the overall thermal capacitance of the device and cooling group. These form the time constants, which gives insight about how fast device temperatures would rise in reaction to applied currents and losses. The interrelationship between fault and thermal characteristics is summarized in Figure 32.

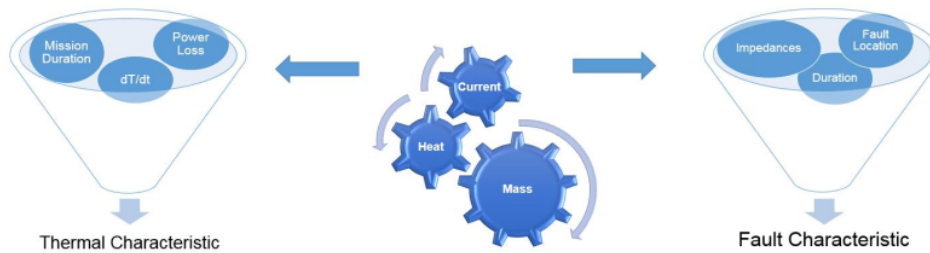


Figure 32 – Thermal and Fault interconnection

A novel cooling approach was presented in the concluding part of this chapter. This was an improvement on earlier presented works. The presented idea intends to limit the cooling needs of the propulsion system to just heat exchangers, with minimal application of forced air cooling. The heat exchangers can replace part of the duct or be implemented as fins which simultaneously serve as the duct stator.



propulsor size and power. Next, the motor sizing and dynamic performance, then other components sizing follows.

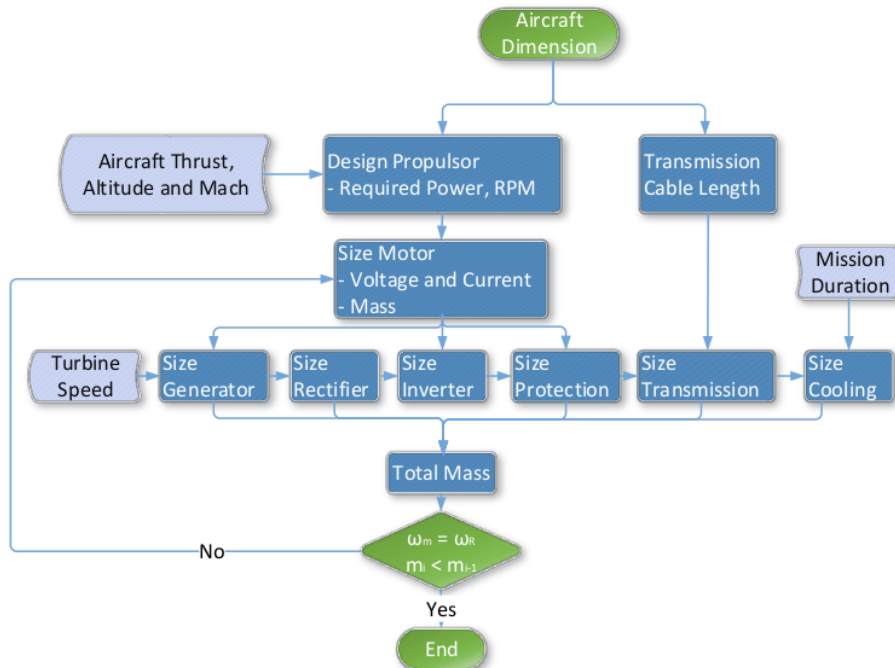


Figure 33 – Flow Chart of Motor Optimized Modelling Procedure

4.1.1 Ducted Fan Modelling

The modelling of the ducted fan is performed to obtain the required power, RPM, diameter and mass of the fan to deliver the needed thrust. This will enable enough evaluation of the designed electrical network, especially with respect to transient and abnormal operation scenarios. The derived power and RPM become the rating of the electrical motor to drive the fan. This effects the aerodynamic and electrical coupling, save for the inclusion of a gearbox.

4.1.1.1 RPM and Power

This section describes the procedure used for obtaining the power and RPM of the fan. Figure 34 is the flow chart representation of the procedure.

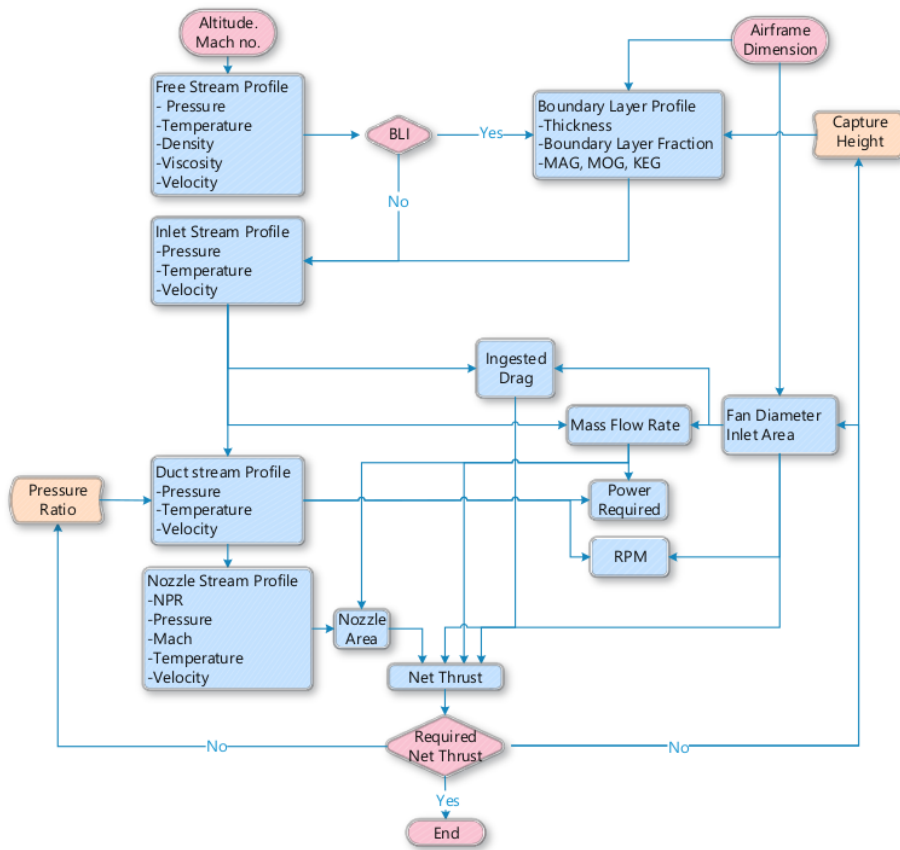


Figure 34 – Power and RPM Derivation Approach

The procedure presented is actualized by considering the thermodynamic and aerodynamic characteristics of the air through the duct in five typical stages; freestream, inlet, duct stream (before and after the fan), and nozzle. The net thrust F_n produced by the ducted fan is expressed by equation (4-1)[117].

$$F_n = \dot{m}(V_8 - V_0) + A_8(p_8 - p_0) \quad (4-1)$$

Where $\dot{m}$ is the mass flow rate of the air into the duct, V_0 and V_8 are the freestream and exit velocities of the air, A_8 is the area of the nozzle while p_0 and p_8 are the static pressures of the freestream air and nozzle respectively. The mass flow rate is given by equation (4-2).

$$\dot{m} = A\rho_0V_0 \quad (4-2)$$

Where A is the cross-sectional area of the duct inlet and ρ_0 is the density of the freestream air. The nozzle exit velocity V_8 can be derived using equation (4-3).

$$V_8 = \sqrt{2C_p(T_8 - t_8)} \quad (4-3)$$

Where C_p is the isobaric specific heat capacity of air, while T_8 and t_8 are the total and static temperatures of air at the duct nozzle. T_8 is assumed to be same as that right behind the fan T_3 which is given by equation (4-4)[21].

$$T_3 = PR_f^{\left(\frac{\gamma-1}{\gamma\eta_p}\right)} \times T_2 \quad (4-4)$$

Where PR_f is the ratio of the air pressure downstream of the fan to that upstream. A value above 1 is initially assumed and adjusted until the an optimal design is achieved. γ is the ratio of specific heat capacity of air at constant pressure to that at constant volume while η_p is the polytropic efficiency of the fan. T_2 is the total temperature at the face of the fan, assumed to be same as the total temperature at the inlet T_1 . This is given by (4-5) [117].

$$T_1 = t_0 \left[1 + (\gamma - 1) \frac{M^2}{2} \right] \quad (4-5)$$

Where t_0 is the temperature of the freestream air and M is the flight Mach number. The computation of t_8 requires a check for choking at the duct nozzle. The condition presented in equation (4-6) indicates a choked nozzle.

$$\frac{P_8}{p_0} > \left(\frac{\gamma + 1}{2} \right)^{\left(\frac{\gamma}{\gamma-1}\right)} \quad (4-6)$$

Where P_8 is the total pressure at the nozzle given by (4-7).

$$P_8 = P_3 PR_{jp} \quad (4-7)$$

P_3 is the total pressure right behind the fan while PR_{jp} is the jet pipe pressure ratio, assumed in this work as 0.99. P_3 can be derived using equation (4-8).

$$P_3 = P_2 PR_f \quad (4-8)$$

Where P_2 is the total pressure at the face of fan, assumed to be same as the total pressure at the inlet P_1 . This is given by (4-9).

$$P_1 = p_0 \left(\frac{T_1}{t_0} \right)^{\left(\frac{\gamma}{\gamma-1} \right)} \quad (4-9)$$

For unchoked nozzle conditions $p_8 = p_0$ and t_8 is obtained by equation (4-10).

$$t_8 = \frac{T_8}{\left(\frac{P_8}{p_8} \right)^{\left(\frac{\gamma-1}{\gamma} \right)}} \quad (4-10)$$

However, when the nozzle is chocked, t_8 and p_8 can be obtained by equation (4-11) and (4-12) respectively.

$$t_8 = \frac{T_8}{\left(\frac{\gamma+1}{2} \right)} \quad (4-11)$$

$$p_8 = \frac{P_8}{\left(\frac{T_8}{t_{8s}} \right)^{\left(\frac{\gamma}{\gamma-1} \right)}} \quad (4-12)$$

The cross-sectional area of the nozzle A_8 can be obtained with equation (4-13) [117].

$$A_8 = \frac{\dot{m}}{\rho_8 V_8} \quad (4-13)$$

Where ρ_8 is the density of air at the exit given by equation (4-14).

$$\rho_8 = \frac{p_8}{R t_8} \quad (4-14)$$

Where R is the specific gas constant 287.052 [J/ (K * kg)].

The power required to produce the net thrust is obtained by (4-15).

$$P = \dot{m} C_p \Delta T \quad (4-15)$$

Where ΔT represents the difference between the total air temperature upstream and downstream of the fan. This is given by equation (4-16).

$$\Delta T = T_3 - T_2 \quad (4-16)$$

With the change in temperature across the fan derived, it is possible to obtain the fan RPM using equation (4-17).

$$RPM = \frac{\sqrt{\frac{C_p \Delta T}{Load}} \times 60}{2\pi r_f} \quad (4-17)$$

Where r_f is the radius of the fan while $Load$ is the fan stage loading coefficient, assumed to be 0.18. This value has been derived as an average from two fans of 3.4m and 1.96m diameter, operating at 2355RPM and 3894RPM respectively. A more detailed procedure for the derivation of the stage loading coefficient would be required for each considered fan. However, such a procedure is beyond the scope of the present study.

4.1.1.2 Fan Mass

The essence of deriving the fan mass is for the computation of the total inertia load on the motor shaft. Electrical machine responses to controls and transients have been found to be influenced by their inertia load[77].

The fan mass estimation is challenging because of insufficient data availability. An averaging method is presented, where a relationship is drawn between the fan dimensions of assumed engines. A reference fan diameter, blade mass, blade length and count are a requirement for the utilization of the method presented here. These have been obtained from public domain[118,119] for two fans, and a computation of the face cone mass is performed. Thus, yielding a total mass for each fan. A scaling is then performed based on three geometrical relationships: The area ratio, diameter ratio and blade length ratio, to obtain three mass values m_{F1} , m_{F2} and m_{F3} for the new fan with a target diameter. Equations (4-18), (4-19) and (4-20) can be used to derive these masses.

$$m_{F1} = m_{ref} \times \frac{D_F}{D_{ref}} \quad (4-18)$$

$$m_{F2} = m_{ref} \times \frac{A_F}{A_{ref}} \quad (4-19)$$

$$m_{F3} = m_{ref} \times \frac{BL_F}{BL_{ref}} \quad (4-20)$$

Where D_F , A_F and BL_F are respectively the target diameter, area and blade length of the new fan, while m_{ref} , D_{ref} , A_{ref} and BL_{ref} are the known mass, diameter, area and blade length of a reference fan. An average of these three masses is computed as the final value for the new fan. m_{ref} can be derived using equation (4-21).

$$m_{ref} = m_B + m_H + m_C \quad (4-21)$$

Where m_B , m_H and m_C are masses of the reference fan blade (total for all blades), hub disc and face cone respectively. The mass of the hub is derived with equation (4-22).

$$m_H = l_{Bb} \times \left[\pi \left(\frac{D_H}{2} \right)^2 \right] \times \rho_H \times k_g \quad (4-22)$$

D_H is the diameter of the fan hub, obtained using equation (4-23).

$$D_H = D_{ref} - 2l_B \quad (4-23)$$

Where l_{Bb} , l_B , ρ_H and k_g are the blade base length, blade length, density of the hub material and groove factor respectively. The mass of the face cone m_C can be derived using equation (4-24).

$$m_C = (V_C - V_{C_v})\rho_H \quad (4-24)$$

Where V_{C_v} is the volume of the vacuum in the cone and V_C is the face cone volume given by equation (4-25).

$$V_C = \pi \left(\frac{D_H}{2} \right)^2 \times \frac{H_C}{3} \quad (4-25)$$

Where H_C is the height of the cone and is given by equation (4-26).

$$H_C = D_H R_{HD} \quad (4-26)$$

R_{HD} is the ratio of the cone height to cone diameter. V

$$V_{c_v} = \pi \left(\frac{D_H}{2} k_v \right)^2 \times \frac{H_C - \frac{D_H}{2} (1 - k_v)}{3} \quad (4-27)$$

Where k_v is the cone vacuum factor.

Table 7 shows the input design parameters used for the fan mass estimation.

Table 7 – Assumed Parameters for Fan Mass Estimation

	Fan 1	Fan 2
Blade mass(kg)	22.63426	4.2223
Fan Blade Height(m)	1.219	-
Fan Blade Base(m)	0.432	-
Fan Diameter(m)	3.2512	1.76
Fan Hub Diameter(m)	-	0.49
Blade Count	22	18
k_g		0.5
R_{HD}		0.85
k_v		0.9

Fan mass estimations are made for diameters up to 6.2m, using approximations from both fans in Table 7. A trend line of the results from the two referenced fans is presented in Figure 35.

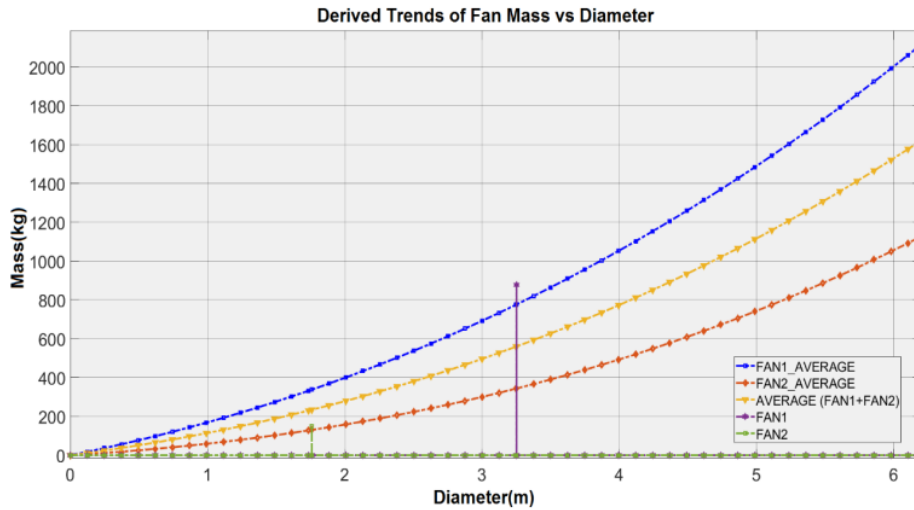
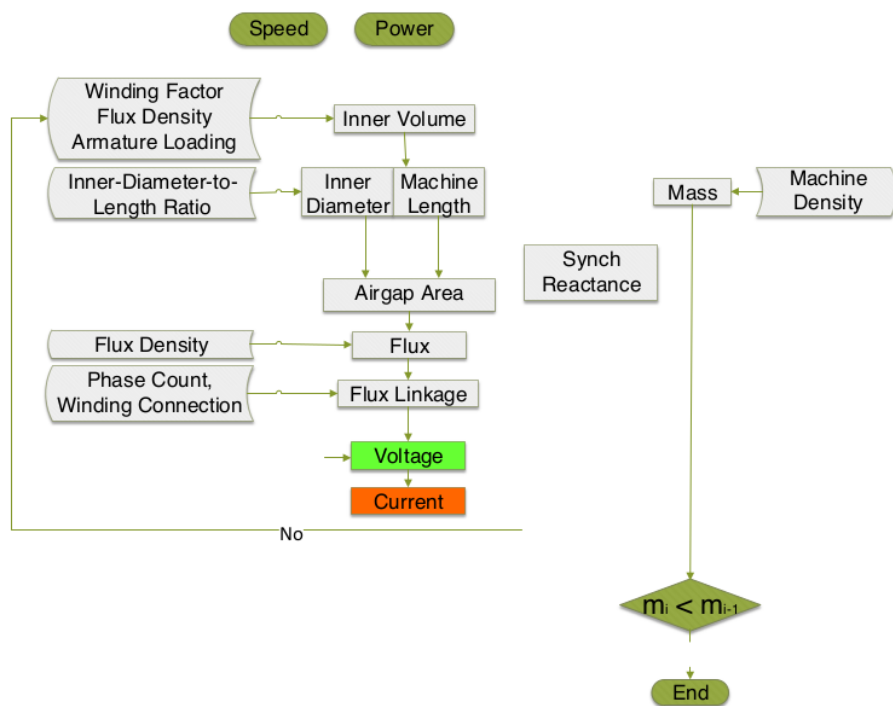


Figure 35 – Derived Trend line of Fan Masses





machine is because of its influence on the flux propagation. The usual aim of optimized motor design, is to fit as much flux required to deliver a given power, in the smallest possible airgap. The inner diameter of the machine D_{in} can be obtained using equation (4-29).

$$D_{in} = \sqrt[3]{\frac{4v_{in}R_{DL}}{\pi}} \quad (4-29)$$

The machine outer diameter D_o is given by equation (4-30).

$$D_o = R_{oi}D_{in} \quad (4-30)$$

The airgap area A_{AG} is given by equation (4-31)[122].

$$A_{AG} = \pi D_{in}L \quad (4-31)$$

The total airgap flux φ and flux linkage per phase λ can be obtained using equation (4-32) and (4-33) respectively.

$$\varphi = B_g \times A_{AG} \quad (4-32)$$

$$\lambda = \frac{\varphi}{N_{ph}k_{ph}} \quad (4-33)$$

Where R_{DL} is the machine inner diameter-to-length ratio, R_{oi} is the machine outer-to-inner diameter ratio, L is the machine length, N_{ph} and k_{ph} are the number of phase and phase connection factor respectively. k_{ph} is 1 for a delta connected winding and $\sqrt{3}$ for a wye connected winding.

The voltage of the motor and hence of the electrical network can be obtained using the induced voltage equation (4-34).

$$V_{Des} = \frac{\omega_e \lambda}{\sqrt{2}} \quad (4-34)$$

Where ω_e is the electrical speed of the machine (rad/s) given by equation (4-35).

$$\omega_e = 2\pi \left(\frac{RPM \times p_s}{120} \right) \quad (4-35)$$

$$I_q = \int \left(\frac{-R_s I_{q,i-1} - \left(\frac{\omega_e p_s}{2} \right) (L_d I_{d,i-1} + \lambda) + V_q}{L_q} \right) dt \quad (4-39)$$

The electromagnetic torque T_e which does the electrical to mechanical power conversion and vice versa, is given by equation (4-40).

$$T_e = \frac{3}{2} \left(\frac{p_s}{2} \right) (\lambda I_q) + (L_d - L_q) (I_q I_d) \quad (4-40)$$

The machine mechanical speed is given by equation (4-41).

$$\omega_m = \int \left(\frac{T_e - T_L - B \omega_{m,i-1}}{J} \right) dt \quad (4-41)$$

The angular displacement of the rotor from a reference point of rest is given by equation (4-42).

$$\theta = \left(\frac{p_s}{2} \right) \int (\omega_m) dt \quad (4-42)$$

Where V_d and V_q are the d and q axis voltages, I_d and I_q are the d and q axis currents, R_s is the stator resistance, L_d and L_q are the d and q axis inductances, T_e is the electromagnetic torque, T_L is the load torque, ω_i is the instantaneous speed, J is the total inertia on the rotor, θ is the angular displacement while B represents the coefficient of viscous damping in the shaft bearings.

4.1.2.2 Parameter Selection

In order to satisfy the condition $\omega_i \geq \omega_{Rated}$ on Figure 36 at the rated torque, it is important to obtain suitable values for the design variables. This section presents the process for selecting the design values used in this work. These parameters are derived or quoted mostly from the SOA machine proposed in [20] for aerospace applications.

4.1.2.2.1 Fundamental Winding Factor and Stator Poles

The winding factor value is useful for improving the EMF generated in a three-phase electrical machine, by the reduction of harmonics. Winding factor reduction are a result of the disparity between the phasor sum and numerical sum of the induced EMF due to the distribution of turns of the same phase in different slots (or positions). By inspection of equation (4-28), it could be seen that a higher winding factor value is important for the mass optimization of electrical machines.

Procedures for winding factor estimation for machine design has been presented previously in [124], based on stator slots to poles relationship. Some of the results of the work are summarized in Figure 37, where certain slot and pole number configurations yield a given winding factor.

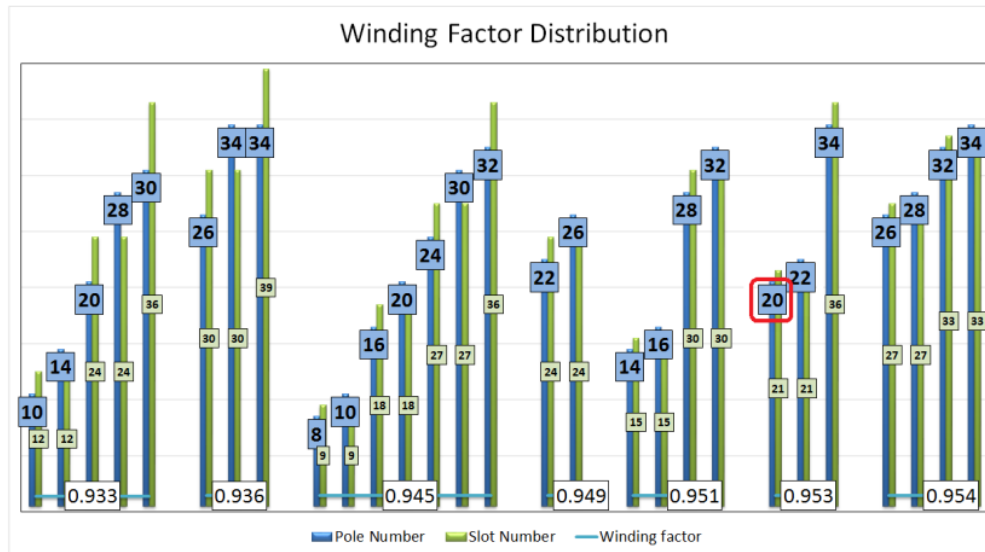


Figure 37 – Variation of winding factor with number of stator poles and slots

In this work, only value of winding factor corresponding to the pole enclosed in the red box of Figure 37 have been considered, since the considered SOA machine has same number of poles.

4.1.2.2.2 Machine Impedances

The derivation of the machine impedance covers the resistance R_s and synchronous reactance X_s from which the d and q axis inductances L_d and L_q are

derived. The inductance L_m of the machine can be obtained using equation (4-43).

$$L_m = \frac{\sin\left(\frac{\phi\pi}{180}\right)\lambda}{I} \quad (4-43)$$

Where ϕ is the phase angle (in degrees). The d and q axis inductances L_q and L_d are obtained with equations (4-44) and (4-45) respectively.

$$L_q = \left(\frac{2}{3}\right)L_m \quad (4-44)$$

$$L_d = \sqrt{L_m^2 - L_q^2} \quad (4-45)$$

The synchronous reactance X_s is given by equation (4-46).

$$X_s = L_m\omega_e \quad (4-46)$$

The resistance R_s can be derived using equation (4-47).

$$R_s = \frac{X_s}{r_{X/R}} \quad (4-47)$$

Where $r_{X/R}$ is the ratio of synchronous reactance to resistance.

The d and q axis voltages V_d and V_q are derived using a d-q transformation procedure in Appendix C with inputs obtained from equations (4-34) and (4-42).

4.1.2.2.3 Machine Inertia

The inertia considered refers to the total inertia of the rotating parts which includes that of the rotor and the fan. The total inertia can be calculated using equation (4-48).

$$J = J_r + J_F \quad (4-48)$$

Where J_r and J_F are the inertia of the machine rotor and the fan respectively.

Table 8 shows the utilized electrical machine design parameters.

Table 8 – Electrical Machine Design Input Parameters

Parameter	Symbol	Unit	Value	Reference
Fundamental winding factor	K_w		0.953	[125]
Airgap flux density	B_g	T	0.93	[20]
Armature loading*	A_L	A/m	48471.61	
Inner diameter-to-length ratio*	R_{DL}		1.132	
Outer-to-inner diameter ratio*	R_{oi}		1.333	
Stator Poles	p_s		20	[20]
Machine Density*	ρ_m	kg/m ³	3854	
Power Factor*	pf		0.997	
Phase angle	$\emptyset$	Deg. Elect	4.2595	[20]
Ratio of synchronous reactance to resistance	$r_{X/R}$		2.516	
* Derived from SOA machine				

A detailed approach for the motor static and dynamic design has been presented. Although a few values have been assumed and deduced from the state-of-the-art machines, this approach ensures that the torque and speed requirement of the propulsor is met, despite the difference between the design specifications and the referenced machines. The specific contribution of this approach is the derivation of the motor voltage and current, which represents the propulsion system specification.

4.1.2.3 Generator Modelling

As already mentioned, a PMSM can also operate as a generator with the application of torque at the shaft rather than voltage at the cable terminals. The generator modelling follows similar procedure, using same equations as the motor modelling. However, for the initially investigated approach, where the motor voltage is imposed on other network components, the static equations are applied in reverse order from (4-35) to (4-28), in order to obtain the volume and then mass of the generator. The inner diameter is selected as the optimization

handle, since the voltage is already known. The *RPM* utilized for obtaining ω_e in this case is the turbine shaft speed driving the generator. Also, due to the reversal of power flow compared to the motor, equation (4-41) is rewritten as equation (4-49).

$$\omega_e = \int \left(\frac{T_e + T_L - B\omega_{e,i-1}}{J} \right) dt \quad (4-49)$$

The inertia derivation for the generator only considers the rotor mass and dimensions. It is also important to state here, that for a motor optimized voltage derivation approach, the motor voltage is retained as part of the generator design inputs, as will be with other components of the propulsion train.

4.1.3 Rectifier Sizing

The rectifier acts to convert the alternating current (AC) from the generator to direct current (DC), to enable it to be controlled. Modern high-power rectifier circuits are often implemented using switches. Figure 38 shows a typical rectifier configuration implemented in MATLAB/Simulink. The In ports 1 – 3 are for the three phases of AC voltage output from the generator. The out ports 4 and 5 are the positive and negative DC terminals respectively. The filter capacitor smoothens the output of the switches to improve the power quality.

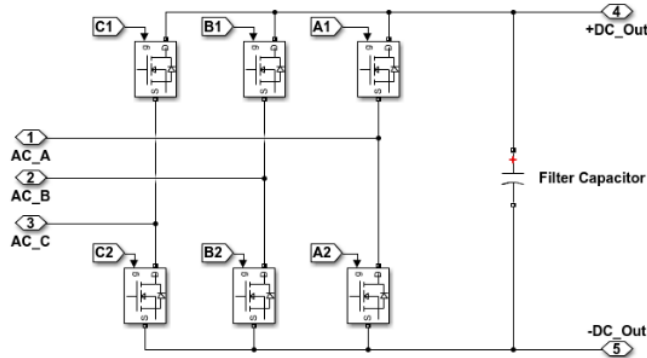


Figure 38- Typical Rectifier circuit configuration.

The rectification is performed based on a present design value (system design voltage), by sending switching control signals to the respective switches using

signalling ports A1 through C2. The details of the control systems are outside the scope of this research. Other supporting components include the heatsinks, other cooling devices and mounting accessories. The total mass of the rectifier is derived using equation (4-50).

$$m_{\text{rect}} = m_{\text{switch}} + m_{\text{cap}} + m_{\text{circuit}} + m_{\text{support}} \quad (4-50)$$

m_{switch} and m_{cap} are the unit masses of the switching device and the capacitor, obtainable from Table 4 and Table 6 respectively. While N_{switch} and N_{cap} are their respective total counts. These are derived using equations (4-51) and (4-52) respectively.

$$N_{\text{switch}} = \frac{V_{\text{sys}}}{V_{\text{switch}}} \cdot \frac{I_{\text{sys}}}{I_{\text{switch}}} \cdot \gamma \quad (4-51)$$

$$N_{\text{cap}} = \frac{V_{\text{sys}}}{V_{\text{cap}}} \cdot \frac{I_{\text{sys}}}{I_{\text{cap}}} \quad (4-52)$$

γ is a multiplier, to account for the mass of the circuit board and support structures, and hence is the switch mass scaling factor. V_{sys} and I_{sys} are the system design voltage and currents respectively, V_{switch} and I_{switch} are respectively the voltage and current handling limit of the switch, while V_{cap} and I_{cap} are respectively the voltage and current handling limit of the capacitor. I_{sys} here refers to the maximum surge current during motor transients.

The efficiency of the rectifier is obtained using equation (4-53).

$$\eta_{\text{rect}} = \frac{\eta_{\text{switch}} + \eta_{\text{cap}}}{2} \quad (4-53)$$

Where η_{switch} and η_{cap} are the efficiency of one switch and capacitor respectively. In normal operation, only 2 switching channels are delivering power at a time, and hence contribute to the instantaneous loss of efficiency of the device. However,

the heat dissipation is estimated for all 6 channels, since the switching interval are so close. Also, when a half bridge switch module is used in place of single switches, then the 6 in equation (4-51) and (4-53) becomes 3.

4.1.4 Inverter Sizing

The inverter performs the speed control of the motor using one of several available techniques. However, common techniques exploit the relationship between current, motor inductance and flux, by means of pulse width modulation (PWM). Figure 39 shows the inverter and the relative functional components. The in ports 1 and 2 connect to the DC source while out ports 1, 2, and 3 are connected to the three phases of the motor. Ports A1 through C3 are the PWM control signals channel.

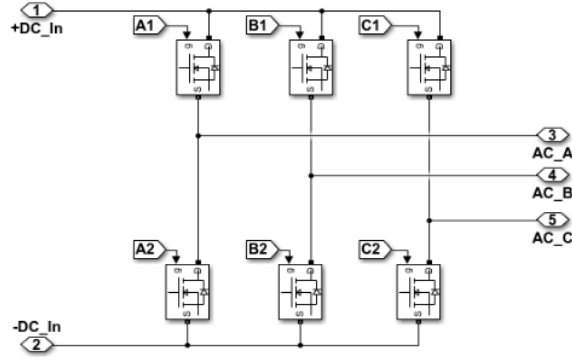


Figure 39 – Typical inverter circuit components

By inspection of Figure 38 and Figure 39 it can be seen that the only physical difference is the presence of the capacitor. Therefore, similar principles are applied for the mass and efficiency estimation. The mass of the inverter is derived by equation (4-54).

$$m_{Inv} = m_S C_{S_Inv} m_k \quad (4-54)$$

The total inverter switch count C_{S_Inv} is given by equation (4-55).

$$C_{S_Inv} = \left[\frac{V_{Des}}{V_{max_S}} \times \frac{I_{Des}}{I_{max_S}} \right] \times 6 \quad (4-55)$$

The efficiency η_{Inv} of the inverter is given by equation (4-56).

$$\eta_{Inv} = \eta_s^{\left(\frac{2C_{S_Inv}}{6}\right)} \quad (4-56)$$

Where C_{S_Inv} is the switch count comprising the inverter while η_s is the efficiency of a single switch and m_k is the switch mass scaling factor.

The I_{Des} here refers to either the maximum surge current for the motor transient or the maximum fault current between the motor and the inverter. Whichever gives the higher number of switches is used. This is because the inverter drive is proposed to perform both the motor speed control and fault protection functions.

4.1.5 Protection Device Sizing

The protection device helps to isolate its principal device from faults occurring upstream or downstream of it. This can be achieved by limiting the magnitude of the current passing through, or by completely blocking off the flow, resulting in device unavailability. Two types of protection devices are considered in this section: the AC circuit and DC circuit protection devices. Figure 40 shows the development and handling of a fault current.

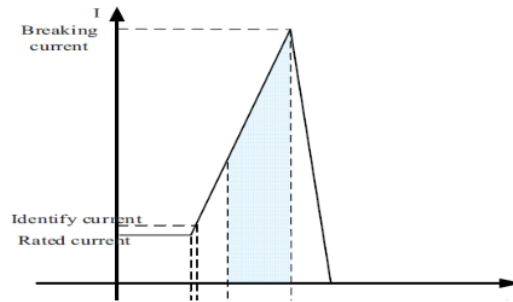
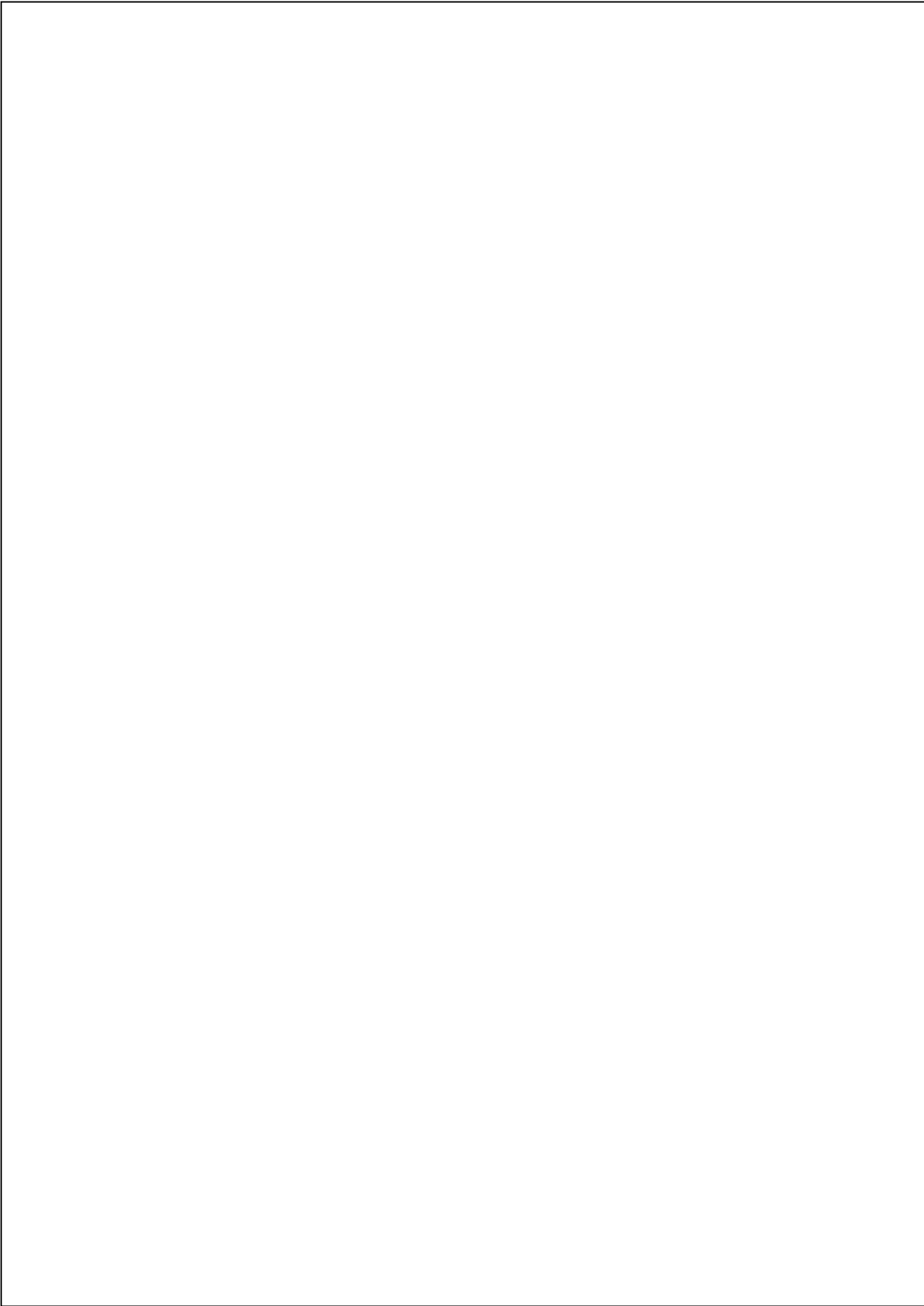


Figure 40 – Fault Current Evolution and Resolution Timing[82]



Based on review performed in [127] for SOA and future protection devices, the entire circuit protection/isolation process can occur in about $800\mu s$ or less. It is shown in [128,129] that fault handling is feasible using converters. However, the application of this strategy was only demonstrated for a converter operating in inverter mode. With rectifiers, the switch incorporated diodes would conduct the fault currents which has pre charged a capacitor as shown in [80]. The mass of the varistor required for each switch is dependent on the magnitude of normal and fault currents and voltages that would attempt to flow through it.

The modelling of the protection system is performed by the addition of these fault handling components to the designated protection zones. The thyristors in Figure 41 are replaced with SiC MOSFETs. Varistors are also added, to perform the energy absorption. Figure 42 shows a sample of the varistor considered in this work. The rating used for this research has been obtained from [130], with a 1465V, 500A continuous and 70kA, 2.9kV surge arresting capability.



Figure 42 – Varistor Sample Considered in this Research

The mass of the protection system m_p deployed per channel at a given point is derived by equation (4-57). The total protection system mass would depend on how many of such devices are deployed. This should vary by aircraft configuration.

$$m_p = m_v C_v + m_s C_s m_k \quad (4-57)$$

Where m_v is the mass of the varistor, m_k is the switch mass scaling factor and C_v is the varistor count, given by equation (4-58).

$$C_v = \left\lceil \frac{V_{Fault}}{V_{max_V}} \times \frac{I_{Fault}}{I_{max_V}} \right\rceil \quad (4-58)$$

Where $V_{\max_V}$ and $I_{\max_V}$ are the varistor surge voltage and current arresting capability respectively. C_S is the switch count comprising the protection device, given by equation (4-59).

$$C_S = \left\lceil \frac{V_{Fault}}{V_{\max_S}} \times \frac{I_{Fault}}{I_{\max_S}} \right\rceil \times 2 \quad (4-59)$$

The I_{Fault} here also considers the maximum surge current or the maximum fault current, whichever gives the higher number, since switches have different current ratings for normal, surge and short circuit operation. These are represented on Table 4 by I_D , I_{DM} and I_{SC} respectively. The RC snubber circuit is not sized separately since it is accounted as part of the mass scaling factor m_k , based on the design presented in [128,129]. Based on same principle, it is seen that the value of fault voltage V_{Fault} can be kept at 1.2 times the rated DC equivalent. The sized protection system is per channel. Therefore, when deployed on a DC line only one is required. However, for a three-phase line, all three channels have one device each, since each channel carries both positive and negative voltages in alternating manner.

4.2 Architecture Optimized Approach

The rationale behind the Architecture optimized approach is to enable devices operate at voltages other than the motor voltage, when having these voltages offer a more optimal configuration. Another reason for applying such an approach is to handle feasibility challenges like combining power sources of different voltages or enable loads which operate at different voltage ratings to be run from the same power source. Figure 43 shows the procedure for the architecture optimized voltage selection. The first difference compared with the motor optimized is the freedom of the generator to run at an optimal voltage adaptable to the turbine shaft speed. Secondly is the introduction of the voltage regulators to support the system flexibility by stepping up/down the voltage to the motor rating. Therefore, there would be a minimum of two voltage ratings for such an architecture: one for the source and another for the loads, with the devices between them enjoying the freedom of operating at any voltage within the range.

This is on the condition that a device's voltage rating is less than or equal to that supplying it.

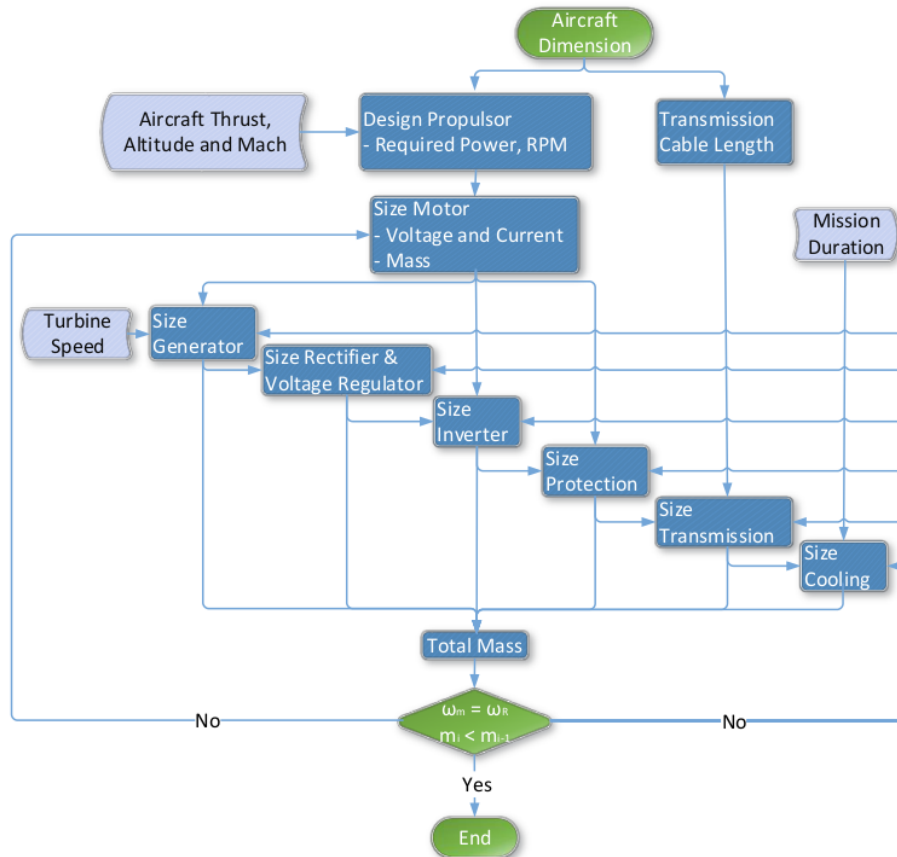


Figure 43 – Flow chart of Architecture Optimized Modelling Procedure

4.2.1 Electrical Machines Sizing

The procedure for sizing both generator and motor are the same as with those presented in section 4.1.2.1. For the generators, the turbine RPM is used, while for the motors the fan RPM is used.

4.2.2 Voltage Regulator Sizing

Voltage regulators serve three main functions; step up, stepdown, or filter of the voltage passing through it. The output of the voltage regulator is the motor's

operating voltage, derived during the motor sizing process. This justifies its presence and is irrespective of the value at its input. In aircraft propulsion systems, a buck-boost voltage regulator is recommended since it meets all the voltage regulation needs, with the fewest components and least incurred mass as shown in [23]. Figure 44 shows an inverting buck-boost converter circuit diagram, comprising the active components. This has one switch branch less than the non-inverting version presented in [23].

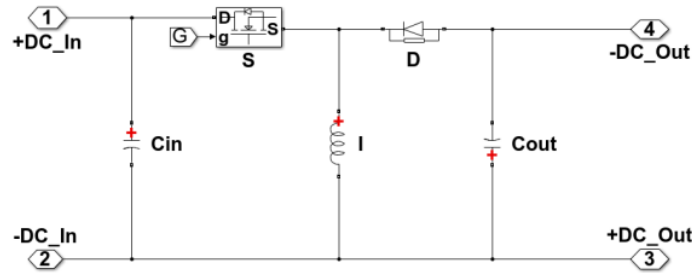


Figure 44- Typical Buck-Boost Converter Circuit Components

S, I, C and D represent the switch, inductor, capacitor and diode respectively. Details of the sizing and efficiency estimation for individual components have been presented in section 3.4.2. However, the capacitance of the input and output capacitors (C_{in} and C_{out}) needs to be derived differently, using equations other than equation (3-41). These are derived using in equations (4-60) to (4-66)[98].

$$C_{in} = \frac{DI_{Ave}}{f \times (V_{in_r_PP} - (ESR \times I_{in_r_PP}))} \quad (4-60)$$

$$D = \frac{V_{out} + V_F}{V_{out} + V_F + V_{in}} \quad (4-61)$$

$$I_{Ave} = \frac{I_{out}}{1 - D} \quad (4-62)$$

$$V_{in_r_PP} = 0.05V_{in} \quad (4-63)$$

$$I_{in_r_PP} = 2(I_{max} - I_{Ave}) \quad (4-64)$$

$$C_{out} = \frac{DI_{out}}{f \times (V_{out_r_PP} - (ESR \times I_{max}))} \quad (4-65)$$

$$V_{out_r_PP} = 0.05V_{out} \quad (4-66)$$

Where D is the switching duty cycle, I_{Ave} is the average of the current handled by the inductor, V_{in} and V_{out} are the regulator input and output voltages respectively. $V_{in_r_PP}$ and $V_{out_r_PP}$ are the input and output voltage ripples respectively. f is the device switching frequency, ESR is the equivalent series resistance of the capacitor obtainable from Table 6. V_{in} , V_{out} , I_{out} and I_{max} are design requirements and should be derived prior to the capacitor sizing. These depend on whether the voltage regulator would be for voltage step up, step down or both. Equations (4-67) can be used to derive the mass of an inverting buck-boost converter for a system of a given rated power.

$$m_{VR} = m_S C_{S_VR} m_k + m_D C_{D_VR} + m_{ind} C_{ind_VR} + m_C (C_{Cin_VR} + C_{Cout_VR}) \quad (4-67)$$

Where m_k is the switch mass scaling factor, while m_S , m_{ind} , m_C and m_D are the masses of the switch, inductor, capacitor and diode respectively. The procedure for obtaining these masses have been covered in 3.4.2, save for the diode which in practice could be replaced by another switch. The voltage regulator switch count C_{S_VR} and diode count C_{D_VR} can be obtained using equation (4-68).

$$C_{S_VR} = \left[\frac{V_{max}}{V_{SD}} \times \frac{I_{max}}{I_{SD}} \right] = C_{D_VR} \quad (4-68)$$

The inductor count C_{ind_VR} can be obtained using equation (4-69).

$$C_{ind_VR} = \left\lceil \frac{L_{Des}}{L_{ind}} \times \frac{I_{max}}{I_{ind}} \right\rceil \quad (4-69)$$

The input capacitor count C_{Cin_VR} can be obtained using equation (4-70).

$$C_{Cin_VR} = \left\lceil \frac{V_{in}}{V_C} \times \frac{I_{in}}{I_C} \right\rceil \quad (4-70)$$

The output capacitor count C_{Cout_VR} can be obtained using equation (4-71).

$$C_{Cout_VR} = \left\lceil \frac{V_{max}}{V_C} \times \frac{I_{max}}{I_C} \right\rceil \quad (4-71)$$

V_{max} and I_{max} represent the respective maximum voltage and currents expected to be handled by the converter. V_{SD} and I_{SD} are the voltage and current capability of the switch and diode, while V_C and I_C are those of the capacitor. L_{ind} represents the inductance of a unit of the deployed inductor while I_{ind} is the unit current capability.

The voltage regulator efficiency η_{VR} can be obtained using equation (4-72).

$$\eta_{VR} = \left(\eta_S^{C_{S_VR}} \right) \times \left(\eta_D^{C_{D_VR}} \right) \times \left(\eta_{ind}^{C_{ind_VR}} \right) \times \left(\eta_{Cin}^{C_{Cin_VR}} \right) \times \left(\eta_{Cout}^{C_{Cout_VR}} \right) \quad (4-72)$$

Where η_S , η_D , η_{ind} , η_{Cin} and η_{Cout} respectively represent the efficiencies of the switches, diodes, inductors, input capacitors and output capacitors.

For all computation of component counts presented in this chapter, it is required that each fraction be rounded to the positive integer of higher value. Attention should also be paid to the basic rules of capacitance and inductance computation in series and parallel circuits.

4.3 Summary

The procedure for selecting the optimal voltage for a turbo electric propulsion system has been presented in this chapter. These were based on two design levels of the propulsion system. The motor optimized approach involves

designing a motor, based on the propulsion needs and imposing the obtained voltage level for the motor on all devices that constitute the propulsion system. The advantage of setting the transmission voltage same as motor voltage is because no voltage regulators would be required. Thereby potentially eliminating the associated mass and inefficiency of extra devices.

However, there exist a possibility that components other than the motor may be better optimized if they operate at a different voltage level. This possibility led to a higher level of voltage selection approach which is architecture optimized. The propulsion system is split into 2 parts comprising the source branch and load branch, with a voltage regulator sitting in between to perform the voltage adaptation to the motor.

The presented approaches do highlight a coupling between the airframe, thrust requirement and the architecture operational voltage, leading to a more integrated but sensitive system. Major governors of the selected voltage are the stage loading coefficient, speed and torque requirement of the fan. The knowledge of these are a requirement to arrive at an optimal voltage for the system.

5 EXPERIMENTS AND DESIGN VALIDATION

This chapter focuses on the validation of the earlier proposed design and sizing methods. These include the fault, transient and thermal characterization, the proposed cooling method and the electrical machine sizing. The validation has been achieved using a combination of empirical procedures and available published data. An experimental facility capable of up to 400W was setup for this validation. This was required to enable the system responses to changing loading and inefficiencies to be verified, as there wasn't enough literature with which to perform such validation. The components of the facility will be presented in the subsequent sections where they have been utilized for the validation of the aspect discussed. Also, simulation tools developed on Simulink have also been utilized to perform the estimates, based on the equations presented in the earlier chapters of this work.

5.1 Fault Characterization Validation

The fault characterization validation is done by comparison of results from the presented procedure with published short circuit derivation in the literature. This is done to validate the tool developed for the fault current derivation.

5.1.1 AC Short Circuit Validation

The validation of the AC short circuit derivation has been performed in comparison with values obtained from [131]. The utilized electrical network specifications are shown on Table 9.

Table 9 – Validation Input Data

Parameter	Specification
Power rating(kW)	46.67
Voltage (V)	380
Current (A)	95
Frequency (Hz)	65
Resistance (Ohms)	0.1
Inductance(H)	35e-6

Figure 45 shows the derived short circuit is comparison with the reference. The top graph has been obtained using the developed tool while the bottom graph is taken from the reference.

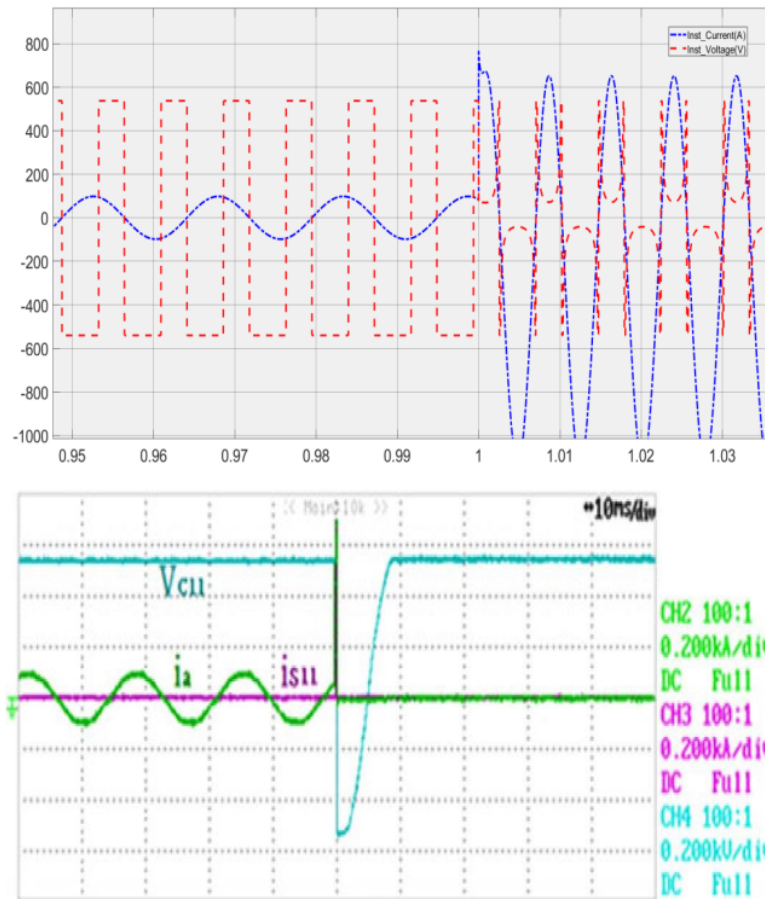


Figure 45 – AC Short Circuit Validation

The derived fault current wave form is seen to be correlative with the referenced, in both waveform and peak value, before and during the fault. The fault is seen to persist in the derived graph since fault handling is not considered in this work.

5.1.2 DC Short Circuit Validation

The tool developed for DC fault current investigation based on earlier presented equation, is validated using data from [132]. Table 10 shows the rating of the reference network and the computed network impedances.

Table 10 – Validation Input Data for DC Short Circuit

Parameter	Value	Units
DC Voltage (VDC)	12000	V
Rated Power (P)	1	MW
Current(I)	83	A
Total Network Impedance(Z)	24.2	Ohms
Fault Resistance	0.49	Ohms
Total Inductance	0.04159	H

The impedances have been derived using the RLC modelling approach presented in 3.2.2. Figure 46 (a) shows the referenced fault current (blue line) is comparison with the derived short circuit shown in Figure 46 (b).

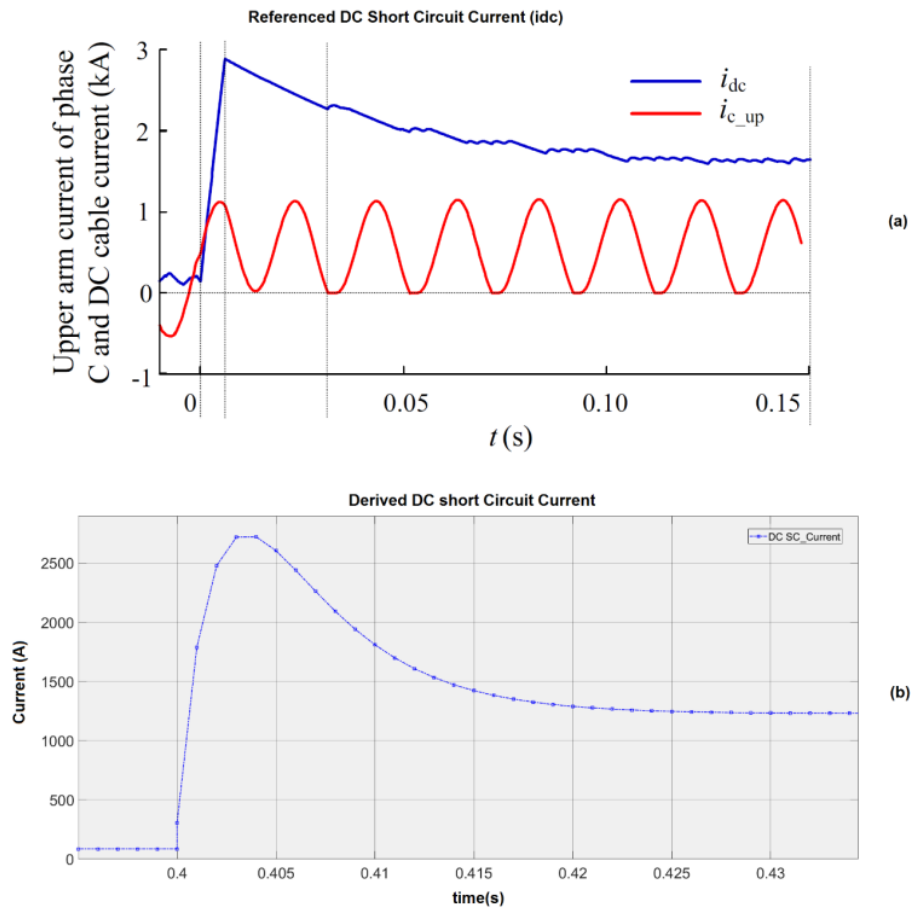


Figure 46 – DC Short Circuit Validation

5.2 Transient Current Validation

The transient current considered in this research is the inrush current at the start-up of the motor, since it exhibits higher value compared to other transients in normal operation. As stated in 3.3, the essence of identifying the magnitude of the transient current is to discriminate it from a fault current. In equation (3-11), it is seen that the transient current relies on the load inertia, which is a function of the fan mass and diameter in aerospace propulsion applications. Other parameters affecting the transient behaviour include the speed change and ramp time. In order to validate the procedure presented for deriving this current, a test is performed for the inrush current of three fans, using the same motor. Table 11 shows the details of the dimensions of the investigated fans.

Table 11 – Investigated Fans Dimensions

	Fan 1	Fan 2	Fan 3
Fan Mass(g)	37.9	9.2	3.3
Fan Diameter(cm)	14.2	25.6	15.5
Voltage	1.95	1.92	2.2
RPM	5113	5164	5769
Power	43.4655	43.008	34.43

Figure 47 shows the setup of the devices and the utilized fans.

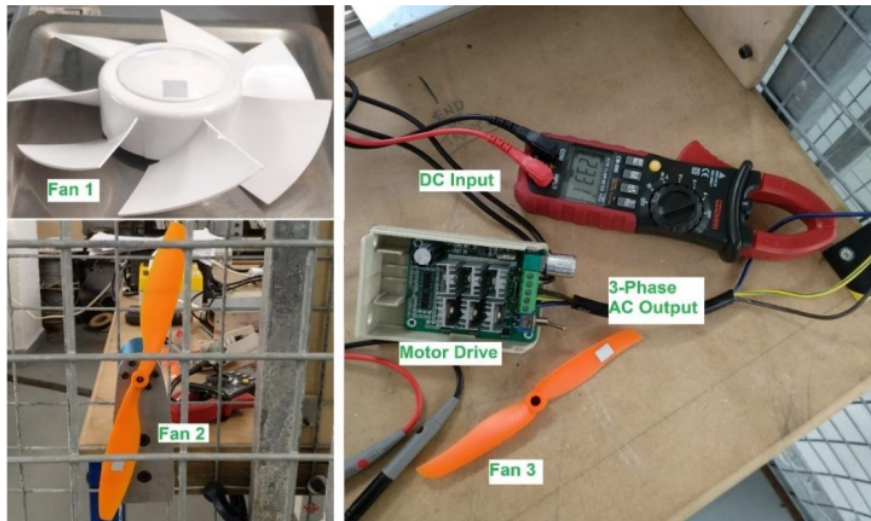


Figure 47 – Inrush Current Test Setup

A hard start and soft start are performed for all three fans and the inrush current is measured using a clamp meter. A noticeable difference in inrush current is observed for all three fans. These scenarios were simulated, to obtain estimated values, as a validation of the developed tool to be used for the investigation of bigger fans. Figure 48 shows a comparison between the measured and estimated inrush currents. Where F1, F2 and F3 respectively refer to Fan 1, Fan 2 and Fan 3 as presented on Table 11.

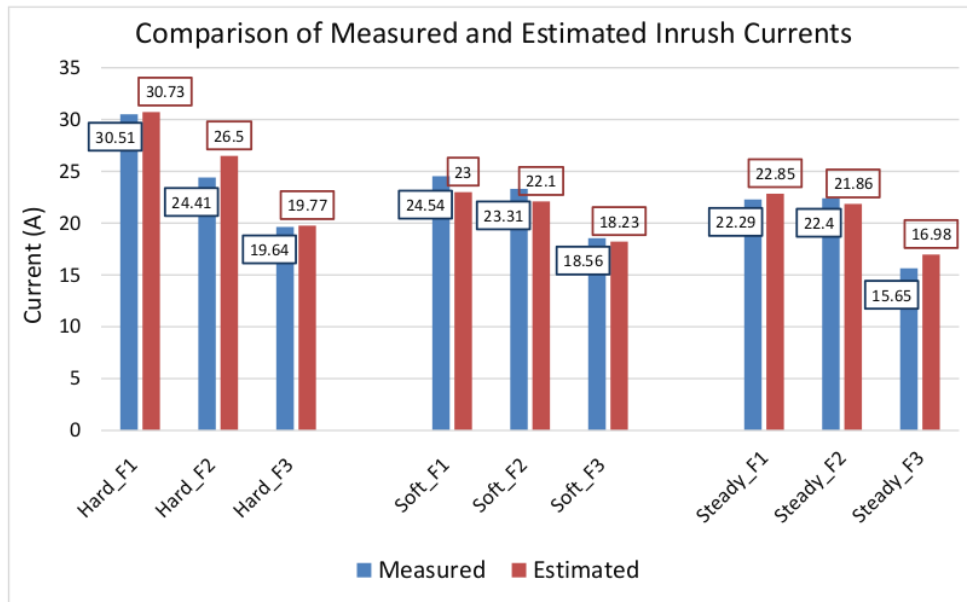


Figure 48 – Measured and Estimated Inrush Currents Comparison

The values are compared for hard starts, soft start and steady state. These represent an uncontrolled ramp to full speed, a controlled ramp to full speed and the steady state operation after full speed. The error in the estimated values appear to be within acceptable limits. These are shown in Figure 49.

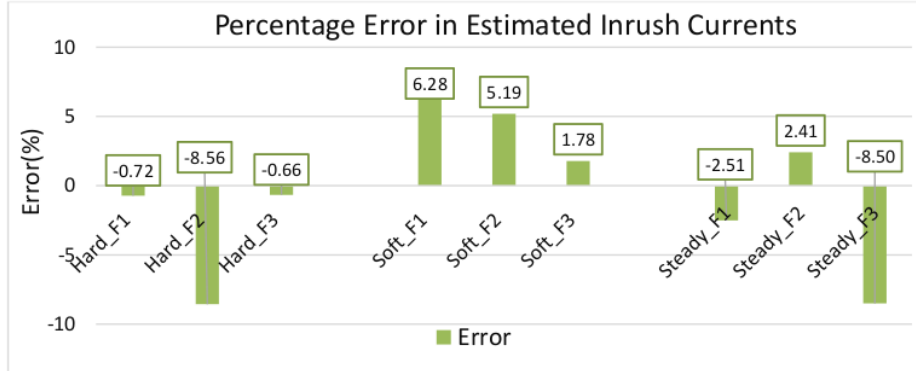


Figure 49 – Percentage Error in Estimated Inrush Currents

5.3 Power Electronics Thermal Validation

This section presents the thermal characterization validation for the power electronics components. This also has been done experimentally and is focused on validating the earlier proposed cooling strategy for the switches. The subsequent subsections present the test rig configuration, mass, experiment procedures and results. The validation does not cover the electric machines deployed on the rig, because they are not scalable to aerospace applications.

5.3.1 Test Rig Structure

The test setup consists of two-channel power sources and one load as shown in the block diagram of Figure 50.

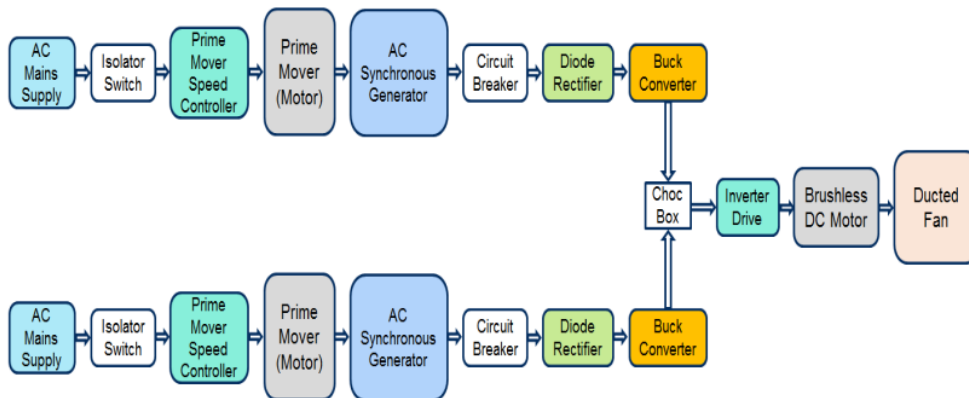


Figure 50 – Block Diagram Representation of a 400W Turboelectric Test Facility

The power flow is from the single-phase AC mains through the isolator switches to the prime mover/generator speed controller. Two 0.75hp induction motors (Black and Red) serve as the rig prime movers, emulating the turbines. Each prime mover drives a 400W rated AC synchronous Machine (white) that performs the 3-phase electric power generation. Figure 51 shows the broad view of the rig AC supply section.

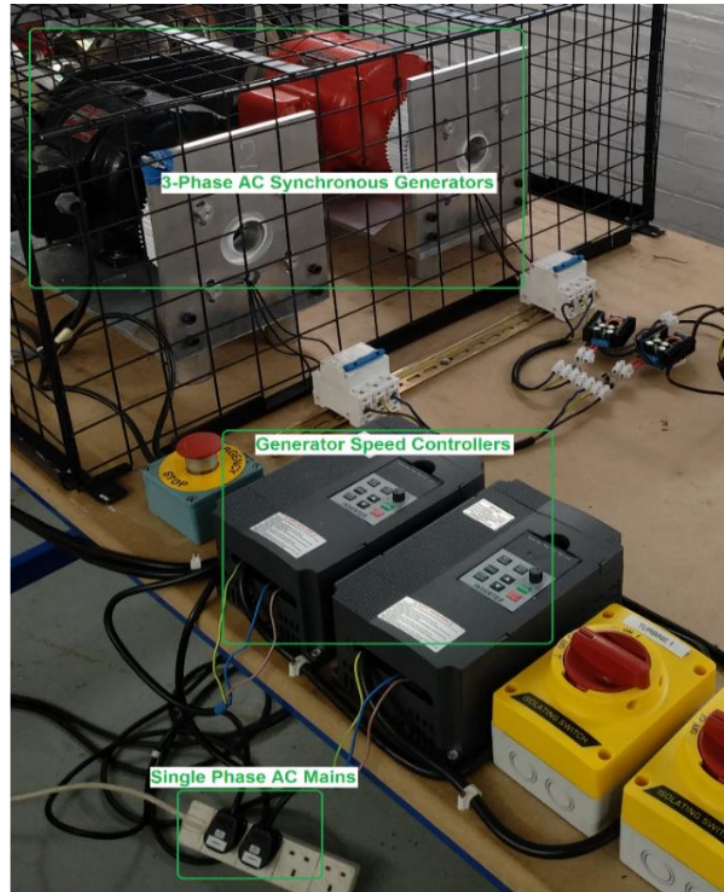


Figure 51 – AC Supply Segment of the Experimental Facility

The 3-phase AC power is routed through circuit breakers and then rectified to DC voltage by diode rectifiers. The DC voltage from the two rectifiers is regulated using 2 buck converters. These balance the voltages from the two- channel sources before they are tied up. Next, the combined DC sources are connected

to the motor 3-phase inverter for driving the ducted fan module. Figure 52 shows the DC and load section of the test facility.

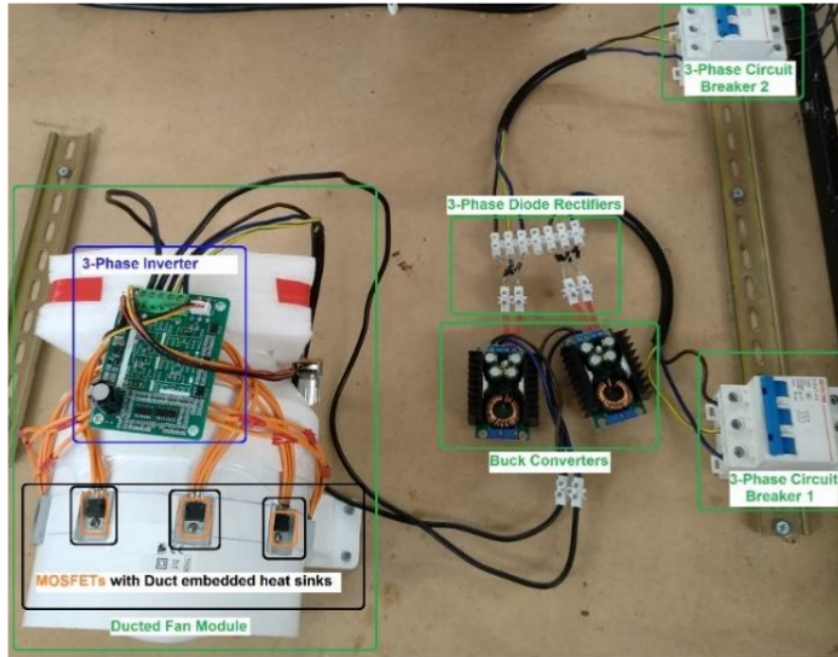


Figure 52 – DC and Load Segment of the Experimental Facility

The rating and mass of the utilized power devices are presented on Table 12

Table 12 – Ratings of Devices in the Experimental Setup

S/N	Component	Rating(W)	Voltage(V)	Current(A)	Mass(kg)	Manufacturer
1	Speed Controller	2200	220	12	-	Gecheer
2	Induction Motor	560	440	1.7	-	Crompton Parkinson
3	Generator	400	48	10	3.937	OS
4	Circuit Breakers	800	400	20(3000*)	0.2365	Tomzn
5	Diodes	600	36	-	-	Diotec
6	Buck Converter	336	28	12	0.0638	OS
7	3-Phase Inverter	360	36	23	0.0773	OS
8	Motor	902	22	41	0.213	SurPass
9	Heat Sink	-	-	-	0.0035	OS
10	Ducted Fan	29	-	-	-	BlauBerg

OS – Open Source. * - Fault Current Capability

5.3.2 Devices and Functions

The devices are presented in subsequent subsections, enumerating their specific functions in the experimental process.

5.3.2.1 Buck Converter

The buck converter is selected to investigate the thermal behavior of the inductor and capacitors. Although an inverting buck-boost converter is recommended for future aerospace applications, due to its dual step-up and step-down functionality, the non-availability of one adaptable for experiments has resorted in the use of a buck converter instead. However, this does not likely undermine the obtained results, since the both devices possess same components but only differ in power flow path and size. Figure 53 shows the devices used in this experiment.



Figure 53 – Buck Converters

The output voltage of the converter is regulated to 12V irrespective of input voltage. This choice is for two reasons; the first is to maintain a balance between the two supply channels, creating the feasibility of tying them up. The second is to keep the voltage value within safe limits of the motor and inverter drive, as would be required in a real propulsion system configuration.

5.3.2.2 3-Phase Inverter Drive

The 3-phase inverter drive has been modified to enable the experimental deployment. This involved decommissioning the MOSFETs and heat sinks and then extending them using cables, in order to space them properly along the duct

circumference. Figure 54 shows the original and modified versions of the inverter drive.

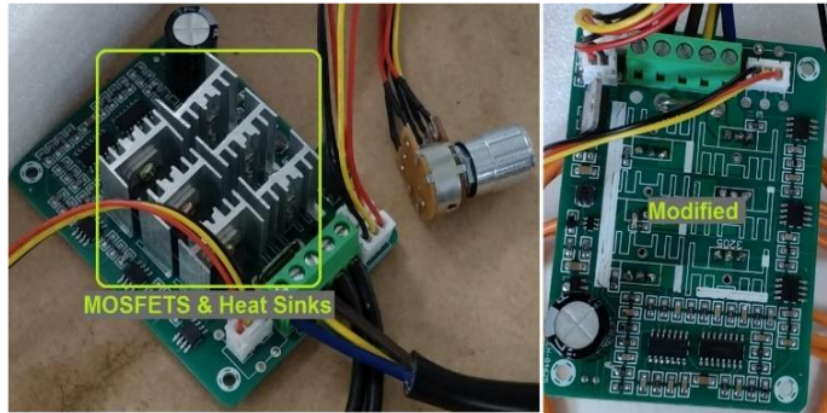


Figure 54 – Inverter Drive Experimental Adaptation

5.3.2.3 Ducted Fan Module (propulsor)

The ducted fan serves as both the electrical load and cooling device for the inverter drive switches. Figure 55 shows the ducted fan and its dimensions.

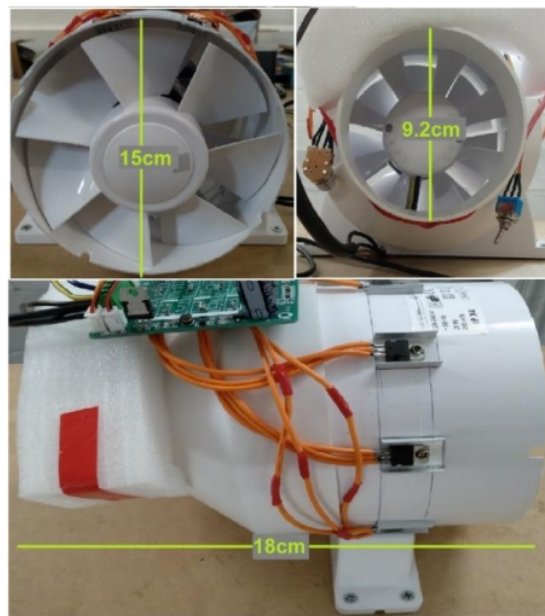


Figure 55 – Ducted Fan Dimensions

It consists of an extractor fan assembly whose single-phase AC motor has been replaced with a brushless DC motor. The brushless DC motor can ideally produce 2300RPM per volt input, up to a maximum of 22 Volts. Figure 56 shows the motor and its dimensions.



Figure 56 – BLDC Motor Dimensions

The heat sinks embedded in the duct is presented in Figure 57. showing the embedded region with arrows pointing downwards. The dimension of the heat sink is 15x10x25mm.

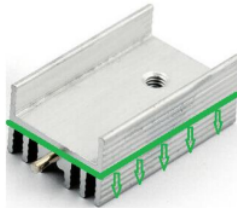


Figure 57 – Duct Embedded Heat Sink

Although a pin finned heat sink is recommended and modelled in 3.5, the non-availability of one matching the facility has led to the use of the same plate finned heat sinks, which have been preinstalled on the 3-Phase inverter drive before purchase.

5.3.3 Experimental Procedure

As stated earlier, the scope of this experiments only covers the power electronics. The experiment is performed in two phases following the aims, which are to:

1. Validate the switch, inductor and capacitor temperature characterization procedure presented in 3.4.2
2. Validate the extent of heat evacuation using heat sinks.
3. Validate the novelty of duct embedded heat sink cooling presented in 3.5
4. Derive a mass scaling factor for the overall mass estimation of power electronics devices.

The two phases of experiment are the cooled and uncooled phases.

The purpose of the uncooled run is to determine the temperature rise without auxiliary heat evacuation. For this phase of the experiment, the duration was determined by either steady state component temperature or an attained maximum temperature of 95°C, whichever comes first. It is important to state also that the experiment was performed with the duct enclosed and free of any embedded devices.

In the cooled phase of the experiment, only the MOSFETs are cooled by means of the attached heat sink being partially immersed in the duct airflow, by embedding them in the duct body as shown in Figure 55.

For both phases, the fan RPM was varied between 2000RPM and 5000RPM. For these speeds, the motor voltage was between 1.2V to 2.36V. Higher voltages and speeds were not used because the deployed motor has a kv value of 2300, which defines the RPM per volt of a motor. Therefore, not more than 2.2V is ideally required for 5000RPM. In order to exceed 5000RPM, more than 23A current had to be drawn, which is the design current limit of the deployed converter drive. The steady state current value is determined by load on the motor shaft, which in this case is a function of the inertia and mass flow characteristics. The thermal results are presented in the next section, in comparison with simulated values.

Figure 58 shows the layout of the simulation setup on Simulink. This has been used to estimate the efficiency and temperature of the components.

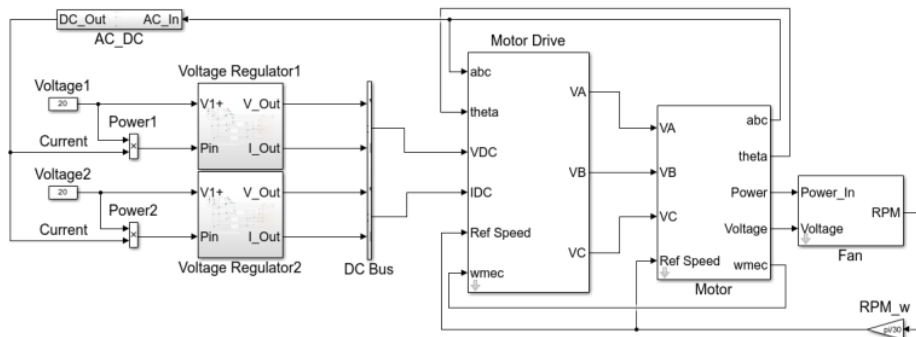


Figure 58 – Top Level View of Simulation Tool on Simulink

5.3.4 Experimental Results Comparison

The results obtained from the various experiments are presented in this section. The temperature values are compared with simulated values for the same power and speed ratings, as a validation of the earlier presented principles and assumptions. Focus is given first to the motor drive, since it carries the switches that are cooled in one of the phases.

The measured power and efficiency of the motor drive is shown in Figure 59.

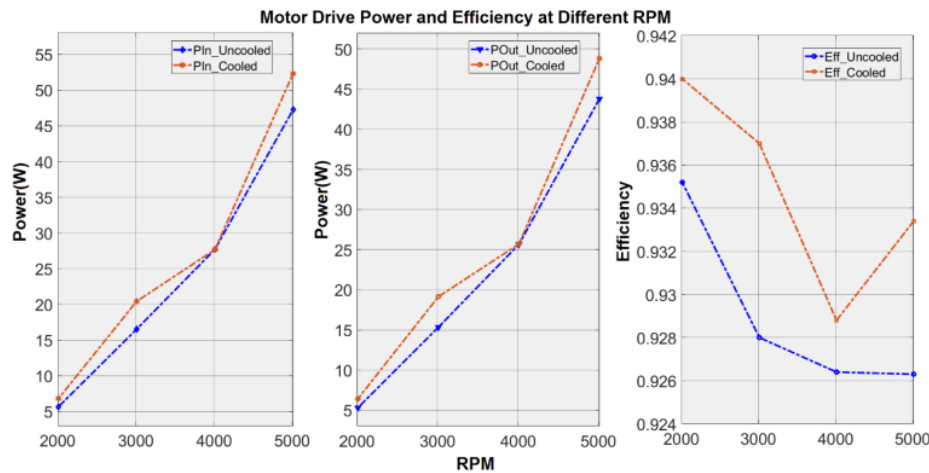


Figure 59 – Measured Motor Drive Performance Parameters

The trend of power input and output increases proportionally with increasing motor speeds for both the uncooled and cooled run. However, the motor drive

efficiency is more degraded in the uncooled run, owing to the performance degradation in the switches, due to lack of heat evacuation.

A comparison of the measured temperature values with the simulated values for the motor drive switches is presented in Figure 60. The trend lines with suffix 'S' represent the simulated while that with suffix 'M' represent the measured values from the conducted experiment.

A single maximum value is recorded from among the inverter drive switches, since the hottest component must be within its temperature limit to guaranty the device's availability. Ideally, the temperature values for all switches should be the same, but this was not the case and is likely due to the practical reality of unbalanced phase loading in electrical systems. A considerable difference is seen between the temperature profiles in the uncooled and cooled phase of the experiment. A gradual increase over time is observed for the cooled phase while in the uncooled phase, the drive switch temperature is seen to reach about 50% of the maximum temperature within one minute. A good correlation is also seen between the measured and simulated values, with the maximum deviation of about 6°C occurring for the cooled run at 4000RPM. This is a validation of the viability of the proposed cooling method and the switch thermal characterization.

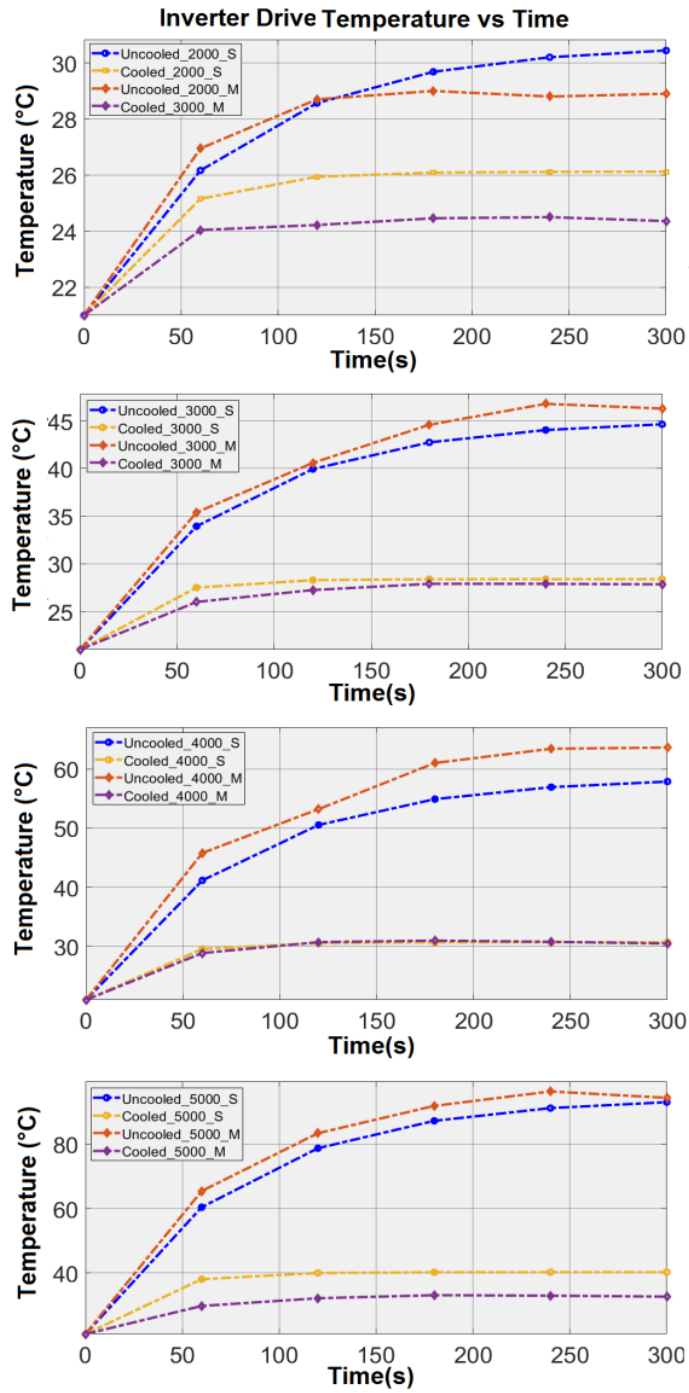


Figure 60 – Inverter Drive temperature

Other components temperatures have been investigated, but only for a motor speed of 5000RPM. These devices have not been cooled by the proposed approach using the duct air flow and have relied on natural convection to the ambient. Figure 61 shows the temperatures of the capacitors and inductor in the motor drive and Buck converter.

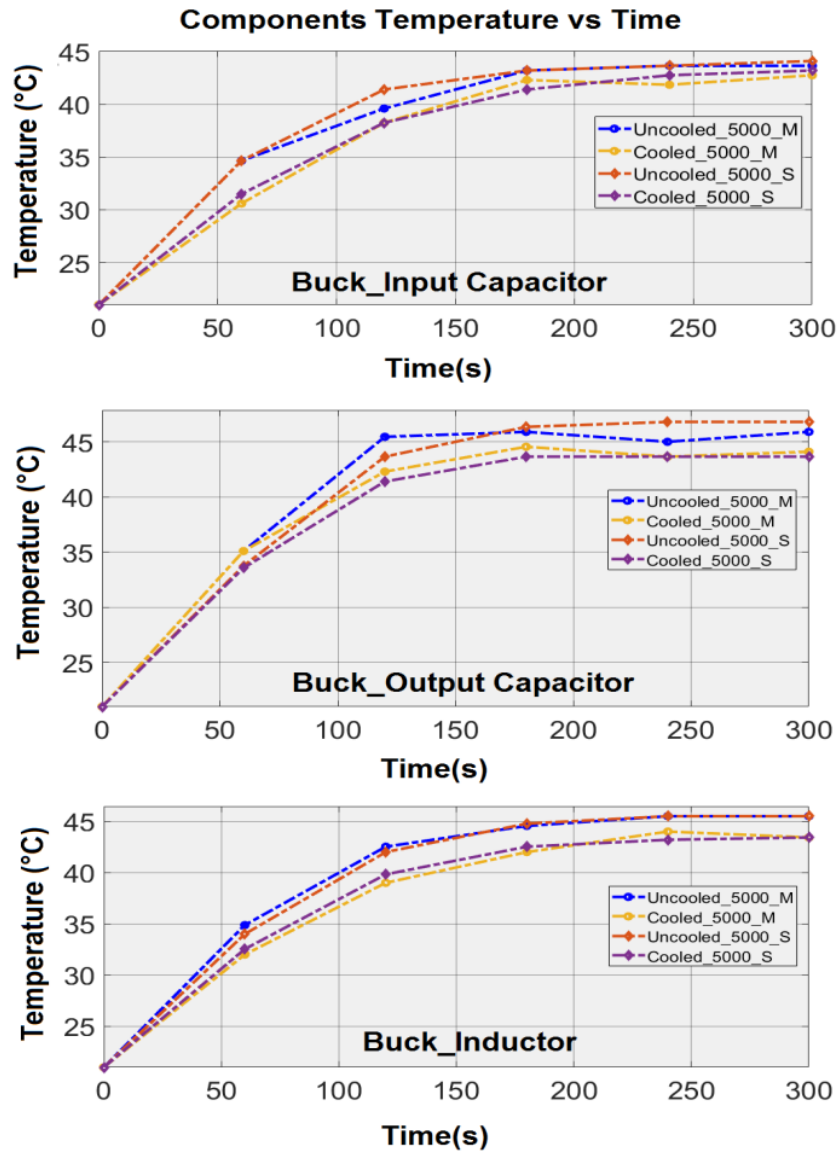


Figure 61 – Buck Converter's Capacitors and Inductor Thermal Performance

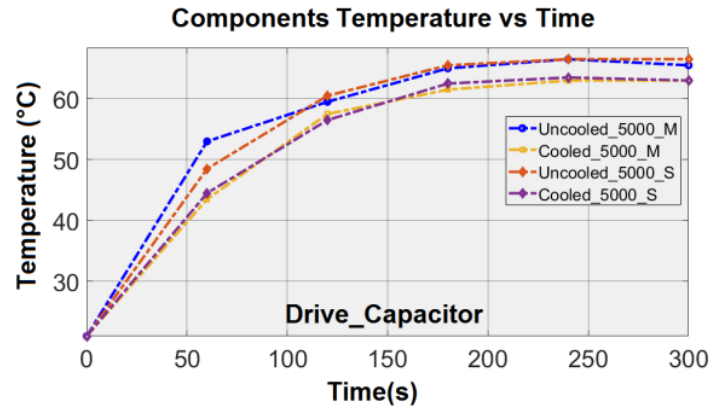


Figure 62 – Inverter Capacitor Thermal Performance

The devices exhibit slightly higher temperatures for the uncooled run. This is due to higher current drawn in that phase of the experiment as the system struggles to balance the degrading temperatures at the drive switches with the speed command. The simulated values are also within acceptable range of the measured, for both the cooled and uncooled phases.

The cables used for the AC and DC interconnection was of 2.55mm diameter with the internal conductor having a diameter of 1.58mm. For all scenarios tested, the simulated and measured conductor temperature change was less than 2°C.

5.3.5 Internal Drag Characteristic Due to Embedded Heat Sink

In order to estimate for the drag penalty caused by the addition of the heatsinks, the intake and exhaust air velocity is compared for both the uncooled run (normal duct) and cooled run (heat sink embedded duct). The aim here is to derive the additional power that must be added to maintain the air velocity. Figure 63 shows the air velocities and the associated static thrust for a range of input power and the respective fan RPM. The air velocities were measure near the fan face and right after the nozzle, to obtain the inlet and exit velocities respectively. Graphs (a) and (b) respectively show the intake and exhaust air velocities. Similar trend is exhibited for both the normal and heat sink embedded duct. A significant difference is observed for the intake velocity at 5000RPM. Also, the exit air velocity at 2000RPM is seen to be the same value for both the cooled and

uncooled configuration. This results in the cooled run providing slightly more thrust for as can be seen in (c).

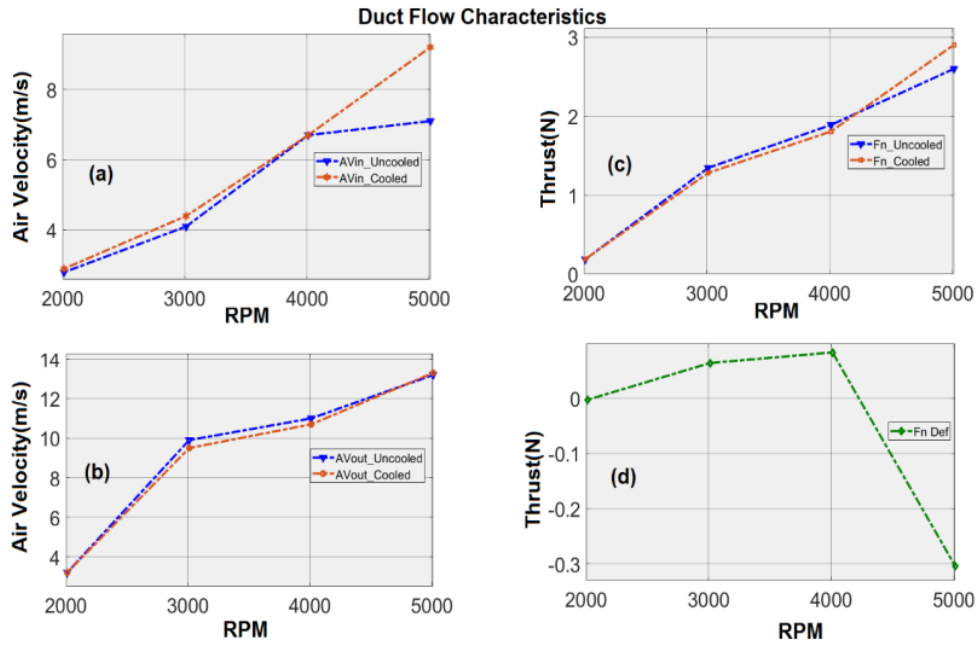


Figure 63 – Duct Flow Characteristics

The thrust deficit is presented in (d), slightly negative at 2000RPM, positive for 3000RPM and 4000RPM and very negative for 5000RPM. For fan speeds with positive deficits, it was required to derive the adjusted power that will produce same amount of thrust as the uncooled configuration. This was achieved using equation (5-1).

$$P_{adj} = P_{init} \left(\frac{F_{S_Def}}{F_S} + 1 \right) \quad (5-1)$$

Where, F_S is the thrust that needs to be compensated, F_{n_Def} is the deficit of F_S and P_{init} is the power yielding F_S .

Figure 64 shows the adjusted power (P_{New}) for all fan speeds investigated, in comparison with the initial power (P_{Old}). A saving in power is achieved for speeds that yield negative thrust deficits.

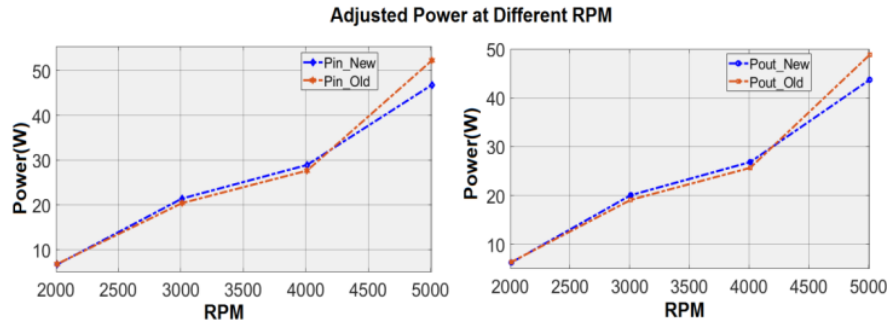


Figure 64 – Adjusted Power for Thrust Compensation

The lack of pattern in the manifestation of thrust deficit makes it difficult to define a scaling factor for the additional power required to overcome the deficit. However, it does indicate a potential for the proposed method to be utilized, following optimal design of the duct profile and the location of the embedded fins relative to the inlet, fan and exhaust.

5.3.6 Heat Sink Sizing Validation

The heat sinks sizing procedure has also been validated against the measured values. First, the thermal resistance was estimated, using the method presented in [133] and then the mass of the heat sink to yield same thermal resistance is obtained. Figure 65 shows the mass comparison for both the motor drive and buck converter incorporated heat sinks respectively. Values obtained are seen to be within acceptable range.

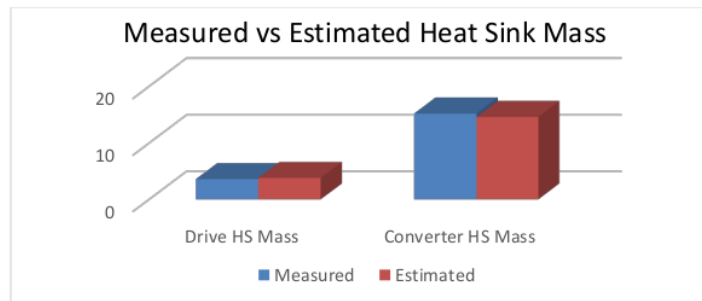


Figure 65 – Heat Sink Sizing Validation

5.3.7 Power and Mass Scaling factor Selection

The derivation of mass scaling factor is a means to account for the overhead accessories which are required to aid the power electronics perform optimally. These include the control boards, interconnecting devices, sensors, support structures, casing and sometimes intermediate cooling aids. These accessories together constitute more mass than the actual power electronics which they support. Therefore, a properly derived scaling factor is required to obtain realistic total mass estimates for the power electronics devices. The approach adopted in this research is to express these accessories masses as a function of the switch count. Table 13 shows data for the three converters considered for this derivation. The three converters have been selected because the design principles presented in this work are related to them. The *New mass* refers to the converter mass excluding the capacitor and all cooling accessories. This is done because the excluded are have sized separately, based on propulsion system requirement.

Table 13 – Mass Scaling Factor Data

Parameter	Rig Converter	X. Ding et al[96]	CRD300DA12E-XM3[128]
Power(kW)	0.36	30	300
Total mass(g)	77.3	16900	6200
Switch Count	6	3	3
Total Switch mass(g)	11.4	900	525
Capacitor mass(g)	3	900	1230
Heatsink mass(g)	3.6	5238.780624	1082
Cooling fan	0	0	62.14
New mass(g)	59.3	9861.219376	3300.86
Scaling factor	5.201754386	10.95691042	6.287352381

The scaling factor for each converter is obtained by dividing the *New mass* by the total switch mass. Due to lack of more adaptable data, only three scaling factors could be computed, making it difficult to generate a trend. Therefore, an average of the 3 values is taken as the switch mass scaling factor. A value of 7.5 is adopted for all devices comprising switches in the designs presented in the next chapter.

5.4 Electrical Machine Voltage Derivation Validation

The voltage and current selection approach presented previously for the electrical machines, is validated by comparing the values estimated with published motor parameters. the high specific power slotless permanent magnet motor presented in [20] is the SOA electrical machine and is designed specifically for aerospace propulsion application. Figure 66 shows a comparison of values estimated using the developed motor design tool with those published.

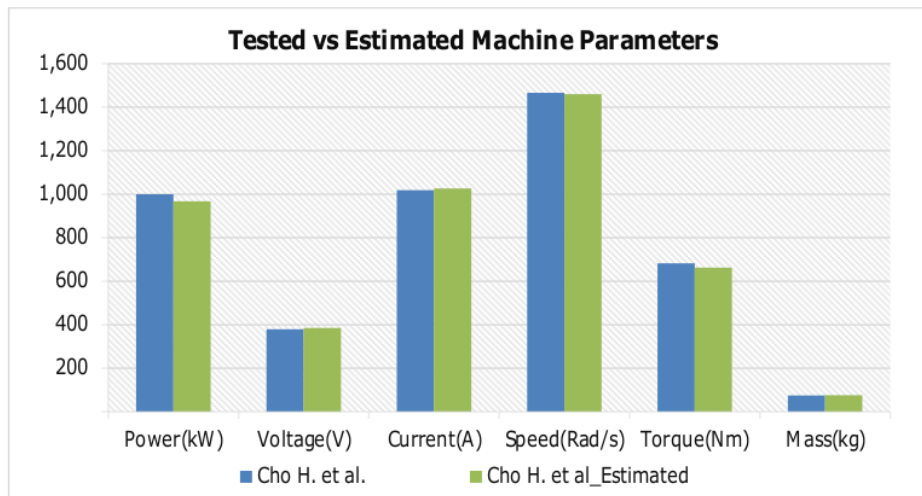


Figure 66 – Electrical Machine Sizing Validation

Obtained values are consistent, to the extent that this approach can be relied upon, in the selection of operational voltages and currents for the electric machines. Provided, the other propulsion system components sizing and efficiency, are not greatly impacted by the imposition of the motor's voltage and current.

5.5 Electrical Machine Temperature Validation

The electrical thermal characterization approach presented earlier has been validated also using the SOA motor described previously in [20]. There it is said that the maximum airgap flux density is 0.93T when the machine operates at 100°C. It is also stated the machine can deliver up to 96% efficiency. Hence, these values are used to deduce the maximum temperature of the windings.

It was important to first compute the thermal capacitances and resistances for each concentric layer that make up the machine. The dimensions of these layers have been inferred from details of Figure 67, presented in [20], and matched with Figure 68.

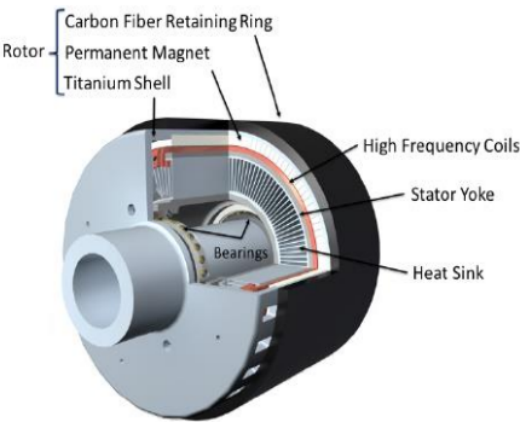


Figure 67 – Proposed PM Motor for Turboelectric Propulsion System

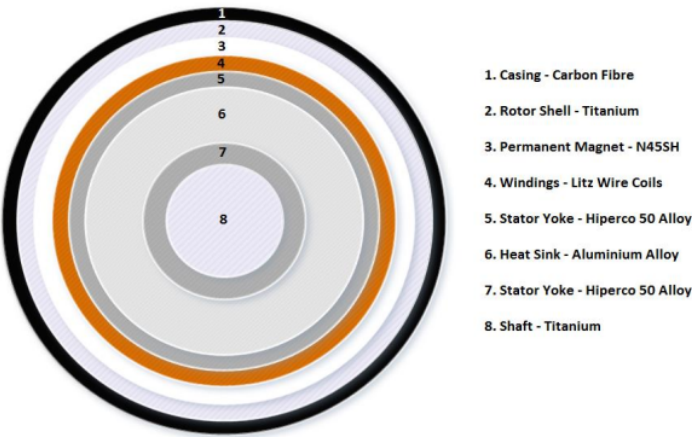


Figure 68 – Assumed Concentric Layers of the Proposed PM Motor

Using an iterative process with 75kg total mass as the convergence target, the outer diameter and other thermal constants for each layer was obtained and are presented on Table 14. The last column expresses the diameter of the outer surface of a given annulus as a percentage of the machine diameter.

Table 14 – Design Thermal Constants and Machine Percentage Diameters

	Specific Heat Capacity- Cp(J/(K kg))	Thermal Conductivity- λ (W/m·K)	Density- ρ (kg/m ³)	Percentage Diameter
Casing - Carbon Fibre	1600[134]	82.6[134]	1770[134]	1
Rotor Shell - Titanium	544[135]	20.4[136]	4510[137]	0.88
Permanent Magnet - N45SH ^a	502.4[138]	8.96[138]	7500[138]	0.83
Windings - Litz Wire Coils	281.8 ^b [139]	7.4 ^c [140]	2032 ^b [139]	0.75
Stator Yoke - Hiperco 50 Alloy ^a	420[141]	29[141]	8110.2[141]	0.7
Heat Sink - Aluminium Alloy	900	218	2690	0.65
Stator Yoke - Hiperco 50 Alloy	420[141]	29[141]	8110.2[141]	0.35
Shaft - Titanium	544[135]	20.4[136]	4510[137]	0.3

^a Converted unit, ^b Calibrated, ^c Per Phase

The thermal characteristic of the machine is shown in Figure 69. A maximum temperature of 104°C is reached after 5 hours.

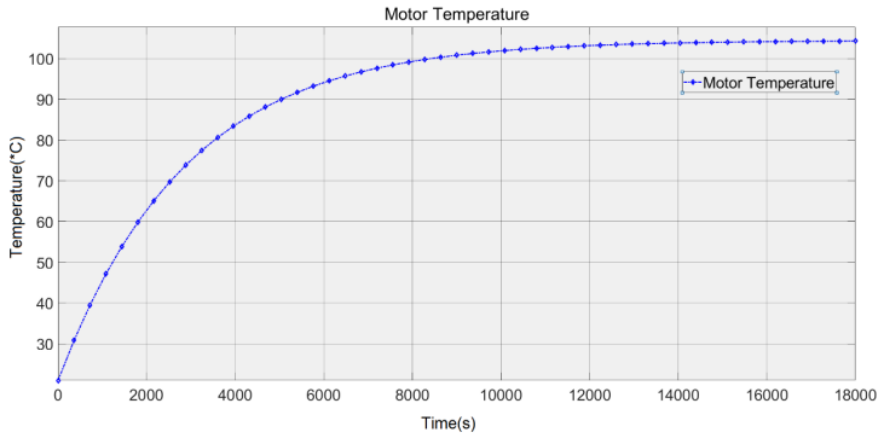


Figure 69 – Thermal Characteristic of the SOA Motor

5.6 Transmission Modelling Validation

The transmission sizing and thermal validation is done by comparing with data obtained in [49]. Figure 70 depicts a comparison between temperature trendline obtained with those published. The trend and maximum values are seen to be within range, although the rate of increase from the two methods appear to differ a little. This is likely to a difference in the solvers used.

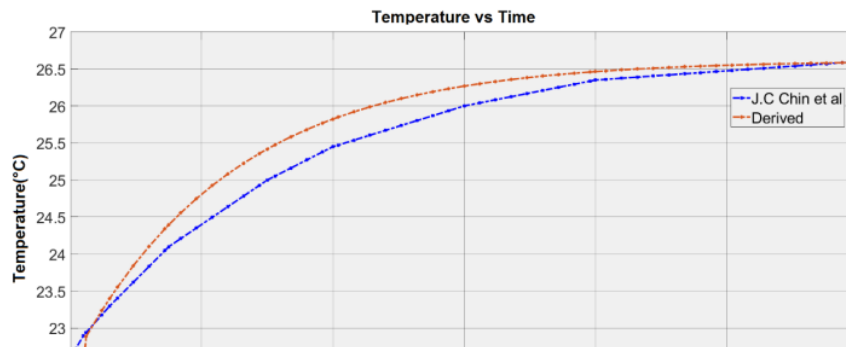


Figure 70 – Transmission Cable Temperature comparison

5.7 Summary

This chapter presents a validation of the feasibility design principles, presented in the earlier parts of this research, for application on future aerospace electrical propulsion train designs. Some of the validation has been done experimentally with others relying on published data.

The transient reaction of an electric network in response to the fan dynamics has been modelled and validated. This enables an estimation of a limit point for normal operation, beyond which protection systems should kick in and arrest a fault. The thermal behaviours of the power electronics devices in reaction to power flow has also been modelled and validated. Specifically, the switches were cooled using the proposed approach which appears promising despite requiring further investigation to establish the extent of drag penalties. A mass scaling factor has also been derived, based on switch count and total mass, using 3 existing converters as reference.

6 ELECTRICAL NETWORK DESIGN FOR THE PROPULSIVE FUSELAGE AIRCRAFT CONCEPT

This chapter presents the application of the earlier presented design principles. The propulsive fuselage concept earlier reviewed in 2.2.5 is selected, because of the availability of enough literature to advance the intention of this research. Also, the reason for adopting an electrical power system, rather than a mechanical drive for this airframe concept has been stated. The major implementation challenge being the added mass and thermal management.

The intended contribution of this chapter is to demonstrate the impact of faults, transients and thermal consideration in the sizing of the considered propulsion system. Subsequent sections present the stages of the performed investigation and the candidate configurations considered.

6.1 Architecture Layout and Components

Figure 71 shows a power flow layout of the electrical components for the considered concept.

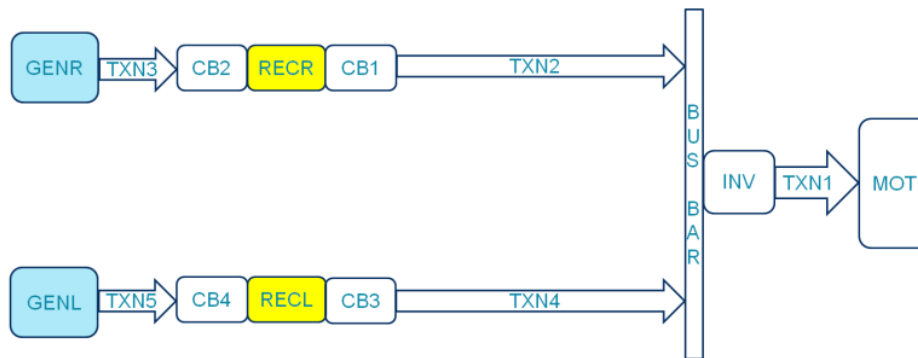


Figure 71 – Electrical Layout of the Propulsive Fuselage Concept

The comprising devices are the generator-GEN, transmission-TXN, rectifier-REC, circuit breaker-CB, inverter-INV and motor-MOT.

The power flow is in the direction of the arrows, from the generators through the bus bars to the motor. An inverse approach is adopted, since the propulsor driven

by the motor requires a certain amount of power and speed to provide the needed thrust. Hence, the motor is sized first with each successive upstream device sized to deliver enough power downstream. This way, enough power is eventually provided to the propulsor. The transmission system for the investigated airframe is divided into five segments as shown in Figure 71. Following convention, the electrical system is designed for the maximum power demand and rated as such.

6.1.1 Fan Parameters

Fan parameters have been obtained from [62] and compared with values estimated using methods presented earlier in 4.1.1. These are presented on Table 15 as a guide for the electrical network design and investigation.

Table 15 – Assumed Propulsion system Design Inputs

Parameter	Published (SI Units)	Estimated
Thrust(lb)	3210(14278)	14009(N)
Altitude(ft)	37574(11453)	11400(m)
Power(hp)	3500(2610000)	2699859(W)
Mach	0.785	0.78
Fan Pressure Ratio	1.25	1.25
RPM	1920*	2813
Fan Diameter(in)	81(2.057)	-
Fan Hub Diameter (in)	24(0.61)	-
Fan Mass(kg)_Aluminium hub	-	289.3
Fan Mass(kg)_Titanium hub	-	358.7
* Take-off [63]		

6.1.2 Electrical Parameters

For the design of an electrical system of known power, the rated voltage and current is first obtained for the machines which constitute the power source and load. These are the motor and the two wing mounted generators. The investigation is performed for motor speeds from 1920 RPM to 14000RPM at 2000RPM intervals. The lower limit of the speed range has been selected based on Table 15, while the upper limit is based on the design value for the SOA electrical machine in [20], with the estimated speed of 2813RPM in-between. The proposed speed for the generator according to [63] is around 4200RPM. Figure 72 shows the derived motor voltage and currents for the considered RPMs.

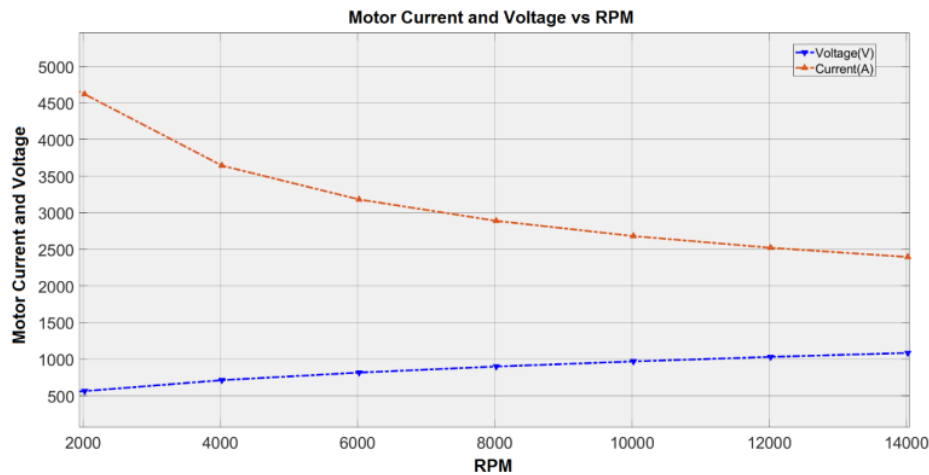


Figure 72 – Electrical Machines Current and Voltage vs RPM

These voltages and current have been derived using methods presented in 4.1.2. An investigation is performed next with these derived parameters for the proposed aircraft concept, for a power rating of 2.6MW.

6.2 Start-up and Transients Characteristics

The importance of performing an evaluation of the electrical system behavior at startup and transients is because of the susceptibility of systems to failure under such scenario. This is especially because of the magnitude of power required for the short ramp up time required for aerospace propulsion systems. First, the surge current of the motor at startup is determined, since it endures a longer duration than short circuits. As stated in section 3.3 the current drawn during motor transient is reliant on the load inertia and acceleration. The magnitude of this current has been investigated for range of motor design RPMs. A five seconds ramp up to full speed duration is used. This is in accordance to FAA requirements, that the thrust of an engine must be able to increase from flight idle to 95% rated power within five seconds[63]. Figure 73 shows the inrush current for the motor rated at 1920RPM. The five seconds setting has been applied using a soft start approach, gradually increasing the voltage input. The inrush of current on start-up results in about 8kA surge I_s , a 180% increase compared to the nominal current I_n of 4.3kA. The surge duration is seen to span about 3.2

seconds, a duration that is beyond the typical overload limit of switches. Therefore, an increase in switch count is required, purposely to handle this surge current.

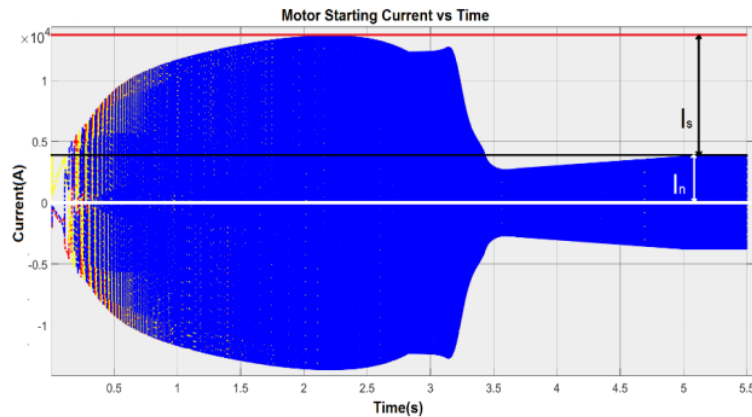


Figure 73 – Current Surge Magnitude for 5 seconds Start-up

The presented surge current is representative of the optimized scenario as earlier highlighted in section 3.3. Other motor speed ratings up to 14000 RPM have also been investigated, for 5s and 10s ramp duration for a titanium and aluminium based fan hub. This is to reflect the translation of increasing fan mass through inertia to higher transient current profiles. Figure 74 shows the inrush current evolution for four combinations of fan masses and ramp up durations.

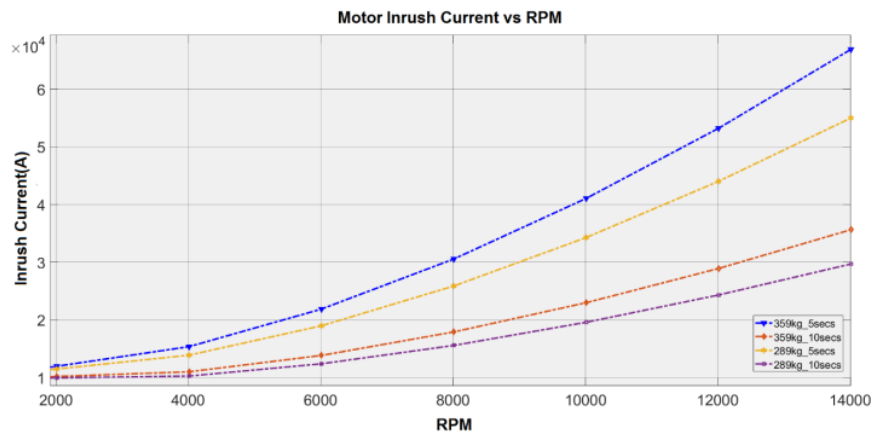


Figure 74 – Current Surge Magnitude for different Speed Ratings.

Further investigations will be performed, to see the influence of these trend on the overall mass and efficiency of a propulsion system.

6.3 Architecture Mass

The architecture mass is presented for a range of motor RPM and generator RPM combination for 5 seconds and 10 seconds ramp up time. These yield configurations with different mass profiles as listed in Table 16.

Table 16 – Investigated Architecture Profiles

Fan Mass(kg)	Ramp Time(secs)	Generator Speed (RPM)	Designation
359	5	4200	359kg_5secs_4200GenRPM
359	10	4200	359kg_10secs_4200GenRPM
289	5	4200	289kg_5secs_4200GenRPM
289	10	4200	289kg_10secs_4200GenRPM
359	5	14000	359kg_5secs_14000GenRPM
359	10	14000	359kg_10secs_14000GenRPM
289	5	14000	289kg_5secs_14000GenRPM
289	10	14000	289kg_10secs_14000GenRPM

The mass for each of the profile was investigated for 7 different motor RPMs. Figure 75 shows the mass of the electric machines for the investigated profiles.

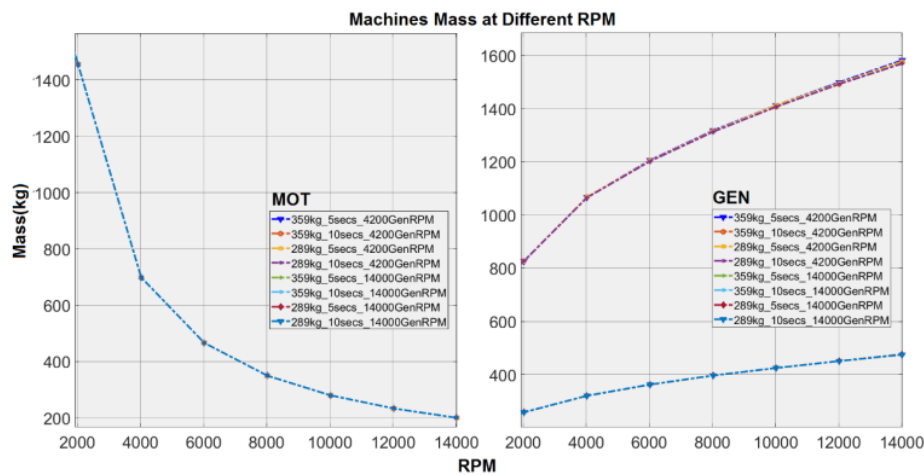


Figure 75 – Electrical Machines Mass Profile

It can be observed that the motor mass has remained the same for all profiles. This is because the motor is sized before a surge current could be derived. It is possible to optimize the motor mass side-by-side the surge current. However, the intension here is to show the reaction of a motor to varied loading conditions. A considerable reduction in motor mass is observed with increasing rated RPM. The generator on the other hand share similarity across the eight investigated configurations, having broadly two categories. One category is for the profiles operated on 4200RPM generator speed and the other for 14000RPM. In each category there exist little differences, which ranges from about 2kg to 10kg. Therefore, ramp time and fan mass profile does appear to have only little effect on the generator mass. The generator mass for a profile having a low motor speed is significantly lighter than that when the motor speed is high. The large difference is due to the motor's voltage being imposed on all the electrical devices. As seen in Figure 72, higher motor speeds correspond to higher voltages. However, the generator speed is fixed by the turbine. With fixed generator speed and number of poles, achieving higher voltages require higher magnetic fluxes, hence bigger magnets and more mass.

Figure 76 shows the inverter drive and rectifier masses for the investigated scenarios.

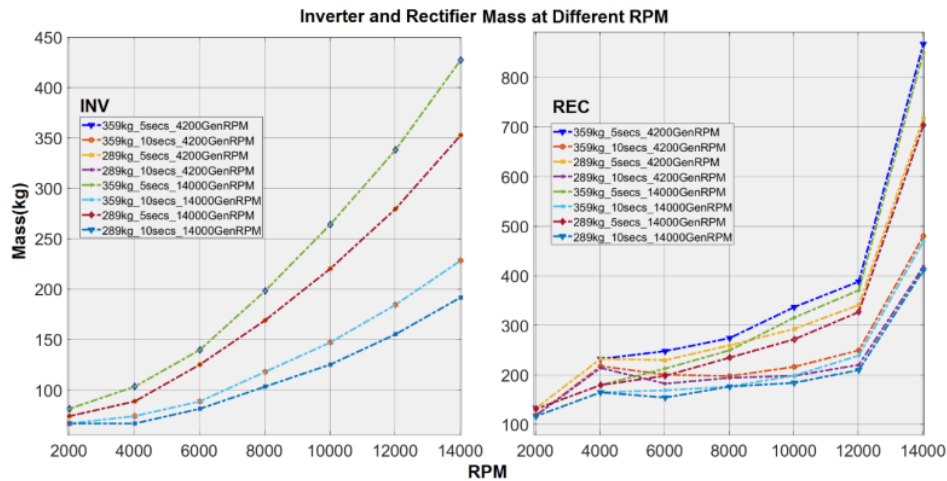


Figure 76 – Inverter Drive and Rectifier Mass Profiles

segments because it has the longest distance. *TXN1* and *TXN3* are 2m each, while *TXN2* is assumed to be 37m.

Figure 78 shows the mass profiles of the circuit breakers. It is important to mention here that the inverter also serves a protection function but has been presented earlier in Figure 76. *CB1* and *CB2* exhibit similar trend with the inverter, having four mass profiles. This is because the magnitude of currents used in the sizing is closely related to the inrush current.

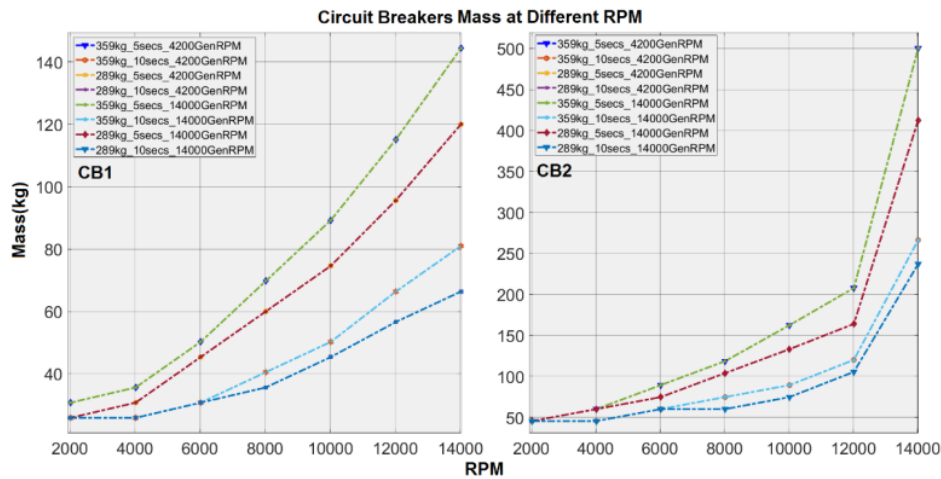


Figure 78 – Circuit Breaker Mass Profiles

The effect of operating at voltages higher than 1200V is also seen with configurations above 12000RPM. Although the graph of Figure 72 doesn't indicate voltages above 1200V for the electric motor, the devices upstream of the motor are required to operate at slightly higher voltage, to be able to deliver the voltage required by the motor despite transmission related voltage drops.

Figure 79 shows the cooling mass profiles of the investigated scenarios. The presented mass is the total for all the deployed heat sinks, for both switches and capacitors. The length and width of the heatsinks were selected in proportionality with the cooled components' dimensions. The optimization was done based on the flow chart in Figure 31.

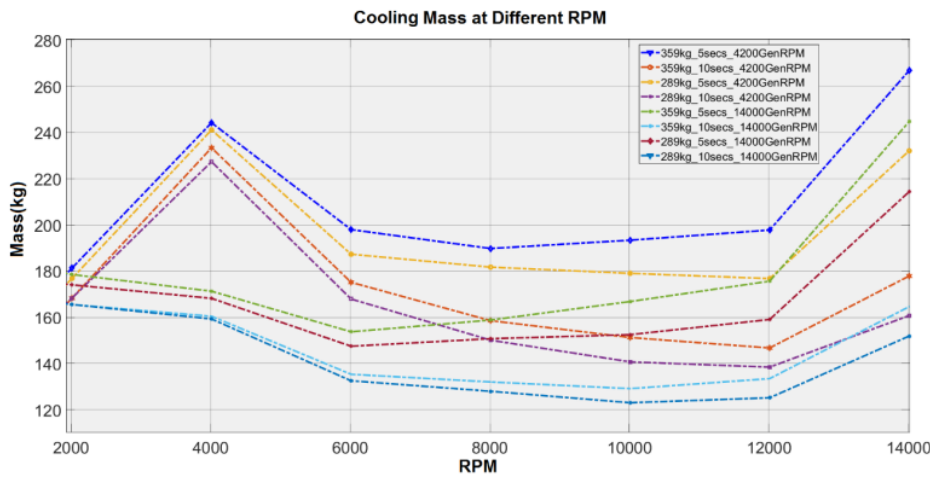


Figure 79 – Cooling Mass Profiles

The sudden increase observed for the 4000RPM is traced partly to the doubling of the filter capacitors in the rectifier, for voltages higher than 700V which is its maximum rating. Since these capacitors are also cooled, their total count would also influence the heat sink mass. The architecture voltage rating at 1920RPM motor speed is less than 700V and is hence unaffected. Another reason for the sharp increase is the gradient in current trendline between 1920RPM and 4000RPM, as can be seen in Figure 72. The changing proportionality of voltage to current with increasing RPMs explains why the trend is not maintained after 4000RPM. The increase in cooling mass due to components count is also current dependent, as a paralleling of several units of a given component is required, to meet the current handling needs of the architecture.

The architecture total mass comparison for the different configurations is shown in Figure 80. The trend shows that the optimal mass profiles are towards the median point of the investigated range of motor speeds. This is because at lower speeds the power electronics equipment has better mass, but the electrical machines are significantly heavier. This tendency is switched at very high motor speed configurations. The best-case configuration was found to benefit from lower fan mass, high ramp duration and high generator rotational speed.

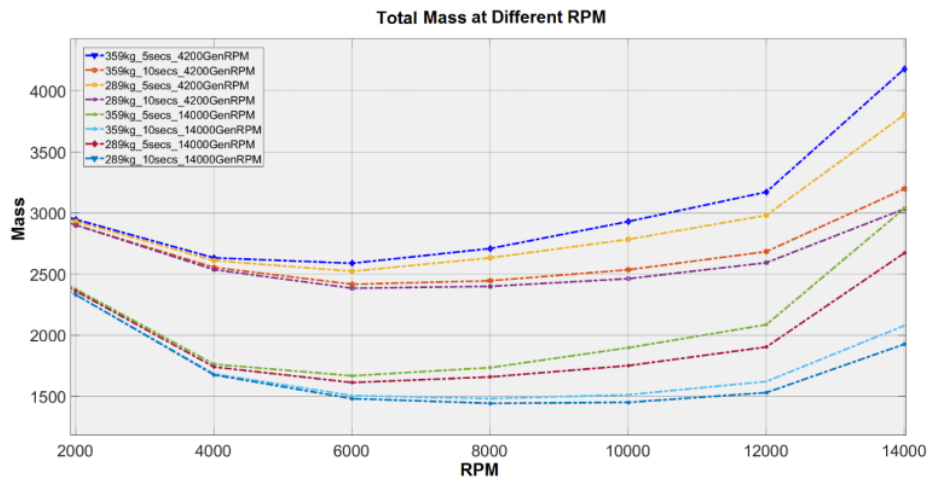


Figure 80 – Architecture Mass Comparison

6.4 Fault Current Characteristics

The fault characteristics of the architecture is presented in this section. The aim of this section is to achieve an optimal protection system size, using the fewest possible switches. This is enabled by the possibility to intercept the fault before it reaches its peak value. The fault characterization makes it possible to know earliest possible fault interception time from its occurrence. The boundaries for this choice are the switch short circuit capability and the maximum non-fault current, like the inrush current for example.

Five protection zones have been identified as shown in Figure 81 and the magnitude of a symmetric short circuit is estimated for these points.

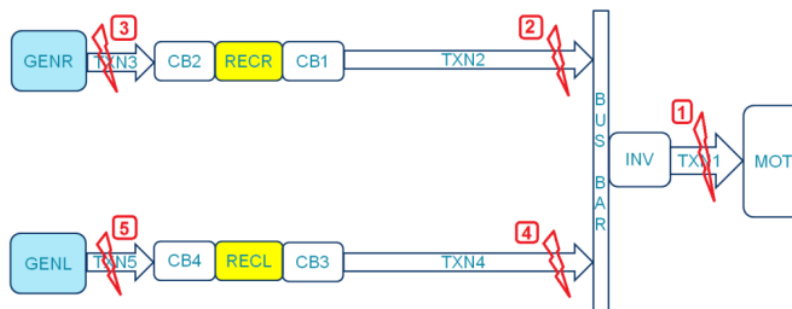


Figure 81 – Fault Areas of the Investigated Electrical Network

As done previously, only faults at point 1, 2 and 3 are shown, since Points 4 and 5 are ideally a replica of 2 and 3 respectively.

Following the procedure presented in section 3.2. An RLC computation is done for the network, to obtain the impedances between the protection devices and the assumed fault points, since the power factor, frequencies and efficiencies have been derived.

In order to keep focus, only the default configuration is considered for the fault investigation. The *359kg_5secs_4200GenRPM* is assumed to be the default configuration, which only holds the fan mass, ramp up duration and generator speed constant. This is set at 359kg of fan mass and ramped up duration of 5 seconds, while the generator is operated at 4200 RRPM. The investigation is hence performed for the range of RPMs earlier considered. Figure 82 shows the magnitudes of the fault currents occurring at point 1 for the different motor speed configurations.

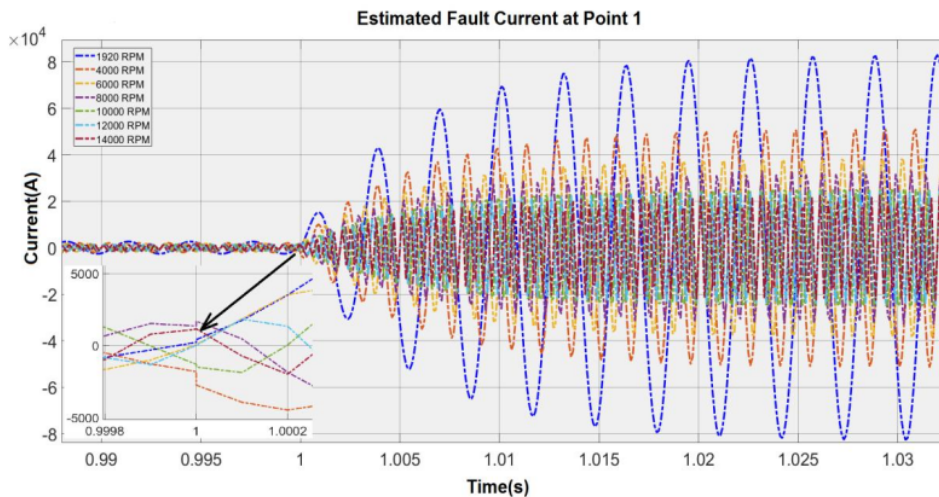


Figure 82 – AC Fault Current Evolution between Inverter and Motor

Higher motor RPM configurations are seen to have lesser fault current values. The maximum value of the fault current for most of the investigated RPMs reach the peak after about 30ms. However, it is possible to intercept this fault earlier, since the switches can detect a fault within 1 μ s. The fault handling relies on the

switching frequency of the protection system which should be preferably high. A balance is required, since such high frequencies have efficiency and hence thermal penalties. The switching frequency of the devices in this research is selected as a multiple of the fundamental frequency of the AC source or load, owing to a multiplicity of optimal choices as stated in[142]. However, a range of 5kHz to 20kHz has been used. The zoomed portion of graph in Figure 82 shows that before the $1\mu\text{s}$ duration is reached, the fault currents at that point of the network is still lower than the derived transient currents. Therefore, it is possible to intercept the fault current once its value is slightly above the maximum transient current. A multiplier of 1.1 times the derived transient current has been used for the protection device sizing.

Figure 83 shows the estimated fault currents for fault point 2. The trend of peak values is seen to vary inversely with motor RPM. Very high magnitude currents are observed, compared to the AC values. For all motor RPM configurations, values are above 100kA with the 14000RPM reaching about 240kA. However, these peak values are not reached before the $1\mu\text{s}$ interception capability of the switches. Therefore, the pre-set trigger point of 10% above the optimized inrush current level of the motor can be used to achieve an optimal design of the protection.

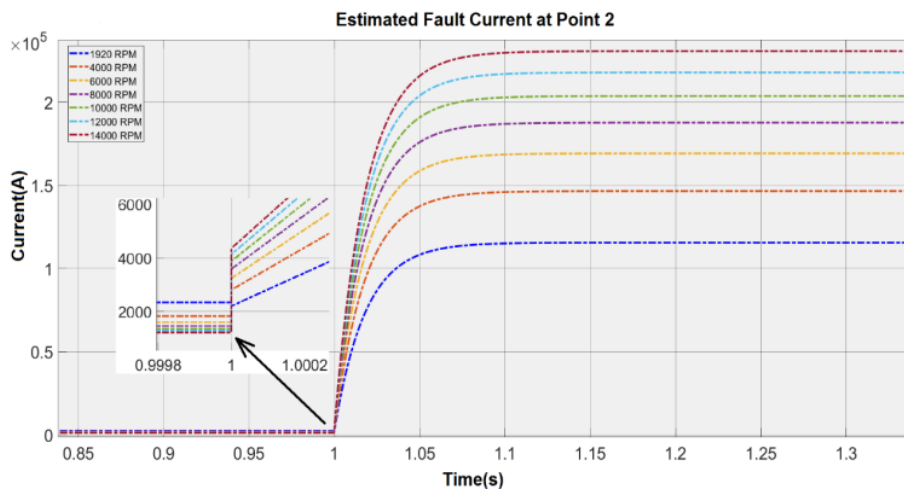


Figure 83 – DC Fault Current Evolution between Rectifier and Inverter

Figure 84 is a representation of the fault current characteristics at fault point 3. The values obtained are slightly higher than at point 1, despite carrying half of the power. This is mostly due to the proximity to the power source and hence lower impedance on the fault current path.

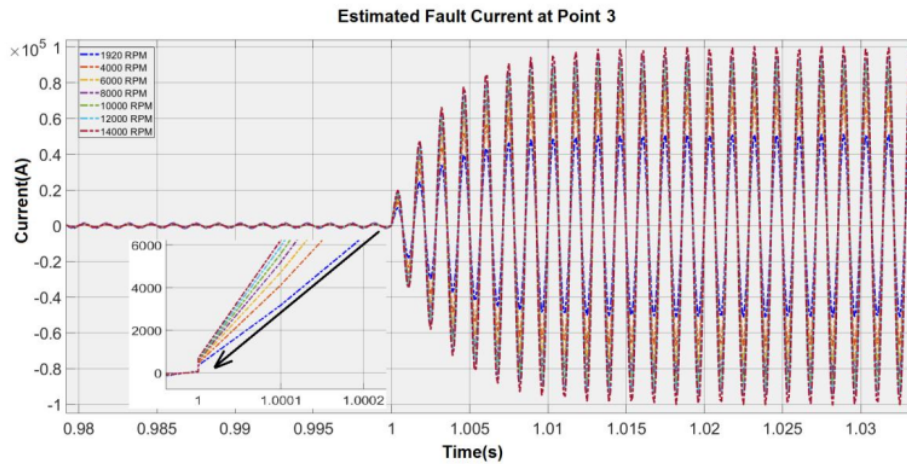


Figure 84 – AC Fault Current Evolution between Generator and Rectifier

From the investigation, the utilized switches possess the capability to intercept the fault currents before they reach 5kA. However, a requirement to consider the maximum transient current values have led to the oversizing of the protection devices for currents 10% higher than the maximum transient values, to avoid false shutdowns. This further shows the influence of these transient currents magnitude on the systems size.

6.5 Efficiencies and Thermal Behaviours

Investigation of the thermal behaviour of the devices is presented along with the device efficiency profiles, since they have a close interrelationship through power loss. Although the efficiencies of all eight configurations are presented, only temperature profiles for the default configuration is shown, for motor speed variation from 1920 to 14000RPM. A total of 5 flight hours have been investigated, with ambient temperatures changing from take-off altitude through cruise to landing. The long period was to observe the maximum temperatures of each component. Faults have also been incorporated at the earlier investigated points

from 3000s to 3800s during the simulations. Figure 85 shows the changing characteristics for the cooling air flowing past the duct embedded heat sinks. The temperature used for the power electronics devices is that before the fan.

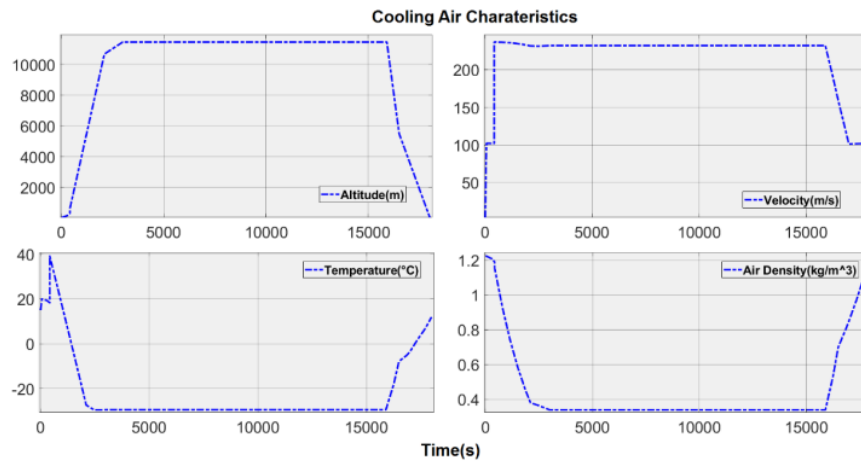


Figure 85 – Cooling Air Characteristics for the Different Stages of Flight

The various components are discussed in subsequent subsections, showing the thermal trend both at normal and fault conditions.

6.5.1 Electrical Machines

The efficiency of the Electric machines is shown in Figure 86.

Figure 86 – Electrical Machines Efficiency vs RPM

The efficiency is seen to increase with increasing machine speed, but over a small range. The motor shows a superior efficiency profile compared to the generator. The efficiency of the motor is rarely affected by the changing fan mass and ramp up duration, the differences for all cases were in the maximum order of $1e-4\%$. The generator shows more variation because it is sized differently for an architecture of different blade mass and ramp up times. These Characteristics are seen to impact on the size and efficiency of all devices upstream of the motor. Hence the generator serving as the power source, must be sized to match the changing efficiencies. Although the estimated efficiency of the generator appears very close, a slight improvement is seen when operated at 14000RPM as compared to 4200RPM.

The temperature is directly related to these efficiencies through the respective heat losses. The temperature profile of the electrical machines is presented for the assumed default configuration. Figure 87 shows the temperature profile of the motor over the investigated speed range and mission duration. For the investigated aircraft configuration, the motor is housed in the aircraft and does not benefit from the cold ambient at high altitudes. Therefore, an ambient temperature of 21°C is assumed.

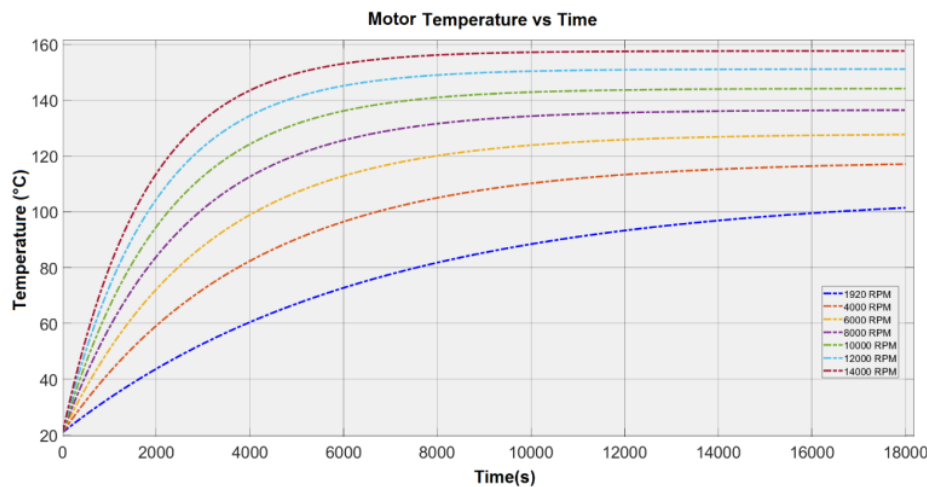


Figure 87 – Motor Temperature Variation with Time

The higher speeds show higher temperature dissipation despite superior efficiencies, due to their higher power density and hence lesser heat capacity. Based on the referenced SOA machine, upon whose principles the machines were modelled, the maximum recommended temperature is 100°C. Therefore, only at 1920 RPM can the motors be deployed at maximum power for the entire flight. Other RPM models can be deployed, provided additional cooling is applied, or an adjustment of the power demand is implemented to keep the temperatures within limits. The generator temperature profile is shown in Figure 88, which are all within the limits for the investigated scenarios.

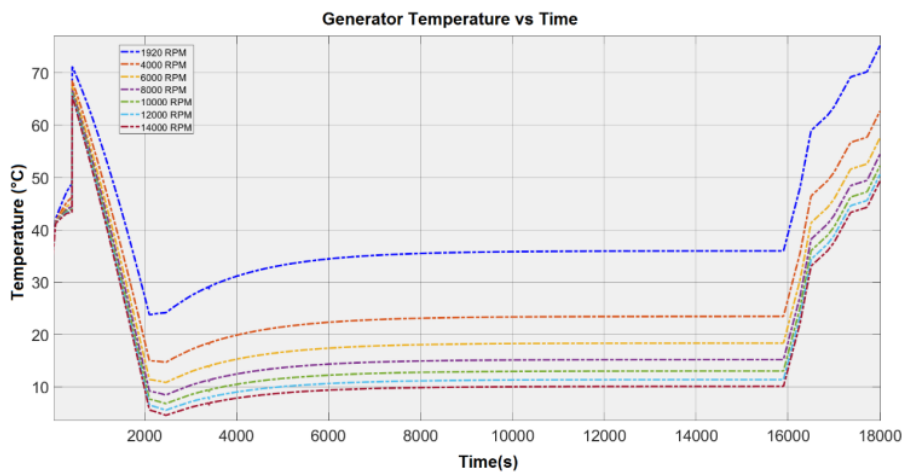


Figure 88 – Generator Temperature Variation with Rated Speed

Compared to the motor, the generator has a better temperature profile since it benefits from the cold ambient air. It is important to state that the temperature assumed as the ambient for the generator is that behind the fan, since it is proposed to be installed concentrically with the turbo fan low pressure shaft.

6.5.2 Power Electronics

Investigating the thermal behavior of the power electronics-based devices for aerospace propulsion is one of the main intensions of this work. This is because of their important role in control, protection and regulation of the entire propulsion system. The temperatures presented here are only of the switches, capacitors

and inductors, since they are the components from which the most heat emanate. These are presented for one unit in each functional group (branch) of a device.

In order to properly investigate the thermal behavior of these components, their instantaneous power handling must be derived. This is a function of current and voltage which must be evenly distributed across the available components. Therefore, a count of these component must be derived based on the power demands of the electrical network, then it is possible to determine how much power each must handle. The temperature of each component is a function of how much of power into it is lost as heat due to its inefficiency. The efficiency of the inverter and rectifier are shown in Figure 89.

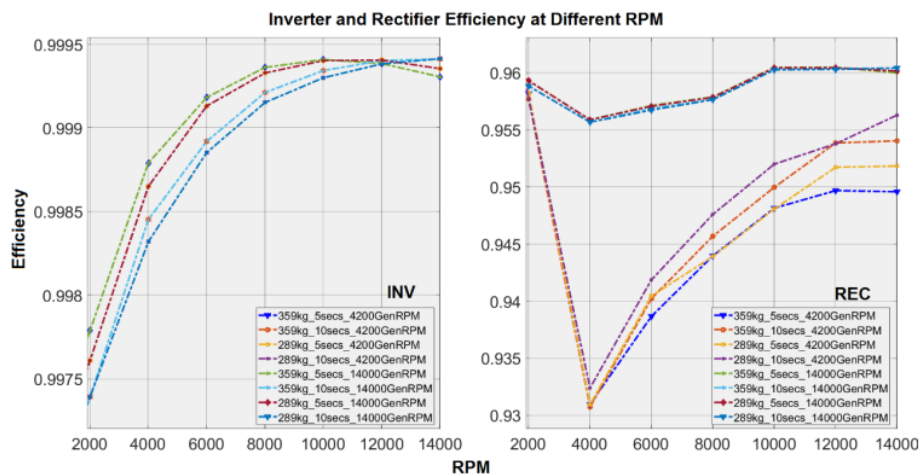


Figure 89 – Inverter and Rectifier Efficiency Profile

A trend of increasing efficiency with motor RPM is observed for the inverter (left graph). Configurations for a set ramp time of 5 seconds are seen to have slightly better efficiency characteristics. The rectifier (right graph) exhibits some irregular patterns in its efficiency variation for different configurations with increasing motor speeds. A sharp decrease is noticed for the lower generator speed configurations, especially at rated motor speed of 4000RPM. This is caused by a doubling of the filter capacitors for voltages higher than 700V as mentioned earlier. This results in an increase in number of devices contributing to the power

loss. It is also noticeable that profiles coinciding with a 14000RPM generator speed demonstrate better efficiency.

Figure 90 shows the thermal profile of the inverter for the full mission duration. A fault induced temperature increase can be seen at 3000s. The magnitude of the momentary increase is seen to exceed a temperature of 140°C for a configuration where the motor is operated at 14000RPM. This is however, below the limit of 150°C. The temperature rise due to fault handling was observed to be lower for lower motor speed configurations except for the 1920RPM and 4000RPM. Also, a significant temperature increase is observed at the end of the mission. This is due to the rapid increase in ambient temperature as the aircraft descends. However, the applied cooling is seen to be enough for all stages of the flight.

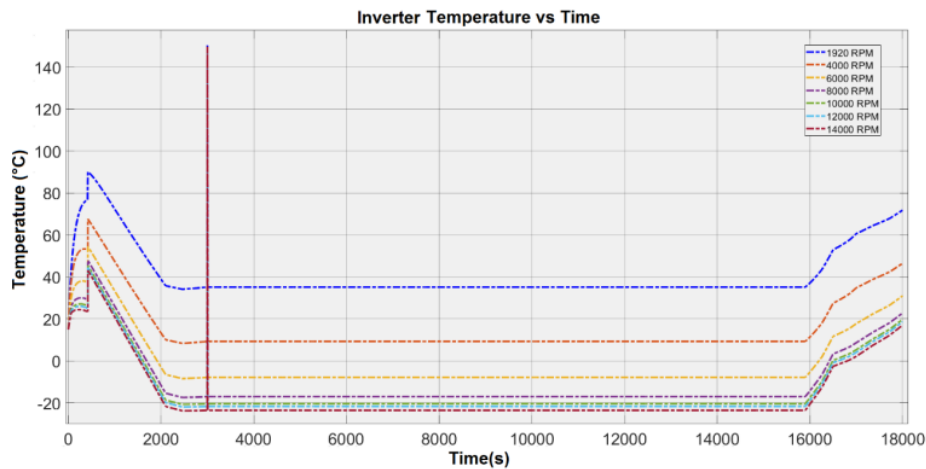


Figure 90 – Inverter Temperature Variation at different Rated RPM

Figure 91 shows the rectifier thermal behaviour for the investigated duration. Similar trend as the inverter switch is also observed for the rectifier switch at the descent. The capacitor temperatures are seen to be more impacted by the rapid change in ambient temperature. The overall rectifier temperatures are within acceptable values. The performance of the alternate rectifier is assumed to be same since they operate in a twin configuration, with each handling half of the system power.

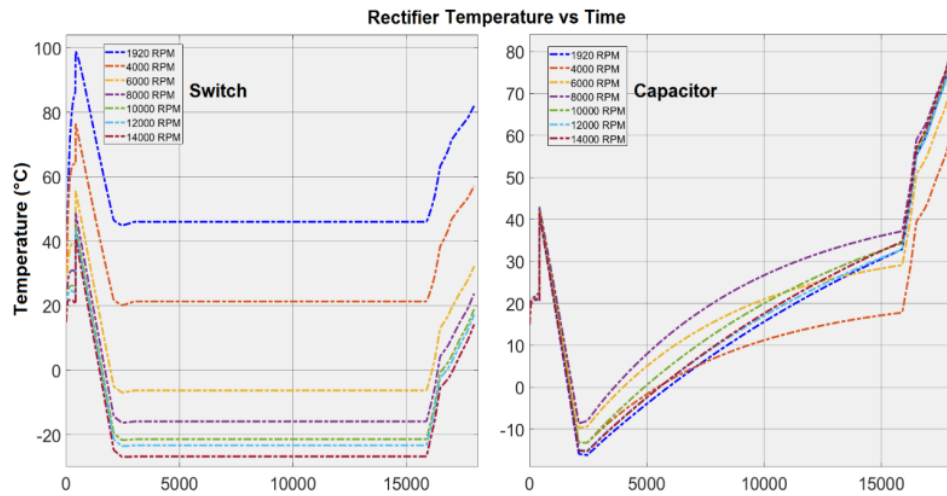


Figure 91 – Rectifier Temperature Variation at different Rated RPM

Figure 92. shows the circuit breakers efficiency profile. *CB3* and *CB4* are not shown because they are a replica in size and performance of *CB1* and *CB2* respectively. *CB1* is a DC circuit breaker while *CB2* is an AC breaker. Like the inverter and rectifier, the circuit breakers efficiency is seen to increase with increasing RPM for all fan mass and ramp profiles.

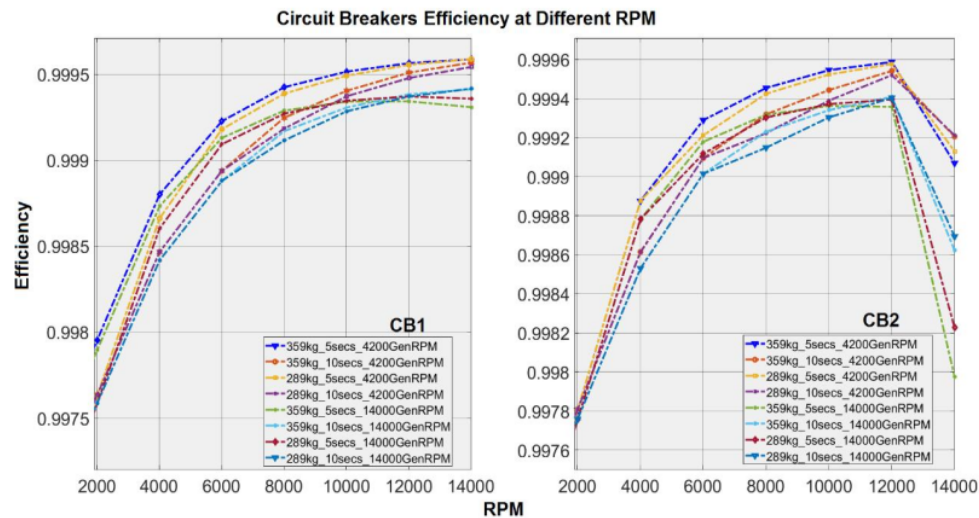


Figure 92 – Circuit Breakers Efficiency Profile

CB1 has broadly four efficiency profiles, mapped either to the ramp times or the two generator RPMs. This is as a result of the switching frequency selected to be related to the generator frequency. With constant machine pole count, higher RPM implies higher frequency and more switching losses. This explains why the circuit breakers in configurations mapped to a 14000RPM generator speed, have lower efficiencies. *CB2* also exhibit the four categories, with a drop occurring for a configuration where the motor speed is 14000RPM. This is a result of switch doubling in the series connection branch for higher than 1200V, as explained earlier. The effect of imposed voltages also appears to have more effect on *CB2*, owing to more efficiency variation with increasing motor speed configurations.

The thermal profiles of the circuit breakers are presented in Figure 93. Only the switch temperatures are shown because they are continuously active.

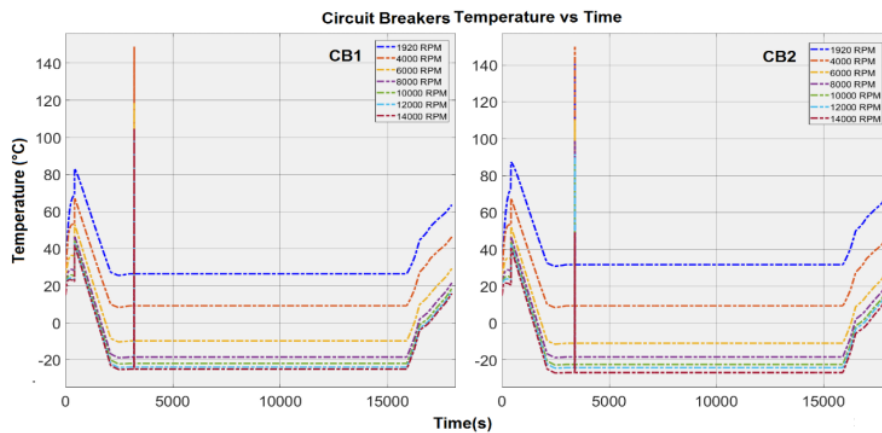


Figure 93 – Circuit Breakers Temperature Profile

CB1 and *CB2* operate within acceptable temperatures in normal operation but experiences a thermal surge while handling the fault. Unlike in the inverter, the thermal response to fault handling is seen to be a bit constant across motor speed configurations.

6.5.3 Transmission Cables

The efficiency profile of the transmission segments is presented in Figure 94. Again, *TXN4* and *TXN5* have been omitted since they are like *TXN2* and *TXN3*

respectively. There is little variation in the efficiency trends of the transmission segments, despite changing fan mass and ramp up speeds. However, there are two categories for *TXN3*, corresponding to the different generator speeds. *TXN2* carries DC voltage and its efficiency is reliant mostly on the core resistance. However, *TXN1* and *TXN3* carry AC voltage and their efficiency rely both on the resistance and reactance, which are functions of the core diameter and frequency. As shown in Figure 94 a general trend for decreasing efficiency is observed with increasing motor speed for *TXN1*, while the reverse is seen for *TXN2* and *TXN3*.

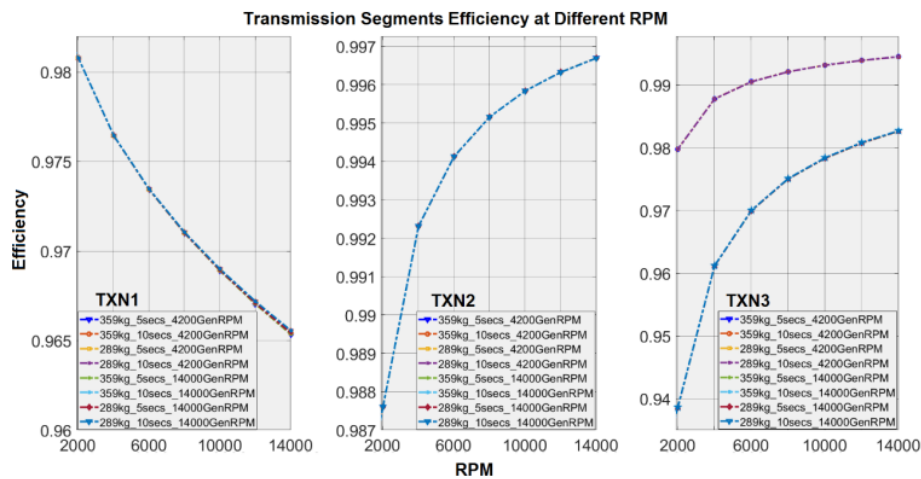


Figure 94 – Transmission Segments Efficiency

Figure 95 shows the temperature trend for the three transmission segments. The temperatures at normal and fault conditions are within acceptable limits. The recommended limits for normal operation is 90°C, while for fault condition is 250°C[105]. The transmission lines show a constant temperature distribution compared to the other devices. This is because they are not cooled by the very cold ambient and so rely more on their heat capacities at a selected constant ambient of 21°C. The temperature rise in reaction to fault is also observed to be within acceptable limit.

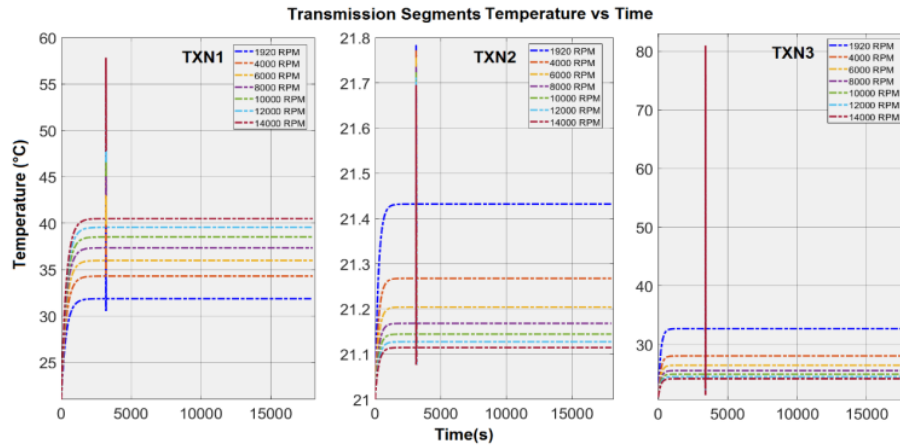


Figure 95 – Transmission Segments Temperature Variation with RPM

Figure 96 show the temperature for all devices at start-up of the system. The effect of the inrush on the thermal behaviour of the system is seen to be minimal as no device exceeded its limit.

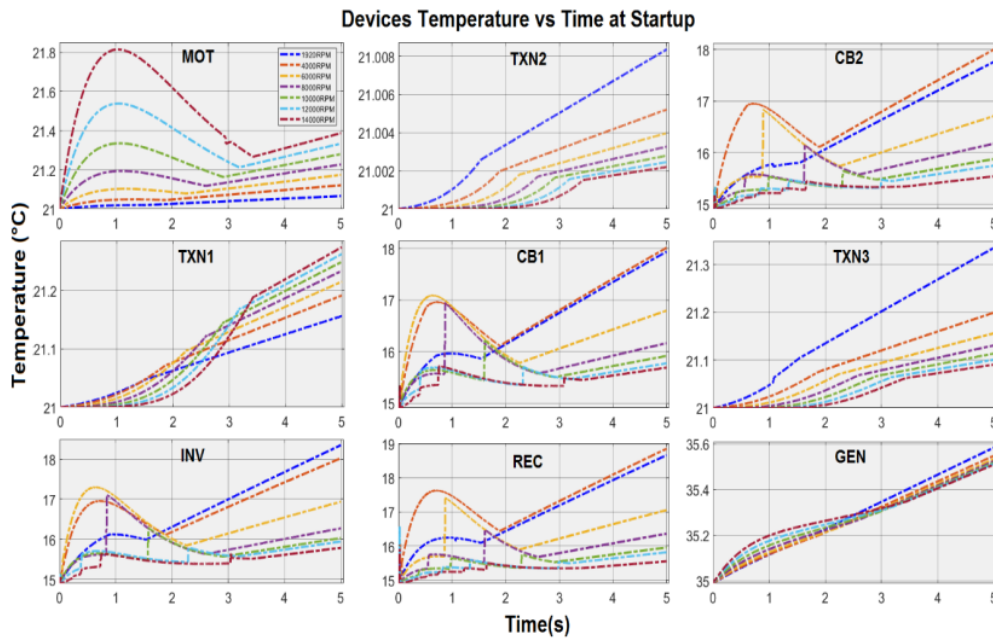


Figure 96 – Devices Temperatures at System Start-up

6.6 Voltage Selection

The efficiency, temperature and architecture mass presented so far are an exposition of the design space for this airframe concept. However, the existing literature aligns more to the selected default configuration in terms of speed specifications of the electrical machines. Voltages have also been derived for other configurations investigated. It should be noted that all devices can rarely be operated at the exact same voltage, since voltage drops across every device as it is transmitted from source to load. Therefore, the difference between the source and load operating voltage depends on the extent of overall system inefficiency which causes the voltage drop. Hence, it is referred to as voltage level. For all configurations investigated, the derived motor voltage is imposed on the entire propulsion system. Such imposition may not yield the optimal mass or efficiency in some cases. In this section an attempt is made to estimate an optimal operating voltage for the considered aircraft concept. This is based on the two main approaches presented in Chapter 4.

6.6.1 Motor Optimized Approach

The motor optimized voltage selection refers to the voltage derived for the design speed of the motor and generator, based on fan thrust requirement and turbine spool speed. This has been covered in the earlier part of this chapter and is represented by *359kg_5secs_4200GenRPM* at a motor rated speed of 1920 RPM.

6.6.2 Architecture Optimized Approach

The purpose of performing an architecture optimized sizing and voltage selection is to explore further areas of merit, especially, where certain components have a potential to be lighter and more efficient while performing at a different voltage level. The architecture optimized voltage selection refers to all other configurations which aren't the motor optimized. Most of which have been presented too. However, since these refer to configurations outside of the thrust specifications, they require an adaptation device(s) to be accomplished. For instance, in order to run the motor at speeds other than 1920RPM, a gearbox is

required. Same applies when it is intended to run the generator at speeds other than the LP turbine spool speed.

This subsection attempts to optimize the mass and efficiency of the default configuration by retaining the shaft speeds of the electrical machines to match those specified for the airframe configuration. This is achieved by the accommodation of two voltage levels, letting the generator and motor operate at voltages optimized to their speeds. Hence, the motor voltage is not imposed on the generator.

In order to achieve this electrically, a buck-boost converter has been considered, to achieve the adaptation by stepping up or down voltage according to individual design requirements. The same default configuration of *359kg_5secs_4200GenRPM* at 1920RPM motor speed is considered. For this arrangement, the rectifier is sized smaller by excluding the filter capacitors. This is made feasible by, the availability of input and output filter capacitors on the introduced buck-boost converter. Figure 97 shows the layout as proposed for the implementation of the optimization. The Converters has been positioned next to the inverter. Two units are required, to receive input from the two DC power sources.

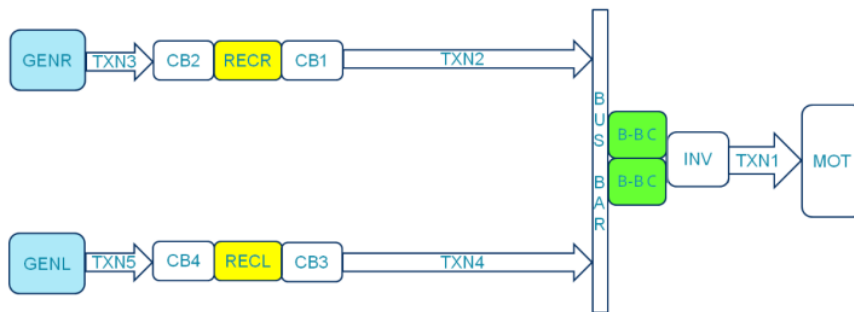


Figure 97 – Modified Layout for the Architecture Optimized Voltage Selection

Other important function of the introduced converter is its ability to limit the inrush current handling to the converter and to manage the changes in disparity of voltage levels of two power sources, arising from the speed variation in the turbine spool.

Figure 98 shows the estimated mass and efficiency of the devices sized for this configuration.

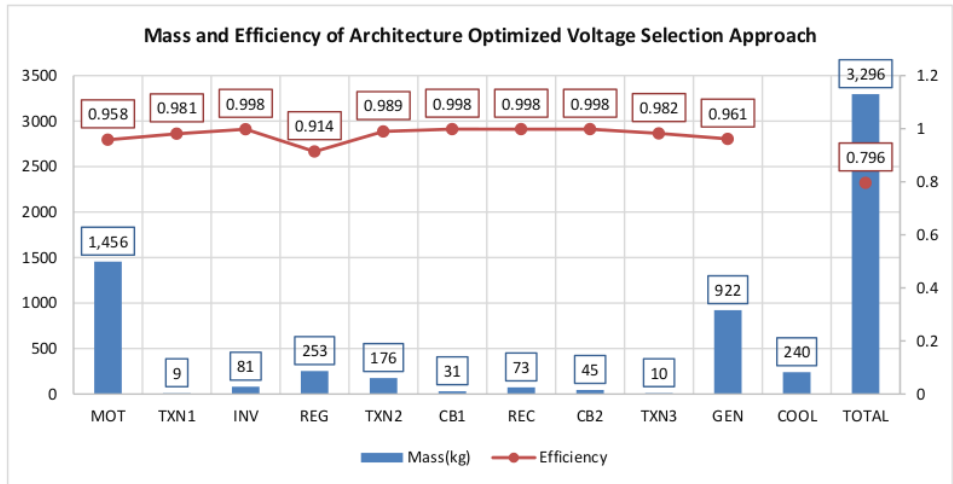


Figure 98 – Components Mass and Efficiency for Architecture Optimized Configuration

An investigation of the thermal performance is also performed for normal and fault operation. Figure 99 shows the electrical machines temperature change for the considered hours.

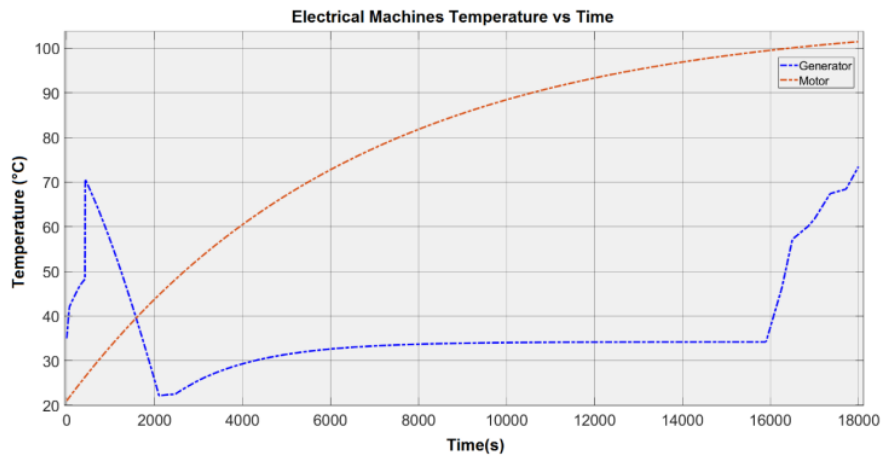


Figure 99 – Electrical Machines Temperature

These are seen to be within acceptable limits and lower than most of the motor optimized configuration. The buck boost converter thermal characteristics are presented in Figure 100. A sudden rise is also noticed in the temperature of the input capacitor due to fault at the upstream part of the network.

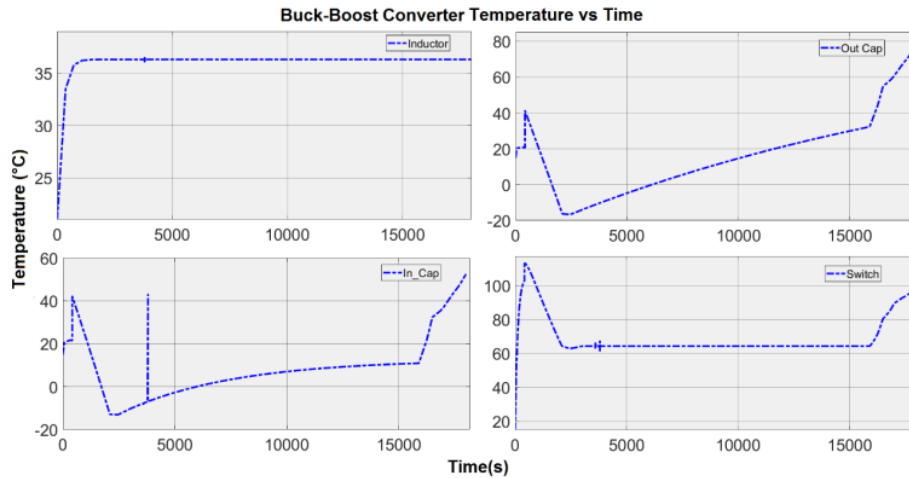


Figure 100 – Buck Boost Temperature Profiles

Only minimal effect of this fault is seen in the temperatures of the switch and the inductor. The inductor exhibits about 14°C temperature rise for the mission, while the output capacitor experiences a continued rise and required the heat sinks to be further optimized to accommodate this rise.

Figure 101 shows the temperature trends for the circuit breakers and rectifier considered in this configuration. The circuit breakers exhibit similar temperatures to a matching motor optimized configuration, both at fault times and normal steady state operation. The rectifier also exhibits slight temperature reaction to faults and performs normally all through the investigated mission duration.

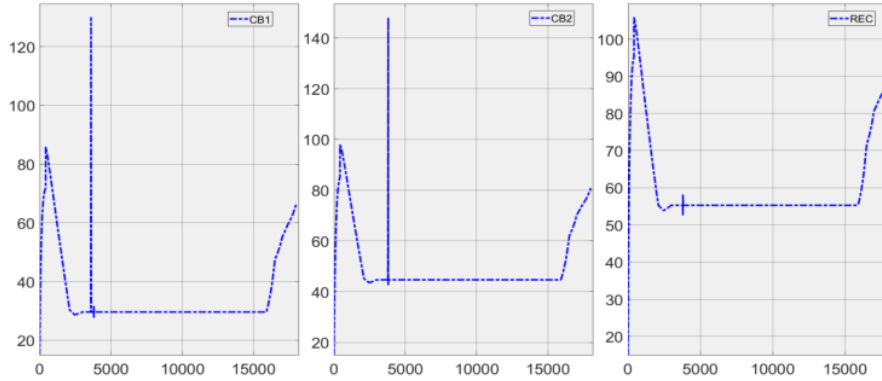


Figure 101 – Rectifiers and CBs Temperature

The transmission line temperatures are shown in Figure 102. These are shown to be within acceptable margin both in normal and fault operation.

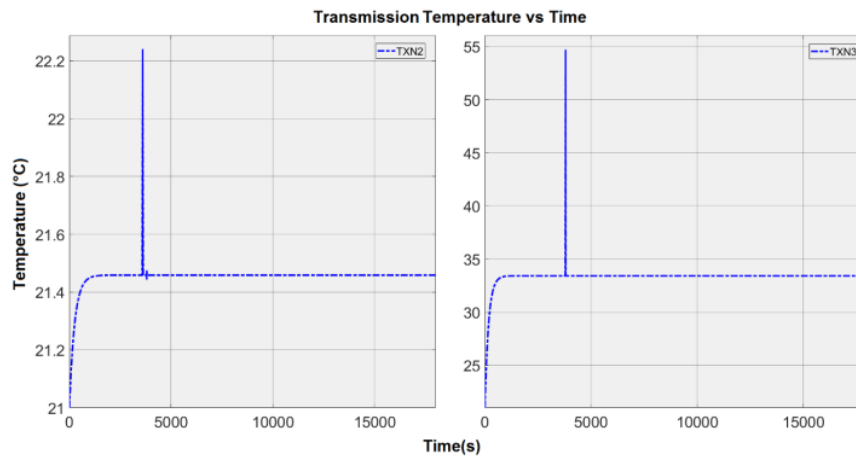


Figure 102 – Transmission Line Temperature

6.6.3 Comparison of Motor vs Architecture Optimized

A side by side comparison of the investigated voltage selection procedures is shown in Figure 103.

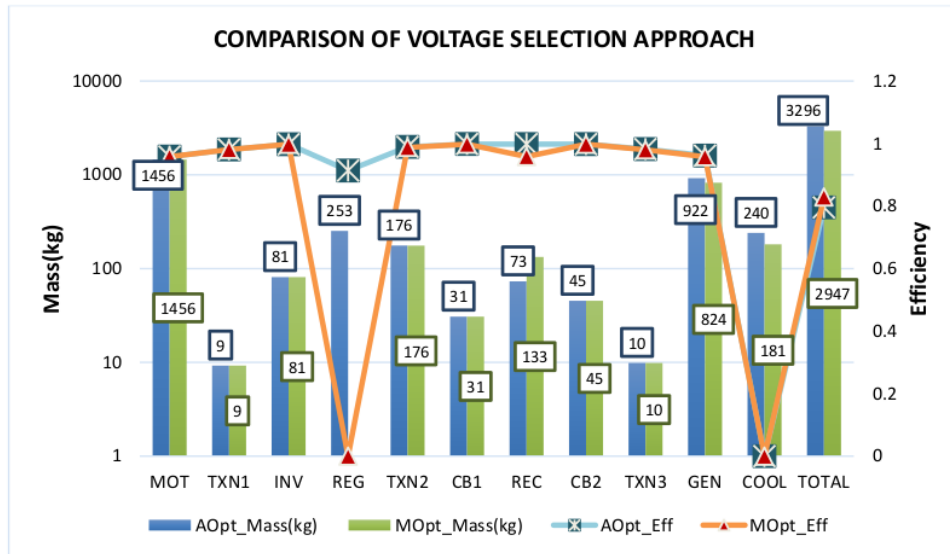


Figure 103 – Comparison of Voltage Selection Design Outcomes

A significant difference of about 350kg is seen, with the motor optimized configuration having better mass and efficiency. This is due to the introduction of the buck boost converter and significant difference in the generator and cooling mass. The generator in the architecture optimized configuration has a higher diameter compared to the motor optimized. The developed sizing procedure relied on a preset length to diameter ratio computed from the volume. This value is based on the referenced SOA machine and was not used in the motor optimized configuration. Hence the diameter had to be used as a design handle since the voltage was imposed. The increase in cooling mass is associated with the voltage regulator, which contained components that require cooling.

For the considered default configuration, the mass penalty resulting from an attempted optimization of the generator and transmission mass, clearly outweighs the benefits of optimal voltage selection. However, this may not be the case for all aircraft configurations, as it could also be noticed from Figure 103 that the rectifier is significantly lighter, indicating a possible tilt in system mass profile, especially at other voltages.

6.7 Results Discussion

The results presented in the preceding section of this chapter were aimed at reflecting the impact of transient, fault and thermal characteristics consideration on the design space for a given aircraft configuration. Figure 104 shows a summary of total mass and efficiency for all the investigated motor optimized architectures. A comparison is made with previously presented mass and efficiency estimates in [61], where no details of faults and thermal limits were presented along with the sizing. This is shown by the straight red line.

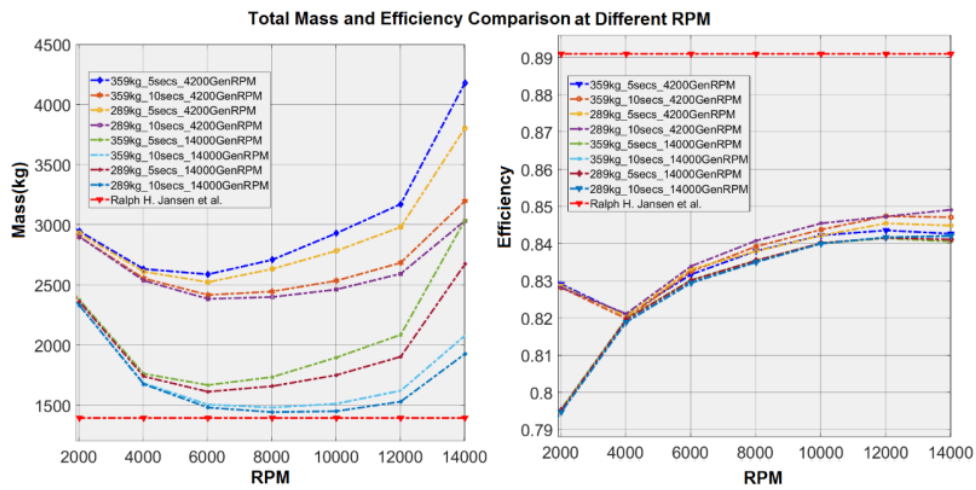


Figure 104 – Design Space Mass and Efficiency Comparison

A significant departure in total systems mass from the proposed values can be seen, due to incorporating the transient, faults and thermal concerns in the design process. The closest similarity was found to require 10seconds of ramp time and the generators be run at 14000RPM. These would require deeper feasibility and regulatory concerns to be addressed. The presented efficiency was also significantly higher.

Further comparison is made for total architecture mass, computed with less consideration to transient and fault characteristics. This is similar to attempt presented in [23] for a hybrid electrical system. There, a multiplier of 2 times the system's current rating, was used to arrive at the maximum possible current in a

system. The results obtained for such an assumption is presented in Figure 105 only for the default configuration. This is because such an approach does not consider fan mass, ramp time or generator speeds, which are the factors necessitating the investigation of the alternate configurations.

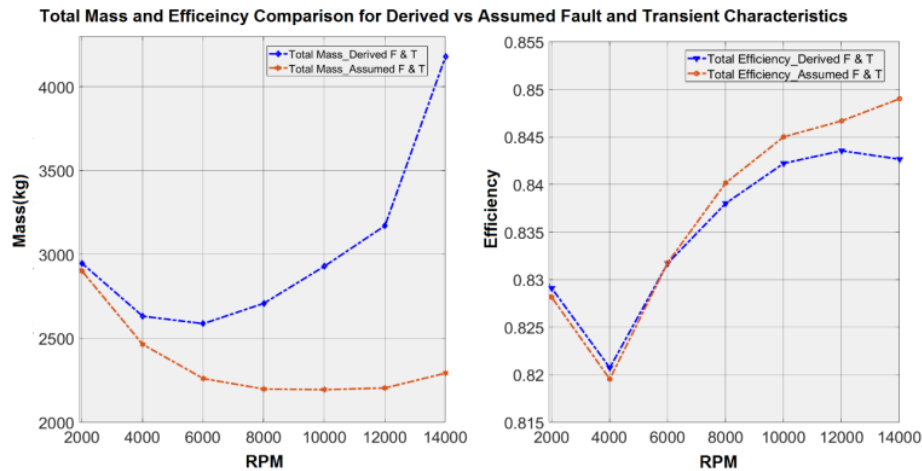


Figure 105 – System Mass Comparison for Derived vs Assumed Fault and Transient Values

Once more, it is observable that the improper consideration of these feasibility aspects would lead to the derivation of optimistic system masses. The trend of disparity is more with increasing motor speeds, both in terms of mass and efficiency.

In terms of cooling, the maximum mass derived in this work based on the novel cooling method is about 268kg. This is less than the 470kg cooling mass proposed in [61] for same aircraft size. A mass comparison is attempted for the designed cooling system with similar cooling strategy proposed in [39]. There it was stated that the mass of the cooling system was about 20% of the cooled device mass. Also, the total cooling mass presented in in [23], was shown to be between 21% to 23% of the total system mass. Figure 106 shows the ratio of the total cooling mass to system's mass.

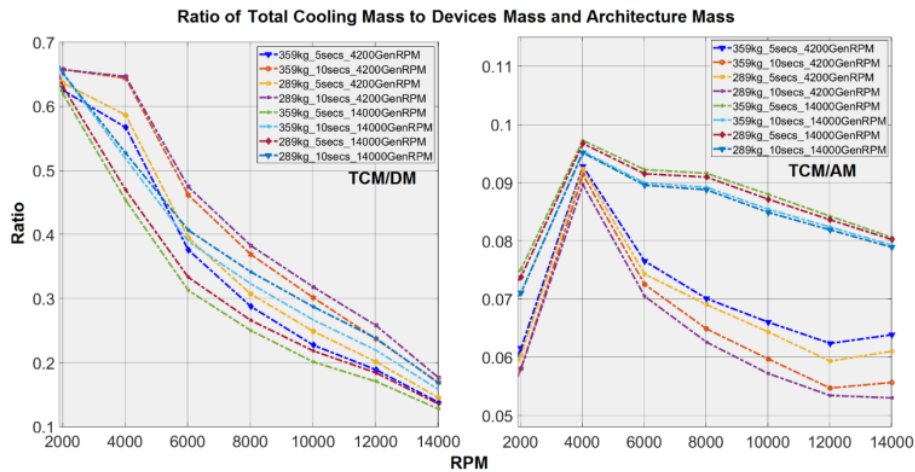


Figure 106 – Cooling Mass Comparison with Devices and Architecture Mass

The left graph is a ratio of the total cooling mass to the total mass of all the cooled devices (TCM/DM) while the right graph is a ratio of the total cooling mass to the total architecture mass (TCM/AM). Only above 10000RPM motor speed configuration was the total cooling mass equal or less than 20% of the total cooled devices' mass. However, from a total system mass point of view, the proposed cooling system was found to be less than 10% of the total systems mass for all investigated cases.

6.8 Summary

This chapter presented an application of the earlier proposed methods to the design feasibility of a turboelectric aircraft concept. The load dynamics is seen to have more impact on the mass and efficiency of the power electronics device than it has on the cables and electrical machines. This is because for power electronics, the sizing was dependent on the switch count per device, which was decided by the motor surge currents. The increase in total system mass was also seen in most cases, to be directly proportional to increasing motor speeds.

Fault current values and protection device mass at different points on the network was also observed to be significantly different. This implies that earlier proposed methods of sizing protection system based on power rating or simple multiples of

the rated current would be insufficient. Such approaches would yield same mass for all protection devices for a given aircraft configuration, leading to an over optimistic size estimate. For the same power rating, the faults closer to the power sources were seen to have higher values, with DC branches of the network showing the highest magnitudes.

The thermal behaviour of the devices has also been investigated, both for normal and fault conditions. For the investigated cases, AC fault paths had more temperature rise during faults than DC. The circuit breakers' temperatures were near absolute maximum ratings in some cases, despite the faults occurring at a time of negative instantaneous temperature value. There also appears to be a rapid temperature increase for most devices at the final stages of the mission due to rapidly increasing ambient temperature.

From the performed investigation, the default configuration (*359kg_5secs_4200GenRPM*) at motor speed of 1920RPM appears to be the most promising, based on existing regulations and thermal behaviour.

7 CONCLUSION AND FUTURE WORKS

7.1 Conclusions

From the performed investigation, one major feasibility challenge with future aerospace propulsion design is the very high transient currents and the associated thermal dissipation. The earlier approaches to protection system sizing using specific power scaling[61] and multiples of rated current[23] did not suffice, as the fault currents are more a function of the line impedances than they are of the circuit current or power ratings. More so, the value of fault currents was found to be between 9.5 to about 100 times the rated current. This appears to be traceable to the very low impedance paths and short distance between power source and fault points. However, it was possible to achieve smaller protection device sizing since the faults could be arrested before the peak values. The selected time used was when it reaches 10% above the maximum transient current value.

Like with electrical machines, power electronics-based devices come in a myriad of configuration even for the same function. This implies several sizing approaches in pursuit of aerospace propulsion train. However, these all rely on the operational details of switches, inductors and capacitors. These were the basis of the sizing performed in this research, together with their fault and thermal characteristics. The trend of behaviour shows the highest mass penalty was incurred by devices which include capacitors. The switches dissipate the most heat, despite boosting the highest efficiencies. This is related to their very low thermal capacitance compared to their thermal resistances.

The developed cooling approach can maintain the devices' temperatures below their limits. It was required to perform several rounds of optimization for the entire mission duration to achieve the final cooling mass. The mass ratios of the cooling system to the cooled devices was between 5% and 65%. However, when compared to the total system mass it was less than 10%.

The *motor optimized* voltage selection approach appears to be the better option, since it eliminates a device. This offers less mass penalty and improved efficiency gains but less robustness.

Based on thrust and power requirement, the design feasibility of the electric part of a turboelectric propulsion system, appears to be hinged more towards the power electronics thermal limits, as these raise reliability concerns. Targeting lower temperatures at faults seems to be of foremost concern, to prevent degradation over time or catastrophic loss of the system. Having a cooling system with the required instantaneous response seems difficult. Therefore, the sized system should be able to withstand this temperature rise with normal cooling. The specific configuration of the aircraft also plays an important role as it decides the size of the fan. The size of the fan is seen to impact the electrical system size through the RPM, which plays a great role in the voltage selection and transient current magnitude.

A summary of the core findings of this research is presented below;

1. The switch counts for protection devices were decided by the magnitude of surge current during motor transients. More switches imply less loading per switch and better thermal profile.
2. Operating below the switch voltage capability significantly reduces the system mass, as anything above would double the switch count and the device mass, even for the same power.
3. The overall efficiency of the entire system only reaches about 85% for all investigated configurations.
4. The effect of transient and thermal consideration on architecture mass is more significant in designs with high motor speeds ($>2000\text{RPM}$). The optimal size with such consideration was observed at motor speeds between 4000RPM and 8000RPM , with best cases in between.
5. The motor voltage rating can be safely adopted as the operational voltage of a turboelectric propulsion system.
6. There is a great potential for cooling power electronics devices for a propulsion system using a combination of heat sinks and high velocity air in

the fan duct. This was based on experimental validation, which showed a good agreement with the presented theories and could be relied upon for scaled investigations.

7. Power flow integrity also appears to be an important consideration. It was noticeable in the experimental process that balancing and combining multiple power sources to drive the same load could be challenging, especially at higher power ratings with load fluctuations.

The following are the contributions to existing knowledge made by this research.

1. A procedure for the selection of optimal operational voltage for a turbo electric propulsion system has been proposed. This is based on the required motor flux linkage to deliver the rated torque and speed. The impact of this contribution is a more optimal propulsion system design which is based on the thrust requirement.
2. An approach for turboelectric components sizing and efficiency estimation considering the devices thermal limits has been proposed. Correlation of simulation and experimental data, obtained from the test rig developed within the context of this research, shows the correctness of the modelling approach. Hence, a detailed thermal characterization of turboelectric network has been performed and validated.
3. An approach to transient current determination for turboelectric network system has been proposed, based on the load inertia and imposed duration from idle to full thrust power. The impact of this contribution is the correct sizing of the drives to accelerate the motor and the fan without oversizing or under sizing. Also, it enables correct system settings to discriminate between fault currents and system transient currents, in order to perform an early interruption.
4. An approach to fault current determination for turboelectric network system has been proposed. This was based on faults along the transmission lines. The knowledge of fault current magnitude is important for appropriate protection system sizing.
5. A novel method has been proposed for the implementation of propulsion system electronics cooling. This is based on sinking the dissipated heat into

the propulsor duct flow. The method demonstrates good potential to solve the mass and complexity challenge posed by existing cooling methods and could therefore be extended to large aircrafts for the future.

7.2 Future Works

- For the proposed cooling method, more investigation would be required on its feasibility especially regarding the impact on the aircraft thrust, aerodynamics and overall efficiency both at design and of design points. Also, the effects of degradation and airborne contaminants on the cooling fins need to be investigated. Further optimization techniques need to be derived for the heat sinks, with due consideration to their shape, geometry, material and the flow characteristics. The possibility of using the duct stators simultaneously as part of the heat sink also needs to be investigated further.
- The fault current estimation approach presented, is adaptable to a bus topology network. The possibility of applying part or whole of the approach, to propulsion systems involving other topologies needs to be investigated. It is also required to investigate further the level of interconnection that would be optimally feasible for an aircraft propulsion system. Several challenges are inherent from such interconnection in pursuit of system redundancy benefits. These may complicate the power flow, resulting in losses and lead to further fault handling complications due to several possible flow paths.
- The ability to deploy the inverter drive simultaneously as a circuit protection device offered significant mass reduction. It would be necessary to investigate the possibility of reducing the power electronics devices by same merging functions approach.
- Owing to the significant mass and inefficiency contribution of capacitors, investigation is required to see the possibility of designing a propulsion system which could operate without them. This would offer more feasibility to the deployment of turboelectric propulsion systems.
- The proposed voltage selection approach is for a configuration where the motor shaft is directly connected to the fan. Where this is not the case as

with the inclusion of a gearbox for higher speed adaptation, an adjustment of the proposed approach would be required. The effect of this adjustment on optimal sizing also needs to be investigated.

- The generalization of the design approaches presented in this work does not entirely fit with hybrid propulsion systems. This is due to the difference between having batteries as power source and having a generator. The fault and transient current characteristics considering battery responses would need to be investigated. The thermal and stability impacts of battery voltage drop over the mission duration also need to be quantified.

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APPENDICES

Appendix A SiC MOSFET Data

Table 7-1 is a presentation of data for SiC MOSFETs from which a choice was made for this research. The data have been extracted directly from their published datasheets.

Table 7-1 – Details of Investigated SiC MOSFETs

	Voltage	V _{Isol}	Current	I _{sc} (A)	I _D pulse or I _{DM}	V _{GS} /V _{DS}	V _F	I _F	V _{F0}	R _{DSon}	R _{th_JH}	E _{on}	E _{off}	Mass	E _{rec}	T _{jmax}
Model	(V)	(V)	(A)			(V)	(V)	(A)	(V)	(Ohm)	(K/W)	(mJ)	(mJ)	(g)	(J)	(°C)
FF45MR12W1M1_B11	1200	3000	25	210	50	3.45	4.6			0.045	1.54	0.3	0.033	24		150
FF23MR12W1M1P_B11	1200	3000	50	420	100	3.45	4.6			0.0225	0.748	0.49	0.094	24		150
FF23MR12W1M1_B11	1200	3000	50	420	100	3.45	4.6			0.0225	0.8	0.49	0.094	24		150
FF11MR12W1M1_B11	1200	3000	100	840	200	3.45	4.6			0.0113	0.553	1.4	0.0647	24		150
FF8MR12W2M1_B11	1200	3000	150	1260	300	3.45	4.35			0.0075	0.346	2	0.76	39		150
CAS120M12BM2	1200	5000	193		480	1.8	1.5			0.013	0.125*	1.7	0.4	290		150
FF6MR12W2M1_B11	1200	3000	200	1680	400	3.45	4.35			0.00563	0.328	2.5	1.2	39		150
SK250MB120SCTE2	1200	2500	249		960	4.05	4.05		2.6	0.0075	0.15	6.8	4	35	0.0025	150
CAS300M12BM2	1200	5000	423		1500	1.8	1.6			0.0042	0.07*	5.8	6.1	300		150
CAS325M12HM2	1200	1200	444			2	1.6			0.0037	0.1*	5.6	3.7	140	0.0028	170
CAB450M12XM3	1200	4000	450		900	1.8	4.2			0.0026	0.11*	11	11.3	175	0.0002	175
CAS300M17BM2	1700	5000	325		900	1.8	1.7	556		0.008	0.067*	13	10	300		150
SKM350MB120SCH15	1200	4000	523		1280	6.4	6.4		3.9	0.0056	0.08	10.3	4.7	325		150
SKM350MB120SCH17	1200	4000	523		1280	1.6	1.4	220	0.95	0.0056	0.075	8.65	7.98	325		150
SKM500MB120SC	1200	4000	541		1920	1.6	0			0.00375	0.105	10.3	4.7	325		150

* Junction to Case

Appendix B Inductors Core Data

Table 7-2 – Details of Investigated Inductor Cores

Table 7-2 contains specification of the inductor cores considered for this research. The information have been extracted from their individual datasheets obtained from [143].

Sample Number	Core type	Permeability	Watt Loss @ 100 kHz max (mW/cm ³)	Voltage Breakdown wire to wire min (VAC)	mass (g)	OD max (mm)	ID min (mm)	Height (mm)	Volume (mm ³)	Path Length (mm)	Cross Section (mm ²)
0	High Flux	14	2600	1000	2.3	13.5	6.99	5.51	365	31.2	11.7
1	High Flux	14	2600	1000	5.4	17.3	9.53	7.11	806	41.2	19.56
2	High Flux	14	2600	1000	6.2	18	9.02	7.11	960	41.4	23.2
3	High Flux	14	2600	1000	7.1	21.1	12.1	7.11	1194	50.9	23.47
4	High Flux	14	2600	1000	12	23.6	13.3	8.4	1863	56.7	32.85
5	High Flux	14	2600	1000	15	24.3	13.8	9.7	2281	58.8	38.8
6	High Flux	14	2600	2000	27	27.7	14.1	11.9	4322	65.4	66.1
7	High Flux	14	2600	2000	35	33.8	19.3	11.4	5509	81.4	67.7
8	MPP	14	2600	2000	28	35.2	22.6	9.8	4215	89.5	47.1
9	High Flux	14	2600	2000	39	36.7	21.5	11.4	6136	89.8	68.3
10	High Flux	14	2600	2000	70	40.8	23.3	15.4	10880	98.4	110.6
11	High Flux	14	2600	2000	96	47.6	27.9	16.1	15502	116.3	133.3
12	High Flux	14	2600	2000	140	47.6	23.3	18.9	21235	107.4	197.7
13	High Flux	14	2600	2000	110	51.7	30.9	14.4	15832	127.3	124.4
14	High Flux	14	2600	2000	140	58	34.7	14.9	20918	143	146.3
15	High Flux	14	2600	2000	180	58	25.6	16.1	28625	125	229
16	High Flux	14	2600	2000	348	62.9	31.7	25.9	51800	144	360
17	High Flux	26	900	2000	345	69.4	34.7	21.4	49700	158	314
18	MPP	14	2600	2000	565	78.9	38.3	26.8	81500	170	478
19	High Flux	14	2600	2000	290	78.9	48.2	13.9	34500	196	176
20	High Flux	14	2600	2000	339	78.9	48.2	17.1	43400	196	221
21	High Flux	14	2600	2000	570	103	55.8	17.9	86900	243	358
22	High Flux	14	965	2000	1300	134	77.2	26.8	220000	324	678
23	High Flux	26	625	2000	2600	166.5	101	33.15	407000	412	987

Appendix C ABC-DQ Transformation

The transformation of a design voltage V into the direct axis component V_d and quadrature axis component V_q , can be achieved using equations (7-1) to (7-4)

$$V_A(t) = V \cos \omega_e t \quad (7-1)$$

$$V_B(t) = V \cos(\omega_e t - 120^\circ) \quad (7-2)$$

$$V_C(t) = V \cos(\omega_e t - 240^\circ) \quad (7-3)$$

$$\begin{bmatrix} V_d \\ V_q \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos \theta & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ -\sin \theta & -\sin\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) \end{bmatrix} \times \begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} \quad (7-4)$$

Appendix D Top View of Design Tool

The model presented in Figure 107 was utilized to perform the investigation and generate the results presented in Chapter 6. The blue enclosure was used for the motor optimized voltage approach, while the green enclosure was used for the architecture optimized voltage approach. Only the MOT, TXN1 and INV handle full power, all other devices handle half power and their results have to be doubled to represent the full system power.

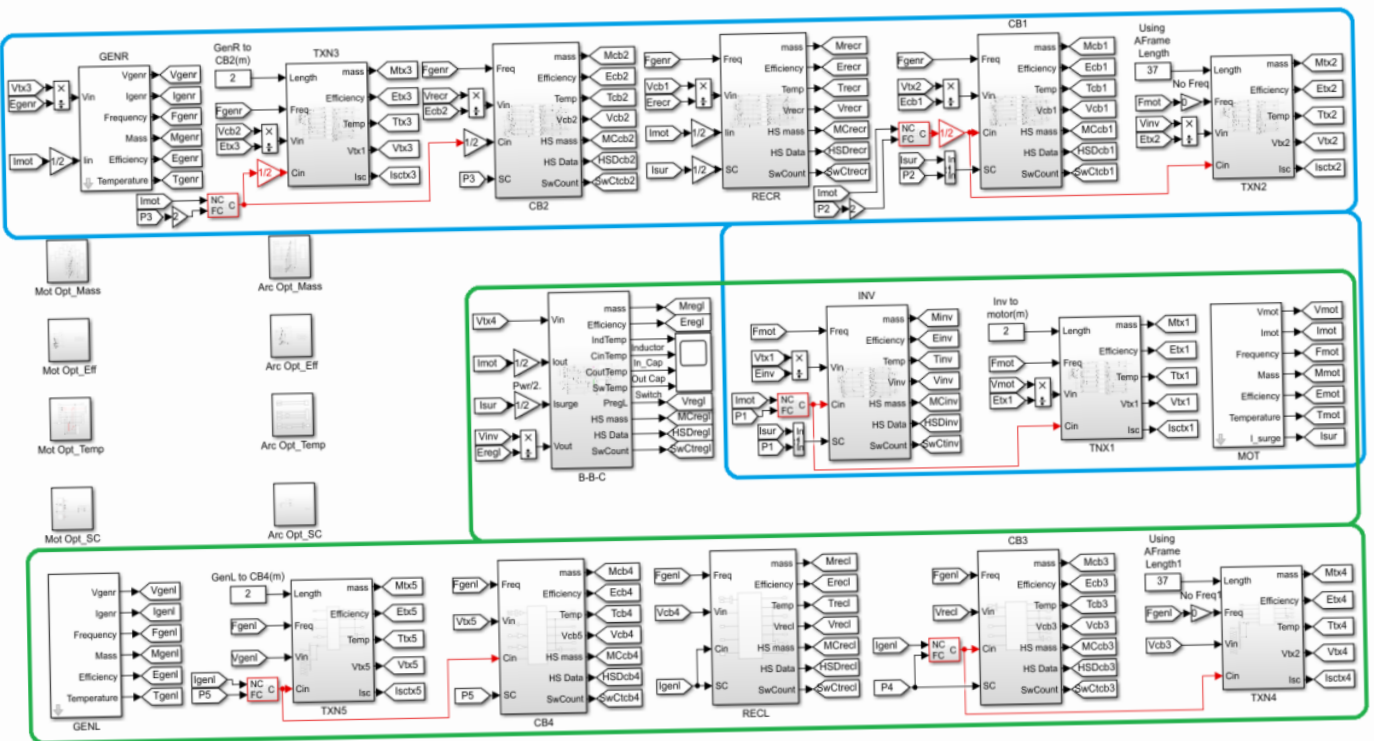


Figure 107 – Top View of the Simulink Model Utilized For This Research