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**Electricity Generation with Direct Drive Wave Energy
Converters: Resistive Loading Control, Grid Integration
and Wave Prediction**

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Author: **Hugo Rocha Mendonça**

Master in Electrical Engineering

Supervisor: **Sergio Martínez Gonzalez**

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To my parents, wife and my boys.

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Resumen

Existe un consenso global sobre la generación eléctrica a partir de fuentes de energía renovables. En este marco, la energía oceánica presenta un gran potencial y puede ser explotada de distintas formas. Prueba de esto es que la energía oceánica ha sido tema de muchas investigaciones recientes. Hay un grupo significativo de tecnologías empleadas para la conversión undimotriz mediante convertidores de energía de las olas (WEC – del inglés: wave energy converter). Esta tesis se centra en el sistema de accionamiento directo de un convertidor del tipo absorbedor puntual, en el que una boya está conectada a un sistema de extracción de energía (PTO – del inglés: power take-off) en el lecho marino. La conversión de energía en el PTO se realiza mediante un generador lineal de imanes permanentes.

De manera general, el gran desafío para la explotación de las energías renovables es aumentar su eficiencia. En la generación undimotriz no es diferente. La operación del convertidor debe ser optimizada para que sea factible económicamente su implementación. Esta tesis propone una estrategia basada en un control de fuerza resistiva de un absorbedor puntual, donde la potencia generada es maximizada mediante el control de la fuerza electromagnética del generadora través de un sistema emulador de resistencia.

Estos WEC se conectan a la red eléctrica mediante un sistema de electrónica de potencia, que se aprovecha para implementar el sistema emulador de resistencia mediante un controlador de corriente. Este debe ser capaz de manejar grandes variaciones en las magnitudes eléctricas, además de intervalos de tiempo largos en que estas presentan un valor bajo. Para ello, se ha desarrollado un algoritmo de fuente de corriente no lineal (banda de histéresis) mediante vectores espaciales con frecuencia de conmutación constante para controlar el generador eléctrico. Además, se ha implementado un controlador tradicional por banda de histéresis y un controlador PWM por vectores espaciales basado en un control orientado a la tensión para comparar sus prestaciones desde el punto de vista de la eficiencia del WEC y su respuesta dinámica.

Muchas de estas estrategias de control son no causales, es decir, dependen de datos

futuros del oleaje. Sin embargo, la predicción del oleaje en un parque marino puede que no sea económicamente factible, ya que se necesitarían muchos instrumentos de medida. Como parte de un sistema de predicción, en esta tesis se plantea un estimador espacial del oleaje en un parque marino basado en un único punto de medida. Se ha implementado este planteamiento mediante una red neuronal artificial con retraso temporal. En este estudio se evalúa el desempeño de dicho estimador y su sensibilidad a distintas situaciones.

Por otro lado, el nivel de penetración de energías renovables en la red eléctrica está en aumento día tras día. La eólica y solar han alcanzado un alto nivel de madurez y, por tanto, sus impactos están bien definidos y estudiados. Pero este no es el caso de otras fuentes de energía emergentes como es la undimotriz. Así, en esta tesis, también se estudia el impacto de la potencia fluctuante inyectada por un WEC en una red de distribución eléctrica. Además, se propone una estrategia para mitigar las fluctuaciones de tensión resultantes. El método de corrección propuesto utiliza la propia electrónica de potencia del WEC para controlar la potencia reactiva disponible en el convertidor del lado de la red. También se plantea el uso de un STATCOM como dispositivo adicional para la compensación de reactiva.

Abstract

There is a global consensus on the use of renewable energies. In this context, the sea has great energy potential to be exploited in different ways, and it has been the subject of many studies. A significant group of technologies is based on the conversion of the ocean wave energy into electricity by a Wave Energy Converter (WEC). This thesis is focused on the direct drive system of a point absorber converter, where a buoy is connected to a Power Take-Off (PTO) system on the seabed. The energy conversion in the PTO is made through a Linear Permanent Magnet Synchronous Generator (LPMSG).

In general, a major challenge for the exploitation of renewable energies is to improve their efficiency. In electricity generation from the energy of ocean waves, not unlike other technologies, the converter must be optimised to make the energy harvesting economically feasible. This thesis proposes a resistive loading control strategy of a point absorber in which the generated power is maximised by controlling the electromagnetic force of the generator with a resistance emulation approach.

As a power electronic interface is required for grid connection, the resistance emulation approach has been implemented by a current controller. It must be able to handle with large variations of voltage and current, and also with large periods where these variables present low values. Therefore, a current source with a space-vector hysteresis-band and a constant switching frequency has been developed to be used with surface-mounted LPMSG. Also, a traditional hysteresis-band controller and a space-vector PWM controller based on voltage oriented control are implemented. This work assesses the performance of the current controllers mentioned above from the point of view of the overall conversion efficiency and dynamic response.

Most control strategies to maximise the captured power by wave energy converters depend on the knowledge of the incoming wave in a short-term future. However, the wave prediction may be unpractical in a wave farm, since many measurement systems would be required. As a part of a prediction system, this thesis presents an estimator of the spatial sea surface elevation in a wave farm based on a single measurement point. The approach

is based on a time delay artificial neural network. This work also explores the performance of the estimation for a given farm and the sensitivity to different sea characteristics.

On the other hand, renewable energy is steadily increasing its penetration level in electric power systems. Wind and solar energy have reached a high degree of maturity, and their impacts on the grid are well known. However, this is not the case for emerging sources like wave energy. In this sense, this work explores the impact of the fluctuating power injected by a wave energy converter on the distribution grid voltage and proposes a strategy for mitigating the induced voltage fluctuations. The proposed corrective action makes use of the hardware resources already available in the WEC, employing a control strategy on the reactive capability of the grid-side converter. The use of a STATCOM as additional reactive compensation equipment is also explored.

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Introduction

Sustainable development and climate change are amongst the biggest challenges facing 21st-century engineering. Facing both issues, since 1992, under the supervision of the United Nations Frame-Work Convention on Climate Change (UNFCCC), several global processes and meetings have taken place. From this convention, the Kyoto Protocol subsequently arises, in 1997, whose purpose is to control and reduce the emission of polluting gases into the atmosphere. According to the International Energy Agency (2020) (IEA), in 2007, electricity and heat are responsible for 27% of total carbon dioxide (CO₂) emission into the atmosphere. In 2018, this value increased to 41%. In this context, the European Environment Agency (EEA) is adamant in stating that renewable energy sources play a crucial role in reducing CO₂ emission in the energy sector. The European Union (EU) has been encouraging investment in this area. For example, between 1992 and 2008, the EU achieved 17% reduction in CO₂ emissions related to the generation of electricity and heat. Although more modest, the reductions achieved by the United States and China were 6% and 5%, respectively. Despite these reductions, the global trend is still an 8% up.

Given this scenario, there is currently a worldwide consensus regarding the generation of electric energy through renewable energy sources. For example, the EU, through the Horizon 2020 work programme, has the goal of reaching 20% of all electricity consumption covered by renewable energies by the year 2020. So, research on the integration of these types of energy has been intensified from the point of view of economic and technical conditions and there has been a significant increase in their use. Although wind energy has experienced the most significant growth, other forms of renewable energy stand out due to their great resource potential and are also object of current interest, such as solar and ocean energy. The latter presents an estimated potential range from 20000 TWh to 80000 TWh of electricity a year (Mofor, Goldsmith, and Jones, 2014). The main forms of ocean energy are waves, tides, marine currents, salinity gradient and temperature gradient.

Mofor, Goldsmith, and Jones (2014) also pointed out that wave and tidal energy can be considered the most mature technologies. They present an estimated potential of 29500 TWh/year and 1200 TWh/year, respectively.

To illustrate the potential of wave energy resource, let us consider the wind resource for comparison. The wind speed profile extends many km upwards. The higher, the more energy can be obtained. However, a wind turbine has a limited height and, therefore, just a minimal part of the available wind energy is captured. On the other hand, the wave energy is more concentrated close to the sea surface and, therefore, a floating structure may absorb a large part of wave energy flow. Fig. 1.1 illustrates both situations.

Figure 1.1 Comparison between the wind and wave energy profile captation.

1.1 Wave Energy

Waves are originated by winds that blow over the sea surface. Usually, waves are created in storm zones in the middle of the ocean. The wind transfers part of this energy to waves that travel over long distances and carry this amount of power. The wave energy is available at any part of the sea with more or less potential. Its exploitation has some advantages such as a high predictability, little environmental interference, diversity in terms of the technology used and closeness between resource and demand. On the other hand, some drawbacks are remarkable: devices are subject to severe weather conditions; high cost of installation and maintenance; and most of all technologies are in a development stage.

Many studies on the wave energy resource have been carried out throughout the world (Rusu and Onea, 2017). Just to cite some cases: the Atlantic Marine Energy Test Site (AMETS) on the west coast of Ireland (Atan, Goggins, and Nash, 2016), the Sicilian coastline (Franzitta, Catrini, and Curto, 2017), Northern Spain (Bento, Martinho, and Soares, 2015), the Maltese coasts (Franzitta, Curto, and Rao, 2016), the Australian south-east shelf (Morim et al., 2016), the East China Sea (Wu, Liu, and Chen, 2015), the Indian seas (Kumar and Anoop, 2015), Brazil (Contestabile, Ferrante, and Vicinanza,

2015), Peru (López, Veigas, and Iglesias, 2015) and Uruguay (Alonso, Solari, and Teixeira, 2015). Therefore, due to the great energy potential of waves, the interest in this topic has been increasing worldwide, and the main challenge is to make it competitive against more affordable generation resources. Unlike wind energy, for example, there are different technologies to convert the wave energy into electricity. These devices, wave energy converters (WEC), are in different development phases, and some of them are even connected to the grid in test sites (*EMEC - The European Marine Energy Centre*; López et al., 2013; *Marine and Hydrokinetic Technology Database*).

WECs are usually classified according to their working principle, location and orientation; and the grouping of these devices can be found in (Lehmann et al., 2017; López, Veigas, and Iglesias, 2015) and the references therein. On the working principle, the WEC technologies are divided mainly into: oscillating water column (OWC), submerged pressure differential, overtopping devices, floating structures and impact devices; in terms of location, they can be classified as onshore, nearshore and offshore devices; and finally, according to their orientation: attenuator, point absorber and terminator. Another common classification derived from the working principles is presented in (Falcão, 2010), where the devices are divided into three great groups: oscillating water column, oscillating bodies and overtopping, with other subclassifications.

1.2 Motivation and Objectives

An important issue to be considered for the feasibility of harnessing wave energy is the control strategy upon the WEC, on which both design and operation are highly dependent. From the operation point of view, many studies are addressed to optimise the energy absorbed by a WEC. In this sense, this thesis intends to improve the WEC efficiency by proposing a novel control strategy. Therefore, the first step is to develop a WEC model that interacts with the waves.

It is also relevant to study aspects related to the interaction between the WEC (point absorbers) and the electric power system. Unlike conventional electrical generation (fossil fuels, nuclear, hydraulic), wave energy is an uncontrollable energy source, such as wind power. However, in comparison with the latter, the wave energy generation has less short-term variability and better predictability. Therefore, the potential for integrating generation from wave energy sources into the electrical system could be high. Nevertheless, the penetration level will be limited by its impacts on the said system and by the technologies available to mitigate them.

On the other hand, for a possible wave energy exploitation, tens or hundreds of WECs (in the case of point absorbers) must be gathered forming a wave farm. The sea surface estimation for each WEC is required for both single WEC control and centralised control. Therefore, a prediction model should be proposed to provide this estimation.

1.2.1 Specific Objectives

More specific objectives defining the scope of the thesis are detailed below:

- To develop a WEC model to reproduce its dynamic behaviour with an adequate degree of detail to conduct control studies.
- To formulate a control strategy to maximise the electrical power generated by the WEC.
- To develop a wave-to-wire model to be connected to an electrical grid.
- To analyse possible impacts on the electrical grid and to formulate some mitigation strategies.
- To develop an estimator for the spatial propagation of wave within a wave farm.

1.3 Research Outcomes

During the period of development of this thesis, the following works, presented in chronological order, have been published in international conferences and JCR index journals:

- S. Martinez, J. A. Sanchez, H. Mendonça, R. Alvaro, A. Cuesta, and A. Diez. Impact of a wave energy farm connected to a distribution grid in voltage power quality. In *Proceedings of the CIGRÉ Lisbon Symposium*, 2013. p. 1–8. (Martinez et al., 2013).
- H. Mendonça and S. Martinez. Modeling of a wave energy converter connected to a resistive load. In *4th International Youth Conference on Energy (IYCE)*. IEEE, 2013. p. 1–6. (Mendonça and Martinez, 2013).
- H. Mendonça, S. Martinez, and D. Ramirez. Comparison of current control strategies applied to a boost-rectifier connected to a direct drive wave energy converter. In *International Conference on Renewable Energy Research and Applications (ICRERA)*. IEEE, 2015. p. 1259–1264. (Mendonça, Martinez, and Ramirez, 2015).
- H. Mendonça and S. Martinez. A resistance emulation approach to optimize the wave energy harvesting for a direct drive point absorber. *IEEE Transactions on Sustainable Energy*, 2016, vol. 7, no 1, p. 3–11. (Mendonça and Martinez, 2016a).
- H. Mendonça and S. Martinez. Energy Conversion Efficiency Assessment of a Direct Drive Wave Energy Converter with Different Current Controllers. In *IEEE International Energy Conference (ENERGYCON)*. IEEE, 2016. p. 1–6. (Mendonça and Martinez, 2016b).

- A. Medina, H. Mendonça, and S. Martinez. Study of Voltage Fluctuations Caused in a Distribution Grid by the Connection of a Wave Energy Converter and Corrective Actions Based on Reactive Power Compensation. *Procedia Computer Science*. 2016, vol. 83, p. 832–838. (Medina, Mendonça, and Martinez, 2016)¹.
- H. Mendonça, R. M. de Castro, D. Montalbán, and S. Martínez. Voltage impact of a wave energy converter on an unbalanced distribution Grid and corrective actions. *Sustainability*. 2017, vol. 9, no 10, p. 1844. (Mendonça et al., 2017).
- D. Ramirez, H. Mendonça, M. Blanco, and F. Martinez. Non-linear vector current source for the control of permanent magnet synchronous generators in wave energy applications. *IET Renewable Power Generation*, 2019, vol. 13, no 13, p. 2409-2417. (Ramirez et al., 2019).

In addition, part of the development of this thesis has resulted in another JCR publication outside the scope of the presented work:

- F. Martinez-Rodrigo, D. Ramirez, H. Mendonça, and S. de Pablo. MMC as non-linear vector current source for grid connection of wave energy generation. *International Journal of Electrical Power & Energy Systems*. 2019, vol. 113, p. 686–698. (Martinez-Rodrigo et al., 2019).

1.4 Layout of the Thesis

This thesis is organised in five chapters. The first chapter provides general information and a brief introduction about the panorama of wave energy in the world. Besides, the motivation and objectives sought by this research are detailed. Finally, this chapter also lists publications related to this thesis. The literature review is presented throughout the chapter in which the specific topic is discussed.

Chapter 2 offers an essential background of hydrodynamic theory, which includes environmental disturbances and the dynamics of a floating structure. On the ocean waves, this chapter presents their model based on the linear theory and their statistical representation. On the other hand, this chapter deals with the interaction between the fluid and the floating structure.

Control strategies applied to WECs are presented in Chapter 3. In addition, the developed strategy to maximise the generated power by using a current controller is presented, along with its performance with a specific model of the WEC. The implementation of the proposed strategy is carried out with three different types of controllers. All of them are assessed with respect to dynamic response and energy efficiency.

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¹Best Paper Award for *The 6th International Conference on Sustainable Energy Information Technology*.

The full developed model with the proposed control strategy is extended to a wave-to-wire model, where the whole conversion chain from wave energy to electricity is modelled in detail in Chapter 4. At this point, the WEC is connected to an electrical grid. This connection may lead to voltage fluctuation. In this sense, this disturbance to the electrical grid is also assessed in this chapter. In addition, a methodology is proposed to mitigate this effect.

In Chapter 5, a spatial wave estimation is presented. By using an artificial neuron network approach and up-wave measurements, this thesis proposes an estimation of the sea surface elevation at any point of a wave farm. The proposed methodology is assessed for different scenarios of wave climate.

The final chapter summarises the most important conclusions of this thesis and its contribution to the state of the art in the wave energy field. Besides, some suggestions for future works are presented based on this thesis achievements.

Dynamics of a Point Absorber Converter

In this chapter, the interaction between a floating structure and the waves is discussed. First of all, a mathematical description of the movement of the sea surface is required to study that interaction. So, this chapter describes a wave model used to generate a time series for the elevation of the sea surface in a given area of the sea. The propagation of waves on the sea surface can be represented by a linear model based on the Airy's wave theory (McCormick, 2009). This theory is developed from a monochromatic, or regular, wave that is a simple harmonic wave traveling in one direction. It can be extended to the case of irregular waves, that, in turn, can be represented by a combination or superposition of a large number of components of regular sine waves with different wavelengths, frequencies, amplitudes, and directions. The waves are usually classified based on the wave period as shown in Fig. 2.1 and Table 2.1.

Figure 2.1 Classification of ocean waves based on the wave period (Munk, 1950).

On the other hand, when exciting by incoming waves, a free floating structure on the sea moves with six Degree of Freedom (DoF), where three of them are translational and

Table 2.1 Classification of ocean waves (Pinet, 2011).

Wave	Period	Wavelength	Wave Type	Cause
Capillary wave	< 0.1 sec	< 2 cm	Deep to shallow	Local winds
Chop	1–10 sec	1–10 m	Deep to shallow	Local winds
Swell	10–30 sec	Up to hundreds of m	Deep to shallow	Distant storm
Seiche	10 min–10 hr	Up to hundreds of km	Shallow or intermediate	Wind, tsunami, tidal resonance
Tsunami	10–60 min	Up to hundreds of km	Shallow or intermediate	Submarine disturbance
Tide	12.4–24.8 hr	Thousands of km	Shallow	Gravitational attraction of sun and moon
Internal wave	min to hr	Up to hundreds of m	Deep to shallow	Disturbance at pycnocline

the others are rotational. Both linear and rotational motions are illustrated in Fig. 2.2.

The motion of a floating structure in waves is studied by the seakeeping theory. As stated by Perez (2006), if the floating structure is not exposed to extreme conditions, its motion can be considered a linear system. In this sense, the linear seakeeping theory assumes that the sea surface elevation is a process completely described by its power spectral density (wave spectrum); the wave excitation forces and the motion of a floating structure are linear; and the motion is in steady state, i.e., there is no forward speed or it moves at constant speed. This system can be seen in Fig. 2.3a, where the linear forces are composed by originating forces of the incoming waves and the floating structure itself. The assumption of linearity is indeed a limitation, since the nonlinearities are not taken into account. However, the linear assumption could be seen as the first step to further studies, in addition to making a basis upon which nonlinearities can be incorporated separately. Fig. 2.3b shows the complete system of a floating structure whose only difference compared to the linear system is the incorporation of nonlinear forces by force superposition.

Examples of nonlinear forces are the viscous damping force, mooring force and, particularly in the case of a WEC, the PTO force. According to Ringwood, Bacelli, and Fusco (2014), the viscous damping force becomes significant with increasing structure speed and it is particularly relevant if the body surface contains irregularities, such as flanges. Also,

Figure 2.2 Linear and rotational motions of a free floating structure.

Figure 2.3 General view of a floating structure system: (a) linear and (b) nonlinear systems.

Perez (2006) says that this nonlinearity cannot be neglected for the roll motion. In the case of a point absorber converter, however, much of the conducted studies found in the specialised literature do not include the viscous damping force. In addition to this case, the mooring force can be associated with the coupling mechanism between buoy and PTO system.

In this work, the WEC studied is a point absorber converter that, in most cases, is composed of a buoy connected to a PTO system on the seabed, as the simplified model schematically represented in Fig. 2.4. From this point, this work will refer to the floating structure as a buoy with 1-DoF in heave motion, i.e., the buoy has only the vertical motion. In addition, the viscosity is neglected what means that the buoy is in an ideal fluid and, for a while, buoy and PTO are assumed to be rigidly connected.

Figure 2.4 Simplified schematic representation of a point absorber converter.

2.1 Regular Waves

Despite the fact that regular waves are not found in a real sea, it is important to study them, since a more realistic wave propagation can be described with the superposition of several regular waves. In addition, the first steps in the design of a control strategy for the WEC are carried out in very controlled sea conditions, as those resulting from this

elementary waveform. After applying the Airy theory (known as deep water wave), the elevation of the sea surface related to the still water level ($z = 0$) is given by:

$$\zeta(x, t) = \zeta_a \cdot \sin(2\pi ft - kx + \varepsilon) \quad (2.1)$$

where ζ_a , f , k , and ε are a constant amplitude, frequency, wave number and phase angle, and x is the position of the sea surface elevation. The wave height is defined as $H = 2\zeta_a$ and the wave period $T = 1/f$. The wave-energy transport of regular waves is defined by (Falnes, 2007):

$$J = \frac{\rho g^2}{32\pi} T H^2 \quad (2.2)$$

given in [W/m], where ρ and g are fluid density and acceleration due to gravity.

2.2 Irregular Waves

Because of the non-stationary and non-deterministic nature of ocean waves, it is impossible to determine the elevation of the sea surface at a pre-determined point in time. However, the amount of energy can be represented statistically by the wave spectrum. More formally, the variance density spectrum, that is usually related to the wave frequency, f , describes the sea surface elevation. The general condition of the sea surface is known as sea state which can be represented by two parameters: significant wave height, H_s , and energy period, T_e . Both parameters can be calculated directly from the wave spectrum and they are also known as statistical wave parameters. In this case, the wave-energy transport is (Falnes, 2007):

$$J = \frac{\rho g^2}{64\pi} T_e H_s^2 \quad (2.3)$$

One of the most widely used wave spectra is the Bretschneider spectrum. A modified version with two parameters, ITTC spectrum, which can be found in (The Specialist Committee on Waves (chaired by C. T. Stansberg), 2002), is used in this work and can be described in terms of H_s and T_e as follows:

$$S(f) = \frac{A}{f^5} \exp(-B/f^4) \quad (2.4)$$

with:

$$A = \frac{0.0081}{K^4} g^2; \quad B = \frac{0.0081}{K^4} \frac{4g^2}{H_s^2}; \quad K = \frac{T_e}{2.137} \sqrt{\frac{g}{H_s}} \quad (2.5)$$

A realisation of the sea surface elevation in a wave farm can be generated by the superposition of a large number of regular waves, each with an amplitude taken from the

given spectrum and a random phase angle:

$$\zeta(x, y, t) = \sum_{n=1}^N \sqrt{2S(f_n)\Delta f} \cdot \sin[2\pi f_n t - k_n(x \cos \chi - y \sin \chi) + \varepsilon_n] \quad (2.6)$$

where f_n , k_n , and ε_n are frequency, wave number and phase angle for each harmonic n , Δf is band frequency, x and y the coordinates at the wave farm, and χ the wave front angle. This model is known as long-crested sea, where regular swell waves travel in the same direction and they have longer periods and lengths than in chaotic seas within the fetch area, where waves are originated (Fig. 2.5). In the dispersion area, waves become more regular, changing to swell waves whose periods range from 5 to 30 seconds and they are usually limited by shorelines. In addition, swell waves carry the most part of the energy generated in the fetch area.

Figure 2.5 Irregular seas in the fetch and dispersion areas (based on Government of Canada, Environment Canada, 2014).

On the other hand, more than one direction of wave propagation might appear in the sea resulting in a short-crested sea. Despite this type of sea is more common in fetch areas, it is possible to find swell waves travelling with different directions whose interactions result in short-crested seas outside the fetch area, with waves characterised by shorter lengths and periods. Therefore, seas with different periods, heights and lengths propagating in different directions appear (Fig. 2.5). This sea can be modelled from a composition of wave and directional spectra: $S(f, \theta) = S(f)G(\theta)$, where the spreading function G depends on the direction of the wave propagation θ . The Specialist Committee on Waves (chaired by C. T. Stansberg) (2002) defines as the most common spreading function:

$$G(\theta) = \frac{2^{2s-1}}{\pi} \frac{\Gamma^2(s+1)}{\Gamma(2s+1)} \cos^{2s} \frac{1}{2}(\theta - \theta_1) \quad (2.7)$$

where θ_1 is the main direction of wave propagation and s is the width of the directional

spread. The realization of the sea is similar to Eq. (2.6), but now it takes into account the angles of wave propagation:

$$\zeta(x, y, t) = \sum_{n=1}^N \sum_{m=1}^M \sqrt{2S(f_n, \theta_m) \Delta f \Delta \theta} \times \sin[2\pi f_n t - k_n(x \cos \chi_m - y \sin \chi_m) + \varepsilon_n] \quad (2.8)$$

where m is the number of directions of wave propagation.

In order to illustrate a real wave spectrum, let us consider the directional spectrum shown in Fig. 2.6, corresponding to observations from buoy 43 of The Coastal Data Information Program (*CDIP Historic Data*) located in the West coast of the United States (GPS coordinates: $33.2198^\circ, -117.4394^\circ$). The directional spectrum was calculated from wave measurements on 12-11-2016 from 11:02 to 11:32 and it is free available in The Coastal Data Information Program (*CDIP Historic Data*). It is possible to identify the direction to which waves are travelling and that the highest energy is spread in a narrow band around 240° . Thus, the most part of the energy is carried by a predominant wave in this direction of propagation.

Figure 2.6 Directional wave spectrum of CDIP buoy 43 (data from CDIP, Scripps Institution of Oceanography).

2.3 Equation of Motion

Assume the WEC in Fig. 2.4 is in equilibrium, where part of the buoy is immersed in water. The buoy position with respect to the Still Water Level (SWL) is $z = 0$, the wet surface of the buoy is S_b and $\mathbf{n}$ is the unity normal vector to S_b at the water with components $(n_x, n_y, n_z) \equiv \mathbf{n}$. If the buoy has a translational motion, $\mathbf{u}$ is its velocity

vector, where $u_n = \mathbf{n} \cdot \mathbf{u}$. Due to the linearity assumption, the hydrodynamic analysis of the buoy can be divided into two systems and both can be analysed separately. In the first one, the buoy is fixed at its equilibrium position, completely restrained without any degree of freedom, and subjected to an incident wave. In the second system, the buoy is forced to oscillate in still water.

When a fixed buoy is subjected to an incident wave, two forces act on the wet surface of the buoy. The composition of these forces is known as the linearised wave excitation or 1st order wave excitation force. The linearity allows that results for sinusoidal excitation can be extended to solve non-sinusoidal cases. This incident wave induces an unsteady pressure on the wet surface that, in turn, particularly in this case, corresponds to a vertical force component, the excitation force f_e . This force on the infinitesimal surface dS_b is $df_e = -p_e n_z dS_b$, where p_e is the hydrodynamic pressure due to the incident wave and n_z is the vertical component of $\mathbf{n}$. So, integrating the excitation force over the wet surface results in the total excitation force:

$$f_e = - \iint_{S_b} p_e n_z dS_b \quad (2.9)$$

This excitation force can be decomposed into two components: the Froude-Krilov force and the diffraction force. The Froude-Krilov force results of the undisturbed pressure field, i.e., the buoy does not disturb the wave field, with a velocity potential ϕ_i , the same as the undisturbed incident wave. On the other hand, the change in the wave field generates the diffraction force due to the diffracted wave with velocity potential ϕ_d . To calculate the excitation force, in terms of the pressure on the wet surface,

$$p_e = -\rho \frac{\partial(\phi_i + \phi_d)}{\partial t} \quad (2.10)$$

a boundary problem must be solved. In this case, both potentials have to satisfy the boundary conditions on the seabed (there is no flow velocity component normal to the sea bottom) and on the free surface (water pressure equals the atmospheric pressure on the free surface) (Perez, 2006). As there is no fluid flow through the limits of the wet surface, the potential is given by:

$$\frac{\partial(\phi_i + \phi_d)}{\partial n} = u_n \quad (2.11)$$

However, as the buoy is fixed, the normal component of the total velocity of the wet surface must be zero ($u_n = 0$). As a result, the diffraction problem also has to satisfy the following boundary condition:

$$-\frac{\partial \phi_d}{\partial n} = \frac{\partial \phi_i}{\partial n} \quad \text{on } S_b \quad (2.12)$$

Falnes (2002a) highlights that the diffraction term ϕ_d could be neglected, in particular, if the extension of the immersed part of the buoy is very small compared with the wavelength. In this case, only the Froude-Krylov force is considered. So, there is no need to solve the boundary value problem to find the diffraction potential ϕ_d .

Now, let the buoy oscillate vertically about its equilibrium position in the absence of an incident wave. As a result of this motion, a radiated wave with velocity potential ϕ_r is produced. This radiated wave reacts with a force on the buoy known as radiation force f_r . Similarly to the wave excitation force, the vertical component of the radiation force is:

$$f_r = - \iint_{S_b} p_r n_z \, dS_b \quad (2.13)$$

where

$$p_r = -\rho \frac{\partial \phi_r}{\partial t} \quad (2.14)$$

is the change of pressure on the buoy due to the outgoing waves. As in the case of the excitation force, a boundary value problem for the velocity potential ϕ_r must be solved to find the radiation force. So, the same boundary conditions of the free surface and the seabed must be satisfied. In this case, the buoy has only vertical component of velocity: $u_n = (dz/dt)n_z$. So, for a displacement z small compared to the wavelength, the following boundary condition must also be satisfied:

$$\frac{\partial \phi_r}{\partial n} = \frac{dz}{dt} n_z \quad \text{on } S_b \quad (2.15)$$

Finally, the last force that acts on the buoy under the linear assumption is the buoyancy force f_b that is a restoring force associated with gravity and buoyancy and it tends to restore the equilibrium. Considering cylindrical a buoy in its equilibrium position and in the absence of waves, by the Archimedes' law, the buoy weight is given by $\rho g D \pi r^2$, where D is draft of the buoy, i.e., the height of the immersed part of the buoy, and r the buoy's radius. Moving the buoy from its equilibrium point to a position z , the upward buoyancy force $\rho g (D - z) \pi r^2$ combined with the buoy weight yields the buoyancy force:

$$f_b = -\rho g S_w \cdot z \quad (2.16)$$

where S_w is the waterplane area, in other words, the cross section of the buoy defined by the water plane at its equilibrium point. So, the buoyancy coefficient is defined as $K_w = \rho g S_w$.

Once described all the linear forces on the buoy, its motion can be defined applying the Newton's second law:

$$m \frac{d^2 z}{dt^2} = f_b + f_r + f_e + f_{pto} \quad (2.17)$$

where f_{pto} is the vertical force applied by the PTO system upon the buoy. Note that if $f_{\text{pto}} = 0$, Eq. (2.17) represents the dynamics of a free floating structure. On the other hand, if the PTO force is considered linear, according to Falcão (2010), it can be given by:

$$f_{\text{pto}} = -B_{\text{pto}} \frac{dz}{dt} - K_{\text{pto}} z \quad (2.18)$$

where f_{pto} consists of a linear damping term proportional to velocity through coefficient B_{pto} and a stiffness term proportional to displacement through coefficient K_{pto} .

2.4 The Frequency-Domain Model

To proceed with the frequency domain analysis, all the linear assumptions in the previous section are considered. The buoy is excited by a simple harmonic (monochromatic) wave with unit amplitude whose representation in the frequency domain is $\zeta(j\omega)$ with angular frequency ω . As the system is under sinusoidal oscillations, the velocity potential of the fluid flow field surrounding the wet surface can be written in complex terms as:

$$\phi(t) = \text{Re} \left\{ \hat{\phi} e^{j\omega t} \right\} \quad (2.19)$$

where $\hat{\phi} = \phi_0 \cos(\omega t + \varphi)$ is the complex amplitude of the velocity potential. Note that the physical quantity is the real part of the complex amplitude, therefore, from now on, the term $\text{Re} \{ \cdot \}$ will be omitted. So, the velocity potential ϕ can be written as a composition of the potential due to incident, diffraction and radiation waves:

$$\hat{\phi} e^{j\omega t} = \left[(\hat{\phi}_i + \hat{\phi}_d) + \hat{\phi}_r z \right] e^{j\omega t} . \quad (2.20)$$

Thus, the pressure due to the motion of the buoy is given by:

$$P = -\rho \frac{\partial \phi(t)}{\partial t} = -j\omega \rho \hat{\phi} e^{j\omega t} \quad (2.21)$$

where the pressure P can be decomposed into:

$$P_e = -j\omega \rho (\hat{\phi}_i + \hat{\phi}_d) e^{j\omega t} \quad (2.22a)$$

$$P_r = -j\omega \rho \hat{\phi}_r z e^{j\omega t} \quad (2.22b)$$

where P_e is the pressure due to the incident and diffraction waves and P_r due to the radiation wave.

So, the vertical force per unit amplitude due to undisturbed waves can be calculated

by substituting Eq. (2.22a) in Eq. (2.9) that yields:

$$\begin{aligned}\hat{F}_{ei}e^{j\omega t} &= \left(-j\omega\rho \iint_{S_b} \hat{\phi}_i n_z dS_b \right) e^{j\omega t} \\ \hat{F}_{ei} &= -j\omega\rho \iint_{S_b} \hat{\phi}_i n_z dS_b .\end{aligned}\quad (2.23)$$

Similarly, the vertical force due to diffracted waves is given by:

$$\hat{F}_{ed} = -j\omega\rho \iint_{S_b} \hat{\phi}_d n_z dS_b . \quad (2.24)$$

As a result, under the frequency domain analysis, the composition of these forces results in a frequency-dependent term known as excitation wave coefficient or wave-to-force Response Amplitude Operator (RAO) $H_e(j\omega) = \hat{F}_{ei} + \hat{F}_{ed}$ given in N/m for a given oscillation at ω rad/s. So, the wave-to-force RAO can be also represented in complex terms as $H_e(j\omega) = H_e(\omega)e^{j\varphi_e(\omega)}$, where $H_e(\omega)$ and $\varphi_e(\omega)$ are the magnitude and phase of the wave-to-force RAO, respectively. Finally, the excitation force is given by:

$$F_e(j\omega) = H_e(j\omega) \cdot \zeta(j\omega) \quad (2.25)$$

On the other hand, the vertical component of the radiation force originated by the vertical oscillating motion with unit amplitude of the buoy is:

$$\frac{F_r(j\omega)}{Z(j\omega)} = -j\omega\rho \iint_{S_b} \hat{\phi}_r n_z dS_b \quad (2.26)$$

Expressing the radiation wave potential $\hat{\phi}_r$ in real and imaginary parts and substituting into Eq. (2.26) results in:

$$\begin{aligned}\frac{F_r(j\omega)}{Z(j\omega)} &= -j\omega\rho \iint_{S_b} \left[\text{Re}(\hat{\phi}_r) + j\text{Im}(\hat{\phi}_r) \right] n_z dS_b \\ &= \omega\rho \iint_{S_b} \text{Im}(\hat{\phi}_r) n_z dS_b - j\omega\rho \iint_{S_b} \text{Re}(\hat{\phi}_r) n_z dS_b \\ \frac{F_r(j\omega)}{Z(j\omega)} &= \omega^2 A(\omega) - j\omega B(\omega)\end{aligned}\quad (2.27)$$

with

$$A(\omega) = \frac{\rho}{\omega} \iint_{S_b} \text{Im}(\hat{\phi}_r) n_z \, dS_b \quad (2.28a)$$

$$B(\omega) = \rho \iint_{S_b} \text{Re}(\hat{\phi}_r) n_z \, dS_b \quad (2.28b)$$

where the coefficients $A(\omega)$ and $B(\omega)$ are known as added mass and potential damping, respectively.

This group of frequency-dependent coefficients (wave-to-force RAO, added mass and potential damping), also called hydrodynamic parameters, exclusively depend on the buoy geometry. Traditionally, these parameters have been determined experimentally in towing tanks. On the other hand, they can be also found mathematically; however, as commented in the previous section, finding these hydrodynamic parameters requires to solve a boundary-value problem for the velocity potential, what it is not possible in terms of elementary functions. Thereupon, numerical approaches have been developed to evaluate this problem and, with advances in computer-aided engineering, it is becoming increasingly more common the migration of experimental approaches to numerical ones. For example, the strip theory (2-D potential theory), presented by Faltinsen (1993), is especially useful for slender bodies. The wet part is divided into a finite number of strips and each cross section is treated as 2-D. So, the hydrodynamic parameters are calculated for each strip and, then, integrated over the length of the wet surface (Journée, 1992). Another common way, it is the panel method (3-D potential theory). In this method, the wet surface is approximated by splitting it into very thin quadrilateral panels so that a 3-D problem is converted into a 2-D one, where each panel is associated with sources and dipoles distributed on the wet surface oriented normal to the surface. Then, discretised equations must be solved for the unknown variables on each panel. Newman (1992) provides more details about this method and a survey of the mathematical basis can be found in references therein. According to Erickson (1990), the great advantage of this method is that there is no need to define a grid over the wet surface and it could be applied to more complex geometries. There are few commercial software tools that use this latter approach, such as WAMIT (WAMIT, Inc., 2016) and Aqwa (AQWA/Ansys, 2013).

To illustrate the aforementioned hydrodynamic parameters, a cylindrical buoy, named as buoy-1, is considered with the following characteristics:

- Diameter: 3 m
- Height: 1 m
- Mass: 3640.32 kg
- Draft at still water: 0.5 m

So, the heave hydrodynamic parameters calculated by Aqwa are shown in Fig. 2.7. Note that all of these parameters are given in the frequency-domain.

Figure 2.7 Frequency-dependent hydrodynamic parameters obtained by Aqwa/ANSYS.

Finally, both buoyancy and PTO forces can be easily analysed in the frequency domain by taking the Fourier transform of Eqs. (2.16) and (2.18), respectively, as follows:

$$F_b(j\omega) = -K_w Z(j\omega) \quad (2.29)$$

$$F_{pto}(j\omega) = -j\omega B_{pto} Z(j\omega) - K_{pto} Z(j\omega) \quad (2.30)$$

2.5 Regular Waves Response

From this point, as the system is constrained by heave motion, the vertical displacement of the buoy is represented by the variable y instead of z . Following the approach presented by Falcão (2007) and assuming that the system is completely linear, an incident regular wave with angular frequency ω , i.e., $\zeta = \zeta_a e^{j\omega t}$, generates single-harmonic forces and motion with the same frequency ω . Therefore, applying the Fourier transform to Eq. (2.17), the heave motion can be described in the frequency domain by:

$$-\omega^2 m Y(j\omega) = F_b(j\omega) + F_r(j\omega) + F_e(j\omega) + F_{pto}(j\omega) \quad (2.31)$$

So, substituting Eqs. (2.27), (2.29) and (2.30) into Eq. (2.31) results in:

$$Y(j\omega) = \frac{F_e(j\omega)}{-\omega^2 [m + A(\omega)] + j\omega [B_{pto} + B(\omega)] + K_{pto} + K_w} \quad (2.32)$$

the buoy's motion response to a regular wave excitation. When $F_{pto}(j\omega) = 0$, the response in Eq. (2.32) is also known as response amplitude operator (RAO) which is the response of a free floating structure to a regular wave excitation. For the buoy with hydrodynamic parameters shown in Fig. 2.7, the resulting RAO is presented in Fig. 2.8.

Figure 2.8 Response amplitude operator (RAO).

Note that Eq. (2.32) is similar to the equation of motion for a damped spring-mass system in frequency domain. Therefore, it is very common, for regular waves, to make the analogy between the buoy motion and the spring-mass system.

2.5.1 Electrical Analogue Model

As commented by Thomas (1979), a mechanical system can be represented by an electrical analogue system. The RLC series circuit in Fig. 2.9a is considered to illustrate this electro-mechanical analogue. The governing differential equation is obtained by applying the Kirchhoff's voltage law (KVL) resulting in:

$$\begin{aligned} e(t) &= v_R(t) + v_L(t) + v_C(t) \\ e(t) &= Ri(t) + L \frac{d}{dt}i(t) + \frac{1}{C} \int_{-\infty}^t i(\tau) d\tau \end{aligned} \quad (2.33)$$

where $e(t)$ is the excitation voltage, $i(t)$ the current through the circuit, R the resistance, L the inductance and C the capacitance.

On the other hand, if a force $f(t)$ is applied to the mass-spring system in Fig. 2.9b in order to move the mass M from its equilibrium position, its motion equation is given by

Figure 2.9 Duality between electrical and mechanical systems. (a) RLC series circuit. (b) Mass-spring system.

the Newton's second law as follows:

$$f(t) = M \frac{d^2}{dt^2} y(t) + B \frac{d}{dt} y(t) + K y(t) \quad (2.34)$$

where $y(t)$ is the mass displacement, B the viscous damping coefficient and K the spring stiffness. As $\dot{y}(t) = \frac{d}{dt} y(t)$ and substituting it in Eq. (2.34), the motion equation of the mass-spring system results in:

$$f(t) = M \frac{d}{dt} \dot{y}(t) + B \dot{y}(t) + K \int_{-\infty}^t \dot{y}(\tau) d\tau. \quad (2.35)$$

Both Eqs. (2.33) and (2.35) are analogue. Therefore, by comparison, the duality is found with the following corresponding quantities:

$$f(t) \equiv e(t) \quad (2.36a)$$

$$\dot{y}(t) \equiv i(t) \quad (2.36b)$$

$$M \equiv L \quad (2.36c)$$

$$R \equiv B \quad (2.36d)$$

$$K \equiv \frac{1}{C}. \quad (2.36e)$$

2.5.2 Absorbed Power

The regular wave response of the buoy can be described from Eq. (2.31) in the time domain as:

$$[m + A(\omega)] \ddot{y}(t) + [B_{pto} + B(\omega)] \dot{y}(t) + (K_{pto} + K_w) y(t) = f_e(t) \quad (2.37)$$

Thus, as shown in Eqs. (2.34) and (2.37), the governing equation of the buoy is similar to that of an oscillating mass on a spring as in the case of Fig. 2.9b. So, the coefficients of the damped mass-spring equivalent is: $M = m + A(\omega)$, $B = B_{pto} + B(\omega)$ and $K = K_{pto} + K_w$.

Figure 2.10 WEC electrical analogue model.

In turn, in Section 2.5.1, the duality between an electrical and a mechanical systems is demonstrated. Hence, the WEC model in Fig. 2.4 with a linear PTO can be modelled as an electrical analogue system as shown in Fig. 2.10. In this system, the buoy is modeled as a hydrodynamic impedance, $\underline{Z}_w$, which is constant and dependent on the buoy's geometry. The PTO is modeled as a PTO impedance, $\underline{Z}_{pto}$, which could be adjusted. Both impedances are given by:

$$\underline{Z}_w = R_w + jX_w = B(\omega) + j \left\{ \omega [m + A(\omega)] - \frac{K_w}{\omega} \right\} \quad (2.38)$$

$$\underline{Z}_{pto} = R_{pto} + jX_{pto} = B_{pto} - j \frac{K_{pto}}{\omega} \quad (2.39)$$

Thus, analysing the WEC electrical analogue circuit in Fig. 2.10, the equivalent current phasor $\underline{\dot{Y}}$ is calculated by:

$$\begin{aligned} \underline{F}_e &= [R_w + R_{pto} + j(X_w + X_{pto})] \cdot \underline{\dot{Y}} \\ \underline{\dot{Y}} &= \frac{\underline{F}_e}{R_w + R_{pto} + j(X_w + X_{pto})} \end{aligned} \quad (2.40)$$

The active power absorbed by the PTO impedance is the average power absorbed by the WEC and it is given by $P_{abs} = R_{pto} \dot{Y}^2$, where $\dot{Y}$ is the rms value of the equivalent current $\dot{y}(t)$. Therefore:

$$P_{abs} = \frac{R_{pto} \cdot F_e^2}{(R_w + R_{pto})^2 + (X_w + X_{pto})^2} \quad (2.41)$$

where F_e is the equivalent rms voltage. If a maximum power transfer is desired, the absorbed power must be derived with respect to the resistance and reactance of the PTO that yields:

$$\frac{\partial P_{abs}}{\partial R_{pto}} = 0 \quad \Rightarrow \quad R_w = R_{pto} \quad (2.42a)$$

$$\frac{\partial P_{abs}}{\partial X_{pto}} = 0 \quad \Rightarrow \quad X_w = -X_{pto} \quad (2.42b)$$

thus:

$$B(\omega) = B_{\text{pto}} \quad (2.43a)$$

$$\omega [m + A(\omega)] - \frac{K_w}{\omega} = \frac{K_{\text{pto}}}{\omega} . \quad (2.43b)$$

As a conclusion, to absorb the maximum power, the PTO impedance must be adjusted to make the excitation force and velocity be in phase. Under these conditions, the system achieves the resonance condition, where the natural frequency of the WEC:

$$\omega_0 = \sqrt{\frac{K_w + K_{\text{pto}}}{m + A(\omega)}} \quad (2.44)$$

is coincident with the incident wave frequency ω . As a result, the maximum absorbed power (Eq. 2.41) yields:

$$P_{\text{max}} = \frac{F_e^2}{4B_{\text{pto}}} \quad (2.45)$$

It can be verified that the maximum absorbed power obtained in steady state by the dual electrical model is the same obtained through the hydrodynamic model, as shown in Falnes (2002a) and Falcão (2008), for example. Note that both references use the peak value of $f_e(t)$ in the complex form, hence the maximum absorbed power is $P_{\text{max}} = \frac{F_e^2}{8B_{\text{pto}}}$. Here, as an electrical analogue model is applied, Eq. (2.45) uses the rms value of $f_e(t)$ instead.

2.6 Capture Width Ratio

An efficient WEC is that one able to absorb as much energy as possible from an incident wave. So, a well-known measure used to assess the performance of a WEC is the capture width, CW , which is defined by the ratio of the power absorbed by the WEC (P_{abs} [kW]) to the wave energy transport (J [kW/m]) as follows (Babarit, 2015; Falcão, 2010; Falnes, 2002a; Price, Dent, and Wallace, 2009):

$$CW = \frac{P_{\text{abs}}}{J} \quad (2.46)$$

given in meters. In Falnes (2002a) and references therein it is demonstrated from Eq. (2.46) that the maximum theoretical absorbable power by a heaving axisymmetric body is:

$$CW_{\text{max}} = \frac{\lambda}{2\pi} \quad (2.47)$$

As pointed out by Falcão (2010), Eq. (2.47) is an important result to define an upper bound for the absorbed power as the Betz is limit for the power coefficient of a wind turbine (Manwell, McGowan, and Rogers, 2010). Babarit (2015) also highlights the importance of $CW_{\max}$ since it shows that the theoretical maximum for the capture width depends only on the wavelength and not on the device geometry.

The quotient between CW and the width d of the floating body results in the non-dimensional measure known as relative capture width:

$$CWR = \frac{CW}{d} = \frac{P_{\text{abs}}}{Jd} \quad (2.48)$$

In other words, it is the ratio between the absorbed power per width unit of the floating body to the wave-energy transport. This measure represents the hydrodynamic efficiency of the system and, from now on, it will be denoted by P_{ratio} .

2.6.1 Upper Bound

As seen in the previous section, the maximum theoretical absorbable power can be calculated with Eqs. (2.46) and (2.47):

$$P_{\max} = \frac{\lambda J}{2\pi} \quad (2.49)$$

Therefore, for regular waves on deep water, where $\lambda = \frac{g}{2\pi}T^2$ and $J = \frac{\rho g^2 T H^2}{32\pi}$, the upper limit for the absorbed power is given by:

$$P_A = \frac{\rho g^3}{128\pi^3} T^3 H^2 \quad (2.50)$$

So, the value of the absorbed power P_{abs} will be below the increasing curve shown in Fig. 2.11, where P_A is calculated for different wave periods and with fixed wave height.

Figure 2.11 Upper bounds of the absorbed power for regular waves.

On the other hand, Falnes (2002a) defines another upper bound: the power-to-volume

ratio, which it is also known as Budal's upper bound. For a regular wave with a given wave height and period, the Budal's upper bound is:

$$\frac{P_B}{V} = \frac{\pi \rho g H}{4T} \quad (2.51)$$

where V is the volume of buoy. This limit is also shown in Fig. 2.11. According to Falnes (2002b), the upper bound could be reached if the following conditions are satisfied:

- (i) The oscillating body is relative small ($V \rightarrow 0$).
- (ii) Most of the incident wave power remains in the ocean.
- (iii) The oscillation has an optimum phase, i.e., the resonance frequency matches the incident wave frequency.
- (iv) The oscillation amplitude equals the design amplitude.
- (v) The oscillation mode has a source-like wave radiation.

In conclusion, a practical device will have its absorbed power below the upper bounds in Fig. 2.11 (the inner grey area). For low wave periods, the absorbed power is limited by the increasing curve P_A , where the volume of the device is practically not used to extract the wave power. On the contrary, for large waves periods, the limit is the decreasing curve P_B , where the absorbed power depends on the immersed oscillating volume.

2.7 The Time-Domain Model

In the presence of irregular waves, the buoy motion is no longer in steady state. Therefore, the frequency-domain analysis (Eq. 2.32) is not valid anymore. On the other hand, even if excited by regular waves, the frequency-domain analysis still can not be used if there are nonlinear forces applied to the WEC, for instance, by a nonlinear PTO system. In such cases, the time-domain analysis must be carried out. Developed by Cummins (1962), for zero speed forward, this latter analysis resulted in an integro-differential equation known as *Cummins equation*, that can be applied to all the 6-DoF (or any of them separately). As only heaving motion is considered in this work, and taking into account the PTO system, the *Cummins equation* is described as:

$$[m + A(\infty)] \ddot{y}(t) + \int_{-\infty}^t h_r(t - \tau) \dot{y}(\tau) d\tau + K_w y(t) = f_e(t) + f_{pto}(t) \quad (2.52)$$

where $A(\infty)$ is the added mass at high frequencies and $h_r(t)$ is the impulse response function (IRF) of the radiation force. The integral term in Eq. (2.52) describes the contribution of the radiation force due to the radiated wave field induced by the vertical

motion of the buoy. This wave field depends on the history of the buoy motion, the reason why is also known as fluid memory effect. The radiation force IRF, $h_r(t)$, for this case, only depends on the buoy geometry. Both domain analysis are related by their parameters (discussed in Section 2.7.1) as established by Ogilvie (1964).

2.7.1 The Radiation Force

As shown by Cummins (1962), the contribution of the radiation force in Eq. (2.52) is given by:

$$f_r(t) = -A(\infty)\ddot{y}(t) - \int_{-\infty}^t h_r(t - \tau)\dot{y}(\tau) d\tau \quad (2.53)$$

i.e., a composition of the pressure forces due to buoy acceleration and the fluid memory model. Some time later, Ogilvie (1964) related the frequency- and time-domain models by applying the Fourier transform. As a result, the added mass and damping coefficient can be calculated by:

$$\begin{aligned} A(\omega) &= A(\infty) - \frac{1}{\omega} \int_0^\infty h_r(t) \sin(\omega t) dt \\ B(\omega) &= \int_0^\infty h_r(t) \cos(\omega t) dt \end{aligned} \quad (2.54)$$

where:

$$A(\infty) = \lim_{\omega \rightarrow \infty} A(\omega) \quad (2.55)$$

Thus, by applying the inverse Fourier transform, the radiation force IRF can be calculated either by the added mass or the damping coefficient as follows:

$$\begin{aligned} h_r(t) &= \frac{2}{\pi} \int_0^\infty B(\omega) \cos(\omega t) d\omega \\ h_r(t) &= -\frac{2}{\pi} \int_0^\infty \omega [A(\omega) - A(\infty)] \sin(\omega t) d\omega \end{aligned} \quad (2.56)$$

Also, from the Fourier transform, the radiation force IRF in the frequency domain can be found by:

$$\begin{aligned} H_r(j\omega) &= \int_0^\infty h_r(t) \cdot e^{-j\omega t} d\omega \\ &= B(\omega) + j\omega [A(\omega) - A(\infty)] \end{aligned} \quad (2.57)$$

Since the added mass and the damping coefficient can be obtained by hydrodynamic codes (as commented in Section 2.4), the issue arises in how to realise the radiation force IRF to represent the fluid memory model in Eq. (2.53). Jefferys (1984) and McCabe et

al. (2007), for example, have developed their model identification based on the frequency domain. On the other hand, Kristiansen, Hjulstad, and Egeland (2005) and Yu and Falnes (1995) work on the time-domain-based model identification. The identification problem is not studied in this thesis, since it is only a tool to realise IRF systems. So, the time-domain approach has been chosen due to its availability in MATLAB and straightforward implementation for both radiation and excitation forces. This latter one will be explained in Section 2.7.2.

The convolutional term in Eq. (2.53) can be approximated to a state-space model of order n as:

$$\begin{aligned}\dot{\mathbf{x}}_{\mathbf{r}}(t) &= \mathbf{A}_{\mathbf{r}} \cdot \mathbf{x}_{\mathbf{r}}(t) + \mathbf{B}_{\mathbf{r}} \cdot \dot{y}(t) \\ f_r'(t) &= \mathbf{C}_{\mathbf{r}} \cdot \mathbf{x}_{\mathbf{r}}(t) + \mathbf{D}_{\mathbf{r}} \cdot \dot{y}(t)\end{aligned}\tag{2.58}$$

where the state vector $\mathbf{x}_{\mathbf{r}}(t) \in \mathbb{R}^{n \times 1}$ and the matrices: $\mathbf{A}_{\mathbf{r}} \in \mathbb{R}^{n \times n}$, $\mathbf{B}_{\mathbf{r}} \in \mathbb{R}^{n \times 1}$, $\mathbf{C}_{\mathbf{r}} \in \mathbb{R}^{1 \times n}$ and $\mathbf{D}_{\mathbf{r}} \in \mathbb{R}$. Following the procedure presented by Kristiansen, Hjulstad, and Egeland (2005), the state-space model is calculated from $h_r(t)$ (Eq. 2.56) and realised by an approach based on the Hankel Singular Value Decomposition (SVD) method (Kung, 1978), which is implemented in MATLAB in the function `imp2ss` (Robust Control Toolbox). As a result, a high order model, whose order is proportional to the number of data points, is calculated. In addition, `imp2ss` is able to reduce indirectly the system order by setting the H_{∞} -norm¹ bounds. However, the desired order for the reduced system can be directly specified by using the function `balred` (also from the Robust Control Toolbox), that is also based on the truncated balanced reduction method².

The realised state-space system of the fluid memory model has a high order as stated above. Nonetheless, there are a great number of near pole-zero cancellations (Fig. 2.12a) which is a good indication that most of the states do not contribute significantly to the model. Thus, to define the order system, the Hankel SVD could be used to measure the contribution of each state to the model. For the buoy-1, Fig. 2.12b shows the state energies where, beyond the 5th state, the states could be clearly neglected.

The frequency response of both models, original and reduced (5th order), can be seen in Fig. 2.13. For periods up to 40 seconds, approximately, the responses both in magnitude and in phase are identical and, therefore, the realised reduced model assures the correct dynamics for the convolutional term for this period range. In addition, Fig. 2.14 compares the original and reduced model of the radiation force IRF for different orders. Even with a low order, as the case of order 2, the approximation is very precise. In the case of order 5, the accuracy is practically perfect and compared to the higher order model (order 10), there are no appreciable differences and it bears out the Hankel SVD analysis in

¹ H_{∞} -norm is the maximum over frequencies of the largest singular value of a stable MIMO system.

²More details about this method can be found in Kristiansen, Hjulstad, and Egeland (2005) and Varga (1991) and references therein.

Figure 2.12 Theoretical realised higher-order IRF of the radiation force (h_r) of the buoy described in Section 2.4. (a) Near pole-zero cancellations. (b) Hankel SVD.

Fig. 2.12b.

Figure 2.13 Comparison of the frequency response between the theoretical realised higher-order IRF of the radiation force of the buoy described in Section 2.4 and its reduced-order approximation system (5th order).

2.7.2 The Excitation Force

The excitation force for regular waves has a straightforward implementation from its representation in frequency-domain (Eq. 2.25), by applying the inverse Fourier transform, that yields:

$$f_e(t) = H_e(\omega) \cdot \zeta_m \cos[\omega t + \varphi_w + \varphi_e(\omega)] \quad (2.59)$$

Figure 2.14 Radiation force IRF (h_r). The black lines represent the real IRF and the grey circles, the realized IRF by a state-space model with different orders.

i.e., it is also expressed in terms of the excitation force coefficient. On the other hand, for irregular waves, the excitation force calculation is not so direct and it could be carried out in two different ways as detailed below.

Power Spectral Density of the Excitation Force

This first approach is presented by Perez (2006) and it is also applied by Falcão (2008) and Ricci et al. (2011), for example. In this case, the excitation force can be seen as a superposition of several sinusoidal forces due to the linearity assumption such as the wave surface elevation seen in Section 2.2. Thus, to directly generate the excitation force due to irregular waves, a similar approach to that used to mathematically generate a random wave field can be applied. That is, from the power spectral density of the excitation force which, in turn, is derived from the wave spectrum:

$$S_e(\omega) = |F_e(\omega)|^2 \cdot S_w(\omega, \chi) \quad (2.60)$$

Or, during the generation process of the random wave field, the excitation force is computed for each harmonic and realised as:

$$f_e(t) = \sum_k |F_e(\omega_k)| \zeta_{mk} \cos[\omega t + \varphi_{wk} + \varphi_e(\omega_k)] \quad (2.61)$$

The main drawback of this approach is that its use is restricted to synthetically generated wave fields. That is, it could not be applied to a recorded wave time series.

Excitation Force IRF

If a simulation with a recorded wave time series or a wave estimation is required, then the excitation force can be calculated through the IRF of the excitation force, whose procedure is very similar to the one presented Section 2.7.1. The excitation force in the frequency-domain (Eq. 2.25), $F_e(j\omega)$, is given as a function of the wave-to-force RAO,

$H_e(j\omega)$. Both, excitation force and excitation force IRF, can be calculated through the inverse Fourier transform:

$$h_e(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H_e(j\omega) e^{j\omega t} d\omega \quad (2.62)$$

$$f_e(t) = \int_{-\infty}^{\infty} h_e(\tau) \zeta(t - \tau) d\tau \quad (2.63)$$

Note that both functions are non-causal, since $h_e(t < 0) \neq 0$, which implies that current knowledge of the excitation force requires knowledge of the future incident wave data. Falnes (1995) studied the reason of these non-causality, despite the wave propagation is a causal system. In this case, the real cause of the input and output is usually a distant storm and the non-causality may be explained by the fact that, even before the incident wave hits the buoy, it already applies a force on it, as Falnes (1995) demonstrated. Therefore, modeling the excitation force is not straightforward as the memory function of the radiation force, since a non-causal system cannot be realised exactly (Falnes, 2002a).

Some approaches to identify a continuous-time system have been proposed in the specialised literature. McCabe et al. (2007) identify a continuous-time transfer function whose coefficients are calculated so that its frequency response best fits to that of the model. As the excitation force is a non-causal system, it is divided into two component forces: one stable and other unstable. The realisation of the stable component is straightforward by using a least-squares error method to obtain a fitted model from the data. To realise the unstable force component, a time scaling is applied resulting in a change of the stability characteristics. So, the excitation force would be the sum of the stable and time scaled unstable components. In this formulation, the excitation force is calculated apart from the WEC model, i.e., offline and, then, applied to the rest of the model.

Unlike the above method, where the excitation force is the system input, it is more reasonable to have the incident wave as the input, instead. For this reason, Yu and Falnes (1995) identify the state space model of the excitation force in the same way as the radiation force. Therefore, the excitation force model can be written as:

$$\begin{aligned} \dot{\mathbf{x}}_e(t) &= \mathbf{A}_e \cdot \mathbf{x}_e(t) + \mathbf{B}_e \cdot \zeta(t) \\ f_e(t) &= \mathbf{C}_e \cdot \mathbf{x}_e(t) + \mathbf{D}_e \cdot \zeta(t) \end{aligned} \quad (2.64)$$

where the state vector $\mathbf{x}_e(t) \in \mathbb{R}^{n \times 1}$ and the matrices: $\mathbf{A}_e \in \mathbb{R}^{n \times n}$, $\mathbf{B}_e \in \mathbb{R}^{n \times 1}$, $\mathbf{C}_e \in \mathbb{R}^{1 \times n}$ and $\mathbf{D}_e \in \mathbb{R}$. To calculate the matrices of the state space model, the same procedure described in Section 2.7.1 is applied.

However, to deal with the non-causality, a time shift is applied to the excitation force IRF to transform it into a “causalized” IRF. Assuming that there is a constant t_c so that

$h_e(t < t_c) = 0$, the “causalized” excitation force IRF is:

$$h_c(t) = h_e(t - t_c) \quad (2.65)$$

Hence, the “causalized” excitation force is:

$$f_{ec}(t) = \int_{-\infty}^{\infty} h_c(t - \tau) \zeta(\tau) d\tau \quad (2.66)$$

$$f_{ec}(t) = \int_{-\infty}^t h_e(t - t_c - \tau) \zeta(\tau) d\tau = f_e(t - t_c) \quad (2.67)$$

That is, the excitation force is time shifted with respect to the “causalized” excitation force:

$$f_e(t) = f_{ec}(t + t_c) \quad (2.68)$$

Note that the upper integration limit ∞ may be replaced by t , since it is a causal system.

To illustrate this application, let us consider buoy-1, whose excitation force IRF is shown in Fig. 2.15a, and the causalising time shift $t_c = -5$ s. The Hankel SVD shown in Fig. 2.15b shows that state orders greater than 10 could be neglected. In fact, it can be seen in Fig. 2.15c that the system with order 15 does not present a significant improvement with respect to the order 10.

Now, let us consider the excitation force time series generated from a wave spectrum Fig. 2.16a. The correspondent wave data series is used to calculate the excitation force through the state space models with order 5, 10 and 15. Figs. 2.16b to 2.16d show the response of the models and the error compared to Fig. 2.16a. In addition, it can be seen that the state space models result in a 5 s lagging response.

As a conclusion, any of the presented approaches can be used to calculate the excitation force. However, the latter one seems to present fewer limitations, since it can be applied online, for synthetically generated or recorded wave time series, despite the delay introduced.

2.7.3 WEC State Space Model

Finally, substituting the fluid memory model (Eq. 2.58) in Eq. (2.52) and making $M = m + A(\infty)$, the *Cummins equation* can be represented by its state space model as follows:

Figure 2.15 State-space realization of the excitation force IRF (h_e) of the buoy described in Section 2.4 by the Hankel SVD method.

$$\dot{\mathbf{x}}(t) = \begin{bmatrix} 0 & 1 & \mathbf{0}_{1 \times n} \\ -K_w/M & -\mathbf{D}_r/M & -\mathbf{C}_r/M \\ \mathbf{0}_{n \times 1} & \mathbf{B}_r & \mathbf{A}_r \end{bmatrix} \cdot \mathbf{x}(t) + \begin{bmatrix} 0 \\ 1/M \\ \mathbf{0}_{n \times 1} \end{bmatrix} \cdot [f_e(t) + f_{pto}(t)] \quad (2.69)$$

$$\equiv \mathbf{A} \cdot \mathbf{x}(t) + \mathbf{B}f_e(t) + \mathbf{B}f_{pto}(t)$$

$$\mathbf{y}(t) = \mathbf{C}\mathbf{x}(t) \quad (2.70)$$

where $\mathbf{x} = [y(t) \ \dot{y}(t) \ \mathbf{x}_r(t)]^T$ and n is the state space order of the approximation for the memory term f_r' . The output vector $\mathbf{y}$ is determined by vector $\mathbf{C}$ which, in turn, can be defined according to the variable of interest in the state vector $\mathbf{x}$.

(a) Excitation force generated from the wave spectrum.

(b) State-space model 5th order.

(c) State-space model 10th order.

(d) State-space model 15th order.

Figure 2.16 Comparison of the excitation force generated from a wave spectrum and a known wave data series with the one calculated, in real time, from the same wave data series by means of the IRF realised by a state-space model with different orders.

Control of Wave Energy Converters

One of the biggest challenges that WEC have to face to enter the race for renewable energy is being profitable for investors. A first and such an essential step for this is to make WEC efficient. A crucial task toward efficiency is the WEC design. As seen in Chapter 2, the hydrodynamic parameters of a floating structure depend on its shape. Results obtained by Nazari Berenjkoo, Ghiasi, and Guedes Soares (2019) indicate that the shape directly affects the motion by modifying the hydrodynamic damping of the buoy. In this sense, its geometry is designed to extract the maximum power for a sea condition. The WEC design seeks to achieve a resonance bandwidth for a given sea characteristic. Many approaches have addressed this issue, e.g., Shadman et al. (2018) use the design of experiments methodology, a statistical technique that consists in exploring the relationship between variables and response; Blanco et al. (2018), in turn, make use of evolutionary algorithms to optimise the buoy geometry including technical constraints of a PTO system. Moreover, Garcia-Rosa and Ringwood (2016) study the relationship between geometry and the control strategy applied by the WEC, which must also be considered at the design stage.

Once designed, the floating structure is rarely changed. So, control strategies must be implemented to improve device performance beyond the natural response of the buoy. These strategies act on changing the damping and/or stiffness of the hydrodynamic system through the PTO force. In this way, the WEC can be better adjusted for more specific wave conditions. Therefore, in this chapter, a new approach to WEC control is presented, as well as methods for its implementation. Besides, a detailed model of the WEC and its PTO system have been developed and explained throughout this chapter.

3.1 Control Strategies

In this section, the most relevant control strategies are described and discussed. Also, they have been grouped according to its working principle based on the classification presented by Hals, Falnes, and Moan (2011a). Most of them are time-based strategies and they can be classified according to the time range involved (Ekström, Ekergård, and Leijon, 2015):

Season sea state: control parameters are tuned to a sea state. For example, Duclos, Babarit, and Clément (2006) present two approaches for wave data over a year: i) using prevailing sea state; and ii) using sea state statistics.

Daily fluctuation: control parameters are tuned more frequently than above to the sea state, which may vary cyclically throughout the day due to tidal effects, directly affecting efficiency, according to Tyrberg, Waters, and Leijon (2010).

Hourly control: the current sea state is used to tune the control parameters. Despite it is a slow control, it is faster than the previous ones.

Wave-by-wave control: the fastest control. The control parameters are tuned within wave cycles or short time intervals. This modality requires more sophisticated control systems.

The season sea state and daily fluctuation controls use wave data statistics to tune control parameters. This adjustment has to be kept constant for an extended period, and it may be combined with other control strategies. On the other hand, hourly and wave-by-wave controls require prediction of either the sea state in the next control period or the variation of the sea surface in a time horizon.

3.1.1 Resistive Loading

Resistive loading control consists in adding a resistive mechanical load (damping), R_{pto} , to the hydrodynamic model through the PTO system so that it is modelled as follows:

$$f_{pto}(t) = -R_{pto} \cdot v(t) \quad (3.1)$$

where the PTO force, $f_{pto}(t)$, is in phase with the speed of the buoy, $v(t)$. Note that $R_{pto} = B_{pto}$. This control strategy is represented by its block diagram in Fig. 3.1, where the WEC dynamic is represented by the transfer function $H(s)$.

This control is characterised by uni-directional power flow and, therefore, R_{pto} must be positive. Also, it is only able to regulate the amplitude motion; because of this, it is

Figure 3.1 Simplified scheme of the resistive loading control.

also known as amplitude control (Bubbar, Buckham, and Wild, 2018; Ricci et al., 2011; Shek, Macpherson, and Mueller, 2008; Valério, Beirão, and Costa, 2007).

From the operating point of view, the control can be classified as passive or active. Passive control is simple to implement, since a prediction is not required, and the PTO system provides linear damping. Waters et al. (2011) present experimental results of a WEC connected to a three-phase resistive load, which provide a good insight into energy absorption. Hong et al. (2016a) compare three different approaches of passive control: linear resistive load, passive rectification, and resonance circuit. The results indicate that linear resistive load presents a better energy efficiency than the others, although the study should not be considered as conclusive. On the other hand, active control allows changing the PTO damping, R_{pto} , at any instant of time to obtain better device performance according to the local wave conditions. This control can be carried out in different ways depending on the type of PTO system used. For example, it could be applied by a hydraulic system (Ricci et al., 2011) or by an electrical system (Tedeschi et al., 2011; Tedeschi and Molinas, 2011; Villa Jaén, Andrade, and Garcia Santana, 2013).

In Section 2.5.2, the expression for the maximum captured power has been developed for regular waves, where it is in terms of mechanical load $\underline{Z}_{pto}$. The same procedure can be used to determine the optimal resistive mechanical load, R_{pto}^{op} , for a regular wave, as follows:

$$R_{pto}^{op} = \sqrt{R_w^2(\omega) + X_w^2(\omega)} \quad (3.2)$$

3.1.2 Reactive Control

This control is considered the true optimal control for WECs, and it is based on the impedance matching of electrical circuits, i.e., it consists in cancelling the mechanical reactive load of the hydrodynamic system, achieving resonance. Thus, speed is in phase with excitation force. This concept has already been developed in Section 2.5.2 for regular waves. Unlike the resistive load, off resonance, its power flow is bi-directional, that is, the PTO system is not only power-fed, rather it may provide some power to the hydrodynamic system, as stated by Falnes (2002b). The reactive control is divided into two categories: complex-conjugate control and phase/amplitude control.

The scheme of complex-conjugate control is represented in Fig. 3.2 by its block diagram. For regular waves it is a very straightforward control. However, for irregular wave applications, it could lead to instability problems. The PTO mechanical impedance is designed to operate on the resonance band, which is narrow. Off resonance, the PTO mechanical impedance could be larger than the hydrodynamical impedance and, then, the controlled system is characterised by positive poles, leading to an unstable system. Sub-optimal strategies have been proposed to overcome this issue. For example, Hals, Falnes, and Moan (2011a) suggest that the parameters of the PTO mechanical impedance must be chosen within a margin that ensures the system stability. Another possibility is to use an active control, but, as highlighted by Falnes (2002b), the need for speed prediction might represent a problem for real-time control. Yavuz et al. (2007) compare three active control strategies with other three passive ones. Their active control strategies are based on tuning the natural frequency of the WEC by estimating the wave period. Best results are obtained with a continuously adjustment of the controller parameters.

Figure 3.2 Simplified scheme of the complex-conjugate control.

The phase/amplitude control consists in generating a speed signal to be tracked, which leads to a maximum absorbed power. From Section 2.5.2, the optimal speed can be written, in the frequency domain, as:

$$V(j\omega) = \frac{F_e(j\omega)}{2R_w(\omega)} \quad (3.3)$$

The speed signal tracking can be achieved by a proportional-integral (PI) controller. This control scheme is shown in Fig. 3.3.

If $R_w(\omega)$ varies with frequency, the transfer function of Eq. (3.3) becomes non-causal and, therefore, a foreknowledge of the excitation force is required. The non-causal function can be approximated to a causal function to avoid predicting the excitation force. Valério, Beirão, and Costa (2007) apply a straightforward implementation, where the absorbed energy is maximised by a direct search method. To overcome the non-causality, they

Figure 3.3 Simplified scheme of the phase/amplitude control.

consider Eq. (3.3) as constant; thus, causal. On the other hand, Bacelli et al. (2019) use a system identification technique (grey-box model) to obtain an approximated causal transfer function within a limited band. Anderlini et al. (2018) use a machine learning approach to tune the controller parameters subjected to minimum and maximum values.

However, despite being the only true optimal strategy, reactive control has some drawbacks, which are also commented by Falnes (2002b). The WEC, led to a quasi-resonance operation, can present high oscillation amplitudes, that is completely unrealistic since the floating body is not expected to be completely out of the water. Linked to that, the control must also be designed subjected to mechanical and/or electrical constraints. Finally, the reactive mechanical power may be high and, then, the PTO system must be highly efficient.

3.1.3 Discrete Control

This control strategy is another sub-optimal one. In this case, phase control is applied, but with the advantage that there is no reactive mechanical power involved. The discrete control can also be combined with other strategies, for example, resistive loading. A generic model of this controller is shown in Fig. 3.4, which illustrates the aforementioned combined control system. κ is the control variable, that can assume the values of 0 or 1, and G is a coefficient that depends on the discrete control technique applied: latching control or declutching control.

Figure 3.4 Simplified scheme of generic discrete control.

Latching control is not really a new strategy, and it was initially proposed by Budal and Falnes (1980). It consists in locking the buoy motion at zero crossing of speed and, then, releasing the buoy after a while. As a result, it is possible to shift the speed waveform so that it is in phase with the excitation force waveform. Thus, the WEC can highly increase the absorbed energy. The biggest challenge faced by this control strategy is determining the latching time; that is, the time the buoy must remain locked. In Fig. 3.4, G would be the latching coefficient, and it must be large enough to counteract all the forces of the system. In practice, the latching can be achieved by a mechanical brake, friction coupling or hydraulic circuits.

For regular wave application, the latching time can be analytically calculated as presented by Babarit, Duclos, and Clément (2004). It is seen that the most significant increase in power absorption occurs for frequencies lower than the resonance one. Besides, for irregular waves, they formulate three different strategies based on foreknowledge of the excitation force. Their results do not show a strategy superior to the others, but, whatever it is, presents a noticeable increase in absorbed energy compared to an uncontrolled system. On the other hand, Sheng, Alcorn, and Lewis (2015) propose a strategy where the latching time only depends on the statistical period of the wave and the resonance period. In this case, they use the energy period.

Falcão (2008) proposes a latching strategy in which the latching time is not calculated. Instead, the system is released when some threshold is achieved. In this case, for hydrodynamic forces larger than the PTO force. Thus, a prediction is not required, and its implementation is straightforward. Nevertheless, this study is limited to a hydraulic system. Lopes et al. (2009) use a similar approach to Falcão (2008), but the system is released when the excitation force achieves a chosen threshold. If the threshold is zero, both last strategies are the same.

Latching control seems to be a good solution for phase control, even though there are some drawbacks. The intermittent power could be a problem for a direct-drive system that uses linear generators. Moreover, the linear generator may require a large amount of energy to produce higher damping. Therefore, the final energy balance may be unsatisfactory. Temiz et al. (2018) carry on a study to implement the latching control in a direct-drive system. Despite the increase in absorbed energy, its implementation could put the mechanical and electrical devices to risk. However, they only evaluate the absorbed power and not the electrical power delivered to a grid. As the reactive control, constraints are not handled. In this sense, Villa Jaén, Santana, et al. (2015) proposed an algorithm to carry out the latching control, considering the WEC constraints. However, it is important to note that to do so, a prediction is always required.

The declutching control, also known as “freewheeling” (Salter, Taylor, and Caldwell, 2002), consists in by-passing the PTO system during part of the cycle. As a result, the floating structure moves freely within this time interval. In Fig. 3.4, $G = -R_{pto}$, and,

therefore, no damping is provided by the PTO system. Other noticeable difference is that it is applied directly to the PTO system, while the latching control is applied to the moving part. Babarit, Guglielmi, and Clément (2009) propose an optimisation for the declutching control applied to a hydraulic system. From the PTO hardware point of view, the system only requires an additional valve to carry out this control strategy. On the other hand, the foreknowledge of the excitation force is also required.

Apart from the mentioned strategies, some studies present a combined strategy known as latching-operating-declutching control (Clément and Babarit, 2012). Despite demonstrating the best results in terms of energy absorption, it is not clear how feasible its implementation is. Given the importance of discrete controls, Feng and Kerrigan (2015) propose an algorithm to unify these discrete control strategies in the optimisation problem.

3.1.4 Model-Predictive Control (MPC)

With increasing computing power, MPC is becoming fashionable in many fields where a highly controllable system is required, and it could not be any different from WEC control. It can handle multi-input multi-output (MIMO) systems and, also, a considerable number of constraints. Real-time control is provided, and it is close to constrained optimal (Qin and Badgwell, 2003). This control technique, at each time step, for a given short-horizon prediction, T_h , defines the control by solving an open-loop optimisation problem and, thus, all the control parameters are updated continuously at each new program cycle. For the optimisation problem, a quadratic programming (QP) is usually applied. Fig. 3.5 shows schematically the block diagram of this control system.

Figure 3.5 Simplified scheme of MPC system.

Brekken (2011) presents an initial study of MPC applied to a simple point absorber to track the optimal speed trajectory within the operational constraints of speed, displacement and PTO force. The estimation of the excitation force is carried out by a Kalman

filter. Some strategies to seek a feasible solution are proposed by relaxing some constraints. It is also shown that MPC can be useful to prevent damages under operation in energetic seas. For nonlinear models, a nonlinear MPC could be applied. Indeed, Richter et al. (2013) use it for a point absorber with two bodies moored. They conclude that the presence of only the nonlinear mooring force does not justify the use of nonlinear MPC. However, it should be kept in mind in cases of more substantial nonlinearities. MPC application requires an accurate system model. Therefore, according to Freeman, Dai, and Sutton (2014), OWC-based systems, for example, are not accurate enough for MPC.

Moreover, MPC seems to be a good solution for direct drives, since the linear generator model can be incorporated into the plant system. In this sense, O'Sullivan and Lightbody (2017) and Villa Jaen, Santana, et al. (2014) implement an MPC, where mechanical and electrical constraints are taken into account. Besides, the cost function is not the absorbed power, but the power delivered by the linear generator.

3.2 PTO Systems

Unlike other forms of renewable energy, the conversion of wave energy into electricity can be carried out in many ways. The most suitable PTO system depends on the device used. Furthermore, depending on its characteristics, a control strategy can be better adapted from both points of view: implementation and operation.

The objective of this section is not to detail the operation of all known PTO systems, but to provide a classification of the most common ones. The chain conversion from wave energy to electricity for most PTO systems is shown in Fig. 3.6, which is an adaptation of the scheme presented by Ahamed, McKee, and Howard (2020). Following, some PTO technologies are presented to illustrate it.

A direct mechanical drive system is characterised by having no other intermediate system between the mechanical and electrical powers. The PTO system used by Tedeschi and Molinas (2012) is composed of teeth piston connected to a gearbox responsible of converting the motion from linear to rotary. Finally, a shaft connects the gearbox to a rotary electrical generator.

Some PTO technologies convert the mechanical power of the wave to a fluid power before the electric conversion. Overtopping devices use hydro turbine system whose working principle is similar to hydroelectric power plants. The water flow drives a coupled system turbine-generator. Probably the most known is the *Wave Dragon*. Attenuators (Henderson, 2006) and point absorbers (Falcão, 2008; Ricci et al., 2011) usually are implemented with a hydraulic motor system. Finally, the pneumatic air turbine system is mostly used by OWC devices (Falcão and Henriques, 2016).

The last category is the direct electrical conversion system that, as the name itself says, directly converts wave energy into electrical energy. A piston is connected to the

Figure 3.6 Main PTO systems and their operating principles (adapted from Ahamed, McKee, and Howard, 2020).

moving part of a linear generator that can be permanent magnets (Leijon et al., 2008; Polinder, Damen, and Gardner, 2004) or switched reluctance (Du, Liang, and Liu, 2017; Santos et al., 2012). This system is highly efficient, and the hardware required for grid connection can also be used for WEC control. Furthermore, as there are no intermediate mechanical stages, the maintenance costs are lower. On the other hand, their cost may be high, and they usually require a huge structure.

3.3 Uppsala's converter L1

The model used in this work is based on the point absorber converter developed at Uppsala University (Lejerskog et al., 2011). It is a direct drive WEC that is composed of a buoy connected to a PTO system on the seabed. The energy conversion in the PTO is

made through a Linear Permanent Magnet Synchronous Generator (LPMSG), a complex and somewhat expensive machine. However, by using it, there is no need to apply any intermediate system, such as hydraulic or pneumatic, to convert low speed from ocean waves (frequency around 0.1 Hz) into high speed for conventional rotary generators. In addition, not having such mechanical connections is very interesting for reducing maintenance costs.

The one degree of freedom point absorber, schematically shown in Figure 3.7, has been implemented in SIMULINK/Matlab, where the input to the model is the excitation force of the incoming wave. Relevant specifications are listed in Table 3.1. A cylindrical buoy is connected to the translator by a rope with high stiffness. The translator is the moving part of the LPMSG, where the magnets are set, and the stator is the static part with the windings. The nominal values of this generator related to the nominal speed of the translator (ν_n) are also given in Table 3.1. There are upper and lower end stop springs to limit the maximum stroke and, then, to avoid damage to the PTO structure during the translator motion. In addition, another spring system connects the translator to the bottom of the structure. This system of retracting springs applies a downward force on the translator during wave troughs and an initial force F_0 to reach a specific buoy draft when the system is in equilibrium. The characteristics of the WEC are based on the compilation of information available for the Uppsala's converter L1 (Bostrom et al., 2013; Eriksson, 2007; Tyrberg et al., 2011). Finally, the LPMSG is connected to a power electronic converter.

Figure 3.7 Schematic representation of the WEC.

3.3.1 WEC Dynamic Model

The model of Uppsala's converter L1 can be seen, very similarly, as the one in Section 2.7.3. However, since the buoy is not rigidly connected to the translator, the system consists of the following coupled integrodifferential equations in which x is the translator position

Table 3.1 Main electrical and mechanical parameters of WEC.

Buoy		Linear Generator		Mechanical System	
Diameter	3 m	Nominal speed ν_n	0.7 m/s	Upper spring length l_u	250 mm
Height	0.8 m	Line to line nominal RMS voltage U_n	200 V	Lower spring length l_l	140 mm
Mass	1000 kg	Nominal Power P_n	10 kW	Upper stiffness k_u	243 kN/m
Draft at still water level	0.4 m	Pole width τ_p	50 mm	Lower stiffness k_l	215 kN/m
		Synchronous inductance L_s	11.7 mH	Energy storage stiffness k_s	6.20 kN/m
		Armature resistance R_a	0.44 Ω	Rope stiffness k_w	450 kN/m
		Translator weight	1000 kg	Initial spring force F_0	8.12 kN

and y is the buoy position:

$$M\ddot{y}(t) + \int_{-\infty}^t h_r(t - \tau)\dot{y}(\tau) d\tau + K_w y(t) = f_e(t) - f_w(t) \quad (3.4a)$$

$$m_t\ddot{x}(t) = f_w(t) + f_{pto}(t) + f_s(t) + f_{stop}(t) - W_t \quad (3.4b)$$

Eq. (3.4) related to the buoy is derived from the Cummins equation (Eq. 2.52) where f_w is the rope pulling force. Eq. (3.4b) is related to the translator motion, where x , $\ddot{x}$, m_t , W_t are the displacement, acceleration, mass, and weight force of the translator, and f_{em} , f_s , and f_{stop} are the electromagnetic, retraction, and end stop forces.

The rope pulling force, f_w , on the connection between the buoy and the translator acts as a coupling point between both systems, buoy and electrical generator. Depending on the position of both buoy and translator, f_w can be null, decoupling both systems, as described by:

$$f_w = \begin{cases} -k_w(y - x) & \text{if } y > x \\ 0 & \text{else,} \end{cases} \quad (3.5)$$

where k_w is the rope stiffness.

The PTO force, f_{pto} , in this case, is the electromagnetic force, f_{em} , provided by the linear generator which is further detailed in Section 3.3.2. The retraction force, f_s , is the force applied by the spring system at the bottom of the structure and modelled by:

$$f_s = -F_0 - k_s x \quad (3.6)$$

where k_s is the spring stiffness, and F_0 is the force of the retraction spring in the equilibrium. The end-stop force, F_{stop} , is the force applied by the upper and lower springs to brake the translator motion. F_{stop} is decomposed into:

$$f_u = \begin{cases} -k_u(x - l_u) & \text{if } l_u < x \\ 0 & \text{else,} \end{cases} \quad (3.7a)$$

$$f_l = \begin{cases} -k_l(l_l + x) & \text{if } l_l < -x \\ 0 & \text{else,} \end{cases} \quad (3.7b)$$

where f_u and f_l are the upper and lower end forces, k_u and k_l are the the upper and lower spring stiffness, and l_u and l_l are the upper and lower spring lengths.

Finally, this coupled model with non-linear forces can also be represented in the state-space model as:

$$\dot{\mathbf{x}} = \begin{bmatrix} 0 & 1 & \mathbf{0}_{1 \times n} & 0 & 0 \\ -K_w/M & -\mathbf{D}_r/M & -\mathbf{C}_r/M & 0 & 0 \\ \mathbf{0}_{n \times 1} & \mathbf{B}_r & \mathbf{A}_r & \mathbf{0}_{n \times 1} & \mathbf{0}_{n \times 1} \\ 0 & 0 & \mathbf{0}_{1 \times n} & 0 & 1 \\ 0 & 0 & \mathbf{0}_{1 \times n} & 0 & 0 \end{bmatrix} \cdot \mathbf{x} + \begin{bmatrix} 0 & 0 \\ 1/M & 0 \\ \mathbf{0}_{n \times 1} & \mathbf{0}_{n \times 1} \\ 0 & 0 \\ 0 & 1/m_t \end{bmatrix} \cdot \begin{bmatrix} f_1 \\ f_2 \end{bmatrix} \quad (3.8)$$

$$f_1 = f_e - f_w \quad (3.9)$$

$$f_2 = f_w + f_{pto} + f_s + f_{stop} - W_t \quad (3.10)$$

with the state variables $\mathbf{x} = [x_1 \ x_2 \ \dots \ x_n]^T$, where x_1 and x_2 are the displacement and speed of the buoy, respectively; x_3 to x_{n-2} belong to the state vector $\mathbf{x}_r$; x_{n-1} and x_n are the displacement and speed of the translator, respectively. Fig. 3.8 shows the mechanical responses of the WEC to a regular wave. Note the coupling and decoupling action of the buoy and the translator by the rope force.

Figure 3.8 Mechanical responses of the Uppsala L1's converter.

In case of rogue waves, end-stop springs bound the translator motion and are also

responsible for braking it and so for avoiding any damage to the PTO structure as illustrated in Fig. 3.9, where the translator reaches its maximum stroke and, therefore, the end-stop spring system reacts by stopping the translator motion.

Figure 3.9 Mechanical responses of the Uppsala L1's converter for rogue wave scenario.

3.3.2 The Linear PM Generator

The model of the linear PM synchronous generator is similar to a rotating synchronous generator, and it can be developed in the same way. However, there are small differences between them that must be considered. Quantities used in rotating machines such as angular speed and torque are replaced in linear machines by their equivalents: linear speed and force. In addition, LPMSG may present a situation in which the PM flux linkages could be lower than its rated value. So, in this section, the LPMSG model based on a conventional PM synchronous machine is presented.

Equivalent Rotating Machine

In most cases, a linear machine has its structure derived from a rotating machine. For example, let us consider the rotating generator in Fig. 3.10, where a cut is made on the imaginary dashed line and, then, unrolled. The rotation motion of the rotor becomes translational; hence the moving part is often known as translator.

To define the equivalent between both machines, let us consider the air gap distance and the PM height are negligible. So, the equivalent rotor radius is:

$$r_r \approx \frac{l_s}{2\pi} \quad (3.11)$$

where l_s is the stator length. As the pole pitch is:

$$\tau_p = \frac{2\pi \cdot r_r}{p} \quad (3.12)$$

Figure 3.10 The equivalent rotating machine.

where p is the number of poles. So,

$$p = \frac{l_s}{\tau_p} \quad (3.13)$$

On the other hand, the linear speed, ν , in the periphery of the rotor is related to its angular speed, ω_m , as $\nu = r_r \omega_m$. As the electrical speed is:

$$\omega = 2\pi f = \frac{p}{2} \cdot \omega_m \quad (3.14)$$

the relationship between electrical and linear speeds can be obtained by combining Eqs. (3.13) and (3.14) resulting in:

$$\omega = \frac{\pi}{\tau_p} \cdot \nu \quad (3.15)$$

Stator Reference Frame Model

The voltage equations of the stator windings of a PM synchronous machine can be expressed as:

$$\mathbf{u}_s = \mathbf{R}_s \mathbf{i}_s + \frac{d}{dt} \boldsymbol{\lambda}_s \quad (3.16)$$

where $\mathbf{u}_s$, $\mathbf{i}_s$ and $\boldsymbol{\lambda}_s$ are generalised in the form:

$$\mathbf{f}_s = \begin{bmatrix} f_a \\ f_b \\ f_c \end{bmatrix} \quad (3.17)$$

where f_a , f_b and f_c are the stator voltage, current or flux linkage in phase- a , $-b$ and $-c$; and $\mathbf{R}_s$ is the diagonal matrix with the stator winding resistance per phase:

$$\mathbf{R}_s = \text{diag}[R_s \quad R_s \quad R_s] \quad (3.18)$$

The flux linkage matrix is expressed as:

$$\boldsymbol{\lambda}_s = \mathbf{L}_s \mathbf{i}_s + \boldsymbol{\lambda}_{\text{pm}} \quad (3.19)$$

where the inductance matrix is

$$\mathbf{L}_s = \begin{bmatrix} L_{aa} & M_{ab} & M_{ac} \\ M_{ba} & L_{bb} & M_{bc} \\ M_{ca} & M_{cb} & L_{cc} \end{bmatrix} \quad (3.20)$$

with the self-inductances L_{aa} , L_{bb} and L_{cc} ; and the mutual inductances M_{ab} , M_{ac} , M_{ba} , M_{bc} , M_{ca} and M_{cb} ; the last term of Eq. (3.19) is the vector of flux linkages in phases a , b and c due to the PM on the translator (PM flux linkage).

Assuming a sinusoidal PM flux linkage and a null flux linkage in phase a (q -alignment) at position $x = 0$ (Fig. 3.11), the PM flux linkage can be written as:

Figure 3.11 Rotor flux position is zero for q -axis aligned with phase a -axis.

$$\boldsymbol{\lambda}_{\text{pm}} = \begin{bmatrix} \lambda_{\text{pma}} \\ \lambda_{\text{pmb}} \\ \lambda_{\text{pmc}} \end{bmatrix} = \begin{bmatrix} \hat{\lambda}_{\text{pm}} \cdot \sin\left(\frac{\pi}{\tau_p} y\right) \\ \hat{\lambda}_{\text{pm}} \cdot \sin\left(\frac{\pi}{\tau_p} y - \frac{2}{3}\pi\right) \\ \hat{\lambda}_{\text{pm}} \cdot \sin\left(\frac{\pi}{\tau_p} y + \frac{2}{3}\pi\right) \end{bmatrix} \quad (3.21)$$

where τ_p is the pole pitch and $\hat{\lambda}_{\text{pm}}$ is the peak of the PM flux linkage (a constant parameter for a PM synchronous machine) which, in turn, can be calculated by conventional methods as follows (Wang and Wang, 2012):

$$\hat{\lambda}_{\text{pm}} = k_w N_s l_{\text{sw}} \tau_p \frac{2}{\pi} \hat{B}_{\text{pm}} \quad (3.22)$$

where:

- k_w winding factor;
- N_s number of turns of the winding;
- l_{sw} machine stack width;
- $\hat{B}_{\text{pm}}$ fundamental of the PM normal airgap flux density.

The induced voltages (back electromotive force - emf) by the PM can be obtained by measuring the no-load voltage at the generator terminals. So, the stator currents in Eq. (3.16) are zero, resulting in $\mathbf{e} = d\boldsymbol{\lambda}_{\text{pm}}/dt$, which is the Faraday's law. So:

$$\mathbf{e} = \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} = \frac{d\boldsymbol{\lambda}_{\text{pm}}}{dt} = \begin{bmatrix} \hat{\lambda}_{\text{pm}} \nu \frac{\pi}{\tau_p} \cdot \cos\left(\frac{\pi}{\tau_p} y\right) \\ \hat{\lambda}_{\text{pm}} \nu \frac{\pi}{\tau_p} \cdot \cos\left(\frac{\pi}{\tau_p} y - \frac{2}{3}\pi\right) \\ \hat{\lambda}_{\text{pm}} \nu \frac{\pi}{\tau_p} \cdot \cos\left(\frac{\pi}{\tau_p} y + \frac{2}{3}\pi\right) \end{bmatrix} \quad (3.23)$$

where ν is the linear speed in m/s of the translator. Thus, the back emf is proportional to the PM flux linkage by the synchronous speed, ω_s , given in electrical rad/s, with $\omega_s = \pi\nu/\tau_p$ (Eq. 3.14). Therefore, the amplitude of the back emf per phase can be calculated by:

$$\hat{E} = \omega_s \hat{\lambda}_{\text{pm}} \quad (3.24)$$

Direct-Quadrature Reference Frame Model

The PM synchronous machine can also be represented in the direct-quadrature (dq) reference frame as the rotating machines. Adopting the dq -frame alignment at $t = 0$ shown in Fig. 3.11, i.e., $\theta_r = \omega_s t = 0$, the rotating frame is 90 degrees behind a-axis. Therefore,

the transformation relationship from abc -frame to dq -frame is given by:

$$\begin{bmatrix} f_d \\ f_q \\ f_0 \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \sin(\omega_s t) & \sin\left(\omega_s t - \frac{2\pi}{3}\right) & \sin\left(\omega_s t + \frac{2\pi}{3}\right) \\ \cos(\omega_s t) & \cos\left(\omega_s t - \frac{2\pi}{3}\right) & \cos\left(\omega_s t + \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} f_a \\ f_b \\ f_c \end{bmatrix} \quad (3.25)$$

Assuming a sinusoidal distribution of the windings and a surface magnet translator, the voltage equations for the stator windings transformed into the translator d - q reference frame can be expressed as:

$$u_d = R_s i_d + \frac{d}{dt} \lambda_d - \omega_s \lambda_q \quad (3.26a)$$

$$u_q = R_s i_q + \frac{d}{dt} \lambda_q + \omega_s \lambda_d \quad (3.26b)$$

where the dq flux linkages at the stator windings are defined by:

$$\lambda_d = L_s i_d + \hat{\lambda}_{pm} \quad (3.27a)$$

$$\lambda_q = L_s i_q \quad (3.27b)$$

where L_s is the synchronous inductance. Note below that only the PM flux linkage in d -axis is different than zero; therefore, it is represented by its value.

$$\begin{bmatrix} \lambda_{pmd} \\ \lambda_{pmq} \\ \lambda_{pm0} \end{bmatrix} = \begin{bmatrix} \hat{\lambda}_{pm} \\ 0 \\ 0 \end{bmatrix} \quad (3.28)$$

The electromagnetic force generated by the PM synchronous machine and, therefore, the LPMSG, can be found based on Eqs. (3.26a) and (3.26b). For this, first, the active output power of the LPMSG, that is the PTO active power (p_{pto}), must be calculated. So:

$$p_{pto} = \frac{3}{2} (u_d i_d + u_q i_q) \quad (3.29)$$

Substituting Eqs. (3.26a) and (3.26b) in Eq. (3.29) yields:

$$p_{pto} = \frac{3}{2} R_s (i_d^2 + i_q^2) + \frac{3}{2} \left(i_d \frac{d\lambda_d}{dt} + i_q \frac{d\lambda_q}{dt} \right) + \frac{3}{2} \omega_s (\lambda_d i_q - \lambda_q i_d) \quad (3.30)$$

On the right side of Eq. (3.30), the first term represents the power loss in the conductors, the second term corresponds to the variation of the power stored by the magnetic field and

the last term represents the energy conversion of the machine, and it is the electromagnetic power transferred across the air gap (Kundur, 1994).

Assuring the fundamental energy conversion equation, i.e., $f\nu = ui$, where f is a mechanical force, the electromagnetic force developed by the LPMSG in the PTO that opposes the motion is given by:

$$f_{pto} = f_{em} = \frac{p_{em}}{\dot{x}} = \frac{3}{2} \cdot \frac{\pi}{\tau_p} (\lambda_d i_q - \lambda_q i_d) \quad (3.31)$$

Thus, by substituting Eqs. (3.27a) and (3.27b) in Eq. (3.31), the PTO force can be expressed as follows:

$$f_{pto} = \frac{3\pi}{2\tau_p} \hat{\lambda}_{pm} i_q \quad (3.32)$$

Note that, by taking the PTO force to Eq. (3.10), Eq. (3.32) must be negative.

Active Stator Area

For LPMSG to have a constant amplitude of the PM flux linkage ($\hat{\lambda}_{pm}$), either stator or translator must be long than the other to assure a complete stator-translator overlap (Vining, Lipo, and Venkataramanan, 2009), i.e., translator's PM must cover all the stator windings or the contrary. However, to reduce costs, in most cases, this condition is not satisfied (Polinder, Damen, and Gardner, 2004) and it may result in a lower magnetic flux for a partial stator-translator overlap. In this sense, Eriksson et al. (2007) introduce the active stator area (A factor), A_{fac} , defined as the ratio between stator-translator overlap and the stator areas, which is expressed in function of the translator position x as follows:

$$A_{fac} = \begin{cases} 0, & |x| \geq (l_t + l_s)/2 \\ 1, & |x| \leq (l_t - l_s)/2 \\ \frac{1}{l_s} \left[\frac{1}{2}(l_t + l_s) - |x| \right], & \text{else,} \end{cases} \quad (3.33)$$

where l_t and l_s are translator and stator lengths, respectively.

As it can be seen in Eq. (3.33), the A factor only depends on the translator's position, and its application was experimentally validated by Eriksson et al. (2007) and Hong et al. (2016b). To take into account PM flux linkage decrease, the A factor must be included as a multiplicative factor in both, emf and electromagnetic force.

LPMSG Implementation

The LPMSG model connected to a generic series RLC load can be written as follows:

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ u_{cd} \\ u_{cq} \end{bmatrix} = \begin{bmatrix} -\frac{R_s + R}{L_s + L} & \omega_s & \frac{1}{L_s + L} & 0 \\ -\omega_s & -\frac{R_s + R}{L_s + L} & 0 & \frac{1}{L_s + L} \\ -\frac{1}{C} & 0 & 0 & \omega_s \\ 0 & -\frac{1}{C} & -\omega_s & 0 \end{bmatrix} \cdot \begin{bmatrix} i_d \\ i_q \\ u_{cd} \\ u_{cq} \end{bmatrix} + \begin{bmatrix} 0 \\ -1 \\ 0 \\ 0 \end{bmatrix} \cdot \frac{\omega_s \hat{\lambda}_{pm} A_{fac}}{L_s + L} \quad (3.34)$$

where R , L and C are resistance, inductance and capacitance of the load, respectively. Note that, due to the capacitor on the load, two new state variables arise: u_{cd} and u_{cq} . They are the capacitor voltage in dq -frame. The electromagnetic force can be calculated by solving Eq. (3.34) and incorporating the A factor into Eq. (3.32):

$$f_{em} = \frac{3\pi}{2\tau_p} \hat{\lambda}_{pm} A_{fac} i_q \quad (3.35)$$

This model has also been implemented in Simulink block diagram to facilitate the use of the LPMSG model with Simscape Electrical™ (Simulink's toolbox). The use of one or the other depends on the type of simulation, which is explained later. Fig. 3.12 shows the LPMSG block and the block parameters. There are two signal ports: speed translator (**v**) and set of measurements (**m**). The other three ports are Simscape Electrical™ connection for the three phases.

Figure 3.12 LPMSG block implemented in Matlab/Simulink.

The system inside the LPMSG block (Fig. 3.12) is shown in Fig. 3.13a. Three areas are identified:

Area 1: The open circuit voltage signals or emf are generated from Eq. (3.23), which is implemented inside the emf block (detailed in Fig. 3.13b). The translator speed

(**v**) is converted to synchronous speed (**w_s**) by Eq. (3.14) and, also, it is used to determine the A factor (**A_{fac}**).

Area 2: LPMSG equivalent circuit modelled with Simscape Electrical^{TM} elements (R_s and L_s).

Area 3: the currents measured in area 2 (**i_{abc}**) are transformed to dq -frame (**i_{dq}**) by Eq. (3.25). The electrical angle (**theta**) is calculated by integrating the synchronous speed (**w_s**) that, in turn, is calculated from the translator speed (**v**). The translator speed integration results in its displacement used to determine the A factor (**A_{fac}**). Finally, the electromagnetic force (**f_{em}**) is calculated.

Apart from these three areas, there is a block (measurements) which collects all system measurements in the output (**m**).

To illustrate the operation of the LPMSG model, the translator is forced to a sinusoidal motion that results in a sinusoidal speed whose peak value is 0.7 m/s, the nominal speed (Table 3.1). Thus, at the nominal speed, the line-to-line rms emf must be 200 V. Besides, two different scenarios are considered: (i) $A_{fac} = 1$, that is, the translator is long enough to assure a stator-translator overlap constant; and (ii) the more stator-translator overlap decreases, the lower the value of A_{fac} . Fig. 3.14 shows the translator displacement and speed and, also both emf and electromagnetic force for scenarios (i) and (ii) when the LPMSG is connected to $1\ \Omega$ resistive load.

3.3.3 Full Model Simulation

The mechanical and electrical models of the WEC have been described individually in Section 3.3.1 and ???. The full WEC model, i.e., buoy connected to a PTO system and the latter connected to an electrical load can be achieved by combining inputs and outputs of the models in Eqs. (3.8) and (3.34). It has been also implemented in Matlab/Simulink, where the LPMSG showed in Fig. 3.12 is used.

Fig. 3.15 shows some mechanical and electrical responses of the full WEC model excited by a regular wave with $H = 2$ m and $T = 8$ s. In this example, the average absorbed power is $P_{abs} = 11.07$ kW and the electrical power delivered by the PTO is $P_{pto} = 8.28$ kW. In addition, the currents in dq -frame are presented as well as the PTO current in phase-a.

As it can be seen in Fig. 3.14, the emf varies in both amplitude and frequency, according to the translator speed. Therefore, the PTO current also varies as in Fig. 3.15. To verify its behaviour with a higher level of detail, Fig. 3.16 shows the three phase currents in a shorter time window.

(a) Inside LPMSG block.

(b) emf block detailed.

Figure 3.13 LPMSG Simulink model.

3.4 The Resistance Emulation Approach

In general, the control system of a point absorber oscillating in heave actuates through the PTO system, and it is based on the fact that the position of the actuator and the buoy are directly related. One particularity of the WEC described in Section 3.3 is that there are some situations in which this is not true, particularly when the buoy displacement is less than translator displacement. This situation results in independent dynamics between

Figure 3.14 Simulation of LPMSG with a sinusoidal excitation, $\dot{x}(t) = 0.7 \cos(0.2\pi t)$, connected to a resistive load of 1Ω .

Figure 3.15 Simulation of the full WEC model connected to a resistive load.

buoy and translator; therefore, the control on the translator has no effects on the buoy.

In order to apply a controller that does not take into account the system coupling, this work proposes a sea state based control in a real WEC model whose objective is to maximise the power extraction for each given sea state.

Figure 3.16 PTO phases currents.

3.4.1 Optimal Power Point

At this point, regular waves, i.e., harmonic waves, as described in Section 2.1, are used to excite the system. Their wave-energy transport is defined by Eq. (2.2).

The electromagnetic force developed by the linear generator can be modelled as a viscous damping (Eriksson, Isberg, and Leijon, 2005), which leads to a damping coefficient proportional to the translator speed described by:

$$f_{pto} = -B_{pto} \cdot \dot{x} \quad (3.36)$$

As the electromagnetic force can be calculated with Eq. (3.32), the damping coefficient, B_{pto} , can be found for a given speed. The instantaneous absorbed power can be written as a function of the damping:

$$p_{abs} = B_{pto} \cdot \dot{x}^2 \quad (3.37)$$

According to Nielsen and Plum (2001), the absorbed power, in reality, is hardly above 20% of the wave-energy transport, and, then, many efforts have been devoted to optimising the captured power. Many control strategies, as discussed earlier in Section 3.1, present optimisation methods for generic PTOs by acting directly on the damping coefficient and/or other variables. These methods are usually compared to established control strategies as latching and declutching. However, in this work, the comparison with a latching control is not possible because the buoy is not rigidly connected to the PTO, as generic ones are. The system is unable to “lock” its motion, and the comparison with the declutching control is not feasible due to electrical instabilities in the converter when unloading the LPMSG.

In this work, an optimisation study is proposed based on regular wave responses. To carry it out, the WEC is connected to a three-phase resistive load, R_{load} , as shown in Fig. 3.17 (electric scheme). Realistic seas do not have periodic waves. Usually, they are

described by statistical parameters such as energy period and significant wave height. Using the optimal values obtained from height and period of regular waves, instead of these statistical parameters, as a simplification for irregular waves, as in (Hals, Falnes, and Moan, 2011b; Oskamp and Özkan-Haller, 2012), is the approach used in this work.

Figure 3.17 WEC connected to three-phase resistive load.

Linear PTO system

As a first approximation, the WEC system is considered as totally linear. This means that, unlike the model in Section 3.3.3, nonlinear contributions such as rope and end stop forces are neglected. Besides, the PTO has been modelled according to Eq. (3.36). Consequently, the WEC system that could be either coupled or decoupled now becomes a single system, that is, the buoy is rigidly connected to the PTO system. In this case, the PTO motion is the same as that of the buoy.

The linear WEC system excited by regular waves can, therefore, be represented by its electrical analogue (Section 2.5.1), represented in Fig. 3.18. Current is equivalent to speed $\dot{y}(t) = v(t)$ and forces to voltage sources. Note that there is a sinusoidal source due to regular wave excitation and two DC sources due to the translator weight and the force of the retraction spring in equilibrium.

Figure 3.18 Electrical analogue of the linear WEC system.

Similarly to the approach used in Section 2.5.2, the linear WEC system can be studied in the frequency domain by phasor analysis. The DC sources do not contribute to the steady-state current of the circuit. Thus, the speed phasor, $\underline{V}$, resulting in Fig. 3.18 can

be expressed as:

$$\underline{V} = \frac{\underline{F}_e}{\underbrace{B(\omega) + B_{\text{pto}}}_{R_{\text{eq}}} + j \underbrace{\left[\omega (m + A(\omega) + m_t) - \frac{K_b + K_s}{\omega} \right]}_{X_{\text{eq}}}} \quad (3.38)$$

The active power absorbed by the PTO impedance, $\underline{Z}_{\text{pto}} = B_{\text{pto}}$, is the average power absorbed by the WEC, i.e., $P_{\text{abs}} = B_{\text{pto}} V^2$. Thus:

$$P_{\text{abs}} = \frac{B_{\text{pto}} \cdot F_e^2}{R_{\text{eq}}^2 + X_{\text{eq}}^2} \quad (3.39)$$

where F_e is the rms value of the excitation force $f_e(t)$; R_{eq} and X_{eq} are the total resistance and reactance of the electric analogue of the hydrodynamic system.

Apart from the B_{pto} , other system elements depend on the design characteristics of the WEC and, therefore, cannot be controlled. In this way, B_{pto} assumes the role of the control variable. The maximum absorbed power is obtained by applying the maximum transfer theorem, where the optimal PTO damping is calculated by:

$$B_{\text{pto}}^{\text{op}} = \sqrt{B(\omega)^2 + \left\{ \omega [m + A(\omega) + m_t] - \frac{K_b + K_s}{\omega} \right\}^2} \quad (3.40)$$

As it can be seen in Eq. (3.40), the optimal PTO damping of this linear WEC system only depends on the wave period ($T = 2\pi/\omega$). On the other hand, the absorbed power depends on both, PTO damping and wave period, as shown in Fig. 3.19a for $H = 2$ m. Note that the maximum value of RAO in Fig. 2.8 is at $T \approx 2$ s. Therefore, the maximum absorbed power (Fig. 3.19a) tends to be higher, the closer to the wave period corresponding to the maximum RAO.

Figure 3.19 Linear WEC system.

In conclusion, for regular waves and linear systems, it is quite straightforward to track the maximum power point of the WEC operation by tuning the PTO damping with its

optimum value, corresponding to the period and the height of the wave. This theoretical case is illustrated in Fig. 3.19b, where optimal PTO damping is mapped according to both period and height of a regular wave. Note that the optimal PTO damping has been analytically mapped by using Eq. (3.40).

WEC system

Although the premises in the previous section are correct, the buoy has a boundless motion which does not represent a real situation completely. So the same methodology used to map the optimal PTO damping that maximises the absorbed power is going to be used for the full WEC model.

Regular incident waves result in periodic responses that can easily reach steady-state, even in the presence of nonlinearities. On the other hand, these responses no longer have to be monochromatic, i.e., with only one harmonic. Hence, the phasor solution of an electrical analogue circuit is not valid anymore.

Therefore, the full WEC model is used to obtain the system responses for regular waves. However, in this case, the maximum power point is reached by adjusting the resistance, R_{load} , connected to the LPMSG. The system has been subjected to multiple excitations to find the optimal R_{load} that maximises the absorbed power. Wave period (T) has been varied between 2 and 16 seconds with steps of 0.1 seconds. Wave height (H) has been varied between 0.1 and 3 meters with steps of 0.1 meters. These wide ranges of periods and heights were chosen not taking into account any specific place. For each period and height, the system was loaded with different values of resistance. Fig. 3.20a shows how P_{ratio} varies according to R_{load} for different values of period and height. The decrease in absorbed power compared to Fig. 3.19a is expected, since the translator motion is bounded in the full WEC model.

Due to the LPMSG losses, the useful power, P_{pto} , delivered by the WEC to the resistive load is less than that absorbed by the buoy. In addition, maximum P_{abs} might not coincide with maximum P_{pto} . As P_{abs} is not the useful generated power, P_{pto} has to be maximised instead. Hence, the expression of the power ratio is slightly modified to expressed as a function of the useful power as follows:

$$P_{\text{ratio}} = \frac{P_{\text{pto}}}{Jd}$$

Fig. 3.20b shows the power ratio of the PTO power ($P_{\text{ratio}}R_{\text{load}}$ -curves) for some sets of period and height, where lower P_{ratio} and different optimal R_{load} , compared to Fig. 3.20a can be seen.

For each $P_{\text{ratio}}R_{\text{load}}$ -curves, it is possible to identify the maximum power point and the respective resistive load, that is defined as the optimal resistance, R_{op} . As a result, optimal resistances can be defined in terms of period and height, as shown in the coloured

Figure 3.20 P_{ratio} for different values of R_{load} for regular waves. Left column for $T = 8$ s and right column for $H = 2$ m.

surface of Fig. 3.21a. The optimal resistance ranges from 0.6 to 9.5 Ω and its histogram is shown in Fig. 3.21b. About 70% of optimal resistances are between 0.6 and 1.5 Ω .

Figure 3.21 Optimal resistance for the full WEC model.

In addition, despite the nonlinearity of the system, it is observed that each load is related to a PTO damping coefficient. Each pair $R_{\text{load}}-B_{\text{pto}}$ is marked in Fig. 3.22 as a grey circle, where the higher the electrical load resistance, the lower the PTO damping

coefficient. A unique representation can be obtained by rational fitting with 95% confidence bound. This relationship is independent of H and T , and it is not a generic curve, being highly dependent on the generator characteristics. Furthermore, it is an essential aspect for a direct drive WEC because it is possible to relate both hydrodynamic and electrical systems.

Figure 3.22 Relationship between the three phase resistive load and the PTO damping.

Fig. 3.23 shows the B_{pto}^{op} map obtained from the R_{load} - B_{pto} relationship (Fig. 3.22). The nonlinearities of the system are mainly related to translator displacement. Thus, the higher the wave, the stronger the nonlinearities. This effect on the system can be seen by comparing Fig. 3.23 and Fig. 3.19b. The optimal PTO damping is not affected by the wave height for the linear WEC system. In contrast, nonlinearities of the full WEC model cause a drop of optimal PTO damping.

Figure 3.23 Optimal PTO damping, B_{pto}^{op} , respect to wave height and period of regular waves.

The proposed method has been applied to wave data time series (WDTS) generated mathematically with different characteristics to verify its effectiveness. Fig. 3.24 shows the results for different resistive loads connected to the terminals of the LPMSG, where the black filled circle represents the optimal resistance, R_{op} . It is reasonable to think that the optimal resistances do not always result in maximum generated power, because they have been found for regular waves. However, its application results in values very close

to the maximum generated powers. In some cases, even the maximum one. In the cases of Fig. 3.24, it can be seen that:

- $H_s = 1$ m, $T_e = 8$ s. The maximum generated power is 1.17% higher than the one generated with R_{op} .
- $H_s = 1$ m, $T_e = 15$ s. The maximum generated power is 0.72% higher than the one generated with R_{op} .
- $H_s = 2$ m, $T_e = 8$ s. The maximum generated power is generated with R_{op} .
- $H_s = 2$ m, $T_e = 15$ s. The maximum generated power is 0.13% higher than the one generated with R_{op} .

Figure 3.24 Performance of the proposed strategy applied to irregular waves.

3.5 Power Electronic Implementation

The electrical energy generated by a LPMSG depends on the linear speed of the moving part, the translator, which in turn depends on the buoy motion. Due to the non-regular and oscillating nature of ocean waves, there is no sinusoidal steady-state in the generated electricity, and direct drive systems (as the PTO in this work) require power electronics to interface with the electrical grid.

Grid connection is a challenge for renewable energies, and it can be seen from two different points of view: generator-side and grid-side converters. The generator-side converter is related to the part of the system from the generator to the ac-dc conversion (ac-dc stage). The grid-side converter corresponds to both dc-ac stage and all the implications for connection to the grid, such as the control of voltage, frequency and active/reactive power.

The proposed method to maximise the power generated by the WEC is based on the sea state parameters (wave height and period) for which there are resistive loads that lead to the maximum power point operation. So, the objective is to implement a circuit able to emulate the resistance, as seen by the linear generator, that allows the WEC to extract the desired power. Emulating a resistance means to make current proportional to voltage and, therefore, to reach unitary power factor.

The main idea is to use the hardware resources already available for connecting the WEC to the grid. Therefore, one way to obtain the resistance emulation is by using the three-phase ac-dc boost rectifier (Erickson and Maksimovic, 2007). This converter, whose topology is shown in Fig. 3.25, is connected to the linear generator by a submarine cable. It is made up of six single-pole single-throw (SPST) current-bidirectional two-quadrant switches and requires an inductor in the ac-side to filter high-frequency harmonics. As the object of study in this section is the WEC control, the dc-ac stage can be simplified and modelled as a dc-source U_{dc} . The boost rectifier requires a dc voltage higher than the peak value of the line-to-line input voltage for proper operation.

Figure 3.25 WEC connected to the boost rectifier.

3.5.1 Current Controllers

The control of a LPMSG is similar to the control of its rotary equivalent, which is usually performed by a Voltage Source Converter (VSC). It works as a voltage source regulated

through current feedback to carry out an indirect torque (or force) and field control (Novotny and Lipo, 1996). Another possibility for a more direct and faster control is to use the electronic converter as a nonlinear current source. Despite traditional algorithms, such as hysteresis-band current controllers (HBCCs), present disadvantages such as non-constant switching frequency and unpredictable harmonic content (Kazmierkowski and Malesani, 1998), they offer simplicity, lack of tracking errors, independence of load parameter changes, and unmatched transient response in comparison with other techniques. These characteristics make it advisable to adopt this method in all cases where high accuracy and wide bandwidth are required (Bhuthukuri and Updhyay, 2018; Malesani, Mattavelli, and Tomasin, 1997) such as the highly irregular (in amplitude and frequency) power profile generated in wave power plants.

Recently, different control strategies for VSCs have been introduced, such as model-based predictive control (MPC). This type of control considers the model of the system and, according to a cost function and the control objectives, chooses the optimal voltage vector (Morel et al., 2009; Zhang, Rodríguez, and Kennel, 2017). However, this approach presents a high ripple in the control variables, and the switching frequency, similar to direct torque control (DTC), is variable, although there are alternative strategies to overcome these downsides (Landsmann and Kennel, 2012; Zhang et al., 2013). Another recent control strategy is the direct power control (DPC) which provides a fast control, but presents variable switching frequency and high ripple in torque and power (Morel et al., 2009; Zhang, Rodríguez, and Kennel, 2017).

Several strategies for fault tolerance (Cecati et al., 2015; Zhou, Zhao, and Liu, 2015) have also been developed for double-fed induction generators (DFIG) and PMSG in wind energy. They could be applicable to wave energy, reducing the maintenance cost, increasing the productivity and the survival of buoy-based WECs in circumstances where control would be lost, etc.

As the control principle proposed in this work is based on emulating the resistance R_{op} , the phase voltages at the generator terminals must be proportional to the line currents. In simple words, this means that the current is controlled to be in phase with the phase voltage. So, in this case, the PTO force can be controlled directly by a current controller and its current reference signals are calculated from the phase voltage measurements as:

$$\mathbf{i}_{abc\text{ref}} = \frac{\mathbf{u}_{abc}}{R_{op}} \quad (3.41)$$

where $\mathbf{u}_{abc} = [u_a \quad u_b \quad u_c]^T$ is a vector with the phase voltages at the generator terminals.

In summary, the choice of the current controllers depends on the system requirements. In this work, two traditional current control strategies are implemented and evaluated: HBCC and Space-Vector PWM Current Control (SVPWMCC). Also, a new algorithm for Nonlinear Vector Current Source (NLVCS) based on spatial hysteresis current source

control is proposed. The current reference direction adopted in this work is the one going out of the boost-rectifier, as shown in the simplified scheme in Fig. 3.26.

Figure 3.26 Resistance emulation system.

Hysteresis-Band Current Controller (HBCC)

This is a straightforward strategy where the measured current signals at the generator terminals track the current references within a hysteresis band, β . For this control scheme, one independent controller per phase is used, as shown in Fig. 3.27 for phase- a . The switching signals of Q_1 and Q_2 are complementary, i.e., while Q_1 is conducting, Q_4 is blocked. The emulated resistance, R_{em} , is obtained by controlling the switching. In other words, the switching sequence provides a voltage in the ac-side of the converter that makes voltage and current proportional at the measurement point.

Figure 3.27 Model of hysteresis-band current controller for one phase.

The principle of operation of the hysteresis-band current control (Brod and Novotny, 1985) is shown in Fig. 3.28. The hysteresis control is based on a nonlinear feedback loop with two-level hysteresis comparator. The switching signals are produced when the absolute error, $|i_{ref} - i|$, exceeds the band $\beta/2$. In this way, each measured current (i_a , i_b , or i_c) is guided to follow its reference (i_{aref} , i_{bref} , or i_{cref}) within the tolerance band β .

The upper switch Q_1 turns on when the measured current i_a is higher than the upper limit of the band ($i_{\max}$), and it turns off when i_a is lower than the lower limit ($i_{\min}$). The special advantage of this controller is the fact that there is no tracking errors (very fast response), what makes it ideally suited for this type of generation as the voltage generated is constantly varying according to the incoming wave. The current is supposed to be kept within the hysteresis band. However, if there is no neutral connection in the circuit, in the case of three independent controllers, the actual current could reach up to twice the hysteresis bandwidth (Brod and Novotny, 1985).

Figure 3.28 Principle of operation of the hysteresis-band current control.

Space-Vector PWM Current Control (SVPWMCC)

This strategy uses vector control to drive a VSC. Therefore, unlike the HBCC, the electrical quantities are represented by their space-vector model (Yazdani and Iravani, 2010). Current controllers based on vector control are usually applied to the electrical machine control (Novotny and Lipo, 1996), known as field-oriented control (FOC), and grid-connection control (Yazdani and Iravani, 2010), known as voltage-oriented control (VOC).

Both control strategies, FOC and VOC, are performed in dq rotating reference frame. In the case of FOC, the d -axis must be aligned with the vector of the rotor flux, $\lambda_{\text{pm}} = \hat{\lambda}_{\text{pm}}$, as shown in Fig. 3.29a. As a result, the emf, $\mathbf{e}$, is 90° ahead on the q -axis. Also, Fig. 3.29a shows the distribution of stator vectors: current, $\mathbf{i}_s$, voltage, $\mathbf{u}$, and flux, λ_s . The angle θ_r is usually obtained by an encoder (Rudnicki, 2019) or an estimate (Boussak, 2005; Kim et al., 2010; Nakashima, Inagaki, and Miki, 2000). On the other hand, Fig. 3.29b shows the distribution of the vectors for VOC. In this case, the d -axis must be aligned with the vector of the grid voltage, $\mathbf{u}$, and the angle θ_s is obtained by a phase-locked loop (PLL) responsible for tracking the grid voltage angle (Chung, 2000; Hsieh and Hung, 1999).

The control strategy proposed in this work to maximise the electricity generation depends on the voltage at the generator terminals (Eq. 3.41), which can be easily measured. The circuit in Fig. 3.30 schematically represents the LPMSG-VSC connection, where $\mathbf{u}$ is

Figure 3.29 Diagram of space vectors for vector control.

the stator voltage; $\mathbf{i}$ the current; R and L a series resistance and inductance, respectively.

Figure 3.30 Equivalent circuit of the LPMSG connected to a VSC.

The circuit in Fig. 3.30 is similar to a VSC connected to a grid. So, in this work, VOC is used to emulate a resistance connected to LPMSG. Another point that justifies the VOC application is the fact that the machine is designed to operate, for a short time, with a translator speed above its rated speed. Therefore, the field-weakening study is not addressed in this work. On the other hand, the angle θ_s is required, and it can be obtained from a speed measurement of the LPMSG. This new measurement is a disadvantage compared to the HBCC that needs the voltage measurements. Although VOC is enough to perform the proposed strategy, future works could include a more detailed control of the LPMSG, i.e., torque and field control, and then FOC is required.

The VSC voltage, $\mathbf{u}_c$, is easily calculated by using the space-vector approach in the equivalent circuit of the generator-side converter shown in Fig. 3.30 as follows:

$$\mathbf{u}_c = R\mathbf{i} + L\frac{d\mathbf{i}}{dt} + \mathbf{u} \quad (3.42)$$

As a generic space-vector can be written as $\mathbf{f} = (f_d + jf_q) \cdot e^{j\theta_s(t)}$, Eq. (3.42) can be

decomposed into real and imaginary parts resulting in:

$$u_{cd} = u_d + Ri_d + L \frac{di_d}{dt} - \omega_s Li_q \quad (3.43a)$$

$$u_{cq} = u_q + Ri_q + L \frac{di_q}{dt} + \omega_s Li_d \quad (3.43b)$$

Both equations represent two decoupled, first-order, linear systems. Rewriting them so that:

$$L \frac{di_d}{dt} + Ri_d = \underbrace{u_{cd} - u_d + \omega_s Li_q}_{u_{1d}} \quad (3.44a)$$

$$L \frac{di_q}{dt} + Ri_q = \underbrace{u_{cq} - u_q - \omega_s Li_d}_{u_{1q}} \quad (3.44b)$$

i_d and i_q can be controlled by u_{1d} and u_{1q} . So this current controller can be implemented through two decoupled closed-loop control systems with a proportional-integral (PI) controller as shown in Fig. 3.31.

Figure 3.31 Simplified scheme of the decoupled current control loops.

The reference voltage of the converter, $\mathbf{u}_c$, in Eqs. (3.43a) and (3.43b) can be realised by a considerable number of techniques, as described by Kazmierkowski and Malesani (1998) and references therein, including the HBCC. The current controller described in this section uses space vector-pulse width modulation or, simply, space vector modulation (SVM), which is a prevalent technique. Compared to other traditional techniques, such as HB or sinusoidal PWM, it requires more computational effort. However, SVM provides a high output voltage, constant switching frequency and well-defined harmonic spectrum (Zeng, Chang, and Song, 2004).

The $\alpha\beta$ -plane is divided into 6 sectors of 60° angle by 6 space vectors ($\mathbf{u}_1$ to $\mathbf{u}_6$) whose magnitudes are the same $(2/3)U_{dc}$, as shown in Fig. 3.32a. Also, there are two null space vectors ($\mathbf{u}_0$ and $\mathbf{u}_7$). The binary sequence corresponds to the switching states of Q_1 , Q_2 and Q_3 (1 for on and 0 for off). The maximum voltage, $\mathbf{u}_{ref\ max}$, in the undermodulation

region, is the radius of the circle inscribed in the hexagon: $0.577 \cdot U_{dc}$. Overmodulation operation, i.e., $\mathbf{u}_{ref} > \mathbf{u}_{ref max}$, is not discussed in this work.

Figure 3.32 Principle of SVM.

The SVM realises a reference space vector, $\mathbf{u}_{ref}$, based on the summation of two adjacent space vectors and a null voltage vector in its correspondent sector over the switching period, T_{sw} . The order of each space vector follows the sequence: $\mathbf{u}_0 \rightarrow \mathbf{u}_k \rightarrow \mathbf{u}_m \rightarrow \mathbf{u}_7 \rightarrow \mathbf{u}_m \rightarrow \mathbf{u}_k \rightarrow \mathbf{u}_0$, where k and m are subscripts of adjacent space vectors. Hence:

$$\mathbf{u}_{ref} = \frac{t_k}{T_{sw}} \mathbf{u}_k + \frac{t_m}{T_{sw}} \mathbf{u}_m + (\mathbf{u}_0 \text{ or } \mathbf{u}_7) \frac{t_0}{T_{sw}} \quad (3.45)$$

The active time of each space vector (t_k , t_m and t_0) can be obtained geometrically in Fig. 3.32a. So, these times for each sector can be calculated as follows:

$$t_k = \frac{\sqrt{3} \mathbf{u}_{ref} T_{sw}}{U_{dc}} \cdot \sin \left(\frac{2\pi}{3} - \theta - \frac{n-1}{3} \pi \right) \quad (3.46a)$$

$$t_m = \frac{\sqrt{3} \mathbf{u}_{ref} T_{sw}}{U_{dc}} \cdot \sin \left(\theta - \frac{n-1}{3} \pi \right) \quad (3.46b)$$

$$t_0 = T_{sw} - t_k - t_m \quad (3.46c)$$

where $n = 1, 2, \dots, 6$ is the sector number.

Each sector has symmetrical pulse pattern waveforms according to the sequence of switching states. As an example, for $\mathbf{u}_{ref}$ (sector I) shown in Fig. 3.32a, the pulse pattern waveform for each switching state is shown in Fig. 3.32b. The modulation signal, m , is obtained by comparing a triangular carrier signal to the PWM pulse of each switching. Finally, a detailed control scheme of this current controller is shown in Fig. 3.33.

Figure 3.33 Block diagram of the VOC current controller.

Nonlinear Vector Current Source (NLVCS)

Ramirez et al. (2012) proposed a spatial hysteresis current source control based on the electrical equations of a grid-connected three-phase converter. This technique combines the advantages, as mentioned earlier, of both nonlinear current controllers and SVM. In this simple case, frequency and voltage amplitudes are constant, and the whole system (inputs and outputs) is electrical.

In the case of point absorbers with a direct drive PTO system, the translator moves upward and downward and, therefore, the voltage alternates between the positive and negative sequences. For this reason, the current controller must be able to adapt to both electrical sequences, positive and negative. Thus, this work proposes an improvement in the strategy proposed by Ramirez et al. (2012) to take into account both sequences. Besides, the calculations performed by the NLVCS could be applied to a control based on the dynamic electromechanical equations of the LPMSG (unlike Ramirez et al., 2012), i.e., to control the mechanical force and the magnetic field of the machine from the

stator currents. This new NLVCS allows performing a maximum power point tracking technique, magnetic field weakening, etc. Moreover, it is robust to parameter variations in the electrical generator. This technique is also applicable to rotary machines without changes.

A simplified scheme of the NLVCS is shown in Fig. 3.34. The measured three-phase current is converted into $\alpha\beta$ stationary reference frame and compared to reference currents also in $\alpha\beta$ stationary reference frame. These error signals, along with the dc voltage, are used by the new NLVSC algorithm to calculate the voltages that must be set by the VSC. This voltage is obtained by SVM (the same modulation technique used in SVPWMCC). The NLVSC algorithm is detailed below.

Figure 3.34 Principle of operation of the hysteresis-band current control.

The proposed algorithm is based on the equivalent circuit of the generator-side converter shown in Fig. 3.30 using the space-vector approach. Firstly, the algorithm calculates the direction of the space-vector current increment necessary to get the desired current value in the real generator at the end of the next control cycle, as shown in Fig. 3.35.

Thus, the current increment, $\Delta \mathbf{i}$, is obtained by comparing the reference current vector, $(i_{\alpha \text{ref}}, i_{\beta \text{ref}})$, and the measured current vector, (i_{α}, i_{β}) , both of them expressed in the $\alpha\beta$ stationary reference frame:

$$\Delta i_{\alpha} = i_{\alpha \text{ref}} - i_{\alpha} \quad (3.47)$$

$$\Delta i_{\beta} = i_{\beta \text{ref}} - i_{\beta} \quad (3.48)$$

Next, taking into account the actual value of the generator voltage, $\mathbf{u}$, the voltage vector to be applied by the inverter, $\mathbf{u}_c$, to get the desired $\Delta \mathbf{i}$, is calculated as follows (neglecting any resistive term):

$$\mathbf{u}_c \approx L \frac{d\mathbf{i}}{dt} + \mathbf{u} \quad (3.49)$$

For each current vector location, there are four possible current increments, depending on

Figure 3.35 Calculation of the necessary Δi and the corresponding u_c .

the sign of its coordinates, so there are four possible solutions for u_c . Its value is calculated in each program cycle by using the generator terminal measurements (Fig. 3.36).

The maximum voltage vector, u_c , describes a circumference of radius $R = 0.866 \cdot U_{dc}$. Therefore, u_c to be generated in the next program cycle can be obtained by calculating the intersection between a straight line with the desired slope, di/dt , and the circumference described by u_c , Fig. 3.36.

The algorithm implementation was carried out by studying solutions in the vertical axis, β , and the location of vectors u and Δi . It can be verified that, when $\Delta i_\beta > 0$, the solution for the intersection of the straight line $L di/dt$ and the circumference described by the converter voltage vector is:

$$u_{c\beta} = \frac{q + \sqrt{q^2 - (1 + d^2) \cdot (q^2 - d^2 R^2)}}{1 - d^2} \quad (3.50)$$

Likewise, when $\Delta i_\beta < 0$, the solution is always:

$$u_{c\beta} = \frac{q - \sqrt{q^2 - (1 + d^2) \cdot (q^2 - d^2 R^2)}}{1 - d^2} \quad (3.51)$$

where q is the intersection point of the straight line and the vertical axis, d the line's slope and R the radius of the circumference. Finally, $u_{c\alpha}$ is obtained from the then known line equation as follows:

$$u_{c\alpha} = \frac{u_{c\beta} - q}{d} \quad (3.52)$$

By using this algorithm, it is possible to obtain the correct solution by calculating a

Figure 3.36 There are four main possibilities for $\Delta \mathbf{i}$ in each program cycle.

single square root in each program cycle, which leads to a considerable computation time-saving in the generator control. Also, it provides a more constant cycle time, which is an advantage for predicting the total computation time of the control program, something fundamental when choosing the ADC sampling period and the SVM PWM cycle time.

The duty cycle of the PWM within the SVM is not corrected in all the control cycles since it depends on whether the increment of current space vector exceeds a particular space hysteresis area around the tip of the electric current vector in each program cycle, as shown in Fig. 3.37.

Figure 3.37 Space hysteresis area around the current space vector. Two possible cases: $|\Delta i| > \epsilon$ and $|\Delta i| < \epsilon$ in each program cycle.

Note that in Fig. 3.30 the space-vector of the generator voltage, $\mathbf{u}$, is changed by the emf, $\mathbf{e}$, the resulting model is similar to the electrical machine model (Novotny and Lipo, 1996). So, the field-oriented control (FOC) could also be performed by the new NLVCS algorithm. The only difference is that each cycle program the emf must be estimated by the measurement of the position of the rotor (or translator). However, this approach is beyond the scope of this work, and it is not considered.

3.5.2 Low-Pass Filter

The voltage at the generator terminals is pulsed due to the switching of the converter. As the voltage measurements are used to generate the current reference signals, the voltage signals must be filtered to eliminate the high-frequency harmonic components. For this purpose, a low-pass filter is used, as shown in Fig. 3.26. Within a range of different filter types, this work evaluates two of them:

- A first order filter:

$$H(s) = \frac{1}{\tau \cdot s + 1} \quad (3.53)$$

where s is the Laplace operator, $\tau = \frac{1}{2\pi f_c}$, f_c being the cutoff frequency.

- A second order filter:

$$H(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n \cdot s + \omega_n^2} \quad (3.54)$$

where $\omega_n = 2\pi f_n$, f_n is the natural frequency, and ζ is the damping ratio. In the case of low-pass filter, the cutoff frequency coincides with the natural frequency.

3.5.3 Switching Losses

The total switching losses of an IGBT-module are associated with both IGBT and diode switching losses. The IGBT switching losses are the sum of all turn-on and turn-off energies described as $E = f(I_C, T_{jv})$, where I_C is the collector current and T_{jv} is the junction temperature of the module. On the other hand, in the case of the diode, only

the recovery energy counts and it is described as $E_{rec} = f(I_F, T_{jv})$, where I_F is the dc forward current. Usually, the component datasheet provides these energy relationships for a given nominal collector-emitter voltage, U_{CE} , such as the ones shown in Fig. 3.38. Therefore, according to Infineon Technologies AG (2015), the total switching losses can be calculated as:

$$P_{sw} = (E_{on} + E_{off}) \cdot f_{sw} + E_{rec} \cdot f_{sw} \quad (3.55)$$

Figure 3.38 Typical curves of inverter switching losses for $U_{CE} = 1200$ V and $T_{jv} = 125^\circ\text{C}$ (IGBT-modules FF100R12KS4, Infineon Technologies AG, 2013).

3.6 Current Controller Evaluation

In this section, the WEC system in Fig. 3.26 has been modelled in Simulink/Matlab with the three different techniques of current controller presented in Section 3.5.1. The performance of each of these current controllers implemented with the WEC system is evaluated from different points of view, such as energy efficiency and dynamic response, to get insight of their suitability for this particular application.

3.6.1 Simulation Parameters

The system is simulated with a 30-minute wave data time series (WDTS) mathematically generated from the modified two-parameter Pierson-Moskowitz spectrum (The Specialist Committee on Waves (chaired by C. T. Stansberg), 2002) whose statistical parameters are the significant height $H_s = 2$ m and the energy period $T_e = 8$ s. For these statistical parameters, the optimal resistance associated is $R_{op} = 1.45 \Omega$. The inductance of the series inductor filter between the linear PMSG and the boost-rectifier is $L = 40$ mH. The HBCC has the hysteresis band set to $\beta = 4$ A and both SVPWMCC and NLVCS have the switching frequency, f_{sw} , set to 2.5 kHz, thereby the waveforms are completely comparable and have similar behaviours. In addition, the boost rectifier is composed of

an IGBT module whose nominal values are: collector-emitter voltage $U_{CE} = 1200$ V and collector current $I_C = 100$ A.

3.6.2 Energy Efficiency Assessment

First, to introduce a base value for comparison, the ideal model (Fig. 3.17) has been simulated, obtaining an average power delivered to the resistive load, R_{load} , of 2.54 kW. Now, let us consider the system shown in Fig. 3.26, implemented with a first-order low-pass filter. The current obtained for each current controller is compared to the ideal current (Fig. 3.17) as shown in Fig. 3.39 for the cutoff frequencies 20 and 200 Hz. It is expected that the more frequency harmonic components are filtered, the more phase shift is introduced. This statement can be verified in Fig. 3.39 for all of the three current controllers, where, for $f_c = 20$ Hz, the current presents a higher phase shift than for $f_c = 200$ Hz. Besides, for higher wave frequencies, a delay is introduced due to the filter time response. On the other hand, from the current waveform quality point of view, the lower cutoff frequency, the better the current quality, except for the HBCC case. In Fig. 3.39a, in which the filter provokes a reduction of the effective hysteresis bandwidth that will be explained later.

Similarly, Fig. 3.40 shows the ideal current compared with the current obtained from the system in Fig. 3.26 for the case of the second-order low-pass filter. It can be seen that there is no noticeable difference in the current quality for all of the implemented controllers. Note this time that, in the case of HBCC (Fig. 3.40a), the ripple is similar in both waveforms, thus making use of the whole the hysteresis bandwidth.

To evaluate the WEC system from the energy conversion efficiency point of view, the system showed in Fig. 3.26 has been simulated with first-order and second-order low-pass filters whose cutoff frequencies range from 20 Hz to 300 Hz. Fig. 3.41 shows the average power delivered to the grid-side converter, taking into account the converter switching losses (Eq. 3.55) in 30 minutes, for the three current controllers under study.

By comparing the WEC systems with different filters, preference for the second-order filter is clear. As the proposed strategy to maximise the generated power does not introduce stability issues to the control, a second-order filter would also not increase the complexity of the control system. However, both cases have been analysed below.

In the case of the first-order filter, the SVPWMCC shows a higher level of injected power compared to the other controllers for $f_c > 50$ Hz. The nonlinear controllers are more sensitive to the cutoff frequency; after a frequency threshold, the higher the cutoff frequency, the lower the PTO power. This decrease is more pronounced in the case of HBCC than NLVCS. The fact that the current reference signal presents more harmonic oscillation for higher cutoff frequencies results in an oscillating hysteresis band, as shown in Fig. 3.42, that, in turn, increases the switching frequency and also the switching losses.

Figure 3.39 Line currents in phase- a for the ideal system and for the resistance emulation system with a first-order low-pass filter.

This effect can also be verified in Fig. 3.43, where the boxplot graphic shows the distribution of the switching frequencies with respect to each cutoff frequency. Outliers are omitted as they do not provide any relevant information. Box-and-whisker diagrams on the left of each pair represent the HBCC switching frequency for the first-order filter and the right ones, for the second-order filter. For $f_c = 200$ Hz, the converter has higher switching frequency, about 10 kHz. This is the reason why the current quality is as good as for $f_c = 20$ Hz in Fig. 3.39a, even presenting smaller ripple. Note that the hysteresis

Figure 3.40 Line currents in phase- a for the ideal system and for the resistance emulation system with a second-order low-pass filter.

band is less than 4 A due to the variable band. Lower cutoff frequency yields a cleaner voltage signal, and, therefore, the switching frequency also decreases. Besides, it is more concentrated around one value.

In case of second-order filter, the power delivered to the grid-side converter varies with the natural frequency of the filter (or cutoff frequency), as shown in Fig. 3.41. In general, the three controllers provide almost the same power. In this case, the HBCC presents an almost constant switching frequency. Therefore, the delivered power decreases for lower cutoff frequency due to the phase shift introduced by the filter only, that results

Figure 3.41 Average power delivered to the grid-side converter with switching losses for two implementation cases: first-order and second-order filters.

Figure 3.42 Switching characteristics of nonlinear current sources. First-order filter ($f_c = 200$ Hz).

Figure 3.43 Boxplot of the switching frequencies for first and second-order filters.

in incorrect current reference signals. Despite keeping the right proportionality between voltage and current, it makes them be out of phase. Cutoff frequencies higher than 200 Hz do not provide significantly different results. Beyond this value, for the three controllers, the maximum deviation with respect to the power delivered by the ideal model is about 2%.

3.6.3 Dynamic Response Assessment

Because of the oscillating nature of waves, the speed and the force transmitted from the buoy to the PTO undergo large and steep fluctuations. Therefore, the control system is required to have a fast response to fulfil its objective whatever it is.

To verify the dynamic response of the WEC to abrupt variations in the control system, let us consider the resistive load, ideal case (Fig. 3.17), and emulated resistance (Fig. 3.26) not constant over time. A simulation has been carried out in which the value of both resistances alternates between $0.2\ \Omega$ and $10\ \Omega$ every 2 seconds. As a consequence, in the ideal WEC system, different R_{load} leads to a change in the trend of the current waveform. Similarly, a new value of R_{em} results in a change in the trend of the reference current of the control system.

Fig. 3.44 shows the current at the LPMSG terminals, where the change from $0.2\ \Omega$ to $10\ \Omega$ can be clearly seen at instant $t = 4\ \text{s}$, for the ideal case and emulated resistance changes for each current control. The nonlinear current controllers present a faster response to this change. Both HBCC and NLVCS, generate currents with the same waveform as the ideal one. On the contrary, the PI-based controller, SVPWMCC, presents a slower dynamic response, and it takes time to establish the expected current. This current settling time occurs not only in the face of variations in the control setpoints but, also, during a more pronounced wave, as shown in the first third of the period in Fig. 3.44.

Figure 3.44 Dynamic performance of the current controllers under the proposed strategy to maximise the generated power by the WEC.

3.7 Simulation and Results

The proposed method to maximise the captured power is based on the adjustment of the damping coefficient of the electrical generator applied to the floating system by setting an equivalent resistance as seen from its terminals. An optimal damping can be obtained through an optimal load resistance. Thereby, both optimal resistance and damping have been mapped by exciting the system with regular waves for several combinations of wave period and height (Figs. 3.21a and 3.23).

According to the results presented in Section 3.6, all of the controllers have similar performance with the second-order filter which, in turn, has superior performance than the first-order filter. Also, as previously commented, the higher the cutoff frequency, the worse the waveform. On the other hand, the lower the cutoff frequency, the higher the phase-shift of the waveform. So, by setting $f_c = 200$ Hz, an acceptable waveform and excellent performance are assured. Finally, in overall comparison, the NLVCS seems to present slight advantages over others. For example, NLVCS shares the same functional characteristics of the HBCC and, besides, presents a constant switching frequency. Concerning SVPWMCC, NLVCS has a better dynamic response.

Therefore, the WEC system is implemented with NLVCS and the second-order filter. Some system simulations are presented in this section, and they are intended to illustrate the applicability of the implemented electronic circuit and to verify that the proposed control strategy achieves the maximum captured power. For that, a set of different WDTs of 30 minutes each have been mathematically generated by using the ITTC spectrum (The Specialist Committee on Waves (chaired by C. T. Stansberg), 2002). The statistical parameters used are the energy period, T_e , and the significant wave height, H_s . These WDTs have been grouped as:

- $H_s = 1$ m and $T_e = 8$ s: WDTs1, WDTs2, WDTs3, and WDTs4.
- $H_s = 2$ m and $T_e = 8$ s: WDTs5, WDTs6, WDTs7, and WDTs8.
- $H_s = 1$ m and $T_e = 15$ s: WDTs9, WDTs10, WDTs11, and WDTs12.
- $H_s = 2$ m and $T_e = 15$ s: WDTs13, WDTs14, WDTs15, and WDTs16.

3.7.1 WEC Operation

To assess the proper operation of the boost rectifier, the system illustrated in Fig. 3.26 was simulated with the WDTs5 ($T_e = 8$ s and $H_s = 2$ m). For this specific sea state, according to Section 3.4.1, the load seen from the generator terminals is set to 1.45Ω , which, in turn, is associated with the optimal damping. The electrical generator is connected to the boost rectifier through a cable in series with a filter whose equivalent resistance and inductance are $R_c = 0.54 \Omega$ and $L = 40$ mH, respectively.

The main waveforms resulting from these simulations are presented in Fig. 3.45. The upper figure shows a window time of the PTO voltage and current in phase-*a*. The red frame applies a zoom in on the time axis of the waveforms. It can be seen that the control works perfectly, achieving its goal of keeping voltage and current in phase. The lower figure shows the PTO power profile, where the dashed line is the average power ($P_{pto} = 2.54 \text{ kW}$). Note that in very short time intervals, the PTO power is over its rated value (10 kW). This scenario is acceptable since the generator is designed to operate up to several times its rated power in a short-term overload (Cruz, 2007).

Figure 3.45 Voltage, current and power waveforms of the PTO.

To show the performance of the emulation resistance approach to control the damping coefficient and, thus, to achieve the maximum power point, the results obtained in the previous simulation are contrasted with the ideal case (Fig. 3.17). Fig. 3.46 shows both PTO powers, and it is observed that the waveforms are practically the same. In some occasions, there might be a very tiny phase-shift between them because of the filtered measured voltage, as explained previously. Similarly to the comparison between currents, another difference is the ripple due to switching in the boost rectifier.

3.7.2 Performance of the Proposed Control Strategy

Another group of simulations were carried out to verify the effectiveness of the strategy proposed in this work. For this purpose, simulations use the sixteen WDTs of irregular waves with a duration of 1800 seconds, sorted into the aforementioned four groups according to their statistical parameters. Each wave time series was simulated with the optimal resistance, R_{op} , and four other resistances (R_1 , R_2 , R_3 , and R_4) with values close

Figure 3.46 Comparison between the PTO power of ideal and emulated resistance systems.

to R_{op} . Table 3.2 each wave time series with the emulated resistances (first line) and their respective average power ratios (second line). Note that R_2 and R_3 have minimal variation respect to the optimal resistance. It is verified that the maximum captured power is concentrated in R_2 , R_{op} , and R_3 . The deviation of the PTO power for the optimal resistance does not exceed 5%.

To examine the behaviour of the P_{ratio} - R_{em} curves for irregular waves, one WDTS from each group has been selected: WDTS1, WDTS5, WDTS9 and WDTS13. Each time series has been simulated with different values of emulated resistance, and giving as a result the P_{ratio} - R_{em} curves shown in Fig. 3.47, similar to Fig. 3.24, where the black filled circle represents the respective optimal resistance, R_{op} . Also, the P_{ratio} - R_{em} curves have a similar shape as Fig. 3.20b, in the sense that the PTO power increases with the emulated resistance up to a maximum point and, then, it starts to decrease. It can be noted that the P_{ratio} - R_{em} curves are quite different in P_{ratio} values, because regular waves carry more energy per meter of wavefront. However, for both regular and irregular waves, the optimal resistances are almost coincident. Thus, it is possible to conclude that mapping the maximum power point for a combination of wave heights and periods for regular waves and applying the optimal values of the load seen from the electrical generator to irregular waves improves the efficiency of the system compared to almost any other electrical load.

3.7.3 Simulations with Real WDTS

In this part, the system was simulated with four WDTS taken from real sea measurements in order to validate the results obtained from the numerically generated WDTS in the previous section. For that, the measured data have been collected from a buoy available in *CDIP Historic Data*. Information about the wave data is presented in Table 3.3, where the buoy reference (station), the location in terms of latitude and longitude, the date and hour of measurement, and the significant wave height and energy period are

Table 3.2 Comparison of the power ratio for irregular wave profiles with different values of emulated resistance.

Wave Data Time Series	H_s (m)	T_e (s)	Emulated Resistances (Ω)				
			R_1	R_2	R_{op}	R_3	R_4
			$P_{ratio}(\%)$				
			P_{R_1}	P_{R_2}	$P_{R_{op}}$	P_{R_3}	P_{R_4}
WDTS1	1	8	0.50	0.90	1.00	1.10	2.00
			4.58	6.21	6.37	6.46	6.08
WDTS2	1	8	0.50	0.90	1.00	1.10	2.00
			4.15	5.75	5.91	6.00	5.64
WDTS3	1	8	0.50	0.90	1.00	1.10	2.00
			4.51	6.10	6.23	6.32	5.95
WDTS4	1	8	0.50	0.90	1.00	1.10	2.00
			3.99	5.53	5.69	5.79	5.52
WDTS5	2	8	0.50	1.30	1.45	1.50	2.00
			3.42	5.20	5.24	5.24	5.08
WDTS6	2	8	0.50	1.30	1.45	1.50	2.00
			3.35	5.01	5.04	5.03	4.85
WDTS7	2	8	0.50	1.30	1.45	1.50	2.00
			3.44	5.28	5.31	5.31	5.13
WDTS8	2	8	0.50	1.30	1.45	1.50	2.00
			3.45	5.17	5.20	5.20	5.03
WDTS9	1	15	0.10	0.50	0.65	0.80	1.50
			0.13	1.19	1.33	1.39	1.26
WDTS10	1	15	0.10	0.50	0.65	0.80	1.50
			0.15	1.19	1.32	1.37	1.25
WDTS11	1	15	0.10	0.50	0.65	0.80	1.50
			0.03	0.85	0.97	1.02	0.94
WDTS12	1	15	0.10	0.50	0.65	0.80	1.50
			0.15	1.20	1.33	1.38	1.25
WDTS13	2	15	0.10	0.70	0.80	0.90	1.50
			0.29	1.17	1.19	1.20	1.10
WDTS14	2	15	0.10	0.70	0.80	0.90	1.50
			0.29	1.19	1.22	1.22	1.14
WDTS15	2	15	0.10	0.70	0.80	0.90	1.50
			0.26	1.12	1.14	1.14	1.06
WDTS16	2	15	0.10	0.70	0.80	0.90	1.50
			0.33	1.24	1.26	1.26	1.16

Figure 3.47 $P_{\text{ratio}}-R_{\text{em}}$ curves for WDTs1, WDTs5, WDTs9 and WDTs13.

shown. The spectrum of each real WDTs is calculated according to McCormick, 2009 for 1800 seconds (Fig. 3.48). The statistical parameters of the waves are directly obtained from these spectra. The selected station is 42-meter water depth (a reasonable depth for this WEC type). The selected WDTs present sea states close to the ones presented in Table 3.2 and, also, are sampled at 1.28 Hz.

Table 3.3 Real WDTs information.

Buoy (Station)	Location		Wave Data Time Series	Date	Hour	H_s (m)	T_e (s)
	Latitude	Longitude					
036	46°51.404'(N)	124°14.630'(W)	WDTs17	2019-09-09	00:03	1.02	7.99
			WDTs18	2018-07-12	16:03	2.11	8.06
			WDTs19	2018-10-17	03:33	1.05	15.09
			WDTs20	2019-01-01	20:03	1.92	15.17

The methodology used to assess the results is the same as in the previous section. Fig. 3.49 shows the average power ratio for each WDTs (17, 18, 19, and 20) obtained from the system set with different emulated resistances, including the optimal resistance, R_{op} (black filled circle). It can be seen that the maximum values of the average power ratio are concentrated around R_{op} , as in the case of generated waves. The maximum deviation of the maximum average power ratio with respect to the optimal resistance does not exceed 5%, similar to the case of mathematically generated waves. Additionally, it can be seen that the $P_{\text{ratio}}-R_{\text{em}}$ curves of the real WDTs are similar to the correspondent

Figure 3.48 Real WDTs spectra.

$P_{\text{ratio}}-R_{\text{em}}$ curves of Fig. 3.24. On the other hand, despite the fact that the sea states of the real WDTs are slightly different from those of the synthetic WDTs, the $P_{\text{ratio}}-R_{\text{em}}$ curves are very similar both in magnitude and in shape, showing the validity of the proposed strategy for real seas.

Figure 3.49 $P_{\text{ratio}}-R_{\text{em}}$ curves for WDTs17, WDTs18, WDTs19 and WDTs20.

Wave-to-Wire Model: Grid-Connected Application

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according to the Ocean Energy Glossary of the International Energy Agency (IEA) OES Agreement (Wave Energy Centre, 2007) a wave-to-wire model is defined as: “mathematical model in the time domain that simulates the energy conversion chain”. Although many works in the literature are focused on the energy source conversion and its control system, interest in electrical grid connection has been experiencing a substantial increase. The U.S. Department of Energy’s Marine and Hydrokinetic Technology Database (*Marine and Hydrokinetic Technology Database*) provides up-to-date information on WEC projects, where it is possible to verify that a few projects are already deployed. This growing interest has been leading to the emergence of test sites, where some companies provide test areas including connection to the electrical grid, such as SEM-REV Sea Test Site (<https://sem-rev.ec-nantes.fr>), France, BIMEP (<http://www.waveenergybasquecountry.com>), Spain, and Wave Hub (<https://www.wavehub.co.uk>), United Kingdom.

Penalba and Ringwood (2016) and Wang et al. (2018) present interesting lists with a review of wave-to-wire works for different WEC technologies and control systems for both, generator- and grid-side converters. The wave-to-wire model validations are mostly by simulation, although a few ones can be found by test lab, and, barely, by offshore experimental tests.

Despite the full conversion chain, from WEC to the electrical grid, most of the studies seem to be centred on the generator-side control, where it is possible to optimise the power generation, and the grid-side model is available only for delivering the generated power (Benreguig et al., 2019; Garcia-Rosa et al., 2013; Kelly et al., 2015). However, the impact on the electrical grid due to the fluctuating nature of the power generated by the WEC has to be identified and studied to allow its connection in a large scale. Some of these impacts to be studied require a detailed wave-to-wire model, as the case of harmonic voltage

distortion introduced by power electronic converters. Other studies, such as slow and fast voltage fluctuation (VF), can be carried out using quasi-static simulations and, therefore, the average wave-to-wire model can be used. This model performs longer simulations with much less computational effort, and, therefore, in a considerably shorter time. For these reasons, in this chapter, on the one hand, the wave-to-wire model in detail and its average equivalent model are developed. Moreover, both models are contrasted in order to validate the use of the average model.

On the other hand, the application of the wave-to-wire model connected to an electrical grid is also studied in this chapter. For this reason, let us, firstly, state the general scenario of renewable energies and the electrical grid.

As a consequence of the increasing penetration of renewable energy, the electrical network structure has been changing, migrating from a centralised generation model, where the power is generated in a few and large power plants, far from the loads, to a hybrid model with distributed generation (DG), where many smaller plants are closer to the loads. This proximity between generation and consumption brings some benefits such as the increasing of electric system reliability and the reduction of losses in distribution and transmission systems (Momoh, Meliopoulos, and Saint, 2012). DG systems can also aid in the grid support with ancillary services and in environmental issues, since most of these primary sources are non-polluting, as photovoltaic, wind, hydro and wave generation. The latter technology, unlike the others, is still under development, and its impact on the electrical grid and its compliance with the connection requirements are the subjects of current research.

DG systems are mostly connected to distribution grids, which are becoming a little more independent from the transmission networks. In this context, the question arises: is the distribution grid ready for DG systems? There is still much work to be done, since DG covers different types of primary resources, and each one has its singularities, not only in technical issues but also in the grid codes that define the connection requirements. As stated by Braun et al. (2012), despite the advances in the development of adequate grid codes, they are still delayed with respect to the continuous increase in DG penetration levels. Furthermore, the different development stage of DG in different countries results in different grid connection requirements around the world. A review focused on weak power grids was presented by Etxegarai et al. (2015), where the differences between the grid codes of interconnected and weak power systems are highlighted. From the point of view of wave energy, as the technical development is in an early stage, there are few grid codes specifically addressed to it. However, wind energy grid codes can be applied to wave energy with minor changes due to their similarities, as shown by Santos-Mugica et al. (2014). In addition, O'Sullivan, Dalton, and Lewis (2010) highlight that grid codes have a strong influence on the WEC's design and, therefore, on its cost.

However, taking advantage of the positive impacts of DG on the electrical grid is quite

a hard task in practice, and some issues arise, as voltage regulation, flicker, impact on short-circuit levels, etc. (Barker and De Mello, 2000). In the case of WEC, an important point concerning power quality is the fluctuating characteristic of the generated power. The injection of such power into the grid results in voltage fluctuations. For medium and large wave power plants, these fluctuations are naturally smoothed, and they could be seen as flicker. For example, Blavette et al. (2014) assess the flicker level of a wave power plant with an installed power of 19.4 MW, and show that, for that specific power plant topology, there are some test cases in which the flicker levels exceed the limits recommended by IEC 61400-21 (2008). Thus, a mitigation method is required. On the other hand, in small power plants, voltage fluctuations are smoothed to a lesser degree. Some approaches to mitigate them are presented by Lafoz, Blanco, and Ramirez (2011), Moreno-Torres et al. (2015), and Santos-Mugica et al. (2014) based on controlling the PTO system and/or using electrical storage systems.

In this thesis, the possibility of achieving the mitigation of voltage fluctuations by reactive power compensation is explored. Particularly, a control strategy applicable in real unbalanced distribution grids is developed. It seeks the mitigation of voltage fluctuations by using the WEC's own reactive resources. To carry out the proposed strategy, it is necessary to know how the unbalanced grid responds to the active and reactive powers from the WEC. To do so, an unbalanced three-phase load flow, based on (Mayordomo et al., 2002), that is able to take into account the unbalance of loads, the asymmetry due to un-transposed lines and the presence of single-phase and two-phase lines has been implemented. The performance of the proposed strategy is assessed, with a continuation load flow (Ajjarapu and Christy, 1992), in a test distribution grid with a wave-to-wire model of a WEC.

4.1 Grid-Side Model

The grid-side converter of the WEC is modelled in a way similar to the generator-side converter (Fig. 3.25). However, now, the power flows from the converter to the grid, and the generator-side converter is initially modelled as a power source, as shown in Fig. 4.1. Plus, the ac side is represented by its spatial-phasor model, where: $\mathbf{u}$ and $\mathbf{e}$ are the voltage at the converter terminals and the grid voltage; $\mathbf{i}$ is the current at the PCC (point of common coupling); and R and L are the resistance and the inductance of the filter's converter. The power source is connected to the converter by a capacitive dc-link, where: i_{ext} is the dc current injected by the power source; i_C is the capacitor current; i_{dc} the current injected to the converter; and, u_{dc} is the capacitor voltage.

An essential step to model the grid-side converter is to determine its filter. Its function is to filter the high-frequency harmonics of the current injected by the converter (Ahmed, Finney, and Williams, 2007). In the literature, one can find different filter topologies.

Figure 4.1 Grid-side converter scheme. ac side represented in spatial phasors.

Beres et al. (2015) present the most common passive filters used for a grid-connected converter, such as L -filter, low-pass LC -filter, low-pass LCL -filter and high-order filter models, mainly focused on the design of filters of second and higher orders.

The filter design optimisation is beyond the scope of this work. For this reason, the L -filter (first-order) is chosen due to its straightforward implementation, and it is enough for the wave-to-wire model presented here. However, future work on filter design may be carried out for connections to weak grids, since a high grid impedance, may lead to system instability.

The ac side can be schematically represented by the single-phase circuit in Fig. 4.2, where $\underline{Z}_f$ is the filter impedance. The L -filter has a minimal resistance which is usually neglected. $\underline{U}$ is the inverter phasor voltage; therefore, it is controllable. As a consequence, the power flow can also be controlled as follows:

Figure 4.2 ac-side equivalent circuit.

$$P = \frac{U \cdot E}{X_L} \sin \delta \quad (4.1a)$$

$$Q = \frac{U^2 - UE \cos \delta}{X_L} \quad (4.1b)$$

where $\delta = \delta_u - \delta_e$. As pointed by Lasseter, Tomsovic, and Piagi (2000), X_L must be roughly sized taking into account the limits on U , defined by the current controller used,

and the angle δ . In this work, the SVPWM modulation is applied and, thus:

$$U_{\max} = \frac{U_{\text{dc}}}{\sqrt{3}\sqrt{2}} \quad (4.2)$$

On the other hand, choosing $\delta_{\max} = 30^\circ$, linear power control is assured.

The dc-voltage used in the converters is 1000 V. Let us consider the grid voltage 400 V, so $E = 230.94$ V, and $P_{\max} = 100$ kW. As a result, the filter inductance is $L = 1.5$ mH. Additionally, a resistance, $R = 2.5$ m Ω , is considered.

4.1.1 Inner Current Control Loop

The current $\mathbf{i}$ injected into the grid can be controlled by regulating the converter voltage $\mathbf{u}$. As the approach used for the grid-connected converter is the VOC, the mathematical model of the system is quite similar to the one presented in Section 3.5.1 (SVPWMCC). Instead of the terminal voltage of the electrical machine, the d -axis is aligned to the space phasor of the grid voltage. Its angle can be tracked by a PLL (as also commented in Section 3.5.1) and it replaces the rotor angle. Therefore, the dynamics of the ac side in space phasor is described by:

$$\mathbf{u} - \mathbf{e} = L \frac{d\mathbf{i}}{dt} + R\mathbf{i} \quad (4.3)$$

Eq. (4.3) can be decoupled in d and q components as follows:

$$u_d - e_d = L \frac{di_d}{dt} + Ri_d - L\omega i_q \quad (4.4a)$$

$$u_q - e_q = L \frac{di_q}{dt} + Ri_q + L\omega i_d \quad (4.4b)$$

where ω is the grid angular frequency. These equations, in Laplace domain, are written as:

$$\frac{I_d}{U_{d1}} = \frac{1}{Ls + R} = G_{pg} \quad (4.5a)$$

$$\frac{I_q}{U_{q1}} = \frac{1}{Ls + R} = G_{pg} \quad (4.5b)$$

which represent the ac-side plant, $G_{pg}(s)$, where:

$$U_{d1} = U_d - E_d + L\omega I_q \quad (4.6a)$$

$$U_{q1} = U_q - E_q - L\omega I_d \quad (4.6b)$$

On the other hand, the converter voltage can be obtained by Eqs. (4.4a) and (4.4b), where a PI controller is responsible for regulating the current terms (not the coupling

term). So the converter voltage can be realised by several modulation techniques, including those developed in Section 3.5.1. Adopting the SVPWM, the control system is identical to the one shown in Fig. 3.33. The average model avoids the switching dynamics; thus, the output signal of the PI controller, u_{1d} or u_{1q} , corrected by the feedforward signals, provides the converter voltage directly. This last one is the input to the grid plant. Fig. 4.3 shows the control block of the average model of the inner control loop. Note that Fig. 3.33 is represented by the controller blocks and the feedforward signals in Fig. 4.3.

Figure 4.3 Current control loop with feedforward contributions.

The closed-loop current control system for the grid-connected converter is shown in Fig. 4.4. Note that the coupling terms and the feedforward voltages are disturbance inputs, and therefore they are not considered in the controller tuning. Also, the PI controller is the same for both dq -currents.

Figure 4.4 Current control loop.

The PI controller can be expressed as:

$$G_{ci}(s) = \frac{k_{Pi}s + k_{Ii}}{s} \quad (4.7)$$

where k_{Pi} and k_{Ii} are proportional and integral gains, respectively. So, the open current loop gain is:

$$\ell_i(s) = G_{ci}(s) \cdot G_{pg}(s) = \frac{k_{Pi}s + k_{Ii}}{s} \cdot \frac{1}{Ls + R} = \frac{k_{Pi}}{L} \cdot \frac{s + k_{Ii}/k_{Pi}}{s} \cdot \frac{1}{s + R/L} \quad (4.8)$$

The plant pole is close to the origin and, therefore, the grid plant presents a slow response. Nevertheless, this pole can be canceled by the controller zero with:

$$\frac{k_{Ii}}{k_{Pi}} = \frac{R}{L} \quad (4.9)$$

This pole cancellation results in a new open-loop gain $\ell_i(s) = k_{Pi}/(Ls)$, which leads to a closed-loop first order model:

$$G_i(s) = \frac{1}{\tau_i s + 1} \quad (4.10)$$

where $\tau_i = L/k_{Pi}$. according to Yazdani and Iravani (2010), the time constante, τ_i , is selected between 0.5 and 5 ms. Besides, τ_i must assume a value such that the G_i bandwidth ($\omega_{ci} = 1/\tau_i$) is at least 10 times smaller than the switching frequency in rad/s.

To verify the performance of the inner current controller, let us select $\tau_i = 1$ ms for a 5 kHz frequency switching. This time constant results in $k_{Pi} = 1.5005$ and $k_{Ii} = 2.5$. Fig. 4.5 shows the the step response for both dq -currents. Note that the average model can replace the detailed switched circuit as long as there is no interest in switching dynamics.

Figure 4.5 Current control loop responses (detailed switched and average models).

4.1.2 Outer Control Loop

This is a cascade control scheme where an outer loop processes another and much faster control system. In the case of grid-connected VSC, the outer control systems regulate the active and reactive powers and can also control the dc-voltage.

dc-Voltage control

The dc-voltage control is based on the premise that all generated power by the primary source has to be injected into the electrical grid, i.e., $p_{dc} = p_g$. For this goal, the dc-voltage on the capacitor, C , must be kept constant. So, considering the circuit in Fig. 4.1 and neglecting the losses in the filter, the power balance equation is:

$$u_{dc}i_{\text{ext}} - u_{dc}C\frac{du_{dc}}{dt} = \frac{3}{2}(e_d i_d + e_q i_q) \quad (4.11)$$

As the d -axis is aligned to the space phasor of the grid voltage, $e_q = 0$. Consequently:

$$\frac{du_{dc}}{dt} = \frac{1}{C} \left(-\frac{3}{2} \frac{e_d i_d}{u_{dc}} + i_{\text{ext}} \right) \quad (4.12)$$

Transforming the above equation into the Laplace domain:

$$U_{dc}(s) = \frac{1}{C_s} [CK_{dc}I_d(s) + I_{\text{ext}}(s)] \quad (4.13)$$

where:

$$K_{dc} = -\frac{3}{2} \frac{E_d}{U_{dc}C} \quad (4.14)$$

Eq. (4.13) indicates that the dc-voltage can be controlled through the current in the ac-side circuit, considering I_{ext} a disturbance signal. Thus a PI controller, $G_{cu}(s)$, can regulate the current in the form: $I_{d\text{ref}} = G_{cu}(s) \cdot (U_{d\text{ref}} - U_{dc})$. Fig. 4.6 shows the dc-voltage control loop where the dynamics of the inner current loop is modelled by the transfer function $G_i(s)$ (Eq. 4.10). Besides:

$$G_{dc}(s) = \frac{1}{C_s} \quad (4.15)$$

A first-order filter, $G_{fu}(s)$, where τ_{fu} is its time constant, can be introduced in the feedback signal for filtering the measured dc-voltage. It can also be used to improve system stability.

Figure 4.6 dc-voltage control loop.

The PI controller can be expressed as:

$$G_{cu}(s) = \frac{k_{Pu}s + k_{Iu}}{s} = \frac{k_{cu}}{s}(\tau_{cu}s + 1) \quad (4.16)$$

where k_{Pu} and k_{Iu} are proportional and integral gains, respectively; and, also:

$$k_{Iu} = k_{cu} \quad (4.17a)$$

$$k_{Pu} = k_{cu} \cdot \tau_{cu} \quad (4.17b)$$

So, the open dc-voltage loop gain:

$$\begin{aligned} \ell_u(s) &= G_{cu}(s) \cdot G_i(s) \cdot CK_{dc} \cdot G_{dc}(s) \cdot G_{fu}(s) \\ \ell_u(s) &= \frac{k_{cu}}{s}(\tau_{cu}s + 1) \cdot \frac{1}{\tau_i s + 1} \cdot \frac{K_{dc}}{s} \cdot \frac{1}{\tau_{fu}s + 1} \end{aligned} \quad (4.18)$$

According to Miret et al. (2009), the PI controller must be tuned so to provide a slow dc-voltage transient response and high stability margin. So, the symmetrical optimum (SO) method (Åström and Hägglund, 1995) could be applied to tune it, which leads the system to a good performance with a phase margin $\varphi_{Pm} \approx 36^\circ$. However, this would be unacceptable for variations of K_{dc} and very sensitive to disturbance signals. To overcome this issue, Preitl and Precup (1999) propose the extended symmetrical optimum (ESO) method, where it is possible to maximise the phase margin.

In the ESO the open-loop gain can be approximated to:

$$\ell_u(s) = \frac{k_0(\tau_{cu}s + 1)}{s^2(\tau_\Sigma s + 1)} \quad (4.19)$$

where $k_0 = k_{cu}K_{dc}$ and τ_Σ is a time constant corresponding to the sum of parasitic time constants ($\tau_\Sigma = \tau_i + \tau_{fu}$). From the frequency analysis, the crossover frequency (bandwidth, ω_{cu}) and the phase margin are given by:

$$\omega_{cu} = \frac{1}{\sqrt{\beta}\tau_\Sigma} \quad (4.20a)$$

$$\varphi_{Pm} = \arctan\left(\frac{\beta - 1}{2\sqrt{\beta}}\right) \quad (4.20b)$$

where $\beta \in [4, 16]$, i.e., $36^\circ \leq \varphi_{Pm} \leq 60^\circ$. Besides, the filter $G_{fu}(s)$ can be designed to provide a desired $\ell_u(s)$ bandwidth through the time constant τ_Σ . Finally, the tuning relations are:

$$k_{cu} = \frac{1}{\beta\sqrt{\beta}t_\Sigma^2 K_{dc}} \quad (4.21a)$$

$$\tau_{cu} = \beta\tau_\Sigma \quad (4.21b)$$

Reactive Power Control

The reactive power injected into the electrical grid by the grid-connected VSC is given by:

$$q_g = \frac{3}{2} (-e_d i_q + e_q i_d) \quad (4.22)$$

As the d -axis is aligned to the space phasor of the grid voltage, $e_q = 0$, so:

$$q_g = -\frac{3}{2} e_d i_q \quad (4.23)$$

transformed into Laplace domain as $Q_g(s) = K_q I_q(s)$, where $K_q = -3E_d/2$.

Therefore, the current $I_q(s)$ can be regulated by a PI controller to achieve a desired reactive power injected into the grid. The reactive power control loop is shown in Fig. 4.7. The procedure to tune the PI controller $G_{cq}(s)$ is the same one presented in Section 4.1.1 (inner current control loop).

Figure 4.7 Reactive power control loop.

Considering the proportional and integral gains, k_{Pq} and k_{Iq} , let $k_{Pq}/k_{Iq} = \tau_i$. Thus, the closed-loop reactive power control is a first-order system:

$$G_q(s) = \frac{1}{\tau_q s + 1} \quad (4.24)$$

where $\tau_q = 1/(K_q k_{Iq})$. As in the case of dc-voltage control, $G_q(s)$ bandwidth, ω_{cq} , must be much slower than the inner control bandwidth.

Performance of the Outer Control Loop

As previously commented, the outer control loop must have a bandwidth much slower than the inner control loop. In the practical design, the bandwidth for the outer control loop should be not higher than the one-tenth of the inner control bandwidth to avoid transmitting low-frequency oscillations from the dc-link to the reference current (Zarif and Monfared, 2015).

Let us select a phase margin $\varphi_{Pm} = 45^\circ$ and a bandwidth $\omega_{cu} = 50$ rad/s for the dc-link control loop, whose capacitor is $C = 8980 \mu\text{F}$. To achieve this bandwidth, $\tau_\Sigma = 8.3$ ms; hence, the time constant of the feedback filter is $\tau_{fu} = 7.3$ ms, and the PI gains: $k_{Pu} = -0.9165$ and $k_{Iu} = -18.9817$. Fig. 4.8 shows the Bode diagram of the

loop gain for both transfer functions, the original (Eq. 4.18) and the reduced (Eq. 4.19). The crossover frequency and the phase margin for the original transfer function are, respectively: $\omega_{cu} = 50.6$ rad/s and $\varphi_{Pm} = 44.6^\circ$; practically the same as the designed.

Figure 4.8 Bode plot of the dc-voltage loop gain for the original and the approximated transfer functions.

In addition, the closed-loop dc-voltage control system, $G_u(s)$, presents double integrating terms which lead to a significant overshoot. Fig. 4.9 shows its step response where the overshoot is about 40%. This effect could be mitigated by introducing a lag filter to the input signal:

$$G_{f\text{lag}} = \frac{1}{\tau_{cu}s + 1} \quad (4.25)$$

As a result, despite a longer rise time, the settling time is a bit shorter, and the overshoot is almost eliminated.

Regarding the reactive power, there is hardly a sudden change. However, it can be avoided by regulating i_q , as previously shown. Let us select a bandwidth $\omega_{cq} = 10$ rad/s. It is even slower than the dc-voltage control loop bandwidth and, in this way, the reactive power control loop hardly affects the dc-voltage control. So, the PI controller gains are: $k_{Pq} = -2.0412 \times 10^{-5}$ and $k_{Iq} = -0.0204$.

Active and reactive power responses injected into the grid and the dc-voltage for a power step in the VSC are shown in Fig. 4.10. Again, both detailed switched and average responses are presented, and they allow to validate the use of the average model.

Figure 4.9 Step responses for the dc-voltage control loop with and without lag filter implementation.

Figure 4.10 Grid-side converter responses (detailed switched and average models).

4.2 Wave-to-Wire Model

So far, generator-side and grid-side models have been developed separately. However, the wave-to-wire model represents a full WEC system from the wave energy conversion to the power delivered to the electrical grid. Therefore, if each side of the system is correctly modelled, there should be no problems in their connection. The wave-to-wire model is schematically represented in Fig. 4.11, where: the WEC system block represents all the hydrodynamical and mechanical models of the WEC; the generator-side converter represents the LPMSG connected to the VSC; and the grid-side converter represents the grid-connected VSC. Plus, it shows the mechanical and electrical quantities exchanged between each block.

Figure 4.11 Scheme of the wave-to-wire model.

The generator-side converter operating under the resistance emulation approach can be modelled as the ideal case, where the resistance R_{em} is connected to the generator terminals. The voltage equations of the LPMSG connected to R_{em} in the dq -frame can be obtained by combining Eqs. (3.26a), (3.26b), (3.27a) and (3.27b), resulting in the equivalent circuit shown in Fig. 4.12.

Figure 4.12 LPMSG dq -axis equivalent circuits.

Therefore, the active power delivered by the LPMSG to the grid-side converter is given by:

$$p_{pto}(t) = \frac{3}{2} R_{em} [i_d(t)^2 + i_q(t)^2] \quad (4.26)$$

Besides, the voltage equations of the generator-side converter can be represented in the Laplace domain, whose plant is a time-varying transfer function because of R_{em} . Therefore, the generator-side converter can also be modelled as shown in Fig. 4.13, where the back emf is calculated with Eq. (3.23). This model simulates faster than the electrical model, so it fits better with the average model.

Table 4.1 summarises the grid-side converter parameters that have been used throughout this chapter. Moreover, Figs. 4.14 to 4.16 show the simulations of the wave-to-wire models, both detailed switched and average. Finally, the results suggest that the detailed switched model can be replaced by the average model as long as the switching dynamics are not required with no loss of information. The benefit in terms of computational effort is incomparable, and this is an essential point since wave energy simulations are usually long. A 30-minute WDTS is simulated in seconds with the average model, while the

Figure 4.13 Generator-side converter modelled in Laplace domain.

detailed switched model might take several hours (not to mention days, depending on the computer).

Table 4.1 Generator-side converter parameters.

Symbol	Quantity	Value
E	grid voltage (line-to-neutral)	230.94 V
f	grid frequency	50 Hz
S	rated power	100 kVA
f_{sw}	switching frequency	5 kHz
R	filter resistance	2.5 m Ω
L	filter inductance	1.5 mH
C	dc-link capacitance	8980 μ F
U_{dc}	dc-voltage	1000 V
k_{Pi}	proportional current control gain	1.5005
k_{Ii}	integral current control gain	2.5
k_{Pu}	proportional dc-voltage control gain	-0.9165
k_{Iu}	integral dc-voltage control gain	-18.9817
k_{Pq}	proportional reactive control gain	$-2.0412 \cdot 10^{-5}$
k_{Iq}	integral reactive control gain	-0.0204
τ_{cu}	dc-voltage PI time constant	0.0483
τ_{fu}	feedback voltage filter time constant	0.0073

Figure 4.14 Full wave-to-wire model: generator-side responses.

Figure 4.15 Full wave-to-wire model: grid-side responses.

4.3 Voltage Fluctuation

Assuming, in a first approach, a distribution grid as a balanced system, it can be represented by its Thévenin equivalent circuit from the PCC, as shown in Fig. 4.17a. The feeder

Figure 4.16 Full wave-to-wire model: dc-voltage response.

has two buses connected by a series impedance with resistance R and reactance X . The voltage at the beginning of feeder (infinite bus) is $\underline{U}_1$, and the voltage at PCC is $\underline{U}_2$. The WEC is modelled as a PQ-load, i.e., the active and reactive powers injected by the WEC (P_{wec} and Q_{wec}) are independent of voltage. P_{wec} is a nonnegative quantity, while Q_{wec} could be both positive or negative. In addition to the WEC, a load that absorbs constant active and reactive powers (P and Q) is connected to the PCC.

Figure 4.17 (a) Thévenin's equivalent circuit. (b) Resulting phasor diagram of the Thévenin's equivalent circuit.

The phasor diagram of the equivalent circuit is shown in Fig. 4.17b, where the voltage of the infinite bus is $\underline{U}_1 = \underline{U}_2 + (R + jX)\underline{I}$. With $\underline{U}_2 = U_2$ and the complex power absorbed at PCC, $\underline{U}_2 \underline{I}^* = (P - P_{\text{wec}}) + j(Q - Q_{\text{wec}})$, the voltage at the infinite bus can be expressed in terms of power as follows:

$$\underline{U}_1 = \left[U_2 + \frac{R(P - P_{\text{wec}}) + X(Q - Q_{\text{wec}})}{U_2} \right] + j \left[\frac{X(P - P_{\text{wec}}) - R(Q - Q_{\text{wec}})}{U_2} \right] \quad (4.27)$$

In distribution systems, the load angle δ (Fig. 4.17b) is usually small. Hence, the imaginary part of Eq. (4.27) is approximately zero, then $\underline{U}_1 \approx U_1$. As U_1 is constant, U_2 is only a function of P_{wec} and Q_{wec} , which can be expressed as:

$$R(P - P_{\text{wec}}) + X(Q - Q_{\text{wec}}) - (U_1 - U_2)U_2 = 0 \quad (4.28)$$

Similar to Weedy et al. (2012), the voltage variation at PCC due to differential varia-

tions in active and reactive powers can be calculated as:

$$dU_2 = \frac{R}{2U_2 - U_1} \cdot dP_{\text{wec}} + \frac{X}{2U_2 - U_1} \cdot dQ \quad (4.29)$$

In transmission grids, the X/R ratio is high, so the voltage is mainly sensitive to reactive power variations. However, in distribution grids, the X/R ratio is lower, and the PCC voltage is sensitive to variations in both active and reactive powers. Therefore, the natural fluctuations in the active power generated by a WEC lead to voltage fluctuations at the PCC. On the other hand, the sensitivity to the reactive power allows one to use the reactive capability of the WEC converter as a countermeasure to mitigate the voltage fluctuation.

4.3.1 Voltage Fluctuation Metrics

The voltage fluctuation for balanced networks can be quantified as shown by Viawan and Karlsson (2007). This concept can be extended to an unbalanced three-phase system as follows:

$$VF_{k\phi} = \frac{1}{T} \sum_{t=1}^T |U_{k\phi}(t) - U_{k\phi}(t-1)| \quad (4.30)$$

where $VF_{k\phi}$ is the voltage fluctuation index at bus k and phase ϕ along the T time samples. It is clear that, for a given period, the more the time samples, the lower the index. Therefore, the sampling time must be chosen as high as compatible with no significant loss of information on the voltage fluctuation.

In addition, a global metrics that quantifies the voltage fluctuation of the grid is presented by Ding et al. (2017), the system average voltage fluctuation index (SAVFI):

$$\text{SAVFI} = \frac{1}{N} \sum_{n=1}^N \text{SAVFI}_k \quad (4.31)$$

where N is the number of grid buses and SAVFI_k is the SAVFI per bus, an intermediate index defined by:

$$\text{SAVFI}_k = \frac{1}{n_\phi} \sum_{\phi \in \kappa} VF_{k\phi} \quad (4.32)$$

where n_ϕ and κ are, respectively, the number and set of phases at bus k .

4.4 Unbalanced Three-Phase Load Flow

The three-phase load flow can be carried out using two different approaches: the phase and the sequence reference frames. The performance of both formulations, in terms of computational effort, depends on the network topology and the unbalance degree. The phase frame approach deals with the *abc* representation of voltages and currents, while the sequence frame decomposes the system into positive-, negative- and zero-sequence networks, which are decoupled from each other in the case of a balanced system. Unfortunately, if a power system includes elements such as un-transposed lines, the network cannot be fully decoupled by means of symmetrical components. In spite of the absence of total uncoupling, the advantage of applying the sequence component approach instead of the phase component one is that the size of the problem is effectively reduced (Balamurugan and Srinivasan, 2011).

4.4.1 Three-Phase Load Flow Formulation

The load flow approach used in this thesis is based on a current residual formulation in the sequence frame as presented by Gómez-Expósito, Conejo, and Cañizares (2017) and Mayordomo et al. (2002). Its main characteristics are:

- The problem formulation is based on a nodal analysis by application of Kirchhoff's current law and uses the Newton–Raphson method based on current injection mismatches.
- Voltages and currents are written in rectangular coordinates and in the sequence frame to improve the sparsity of the Jacobian matrix.
- Several types of load are possible: constant power, constant impedance, constant current and rotating load, balanced or unbalanced in the delta or wye grounded configuration.
- The constraints for PV buses are active power and voltage magnitudes, specified in positive sequence.

This formulation assumes that the unknowns are the bus voltage magnitudes. Therefore, there are six unknowns at each node, three real parts and three imaginary parts of positive, negative and zero sequence voltages. Therefore, for a network with N buses, the total unknowns are $6N$, corresponding to $6N_g$ current mismatch equations for N_g PV buses and $6(N - N_g)$ current mismatch equations for PQ buses.

Then, the current mismatch equations can be written as:

$$\Delta \underline{I}_i^{(j)} = \underline{I}_{i_{ls}}^{(j)} + \underbrace{\sum_{k=1}^N \sum_{m=0}^2 \underline{Y}_{ik}^{(jm)} \underline{U}_k^{(m)}}_{\underline{I}_{ik}^{(j)}} \quad \text{with } i, k = 1, 2, \dots, N ; j, m = 0, 1, 2 \quad (4.33)$$

where:

- j, m are the indices for sequence components 0 (zero), 1 (positive) and 2 (negative).
- i, k are the subscripts for buses i and k , respectively.
- $\underline{I}_{ik}^{(j)}$ represents the total system current injected into bus i .
- $\underline{I}_{i_{ls}}^{(j)}$ is the demanded current at bus i for sequence j due to the loads and generators connected to the bus.
- $\underline{U}_k^{(m)} = U_{r_k}^{(m)} + jU_{x_k}^{(m)}$ represents the complex voltage at bus k and sequence component m .
- $\underline{Y}_{ik}^{(jm)} = G_{ik}^{(jm)} + jB_{ik}^{(jm)}$ is the sequence component admittance matrix element that connect bus i with bus k .

This formulation can be expressed in terms of real and imaginary parts to obtain the problem equations:

$$\begin{aligned} \Delta I_{r_i}^{(j)} &= I_{r_{i_{ls}}}^{(j)} + \sum_{k=1}^N \sum_{m=0}^2 \left(G_{ik}^{(jm)} V_{r_k}^{(m)} - B_{ik}^{(jm)} V_{x_k}^{(m)} \right) \\ \Delta I_{x_i}^{(j)} &= I_{x_{i_{ls}}}^{(j)} + \sum_{k=1}^N \sum_{m=0}^2 \left(B_{ik}^{(jm)} V_{r_k}^{(m)} + G_{ik}^{(jm)} V_{x_k}^{(m)} \right) \end{aligned} \quad (4.34)$$

The solution of Equations (4.34) is performed by a Newton–Raphson algorithm where the current mismatch must be forced to zero by means of an iterative process. The currents $\underline{I}_{i_{ls}}^{(j)}$ present a sequence voltage dependence according to the type of load/generator.

This approach represents a compact and flexible formulation, which has been validated in several transmission test systems with unbalanced loads and un-transposed lines (Gómez-Expósito, Conejo, and Cañizares, 2017). In the case of distribution networks, this approach faces a difficulty when two-phase and single-phase lines must be considered. These unbalanced laterals lead to an ill-conditioned positive sequence problem. To avoid this, two-phase and single-phase lines are modelled as three-phase lines by using one and two dummy lines and dummy nodes, respectively. The dummy nodes do not exist in the real network and are unloaded, so no current flows through the dummy lines, and the voltage drop across them is zero. Hence, there is no coupling between the

dummy lines and the other real phases of the line, and the problem is solved by setting the series impedance of the dummy line to any arbitrary value.

4.5 Voltage Impact

With the three-phase load flow formulation described in Section 4.4 and the average wave-to-wire model presented in Section 4.2 and shown in Fig. 4.18, it is possible to study the connection of the WEC to a distribution network and to assess the impact of its power injection on the voltage at different nodes. To illustrate its application, the following is an analysis in which the values of the voltages in all the nodes of a test distribution network are determined in two different cases: null and maximum power injection from the WEC. These cases correspond to extreme operation points that can be reached in just a few seconds.

Figure 4.18 Wave-to-wire model.

4.5.1 Test Distribution Network

The test distribution network is based on the 4.16 kV IEEE 13-bus test feeder (Kersting, 2001) shown in Fig. 4.19. The feeder presents an important unbalance not only due to the grid (two three-phase unbalanced laterals, two single-phase laterals and two two-phase laterals), but also to the unbalanced loads of different types (constant power, constant impedance and constant current). The test network includes the following modifications to the original one: the regulator is replaced by the delta-grounded wye three-phase substation transformer with taps in the primary winding adjusted to -7% , in series with the line segment of the original regulator; the distributed load along the line connecting buses 671 and 632 is modeled as two spot loads of half the total power at buses 671 and 632; and the WEC described in Section 4.2 is connected to bus 634.

It is relevant to mention that in the operation point with no power injection from the

Figure 4.19 Modified IEEE 13-bus radial distribution feeder (Kersting, 2001).

WEC, the 500 kVA transformer that connects buses 633 and 634 is highly loaded with a 494 kVA unbalanced load.

4.5.2 Network Voltage Profiles

The results of the analysis are shown in Fig. 4.20, where the black bars represent the per unit (p.u.) voltage for each phase at all the network buses when the active and reactive powers injected by the WEC are null. Note that buses 680 and 692 are not shown, since bus 680 is unloaded, and a breaker connects buses 692 and 671. Similar to Eq. (4.29) for the balanced case, an increase in the power injection from the WEC leads to an increase in voltage at the PCC, that also propagates to the rest of the buses. The detail for the unbalanced test network is represented with white bars in Fig. 4.20, which shows the case in which the WEC generates $P_{\text{wec}} = 500 \text{ kW}$, without reactive injection. Note that in this voltage analysis, the maximum generated power by the WEC has been scaled up to 500 kW; the rated power of the LPMSG is 10 kW.

4.6 Corrective actions Based on VAR Compensation

As shown in Section 4.2, the WEC connects to the grid through power electronic converters. The grid-side converter is responsible for keeping the dc-voltage U_{dc} constant by transferring the active power from the electrical generator to the grid. Although this

Figure 4.20 Voltage profiles of the test network with 0- and 500-kW injections from the WEC.

is its main task, it is important to note that the grid-side converter is also capable of interchanging (generating or absorbing) reactive power with the grid, within given limits that are defined by a maximum current, $I_{\max}$, or a Volt-Ampere (VA) constraint $S_{\max}$. Thus, the reactive limits depend on the active power injected by the WEC:

$$Q_{\max} = \sqrt{S_{\max}^2 - P_{\text{wec}}^2} \quad (4.35)$$

The dotted line of Fig. 4.21a represents these limits in the case of the converter used in this work. However, the real active power is bounded by the WEC rating (dashed line). Stability and other practical issues impose other limits on the reactive power (Teodorescu, Liserre, and Rodriguez, 2011), so that the real active and reactive capabilities of the converter are limited to the hatched area of Fig. 4.21a.

Figure 4.21 (a) active and reactive power operational limits of the the grid-side electronic converter of the WEC. (b) Voltage fluctuation at bus 634.

To illustrate the inverter's ability to generate or absorb reactive power according to

the generated active power, let us consider the power production from a regular wave:

$$P_{\text{wec}}(t) = \frac{1}{2}P_{\text{max}} \left[1 - \cos \left(\frac{2\pi}{T_p} \cdot t \right) \right] \quad (4.36)$$

where $P_{\text{max}} = 500 \text{ kW}$ and $T_p = 4 \text{ s}$ is the power period. The solid line of Fig. 4.21b represents the voltage fluctuation at PCC due to the injection of this fluctuating active power with no interchange of reactive power. The dashed and dotted lines represent the corresponding voltage evolution if the reactive power is set to its maximum and minimum limits, respectively. If the reactive power is set to operate in any other value within these limits, the voltage values would be within the area defined by these lines in Fig. 4.21b for this particular power profile. Therefore, any voltage trajectory within this area (for example, with less fluctuations) is reachable by controlling the reactive power. A desirable situation is to keep the voltage constant.

In real unbalanced systems, as the one described in this work, the bus voltages are different in each phase. On the other hand, the grid-side converter is composed of one three-phase inverter, which is only able to generate three-phase balanced reactive power from just one control variable. To implement this control, this work proposes to use the mean bus voltage defined as the mean of the rms voltage in each phase of a bus. For instance, the mean bus voltage of bus 634 is $U_{634} = (U_{634a} + U_{634b} + U_{634c})/3$.

As the electrical grid dynamics is much faster than the generated power dynamics, the distribution grid in Fig. 4.19 can be simulated as a quasi-static model, i.e., the voltage phasors in each bus and phase are calculated in steady-state for a given $P_{\text{wec}}(t)$ that varies in time. Finally, the amount of reactive power generated by the inverter is calculated in the reactive power control loop inside the VSC controller shown in Fig. 4.22. The plant is the power flow discussed in Section 4.4, where the outputs are the three-phase rms voltages. The PI controller has been tuned to be as fast as possible. Besides, as the reactive power is bounded, an anti-windup system has been implemented. Note that, in the reactive power control loop, the dynamics is only due to the PI controller and the anti-windup system.

Sometimes, the reactive capability of the inverter is not high enough to completely mitigate the voltage fluctuation. In this case, a STATCOM could be used to provide extra reactive power. In this work, a 50-kvar STATCOM is implemented as a reactive power load at bus 634. The combination of both inverter and STATCOM results, in practice, in an expansion of the operating area of the reactive compensation as shown in the grey area in Fig. 4.21a. In this case, the reactive power of the STATCOM is $Q_{\text{statcom}} = Q_{\text{PI}} - Q_{\text{wec}}$, with $|Q_{\text{statcom}}| \leq 50 \text{ kvar}$. Once the STATCOM reaches its limit, the voltage becomes uncontrolled. The STATCOM is only on when $Q_{\text{PI}} > Q_{\text{wec}}$, i.e., the grid requires more reactive power to keep the voltage level than the grid-side converter capability.

To illustrate this controller behavior, the sinusoidal power in Eq. (4.36) has been

Figure 4.22 Reactive power control loop.

injected into bus 634. The black solid line in Fig. 4.23 is the mean bus voltage fluctuation at this bus (U_{634}). The reference voltage $U_{634\text{ref}}$ controller is set to the value of U_{634} for the default network. If only the inverter is operating and its controller is enabled, U_{634} is kept at its reference value for almost all the time, as shown in Fig. 4.23 by the dashed line. After about 1.5 s, the inverter reaches its reactive power limit, and therefore, the voltage is uncontrolled and varies according to p_{wec} . On the other hand, when both the inverter and STATCOM are operating, the voltage fluctuation is completely mitigated as shown in Fig. 4.23 by the grey line. In this latter case, $Q_{\text{statcom}}(t = 2\text{ s}) = 25\text{ kvar}$, i.e., the STATCOM absorbs 25 kvar at this instant; therefore, it is half loaded.

Figure 4.23 Mitigation of the mean bus voltage U_{634} for a sinusoidal power generated by the WEC.

4.7 Simulation Results

In this section, the voltage fluctuation due to a very variable power injection into the grid, as well as the proposed actions to mitigate this fluctuation, are assessed. For this purpose,

the WEC is connected to the electrical grid as shown in Fig. 4.18 through bus 634, and it is excited by an incoming wave mathematically generated from the ITTC spectrum (The Specialist Committee on Waves (chaired by C. T. Stansberg), 2002), defined by its statistical parameters: significant wave height ($H_s = 2$ m) and energy period ($T_e = 8$ s). The electrical power delivered to the grid is shown in Fig. 4.24. As described in Section 4.6, the correction actions assessed are based on the generation or absorption of reactive power by the grid-side converter or the combination of both grid-side converter and STATCOM.

Figure 4.24 active power injected into the grid by the WEC.

The fluctuation of the mean bus voltage caused by the WEC on the PCC without VAR compensation is shown in Fig. 4.25 (grey line). The waveform of the voltage fluctuation follows the injected power waveform. When the injected power reaches its highest value, the voltage is about 1.7% higher than the default case. This value may seem low, and in addition, it is within permissible deviation ranges. However, areas with greater waves generate higher fluctuations. Furthermore, according to the first term in Eq. (4.29), the voltage increment depends on the the grid stiffness. Therefore, in weaker grids, with higher R , these effects are amplified. Fig. 4.25 also shows the mitigation effectiveness with the use of VAR compensation by the grid-side converter according to the proposed control strategy. The corrected mean bus voltage U_{634} (black line) is kept constant almost all the time. Only when the power excursion experiments high values, as between 400 and 450 s, the WEC is not able to keep U_{634} at its reference value.

The detail of the the mean bus voltage U_{634} within the interval between 400 and 450 s and the VAR compensation is shown in Fig. 4.26 (left side). The black line is U_{634} , and the grey line is the amount of reactive power injected by the grid-side converter. As the active power generated by the WEC drives U_{634} above its reference value, the grid-side converter operates absorbing (negative injection) more or less reactive power to mitigate the fluctuations. As previously commented, in this interval, the grid-side converter has reached its limit twice; then, during these short periods, the voltage is no longer controllable, and it varies according to the WEC active power. On the other hand, if, in addition

Figure 4.25 Mitigation of the mean bus voltage U_{634} with VAR compensation by the grid-side converter.

to the grid-side converter, the VAR compensation is complemented by the STacOM, the voltage fluctuations are completely mitigated as shown in Fig. 4.26 (right side). It is clear that the greater reactive power contribution is done by the grid-side converter, and the STATCOM just occasionally complements the VAR compensation. Indeed, in this scenario, the STATCOM is practically not necessary, since the VAR compensation by the grid-side converter results in a negligible voltage fluctuation.

Figure 4.26 Mean bus voltage U_{634} fluctuation with VAR compensation: by the grid-side converter (left); and by a combination of the grid-side converter and STATCOM (right).

So far, only the mitigation of the mean bus voltage has been assessed. However, this is not the real voltage. As an unbalanced system has been considered, each phase at each bus may have different fluctuations depending on the grid topology and the load level per phase. The voltage fluctuation may be assessed by the SAVFI per bus (Eq. 4.32) as shown in Fig. 4.27, where it is possible to see how the voltage fluctuation propagates along the buses. As expected, the PCC (bus 634) presents the highest fluctuations, and the combination of grid-side converter and STATCOM does not represent a significant improvement for the grid. However, it is interesting to have its usage in mind, since the WEC reactive capability may be reduced by the WEC commitment with some other grid requirements (Etxegarai et al., 2015), for example to comply with low-voltage ride-

through requirements. It is also easy to see that the buses farther from the infinite bus are more susceptible to fluctuations, and at the same time, they present a much better response to VAR compensation. Table 4.2 summarizes the SAVFI reduction per bus after VAR compensation. Only buses 651, 645 and 646 present a reduction lower than 50% due to a stiffer connection to the infinite bus. The global SAVFI, in turn, is reduced by 65.95%.

Figure 4.27 SAVFI per bus for the grid without VAR compensation, with VAR compensation by the grid-side converter and with VAR compensation by the combination of grid-side converter and STATCOM.

Table 4.2 $SAVFI_k$ (system average voltage fluctuation index) reduction after VAR compensation.

buses										
651	632	671	645	646	633	634	684	652	611	675
31.47%	50.61%	50.63%	31.20%	31.26%	72.65%	92.33%	71.35%	77.33%	65.01%	50.73%

Fig. 4.28 shows the per-phase voltage fluctuation at bus 634 without and with VAR compensation, where the left column is for the grid-side converter and the right one is for both the grid-side converter and STATCOM. As it was verified for the mean bus voltage, the STATCOM backs the grid-side converter up just during high peaks of power. In addition, the fluctuation reduction in phase-b is lower than phase-a and -c. In any case, the VF index achieves a great reduction. For example, at bus 634, the VF is reduced by 94.43%, 86.50% and 94.73% in phase-a, -b and -c, respectively.

Finally, to quantify the voltage unbalance in the test grid, the percent voltage unbalance factor defined as the ratio of the negative sequence voltage component to the positive sequence voltage component is used (Pillay and Manyage, 2001). Fig. 4.29 shows

Figure 4.28 Voltage fluctuations per-phase at bus 634 with VAR compensation: by the grid-side converter and (left column) by the combination of the grid-side converter and STATCOM (right column).

the voltage unbalance factor in each bus of the test grid for null active power injected by the WEC and for the maximum active power injected by the WEC. In addition, for the latter case, three different scenarios are shown: no VAR compensation, VAR compensation by the grid-side converter and by the combination of grid-side converter and STATCOM.

Figure 4.29 Voltage unbalance factor for $P_{wec} = 0$ and maximum active power injection from the WEC with: no VAR compensation, VAR compensation by the grid-side converter and by the combination of grid-side converter and STATCOM.

It can be seen that the test grid has an important voltage unbalance for $P_{wec} = 0$,

even with values exceeding the typical ones: 0.5–2% (IEEE Std 1159™-2009, 2009). The injection of a balanced active power by the WEC contributes to a small reduction of the voltage unbalance as shown in Fig. 4.29 for the maximum P_{wec} . The consumption of reactive power by the grid-side converter due to the corrective action adopted to mitigate voltage fluctuations results in a very small increase of the unbalance voltage factor with respect to the case without VAR compensation. Furthermore, the effect of an extra VAR compensation by the STATCOM on the unbalance voltage is insignificant in this case. Considering the unbalance voltage along time, Fig. 4.30 shows the behavior of the unbalance voltage factor at PCC according to the injection of active and reactive powers by the WEC. The mean of the voltage unbalance factor within the period (500 s) is 1.928% without VAR compensation, whereas it slightly increases to 1.933% if the proposed fluctuation compensation technique is applied. As conclusion, the technique is effective at compensating voltage fluctuations without practically affecting the unbalance. Further research is suggested to expand the capability of the technique to incorporate improvements on the voltage unbalance.

Figure 4.30 Instantaneous voltage unbalance factor at bus 634.

Artificial Neural Network Approach for Wave Prediction

One of the challenges in the field of wave energy is to improve the efficiency of its conversion into electrical energy. As a rule, the wave energy converter (WEC) should operate in an ideal operating point (Ilyas et al., 2014). This is an essential factor to make wave power plants able to recover the investments through fuel cost savings (Glendenning, 1978). One way to achieve this improvement is to maximise the power generated by a WEC. In this sense, many control strategies have been proposed – some of them are general and others specific to particular devices. The main strategies are commented in Chapter 3 and more extended comparisons among them can be found in Hals, Falnes, and Moan (2011a) and Xie and Zuo (2013). It can be seen that, for example, to implement discrete controls, latching and declutch, the prediction of the excitation wave force is required.

For feasible exploitation of the wave energy, large wave farms will be required, like wind energy. So, in the case of point absorbers, a wave farm would be composed of tens or hundreds of WECs. These amount of converters result in the so-called park effect (Babarit, 2013), where the total power of the farm is less than the total power produced by one WEC multiplied by the number of WECs in the farm. Hence, it is even more critical to ensure the WEC operation is in, or very close, to the ideal operating point. In this sense, the improvement of the efficiency of every single converter is closely related to the wave prediction in a short time horizon. Cruz et al. (2010) presented another approach where the wave farm controller must take into account two aspects: to maximise the effects of constructive interferences of the waves caused by the WECs and to minimise the effects of its destructive interference. At this point, it may be possible that prior knowledge of the incoming waves contributes to better maximisation results. In addition, Korde (2013) points out the importance of up-wave surface elevation knowledge to provide a smooth hydrodynamic control.

So far, wave prediction has proved of great importance to the development of wave energy technologies. The wave prediction can be mainly carried out by two methods: **(i)** the first one is based on single-point measurements, where the WEC is responsible for recording the observed wave and, also, to provide a wave prediction for its own location; and **(ii)** the second one is based on a spatial reconstruction of the wavefield with available up-wave information of one or more nearby sensors.

According to Fusco and Ringwood (2010), the approach **(i)** has considerable advantages with respect to **(ii)**, such as simpler forecast model and less amount of instrumentation required. They present different models for short-term wave prediction, among them are cyclical models, extended Kalman filter, autoregressive (AR) model and neural networks. All of them provide a wave prediction within a limited horizon time by using past wave elevation data at the WEC location. They conclude that a simple AR model is enough to provide a very accurate prediction. Nevertheless, their approach applies a low-pass filter to the WDTs by a zero-phase filter, i.e., no phase shift is added and, therefore, the original signal is preserved. So, for real-time application, an adequate filtering process must be studied.

The approach **(ii)** lies in the deterministic sea wave prediction (DSWP). This technique is capable of predicting the sea surface propagated by measuring the sea surface elevation at some distance (Belmont et al., 2006). This prediction is usually within a time horizon of tens of seconds. Belmont et al. (2006) use linear filter to provide the dispersive phase shift to the DSWP model. The filter methodology reduces the computational effort compared to other methods as the one developed by Hilmer and Thornhill (2015). On the other hand, Tedd, Frigaard, et al. (2007) use digital filters to translate the pressure and wave phase measured at the up-wave sensor to determine the resulting surface elevation and phase of the wave propagation at the WEC. According to Abusedra and Belmont (2011), DSWP models are valid when characteristics of the measured wave are predominant at the prediction site. They also presented a space–time diagram that provides a guideline to assess a DSWP model in terms of prediction time, prediction space and accuracy.

X-band marine radar has been used to obtain the sea surface data for DSWP models as in the cases of Hilmer and Thornhill (2015), Lund (2006), and Simpson et al. (2020). This technology scans a sea area and it provides high-resolution images of the sea surface in both spatial and temporal variations. In addition to X-band radars, that operate with electromagnetic microwaves, LIDAR technology can also be found for this application. A sea surface image is formed from laser pulses (Belmont et al., 2007; Nougier, Grilli, and Guérin, 2013). Another way to obtain these data is from up-wave sensors, as the case presented by Monk et al. (2013), where a hydrostatic wave pressure sensor (a near-shore air-chamber at about 60 meters of distance) was used to determine the sea surface elevation at this point.

Nevertheless, what if the sea surface elevation measurements of the WEC and up-

wave sensor are available? This is the case of Pico OWC shown by Monk et al. (2013). They use artificial neural networks (ANN) to forecast the chamber hydrodynamics from the hydrodynamic data measured by the hydrostatic wave pressure sensor. As a result, hydrodynamic forecasts present a satisfactory accuracy.

On the other hand, Paparella et al. (2014) focused on the sea surface elevation only. They brought together the AR model for short-time prediction (Fusco and Ringwood, 2010) and the model used to predict the sea surface elevation at the OWC from the up-wave measurements and, finally, built an autoregressive with exogenous (ARX) input model. It is important to highlight that the contribution of extra data to the AR model, in this case, the exogenous input, might make the prediction model more robust and then improve forecast accuracy.

A measurement system for every single WEC would have a significant impact on the cost of the farm. Facing this scenario, this thesis proposes the instantaneous up-wave forecast within a wave farm from a measurement spot employing an ANN model. Using this tool as part of a prediction system would allow the deployment of just one measurement and prediction system for a farm. This approach is justifiable in regions where the wave climate presents small changes and a well-defined swell component. Thereby, a wave measurement campaign should be carried out to provide different possible data scenarios and, thus, to avoid the development of new models.

ANNs have been widely used for system identification presenting almost always optimal results. As the “real” model between the measurement spot and any spot at the wave farm is unknown, the ANN can be responsible for building the nonlinear dynamic model from input/output wave data and then to estimate the sea surface elevation within a wave farm. The ANN architecture is usually represented by the scheme [(a)-(b)-(c)], known as multi-layer network, where (a) is the input layer, (b) represents the hidden layers, and (c) the output layer. This architecture is a universal function approximator (Hornik, Stinchcombe, and White, 1989) because can be easily adapted to process time series by introducing a memory structure to the model. This last structure was first introduced by Waibel et al. (1989) for phoneme recognition where each weight is associated with a delay. Thus, an interesting ANN architecture is designed taking into account the temporal aspects of the input: the time-delay neural network (TDNN). In addition to this, there are ANN architectures more robust for nonlinear identification systems, as it is the case of the recurrent neural networks (RNN). More specifically, the Nonlinear Autoregressive models with exogenous inputs (NARX) (Lin et al., 1996), which showed to be well suited for this task in many different fields, including time series. NARX models have been applied for nonlinear identification in different fields (Deng, 2013; Prasad and Bequette, 2003; Tijani et al., 2014). The input of the NARX model, besides taking into account past values of the input signal, includes past values of the output. However, the central idea of this study is to estimate instantaneously the sea surface elevation within

a wave farm with the minimum possible measurements: only one. Thus, this condition rules out any RNN, despite having more accurate results.

5.1 Artificial Neural Network

The most simple structure of an ANN is the perceptron, presented by McCulloch and Pitts (1943), in which there is just a single unit (artificial neuron) able to classify objects according to their classes. The artificial neuron (or cell-body) is composed of a sum block and an activation function and its response, therefore, it is a function of the weighted sum of all inputs and a bias. This activation function can be defined in different forms, such as linear, step, ramp, sigmoid, hyperbolic tangent and gaussian, according its application. Fig. 5.1 shows the model of the perceptron neural network, where the subscript i corresponds to the number of inputs x (1 to m) and w is the weight related to each input and neuron. The output y can be written in both scalar and matrix forms as shown in Eq. (5.1). Note that, in this case, the bias b is translated to the input as $x_0 = 1$, so the weight w_0 works as the bias.

Figure 5.1 Perceptron neural network.

$$y = f \left(\sum_{i=0}^m w_i x_i \right) = f(\mathbf{w}^T \mathbf{x}) \quad (5.1)$$

A combination of two or more neurons results in a network with one or more layers. This is the most common neural architecture known as multi-layer perceptron (MLP). A MLP is a feed-forward network composed of perceptrons arranged in layers. These layers are: an input layer, one or more hidden layers, and an output layer. Therefore, a MLP has 3 layers minimum. In a general way, the MLP is illustrated in Fig. 5.2 with one hidden layer. Note that the neuron structure is represented by a filled grey circle for simplification. Following the same methodology to determine the perceptron, the MLP outputs can be written as shown in Eq. (5.2), where: h is the number of neurons on in the hidden layer; k is the number of outputs; f and g are the activation function of the

output and hidden layers; $w^{(1)}$ and $w^{(2)}$ are the weights of the hidden and output layers. Again, each neuron input with subscript 0 is assumed to be 1, so its associated weight corresponds to the bias. In this network, the sigmoid and the linear functions are used in the hidden and output layers, respectively.

Figure 5.2 MLP structure with three layers: input, hidden and output.

$$y_k = f \left[\sum_{j=0}^h w_{kj}^{(2)} \cdot g \left(\sum_{i=0}^m w_{ji}^{(1)} x_i \right) \right] \quad (5.2)$$

5.1.1 Dynamic Neural Architectures

If the perceptron is adapted to receive a temporal information, their independent inputs are changed into time-lagged inputs as shown in Fig. 5.3, where $y[k]$ is the current data, z^{-1} is the unit delay, w_{ji} is the weight from the i th input neuron to the j th neuron ($j = 1$ for the perceptron), and $f(\cdot)$ is the activation function. Thus, the response to a tapped delay line input is given by:

$$y[k] = f \left(\sum_{i=0}^p w_{ji} x[k-i] + b_j \right) \quad (5.3)$$

In addition, if the activation function of the ANN in figure 5.3 is linear, one can conclude that this model is a neural equivalent to the ARX model presented by Paparella et al. (2014) with no poles. Several dynamic neural models can be found in Haykin (2001).

Figure 5.3 Perceptron with a time-lagged input.

Time Delay Neural Network (TDNN)

The TDNN architecture is based on a typical MLP feedforward network, but with a tapped delay line. In addition, most of TDNN systems also present feedback inputs. TDNN has been mostly used to speech recognition applications (Sugiyama, Sawai, and Waibel, 1991).

Nevertheless, the TDNN application has been extended to other fields as commented before. For example, it could be applied to predict wind speed, solar radiation, rainfall, traffic flow, etc. (Abhishek and Misra, 2016; Benmouiza and Cheknane, 2016; Shamshirband, Rabczuk, and Chau, 2019; Sulaiman and Wahab, 2018; Sushmitha et al., 2019; Wang, Tsapakis, and Zhong, 2016). Also, TDNN could be used to dynamic system identification, i.e., to determine an output for a given input. Tijani et al. (2014) shows that, despite the high nonlinearities presented in an unmanned helicopter system, TDNN can lead to very accurate models. Moreover, other techniques can be used with TDNN to improve its performance as commented by Coban (2013) and to reduce model as shown by Prasad and Bequette (2003).

In this sense, this thesis seeks to identify the nonlinear model of the wave-field propagation. The model must be able to estimate the sea surface elevation at any point within an area from a measurement of the sea surface elevation at a single spot. Therefore, each estimation spot has its own estimation system with a single signal input from the measurement spot.

The TDNN structure for the estimation model is presented in Fig. 5.4. This is a

typical MLP feedforward network, but with a tapped delay line. The estimated value $\hat{y}[k]$ depends on the previous values of the input $x[k]$. Therefore, the system can be described by:

$$\hat{y}[k] = f(x[k], x[k-1], x[k-2], \dots, x[k-p]) \quad (5.4)$$

where p is the tapped delay line length. This architecture is modelled as a composition of weighted sum as follows:

$$\hat{y} = f \left[\sum_{j=0}^h w_j^{(2)} \cdot g \left(\sum_{i=0}^m w_{ji}^{(1)} x_i \right) \right] \quad (5.5)$$

Note that Eq. (5.5) is the same as Eq. (5.2) reduced to a single output. In this network, the sigmoid and the linear functions are used in the hidden and output layers, respectively.

Figure 5.4 Feedforward neural network structure with a tapped delay line.

The learning algorithm used is the Levenberg-Marquardt (Marquardt, 1963), that is a usual optimisation technique to solve nonlinear least squares problems, which, in this case, results in the adjustment of the network weights. The learning method must minimize a

cost function, for example, the mean square error (MSE):

$$\text{MSE} = \frac{1}{N} \sum_{k=1}^N (y[k] - \hat{y}[k])^2 \quad (5.6)$$

In addition, it can be said that this learning algorithm has been shown to be fast, accurate and efficient, even applied to a simple structure of the ANN model as verified by Mirzaee (2009), where a simple configuration was used to predict a chaotic time series.

5.2 TDNN Application to Wave Estimation

To assess the quality of the wave estimation in different spots within a wave farm from a measurement spot, wave data series for long-crested seas have been mathematically generated for a duration of 2000 seconds, sampled at 2 Hz, with the following wave statistical parameters: $H_s = 2$ m and $T_e = 8$ s. In addition, the wave data series have been generated with different wave front angles. Fig. 5.5a shows the wave farm topology analyzed in this work, whose area is 8000 m². The estimation spots are numbered from 1 to 100 and the measurement spot is marked with a cross. The spots stand 10 m apart from each other in both axis, x and y . The TDNN is then configured and trained to estimate the sea surface elevation within the wave farm. The first 1000 s of the wave data are used to train the TDNN and the remainder is used to test it and to evaluate its performance. As mentioned before, the Levenberg-Marquardt algorithm is used to train the TDNN and the training results are evaluated by the MSE. The training stops when, during the validation process, the MSE starts growing.

Once the TDNN is trained, it is tested with the other 1000 s of the wave data. The estimation at each spot can be assessed by comparing the estimated with the pseudo-measured data. The assessment is carried out according to the R^2 value which is calculated as follows:

$$R^2 = 1 - \frac{SS_{res}}{SS_t} \quad (5.7)$$

where $SS_t = \sum_k (y_k - \bar{y})^2$ is the total sum of squares, $SS_{res} = \sum_k (\hat{y}_k - y_k)^2$ is the residual sum of squares, y is the measured data, $\bar{y}$ the mean of the measured data and $\hat{y}$ is the estimated data. The higher R^2 value (the closer to 1), the better estimation (Samarasinghe, 2016). So the R^2 value is calculated for each estimation spot and, to evaluate the overall estimation in the wave farm, the average of all R^2 values is also calculated.

(a)

(b)

Figure 5.5 (a) Wave farm topology with one hundred estimation spots from a measurement spot (marked with a cross). (b) Incident wave fronts.

5.3 Effects of the ANN Configuration

In any ANN architecture, its configuration plays a key role in the performance. The main drawback of the ANN is its black box model where there is no control on the calculation. So, usually, the best parameters are obtained after several trials by selecting the network with the best performance. In this sense, this section studies the effect of the number of neurons and tap delays on the estimation performance. Assuming that the wave front is perpendicular to axis- x , i.e., it is at $\chi = 0^\circ$, all the spots along axis- y have the same sea surface elevation, since a long-crested wave is considered. Therefore, in this section, the TDNN is trained with the spot array from 41 to 60 and the measurement spot at the position (0, 20).

The ANN has been trained and tested to between 1 and 50 hidden neurons, and to between 10- and 100-tap delays. Sweeping all configurations, the average R^2 value is calculated from each spot estimation and it is shown in Fig. 5.6a. It is verified that the longer the tap delay, the better the estimation. For short tap delays, the results indicate a nonlinear relationship for the model. Besides, it is possible to see that the improvement of the R^2 value for tap delays longer than 80 is not significant, what leads to conclude that a much longer tap delay line results in a higher computational effort without a considerable estimation improvement. In addition, for each tap delay curve in Fig. 5.6a, it is seen that, from a certain number of hidden neurons, the estimation performance is not affected.

The estimation performance is not only affected by the TDNN configuration, but also by the distance of the estimation spot from the measurement spot. According to Fig. 5.6a, the best TDNN architecture is [100-20-20], i.e., one hundred neurons in the input layer (tapped delay length); one hidden layer with twenty neurons, since, for the best fitted model (100-tap delay), this number of hidden neurons is enough; and one output layer with twenty neurons (the spot array from 41 to 60). For this TDNN configuration, Fig. 5.6b shows the R^2 values with respect to the number of tap delays and the distance from the measurement spot. It is clear that the closer the estimation spot to the measurement spot, the better the estimation. However, for taped delays longer than 80, it is possible to get a very good estimation at the farthest spots. Thus, this TDNN architecture ([100-20-20]) shows to be adequate to estimate the wave surface elevation in the wave farm from a measurement spot.

5.4 Sensitivity Analysis of the Wave Estimation

In this section, the sensitivities of the the wave estimation with respect to the wave front angle and the sea state are analyzed. To assess the estimation for a given spot in the wave farm, the TDNN architecture with twenty hidden neurons is used. However, unlike the previous model, the output layer has one hundred neurons (the total number of estimation

Figure 5.6 Evaluation of the TDNN architecture. (a) Average R^2 for the number of hidden neurons and tap delays. (b) R^2 value with respect to the number of tap delays and the position in the wave farm.

spots). In other words, the new TDNN is [100-20-100]. The first analysis is carried out with three wave data series whose wave front angles are at 0° , 45° and -45° with respect to the axis- x , as shown in Fig. 5.5b. In addition, a comparison between the TDNN with 100- and 70-tap delay line is presented. Note that, for the 70-tap delay, some estimation spots present R^2 values below 0.8, as it can be seen in Fig. 5.6b. The TDNN has been trained for each wave front angle and for both tapped delays.

As a result, Fig. 5.7 shows the R^2 value for each scenario. It can be seen that, for 100-tap delay, the wave estimation has a very good quality for all of the three wave front angles. Also, it can be said that for different angles, the estimation is not as good as for 0° , but it still has a good estimation quality. For example, Figs. 5.7c and 5.7e show the R^2 values for wave front angles at 45° and -45° being mostly around 0.9, except for the extremes where the incoming wave reaches the estimation spot before the measurement

Figure 5.7 R^2 value for the wave estimation in the wave farm with 100- and 70-tap delays, and wave front angle at 0° , 45° and -45° .

spot. An interesting fact is that those extreme spots present a better quality estimation for the 70-tap delay, but the overall quality of the estimation is better for the 100-tap delay line. Note that, for wave front angles at 45° and -45° , the R^2 values might be symmetrical. However, it is not perfectly symmetrical, as in the case of Fig. 5.7, because for each training of the TDNN, different solutions are given.

Finally, the worst estimation spots are selected by comparing Figs. 5.5a and 5.7. For a wave front angle at 0° , they are 20, 40, $\dots$, 100, and for wave front angles at 45° and -45° , they are 1 and 81, respectively. The scatter plot of each spot for 100- and 70-tap delay is shown in Fig. 5.8 where the line represents the perfect estimation, i.e., $y(t) = \hat{y}(t)$. It is seen that all of those worst cases present a dispersion-cloud around the line with no great deviations in the estimation results.

Figure 5.8 Scatter plot for the worst estimation spot.

5.4.1 Wave Front Angle

To evaluate how the wave front angle affects the wave estimation, the TDNN is firstly trained for a wave data series in which the wave front angle is 0° . Then, wave data series have been generated with wave front angles between -45° and 45° within the wave farm in Fig. 5.5a. The average R^2 value of the wave estimation in the farm is shown in Fig. 5.9. Note that R^2 value is symmetrical as said in Section 5.4. In addition, it is possible to conclude that the wave estimation within the wave farm presents a high dependency on the main direction of the front wave. Nonetheless, under these conditions, it is possible to say that for wave front angle deviations up to $\pm 10^\circ$, the estimation quality is still very good. For example, for wave front angle -10° , the average R^2 value is about 0.9. Note that a wave farm must be settled in a sea area previously studied with several weather conditions recorded. Based on this, it would be possible to have the ANN model trained

for different scenarios of directional spectrum.

Figure 5.9 Sensitivity of the wave estimation to the incident wave front angle.

5.4.2 Sea State

Although the wave statistical parameters tend to remain constant for a long period of time, they can change, albeit slightly. In other cases, they may undergo major changes over a day, for example, during a storm. So it is important to consider how these changes may affect the wave estimation.

First, a TDNN is trained for wave data series whose wave statistical parameters are $H_s = 2$ m and $T_e = 8$ s with a wave front angle at 0° . Then, several wave data series with the same wave front angle, but different wave statistical parameters have been generated. After estimating the sea surface elevation within the wave farm, the average R^2 value is calculated for each combination of significant height and energy period. These results are shown in Fig. 5.10a, where it can be seen that most of the average R^2 values are between 0.9 and 1, with a minor portion from 0.8 to 0.9. Thus, one can conclude that the overall estimation is quite satisfactory. On the other hand, if the same TDNN is applied to estimate wave series with different wave statistical parameters and different wave front angles, the estimation shows a worsening compared to the previous one, as it can be seen in Fig. 5.10b for the case of 10° . None average R^2 value is close to 1, and, for the most part of the farm, the values are between 0.6 and 0.9.

5.5 Short-Crested Sea Estimation

Despite the fact that in many cases short-crested seas can be ignored because they do not much affect a floating structure motion (Lloyd, 1989) and the wave farms might be located in a dispersion zone, outside the fetch area (see Fig. 2.5), where there is almost no energy dispersion, in this section, the quality of the wave estimation is assessed with respect to the degree of wave spreading. The wave data series generated for this analysis

Figure 5.10 Sensitivity of the wave estimation to the statistical wave parameters for an ANN trained with a wave front at 0° .

have the same characteristics as the case of long-crested seas. However, here, the waves have more than one propagation direction.

In a first analysis, the wave spectrum has $\theta_1 = 0$ and the energy is spread in one direction θ only, that is varied from 5° to 85° . The sensitivity of the estimation to θ is shown in Fig. 5.11a, where it is verified that the best estimation is achieved for wave spreading close to the main wave propagation direction and its orthogonal direction. Similarly, the estimation quality when the energy spreads in two symmetrical directions is shown in Fig. 5.11b. The results are similar to the obtained for the one direction case, but the average R^2 value found is slightly worse. However, for both cases, directions of wave spreading up to 15° present a very good estimation with average R^2 values greater than 0.9, in general.

In addition to these cases, the estimation quality is assessed for a chaotic sea, i.e., the wave farm is assumed to be inside a fetch area. In this scenario, the energy spreads

Figure 5.11 Estimation quality in a wave farm for a sea state with wave spreading.

in many directions. Fig. 5.11c shows the R^2 values over the wave farm. It is seen that beyond 100 m in the axis- x the estimation is not satisfactory. The estimation quality can be also evaluated by Fig. 5.12, where the scatter plot of the estimation is presented. The dispersion-cloud becomes even wider, the farther from the measurement spot the estimation spot is. For example, at spot 41, the dispersion-cloud is over the line $y(t) = \hat{y}(t)$, while at spots 50 and 60, it is spread out from the line $y(t) = \hat{y}(t)$. As a conclusion, even in chaotic seas the instantaneous wave estimation can be satisfactory depending on how far the estimation spot is from the measurement spot.

Figure 5.12 Scatter plot for the chaotic sea.

Conclusions and Future Works

A good control strategy applied to a WEC is important to make its operation efficient. In fact, many control strategies have been developed for different types of WEC to maximise the absorbed power. The challenge in designing these types of controllers is to tune them by taking into account the non-deterministic characteristics of the waves.

The proposed method, that is defined by mapping the maximum power point for a combination of wave heights and periods for regular waves and applying the optimal values of the load seen from the electrical generator to irregular waves improves the efficiency of the system compared to almost any other electrical load. Therefore, it proved effective when applied to real seas so long as the statistical parameters are identified. Simulations with different wave profiles have shown that the maximum generated power is around the optimal resistance previously identified, and that the $P_{\text{ratio}}-R_{\text{em}}$ curves of an irregular wave and the ideal case have similar behaviour. In addition, the proposed control strategy has been applied to real waves. Despite the fact that the sea states of the real WDST are slightly different from those of the synthetic WDST, the $P_{\text{ratio}}-R_{\text{em}}$ curves are very similar both in magnitude and in shape, showing the validity of the proposed strategy for real seas.

The maximum generated power by the LPMSG is obtained by leading the PTO force to an optimal power point. The PTO force is directly controlled by the PTO current. Therefore, to carry out the proposed method, a current controller is required. Two standard current controllers have been implemented: hysteresis-band current controller and space-vector PWM current control. The latter is implemented by using the voltage-oriented control approach.

In addition, a novel nonlinear vector current source has been proposed to implement the current controller. The cycle time of the current source algorithm is constant, and thus predictable, which makes programming easier in a microprocessor. Also, the algorithm works in both generator electrical sequences. The fast response of the nonlinear current

source turned out to be very suitable to impose the force (or torque, in case of rotary machines) and magnetic field references in wave energy applications. This is even more relevant if it is considered that in these applications the translator speed experiences large variations that must be controlled in order to impose the maximum efficiency speed on the PTO.

The comparison between the implemented current controllers shows that all of them are very sensitive to the filter applied to the voltage measurement. A lousy filter choice can result in a decrease in system efficiency. On the dynamic response, both, HBCC and NLVCS, generate currents with the same waveform as the ideal one, while the PI-based controller, SVPWMCC, presents a slower dynamic response, and it takes time to establish the expected current.

On the other hand, despite the incipient development status of WEC technologies, some of them have been already connected to the electrical grid in test sites and more are expected in the near future. Thus, studies on the impact of their connection and compliance with grid requirements are needed. The WEC locations and their power level lead to a connection to distribution grids, which are normally unbalanced and more sensitive to active power variations than transmission grids.

In this work, a detailed wave-to-wire model has been developed. This model provides a complete response of the system for both, generation and grid connection. However, for grid connection studies, the switching detail is not necessary in many cases and, then, an average model can be used. This brings a great advantage, since the simulation time is greatly reduced. For this reason, parallel to the detailed model, an average model has also been developed and validated.

The connection of a WEC to an unbalanced distribution grid has been analysed showing that the oscillating active power injection from the WEC leads to a similar fluctuation in the voltage of the different phases and buses. A strategy for mitigating these fluctuations based on using the VAR capability of the grid-side converter of the WEC has been proposed. As grid-connected converters are typically three phase, the reactive power control is made by means of just one control variable. The proposed mean bus voltage has been demonstrated to be a simple yet effective choice. In some cases, as the one presented in this paper, the WEC converter is able to mitigate the voltage fluctuations. In other cases, a STATCOM can be used to provide an extra reactive support that, in addition, might be useful if the WEC is committed to other grid requirements that would reduce the WEC reactive capability.

Finally, it is often important to know the incoming wave in a short-term future within a wave energy field. This knowledge could be used for control strategies to maximise the power generated by a unit (only one WEC). It could also be useful for the farm controller to accomplish some grid requirement. As a wave farm can be composed of tens or perhaps hundreds of WECs, it may be very expensive and unpractical to have a

measurement system for each one. In this sense, this thesis presents a study about the spatial wave estimation in a wave farm from a single buoy.

The TDNN has been used to model the behaviour of the waves between the measurement and the estimation spots. This ANN architecture seems to be quite satisfactory for this applications, since it is able to accurately estimate the sea surface elevation in many spots within a wave farm from just one point, even if the farm is inside a fetch area. The influence of the distance of the estimation spot from the measurement buoy and the tapped delay length has been assessed. Naturally, the estimation quality is worse as the distance from the measurement buoy increases. However, depending on the tapped delay length, a very good accuracy in the estimation can be achieved. On the downside, a longer tapped delay leads to an increasing in the TDNN training time. In addition, it is fair to say that its accuracy is highly dependent on the wave front direction and somewhat of the wave statistical parameters. But both training time and its dependence should not be such a problem, since wave farm sites, as in the case of wind farms, should be thoroughly studied before its exploitation and, then, the TDNN must be trained for each possible scenario in order to make this method as effective as possible.

6.1 Contribution of This Thesis

The main contributions presented in this thesis are listed below:

1. Maximisation of the power generated by the WEC.
  - (a) The WEC model connected to a resistive load was published in (Mendonça and Martinez, 2013).
  - (b) A resistive loading strategy is proposed by defining a relation between the electrical power and a resistance connected to the LPMSG terminals. This contribution was published in (Mendonça and Martinez, 2016a).
  - (c) A novel nonlinear vector current source has been proposed to implement the current controller for wave energy application. This contribution was published in (Ramirez et al., 2019).
  - (d) The assessment of three different current controllers with respect to efficiency and dynamic response of the WEC. It has been showed that the nonlinear controllers are best suited. This contribution was published in (Mendonça, Martinez, and Ramirez, 2015; Mendonça and Martinez, 2016b).
2. WEC connection to the grid.
  - (a) A detailed and average representation of the wave-to-wire model. Pending publication.

- (b) Mitigation of the voltage fluctuation caused by a WEC connected to a distribution grid. This contribution was published in (Martinez et al., 2013; Medina, Mendonça, and Martinez, 2016; Mendonça et al., 2017).

3. Wave estimation.

- (a) A up-wave estimation model based on ANN architecture. Pending publication.

6.2 Future Works

There is still much to research in the field wave energy. So, some suggestion of ideas that can be considered for future work or continuation of this thesis are listed below:

- The control strategy proposed in Chapter 3 can be adapted to other time base, for example, a wave-by-wave control. The optimal values could be applied to the harmonic components of the wave.
- The current controller of the generator-side converter could replace the VOC for the FOC. Thus, this controller could perform torque and field control and prevent damage to the LPMSG. Also, to assess the influence of the said control on the WEC efficiency.
- The proposed generator-side converter is able to exchange reactive power. A future work could adapt the proposed resistive loading control to a reactive control.
- Although simple, but functional, the proposed strategy to mitigate voltage fluctuation by using a three-phase balanced reactive compensation could be replaced for an unbalanced compensation. Results should be assessed.
- A study on the suitable filter for WEC connected to weak grids.
- The WEC support to other ancillary services can be studied. Such as: frequency control and voltage unbalance mitigation.
- Implementation of a measurement system with RADAR or LIDAR to provide real sea surface information within an area.
- To assess the up-wave estimation proposed in Chapter 5 combined with a short-time forecast model.

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