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ESCUELA TÉCNICA SUPERIOR DE INGENIEROS
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**Isolated Swiss-Forward Three-Phase
Rectifier for Aircraft Applications:
Analysis, Design and Validation**

TESIS DOCTORAL

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Máster en Electrónica Industrial, Universidad Politécnica de
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*To my parents
Magdalena y Sergio.*

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Abstract

In the last decades, there has been important interest in improving electrical system both in commercial and military aircrafts. These improvements have been driven by economic and environmental issues because replacing mechanic, pneumatic or hydraulic actuator by electric systems bring efficiency improvements reducing CO₂ emissions and flying costs. The tendency of replacing mechanical, pneumatic and hydraulic system by electric one is commonly called MEA or more electric aircraft. In addition, electric actuators are more reliable and require less costs in maintenance.

In aircraft applications, the electric power distribution is done using either three-phase or dc sources. Classically, the ac power line uses 115Vac and in dc the voltage used is 28Vdc. However in some of the modern aircrafts, the voltage levels have been increased to 230Vac and 270Vdc and 540vdc, in order to reduce the weight of the wiring. Following the MEA concept, this work analyses and develop a new converter topology of an isolated three-phase rectifier focusing in aircraft applications. As reference design for the analysis and development of an experimental demonstrator prototype, this work considers a three-phase main of 115Vdc @ 400Hz and 270Vdc for the output voltage and a load power of 3.3kW.

In this work, different solutions of an isolated three-phase rectifier have been studied including passive rectifiers, a two-stage topology or a single stage configuration. The two-stage topology can be assemble using an active rectifier with an isolated dc-dc converter or using a diode bridge plus an active power filter with an isolated dc-dc converter. On the other hand, in a single stage configuration both the ac input current shaping and the output voltage control are performed by the same converter leading to a less complex and more reliable and more efficient system. In this work, a single stage solution has been selected because this kind of configuration have several benefits in terms of efficiency, simplicity and power density.

There are several single stage isolated three-phase rectifiers in the power electronic literature, they can be classified in three groups: the DCM type, Matrix-type and Unidirectional family. In aircraft applications, the unidirectionally is needed because it is not allowed regenerate power back to the ac

side, for this reason Matrix-type rectifiers have been discarded. DCM rectifiers do not have current-voltage angle compensation therefore at light load these systems have a poor power factor. In this work, a new isolated three-phase rectifier that belong to the unidirectional family has selected. The Swiss Forward rectifier is an adaption of the novel non-isolated Swiss rectifier. In order to add the isolation feature in the system, a forward converter with resonant reset has been used because this topology allows to take advantage of the new semiconductor material such as the SiC.

In this work, the principle of operation of the Swiss-Forward rectifier is discussed as well as a converter modulation that allows the current-voltage angle compensation. In addition, in this work is discussed a reduce order model of a buck-type rectifier system is used to design of the control. Finally, a demonstrator prototype has been designed and the experimental results obtained with prototype confirm that the proposed rectifier topology is a good alternative to classical single stage topologies.

Resumen

En las últimas décadas, ha habido un gran interés importante en mejorar el sistema eléctrico tanto en aviones comerciales como militares. Estas mejoras han sido impulsadas por temas económicos y también ambientales ya que el reemplazar actuadores mecánicos, neumáticos o hidráulicos por sistemas eléctricos trae mejoras en eficiencia que reducen las emisiones de CO_2 y los costos de vuelo. La tendencia de reemplazar el sistema mecánico, neumático e hidráulico por uno eléctrico comúnmente se llama MEA o “aviones más eléctricos”. Además, los actuadores eléctricos son más fiables y requieren menos costos en mantenimiento.

En las aplicaciones aeronáuticas, la distribución de la energía eléctrica se realiza utilizando fuentes trifásicas o de corriente continua. Clásicamente, las líneas de alimentación en alterna utilizan 115Vac y continua la tensión utilizada es 28Vdc. Sin embargo, en algunos de los aviones modernos, los niveles de tensión han aumentado a 230Vac en alterna y 270Vdc o 540vdc en continua, con el fin de reducir el peso del cableado. Siguiendo el concepto de MEA, este trabajo analiza y desarrolla una nueva topología de un rectificador trifásico con aislamiento galvánico enfocado a aplicaciones aeronáuticas. Como diseño de referencia para el análisis y desarrollo de un prototipo demostrador experimental, este trabajo considera una tensión trifásica de 115Vdc a 400Hz y 270Vdc para la tensión de salida y una potencia de carga de 3,3kW.

En este trabajo se han estudiado diferentes soluciones de un rectificador trifásico aislado, incluyendo rectificadores pasivos, una topología de dos etapas o convertidores de una única etapa. La topología de dos etapas se puede montar utilizando un rectificador activo con un convertidor dc-dc aislado o utilizando un puente de diodo más un filtro de potencia activa con un convertidor dc-dc aislado. Por otra parte, en una configuración de una etapa, tanto el control de la corriente de entrada como el control de tensión de salida son realizados por el mismo convertidor que conduce a un sistema menos complejo, más fiable y más eficiente. En este trabajo, se ha seleccionado una solución de una etapa ya que este tipo de configuración tiene varios beneficios en términos de eficiencia, simplicidad y densidad de potencia.

Existen varios tipos de rectificadores trifásicos de una etapa con aislamiento

en la literatura electrónica de potencia, estos pueden clasificarse en tres grupos: el tipo DCM, el tipo Matriz y la familia de los Unidireccionales. En aplicaciones aeronáuticas, es necesario que los rectificadores trifásicos sean unidireccionales porque no se permite regenerar la energía hacia la parte alterna, por esta razón los rectificadores bidireccionales de tipo Matrix han sido descartados. Los rectificadores de DCM carecen de compensación de ángulo de corriente-voltaje en la entrada por lo tanto a la baja carga estos sistemas tienen un factor de potencia muy pobre. En este trabajo, se ha seleccionado un nuevo rectificador con aislamiento que pertenece a la familia de los unidireccionales. El rectificador Swiss Forward es una adaptación del nuevo rectificador Swiss no aislado. Para implementar el aislamiento se ha utilizado un convertidor forward resonante con el cual se puede aprovechar las ventajas que traen los nuevos materiales de los semiconductores como lo es el carburo de silicio o SiC.

En este trabajo, se discute el principio de funcionamiento del rectificador Swiss-Forward así como la modulación del convertidor que permite la compensación del ángulo de corriente-voltaje. Además, en este trabajo se discute un modelo de orden de reducido de un rectificador de tipo reductor para el diseño del control del sistema. Por último, se ha diseñado un prototipo demostrador en donde sus resultados experimentales obtenidos confirman que la topología de rectificación propuesta es una buena alternativa a las topologías clásicas de una sola etapa.

Abbreviations

<i>ABD</i>	Airbus Directives.
<i>AC</i>	Alternating Current.
<i>APU</i>	Auxiliary power supply.
<i>ATRU</i>	Auto Transformer Rectifier Unit.
<i>Batt</i>	Battery.
<i>BJT</i>	Bipolar Junction Transistor.
<i>CISPR</i>	International Special Committee on Radio nterference.
<i>CM</i>	Common Mode.
<i>CoolMOS</i>	Infineons silicon power MOSFET technology.
<i>DC</i>	Direct Current.
<i>DM</i>	Differential Mode.
<i>DSP</i>	Digital Signal Processor.
<i>EMC</i>	Electromagnetic Compatibility.
<i>EMI</i>	Electromagnetic Interference.
<i>FOC</i>	Field-Oriented Control.
<i>FOM</i>	Figure-of-Merit.
<i>FPGA</i>	Field Programmable Gate Array.
<i>GaN</i>	Gallium Nitride.
<i>GND</i>	Ground.
<i>HF</i>	High Frequency.
<i>IGBT</i>	Insulated Gate Bipolar Transistor.

<i>JFET</i>	Junction Field Effect Transistor.
<i>LF</i>	Low Frequency.
<i>LISN</i>	Line Impedance Stabilizing Network.
<i>MEA</i>	More Electric Aircraft.
<i>MIL – STD</i>	Military Standard.
<i>MOSFET</i>	Metal Oxide Field Effect Transistor.
<i>MTTF</i>	Mean Time To Failure.
<i>OP</i>	Operating Point.
<i>PC</i>	Personal Computer.
<i>PCB</i>	Printed Circuit Board.
<i>PE</i>	Protective Earth.
<i>PFC</i>	Power Factor Correction.
<i>PLL</i>	Phase Locked Loop.
<i>PWM</i>	Pulse Width Modulation.
<i>RAT</i>	Ram Air Turbine.
<i>RB – IGBT</i>	Reverse Blocking IGBT.
<i>RC – IGBT</i>	Reverse Conducting IGBT.
<i>RMS</i>	Root Mean Square.
<i>RTCA</i>	Radio Technical Commission for Aeronautics.
<i>SBD</i>	Schottky Barrier Diode.
<i>Si</i>	Silicon.
<i>SiC</i>	Silicon Carbide.
<i>SVM</i>	Space Vector Modulation.
<i>T&FS</i>	Trench and Field Stop.
<i>THD</i>	Total Harmonic Distortion.
<i>TRU</i>	Transformer Rectifier Unit.
<i>VSD</i>	Variable Speed Drive.

<i>VSI</i>	Voltage Source Inverter.
<i>VSR</i>	Voltage Source Rectifier.
<i>ZCS</i>	Zero Current Switching.
<i>ZVS</i>	Zero Voltage Switching.
CSR	Current Source Rectifier.
CSVM	Conventional Space Vector Modulation.

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Introduction

Creation is a bird without a flight plan.
It never flies in a straight line.

Violeta Parra

The Air traffic market has been constantly growing since the 60s. The average growth rate has been around 9% since 1960 and it is expected to continue growing at a rate between 5% and 7% until 2020 in just passenger traffic; meanwhile air cargo traffic will grow even faster[1].

Owing to the large number of flights that take place daily, today the air transport produces the 2% of the man-made CO₂ emissions and it is expected that in year 2050 it will be 3% [1]. It is claimed that saving 1 kilogram in each flight would save 1700 tons of burn fuel and 5400 tons of CO₂ per year for all air traffic [2].

In this context there are many commercial and environmental pressure on aircraft manufactures to improve the performance of future aircraft by reducing overall costs (design costs, operational and maintenance cost), fuel consumption and CO₂ emissions in future aircrafts. To tackle these goals, aircraft manufactures and governments have invested large amount of money in these last 2 decades in research and develop of new aircraft technologies, especially at the level of power distribution architecture. The current aircraft design trend indicates that aircraft will be increasingly more and more electric in the future due to the concept call "More-Electric-Aircraft

In the figure 1.1 the evolution over 50 years of total electric power demand in conventional aircraft is shown. As it can be seen the electric power demand has drastically increased in the latests aircrafts Airbus 380 and Boing 787.

This large increase in the electric demand in airplanes has been boosted by European financing over several years. Some of the most important projects

Figure 1.1: Electrical generation trends in large commercial aircraft in the last 5 decades [3]

are the following

European projects :

- 2000 MESA Magnetostrictive equipment and system for more electric aircraft
- 2002 POA Power optimized Aircraft
- 2004 MESENA Magnetostatic Energy System for even More electric Aircraft
- 2006 MOET More open Electrical Technologies
- 2008-2013 Clean Sky is the largest European research programme developing innovative, cutting-edge technology aimed at reducing CO₂, gas emissions and noise levels produced by aircraft (€1.6 billions)
- 2014-2016 Clean Sky 2 (€4.0 billions)

1.1 The More Electric Aircraft concept

In a large aircraft the main energy source is the jet fuel, the aircraft turbines transform this energy in the thrust needed for the aircraft to take-off, flying or landing. However, turbines not just supplied the thrust but also they generate the power required for the rest of loads in the aircraft. The whole non-propulsive power generated by turbines are called secondary power distribution. In a conventional civil aircraft the secondary power distribution is

Figure 1.2: Comparison of the secondary power distribution in a conventional aircraft and a More-Electric-Aircraft [4]

typically composed by four different power sources: the electrical, the pneumatic, the hydraulic and the mechanical. In the figure 1.2 (a) an aircraft turbine is shown illustrating the different power of sources attached to the turbine in a civil aircraft.

The power required for an aircraft can be classified in Electrical, Pneumatic, Hydraulic and Mechanical. Typically the sources and loads are distributed in this way.

- Mechanical system is used for fuel and oil pumping and it is driven from an engine gearbox
- Pneumatic system derives pressure from turbines off-take and provides heat and pressure for anti-ice protection engine start and cabin environmental control.
- Hydraulic systems primarily provide actuation of flight surfaces, landing gear and doors.
- Electric system provide power for avionic, lighting and galleys, etc.

The figure 1.3 gives an idea of how the four type of sources in the secondary power distribution coexist in a conventional commercial civil aircraft. This complex architecture with several types of sources and loads has begun to be questioned. It is expected that a significant improvement in the cost and in the performance in the system can be achieved by consolidating the whole

Figure 1.3: Conventional architecture with four different secondary power sources [3]

secondary power distribution into a single system. A reduction of complexity and improvement in maintenance are the immediate benefits.

In the hypothetical idea of unifying all type of loads and sources, the electrical system would be the chosen one because it has several advantages in terms of flexibility and application range, in fact the electrical system is the only one that has the potential to perform the tasks of all the other power system. Furthermore, the electrical system bring improvements in maintenance which considerable reduces aircraft cots. Another benefit in the medium-long term of electrical systems is their potential in new technologies because although at the level of architecture electrical systems are very mature, at the level of elements (semiconductors, capacitors and inductors) there are still significant technological improvements that can bring better performances in near future aircrafts. In the Table 1.1 are listed the main characteristics of electrical, hydraulic, mechanical and pneumatic system.

As it is shown in the figure 1.2, for a conventional civilian aircraft the nominal thrust power in a turbine is 40MW. The propulsive part corresponds to 95% of the full power, the rest of the power is distributed in electrical power (200kW), pneumatic power (1.2MW), Hydraulic power (240kW) and mechanical power (100kW). Unifying all the secondary power sources using just one single electrical system could bring benefits in efficiency as it is shown in the figure 1.2 (b).

The main goal of MEA is to reduce the operating costs, the fuel burn and the environmental impact of air travel. Replacing the pneumatic system removes the bleed air on the jet fuel turbine significantly improving the system efficiency.

Table 1.1: Comparison of aircraft power distribution systems

System	Complexity	Maintenance	Technological Maturity
Electrical	Complex	Simple ✓	System (Mature), New Tech. (Immature)✓
Hydraulic	Simple	Complex and Hazardous	Mature
Mechanical	Very Complex	Frequent and Slow	Very Mature
Pneumatic	Simple	Complex	Very Mature

The hydraulic system is mainly used for the primary and secondary flight control. This system is spread along the aircraft in the wings, horizontal and vertical stabilizer. This system is very robust but it is heavy, inflexible and it requires regular maintenance. Because primary control requires high levels of safety, this system requires redundancy. In some aircrafts, the hydraulic system possesses three full redundant hydraulic networks which means heavy networks with dangerous fluids all along the aircraft. The replacement of hydraulic actuators by electrical systems is highly desirable and one of the biggest part of MEA in order to reduce the overall weight and reduce the schedule and unscheduled maintenance costs

The MEA could also bring benefits in the maintenance costs since electrical systems offer far more options for reconfigurability as well as for advanced prognostics and diagnostics. These prognostics and diagnostics systems could help improve aircraft availability and reduce the need for unscheduled maintenance

The most important enabling technology for the More Electric Aircraft has been the power electronics. Without power conversion, none of the benefits of this technology could be possible [5].

1.2 Conventional electrical power distribution in an aircrafts

Typically, the power distribution in medium and large aircrafts contain 2 AC main bus and 2 DC main bus to supply power to the whole plane. The voltage levels employed in an aircraft are:

- 28 Vdc for low power loads and small aircrafts

Figure 1.4: Typical electrical system architecture

- 270 Vdc (± 135 V) in military aircraft
- 115 Vac at 400 Hz in civilian aircraft

These voltage level convention has been used since 1950, however due to the large increase in electrical power demanded imposed by the MEA concept, new aircrafts such as Boeing 787, have begun to increase the voltage in their generator to reduce the rating current. This allows to reduce significantly the wiring weight. As a drawback is worth mentioning that the use of higher voltages for power distribution requires new design restriction to avoid the possible effects of partial discharge, commonly called as corona effect.

The new voltage levels that are considered to be implemented in modern aircraft are the following:

- 540 VDC (± 270 VDC)
- 230 VAC at 400 Hz
- 230 VAC variable frequency (360Hz - 800Hz)

Today, most of the civil aircrafts have similar electric power architectures as it is shown in figure 1.4. The electrical architecture contains two main generators to supply AC power to two independent AC buses. The AC power is

three-phase 115V at 400Hz. The power level of the generator varies depending on the aircraft, for civilian aircraft, the generators power range is between 40kVA for old medium size aircraft (Boeing 717) to 250kVA in modern and large aircrafts (Boeing 787).

In addition to the main generators there is an auxiliary generator called APU (auxiliary power supply unit). This generator also provides three-phase 115VAC at 400Hz but it is independent of the main engines (thrusters) to diversify the sources of power. This generator is used on the ground but it can also be used in the air on certain installations under failure conditions. There is also a port to connect an external 115VAC power source meant to be used when the plane is parked.

There is also a generator specifically designed for emergency purposes when the conventional power system fails or is unavailable for some reason. This generator is called RAT (ram air turbine) and it uses the wind power outside of the aircraft. This is a smaller generator that can supply up to 15kW to power the crews essential flight instruments and a few other critical services for a limited period of time.

While the more demanded load is in AC frame, there are also DC load that are supplied by transformer rectifier unit (or TRU). Similarly, then in the AC side, in the DC side there are two power buses for normal operation and one extra exclusive to supply power to essential loads. The TRU are passive three-phase rectifier based on low frequency transformer and a diode bridge. The transformer has two purposes, to step down the voltage from 115VAC to 28VDC and divide the output in two windings in delta and star connection. This delta and star configuration helps to mitigate certain AC input harmonics when a diode bridge is used to rectify the voltage.

The main advantage of the TRU topology is the excellent reliability in combination with a reasonable good efficiency. This is mainly due to this topology does not possess high frequency switching mode transistors or diodes. The reliability of TRUs are very high because it is a simple system that just contain passive elements and each of these elements are very reliable such as low frequency magnetics and low frequency diodes.

While a TRU has unbeatable reliability it also has certain weaknesses compared to switched converters such as in power density. Selecting a switching frequency much higher than the main frequency the size and weight of magnetics can be drastically reduced. Additionally, in a modern switching rectifier, the harmonics of the main frequency (low frequency harmonics) are much attenuated than in a passive diode bridge topology which can significantly improve the quality of the demanded power and reduces the current stress in generators. In addition, switching mode rectifier generally possess the control capability of the output voltage unlike passive RFUs that present a voltage drop depending on the power demanded. This control capability

in modern rectifiers would allow improvement in the power architecture at system level because a DC load with a narrow input voltage range is easier to be optimized in terms of weight and volume than a load that has to operate for wide input voltages range.

In case of certain emergencies, there is a three-phase inverter that can power the essential ac bus using the 28V battery. This battery has three main proposes, to assist the dc power in damping transient, to provide power to specificity essential loads under emergency condition and to provide power in system startup modes when no other power source is available.

1.3 Requirements for Equipment Connected to the Aircraft Mains

As any AC load, three-phase rectifier have to comply with aircraft standards. Primarily the airborne standard RTCA DO-160C and the military standard MIL-STD-704E have to be fulfilled. As reported in [6] aircraft companies create their own standard with partly more stringent limitations than the ones listed in the mentioned airborne standard as an example Airbus has created the standard ADB-010 which is based on the DO-160C but is more permissible. In the following the main requirements will be discussed briefly.

1.3.1 Power factor requirements

As previously stated, there are different aircraft quality power standards for military and civil aircrafts but also depend on the country of manufacturing (US or Europe in general). In general, for civil aircrafts the US and the Europe standards for quality power used are RTCA DO-160C and ADB-0100 respectively. For military aircraft, the quality power standard used in general is MIL-STD-704E. In general terms, these standards contain equivalent requirements to comply with, however, as far as the power factor is concerned, this is not the case.

Among these standards, the MIL-STD 704 is the strictest in terms of the power factor requirements. This states that power factor of AC equipment greater than 500 VA shall be between 0.85 lagging and unity when operating at 50% or more of its rated load current in steady state condition. AC equipment shall not have leading power factor when operating at more than 100 VA.

This standard is especially strict in terms of leading power factor since it does not allow any capacitive reactive power when the apparent power is higher than 100VA in order to prevent self-excitation of synchronous generation sources.

On the other hand, in DO-160 is more permissive in the limits for leading and lagging power factor. In this case, for equipment with a power greater

Figure 1.5: Power factor limits for MIL-STD 704, DO-160 and ADB 0100

than 150VA, the steady-state full load power factor for each load shall be between 0.800 lagging and 0.968 leading. Power factor shall be higher than 0.2 either lagging and leading for equipment rated a very low power (smaller than 20VA) as it can be seen in the figure 1.5. In this figure is also shown the power factor limits for an Airbus standard ADB0100 used in aircrafts such as A350.

As the figure 1.5 illustrates, the MIL-STD 704 is the most strict standard for the power factor in the whole power range. In addition to this, the limits for MIL-STD 704 have to be fulfilled not only at nominal power but also from light load. Instead for standards RTCA DO-160C and ADB-0100 the power factor requirements must only be complied at nominal power. This difference has a great implication in the selection of the topology of the rectifier because unidirectional rectifiers lose the power factor correction capability at light load. Then, in order to comply with the MIL-STD 704 only bidirectional rectifiers, such as 6-switch boost rectifier, have to be consider in the topology selection analysis.

The reason why unidirectional rectifier loses the power factor capability is the following. As any switching converters, three-phase rectifiers need to have an EMI filter in the input of the converter. The EMI filter generates reactive power since it is implemented as an $L-C$ filter. As it will be shown in the chapter 3, unidirectional rectifiers have a limited reactive power compensation capability and this limitation increases at light load. Therefore, when the converter works at very light load the input impedance of the converter is defined by the input filter and consequently the power factor in this situation is very poor.

Figure 1.6: Simplified single phase low frequency rectifier model

In order to illustrate the power factor as a function of the output power, a simplify low frequency model of the rectifier is shown in figure 1.6 . In this figure, the resistor models the active power demanded by the rectifier and the inductance (L_{mod}) exemplifies the capability of reactive power compensation of the rectifier. As long as the output power is lower, the equivalent resistance, and consequently the impedance of the rectifier is higher. Since the rectifier impedance is in parallel with the capacitor of the filter, at light load the effect of the rectifier disappears then the power factor of the system only depends on the capacitance of the filter.

In the figure 1.7 the typical behavior of the power factor of a buck-type rectifier as a function of the output load is shown in conjunction with aircraft standards. As it can be notice, the power factor is capacitive at light load and te amount of reactive power depends on the filter capacitance, the grid voltage and frequency.

In order to design an unidirectional buck-type rectifier that complies with the MIL-STD 704 requires to add a low frequency inductance to compensate the leading power factor however this inductance would bring unacceptable increased in the weight and volume of the system therefore a unidirectional buck-type rectifier is not suitable in this case.

In military aircrafts when the MIL-STD 704 standard must be comply with, the common rectifier used the passive 12-pulse converter or RFU. Since this converter is passive does not require an EMI filter then this system does not have leading power factor at light power. Other option is to use the widely used 6-switch boost type bidirectional rectifier. Since this converter is bidirectional and the modulation of this converter does not depend on the input voltage or current sectors, this rectifier can compensate the reactive power of the filter in any condition of the load. However in aircraft applications, rectifiers are not allowed to regenerate power to the main for safety reasons.

In principle, by adjusting the control of a bidirectional converter, this can work as a unidirectional rectifier, however in the case of failure in the control, the converter has the capability to push power back into the main causing a problem in the main ac bus of the aircraft. For this reason unidirectional rectifier are preferred because the topology itself prevent the regeneration.

Figure 1.7: Typical power factor of a 10kW unidirectional buck-type rectifier including its EMI input filter.

On the other hand, in figure 1.7 it can be seen that using a buck-type rectifier with reactive power compensation, the system comply with the standards RTCA DO-160C and ADB-0100 for converter with a nominal power higher than 400VA and without compensation applications with a nominal power higher than 2kVA also comply with this aircraft standards.

Therefore a three-phase buck-type rectifier is suitable for civil aircraft application where the standards allow leading power factor at light load. However in the case of military applications a buck-type rectifier cannot be used due to the inevitably leading power factor at light load.

1.3.2 Aircraft Requirements for low and high frequency harmonics

The harmonics requirements can be separated in 2 groups, the low frequency harmonics up to the 40th harmonic of the main frequency and the high frequency (EMI requirements) for frequencies from 10kHz up to 10MHz. In the table 1.2 the harmonic requirements for the DO160 and ADB0100 standards have the same limits and their levels are shown. Additionally, for more clarity in the figure 1.8 these same limits are shown in a graph.

The harmonic limits are specified in such a way, that passive 12-pulse rectifiers, with their characteristic harmonics at $n = 12 \pm 1$, fulfill the specification. Boeing, however, defined more stringent harmonic limits for the 11th and 13th so that the limits cannot be fulfilled with passive rectifiers.

With regard to the limits of high frequency emissions also called EMI emissions, civil and military standard have certain differences. In the DO160

Figure 1.8: DO-160 harmonics limits

Table 1.2: Current harmonics limits for balanced three-phase electrical equipment according to DO160 and ADB0100

Harmonic Order	Limits
Odd Harmonics 3 rd , 5 th , 7 th	0.02 I_1
Odd Triple Harmonics (9 th , 15 th , 21 th , ... 39 th)	0.1 I_1/n
Odd Non-Triple Harmonic 11 th	0.1 I_1
Odd Non-Triple Harmonic 13 th	0.08 I_1
Odd Non-Triple Harmonics 17 th , 19 th	0.04 I_1
Odd Non-Triple Harmonics 23 th , 25 th	0.03 I_1
Odd Non-Triple Harmonics 29 th , 31 th , 35 th , 37 th	0.3 I_1/n
Even Harmonics 2 nd , 4 th	0.01 I_1/n
Even Harmonics 6 nd , 8 th , 10 th , ... 40 th	0.0025 I_1

Figure 1.9: EMI limit defined in DO-160

Figure 1.10: EMI limits for MIL-STD 461

and ADB0100 standard the input current is measured directly using a current probe. These emissions are measured in $\text{dB}\mu\text{A}$ and the frequency range of the standard begins at 150kHz and ends at 152MHz. The figure 1.9 shows the corresponding limits respect to the type of category.

On the other hand, in military aircraft the input current measurement is performed indirectly through the 50Ω resistor of the Line Impedance Stabi-

lization Network LISN. The LISN is a normalized and known impedance that is placed between the grid and the rectifier or device under test when all the EMI emission tests are performed.

Because the current is measured in a resistor, the emissions for this case are specified using $\text{dB}\mu\text{V}$ units. The figure 1.10 shows emission limits for the MIL-STD 461 and the limit for the industrial standard CISPR. The frequency range that is considered in the MIL-STD starts from 10kHz and ends at 30MHz.

The frequency range of the MIL-STD has an important effect in the volume and weight of the EMI filter for a military aircraft application in comparing with an EMI filter designed for a civil aircraft or an industrial application. In civil aircrafts since limits for the respective standard begins at a relative high frequency i.e. 150kHz, the switching frequency of the converter can be properly selected below 150kHz. In this way, the filter attenuation requirement is set by the first harmonic of the switching frequency that is beyond 150kHz. For example, if the switching frequency is 28kHz, only harmonic higher than 6th (168kHz) are considered and from the first and the 5th harmonics are not considered at all in the design and the EMI tests.

On the other hand, since in the MIL-STD the frequency range of the standard begins in a relative low frequency (10kHz) this technique of hiding switching harmonics cannot be used. More details about this are covered in the chapter 3.

1.3.3 DC output voltage regulation requirements

Similarly, that happens with the requirement of low frequency harmonics, the output voltage regulation in aircraft applications is meant to be applied in passive rectifiers, for this reason the steady state voltage range is relatively wide from 250Vdc to 280Vdc. During transients, the output voltage can vary between 200V to 330V during the first 10ms. The figure 1.11 shows the envelope of the allowed voltage in transients and in steady state.

1.11 shows the envelope of the allowed voltage in transients and in steady state.

This requirements are taken into account in the design of the control of the output voltage to set the minimum bandwidth of the control and the maximum output impedance.

1.4 Passive transformer rectifier unit (RFU)

Nowadays the most used rectifier used in aircraft application is the passive 12-pulse shown in the figure 1.12. In part, all the requirements of electrical quality in aeronautical applications for the rectifiers have been created thinking about this type of converters.

Figure 1.11: Envelope of normal voltage transient for 270 volts DC system in MIL-STD-704F

Figure 1.12: Tranformer Rectifier Unit (RFU) based on a 12-pulse transformer rectifier

This rectifier possesses a quasi-sinusoidal input current (in comparison with other passive solutions) and since it does not have neither control nor transistors the reliability of this converter is very high. However, the required mains frequency magnetic components result in a low power density and high system weight.

The waveform of the input currents of this converter are shown in the figure 1.13. Due to the characteristics of passive multi-pulse rectifier systems the input current harmonics are occurring at multiples of the pulse number, i.e. a 12-pulse system will show current harmonics at ordinal numbers $n=11, 13, 23, 25$ etc. One of the requirements for the rectifier system is a limitation of the amplitudes of the low frequency input current harmonics, where the 11th and 13th harmonics must be lower than 10

Figure 1.13: Voltage and current waveform of the twelve-pulse rectifier

Figure 1.14: Harmonic spectrum of the 12-pulse passive rectifier with selective filter

In the figure 1.14 the harmonics generated as a function of the first harmonic are shown in conjunction with the limits of the DO-160. As it can be seen, this converter fulfills the requirements in the harmonics.

Due to the RFU does not have control at all, the output voltage varies depending on the load. As long as the output current increases, the output voltage decreases, therefore the design of a RFU have to consider that the output voltage cannot exceed 280Vdc at very light load and it cannot be below 250Vdc at maximum power.

Figure 1.15: Output voltage behavior of twelve pulse transformer rectifier under different loads.

Figure 1.15(a) shows the value of the output voltage in steady state for different load resistances. This series of simulations have considered the turns ratio, inductances and output capacitance shown in the figure 1.12. For a very light load the output voltage has the highest value, as long as the load is getting higher the output voltage decreases as it expected. The output power increases as long as the resistance value decreases until a certain point. After this point, the output power decreases as long as the resistance is smaller because the voltage drop has a higher effect than the increase in output current.

In order to comply with the minimum steady state output voltage requirement of the rectifier, the maximum power for the converter is 6kW and 4kW for 400Hz and 800Hz respectively.

The figure 1.15(b) shows the power factor of the converter for different load resistance, as it can be seen the power factor is inductive in the whole power range. The value of the power factor decreases with the power, due to the minimum lagging power factor is 0.8, the limit for the maximum power is 12kW and 7kW for 400Hz and 800Hz respectively.

1.5 Contributions and Objectives of this Work

The objective of this work is to study, select and analyze a new power topology of an isolated three-phase rectifier which are suitable to fulfill the requirements for aircraft applications. Because of the high-power density and efficiency that a single stage configuration can reach, this work is focused in single stage rectifiers.

The topology selected in this work is the Swiss-Forward rectifier with resonant reset. This topology has been proposed in this work and this is not just suitable for aircraft application but also in industrial applications.

The main new contributions of this work are:

1. A modification of the non-isolated Swiss rectifier has been proposed in this work in order to transform this rectifier into an isolated single stage three-phase rectifier.
2. A New modulation for Swiss and Swiss-Forward rectifiers has been proposed in order to compensate the reactive power of the EMI input filter for light load operation.
3. Guidelines design of the proposed topology using analytical equations for calculation of voltage and current stress in semiconductors as well as a new design criteria for the LC input filter in buck-type rectifiers in application where the main frequency is 400Hz or higher.

The Objective of this work are:

1. Evaluation of isolated three-phase rectifier systems either using a 2 stages configuration or using a single stage topology.
2. A modification of the non-isolated Swiss rectifier has been proposed in this work in order to transform this rectifier into an isolated single stage three-phase rectifier.
3. A New modulation for Swiss and Swiss-Forward rectifiers has been proposed in order to compensate the reactive power of the EMI input filter for light load operation.
4. Design equations for the EMI differential mode filter in buck-type rectifiers in application where the grid frequency y 400Hz or higher.
5. Derivation of a reduce order (R.O) averaged model for the Swiss-Forward rectifier for control purposes .

6. Design and implementation of a digital control using a low cost DSP.
7. Successful design and construction of a 3.3kW demonstrator prototype of an isolated three-phase Swiss-Forward rectifier.
8. An experimental validation of the new isolated rectifier topology proposed in this work has been done. The demonstrator has been designed and tested to comply with aircraft applications for steady state conditions (THD, power factor and efficiency) and transient conditions (under, over voltage, settling time).

In the course of this dissertation the following conference and journal papers have been published:

1.5.1 Conference Papers

1. M. Silva, N. Hensgens, J. Oliver, P. Alou, Ó. García and J. A. Cobos, “New considerations in the input filter design of a three-phase buck-type PWM rectifier for aircraft applications,” 2011 *IEEE Energy Conversion Congress and Exposition*, Phoenix, AZ, 2011, pp. 4087-4092.

2. M. Silva, N. Hensgens, J. M. Molina, M. Vasić, J. Oliver, P. Alou, Ó. García and J. A. Cobos, “Interleaved multi-cell isolated three-phase PWM rectifier system for aircraft applications,” 2013 *Twenty-Eighth Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*, Long Beach, CA, USA, 2013, pp. 1035-1041.

3. M. Silva, N. Hensgens, J. Oliver, P. Alou, Ó. García and J. A. Cobos, “Isolated Swiss-Forward three-phase rectifier for aircraft applications,” 2014 *IEEE Applied Power Electronics Conference and Exposition APEC* 2014, Fort Worth, TX, 2014, pp. 951-958.

4. N. Hensgens, M. Silva, J. A. Oliver, J. A. Cobos, S. Skibin and A. Ecklebe, “Optimal design of AC EMI filters with damping networks and effect on the system power factor,” 2012 *IEEE Energy Conversion Congress and Exposition (ECCE)*, Raleigh, NC, 2012, pp. 637-644.

5. N. Hensgens, M. Silva, J. A. Oliver, P. Alou, Ó. García and J. A. Cobos, “Analysis and optimized design of a distributed multi-stage EMC filter for an interleaved three-phase PWM-rectifier system for aircraft applications,” 2012 *Twenty-Seventh Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*, Orlando, FL, 2012, pp. 465-470.

6. S. Zhao and J. M. Molina and M. Silva and J. A. Oliver and P. Alou and J. Torres and F. Arévalo and O. Garcia and J. A. Cobos, “Design of energy control method for three-phase buck-type rectifier with very demanding load steps,” 2014 *IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, Fort Worth, TX, 2014, pp. 2713-2718.

7. U. Borović and S. Zhao and M. Silva and Y. E. Bouvier and M. Vasić and J. A. Oliver and P. Alou and J. A. Cobos and F. Árevalo and J. C. García-Tembleque and J. Carmena and C. García and P. Pejović, “Comparison of three-phase active rectifier solutions for avionic applications: Impact of the avionic standard DO-160 F and failure modes,” 2016 *IEEE Energy Conversion Congress and Exposition (ECCE)*, Milwaukee, WI, 2016, pp. 1-8.

8. S. Zhao, M. Silva, J. A. Oliver, P. Alou, Ó. García and J. A. Cobos, “Analysis and design of an isolated single-stage three-phase full-bridge with current injection path PFC rectifier for aircraft application,” 2015 *IEEE Energy Conversion Congress and Exposition (ECCE)*, Montreal, QC, 2015, pp. 6777-6784.

9. P. Cortes, L. Fässler, D. Bortis, J. W. Kolar and M. Silva, “Detailed analysis and design of a three-phase phase-modular isolated matrix-type PFC rectifier,” 2014 *International Power Electronics Conference (IPEC-Hiroshima 2014 - ECCE ASIA)*, Hiroshima, 2014, pp. 3864-3871.

10. P. Cortes, J. Huber, M. Silva and J. W. Kolar, “New modulation and control scheme for phase-modular isolated matrix-type three-phase AC/DC converter,” *IECON 2013 - 39th Annual Conference of the IEEE Indus-*

trial Electronics Society, Vienna, 2013, pp. 4899-4906.

11. J. M. Molina, P. Alou, J. A. Oliver, M. Silva and J. A. Cobos, “Three-Phase Buck type Rectifier topology integrated with Current Fed Full-Bridge,” 2015 *IEEE Applied Power Electronics Conference and Exposition* (APEC), Charlotte, NC, 2015, pp. 84-91.

12. J. M. Molina and S. Zhao and M. Silva and J. A. Oliver and P. Alou and J. Torres and F. Arévalo and Ó. García and J. A. Cobos, “Power distribution in a 13 kW three-phase rectifier system: Impact on weight, volume and efficiency,” 2014 *IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, Fort Worth, TX, 2014, pp. 906-911.

1.5.2 Journal Papers

1. M. Silva, N. Hensgens, J. A. Oliver, P. Alou, Ó. García and J. A. Cobos, “Isolated Swiss-Forward Three-Phase Rectifier With Resonant Reset,” in *IEEE Transactions on Power Electronics*, vol. 31, no. 7, pp. 4795-4808, July 2016.

2. S. Zhao, J. M. Molina, M. Silva, J. A. Oliver, P. Alou, J. Torres, F. Arévalo, Ó. García and J. A. Cobos, “Design of Energy Control Method for Three-Phase Buck-Type Rectifier With Very Demanding Load Steps to Achieve Smooth Input Currents,” in *IEEE Transactions on Power Electronics*, vol. 31, no. 4, pp. 3217-3226, April 2016.

1.6 Thesis outline

This thesis is divided in 5 chapters starting from the explanation of the problem that is address in this thesis, the solution proposed and finally the results obtained.

In the **chapter 2** a study and classification of the different alternatives for isolated three-phase rectifier are shown. In this rectifier survey both single and 2 stage topologies are included as well as passive and active solutions.

In the **chapter 3** an in-depth analysis of the proposed topology for this work is performed. This analysis includes the principle of operations of the converter, two different modulations of the rectifier, equations for the stress in semiconductors, design equations for the differential mode filter, averaged model and control loops of the converter.

All experimental results are shown in the **chapter 4**. This chapter begins with the description of the experimental setup used throughout the development of this work. In the experimental results figures for the system operating in steady state as well as during transients using feedback control loops are shown.

Finally, in the **chapter 5** the conclusions of this work are presented. In addition, guidelines for future work are discussed.

Since the develop of this thesis has been intensive in hardware design for the demonstrator, in the **appendix** the electric schematic of the main PCBs are included.

Classification of Isolated Three-phase Rectifier Topologies

This work is focused in isolated three-phase rectifier for aircraft applications. In the state-of-the-art of power electronic there are many possible interfaces to convert electric power from a three-phase source to a DC load. In general, this power conversion is faced by a 2 stage configuration [7, 8]. In a 2-stage topology, the active front end or rectifier, is in charge mainly of the power factor correction i.e. low input current distortion and currents in phase with their respective voltages. On the other hand, the secondary stage has two main tasks, one is to set the appropriate output voltage and provide the galvanic isolation. Systems based on 2 stages are widely used because of the benefit of dividing the system in to parts brings some advantages in terms of optimization.

However, in the literature of power electronics it is possible to find alternative topologies to obtain a three-phase isolated system. One option is to replace the active rectifier by a passive diode bridge, and a power filter (preferably active filter) to minimize the input current distortion. The active power filter could bring a potential improvement in the power density since this filter has to process a fraction of the load power leading to a smaller size.

The third alternative is to use a single stage topology. Unlike in single phase grid, in three-phase grids the active power is constant therefore in a single-stage topology it is possible to provide a constant output power without the energy storage requirement. A single stage configuration could bring benefits in terms of power density since it has a lower number of reactive components than a 2-stage especially the DC-link Capacitor.

Figure 2.1: Classification of three-phase rectifier system with a high frequency transformer isolation

This chapter explores and compares different possible topologies of isolated three-phase rectifiers without neutral point connection, with unidirectional power direction. The figure 2.1 shows a classification of isolated three-phase rectifier. In this classification includes the most important topologies in the literature.

2.1 Passive System with isolation

The most used passive three-phase rectifier is the unidirectional 6-diode rectifier with an output capacitance to smooth the output voltage and inductor either in the ac or dc side to filter the current. This is the simplest rectifier topology and consciously it is very robust because it does not need control, current or voltage sensors, auxiliary power supplies, or EMI filters. However, the diode bridge has some drawbacks especially the non-sinusoidal input current waveforms and the unregulated output voltage.

Although it is true that increasing the capacitance and/or inductance values of the rectifier filter, the power factor is better, however since the input current waveform are not sinusoidal the minimum THD in the input current

Figure 2.2: Passive Three-phase rectifier. Transformer rectifier unit (TRU)

is limited. The main problem with the current waveform of this diode is due to the fact that the diode bridge prevents the existence of current in the three phases at the same time (if there is no inductance in the ac side of the diode bridge). In the diode bridge only the 2 phases with maximum and minimum instantaneous voltages are active and the third phase is disconnected.

In order to tackle this drawback of the diode bridge there are 2 widely used techniques in passive rectifier, the current injection method [9, 10, 11] and the multi-pulse rectifier (MPR) technique [12, 13, 14]. Among these 2 methods the multi pulse has been widely used in aircraft applications [15, 16, 17]. The figure 2.2 shows the 12-pulse rectifier using a transformer interphase. As it was explain in the previous chapter, currently this converter is the preferred in aircraft application because of its simplicity and the robustness since this converter does not require sensing, transistors, gate signal and control. In addition this converter complies with aircraft standards since the limits for this standards in many cases have been defined using the 12 pulse rectifier figures (harmonics, PF, etc) as a reference.

Although the 12-pulse rectifier is the most used rectifier in aircraft application this topology possesses the same drawbacks that any passive topology. First the inter-phase transformer is a low frequency element i.e. this is a bulky and heavy transformer and secondly the lack of control requires a regulated AC grid in order to provide a regulated dc bus. Using an active rectifier a much wider grid voltage could be allowed in future aircraft bringing some benefits in the design of the electric generators.

In this work, passive solutions have been discarded mainly because of the low power density limited by the transformer. In addition, this work attempts to include a controlled topology because of benefits that this feature incorporates to the global electric architecture in future aircraft.

2.2 Two-stage topologies

The 2-stage topologies are the classical approach for a PFC rectifier that requires galvanic isolation. In a first stage or front end is used typically an active rectifier that is designed to comply with all the input steady state requirements of the application such as power factor and input current distortion. Although for this purpose generally an active rectifier is used, a power filter (active or passive) in conjunction with a diode bridge can be also used. The benefits of using an active power filter is in the power density of the system since the power that have to be handle by the filter is a fraction of the rated power of the system.

The second stage is an isolated dc-dc converter. This converter provides the isolation as well as the proper output voltage regulation. This converter has to be designed in order to comply with the dynamic requirements of the application. In a 2-stage converter, the input and output dynamics can be adjusted with the bandwidth of the corresponding converter since in this topology there is always an energy storage component in the middle of the 2 converters (DC link capacitor or inductor). This energy storage component decouples the dynamic of the AC with the dynamic of the DC side facilitating the control design of the system. Other important advantage of the 2-stage topology takes a place in the design optimization, since the two submodules have independent requirements and specifications.

2.2.1 Modular-phase rectifier

A phase modular converter can also be considered as a 2-stage topology[18, 19]. A phase modular converter is composed by three isolated single-phase rectifiers connected either in star or delta connection in the input. Additionally, in the case of isolated phase modular rectifier the output of the modules can be connected in series (for higher output voltages) or in parallel (for higher currents).

Modular rectifier brings a great flexibility to the system, a three phase modular phase rectifier could be transformed in a 2-phase or single phase rectifier by reconnecting the input ports and modifying control scheme of the system. Another feature in modular systems is the scalability in power, especially when the system has input-output isolation because the modules can be stacked in parallel and/or in series to increase the current rate or the voltage rate of the rectifier. All of these benefits are attractive in some industrial application such as programmable power sources commonly used in electric and electronic power laboratory.

However, in applications where the modularity is not required the additional effort in the measurements, AC power filtering and EMI filter limit the global performance of a three modular phase rectifiers in compare with a

Figure 2.3: Two-stage modular phase converter. a) Three Y-connected single phase rectifiers and Three Full-bridge converters connected in parallel at the output. a) Three Δ -connected single phase rectifiers and Three Full-bridge converters connected in parallel at the output

single three phase system.

The main drawback of a phase modular system is in the fact that each rectifier subsystem is processing pulsing power since they are single phase rectifier. The pulsating power implies oversizing certain element of the systems, especially the transformer. The input voltage of the isolated converter has a rectified sinusoidal waveform and this is the voltage applied to the transformer. The duty cycle of transformer is also proportional to the absolute value of the input voltage therefore the volt times second of the transformer has a $\sin^2(\omega t)$ waveform. This means that the magnetic excursion of the transformer varies during one period making its design more complex and, more importantly in this kind of systems, the maximum power of each module is one half of the total rated power of the system instead of a third. Therefore, a phase modular system is over-sized in power leading to a potential decrease in power density.

Since in this work the rectifier power density is the most important feature

and the flexibility that a phase modular system brings is not critical, these kind of systems have been discarded for the application of this work.

2.2.2 Three-phase rectifier plus Isolated DC-DC converter

An active three-phase rectifier in conjunction with an isolated dc-dc converter is by far the most used configuration for an isolated three-phase rectifier system in the literature and in the non-aircraft industry. In the literature of active three-phase rectifier many different topologies have been studied [8]. In this work only the most popular topologies are analyzed. Since the scope of this work is in three-phase rectifiers, this section is only focused in the family of rectifiers.

The two main groups of active three-phase rectifier topologies are the DCM and PWM (or CCM). Dividing the active rectifiers between DCM and PWM is part of a global classification of three-phase PFC rectifier that has been done in [8].

DCM Rectifiers

The DCM or discontinuous conduction mode rectifiers take advantage of the natural resistive behavior of a boost converter when this works in DCM. Using a constant duty cycle in a boost converter, the average inductor current is proportional to the input voltage. In DCM three-phase rectifiers, using only one transistor the three input currents are proportional (ideally) to their respective voltage. However, in this converter the average current is perfect proportional to the corresponding voltage and this rectifier possesses a low harmonic distortion. This distortion can be attenuated by modulating the duty cycle of the transistor, this distortion compensation works better in high gain application (high output voltage), however the distortion cannot be completely eliminated [20]. The figures 2.4 and 2.5 show the power circuit of the single switch and double-switch DCM rectifier respectively.

These systems are characterized by their simplicity since they do not require measurement of the input current or voltage and there is only one (or 2) transistor to command facilitating the gate driving circuit and the control. This type of rectifier is appropriate in relative low power application because since the converter works in DCM, the RMS current of the input inductors, as well as in the diodes and transistors, are considerable higher than in other converter such as PWM (CCM) rectifiers.

PWM Rectifiers

PWM rectifiers are the most used kind of active front end in application for a few kilowatts to some hundreds of kilowatts. These rectifiers can be classified

Figure 2.4: DCM three-phase single switch boost rectifier [21] and [20].

Figure 2.5: DCM Two-Switch Three-Phase [22]

topologically in boost-type and buck-type. Similarly, as with dc-dc converters, the available output voltage for a buck-type rectifier is lower than the input voltage ($3\sqrt{2}U_{N,RMS}/2$) and for a boost-type rectifier the output voltage has to be higher than the input voltage ($\sqrt{6}U_{N,RMS}$).

The figure 2.6 illustrates the available output voltage area for the boost (blue area) and buck type (red area) rectifier as a function of the RMS line to neutral voltage of the grid. The operation limits of these two types of converters do not overlap between them, which means there are certain operation points (output/input ratio) that cannot be satisfied with either type of rectifier. Particularly in aircraft applications a PWM rectifier for a non-isolated solution cannot be used when the grid voltage is 115Vac and the dc voltage required is 270V, similarly happens for the newer voltage levels 230Vac and 540Vdc.

Among the PWM boost-type rectifiers most used both in industry and power electronic literature are the bidirectional six-switch and the Vienna rectifier. The figure 2.7 shows the schematic circuit of the boost-type six-switch and the figure 2.8 shows the Vienna rectifier with three possible semiconductor realizations. The three VIENNA rectifier implementations keep the same principle of functionality with some differences in the modulation and in the conduction losses of the system.

The six-switch rectifier is the most used active rectifier because it has

Figure 2.6: Output voltage range of the three-phase Boost and Buck type rectifier [8]

Figure 2.7: PWM Six-switch boost Rectifier [23]

several advantages. The modulation of this converter is straightforward and it does not need information about the input voltage sectors as it happens with unidirectional rectifier. Since this is a bidirectional rectifier it can be used as an three-phase inverter. While bidirectionality is an advantage in many applications, in aircraft applications it is a disadvantage because in this application the rectifiers are not allowed to regenerate power back to the ac side. A bidirectional converter can work as it were a unidirectional adjusting the control of the system, however under a failure on the system the converter could potentially regenerate power and propagate the failure in the entire electrical system of the airplane. For this reason, the unidirectionality is a must in aircraft applications.

The Vienna rectifier is a unidirectional rectifier and as, it is explained in detail in [25], it is a great alternative for non-isolated rectifier in future aircraft applications. This converter has been widely studied and it presents some

Figure 2.8: PWM VIENNA Rectifier [24]

benefits in terms of power density. Unlike the six-switch rectifier, the Vienna rectifier is a three level converter i.e. the switching side of the inductor has three possible voltage levels, this side can be connected to positive, medium and negative voltage of the output. This feature considerably reduces the input current ripple leading to smaller input inductances requirements. The control and modulation of the Vienna rectifier is more complex because it has to include the control of the voltage balance of the output capacitors.

On the other hand, in the PWM buck type three-phase rectifier family there are also two main topologies, the six-switch and the Swiss rectifier. Both power circuits are illustrated in the figures 2.9 and 2.10 respectively. These two topologies are three-level current fed unidirectional rectifiers therefore both require equivalent filtering effort. Additionally, these two converters have the same available space vectors which means that both have the same reactive power compensation capability.

Therefore, these two systems have quite similar performance, the main difference between these rectifiers is in the modulation. In the six-switch rectifier all its transistors switch at high frequency while in the Swiss rectifier the three bidirectional are low frequency devices and 2 high frequency transistors. For purposes of modulation the Swiss rectifier can be considered as if there were two dc-dc buck converters while this simplification cannot be done that easily in the six-switch converter.

Thanks to this feature, the non-isolated Swiss Rectifier can be transformed into an isolated one by replacing the two buck converters by a buck-type isolated converter such as, forward, full-bridge or a push-pull converter. Making

Figure 2.9: PWM Six switch buck-type rectifier with freewheeling diode [26]

Figure 2.10: Swiss rectifier [27, 28]

this circuit modification a single stage isolated three-phase rectifier can be obtained.

During the development of this work, a two-stage configuration has been deeply analyzed and developed using a six-switch buck type rectifier in a multi-cell configuration, the study and results are summarizing in [29]. Basically, this paper shows an isolated three-phase rectifier system with interleaving which reduces the filtering effort. Additionally, the multi-cell configuration can improve the availability of the system since when a semiconductor fails, the corresponding cell can be disconnected allowing the system to continue working at lower power.

However, a two-stage topology has potential drawbacks in terms of efficiency and reliability. In a two-stage topology the whole load power has to be processed twice therefore the global efficiency is limited. In addition, in a two-stage topology the number of element is relatively high since the system is composed by two full converters, the reliability is lower as long as the number of components is higher, especially in system with a large number of high frequency transistors.

Figure 2.11: Voltage source active power filter plus isolated DC-DC converter

2.3 Active Power Filter

In power electronic, there are passive rectifier base on diode bridge and active rectifiers. The three-phase diode bridge has several benefits such as low cost, it is very reliable and it is highly efficient, however this rectifier injects significant current harmonics into the grid. This prevents this type of converters to meet industrial or aircraft quality standards. On the other hand, active rectifiers generate much lower harmonics (in high frequencies) and they have the capability to control both the output voltage and the input current. However, active rectifiers are more expensive, less reliable and they possess lower efficiency than a passive solution.

Despite the three-phase diode bridge does not meet quality requirements, because of the low frequency harmonics because of the non-sinusoidal input current waveforms, the THD of the current can be improved adding a L - C filter. This filter deteriorates the power density because low frequency harmonics have to be attenuated. Another option is to use an active power filter which it is essentially an active bidirectional rectifier [16, 30].

The active power filter is in charge only of the current shaping to transform the distorted input current of a classical three-phase diode bridge into a sinusoidal (or quasi sinusoidal) current. The idea of this solution is to combine the benefits of the passive with the benefits of the active rectifiers. The key of this approach is based on the fact that the power processes by the active power filter is just a fraction of the load power (around the 30% [8]). In this way, since the diode bridge has a high-power density and the active power filter has also a relatively high-power density since this have to handle just a small part of the power, the power density of the full system preserves the high-power density as well.

The figure 2.11 illustrates an active power filter with a diode bridge and an isolated dc-dc converter as a secondary stage. The active power filter can be implemented by any bidirectional rectifier, in this figure a boost-type six-switch rectifier has been used because in the literature this topology is the most used for this purpose.

The principle of operation of the active power filter consists in the following. The output capacitor of the active power filter is not connected to any load, therefore during one part of the grid period, this converter has to absorb and in the other part it has to return the same energy back to the grid. The control of this system is analogous to that used in active rectifiers, the output voltage is controlled in an outer loop of a cascade control architecture, and the input current of the system in the inner loop.

The active power filter together with a diode bridge is a good alternative to an active rectifier specially when an isolated dc-dc converter is placed as a secondary stage because the drawback of the non-controlled output voltage is not relevant in this case. However the active power filter has a drawback in the way how the power is processed, because the output of the filter is not delivering power to the load, this power have to be processed again to put it back to the grid i.e. there is circulating power that is not being used. Thus, the way that this system processes the power is not extremely efficiency.

Following the main idea of taking advantage of benefits of the passive and active rectifier, hybrid rectifiers were introduced in the power electronic literature. Hybrid rectifier are characterized by the parallel connection of a passive and an active rectifier [31, 32, 33]. The passive low frequency converter processes most of the load power while the active rectifier is used to adjust the power factor of the system. As a consequence of this combination a system with a quasi-sinusoidal input current and high efficient system is obtained.

In the figure 2.12 an active power filter topology is compared with a hybrid rectifier topology. The hybrid rectifier is composed by two boost converters, the diode rectifier with a dc-dc boost converter works in continuous conduction mode (CCM) in order to minimize the RMS current in the boost inductors as well as in the transistor and in diodes. Since the dc-dc converter is in CCM it generates low frequency distortion in the input current, this distortion is compensated by the six-switch three-phase rectifier. Since the three-phase rectifier just adjust the input current waveforms, this converter processes a ratio of the load power. Additionally, the six-switch three-phase rectifier is directly connected to the output therefore there is no recycling power as it happens in a classical active filter power.

The hybrid rectifier shown in figure 2.12 is composed by two active rectifiers then, in the case of one of them fails, this one could be disconnected from the system and the other rectifier can still operate and generate power to the output with sinusoidal input currents. However, to take advantage of

Figure 2.12: Comparison topology between an active power filter and a hybrid Rectifier

this kind of system redundancy, additional circuits and control logic have to be implemented of identify a failure in one converter and then disconnect the broken converter from the grid and the output.

An active filter or an hybrid rectifier may have perform better in terms of efficiency or power density than an active rectifier but this difference is jut marginal, therefore in a two stage configuration the improvement of replacing the active rectifier with an active power filter is very small.

A rectifier system using a rectifier with an active power filter or a hybrid rectifier in conjunction with an isolated dc-dc converter has been discarded in this work for the same reason than a typical two-stage configuration has been discarded. These systems have a relatively large number of semiconductors and filtering components, this negative affect the reliability of the system. In addition, since in this configuration there are two converters in cascade connected, all the power is processed by the two converters then the efficiency is limited. The problem with the large number of components (semiconductors, inductors, capacitors) and the limited efficiency can be improved by simplifying the system into an isolated single stage three-phase rectifier.

Figure 2.13: Line time of single-stage isolated three-phase rectifier.

2.4 Single stage topologies

The advantage of these topologies is that the whole output power is processed only once which can bring benefits in terms of the efficiency of the system. In addition, these systems lack of a large energy storage element which, in principle, can improve the power density but at the same time also brings disadvantages in terms of the ability to compensate for pulsed power. Highly unbalanced three-phase grids generate pulsating power (not constant) therefore single stage rectifier solutions are not suitable in this kind of applications. In aircraft application the ac grid is fairly balanced therefore single stage rectifier can be suitable alternatives.

The single stage isolated rectifier family is classified in three groups, the DCM, the matrix-type and the unidirectional. These three families of rectifiers have been developed over time as shown in the figure 2.13.

The first single stage isolated rectifier was published in 1985 [34, 35, 36] and this converter is a Matrix-type rectifier, seven years later a boost-type with ZVS version of this rectifier was published [38, 37]. These two matrix-type are the more important rectifiers, however in others applications such as ac-ac, these matrix converters among others are widely used.

Both the single-switch and the unidirectional families have more topologies associated, and the time separating between them is shorter. This means that there was a special interest in the power electronics community to research and develop new topologies of single stage rectifiers with isolation from the nineties.

Figure 2.14: Isolated three-phase boost type rectifier based on a Matrix converter [34, 35, 36]. (a) Bidirectional-switch configuration and (b) six transistor boost converter plus a full-bridge converter without DC-link capacitor

2.4.1 Matrix type Rectifier

Matrix converters are used in some AC-AC applications. An isolated rectifier needs to generate AC to feed the transformer that provides the galvanic isolation. Therefore an isolated rectifier can be created using a matrix converter in conjunction with a transformer and a rectifier.

Based on this idea, there are two types of matrix rectifiers used as rectifiers that can be highlighted.

Boost-type Matrix converter

The single-stage matrix-type rectifier are the oldest in this category. The boost-type version was published in [34, 35, 36]. The figure 2.14 shows two versions of this converter. In (a) the converter is formed by 6 bidirectional switches (2 anti-series transistors for each bidirectional switch) and in (b) it is shown the version of a synchronized three-phase rectifier with a full-bridge converter. Although at first sight it seems that they are different topologies, in reality these two converters are equivalent. Each of these rectifiers has its own modulation however the waveforms of the voltage in the transformer as well as the current in the input current are identical. One drawback of these topology is the lack of ZVS which has a negative impact in the efficiency and EMI emissions, however the buck-type version of this rectifier has ZVS.

Figure 2.15: Isolated buck-type three-phase rectifier based on a matrix converter[38, 37]

Buck-type Matrix converter

The single stage Isolated Six-bidirectional-switch Buck-type Rectifier topology is shown in the figure 2.15. This converter was published in 1995 in [37].

This power circuit consists in a matrix of six bidirectional switches, a high frequency transformer and a diode rectifier with an output filter. Basically this converter works similarly than the classic unidirectional three-phase buck-type rectifier, however the bidirectional switches allow to connect positive and negative voltage to the primary side of the transformer for a proper magnetization. In addition, since in each switching cycle this converter works as a dc-dc voltage fed full-bridge converter using gate signals with a proper dead time, the rectifier can work with zero voltage switching improving the efficiency of the system.

The two versions of the matrix converters are good options to be used as a rectifier in aeronautical applications, especially the buck type due to the benefits of the ZVS in efficiency and noise reduction. However, the relative high number of high frequency transistors that have matrix converters impairs the reliability of the system.

The main objective of studying one-stage topologies is to obtain better results in terms of efficiency and reliability than a two-stage system, but since in any case the number of semiconductors is high in a matrix converter, in this work they are discarded.

There is a family of three-phase rectifiers that have only one transistor that are called single-switch DCM rectifiers. Since the number of high frequency transistor is reduced in these topologies they are promising in applications that require high reliability.

2.4.2 Isolated Single-Switch Discontinuous Conduction Mode three-phase Rectifier type

This family of three-phase rectifier is based on a non-isolated three-phase rectifier called single-switch rectifier published in [39],[21]. The figure 2.16 shows the structure of a three-phase single-switch discontinuous-mode boost converter.

The principle of operation of this rectifier is based on the resistive behavior of the input current that DCM boost converters offer because in DCM the input current is proportional to the input voltage using a constant on time modulation. Thanks to this characteristic of the DC-DC boost converter, a three-phase rectifier using only one transistor can be developed.

In addition to the simplicity in terms of the number of semiconductors presented by this rectifier, the modulation and control of this topology is simple as well. A simple modulation using a constant duty cycle can be used however there are more complex modulation in the literature that can improve the quality of the input current of the rectifier. However, the low frequency distortion present in this converter cannot be completely eliminated since the smallest phase current in each case always reaches zero prior to the other two currents and, thus, exhibits a zero current interval at switching frequency [40], [41].

Figure 2.16: Single-Switch non-isolated three-phase rectifier [39],[21]

Figure 2.17: Isolated single-switch three-phase rectifier based on the Ćuk converter [42]

Figure 2.18: Isolated single-switch three-phase rectifier based on the flyback converter [43, 44]

Figure 2.19: Isolated single-switch three-phase rectifier based on the SEPIC converter [45]

Figure 2.20: Isolated single-switch three-phase rectifier based on a Full-Bridge converter [46, 47]

The three-phase single-switch rectifiers have several isolated versions where the transformer is introduced adapting the topology using some common isolated dc-dc converter. In the figures 2.17, 2.18, 2.19 and 2.20 the Isolated single-switch three-phase rectifier based on the Ćuk, flyback, SEPIC and Full-bridge converter are shown respectively.

All these isolated versions of the single-switch rectifier share the characteristic of the original, that is, the inevitable current distortion at low frequency and the relative high current stresses present in the semiconductors and the inductors of this converter due to the DCM operation. In addition, in this type of rectifier the phase of input current cannot be controlled since this just can operate in resistive behavior, thus the reactive power of the EMI filter cannot be compensated therefore at light load this rectifier system cannot work at unitary power factor.

Due to the limitation in reactive power compensation, the low frequency current distortion and the high peak and RMS current loading of the power semiconductors and the large EMI filter effort, this family of single stage rectifier has been discarded in this work since the relative high power rating of this application target make more convenient the use of a CCM rectifier.

2.4.3 Unidirectional

This family of rectifiers are based on unidirectional, non-insulated topologies widely used in the literature and in industry, such as VIENNA and the buck-type rectifier. In this section a brief summary of the topologies of this family most named in the literature is made.

Boost-type Unidirectional Rectifiers

In the boost-type version of the unidirectional rectifier with isolation the most known topology is called VIENNA II (see figure 2.21). It was published in [48, 49]. This topology has a simple structure since with only three transistors the three input currents are controlled, and it has another pair of transistors that allows to polarize in both directions the transformer for a correct magnetization.

This structure of a transistor with a diode bridge allows connecting the boost inductors to three different potentials. Since this is a three-level rectifier and the input current has a continuous sinusoidal waveform, the filter effort is lower than other topologies such as a buck-type or a DCM rectifier.

The VIENNA II shares the disadvantages of the classical boost-type rectifier such as the additional circuitry needed for the start-up of the converter. In addition, this rectifier also does not have a direct current limitation under a short circuit at the output. More details about advantages and disadvantages of the topology are shown in [48, 49]

This topology is a general one, it is an excellent option for aircraft applications. However, in this work this topology has been discarded because it has preferred to focus work with buck-type topologies that do not require a start-up circuit. In any case, it should be noted that the VIENNA is a very

Figure 2.21: Isolated boost-type three-phase VIENNA rectifier (VIENNA II) [48, 49]

interesting topology that can be evaluated and studied in detail for future aeronautical applications

Buck-type Unidirectional Rectifiers

Among the isolated buck-type three-phase rectifiers there are three topology that can be highlighted, the VIENNA III, the buck type with flyback demagnetization and the Swiss Forward Rectifier. There three topologies circuit are shown in the figures 2.22, 2.24 and 2.25 respectively.

As shown in [50] VIENNA III is somehow a topological simplification of the buck-type matrix rectifier shown in the figure 2.14. The main advantage of the VIENNA III against the matrix converter is the lower number of transistors and the unidirectionality power which prevents topologically injecting power to the ac side of the rectifier. This characteristic is important in aircraft applications.

The VIENNA III also presents some comparative disadvantages with its matrix counterpart such as higher orders for driving due to the greater number of diodes on the ac side. Furthermore, the smaller space vector of the VIENNA III rectifier has, prevents the magnetization and demagnetization of the transformer from being carried out with the same voltages, which makes modulation of the rectifier difficult since the different voltages have to be compensated with different on times.

In [52] a new buck type rectifier with insulation was published. This converter, although it seems to be different from VIENNA III, is actually very similar topology, because both share the same vector space. The modulations of these converts are very different but the waveforms of the input currents as well as the voltages applied to the transformer are identical. The main

Figure 2.22: Isolated buck-type three-phase VIENNA rectifier (VIENNA III) [50],[51]

Figure 2.23: Unidirectional Isolated buck-type Full-Bridge Rectifier [52]

Figure 2.24: Isolated unidirectional Forward/Flyback Three-phase Rectifier [53, 54]

advantage that the buck-type full-bridge rectifier has over the VIENNA III is the ZVS behavior.

Both the VIENNA III and the Unidirectional Isolated buck-type Full-Bridge Rectifier are excellent rectifier candidates for aircraft applications however in this work they have been discarded because the SWISS Forward rectifier, as shown in the following chapters, is a more reliable topology since it has just 2 high frequency transistors and the modulation and control are simpler than the VIENNA III and the buck-type Full-Bridge Rectifier.

Swiss Forward Rectifier with isolation

The Swiss Rectifier is a new PFC topology proposed by the authors in [27, 28]. This converter is shown in Fig 3.2(c) and belongs to the family of current source, buck-type active rectifiers. As it has been presented in these papers, this topology is a good alternative to the classical six-switch buck type rectifier when a step down PFC converter is required, basically because of lower transistor losses (Fig. 11 [28]). Additionally, it is an easy to control solution where the three input currents as well as the output voltage are controlled with only two high frequency transistors. Therefore, the Swiss Rectifier seems very promising in the research field of three-phase active power factor correction rectifiers.

The Swiss Rectifier is basically formed by a combination of two DC-DC buck converters and an active third harmonic current injection circuit composed by three bidirectional switches. Replacing the buck converters by isolated converters, it is possible to provide the isolation in this power stage as it has been stated in [56]. Hence, a secondary isolated stage is no longer required. Two interesting solutions for the isolated converter could be a Forward or a Full-Bridge converter. Figures 3.3(a) and 3.3(b) show the circuit topology for the Swiss-Forward with resonant reset and the Swiss Full-bridge respectively. In general, in DC-DC applications the Full-Bridge presents better performance in the range of several-kilowatt applications because it can reach ZVS and the voltage stress in the transistors is equal to the input voltage. On the other hand, the Forward converter has hard switching, and higher voltage stress in the transistor. Nevertheless, the Forward converter has just one power transistor what leads to a less complex system with lower conduction losses.

Recent progress in research and design of the semiconductor devices based on new materials, such as SiC, provided commercially available transistors and diodes with high voltage ratings (1200V and 1700V). Comparing to corresponding Si devices, SiC transistors and diodes have much better characteristics in terms of on-resistance and parasitic capacitances. This new technology has changed the design criteria in power electronics because topologies like

Figure 2.25: Swiss Forward With Resonant Resent [55]

the Forward converter with higher stress in the devices can employ these SiC switches and avoid performance degradation.

In this work, a Forward topology with SiC devices has been chosen over the Full-Bridge since it contains lower number of high frequency transistors and the modulation is easier to implement leading to a less complex and more reliable system

Swiss Forward Rectifier

This chapter presents a new isolated single-stage PWM rectifier system based on the recently presented Swiss rectifier topology providing the isolation by replacing a buck with a resonant forward converter. The principle of operation and a new modulation technique which compensates the reactive power generated by the input filter at light load and maximizes the power factor are discussed. Furthermore, the analytical equations for the stress in the semiconductor devices, useful for the system optimization are derived as well as a simplified dc equivalent average model for control proposes is presented.

3.1 Introduction

Modern three-phase power supplies with low THD, high power factor (PF) and isolation are normally composed of two cascade-connected converters. An active-front-end power factor correction rectifier (PFC) is used as a primary stage, whereas an isolated DC-DC converter is employed as a secondary one [8, 57]. In general, the PFC is implemented with a voltage source boost-type PWM rectifier which can be either the classical bidirectional six-switch or the unidirectional Vienna-type rectifier. The secondary stage is connected through a DC-Link capacitor which, in general, is relatively large because it acts as energy storage that allows to decouple the dynamics of two converter stages (cf. Fig. 3.1a).

This approach allows dividing the design problem and optimizing the two sub-systems separately, focusing the rectifier design on the AC steady state performance (PF and THD) and leaving the dynamic requirements to the DC-DC converter [29]. In addition, in applications where the input and output voltage ranges are large and the energy storage capability is required, two-

Figure 3.1: (a) Two-Stage rectifier system. (b) three-phase rectifier with active filter plus isolated DC-DC converter topology, (c) single-stage isolated AC-DC converter

stage topologies are very attractive due to the flexibility given by the voltage in the DC-link capacitor [58, 59].

In this approach, the whole power has to be processed twice, once by the PWM rectifier and once by the DC-DC converter, thus the whole system efficiency is determined by the efficiency of both converters. An alternative is replacing the PWM rectifier by a diode-bridge with an active power filter (cf. Fig. 3.1b). The active filter only needs to process a small part of the total power. Consequently, the global efficiency and power-density can be increased. However, the number of semiconductors still remains high.

Balanced three phase systems do not require energy storage for normal operation, since the active power is constant. In those applications where energy storage is not required for hold-up purposes or for buffering ac power, the power stage can be implemented in a single stage configuration.

In single-stage topologies, the system requirements and specification are defined for one single converter, which complicates the design and optimization of the system, especially in the applications where the input/output voltage range is wide. However, since the power is processed only once, it is expected to reach higher efficiency, this is especially true in applications with a narrow input/output voltage range. Furthermore, since a single-stage

Figure 3.2: (a) Three-phase rectifier system based on a combination of a diode bridge and a DC-DC converter. (b) Boost-type version and (c) Buck-Type or Swiss Rectifier

configuration is basically only one converter, it is expected to employ fewer semiconductors, inductors and capacitors increasing the power density and reliability of the system. Therefore, in applications where input/output voltage range is small, single-stage topologies (cf. Fig.3.1(c)) are attractive for obtaining high efficiency, high power density and high reliability.

In [37] a single stage isolated buck-type three-phase rectifier has been proposed. This converter combines a matrix converter which, by using bidirectional switches, can apply positive and negative voltage to the primary winding of a high frequency transformer. This converter has an equivalent behavior of two full-bridge converters supplied by two maximum line to line grid voltages. In this way, under certain conditions, this converter can work in zero voltage switching reducing the switching losses. Due to the presence of six bidirectional switches, the modulation complexity is relatively high and

Figure 3.3: Isolated buck-type rectifiers topologies based on a diode bridge and an isolated DC-DC converter. (a) Swiss-Forward Rectifier with resonant reset and (b) Swiss Full-Bridge Rectifier [55].

the number of transistors penalizes the reliability.

In [46] and [60] an isolated single-stage topology has been presented. This converter is relatively easy to control as it operates in discontinuous conduction mode (DCM) using only one transistor and providing the proportionality of the three input currents to the line voltages. Furthermore, since this converter is based on a full-bridge topology, it operates in ZVS minimizing the switching losses. This solution is suitable for high frequency and low power application. However, at the power levels considered in this work (3.3kW), DCM converters are not appropriate because they have high RMS currents in the semiconductor devices, inductors and the transformer. In addition to that, DCM topologies do not allow the compensation of the reactive power at light load and therefore are not suitable solution for 400Hz grid applications.

On the other hand, there is a family of unidirectional three-phase PFC rectifiers based on a diode bridge and a DC-DC converter (Fig. 3.2(a)). In the first approach presented in [61], the employed DC-DC converter was boost-type, as shown in fig 3.2(b). Subsequently, the buck-type version called Swiss Rectifier (SR) has been proposed in [27] and [28] and it is shown in Fig. 3.2(c). It should be noted that the grey diode in the boost configuration indicates

that this diode is not required because is in series with the other diode however it has been left there to easily understand that there are 2 boost converter connected in series.

Furthermore, based on this idea, newer PFC rectifier topologies can be employed using other DC-DC converters such as Ćuk, buck-boost or SEPIC [56]. Additionally, as [56] states, it is possible to use an isolated DC-DC converter and therefore obtain an isolated single-stage rectifier topology. However in [56] the galvanic isolation it just mentioned in the introduction as future application of the topology without going into the matter. In this work, the galvanic isolation for the Swiss rectifier is explored in detail both in practical and theoretical level.

In this work, a SR-derived single-stage isolated three-phase rectifier, named the Swiss-Forward Rectifier with resonant reset is presented. The chapter explains the principle of operation and modulation of the converter.

3.2 Principle of Operation of the Swiss Rectifier

The Swiss Rectifier is a new PFC topology presented in [27, 28]. This converter is shown in Fig 3.2(c) and belongs to the family of current source, buck-type active rectifiers. As it has been presented in these papers, this topology is a good alternative to the classical six-switch buck type rectifier when a step down PFC converter is required, basically because of lower transistor losses (Fig. 11 [28]). Additionally, it is an easy to control solution where the three input currents as well as the output voltage are controlled with only two high frequency transistors. Therefore, the Swiss Rectifier seems an attractive solution for three-phase active power factor correction rectifiers.

The Swiss Rectifier is basically formed by a combination of two DC-DC buck converters and an active third harmonic current injection circuit composed by three bidirectional switches. Replacing the buck converters by isolated converters, it is possible to provide the isolation in this power stage as it has been stated in [56]. Hence, a secondary isolated stage is no longer required. Two interesting solutions for the isolated converter could be a Forward or a Full-Bridge converter. Figures 3.3(a) and 3.3(b) show the circuit topology for the Swiss-Forward with resonant reset and the Swiss Full-bridge respectively. In general, in DC-DC applications the Full-Bridge presents better performance in the range of several-kilowatt applications because it can reach ZVS and the voltage stress in the transistors is equal to the input voltage. On the other hand, the Forward converter has hard switching, and higher voltage stress in the transistor. Nevertheless, the Forward converter has just one power transistor what leads to a less complex system with transistors in the conductive circuits leading to potential reduction conduction losses.

Figure 3.4: Principle of operation of the Swiss and Swiss-forward rectifier. (a) Line voltages (phase to neutral), (b) u_{pY} and u_{Yn} quasi-triangular waveform, (c) ideal current waveform in the positive and negative node of the diode bridge for a unitary power factor correction

Recent progress in research and design of the semiconductor devices based on new materials, such as SiC, provided commercially available transistors and diodes with high voltage ratings (1200V and 1700V). Comparing to corresponding Si devices, SiC transistors and diodes have much better characteristics in terms of on-resistance and parasitic capacitances. This new technology has changed the design criteria in power electronics because topologies like the Forward converter with higher stress in the devices can employ these SiC switches and avoid performance degradation.

In this work, a Forward topology with SiC devices has been chosen over the Full-Bridge since it contains lower number of high frequency transistors and the modulation is easier to implement leading to a less complex and more reliable system

The principles of operation of the Swiss and Swiss-Forward Rectifier basically are the same [27, 28]. The only difference between these topologies is the replacement of the buck-type DC-DC by the isolated forward-type DC-DC converter. Thus, two high frequency transformers and the diodes D_+ and D_- are added in Fig. 3.3(a).

The diode bridge sets the voltage u_{pn} to the highest instantaneous voltage difference among the three input phases, while the bidirectional switches are used to connect the remaining third phase. In this way, as can be seen in Fig.3.4, the voltages u_{pY} and u_{Yn} are always positive and present a quasi-triangular waveform from 0V to $1.5\hat{U}_N$. In the resistive behavior operation, the currents i_p and i_n will follow the diode bridge voltages u_{pN} and u_{Nn} (see Fig3.4(c)). Advantageously in buck and forward converters in continuous conduction mode, the average input current is proportional to the duty cycle.

Therefore, in open loop operation the duty cycles for T_+ and T_- must be proportional to u_{pN} and u_{Nn} respectively. In this way a sinusoidal current in phase with the voltage is demanded from the grid.

3.3 Modulation of the Swiss Forward

3.3.1 Three-phase system assumptions and vector operator $\vec{a}$

In order to simplify the mathematical analysis of the rectifier, it is assumed pure sinusoidal symmetrical and balanced input voltages. In this way, the input voltages can be written as follows:

$$\begin{aligned} u_a(\theta) &= \hat{U}_N \cos(\theta) \\ u_b(\theta) &= \hat{U}_N \cos(\theta - 2\pi/3) \\ u_c(\theta) &= \hat{U}_N \cos(\theta - 4\pi/3), \end{aligned} \quad (3.1)$$

where θ is the angle of the grid voltage. Similarly, assuming a balanced and purely sinusoidal three-phase current could be presented as:

$$\begin{aligned} i_a(\theta) &= \hat{I}_N \cos(\theta - \phi) \\ i_b(\theta) &= \hat{I}_N \cos(\theta - \phi - 2\pi/3) \\ i_c(\theta) &= \hat{I}_N \cos(\theta - \phi - 4\pi/3), \end{aligned} \quad (3.2)$$

where ϕ is the phase between the voltages and currents. The angle of the currents can also be named $\psi = \theta - \phi$.

Defining a vector operator $\vec{a}$, where $\vec{a}$ is equal to:

$$\vec{a} = e^{j2\pi/3}, \quad (3.3)$$

One of the important properties of this vector is:

$$1 + \vec{a} + \vec{a}^2 = 0; \quad (3.4)$$

The system of equations formed by the equations 3.1 and 3.2 can be rewritten using the space vector representation, as follows

$$\vec{u}_N = \frac{2}{3} (u_a + \vec{a} \cdot u_b + \vec{a}^2 \cdot u_c) \quad (3.5)$$

$$\vec{i}_N = \frac{2}{3} (i_a + \vec{a} \cdot i_b + \vec{a}^2 \cdot i_c). \quad (3.6)$$

This transformation is widely used to analyze three-phase systems because one bi-dimensional space vector can represent a system of three-phase voltages or currents.

The instantaneous voltages of the individual phases can be obtained through projection of the space vector $\vec{u}_N$ on the axes 1, $\vec{a}$ and $\vec{a}^2$.

$$\begin{aligned} u_a(\theta) &= \Re(1 \cdot \vec{u}_N(\theta)) \\ u_b(\theta) &= \Re(\vec{a}^2 \cdot \vec{u}_N(\theta)) \\ u_c(\theta) &= \Re(\vec{a} \cdot \vec{u}_N(\theta)) \end{aligned} \quad (3.7)$$

Similarly, the input current can be expressed using the projection in the real part of the vectors:

$$\begin{aligned} i_a(\theta) &= \Re(1 \cdot \vec{i}_N(\theta)) \\ i_b(\theta) &= \Re(\vec{a}^2 \cdot \vec{i}_N(\theta)) \\ i_c(\theta) &= \Re(\vec{a} \cdot \vec{i}_N(\theta)) \end{aligned} \quad (3.8)$$

The space vector system has its origin in the field of rotating machines [62], but it is also advantageous used for the analysis of power electronic circuits [63]. The entire system state, through representation of the input voltage or current space vector can be described as a parameter in a structured manner. For this reason, the space vector is used for fundamental studies of the system at the balanced grid.

3.3.2 Symmetry of three-phase systems

Unidirectional three-phase rectifiers such as the buck type six-switch or boost type Vienna rectifier are analyzed dividing the period of the grid frequency in sectors and intervals. In general, one full period is divided in 12 sectors of $\pi/6$ -width or it is divided in 6 intervals of $\pi/3$ -width, as it can be seen in Fig.3.5.

The division is made in such a way that during an interval there are no crossings between phases maintaining a specific order in the voltages values. For example in the interval 1 is fulfilled that

$$u_a > u_b > u_c \quad (3.9)$$

For the definition of the section the rule also includes the sign of the voltages, for the sector 1

$$u_a > 0 > u_b > u_c \quad (3.10)$$

and for sector 2 applies

$$u_a > u_b > 0 > u_c \quad (3.11)$$

Figure 3.5: Sectors and intervals representation over one period of the grid.

The purpose of making this division of the period is to facilitate the analysis of rectifier, the idea is to focus the analysis of the input current in one interval. Since all intervals are equivalent between them assuming that the phases are interchangeable, the analysis of the currents done in one interval is also valid for the rest of the intervals. For this reason, this system is called symmetric respect to intervals.

3.3.3 Equivalent circuits and space vector for the first Interval

The transistors in the Swiss-Forward rectifier can be classified in two groups: the low and high frequency transistors. The low frequency transistors are three bidirectional switches S_{ya} , S_{yb} and S_{yc} (see Fig 3.3(a)). These transistors are complementary ($S_{ya} + S_{yb} + S_{yc} = 1$), and they provide the connection of the phase that is not connected through the diode bridge, allowing the third harmonic injection. For instance, in the interval one and four, the switch S_{yb} remains on during the whole time, and S_{ya} and S_{yc} remain off. Table 3.1 shows logic values for the three bidirectional switches over the whole period of the grid. The switching frequency of these transistors is twice the line frequency because in the table 3.1 each bidirectional is turned on in 2 over 6 intervals of the main frequency.

On the other hand, T_+ and T_- are high frequency transistors which are not complementary. Hence, there are four possible combinations that lead to four different equivalent circuits (see Fig.3.6). In the interval 1 i.e. $u_a > u_b > u_c$, if T_+ and T_- are on, the input current is

Table 3.1: Truth table for bidirectional switches

θ	Sector	Interval	S_{ya}	S_{yb}	S_{yc}
0 to $\pi/6$	1	1	0	1	0
$\pi/6$ to $\pi/3$	2		0	1	0
$\pi/3$ to $\pi/2$	3	2	1	0	0
$\pi/2$ to $2\pi/3$	4		1	0	0
$2\pi/3$ to $5\pi/6$	5	3	0	0	1
$5\pi/6$ to π	6		0	0	1
π to $7\pi/6$	7	4	0	1	0
$7\pi/6$ to $4\pi/3$	8		0	1	0
$4\pi/3$ to $3\pi/2$	9	5	1	0	0
$3\pi/2$ to $5\pi/3$	10		1	0	0
$5\pi/3$ to $11\pi/6$	11	6	0	0	1
$11\pi/6$ to 2π	12		0	0	1

$$i_a = \frac{N_2}{N_1} \cdot I_{DC}, \quad i_b = 0, \quad i_c = -\frac{N_2}{N_1} \cdot I_{DC}, \quad (3.12)$$

where N_2/N_1 is the turns ratio of the transformer and I_{DC} is the DC inductor current which is assumed to be constant in a switching period (T_{sw}). Applying the transformation of the equation (3.6), the rectifier input current space vector for this switching state is

$$\vec{i}_{N(on,on)} = \frac{2}{3} \frac{N_1}{N_2} (I_{DC} + \vec{a} \cdot 0 - \vec{a}^2 I_{DC}) = \frac{2\sqrt{3}}{3} \frac{N_1}{N_2} I_{DC} e^{j\pi/6} \quad (3.13)$$

Similarly, for the other combination of T_+ and T_- the space input current are

$$\begin{aligned} \vec{i}_{N(on,off)} &= \frac{2}{3} \frac{N_1}{N_2} (I_{DC} - \vec{a} \cdot I_{DC} + \vec{a}^2 \cdot 0) = \frac{2\sqrt{3}}{3} \frac{N_1}{N_2} I_{DC} e^{-j\pi/6} \\ \vec{i}_{N(off,on)} &= \frac{2}{3} \frac{N_1}{N_2} (0 + \vec{a} \cdot I_{DC} - \vec{a}^2 \cdot I_{DC}) = \frac{2\sqrt{3}}{3} \frac{N_1}{N_2} I_{DC} e^{j\pi/2} \\ \vec{i}_{N(off,off)} &= \frac{2}{3} \frac{N_1}{N_2} (0 + \vec{a} \cdot 0 - \vec{a}^2 \cdot 0) = 0 \end{aligned} \quad (3.14)$$

Figure 3.6: Equivalent circuits for the four possible combinations of the high frequency transistors. This model uses two ideal dc transformers without magnetizing inductance because this model is focused on the analysis of input currents of the rectifier. The demagnetization process of the transformers is shown in detail later.

Figure 3.7: Space vector representation of the input current for the Swiss Rectifier. (a) Voltage vector in sector 1 and 2 (b) input current vector in sector 1, 2, 3 and 12.

The three no-null vectors have the same magnitude but different angles. Performing a linear combination of these vector in conjunction with the null vector is possible generate any vector inside of the region formed by 2 extreme vectors. These 4 vectors are represented in the complex plane in Fig. 3.7(b). The figure 3.7 (a) shows the vector of the main voltage for the sector 1 and 2. Evaluating the system equation 3.14 for the sector 2 the available current space vector is obtained. The input current space vector is the same for the sector 1 and for the sector 2.

As it can be noticed, when $\vec{u}_N$ is in the sector 1 or 2 ($0 < \theta < \pi/3$), the current vector $\vec{i}_N$ represented as linear combination of the four vectors can be located in the sectors 1, 2, 3 or 12 i.e. $-\pi/6 < \psi < \pi/2$, which it is representing by a gray zone in the figure 3.7(b).

The fact that the angle range for the current is wider than the range of the input voltage means that the input current angle can follow the input voltage angle leading to the power factor correction action. Additionally, since the space vector angle range is 30° wider, this rectifier also possesses the capability to generate or compensate reactive power to the grid. In practice, the rectifier can work with up to 30° of leading or lagging power factor.

However, the input differential mode EMI filter can generate an additional phase shift between the current and voltage, depending on the power demanded by the load. In general, an unitary power factor is required in the input of the rectifier system (EMI filter and rectifier). Therefore, at light load a $\phi < 0$ is required in order to compensate the reactive power of the filter.

3.3.4 Selection of the vector combination for the modulation

The Swiss-Forward rectifier possesses 4 vectors for the input current. However, in order to construct a certain target vector $\vec{i}_N$ just three vectors are needed. To know which vectors are the most suitable to obtain the desired one depends on the angle and the magnitude of $\vec{i}_N$.

If $\vec{i}_N$ is in the sector 1 ($0 < \theta < \pi/6$) the appropriate vectors are the ones around to it i.e. vectors (on,off), (on,on) and (off,off). Similarly if $\vec{i}_N$ is in the sector 2 the used vector are (off,on), (on,on) and (off,off). In these two group of vector the null-vector is used to adjust the magnitude of the vector however it is possible to create the same effect without the null-vector.

When $\vec{i}_N$ is relative large that the modulation index is higher than $\sqrt{3}/3$ (modulation index is described below), $\vec{i}_N$ can be generated from the vector (on,off), (on,on) and (off,on) i.e. without the need of the null-vector. On the other hand, if $\vec{i}_N$ is small ($M < \sqrt{3}/3$), $\vec{i}_N$ is created using the vectors (on,off), (off,off) and (off,on).

Regarding the sequence of the vectors the most important aspect is to use sequences that avoid changing two transistors at the same time.

In summary, there are 4 combinations of three vector possible to generate any $\vec{i}_N$ in the space vector. However depending on its magnitude and angle, only two of these combinations are possible.

The figure 3.8 shows the key waveforms (low and high frequency pulses, input currents, voltage across the output diodes and the inductor current). This figure shows the 4 possible combinations of three vectors, in (a) and (b) are shown the waveforms with an small $\vec{i}_N$ and in (c) and (d) with large $\vec{i}_N$.

So far it has been shown that the input current vector can be generated in two possible ways, however the question now is how the duty cycles of the two transistors can be calculated to generate the required vector modulation. In [27] a formal mathematical demonstration is presented however it is also possible to explain this analysis in a more didactic manner.

As it can be seen in the 4 equivalent circuits for the sector 1 shown in the figure 3.6, when the transistor T_+ (upper one) is in series with the phase a, and the transistor T_- (lower one) is in series with the phase c, therefore the duty cycles of these two transistors control directly the current through the phase a and c respectively. In this way, assuming that the converter is working in continuous conduction mode, to set the rectifier with a resistor behavior the duty cycles for these two transistor must be proportional to the a and c phases for the sector 1 and the maximum and minimum voltage value in general for all sectors. This is the particular case for a resistor behavior but to generalized this concept when reactive power is required it covers in

Figure 3.8: Key waveforms signals for modulations with a small modulation index in (a) and (b) and with a large modulation index (c) and (d)

the following section.

3.3.5 Duty cycle variation of the high frequency transistors

As it was already explained, to generate the duty cycles of the two high frequency transistors T_+ and T_- is required to know the line voltage and the corresponding sector of the grid.

The voltage across the positive and negative side of the diode bridge $u_{p,N}$

and $u_{n,N}$ as a function of θ are

$$u_{p,N}(\theta) = \begin{cases} \hat{U}_N \cos(\theta) & : \text{for intervals 1 and 6} \\ \hat{U}_N \cos(\theta - 2\pi/3) & : \text{for intervals 2 and 3} \\ \hat{U}_N \cos(\theta - 4\pi/3) & : \text{for intervals 4 and 5} \end{cases}$$

$$u_{n,N}(\theta) = \begin{cases} \hat{U}_N \sin(\theta + \pi/6) & : \text{for intervals 1 and 2} \\ \hat{U}_N \sin(\theta - \pi/2) & : \text{for intervals 3 and 4} \\ \hat{U}_N \sin(\theta - 7\pi/6) & : \text{for intervals 5 and 6} \end{cases}$$

Since the average currents through the transistors T_+ and T_- are proportional to their corresponding duty cycles (for continuous conduction mode), $u_{p,N}$ and $u_{n,N}$ are used to generate the PWM signal for these transistors. Besides, a reactive power compensation can be implemented adding a certain phase. This phase is called in this work current-voltage phase of displacement ϕ . In this way, the duty cycles for T_+ and T_- are

$$\begin{aligned} d_{T_+}(\theta, \phi) &= \frac{M}{\hat{U}_N} u_{p,N}(\theta - \phi) \\ d_{T_-}(\theta, \phi) &= \frac{M}{\hat{U}_N} u_{n,N}(\theta - \phi). \end{aligned} \quad (3.15)$$

where M is the modulation index, while the term $\hat{U}_N$ is canceled out with the amplitude of $u_{p,N}$ and $u_{n,N}$. Therefore, the duty cycles are fully determined by the grid angle θ , the compensation current angle ϕ and the modulation index M .

d_{T_+} and d_{T_-} waveforms for a whole grid period are illustrated in Fig. 3.9(a) and (b) with a compensation angle $\phi = 0$ and $\phi = -\pi/6$ respectively. For $\phi = 0$ the duty cycles vary from $M/2$ to M . The duty cycles variations increases with the increase of $|\phi|$. Therefore, the maximum applicable $|\phi|$ is $\pi/6$ because in this case, the duty cycle variation is from zero to M .

As in a conventional buck converter, the output voltage of the rectifier is the average voltage in the free-wheeling diodes. In the interval one, when the T_+ is on, the voltage applied in the secondary of the transformer is $N_2/N_1 u_{ab}$ and this remain during in the duty cycle D_{F+} while in the case when the T_- is on, the voltage is $N_2/N_1 u_{bc}$ during in D_{F+} percentage of the switching period. Hence, the output voltage can be calculated as the integral of these diode voltages in one interval

$$\overline{u_{out}} = \frac{N_2/N_1}{\pi/3} \int_0^{\pi/3} d_{T_+} u_{ab} + d_{T_-} u_{bc}, d\theta \quad (3.16)$$

$\overline{u_{out}}$ is the average output voltage in a switching period. For the grid frequency analysis the output voltage ripple can be neglected then $u_{out} \approx \overline{u_{out}}$.

Figure 3.9: T_p and T_n duty cycles. (a) Resistive behavior $\phi = 0$, (b) Resistive-Inductive behavior $\phi = 30^\circ$

In the interval 1 the duty cycles d_{T+}, d_{T-} and input voltages u_{ab} and u_{bc} are

$$\begin{aligned}
 d_{T+} &= M \cos(\theta - \phi) \\
 d_{T-} &= M \sin(\theta + \pi/6 - \phi) \\
 u_{ab} &= \sqrt{3} \hat{U}_N \cos(\theta + \pi/6) \\
 u_{bc} &= \sqrt{3} \hat{U}_N \sin(\theta)
 \end{aligned} \tag{3.17}$$

Replacing this expressions in the equation 3.16, the output voltage is

$$u_{out} = \frac{3}{2} \cdot \frac{N_2}{N_1} \cdot M \cdot \hat{U}_N \cos(\phi) \tag{3.18}$$

Therefore, the resistive behavior, when $\phi = 0$, maximizes the output voltage under a fixed M , while in the case of $\phi \neq 0$, the output voltage is lower.

This is caused by the fact that the rectifier is using part its active state to handle the reactive power, thus limiting the output voltage.

3.3.6 Impact of the Carrier signals in the DC inductor and bidirectional current stress

So far, the duty cycles are determined by the angle of the input voltage (θ) and the modulation index: M . Regarding the switching frequency of the 2 transistors is convenient to use the same frequency in both transistors since the 2 forward converters share filters and having different switching frequencies would bring additional high frequency harmonics.

The Swiss-Forward converter is in fact the combination of 2 forward converters, these converters share the inductor current. Similarly as happens in interleaved multiphase converters is interesting to study what happens if the carrier signals of both buck-type converters in the Swiss-Forward rectifier are in phase or phase-shifted 180° . The study of intermediate cases where the phase between carriers is different from 0 or 180° are not relevant then they are not considered in this work.

In this work it is called modulation 1 when the carrier signals of the PWM modules are in phase and it is called modulation 2 when both carriers are phase-shifted 180° . As it was stated in [27] there is a trade-off between these two modulations, the modulation 1 possesses smaller current stress in the bidirectional switches and the modulation 2 has smaller current ripple in the inductor as it is shown in detail in the figure 3.10.

Fig. 3.10 illustrates three input currents, the voltage in the free-wheeling diodes and the DC inductor current in the interval 1 for modulation 1 in black lines and for modulation 2 in gray lines. In Fig. 3.10(d) the current through phase b is shown. In the interval 1, this is also the current in the bidirectional switch and it can be noticed that in one switching period for modulation 1 this current has just one direction. However, in the case of the modulation 2, this current always has both directions and in order to keep the same average value this modulation gives higher RMS currents in bidirectional switches.

Fig. 3.10(f) illustrates the freewheeling diode voltages. For the modulation 1, the conduction of the two freewheeling diodes is synchronized, giving a zero voltage applied to one side of the inductor. Therefore, the inductor demagnetization is done with the full output voltage. On the other hand, for the modulation 2, one of the free-wheeling diodes is always off. Therefore, it is preventing the inductor demagnetization with the full output voltage. This leads to a smaller current slope and therefore smaller current ripple as Fig. 3.10(g) shows.

For modulation 1 the maximum current ripple in the inductor happens at $\theta = \pi/6$, when the duty cycle is equal to M . Then the current ripple can be

Figure 3.10: Key signals in the interval 1 ($0 < \theta < \pi/6$) for modulation 1 and 2. (a) Pulses signals to the high frequency transistor for modulation 1. (b) Pulses signals to the high frequency transistor for modulation 2. (c),(d) and (e) Currents through the phase a, b and c respectively. (f) Freewheeling diodes and output voltages. (g) DC inductor and output currents.

calculated with the time of the freewheeling state according to

$$\Delta i_{\text{Mod 1}} = \frac{(1 - M\frac{\sqrt{3}}{2}) \cdot u_{out}}{f_{sw} L}, \quad (3.19)$$

The maximum current ripple for the modulation 2 happens at $\theta = 0$. Similarly, the maximum current ripple for the modulation 2 is

$$\Delta i_{\text{Mod 2}} = \frac{(1 - M) \cdot u_{out}}{f_{sw} L}. \quad (3.20)$$

Therefore, the current ripple ratio for modulation 1 over 2 is $(1 - M)/(1 - M\sqrt{3}/2)$ which means a reduction of the 55% of a maximum current ripple with using a $M = 90\%$. The reduction of current ripple leads to a reduction in the inductance requirement in the rectifier output filter decreasing the system weight and volume.

Another difference between these two modulations is related with the conducted EMI noise. The modulation 1 presents lower level of differential mode noise while the modulation 2 has lower common mode noise [27].

3.4 Resonant Reset Forward Topology

Similarly, that happens with the two Buck converters in the Swiss rectifier, the forward converters have a very particular current and voltage waveform. The input voltage varies from 0[V] to $3/2 \hat{U}_N$ with a kind of triangular waveform as it show the figure correction 3.4, and the current has a piecewise sinusoidal waveform as it is shown in 3.9. Therefore, the forward converter and most precisely, the demagnetization process have to work properly for a wide range of input voltage and current.

As it is usually done in switching rectifiers, the input currents and voltages are considered constants during a whole switching period since the switching frequency is considerable higher than the main frequency.

3.4.1 Transformer Demagnetization

It is a well-known fact that the forward converter needs an extra energy path to demagnetize the transformer when the transistor is off. The most common technique is the application of an additional third winding in the transformer (Fig. 3.11(a) and (b)), in order to properly reset the transformer through the input or the output of the converter [64, 65]. The disadvantage of this technique is a more complex transformer together with the single quadrant operation of the flux in the transformer. In the Swiss-Forward Rectifier the demagnetization through the input is not possible due to the diode-bridge

Figure 3.11: Forward topologies variation, depending on transformer demagnetization method: (a) classical forward converter, (b) forward-flyback converter, (c) active clamp PWM forward converter, (d) resonant-reset forward converter.

which only allows unidirectional current flow. In order to use the output for demagnetization i.e forward-flyback topology, the output voltage needs to be above a specific threshold. This condition is not fulfilled during the converter start-up phase, thus complicates the soft-start procedure of the converter.

The second most used technique for demagnetizing the transformer in the forward converter is the use of an active clamp (Fig. 10(c)) [66, 67, 68]. In this converter the steady state voltage in the clamping capacitor is naturally set according to the input voltage of the converter and the duty cycle of the transistor. However, in this application both the input voltage of the converter and the duty cycles have large variations, resulting in a large ripple in the clamping capacitor voltage. This high frequency ripple disables the proper demagnetization of the transformer in each switching cycle.

A third technique to demagnetize the transformer is the resonant reset [69, 70, 71], which is simple in the sense that it does not need an extra winding or any additional switch, but only one resonant capacitor. However, the voltage stress in the transistor is higher than in the previous configurations. On the other hand, the voltage in the capacitor starts rising from zero, which ensures that the varying triangular input voltage does not affect the functionality of the reset circuit. Therefore, this technique is an appropriate choice for the Swiss-Forward converter.

Fig. 3.12 shows the demagnetization process of the resonant reset network. During $t_0 < t < t_1$ the input voltage is applied to the transformer primary winding and the magnetizing current increases with a constant slope. In $t_0' < t < t_1$ the transistor is turned off, when the voltage in the capacitor reaches the

Figure 3.12: Equivalent circuit and main waveforms of the forward converter with resonant reset transformer demagnetization.

input voltage in $t = t_1$, this capacitor starts to resonate with the magnetizing inductance of the transformer, this occurs in $t_1 < t < t_3$. Therefore, the inductance current decreases while the capacitor voltage increases. In $t_2 < t < t_3$ when the magnetizing current reaches zero, the transistor voltage reaches maximum and starts to decrease while the magnetizing current is negative.

In $t_3 < t < t_4$, when the transistor voltage reaches the input voltage, the diode D_1 is turned on, the secondary voltage is clamped to zero and the magnetizing current remains constant. In this way, regardless of the input voltage and the duty cycle (below a threshold), the capacitor voltage remains zero until the next period when the whole process repeats. Despite the fact that the transistors have to withstand the peak voltage, the turn off happens when the voltage across the transistor is equal to the input voltage which in the worst case has the value of $3/2 \hat{U}_n$. In this way, the switching losses are not affected by the resonant circuit.

3.5 Stresses and Selection of Power Semiconductors

The transistors and diodes in this topology can be classified in two groups: the ones that are switching at low-frequency (twice the line-frequency) and those switching at high-frequency (switching frequency). The low-frequency

devices are the three-phase diode-bridge (D_{N+} and D_{N-}) and the bidirectional switches of the active third-harmonic injection circuit ($S_{y,123}$). The high-frequency devices are the Forward-transistors (T_P and T_N) and the secondary rectifier diodes (D_+ , D_- , D_{F+} and D_{F-}).

3.5.1 Semiconductors voltage stresses

The voltage stress on the rectifier diodes is equal to the maximal line-to-line input voltage. The highest voltage stress on the bidirectional switches occurs at the moment when one line-voltage reaches its maximum amplitude while the other two are equal to minus half of their amplitude, resulting in the following equations:

$$\hat{u}_{DN+/-} = \sqrt{3} \cdot \hat{U}_N, \quad (3.21)$$

$$\hat{u}_{Sy} = \frac{3}{2} \cdot \hat{U}_N. \quad (3.22)$$

The voltage stresses in the high-frequency diodes correspond to the voltage stresses in the low-frequency devices, multiplied by the turns-ratio of the transformer and are given by:

$$\hat{u}_{DF} = \frac{N_2}{N_1} \cdot \frac{3}{2} \cdot \hat{U}_N. \quad (3.23)$$

None of the previously discussed voltage stresses of the devices did not depend on the modulation index M . However, the higher the M is, the shorter the time available for the demagnetization of the transformer is. This has a negative impact on the voltage stress in the transistors. The peak voltage in the transistors is given by the input voltage plus the peak voltage of the resonant circuit. The worst case scenario for this peak voltage is when the input voltage u_{PY} or u_{YN} is maximum ($3/2 \cdot \hat{U}_N$) and at the same time, the duty cycle is also the maximum ($d_{T+} = M$ or $d_T = M$).

The voltage in the primary of the transformer is shown in Fig. 3.13. This voltage corresponds to the input voltage when the transistor is turned on, while in the off state this voltage is defined by the resonant circuit. To minimize the peak voltage in the transformer and also in the transistors, the half of the resonance period is made be equal to $(1 - M)T_{sw}$.

In steady state, the average voltage applied to the transformer needs to be zero, which results in the equality of the areas A_1 and A_2 at Fig. 3.13. Thus, the peak voltage of the resonance is

$$\hat{u}_p = \frac{3\pi}{4} \cdot \frac{M}{1 - M} \cdot \hat{U}_N. \quad (3.24)$$

Therefore, the transistor peak voltage is:

$$\hat{u}_T = \left(\frac{3\pi}{4} \cdot \frac{M}{1 - M} + \frac{3}{2} \right) \hat{U}_N \quad (3.25)$$

Figure 3.13: Transformer voltage in the forward with resonant reset.

Finally, the diodes $D_{+,-}$ withstand the voltage of the resonant reset circuit multiplied by the turn ratio of the transformer. Therefore, the peak voltages in these diodes are:

$$\hat{u}_{D_{+,-}} = \frac{N_2}{N_1} \cdot \frac{3\pi}{4} \cdot \frac{M}{1-M} \cdot \hat{U}_N. \quad (3.26)$$

3.5.2 Semiconductors current stresses

Analytical equations for the average and RMS current for all of the semiconductors are very useful because they provide the relationship for this current respect to different condition of the converter i.e, different output voltage and current. This mathematical expressions are useful to estimate conduction losses in different operation point without the necessity to run a simulation and most importantly, these equation are needed when an optimized system is required because any multi objective optimization algorithms process needs equation to estimate losses in the system.

The equations obtained in this section are based on the following approximations

- Zero current ripple in the DC inductor
- Purely sinusoidal input voltage
- Basic modulation without phase angle compensation.

Figure 3.14: High frequency transistor duty cycle and current waveform.

The figure 3.14 illustrates the waveform of the current through the high frequency transistor together with the waveform of the its duty cycle and carrier signal. The operation point of these signals correspond with the nominal power of the system.

The current through the two high frequency transistors I_T is a PWM signal where the maximum value is the dc inductor current times the turns ratio. The duty cycle is changing according the the input voltage, however, because of the symmetry of the system, the period of this current is $2\pi/3$, then the average current can be written as the following:

$$I_{T,avg} = \frac{3}{2\pi} \int_{-\pi/3}^{\pi/3} I_{DC} \frac{N_2}{N_1} M \cos(\theta) d\theta$$

$$I_{T,avg} = I_{DC} \cdot \frac{N_2}{N_1} \cdot \frac{3M\sqrt{3}}{2\pi}. \quad (3.27)$$

To obtain an expression for the RMS value of the transistor current a propriety of the RMS value is used. For definition the RMS is the square root of the arithmetic mean of the squares of the values [72, 73], i.e,

Figure 3.15: RMS value of a continuous waveform as a function of separated pieces

$$x_{RMS} = \sqrt{\frac{1}{n} (x_1^2 + x_2^2 + \dots + x_n^2)}, \quad (3.28)$$

where n is the number of times the switching period fits the period of the transistor current.

In addition the RMS value of each switching period is the maximum value times the square root of the duty cycle

$$I_{T,RMS,k} = I_{dc} \frac{N_2}{N_1} \sqrt{d_k}, \quad (3.29)$$

where $k = 1, 2, 3, \dots, n$.

Putting together the equation 3.28 and 3.29 the following equation is obtained

$$I_{T,RMS} = I_{dc} \frac{N_2}{N_1} \sqrt{\frac{d_1 + d_2 + \dots + d_n}{n}}, \quad (3.30)$$

the argument of the square root is in fact the average value of the duty cycles along the full low frequency period. Therefore the RMS value can be express as a function of the it average value.

$$I_{T,RMS} = I_{DC} \frac{N_2}{N_1} \sqrt{\frac{3M\sqrt{3}}{2\pi}} \quad (3.31)$$

On the other hand, the high frequency diodes in the secondary side of the transformer have either the same waveform in the current or the complementary current. The forward diode conducts at the same time than the transistor however since this diode is directly connected to the dc inductor without the transformer, the average and RMS values of the current of this diode are :

$$I_{D_{Fo},avg} = I_{DC} \cdot \frac{3M\sqrt{3}}{2\pi}. \quad (3.32)$$

$$I_{D_{Fo},RMS} = I_{DC} \sqrt{\frac{3M\sqrt{3}}{2\pi}} \quad (3.33)$$

The free-wheeling diode conducts when the transistor and the forward diode is turned off, e.i. these diodes are conduct complementary therefore the waveform current of the free-wheeling diode are

$$I_{D_{FW},avg} = I_{DC} \cdot \left(1 - \frac{3M\sqrt{3}}{2\pi}\right). \quad (3.34)$$

$$I_{D_{FW},RMS} = I_{DC} \sqrt{1 - \frac{3M\sqrt{3}}{2\pi}} \quad (3.35)$$

The current stress in the low frequency diode bridge can be obtained similarly. The three upper diodes pf the bridge are in series connected to the high frequency transistor, due to the phase symmetry each diode has one third of the average current of the transistor i.e.

$$I_{D,avg} = \frac{I_{T,avg}}{3}$$

$$I_{D,avg} = I_{DC} \cdot \frac{N_2}{N_1} \cdot M \cdot \frac{\sqrt{3}}{2\pi}. \quad (3.36)$$

The RMS current of these diodes can be calculated from the average current as well as in the others devices

$$I_{D,RMS} = I_{DC} \cdot \frac{N_2}{N_1} \cdot \sqrt{M \cdot \frac{\sqrt{3}}{2\pi}}. \quad (3.37)$$

Finally, for the stress in the bidirectional switches the analysis is equivalent. The bidirectional switches have zero current in average since they have current in both directions. However for looses estimation proposes is important to have the average of the absolute value since the direction of this current does not affects the calculation of the conduction losses when both anti-series transistors are tuned on.

The bidirectional switches conduct current when the corresponding current is in the neighborhood of zero. The bidirectional switch for the phase a conducts when $\pi/3 < \theta < 2\pi/3$ and $4\pi/3 < \theta < 5\pi/3$. Due to the symmetry the average current can be calculated just integrating in the domain $\pi/3 < \theta < \pi/2$

Figure 3.16: Current stress equations summary for the Swiss Forward Rectifier

and multiplying the result by 4

$$I_{S,avg} = I_{DC} \cdot \frac{N_2}{N_1} \cdot \frac{4}{2\pi} \int_{\pi/3}^{\pi/2} M \cos(\theta) d\theta$$

$$I_{S,avg} = I_{DC} \cdot \frac{N_2}{N_1} \cdot M \left(\frac{2 - \sqrt{3}}{\pi} \right). \quad (3.38)$$

And the RMS current of the bidirectional switches is

$$I_{S,RMS} = I_{DC} \cdot \frac{N_2}{N_1} \cdot \sqrt{M \left(\frac{2 - \sqrt{3}}{\pi} \right)}. \quad (3.39)$$

The figure 3.16 shows a summary of the all analytical expressions obtained in the section for the average and RMS current of all semiconductors in the Swiss Forward Rectifier.

In order to validate these equations a comparison between this equations and numeric simulations has been performed. For these tests the nominal converter operation has been set i.e. 115V_{rm} in the input and 270V_{dc} at 3300W. Table 3.2 shows all results obtained in this comparison. This table includes 2 columns for simulations, one is for an ideal simulation of the rectifier without including the EMI input filter either the output filter. In the second-to-last column of the table results for a more realistic simulation is shown,

Table 3.2: Stress current comparison between numeric simulation and analytical equation

Parameter	Analytical eqs[A]	Sim W/O filters [A]	Sim. with filters[A]	Percen. error %
$I_{D,avg}$	3.8137	3.8066	3.8505	0.9557
$I_{D,RMS}$	9.1515	9.1356	9.2500	1.0649
$I_{S,avg}$	1.1800	1.1968	1.1960	1.3378
$I_{S,RMS}$	5.0904	5.1132	5.1653	1.4501
$I_{T,avg}$	11.4413	11.4174	11.4951	0.4680
$I_{T,RMS}$	15.8509	15.8269	15.9855	0.8420
$I_{DFo,avg}$	6.3563	6.3437	6.3862	0.4682
$I_{DFo,RMS}$	8.8060	8.7929	8.8808	0.8423
$I_{DFw,avg}$	5.8437	5.8535	5.8257	0.3090
$I_{DFw,avg}$	8.4435	8.4461	8.4563	0.1514

in this simulation both the input and output filter is included as well as equivalent series resistances for inductors.

As it can be seen in table 3.2 there is a good agreement between the calculus obtained with the equations with the ideal simulation as well as if in the simulation with input and output filters. In the last column of the table the percentage error between the analytical equations and the more realistic simulation are shown. The error in all cases is smaller than 1.5% which means that the approximation with equations is accurate enough to be applied in design procedure and optimization algorithm. It is important to note that although the analytical equations assume ideal conditions in terms of current and voltage ripples, this approximation are also valid when in real design when ripples are relative small in compare with the average values.

3.6 Differential mode EMI filter

The input filter in a PWM rectifier system has three main purposes:

1. to ensure sinusoidally shaped input currents by filtering the switching-frequency harmonics;
2. to attenuate the electromagnetic interference with other electronic systems
3. to avoid susceptibility to electromagnetic emissions from surrounding systems and itself [74], [75] and [76]

In order to design an EMI filter for a power electronic system, the applicable EMI standards need to be considered.

3.6.1 Differences in the EMI standard for Industrial and Aircraft applications

Typically, in industrial applications, the conductive EMI standard to comply with is CISPR 22 class B [77]. The frequency range considered by this standard reaches from 150kHz to 30MHz. In [73] and [78],[79] systems with a switching frequencies of 28kHz and 18kHz respectively have been designed. These switching frequencies have been chosen because they are sufficiently higher in comparison with the grid (50 or 60Hz). In addition, the first, second, third, fourth, and fifth harmonic of the switching frequencies are out of the range of CISPR 22 class B; thus, the first harmonic to consider in the input filter design is the sixth harmonic at 168kHz (when the switching frequency is 28kHz).

In this work new considerations in the input filter design for a three-phase buck-type pulse-width modulation rectifier for aircraft applications. In this application the standard to comply with is MIL-STD-461E [80]. This standard is more restrictive than the CISPR 22, regulating a wider range of frequencies from 10kHz to 10Mhz. Due to the frequency range of the MIL-STD-461E and the fact that switching frequencies below 10kHz would not be an optimal design, the rectifier switching frequency must be inside of the range. Therefore, the input filter must be designed in order to attenuate the switching frequency.

The figure 3.17 illustrates the difference between the design of a EMI filter for an industrial and a aircraft application. In industrial applications, the main frequency is 50Hz or 60Hz and the EMI standard begins to be applied from 150kHz. Since 150kHz is quite far from 10kHz (3000 times) it is possible to set a switching frequency of the rectifier somewhere in the middle of these frequencies. In this way the harmonics of the switching frequency do not disturb the low frequency of the rectifier and first harmonics, which are the strongest, of the switching frequency are not contemplated in noise measurement.

On the other hand in aircraft applications the main frequency and frequency range of the standard are much closer than in industrial applications. As the figure 3.17 shows, the main frequency is 400Hz to 800Hz and the standard begins to be applied at 10kHz. Since these frequencies are very close it is not possible to hide harmonics of the switching frequency and the EMI filter has to be designed to attenuate the first its harmonics.

3.6.2 Differential input filter design equations

Basically, the differential EMI input filter is composed of a single or multiple L - C states.

Since each stage has 2 parameters, to design a filter 2 equations are needed

Figure 3.17: Frequencies scheme for the EMI industrial standard in contrast with the aircraft standard.

to define the filter. The first condition to consider in the design procedure corresponds to the noise attenuation requirements. To know this requirement the maximum noise level allowed for the respective standard and the level of the converter emission are needed. The switching frequency harmonic is usually the more restrictive then the design equation to compute the cut-off frequency ($\omega_c = 2\pi f_c$) of the filter as a function of the attenuation required and the switching frequency is

$$\omega_c = \frac{2\pi f_{sw}}{\sqrt{10^{att[dB]/20n}}} = \frac{1}{\sqrt{LC}}, \quad (3.40)$$

where n is the number of stages of the input filter.

With this equation the product $L-C$ is fixed then using a second condition or equation it is possible to determine the filter.

In power factor correction systems the EMI filter can affect the power factor of the converter. In particular the capacitance consume reactive power and but this power could be compensated with the inductance (ideally) or by modifying the modulation of the active rectifier.

In order to know the influence of the single stage $L-C$ filter on the power factor, the equivalent impedance seen from the grid (see figure 3.18) can be

$$Z_{eq} = Z_L + (Z_C // R)$$

Figure 3.18: Impedance of the input $L - C$ filter with the power equivalent resistance of the rectifier

simplified to:

$$Z_{eq}(j\omega) = \frac{R}{1 + (\omega CR)^2} + j \left(\omega L - \frac{\omega}{R^{-2} + \omega^2 C^2} \right), \quad (3.41)$$

For the main frequency, the unity power factor is obtained when the imaginary part of the equivalence impedance is null. In order to cancel out the imaginary part of the impedance at the main frequency, the capacitor must be

$$C = \frac{1}{R\sqrt{\omega_c^2 - \omega_{main}^2}}, \quad (3.42)$$

where ω_c and ω_{main} are the cut-off frequency of the $L-C$ filter and the frequency of the grid respectively. R represent the power processed by the rectifier in each phase, i.e. $R = U_{N,RMS}^2 / (P_{load}/3)$. As a general rule, R value used for the input filter design is the corresponding at the maximum power of the application to minimized in this condition the reactive power.

Using this expression for the capacitance and evaluating the equivalent impedance (eq 3.41) using the main frequency this is obtained

$$Z_{eq}(\omega = \omega_{main}) = \frac{R \cdot (\omega_c^2 - \omega_{main}^2)}{\omega_c^2} \quad (3.43)$$

Since the main purpose of the input filter is to attenuate the switching frequency and its harmonics and let the main frequency pass, i.e. $\omega_c > \omega_{main}$ then $\omega_c^2 \gg \omega_{main}^2$ therefore the equivalent impedance is equal to

$$Z_{eq}(\omega = \omega_{main}) \approx R \quad (3.44)$$

The figure 3.19 illustrates the effect of converting the equivalent impedance Z_{eq} to the resistance R . This analysis can be extended to a multi-stage filter configuration as it is shown in the figure. Since at the main frequency the $L-C-R$ network can be simplified as it were a single resistance with that keeps the same value, again this simplification can also be deployed again for the

Figure 3.19: Equivalent resistance of the multi-stage input filter

secondary filter stage. Once this process is done for all filter stages, finally the equivalent circuit is a single resistor which is aligned with the power factor correction concept.

In summary, in order to determine the L - C filter the 2 design equation are 3.40 and 3.42. These equation are also valid for multi-stage filters. In [81] these equation are deeply analyzed using a design example.

3.6.3 Including the effect of the Damping network

In general, damping networks are needed in L - C stages to attenuate the peak value of the resonance of the filter. An excessive resonance peak could bring problem in the stability of the system and unwanted oscillations in steady state. Two typical recommended damping networks are shown in figure 3.20 using an additional capacitor with a damping resistor in series or an additional inductor with a resistor in parallel [82],[83].

In a comprehensive damping network, the damping resistor impedance should be negligible at low frequencies (main frequency in particular) otherwise the damping resistor would dissipate significant power deteriorating the efficiency of the system and creating hot spots.

Since the damping resistors are negligible the damping capacitor is in parallel with the capacitance of the filter and similarly the damping inductance is in series with the inductance of the filter. Therefore to include the damping networks in the design filter guidelines presented in this work, the capacitance or the inductance obtained have to be shared between the attenuation component (L or C) and the damping element (L_d or C_d).

In general, to add a capacitive damping bring a degradation in the power factor in buck-type rectifier applications. For this reason normally in this applications the inductive damping is preferred.

Figure 3.20: Equivalent extension for damping networks

3.7 Control-Oriented Modeling of the Swiss-Forward Converter

As a rule, three-phase rectifier systems need feedback control when the output voltage requires a tight regulation regardless of the disturbances such as changes in the input voltage or frequency in the grid or variable loads, especially if the load is pulsating. Besides, in power factor correction rectifiers the input current must be proportional to the input voltage to achieve a resistive behavior. Furthermore, usually power supplies require some additional dynamic requirements such as minimum bandwidth, settling time, voltage or current overshoots or undershoots, etc.

In order to design a converter that comply with all of this kind of specifications or requirements a dynamic model of the system is needed. This mathematical model has to describe how the system behaves under variations in the input voltage, load current or duty cycles of any transistor.

Modeling is the mathematical representation, of physical phenomena. In engineering, and specifically in power electronics, it is considered a proper model that one that capture the most important and dominant behavior of a system neglecting insignificant but complex dynamics.

An averaged model of a three-phase rectifier, and converters in general, is based on equivalent circuit manipulation where all switching devices are replaced by controlled voltage and current sources. The rest of elements in the averaged model will be the same than the original circuit. These new added sources will replace the actual switching device by its average current or voltage. Thus, all the system dynamics in the range of the switching frequency and above is completely neglected such as the switching ripple in the current of inductors and voltage in capacitors.

The objective of an averaged model is to predict the low-frequency behav-

ior of the system. The resulting simplified model contains physical insight into the system behavior. However, a basic model can always be improved including more system dynamics such as delays in analog to digital conversions, PWM modulators and filters in sensors.

The workflow of this section begins with a complex switching model of the rectifier which is useful to deeply understand the system in details such as modulation, switching ripple, overshoot in diodes voltage, etc. However, a switching model is unnecessary more complex than it is needed for control proposes. For control design an averaged model is more appropriated and in the literature of three-phase systems the most used averaged model is implemented in d - q coordinates.

3.7.1 Averaged model in d - q coordinates of the Swiss-Forward Rectifier

In modeling of three-phase systems a common model used is synchronized respect to a rotary frame called d - q coordinate. In three-phase power factor correction rectifiers an average model in d - q coordinate have a number of advantages respect to a static frame such as the a - b - c frame or the α - β frame [84, 85].

A three-phase electric system without neutral line connection is strictly a two-phase mathematical model, this is because the sum of three currents is always zero i.e. the model is linearly dependent. Hence, a three-phase rectifier can be exactly transformed into in a simpler two-phase system without losing any dynamic of the system. The two-phase system could be implemented in a - b - c coordinate just choosing two of the phases, or through a scalar transformation using the α - β frame. Both the a - b - c and the α - β are static frames but there is a third frame call d - q which has the particularly that it rotates synchronously with frequency of the grid [86, 84, 85, 87].

The use of a rotating frame synchronous with the main frequency brings several benefits in the control of the system. Firstly, in d - q variables to control are two dc quantities instead of three ac quantities reducing the control effort. The linear control theory says that if the reference of a system control is sinusoidal, the regulator requires an infinity gain at the sinusoidal reference to obtain perfect tracking in steady state [88]. In order to get this infinity gain two conjugate poles must be included in the regulator. However, d - q transformation involve a translation in frequency where the frequency of the grid moves to dc (0Hz). Therefore, in d - q the infinity gain required is in dc which can be achieved with an integrator. Furthermore, the integral action of the regulator is also needed to reject the cross-coupling disturbances between the d and q components due to the ac capacitor and inductors of the input filter, as it can be seen in the figure 3.21.

Figure 3.21: d , q model of the Swiss Forward Rectifier including the dynamic of the EMI filter

Since in d - q model the rectifier is interpreted as it were a dc-dc converter, all the vast knowledge in the control of dc-dc converter can be also used in the control of three-phase systems. The figure 3.21 illustrates the averaged model in d - q coordinates of the EMI differential filter and three-phase Swiss-Forward rectifier. As it can be seen, the d and q equivalent circuits are very similar, having the same topology. Since the frame is synchronous to the grid by definition the quadrature component of the input voltage is identical equal to zero ($u_{gq} \equiv 0$). In d - q model, each capacitor has a controlled current parallel and each inductor has a controlled voltage source in series, these sources couple the variables in d with the q circuit and they are called cross-coupling disturbances. The output circuit of the d - q model contains the dynamic of the dc-side of the rectifier.

The d - q modeling is strictly a multivariable system (MIMO), since there are two duty cycles, the direct and quadrature component (d_d and d_q) to control the two components of the input current (i_d and i_q). This MIMO system can be controlled using complex multivariable control theory where the transfer function of the system is a matrix instead of an scalar[88, 89]. In MIMO control theory, when a system each output is closely dependent to only

one input, the system can be model as two or more SISO subsystems adding cross disturbances to model the interaction between internal variables.

In particular, the d - q model of the buck-type rectifier system can be controlled using two independent regulators one for d and other for q circuit. Furthermore, the transfer function of the d and the q circuits are identical, the only difference of these subsystems rely on the cross disturbances. However, since these disturbances are in the same frequency range (low frequency), with a proper regulator design this disturbances are well rejected. Therefore, in general the same regulator is used for the d and the q components.

The question is, if the same regulator can be used for the d and for the q component why is a full d - q model needed to design the control? Could a simpler model be enough to analyze the stability, bandwidth or transients figures (settling time, undershoot, overshoot, etc) of the system? Well, in the next part a reduced order model of the d - q is introduced which still keeps the low frequency dynamic of the system needed to design an appropriate regulator.

3.7.2 Averaged model using and equivalent DC forward converter

Up to now, it has been shown that a three-phase system can be unambiguously model as two dc-dc converters where one converter corresponds to the d and the other to the q component of the system. Although there are some interactions between d and q , in a feedback control loop the cross-coupling disturbances are highly rejected by the action of the regulator, therefore the cross-coupling can be neglected in a simplified model of the rectifier. These two quasi-independent systems are almost identical so it is possible to unified them into just one reduced order model to simplify the analysis. In [90, 91] a reduced order model (R.O.) of a buck-type three-phase rectifier is obtained. The R.O. model obtained is very similar to the d component of the d - q model in a large range of frequencies including frequencies relevant to the design of the control loop.

In [90, 91] demonstrate that from the stability point of view the d component of the d - q model play the most important role and the dynamic of the q are practically irrelevant in this analysis. This is because the d component of the current is used in a feedback loop to control of the output voltage, and the outer loop has a constant power source behavior which is equivalent to a negative resistance. Negative resistances have an inherently unstable behavior unlike the q component that does not have a practical effect in the output power i.e. it is more difficult to make the q component unstable.

Therefore, in a typical three-phase cascade control scheme the loop of the q component has always bigger stability margins than the d component, i.e.

the q model is more robust than the d model. Actually, [90, 91] prove that the q component has practically no relevance in the system control stability and dynamics. Based on this analysis in this paper a reduced order model of buck-type three-phase rectifier is proposed and this model is basically just the d part of the classical d - q model of the rectifier.

Following this idea, the three-phase Swiss-forward can be modeled as a simple dc-dc forward converter with an equivalent input voltage and equivalent input ac filter. The find the equivalences of the dc model parameters with the ac model parameter an equivalent in storage energy is used [92, 93].

AC to DC model parameter equivalence

The instantaneous energy storage in the three-phase inductor is equivalent to

$$\begin{aligned} E_{L,3\phi} &= \frac{1}{2} L \left(\hat{I}_N \cos(\omega t + \psi) \right)^2 \\ &\quad + \frac{1}{2} L \left(\hat{I}_N \cos(\omega t - 2\pi/3 + \psi) \right)^2 \\ &\quad + \frac{1}{2} L \left(\hat{I}_N \cos(\omega t - 4\pi/3 + \psi) \right)^2 \end{aligned} \quad (3.45)$$

$$E_{L,3\phi} = \frac{3}{4} L \hat{I}_N^2 \quad (3.46)$$

Assuming that all this energy have to be storage in one inductor, the equivalent inductance have to be

$$E_{L_{eq},DC} = E_{L,3\phi} \quad (3.47)$$

$$\frac{1}{2} L_{eq} \hat{I}_N^2 = \frac{3}{4} L \hat{I}_N^2 \quad (3.48)$$

$$\Rightarrow L_{eq} = \frac{3}{2} L \quad (3.49)$$

Performing the same algebra for the capacitors and resistor, the DC equivalent capacitance and resistance are

$$C_{eq} = \frac{2}{3} C \quad (3.50)$$

$$R_{eq} = \frac{3}{2} R \quad (3.51)$$

In the figure 3.23 (a) and (b) the three-phase model and the equivalent model of the Swiss-Forward are shown. As it can be seen, the simplified model corresponds to a forward converter with theirs repetitive equivalent elements.

Figure 3.22: steady state free wheeling diode voltage and inductor current waveform in the three-phase model and the equivalent DC-DC mode simulation

Both in the modeling and in the control of the rectifier, the state variables to be used are the voltage of the output capacitor as the current in the dc inductor. The figure 3.22 shows the waveform of these variables in steady state for the switching three-phase system and the switching dc-dc equivalent. Although there are differences in these waveforms because of the fact that in the three-phase model two voltages are applied to the free-wheeling diodes, the averaged value of these corresponding signals is similar.

Once an equivalent dc-dc converter is obtained, the next step is to obtain an averaged model of the forward converter. The average model of a forward converter is exactly the same than the buck converter including the turns ratio as it shows the figure 3.23 (c).

As already said the dc model is a simple simplification of the whole system. The figure shows the voltage in the freewheeling diode and the current dc inductor of the three-phase rectifier and the forward converter. As it can be seen there are some differences in the waveform of these signals. These differences appear because the rectifier actually contains two forward what explains the two voltage levels in the freewheeling diode. These two voltage levels affect the current slope in the inductor as it shows the figure.

On the other hand, the forward with resonant reset converter is a mixed

Figure 3.23: Equivalent DC-DC models of the three-phase Swiss-Forward rectifier. a) Three-phase Swiss Forward with resonant reset, b) DC-DC equivalent switching model of the three-phase swiss forward rectifier. c) DC-DC equivalent averaged model

system with slow and fast state variables [94, 95, 96]. The resonant tank has fast state variables, magnetizing inductance current and the voltage in the resonant capacitor. The output filter of the forward converter has slow state variables. Since the resonant voltage of the capacitor is set to zero at the end of every switching cycle (see figure 3.12), after several switching cycles the initial value of the voltage in the resonant capacitor drop to zero and this state variable lose its dynamic importance. The fact that the initial state of this fast variable loses its information, indicate that the resonant voltage does not qualified as a proper state variable of the converter.

These behavior is equivalent that happens to a PWM converter working in DCM. In this converter the inductor current is set to zero in each switching cycle and therefore the effect of the inductor current can be neglected in a

Figure 3.24: Step response comparison between the three-phase Swiss-Forward rectifier model, the equivalent DC-DC switching model and the averaged non linear model.

voltage mode configuration.

Due to the type of fast state variables that the Swiss-Forward with resonant reset possess, using the partial averaging modeling procedure can be used to obtain a reduced-order averaged model where the dynamics of resonant tank are completely neglected since for control purposes they are not relevant. Therefore in the averaged model shown in the figure 3.23 (b) as well as in the following analysis the resonant tank is not included.

The figure 3.24 shows the output voltage and the inductor current of the three-phase model and the dc equivalent switching forward and average model of the forward converter under a step in the modulation index. As it can be seen under a large step, the transient response of the output voltage of three modes is very similar. The inductor current of three-phase rectifier contains an envelope of the sixth harmonic of the grid frequency, however the averaged value of the transient of the three models also have an excellent agreement for the inductor current.

Even though the three models have similar step responses, for linear control a frequency validation model is needed. The figure 3.25 shows the frequency responses of three-phase switching rectifier model, the switching forward converter and the average forward model. In these tests, the input is the modulation index and the output voltage and the dc inductor current are the outputs of these systems. The frequency range of the analysis starts

Figure 3.25: Models comparison in frequency domain.

at 100Hz and ends at half of the sampling frequency of the rectifier (50kHz). The magnitude of the Bode plot shows that the three models match very well in all the frequencies of interest.

On the other hand, in the plot of the phase it can be seen discrepancy at high frequencies. This effect is generated by the delay that the modulation and sensor generate in the system. The modulation delay can be easily included in the linear model. In the design of the controller this delay is very important since this effect can easily make unwanted oscillations or just make the system unstable.

It is important to emphasize that the main objective of this work is to present a new topology for three-phase rectifier applications and to show guidelines for a prototype design as well as guidelines for the control design. In this last aspect, in this chapter a linear reduced-order model of the Swiss Forward rectifier has been shown for the sole purpose of designing controllers. However, modeling of the Swiss Forward converter is not a fundamental objective in this thesis, in this work only a model has been developed with sufficient

Figure 3.26: Linear model of the Swiss-Forward rectifier. a) Linear model base on voltage and current sources. b) Linear block-model base on impedances transfer functions In this model, capital letters are constant values obtained in the linearization of the system, i.e. N is the turn ratio, I is the inductor current, U_C is the voltage in the input capacitor and M is the modulation index in the operating point.

accuracy for control purposes.

In summary, the averaged model of a dc-dc forward converter has very similar behavior than a linear model of the three-phase Swiss-Forward rectifier. Therefore, the much simpler model contains all the important dynamics needed for the control design system. Finally, in the figure 3.26 shows the linear model using electric circuits in (a) and the linear model using block diagrams. The block diagram scheme is useful in control because it explicitly shows the transfer functions of the plants and the disturbances of the EMI filter.

3.8 Control Scheme

As commonly used to control three-phase rectifiers, the control scheme used in this work have two feedback loops in cascade where in the inner loop the inductor current is controlled in the outer loop the output voltage is regulated. The figure 3.27 illustrates the control scheme implemented in the control of the Swiss-Forward rectifier. The $C_{i_L}(s)$ and $C_{u_{out}}(s)$ are the current and voltage regulators respectively.

Since the cascaded control contains an inner control loop, the system has an inherent current protection under short circuit in the output of the converter. This characteristic is mandatory in this aircraft applications.

In a cascade control architecture, ideally the inner loop has to be much faster than the outer loop, thus the dynamic of both systems is decoupled making easier the design of two regulators.

The figure 3.28 shows the Bode plot of the key transfer functions of the system. The blue line shows of the inductor current as a function of the modulation index and the red line shows the relation between the output voltage and the inductor current. In the transfer function of the inductor current can be clearly seen the impedance resonance of the output filter at 400Hz and the effect of the EMI input filter in the neighborhood of 5kHz.

As it is shown in 3.28, a conservative design for a current control would be in the range of some hundreds Hertz up to 2kHz. Above to 3kHz the delay of the ADC and the PWM actuator are not neglectable and therefore the phase margin of the system could be insufficient.

On the other hand, the regulator of the output voltage should have a bandwidth sufficiently lower than the current loop to obtain a decoupled cascaded system. For this reason, the bandwidth target for this control loop is in the

Figure 3.27: Cascade control scheme for the swiss forward rectifier

Figure 3.28: Bode plots of the transfer function of the plants. The blue line shows the frequency behavior of the inductor current as a function of the modulation index. The red line is the output impedance i.e. this is the transfer function of the output voltage respect to the inductor current

range of 50 to 200Hz.

It is this work that has been chosen to use a fairly conservative control design criterion because the application does not require a more performant system. In the next chapter it is shown that the dynamic requirements of the system are widely met.

More detail of the actual regulators implemented in the demonstrator prototype of the Swiss-Forward rectifier are shown in the next chapter, where further experimental stability issues are explained as well.

3.8.1 Limitation of the control scheme in the power factor correction

The control scheme presented in the figure

3.27 uses a simplified model of the three-phase rectifier where this ac-dc converter is treated as it were a dc-dc converter. This control scheme uses the i_L and u_{out} as feedback signals and it does not measure any ac current. Therefore, the power factor correction is performed indirectly assuming that the voltages in the grid are balanced.

This control scheme ensures constant values for $\langle i_L \rangle$ and $\langle u_{out} \rangle$ in steady state, i.e. the output power is also constant in steady state. Since a single-stage converter does not have power storage capability (at least in steady state) the input power has to be constant as well. In a balanced three-phase grid the only way to have constant power is by supplying a balanced three-phase sinusoidal current at the same frequency. Therefore, by ensuring constant output power the input currents will be irrevocably sinusoidal at the same frequency than the voltages.

However, in spite that this control scheme generates sinusoidal currents, the power factor of the rectifier is not compensated in this control loop since it does not have information about the ac currents and consequently the power factor could be very poor especially at a light load. Using this control scheme, the PWM input current of the rectifier are in phase with the corresponding voltages, however the capacitors of EMI filter generate reactive power and they limit the power factor of the whole system.

The Swiss-Forward rectifier has a certain voltage-current phase correction capability to compensate the reactive power of the EMI filter. However, in this thesis this correction is not implemented in the control scheme but it is performed inside of the modulator.

Experimental Work

4.1 Experimental Setup

The figure 4.1 illustrates the laboratory test bench used in the course of this thesis. It consists of the converter prototype, a three-phase linear power source, a dc electric load, a PC to interface with the prototype, measurement equipments, and a thermal camera.

Some of the most important characteristics of the equipment used are the following

- **Power source HP 6834B:** This is a 150VRMS (phase to neutral) and 10A (per phase) three-phase power source which is very suitable for laboratory tests since it is very versatile in voltage, current and frequency range. Since this is a linear source it has a considerable large bandwidth and lower distortion than a switching power source under demanding loads.
- **Electric load Chroma 63204:** This a 5.2kW programmable dc electric load. Can be configured can demand power as a resistor, constant current, constant voltage and constant power. In general the constant current mode and the resistive mode are useful for steady state validation and to study the dynamic behavior of the control loops the constant power mode is used with load power steps
- **Power analyzer WT1800:** In order to analyze three-phase systems, several current and voltage measurement probes are required, as well as a certain data processing capacity to analyze the electrical quality. All of these tasks are done by this power analyzer. This equipment facilitates the measurements and it is very accurate (0.05%) which is extremely important in the measurement of the rectifier efficiency.

Figure 4.1: experimental setup used throughout the development of the thesis

- **Thermal camera Pyroview 640L:** This is a high resolution infrared camera that allows a constant monitoring of the hottest point of the converter. This camera is also useful to estimate the power loss in a certain component using the temperature rise.
- **Monitoring interfaces:** A PC with the software code composer studio has been used to develop the code for the register configuration of the DSP and control algorithms required for the control of the system. To facilitate the debugging process, the code composer studio allows supervise internal variable and to manipulate them on real-time. In order to have a visualization of key internal variables waveform a digital to analog converter (DAC) is connected from the DSP to an oscilloscope. The oscilloscope is also used to measured high frequency signals such as the voltage in transistors, diodes and transformers.

These are the elements that are permanently used in the development of the tests of the converter. However for specific tests additional equipments are required for example in the analysis of the stability in the feedback control a frequency response analyzer is connected to the converter. This part is explained in detail in the section dedicated to the experimental part of the system control.

Table 4.1: Specifications and requirements for the design of a demonstrator prototype of the Swiss Forward Rectifier

Parameter	Value
Aircraft Derating	yes
Galvanic Isolation	yes
Input line voltage	115Vrms
Main frequency	400Hz
Power Factor	$\geq 98\%$
THD	$\leq 5\%$
Switching frequency	100kHz
Output voltage	270Vdc
Output power rated	3.3kW
Semiconductor Voltage derating	75%
Electrolytic capacitors	Not allowed
Temperature derating	70%

4.2 Swiss-Forward rectifier Demonstrator Prototype

In order to illustrate the design and validate the proper functionality of the Swiss-Forward rectifier a demonstrator prototype has been designed according to the specifications and requirements shown in the Table 4.1.

The figure 4.2 shows the prototype particularly constructed for the experimental validation of the topology concept proposed in this thesis. The demonstrator prototype includes the EMI input filter (In the middle of the two heatsinks), the power circuit of the topology, the sensing circuits required for the feedback and MOSFETs driving circuits. Everything in a single PCB, performing the whole converter on only one PCB has the complexity of combining power track with signal track nevertheless it is cheaper and faster than a design with several PCBs. For these last two reason a single PCB configuration has been selected for this work.

The demonstrator has two heat sinks that allow the heat dissipation of all semiconductors. In addition, the coming out air of the heatsink helps the thermal dissipation of transformers and dc inductors. All power semiconductors are thermally connected to one of the heatsinks. Since these elements are prone to fail especially in the debugging process, the PCB has two large slots that allow the component replacement simply and quickly.

In aircraft applications, the electrolytic capacitors are not recommended, for this reason film capacitors have been used in the output filter. These capacitors have lower energy density than electrolytic ones but they have others positive feature such as self-healing, lower series equivalent resistance, con-

Figure 4.2: Developed isolated three-phase 3.3kW 100kHz Swiss-Forward Rectifier 115Vac to 270Vdc hardware prototype using 1200V SiC MOSFETs, TI DSP controller F28069 Piccolo controlSTICK, including differential and common mode EMI filter. Physical Dimensions: 400mmx210mmx90mm. Weight: 9.4kg. Power Density: 0.44kW/dm³.

stant capacitance over the whole voltage range and an excellent performance for a large temperature range.

For the modulation and control of the demonstrator prototype a commercial digital signal processor (DSP) from texas instruments has been used. To speed-up the development period an evaluation board called piccolo control-stick f28069 that contain all the needed component is mount in the center of the PCB as the figure shows. The f28069 DSP contains a floating-point unit that facilitate the data processing especially when trigonometric functions are required. This DSP also has two 12bits analog to digital converters (ADC) and analog multiplexors to obtain up to 16 analog channels and the DSP also have several PWM modules and communication peripheral such as the SPI used in conjunction with the external DAC. A DAC is extremely useful in hardware debugging because this allows the motorization of DSP local variables such as ADC signals, input voltage sectors, PLL output (θ), etc.

In summary, the f28069 is an inexpensive DSP that has all the digital requirements needed for the rectifier in a single chip. Other platforms such as a FPGAs were discarded because the control requirements of the rectifier are typical and the develop time required generate VHDL code to synthesized all features that are natively included in a commercial DSP.

The converter prototype has been designed to be compact while allowing easy access to all key signals. The prototype weight and dimensions are 9.4kg and 400 x 210 x 90mm³ respectively leading to a 0.44kW/dm³ of power density. The table 4.2 contains a list of the main component of the demonstrator.

Although the demonstrator prototype has been designed for an aircraft grid (115Vrms and 400Hz), the topology proposed in this work is also suitable for applications with public grid (110/220Vrms 60/50Hz). The main difference between a public and an aircraft grid with respect to the functionality of the rectifier is given by the grid frequency. At 400Hz the reactive power generated by the input filter is higher thus the purposed modulation with voltage-current angle compensation is highly desirable in aircraft applications at light load.

4.3 Design consideration in the turns ratio of the transformer

In this application both the input and the output voltage can change just a little then both can be considered fixed for this analysis, thus the product of M times N_1/N_2 is given by eq. (??). This allows optimizing the system analyzing the stress in the semiconductors as a function of the M calculated in the equations shown in the previous section.

For a certain input and output voltage and power, the higher the M is, the lower the $I_{T,RMS}$ is. Fig 4.3(a) shows how $I_{T,RMS}$ changes for M varying

Table 4.2: Summary list of component employed in the Swiss-Forward Rectifier prototype

System	Element	Description
EMI Filter	C_{1stage} L_{1stage}	3x0.56 μ F Murata X7T 450V 170 μ H, 3 stacked cores 58548 N= 21 4 parallel wires of 1mm diameter
	C_{2stage} L_{2stage}	6x0.56 μ F Murata X7T 450V 350 μ H, 2 stacked cores 58438 N= 38 4 parallel wires of 0.8mm diameter
Swiss Forward Rectifier	T_P and T_D D_+ and D_- D_{F+} and D_{F-} D_{N+} and D_{N-} $S_{y,123}$ Transformer L_{out} C_{out}	2xSiC C2M0080120D Power MOSFET C3D25170H Schottky diode SiC C4D30120D Schottky diode Si STTH30R04 Ultrafast recovery diode Si IPW65R037C6 CoolMOS ETD59, N1/N2 = 13/24, Litz wire 400x0.07mm. N2 3 parallel wire 2x150 μ H, core E65, gap 0.6mm, N87 9 parallel wire of 0.6mm 6x40 μ F F339M X2 450Vdc

from 40% to 100%, together with $I_{D+-,RMS}$ and $I_{DF+-,RMS}$. Fig 4.3(b) shows the average currents for these semiconductor devices. From this figure it can be seen that the average currents through the transistors is independent on the value of M since they are defined by the input voltage and output power (20).

On the other hand, Fig. 4.3(c) shows the peak transistor voltage for M from 40% to 80%. In order to obtain a design with the lowest possible transistor RMS current, it is necessary to use a high voltage device. Currently due to improved characteristics of SiC comparing to Si devices, 1200V transistors can be used at high frequency expecting a quite good performance. Using devices such as 1200V SiC JFET from Infineon or the 1200V SiC ZFET from Cree and considering 75% of voltage derating, the maximum possible value for M is equal to 63%. This gives a $I_{T,RMS}$ of 16A.

Figure 4.3: Current and voltage stresses in the main transistors: (a) $I_{T,RMS}$, D_{+-} and I_{F+-} as a function of the modulation index M , (b) peak voltage in the transistor as a function of the modulation index

The bidirectional switches S_{y123} are implemented with two MOSFETs in series connected by their sources. These switches are implemented using 650V/80A IPW65R037C6 transistor with extremely low R_{on} minimizing the conduction losses. The switching losses of these devices are quite low since they switch at twice of the grid frequency.

The voltage stress in the diodes D_+ and D_- is the same as in the resonant capacitor, multiplied by the turns ratio of the transformer (1200V peak). To enable low switching losses, these diodes are implemented by 1700V Schottky SiC diodes C3D251704 from Cree. For D_{F+} and D_{F-} also SiC C3D20060D Schottky diodes rated for 600V are used in order to achieve low switching losses. For the low frequency diodes D_{N+} and D_{N-} , Si diodes STTH30R04 are used. In table 4.2 all the main components are shown.

4.4 Experimental results

In this section shows the main experimental results chronologically obtained. The first tests were related to the ADC to measure the grid voltage to feed the digital PLL for the synchronization. Then to isolate power issues the rectifier was connected as it were a dc-dc forward converter with the equivalent full power to check voltage and current stresses. Finally, the converter was characterized at full power monitoring temperature, efficiency, THD and power factor.

4.4.1 Modulation and control Scheme

The control scheme of the Swiss-Forward Rectifier is shown in Fig. 4.4. The input line-to-line voltages are measured and fed into a digital PLL which provides the synchronized angle θ and consequently the sector and interval of the grid. Depending on the value of θ , the bidirectional switches are turned on and off according to the grid sector. For open loop operation, the modulation index M and the current angle ϕ are fixed and the transistor duty cycles d_{T+} and d_{T-} are determined with θ and the interval according to the modulator block (white box) in Fig. 4.4.

The complete control and modulation schemes are implemented in a low cost floating point DSP, i.e. the TI F28069 control stick. It takes $16\mu s$ for the DSP to compute all the functions (PLL, PI controller, trigonometric functions, etc), thus the sampling frequency is conveniently set to 50kHz ($20\mu s$) and the duty cycle of both transistors is updated every two periods to reach the desired switching frequency of 100kHz.

Figure 4.4: Modulation and Control system implemented in a DSP to control the Swiss-Forward Rectifier

4.4.2 Grid synchronization

In order to generate sinusoidal input currents with the lowest distortion possible is important to filter out the voltage distortion of the grid. This voltage distortion can come from the source itself or as an effect of the interaction between the converter and the output impedance of the EMI filter. To face the voltage distortion there are two common alternatives, one is using a digital IRR band-pass filter and the second is using a PLL.

In this work for the grid synchronization a PLL was implemented. In the figure 4.5 a diagram block of the PLL is shown. The three voltages are transform into d and q coordinates and the q component is controlled with PI regulator using a zero as reference value. The output of the PI regulator is the value of θ and this is feeding back to the d - q transformation block. In some PLL scheme a double integration action is included since the theta has an sawtooth behavior, however in this work the single integration was sufficient.

Fig. 4.6 illustrates the grid voltage of the phase a and the output of the digital PLL. The θ goes from $-\pi$ to π and since $u_a = -\hat{u}_a \cos(\theta)$ the PLL is synchronized with the input voltage.

Figure 4.5: PLL Synchronization scheme used in the modulation of the Swiss-Forward Rectifier

Figure 4.6: DSP output signal. Channel 1 (blue), PLL output θ . Channel 2 (light blue), duty cycle of T_P . Channel 3 (red), duty cycle of T_N . Channel 4 (green), line to neutral voltage (v_{aN}). The digital signals show the drive signals to the bidirectional switches

Using the value of θ , the sector and interval of the grid are identified, allowing to generate the gate signals for the bidirectional switches. Also, the duty cycles for the high frequency transistors, d_{T+} and d_{T-} are shown using $\phi = 0$.

4.4.3 Voltage stress in HV transistors

Once the synchronization was done, the second series of tests were related to reach the nominal conditions of the converter. For this propose, the rectifier prototype was connected as it were a dc-dc forward converter. Each forward converter was independently connected to a dc power source using the maximum power condition of the forward converter i.e. 240Vdc at the input, 270Vdc at the output and 1.6kW output power.

In this condition, the high voltage SiC transistor withstand the worst-case

Figure 4.7: Resonance behavior between transformer and parasitic capacitances. Channel 1 (blue), forward diode current. Channel 2 (light blue), freewheeling current diode. Channel 3 (red), transformer primary voltage. Channel 4 (green), Drain source transistor voltage.

voltage and the power converter is equivalent to the average power of each forward converter. Fig. 4.7 shows the signals of the resonant reset circuit. The worst-case of the T_P transistor voltage is identified when $u_{PY} = 3/2\hat{U}_N = 244V$. The peak transistor voltage is 920V while the expected value is 900V. This peak can be reduced by adding more capacitance in the resonant circuit. However, the size of this capacitance is in the order of hundreds of pF and is difficult to adjust perfectly. Furthermore, additional few volts in the resonance circuit does not degrade system performance. The blue and light blue lines show the high frequency diode currents. While the resonance is happening, D_{F+} is conducting and D_{F-} is blocking. When the resonance finishes, both diodes share the inductor current. When the transistor is turned on, unwanted current oscillations occur in D_{F+} due to parasitic capacitances.

During the dc-dc prototype tests, there were some issues related to the voltage in the high voltage transistor. The Swiss rectifier topology has a diode bridge between the input filter (capacitor) and the switching transistor figure 4.8 (a). This diode bridge disconnects the input capacitor of the EMI filter when the resonant capacitor and the magnetizing inductance are resonating. On the other hand, a capacitor that minimizes the parasitic inductances of the circuit transistor and transformer have to be placed in the middle of the diode bridge and the transistor.

Ideally this capacitor should be small enough so as not to disturb the low frequency of the input current and it should have an excellent behavior at high frequency (low series equivalent inductance). Taking these considerations into account, a SMD ceramic capacitor was selected for this purpose as it is shown in the figure 4.8(b). However, this capacitance value is too small and it has a negative effect in the switching losses of the transistor. When the transistor

Figure 4.8: Experimental issues related to the input capacitance between the diode bridge and the transistor.

is open there is always a connection between the drain to source given by the parasitic capacitance (C_{oss}). When the magnetizing current is feeding power back to the primary, the parasitic capacitance of the transistor begins to discharge and this current is closed by the input capacitor and not by the EMI filter. Since this capacitor is too small there is a significant rise voltage in the capacitor but in the transistor as well. Then when the transistor is turned on, in the next cycle the voltage to be blocked is higher than the input voltage.

The simplest solution for this problem is to place a higher capacitance keeping in mind the effect on the input current. For this prototype, increasing the input capacitance a couple of order of magnitude (74nF) is still small enough respect to the 400Hz and big enough to obtain a neglectable voltage ripple as is shown in figure 4.8(c).

4.4.4 Measured efficiency of the prototype

The efficiency for the whole rectifier system in conjunction with the total harmonic distortion of the input current of the rectifier are shown in figure 4.9. These measurements have been performed with a precise power analyzer Yokogawa WT1800 [97] and the measurement of the efficiency include losses in the rectifier itself and the EMI filter as well as the losses due to power consumption of fans, DSP and transistor drivers. The peak efficiency is 93.6% and it is obtained at maximum power.

Figure 4.9: Efficiency of the demonstrator prototype from light load to nominal load.

4.4.5 Measured input current THD of the prototype

The input current THD is relatively high at very light load (600W) but this drastically decreases below to 5% at 1kW of load. From this power to the nominal power the THD varies from 4.0% to 2.5%. Within the input current requirements, it is stipulated that the THD must be less than 5% at nominal power therefore, it is shown that the prototype fulfills this requirement.

4.4.6 Input current waveforms at different power levels

In order to illustrate how are the input current of the rectifier, the figures 4.11 and 4.12 show the input current waveforms of the three phases and the input voltage of the phase (a) to have the reference for θ . These figures show the behavior of the input currents from light load to full load. At is can be seen, the three currents are well balanced in all conditions, although the ac currents are not measured or controlled in the control scheme used in this prototype.

Figure 4.10: Total harmonic distortion of the demonstrator prototype from light load to nominal load.

The figures 4.11 and 4.12 also show that the currents have a certain low frequency distortion when two line voltages are crossing. Similarly, as it happens in the classic Swiss Rectifier the Swiss forward Rectifier also has low frequency distortion in the input currents that affect negatively in the THD of the input current. This distortion is caused by the switching frequency voltage ripple across the input filter capacitors in the moment when the voltage of two lines are crossing. The generation of this distortion is explained in detail in [98, 99, 100].

Basically, the problem is given when the voltage between phases is smaller than the voltage ripple in the input capacitor of the buck-type rectifier. When this happens the three-phase diode bridge clamps the voltage of two of the input capacitors generating a voltage distortion in these capacitors leading to a distortion in the input current of the rectifier. The simplest way to attenuate this distortion is by increasing the switching frequency or by increasing the input capacitance in order to decrease the voltage ripple in the input capacitors, however these solutions have negative impact in the efficiency and in the power factor of the system.

In [100] a modification of the Swiss Rectifier replacing the diode bridge with transistors is proposed. This is a relative complex solution since requires additional hardware (transistors and their gate circuits) and a more complex modulation algorithm and unfortunately it does not prevent completely the distortion but it is highly mitigated.

For this work the limit specified for the THD of the input current is not very strict since the maximum limit is set at 5% and only have to me comply at the nominal power (Chapter 1). The figure 4.10 (b) shows the THD of the

Figure 4.11: Input current waveforms st light load

Figure 4.12: Input current waveforms st high load

demonstrator prototype is below 4% in the most part of the range of power. Considering that the THD limit is wide and that an advanced modulation scheme leads to a considerable high effort in hardware and software, in this work only the simple modulation is considered. The THD obtained

As the figure 4.11 illustrate, at light load the input current and voltages are shifted and the power factor is not unitary. At very light load, the phase between currents and voltages is close to 30 degrees however as long as the load is higher this phase is getting smaller. At full power the phase is close to zero. For the case of this prototype the current leads the voltage i.e. this prototype has a capacitive power factor. As it is shown in the first chapter, in aircraft application a capacitive power factor is very undesired therefore, in this thesis phase compensation is introduced in the modulator to maximized the power factor. Experimental results of the compensation are shown in the next section.

4.4.7 Phase compensation of the input current

In the classical modulation of the Swiss rectifier presented in [56], the rectifier is controlled to behave as a resistor. In general, in rectifiers the resistive behavior is the goal of PFC systems, however there are certain applications where some reactive power is also required to compensate the reactive power of others loads connected to the same power main. As it was mention in the previous chapter, this rectifier has a certain capacity to handle reactive power. The input current in the Swiss rectifier can be 30 degree either leading or lagging the input voltage.

On the other hand, the input filter of the rectifier can also generate some reactive power since this is compose by a L - C network. The amount of reactive power of the input filter depends on the demanded power. For instance, in the prototype of this work, at nominal load the filter does not injected reactive power (it is negligible) while at light load the most part of the input apparent power comes from the filter, in fact when there is no load (zero load current) the equivalent impedance of the rectifier system is determined by the L - C input network therefore all input power is reactive.

In aircraft applications the reactive power of three-phase systems is very limited, therefore is extremely important to modify the standard modulation of the Swiss Rectifier in order to compensate the reactive power of the EMI filter. In work a new modulation is proposed and in this section an experimental validation of it is presented.

The figure 4.13 shows the key signals of the standard modulation. For this test 2 synchronized oscilloscopes are used to analyzed 6 signals at the same time. In the upper part of the figure signals from the DAC (digital to analog converter) are shown. These signals are the duty cycles of the 2 high frequency transistors, the synchronization angle θ and the measured voltage

Figure 4.13: Modulation without phase compensation. Above signals from the DAC (digital to analog converter) that show the duty cycles of the two high frequency transistors, the output of the PLL (θ) and the DSP measured voltage of the phase. Below the input voltage and current of the rectifier

Figure 4.14: Modulation including phase compensation. Above signals from the DAC (digital to analog converter) that show the duty cycles of the two high frequency transistors, the output of the PLL (θ) and the DSP measured voltage of the phase. Below the input voltage and current of the rectifier

Figure 4.15: Power factor of the demonstrator prototype with and without phase compensation modulation

of the phase A of the three-phase input voltage. In the bottom part of the figure, the input current and voltage of the phase A are shown.

As it can be seen in the figure 4.13, the input current of the rectifier system is leading the voltage around 30 degrees. In the standard modulation the rectifier behaves as a resistor therefore the phase is generated by the input filter.

In order to illustrate the effect of the proposed modulation with phase compensation, the figure 4.14 shows the same signals under the same rectifier input and output condition but using the new modulation. For this test the compensation angle is set at 30 degrees ($\phi = 30^\circ$) as it can be seen in the waveform of the duty cycles since they are equal to the shown in the figure 3.9 (b). As a result of this modulation, the input current now is in phase with the voltage obtaining a power factor close to the unity.

In order to illustrate how much the power factor is improved using the proposed modulation, the figure 4.15 shows the power factor of the prototype as a function of the output power using the 2 modulations. As it shown, the power factor is highly improved from 0.5kW to 1.5kW. In addition, with the proposed modulation a practical unitary power factor is obtained from 1.3kW while with the standard modulation it is around 2.5kW.

However, although this new modulation substantially improves the power factor of the rectifier at light load, when the current load is zero, inevitably the power factor of the rectifier system is zero no matter what modulation is used.

4.4.8 Thermal behavior of the prototype

Although the main objective of the prototype has been the practical validation of the Swiss Forward rectifier, in the design of the demonstrator aircraft requirements and specifications have been considered to build a representative rectifier that can be compare with a commercial one. In this context, it is important to study the thermal behavior of the prototype especially in key component such as semiconductors and magnetics.

For all steady state measurement such as power factor, THD and efficiency the prototype needs to reach and stable temperature in all components. However, since this prototype has several bulky and massive elements that slow down the thermal evolution, for each measurement the prototype has to work for a long period.

After several tests, it has been found that the transformers are the devices that take longer to reach the stationary temperature. As it shows the figure 4.16 transformers take 50 min to reach an stable temperature. Therefore, all stationary measures were taken after an hour. The figure 4.16 shows the temperature evolution of the hottest point of the 2 transformers when the system is working at nominal power. This test was performed in summer when the laboratory ambient temperature is around 37°C (initial temperature) and the final transformers temperature after 80min are 70°C and 66°C, thus the

Figure 4.16: Transformers temperature transient

Figure 4.17: Thermal photo of the top view of the prototype after working for 90 minutes at nominal power

temperature rise is around 30°C. In aircraft applications due to the magnetic reliability the maximum core temperature allowed is 110°C then considering an ambient temperature of 80°C (typical case) the temperature rise allowed is 30°C therefore the design of transformer of this prototype comply with this aircraft thermal requirements.

In order to study the temperature of the rest of components of the prototype, in figures 4.17 and 4.18 thermal images of the top side and the bottom side of the rectifier as well as equivalent pictures to illustrate what devices are the critical thermal components. As it can be seen, thanks to the large heatsinks, the hottest point in semiconductors is only 63°C. this means that

Figure 4.18: thermal photo of the bottom view rectifier system after working at nominal power for 90 min

the heatsinks are oversized which means that power density of the demonstrator could be easily improved just optimizing heatsinks.

In the thermal top view of the demonstrator there are three components that are particularly hot. In the down-right corner of the figure the shunt resistor used to measure the inductor current is at 77°C , but this is in the nominal range of this resistor. The other two hottest points in the PCB correspond to a couple of film polyester capacitors. These capacitors are placed to minimize the parasitic inductance of the transistor and primary winding loop shown in the figure 4.8. Similarly, that in the shunt resistor, although the working temperature looks high, this temperature is still lower than the max temperature specified for the capacitor manufacture (125°C).

In the figure 4.18 are shown the heatsinks, the EMI filter, Transformer, DC common mode choke, DC inductors and the output capacitors. Among these components, transformers are the critical elements in terms of temperature however the temperature rise is compatible with aircraft standards. Between both transformers there are 4°C of difference in the final temperature, this is due to differences in equivalent series resistances of transformers. These transformers are custom made and they were built by hand, this manufacture process can bring certain differences in the relative position between primary and secondary windings. Tolerance in the windings-distance affect the leakage inductances and the equivalent series resistance leading to a different transformer dissipation.

4.5 Closed Loop results

In the modelling section of the previous chapter, a simplified model of the three-phase Swiss-Forward rectifier has been proposed. This model states that if the rectifier is being supplied by a balanced grid, the full rectifier system can be model as it were a dc-dc converter. This simplified model contains all relevant dynamics for the study of the stability and control of system. Therefore, for control proposes the rectifier is conserved as a dc-dc converter since just the output dc side of the rectifier is analyzed in this section. The control scheme contains two closed loops, in the inner one, the inductor current is controlled and in the outer loop the voltage in output capacitor is controlled. This control scheme does not include measures and control of the input ac current of the rectifier.

In the modelling section of the previous chapter, a dc equivalent averaged model of the Swiss forward has been presented. This analytical model helps to interpret the dynamics of the system in a theoretical way, emphasizing the dominant dynamics of interest. This theoretical analysis serves to understand in a general way the Swiss Forward rectifier for a wide range of system parameters such as inductances and capacitances. However, in the case of the

Figure 4.19: Open loop gain measurement scheme using the vector Network Analyzer Bode 100

analysis of a specific prototype rectifier, it is more effective to analyze the system in a practical way directly measuring the plant.

In order to synthesized the inductor current and output voltage regulators, a practical approach has been utilized in this work. This practical way to design regulator is very effective due to the direct measurement of the plant and the open loop gain of a system including the phase and the gain margins.

To measure the stability of the prototype, a frequency analyzer called Bode 100 [101] was used. This device uses the signal injection method to characterize the system. The injection signal is inserted into the feedback loop through the current and output voltage sensing circuit. This signal is connected between two operational amplifiers to take advantage of the low and high output and input impedances of the operational amplifier, more detail of the implemented circuit is shown in the appendix of this thesis.

The figure 4.19 illustrates the experimental setup used for the design of the current controller. The Bode 100 through and isolator transformer inject a disturbance, the frequency range of this disturbance is between 10Hz to 25kHz. The ADC of the DSP is sampling at 50kHz and this measurement includes the inductor current information and the disturbance. The DSP with the discretized regulator generate the modulation index and with the measurement of the main voltages the modulator generates the transistor's pulses.

In order to measure the transfer function between the modulation index to the inductor current the system is closed with a proportional regulator and with the measurement of the open loop gain it is possible to obtain the target plant. The proportional regulator should only affect the gain in the open loop, however, since the conversion of data (ADC), the calculation time and

Figure 4.20: Modulation index to inductor current plant measurement using a P regulator

the PWM modulator generate transport delay, this effect is also included in this measurement. Although the delay effect is not part of the plant itself, it is very important to incorporate this dynamic in the design of the regulator since the transport delay of the DSP can generate instabilities in the system if the bandwidth of the loop gain is excessive.

The figure 4.20 shows the open loop gain at nominal conditions (u_{in} : 115Vrms, u_{out} : 270Vdc, P_{out} : 3.3kW). The blue line shows the open loop gain using a proportional regulator ($k = 0.01$), and the red and black lines are also open loop gain using the same proportional regulator with one and two poles at 1kHz and 2kHz respectively. Since the blue line is the open loop gain using a constant gain of 0.01, the plant of the inductor current over the modulation index is the blue line plus 40dB in the whole frequency range. The regulator with low-pass filter (red and black lines) show how the resonance of the input filter can be attenuated. Other effect to point out is drop phase generated by the delay of the DSP. In the middle range of frequencies, the open loop gain in red has higher phase and this corresponds with the fact that the poles of the regulator for this system is twice the pole of the open loop gain shown with the black line. However as long as the frequency increases this difference is smaller due to that the effect of the delay in both system is significantly higher neglecting the effect of the regulator's poles.

The figure 4.21 shows the open loop gain of two implemented regulators, in blue an open loop gain obtained with a PI regulator is shown while in red is shown with a PI plus a pole in 1kHz. The purpose of the additional pole is

Figure 4.21: Bode and Nyquist plot of the open loop gain of the current loop.

to reduce the gain and the phase of the resonance due to the LC input filter. It is important to highlight that both regulators have the same bandwidth (600Hz) however from the stability point of view the PI plus low-pass filter is a far more robust design. The PI with the low-pass filter (red line) is a more robust design however it has lower gain in low frequency than the PI regulator (blue line). An open loop gain with low gain affect the capability of disturbance rejection of the closed loop system however in this work the disturbance rejection has not a strict requirement and therefore a the PI plus low-pass filter regulator is chosen because of its better robustness.

As it was mention before, the resonance of the LC input filter reduces the margins of stability, in particular in the open loop gain with the PI regulator the gain margin is quite small, approximately 6dB, which is insufficient in some applications (industrial or in aircrafts). On way to improve the stability margins is by reducing the bandwidth of the system, however this deteriorates the dynamic response of the system and the output impedance of the rectifier. In order to improve the robustness without harming the bandwidth of the pole of the current loop an additional pole is added in the regulator (line red). This pole only attenuate the high frequency of the system keeping the same gain at low frequencies. The low-pass filter minimizes the gain of the open loop gain but also reduces the phase of the system. As a consequence of this pole both stability margins are substantially improved. In the figure 4.21 (b) a Nyquist plot of the two open loop gain is shown. In this graph can be clear seen how the system with the PI plus low-pass filter is more robust since the red curve is much further to the -1 point than the blue line.

Therefore, using adding a pole in a PI regulator, is it possible to minimize the resonance of the LC filter keeping a high bandwidth in the current loop.

Figure 4.22: Positive step power response from half to the nominal power of the demonstrator

Figure 4.23: Negative step power response from nominal to the half power of the demonstrator

For the design of the regulator of the outer loop the same methodology has been followed i.e. measuring the open loop gain using the output voltage measured and a proportional voltage regulator. With this frequency response, and the dynamic requirements for the outer loop such as the settling time and the voltage drop, a PI voltage regulator has been designed and implemented in the DSP.

In the figures 4.22 and 4.23 the power step responses of the outer voltage loop are shown. In the upper part of the figures the output current and voltage are included, and in the bottom part the ac input current and voltage. These power steps are performed at nominal input and output voltages, the power step is from half to the nominal power and vice versa. As it can be seen, despite of the control is implemented in the dc side, the transient of the input ac current does not possess any strange behavior, in fact this transient behavior is similar to a PFC with a control in input current.

The voltage drop is 25V and the settling time is 20ms. Considering the voltage drop and settling specified in the MIL-STD 704 (see figure 1.11) are 60V and 100ms respectively, it is shown that the dynamic characteristic obtained with the demonstrator full-fills this aircraft standard.

Conclusion and Future Work

Learning grows until it dawns on you.

Moshé P. Feldenkrais

This work presents an isolated single-stage three phase rectifier based on the Swiss Rectifier combined with a forward with resonant reset suitable in applications with a tight input/output voltage range. Besides the benefit of the isolation itself, the transformer allows the up or down scaling of the output voltage of the rectifier, maximizing the applicability of the concept.

The main contributions of this work are the proposed new isolated three-phase rectifier, the new methodology for the design of the EMI filter and the new modulation of this rectifier that compensated the reactive power of the EMI filter. Both the EMI filter design and the new modulation technique can be also apply to the others buck-type rectifiers.

Single stage topologies have certain limitations in the way that the power is processed. The lack of an energy storage element, such as DC-link capacitor or inductor, prevents the compensation of the pulsating power that an unbalanced grid has. However, in aircraft applications the phase balanced grid is regulated therefore a single stage rectifier can be used in these applications. Single stage applications have potential benefits in lower complexity, higher power density and higher efficiency than 2-stage topologies.

The main principle of operation, and the differences comparing to the ordinary Swiss Rectifier have been discussed. An input voltage/current angle compensation has been proposed as a part of the modulation in order to increase the power factor at medium and light load. This modulation can be implemented in the Swiss Rectifier as well. In addition, a design methodology for the design of input EMI filter in applications where the grid frequency

is higher than traditional industrial grids e.i. 400Hz or 800Hz. The main improvement of this new design methodology is minimized the reactive power of the EMI filter in a particular load power maximizing the power factor of the system.

The proposed concept was experimentally verified in a prototype that was designed and built using aircraft specifications (115VAC at 400Hz). Nevertheless, the proposed methodology can also be used in applications with the public grid (50/60Hz), since the grid frequency does not affect the selection of the semiconductor devices or the design of the transformer and the output DC filter. The grid frequency only affects the reactive power generated by the input AC filter and this work demonstrates the reactive power compensation by modifying the modulation.

5.1 Future work

In the field of isolated PWM single stage three-phase rectifier there are several research lines to continue. For example, there are still some gap of improvement in power density, efficiency, costs and durability. To carry out this potential improvement it is necessary to employ multi-objective optimizations rather than optimizing each part of a full complex rectifier system.

In this work, a deep analysis of a new rectifier topology has been done. The result of this analysis has shown that this topology has a serious potential to become a commercial product however there are still many aspects left that have to be analyzed aiming at final product. The most important thing that has to be carry out is the multi-objective optimization to proper design a system that considers harmonically the important figures of a rectifier system such as power density, efficiency, costs and reliability. The only way to make a fair and unbiased comparison between two topologies is after developing multi-objective optimizations using the same specification and requirements for both systems. Unfortunately, this kind of detail comparisons entails a lot of time and effort due to that develop of an optimization algorithm and prototyping process are very time consuming.

In order to improve the power density and the efficiency at light load it could be interesting to divide the system in cells placed in parallel. Turning off cells as long as the load power is being smaller increases the overall efficiency. In addition, using interleaving allows to cancel out an important part of current ripple in the inductor. In this way, the filter effort in the input and in the output of the converter is reduced, and this has a positive effect in the volume and weight of full system improving the power density. Finally, an interleaved forward converter can also have dynamic benefits since the transient response is faster than a classical forward converter.

Therefore, the interleaved Swiss-Forward converter shown in the figure

Figure 5.1: Interleaved Swiss Forward rectifier with resonant reset

5.1 is an excellent option to explore as an isolated single-stage three-phase rectifier for future work. The main benefit and motivation to use interleaving is related with THD of the input current. As it was explained in this work, the swiss converter and all its variants suffer the same problem, when two input voltages are equal or very close, the voltage ripple in the input capacitor avoids a proper modulation and this effect generate a distortion in the ac currents every 60 degrees. With the interleaved Swiss-Forward Rectifier the voltage ripple in the input capacitors are much smaller leading to an smaller input current THD.

One of the main motivation of using a forward with resonant reset converter was motivated by the development of new semiconductors. This forward converter is historically limited for low power applications because it is a topology with hard switching and the resonance for the demagnetization of the transformer oblige the use of high voltage semiconductors. However, the technological advances in semiconductors, especially in SiC, have enabled applications with the high voltage transistors (kV) in conjunction with high switching frequency (several hundred kHz).

Today the SiC semiconductors are already well positioned and already have defined their niche market. In the near future, it is expected that GaN devices will drastically improve the figure-of-merit of semiconductors. The market of high voltage GaN transistors is targeting the range of 600V which is not high enough for the Swiss-forward with resonant reset rectifier. For this reason, as a future work is very interesting to explore other topologies for the swiss converter with lower stress voltage. For example the forward can

Figure 5.2: Isolated Swiss-Push-Pull rectifier

be replaced by two full-bridge converter or a push-pull converter as the figure 5.2 shows.

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Swiss Forward Rectifier

Experience can be merely the repetition
of same error often enough.

John G. Azzopardi

This chapter presents an introduction to different algorithms used in meta-heuristic optimization. The objective of this chapter is to help understanding the underlying principles of these techniques.

A.1 Control Swiss Forward PCB project

Figure A.2: Current mirrors

Figure A.3: Input Voltage sensors

Figure A.5: 4 Chanel DAC connector

Figure A.6: Auxiliary isolated power supply

Figure A.7: Top and bottom layers

A.2 Isolated driver PCB for SiC transistors

Figure A.8: Electrical schematic of the isolated MOSFETs drivers

Figure A.9: Isolated power supply

Figure A.10: Layout

A.3 Output voltage and DC inductor current PCB

Figure A.11: Output voltage measurement schematic

Figure A.12: DC inductor measurement circuit

Figure A.13: Power supply for the output sensors PCB